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RCA REVIEW

SIMULTANEOUS ALL-ELECTRONIC COLOR TELEVISION*

A Progress Report

BY

RCA Laboratories Division, Princeton, N. J.

Summary—*This paper presents the latest progress report on* ***Color Television*** *and the first on the new simultaneous all-electronic system.* ***Basic*** *design and operating characteristics are reviewed. The apparatus for scanning color slides and color motion picture film together with the color television receivers are described.*

IN OCTOBER and November, 1946, the Radio Corporation of America gave several demonstrations of color television to press, industry and Government groups. These demonstrations constituted a progress report on the work done in color television, which follows the program announced at the time of earlier demonstrations in December, 1945.[1] In the current demonstration, important advances in color television were shown. The new system is *all-electronic*, having the potential flexibility inherent in electronic arrangements, and *simultaneous*, all three color images being transmitted continuously. This system has many operating and performance advantages and is compatible with the present black-and-white television. Since each of the three color channels employs the same standards as those now in use for black-and-white transmission, the green channel is suitable for monochrome presentation. Color television of this type can be introduced at any time it is made ready and can be operated interchangeably with black-and-white television; undesirable obsolescence is not created.

The recent demonstrations included television pictures in natural color scanned from kodachrome slides and from 16-millimeter color motion picture film. In order to demonstrate interchangeability, pictures in monochrome using signals of present black-and-white standards were shown on the color receivers; pictures in monochrome, using signals of the simultaneous color transmission, were then demonstrated on a current model black-and-white receiver.

Research work is under way and progress is being made in the radio transmission and reception of simultaneous all-electronic color

* Decimal Classification: R583.

[1] R. D. Kell, G. L. Fredendall, A. C. Schroeder, and R. C. Webb, "An Experimental Color Television System", *RCA REVIEW*, Vol. VII, No. 2, pp. 141-154, June, 1946.

television and in the building of television cameras for studio and outdoor pickup of this system. This work, together with propagation tests and field surveys, is a part of the over-all schedule yet to be fully worked out, but already well along.

Since simultaneous all-electronic color television is of far-reaching importance, the experimental equipment used during the recent demonstrations is described herein. This includes the apparatus for scanning color slides and color motion picture film together with the television receivers for color. Some of the basic design and operating characteristics are also reviewed. Figure 1 is a block diagram of the system.

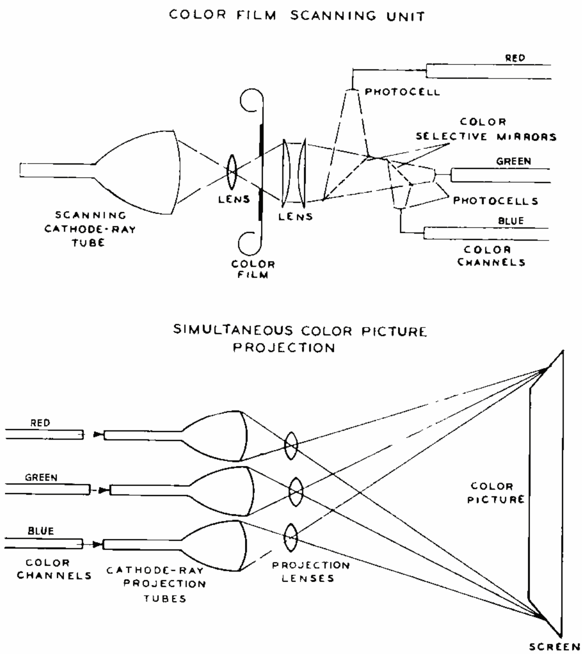

Fig. 1—Block diagram of the simultaneous all-electronic color television system. (Each of the color channels, shown at the right of the upper figure, have the same operating standards as the current black-and-white system. At the receiving end, shown at the left of the lower figure, each color channel is associated with its separate cathode-ray projection tube.)

Stationary Picture Signal Generator

One of the primary needs for the development of a simultaneous color television system is a standard source of tricolor video signals on which one may rely for good resolution, good registration, high signal-to-noise ratio, freedom from spurious signals, and good color fidelity. A special slide scanner utilizing a cathode-ray tube as a flying-spot scanner, a beam splitter, and three photoelectric tubes, were developed for this purpose.

A photograph of this apparatus with a superimposed phantom view of the kinescope and the light paths is shown in Figure 2. The raster formed on the screen of the kinescope is imaged on the slide by means of a lens. The light rays transmitted by the slide are condensed and

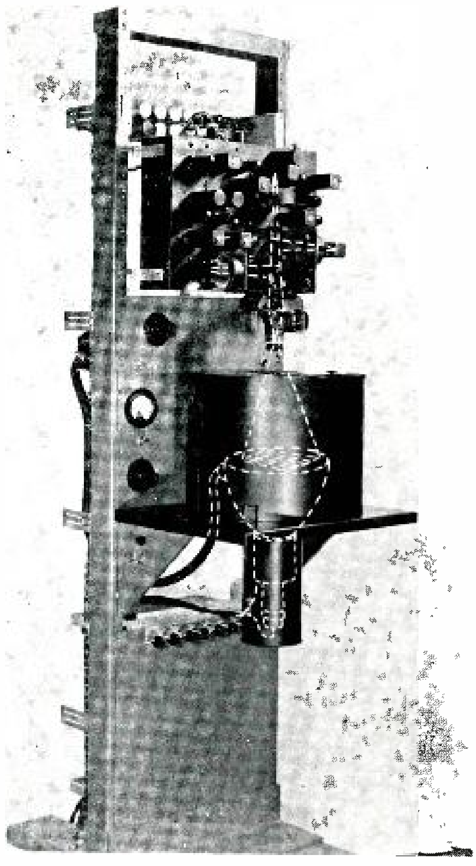

Fig. 2—Stationary picture signal generator.

then divided by dichroic mirrors which pass one color of light and reflect the other colors. The use of dichroic mirrors for a light splitter instead of half-silvered mirrors and color filters, reduces light losses and therefore provides a signal with higher signal-to-noise ratio. The divided light beams are further filtered by color absorption filters, then collected by multiplier type phototubes which convert the varying light intensity of the spot as transmitted by the slide into video signals corresponding to the three primary colors of the slide. The use of multiplier phototubes provides a high video input to the amplifier. The amplifiers are equalized to correct for the decay characteristic of the phosphor used in the flying-spot kinescope.

On the bottom of the rack in the photograph (Figure 2) is the chassis containing the synchronizing, blanking, and deflection circuits for the flying-spot kinescope. Behind the kinescope is the high-voltage power supply which provides approximately 30 kilovolts for the kinescope and a first anode focusing potential variable between four and seven kilovolts. The flying-spot kinescope has a special short persistence phosphor whose light intensity drops to less than 1 per cent of its original value in one microsecond. Facing the kinescope screen is the slide holder with an F:2 objective lens. The whole optical assembly is mounted on the same chassis with the three video amplifiers. The beam splitter is at the lower end of this chassis and a condensing lens system for each channel reduces the beam diameter to that of the photocell aperture after the beam is divided. The photocells are enclosed in shield cans, which also support the color filters.

The voltage for the phototube multipliers may be controlled by the variable power supply directly under the beam splitter chassis, and by this means the video levels of all three channels may be varied simultaneously. The supply voltage of the phototube multipliers of the individual channels may be varied individually by the potentiometers visible at the top of the chassis, to provide the desired color balance. Each of the three video amplifiers contains three stages having a flat frequency response to approximately 5.5 megacycles. Included in each of the amplifiers are the equalizing circuits to compensate for the various phosphor persistences. The output level of the amplifiers is approximately 1 volt peak-to-peak. The small chassis above the beam splitter is for the insertion of the synchronizing signals in the green video signals.

The quality of the signal from this generator is highly satisfactory not only because of the high resolution, but also because the blacks have the unusual characteristic of being practically free from noise. Noise in the picture therefore has the general appearance of the equivalent effect found in motion pictures from photographic grain and dirt. The registration of the three signals is inherently correct.

Motion Picture Film Scanner

The motion picture film scanner was built with no attempt at refinement or optimum design, in order to hasten preliminary tests of reception with moving subjects. Its general scheme is the same as that of the slide scanner, with the film gate replacing the slide holder. A photograph of the apparatus is shown in Figure 3. A standard 16-millimeter home sound film projector was modified by substituting a synchronous motor drive so that the film speed was changed to 30 frames per second (instead of 24). Each frame was then scanned

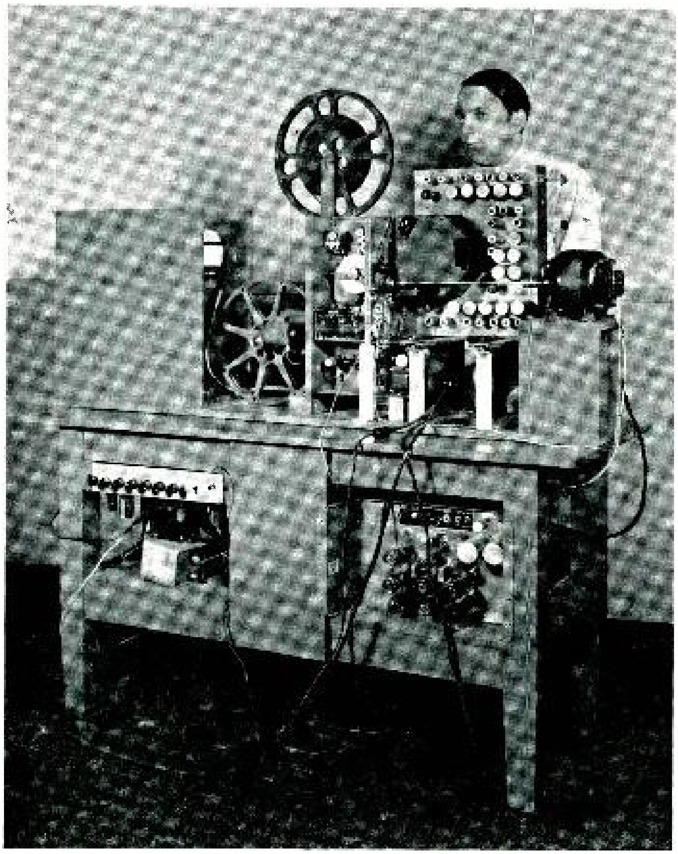

Fig. 3—Motion picture film scanner.

twice to give 60 fields per second. The pull-down mechanism, which was unchanged, is so slow that it was necessary to blank approximately 30 per cent of the field time to avoid showing the distorted picture produced during the film pull-down time. The picture therefore actually contained only about 370 lines, although the nominal number of lines was retained at 525.

The picture quality was judged to be good, particularly when allowance is made for the fact that part of the picture area was missing, due to the compromise in design of the film projector. The sound was usable, but was not very satisfactory due to the improper film speed.

REPRODUCING EQUIPMENT

The picture reproducer, as shown in Figure 4, 5, 6, contains three

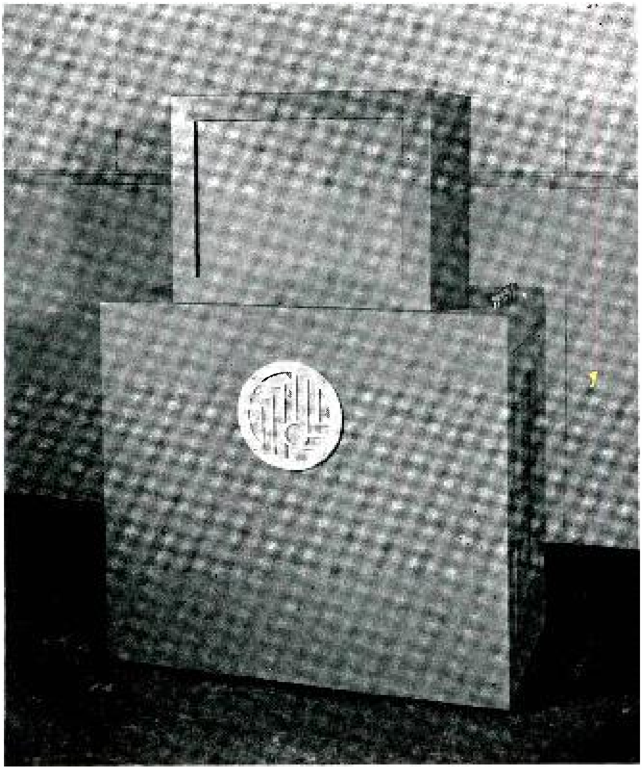

Fig. 4—Laboratory model simultaneous all-electronic color television receiver.

three-inch kinescopes arranged side-by-side in an equilateral-triangular group, each having an associated projection lens and deflection yoke. The kinescopes are identical except that phosphors selected for producing red, green, and blue light, respectively, are used. The kinescopes and lenses are mounted in an assembly frame which also holds the yokes in such a manner that each yoke may be adjusted in rotation and height without disturbing the kinescope mounting position. Each kinescope is provided with the video signal corresponding to its particular primary color, and has a scanning raster which produces light for its primary color image in the completed picture. The lenses project these three pictures simultaneously to the translucent viewing

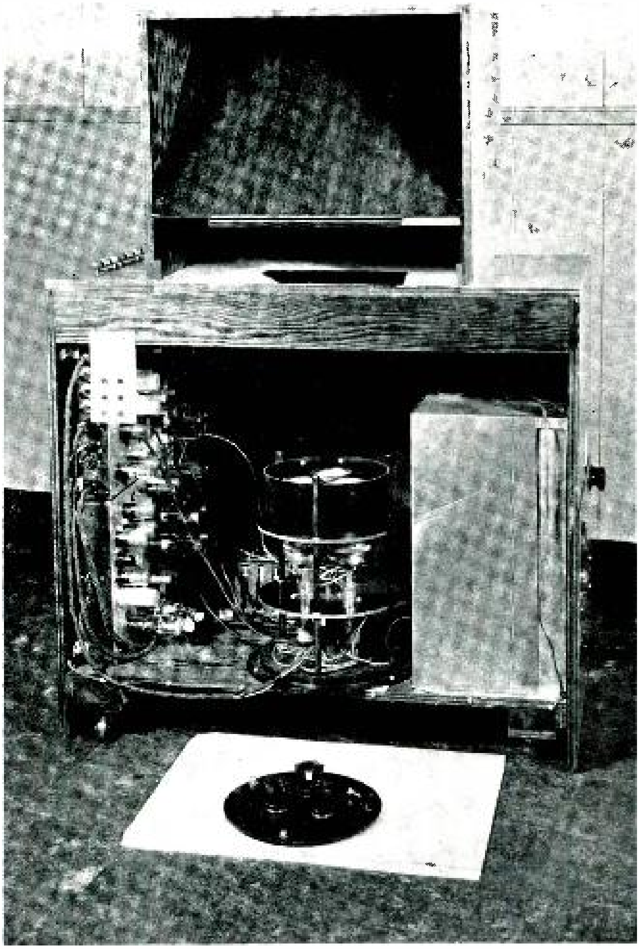

Fig. 5—Receiver with three cathode-ray projection tubes, lenses removed.

screen by way of a 45-degree mirror, as shown by dotted lines drawn on the photograph (Figure 6). The kinescopes are operated at a second anode potential of 25 kilovolts.

The optical system serves to focus and combine the three pictures on the translucent screen. In so doing, the images must not differ from one another in geometric distortion or location. Prevention of such difference is accomplished by placing the three kinescope faces in the same plane and mounting the three lenses above this plane, with their

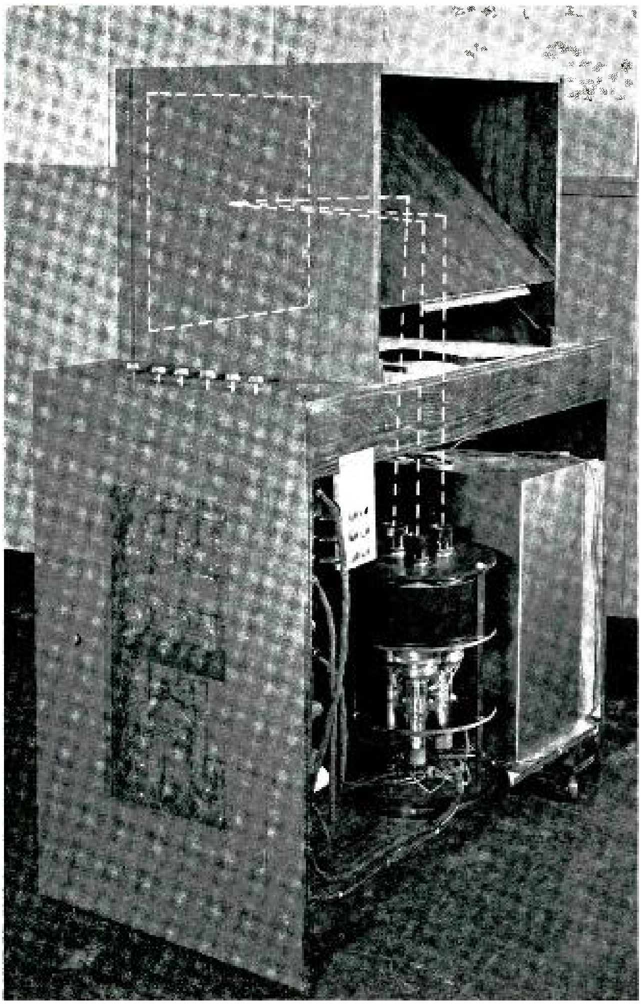

Fig. 6—Receiver with lenses in position, showing projection paths.

axes perpendicular to it. The axis of each lens is offset from the center of the kinescope face toward the center of the assembly by an amount sufficient to bring the three pictures into approximate register on the screen. Exact registry is then obtained by moving the raster on the kinescope face electrically. This offset, which is similar to a rising front on a photographic camera, causes no distortion, but requires extra covering power in the lens. The lenses used are F:2 projection lenses. They are threaded into the lens plate at the calculated positions and the threads serve as the focus adjustment.

The registration requirements are similar to those existing in color printing and color photography. Ideally, the three rasters should be identical and properly positioned within a fraction of the width of a scanning line. Practically, a considerable amount of misregistration may be present without being objectionable.

The scanning rasters are made substantially identical by using three similar yokes, and supplying them with power from the same deflecting circuit. The three yokes are connected in parallel rather than in series. This permits a simple individual centering or positioning arrangement and also insures more nearly identical deflection fields. In a series arrangement, one yoke would operate at a higher alternating-current potential with respect to ground, and would thus be shunted to a greater extent by the stray capacitances.

The positioning or centering arrangements are the usual television centering circuits, the only requirements being that the centering supply voltage must be stable, and that adjustment of the potentiometers must not alter the current waveshape through one yoke with respect to that through the others. To insure the latter, the horizontal centering potentiometers are by-passed very lightly and the vertical ones are unby-passed. Enormous capacities would be required to by-pass the vertical centering potentiometer properly. However, since the vertical circuit is essentially resistive, the addition of more resistance would simply change the amplitude, which is easily corrected.

The procedure used in registering the kinescope assembly is as follows. The kinescopes are adjusted for the proper height and clamped, then the lenses are focused optically. The yokes are then rotated until the edges of the three rasters are parallel. One of the three rasters is then considered standard for horizontal size, and one of the others is adjusted to it by moving the deflecting yoke up or down. Although this, of course, varies the raster size both horizontally and vertically, only the horizontal size is considered during this adjustment. This adjustment is then repeated on the remaining raster until the three horizontal sizes are alike. The vertical sizes are then adjusted by varying the value of small resistors in series with the vertical deflecting coils. The positioning or centering controls are then set to register properly the three rasters which are now of the same size.

Successful registry of the three rasters has been greatly facilitated by the use of aluminized kinescope tubes.[2] The aluminum film insures that the phosphor screen and the glass wall are at second-anode potential and hence do not collect charges that will divert the electron beam erratically.

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[2] D. W. Epstein and L. Pensak, "Improved Cathode-Ray Tubes With Metal-Backed Luminescent Screens", *RCA REVIEW*, Vol. VII, No. 1, pp. 5-9, March, 1946.

TELEVISION EQUIPMENT FOR AIRCRAFT*

By

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Camden, N. J.

Summary—The design considerations involved in the development of lightweight television equipment for airborne use are discussed in Part I of this paper. Following this is a description of Block I television equipment developed in accordance with these considerations. Flight testing of Block I equipment brought to light several difficulties peculiar to the transmission of television signals from aircraft. In Part II a number of these difficulties are discussed as well as methods developed for minimizing them.

Part I

Development and Design of Lightweight Television Equipment

THE first development work on television transmitting apparatus for use in aircraft was undertaken in 1936 under the direction of R. D. Kell. This was a direct outgrowth of Dr. V. K. Zworykin's memorandum to David Sarnoff of April, 1934.[1] Equipment was built using the 1850 type iconoscope and was installed and tested in a Ford trimotor airplane. Results obtained with this equipment clearly demonstrated the potential usefulness of television for the armed services. However, it was apparent that this equipment, although much smaller than other commercial equipment of the same type, was still appreciably larger and heavier than was considered desirable.

The advent of the 1848 iconoscope, a smaller tube than its predecessor, the 1850, made possible the design of television cameras of greatly reduced size. Commercial equipment using this tube and designed specifically for field pickup use was introduced in 1939. All the equipment associated with the camera was built into suitcase-type units, making it convenient to transport.

At this time both the Army and the Navy began to take a very serious interest in the military possibilities of television and equipment quite similar to the suitcase-type commercial design was constructed for airborne use in 1940. This equipment has been described in previous literature.[2] While this apparatus was useful in supplying a need for experimental and training equipment it was not particularly suit-

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* Decimal Classification: R583 X R520.

[1] V. K. Zworykin, "Flying Torpedo with an Electric Eye," *RCA REVIEW*, Vol. VII, No. 3, pp. 293-302, Sept., 1946.

[2] C. J. Marshall and L. Katz, "Television Equipment for Guided Missiles," *Proc. I.R.E.*, Vol. 34, No. 6, pp. 375-401, June, 1946.

able for military use since it had been designed primarily to meet commercial standards.

In 1940 a program was undertaken to develop television equipment specifically for use in military aircraft and having the following design objectives:

(1) Light weight
(2) Compactness
(3) Reasonably low power drain from a 12-volt direct-current source
(4) Reliable line-of-sight range up to 10 miles
(5) Resolution capability close to commercial standards
(6) Unattended operation possible at the camera and transmitter
(7) Overall ease of installation, operation and maintenance.

Naturally, a certain amount of compromise was necessary in order to produce results in reasonable agreement with the above requirements. Some of the fundamental decisions made in order to achieve these results are listed below:

(1) It was decided to place the iconoscope with its associated circuits and the transmitter in a single unit instead of two separate units. Although some difficulty was anticipated because of feedback from the transmitter into the camera circuits it was felt that advantages were to be gained with regard to overall size and weight considerations as well as in simplification of the interconnection problem.
(2) The dynamotor power supply was made a separate unit so that a dry battery supply might be substituted for the dynamotor under certain operating conditions.
(3) The frame frequency was made approximately 40 cycles and the line frequency approximately 14,000 cycles without definite relationship between the two as is necessary for commercial interlaced scanning. This decision greatly simplified the problem of developing suitable synchronizing and other line and frame frequency pulses. The choice of 40 cycles for frame frequency was a compromise. A lower frequency would have resulted in objectionable flicker and a higher frequency would have resulted in lower overall resolution for the available bandwidth. The line frequency of 14,000 cycles was chosen to give 350 scanning lines. This resulted in approximately 275 lines resolution in the vertical direction and 350 lines in the horizontal direction. Equal values of resolution could have been obtained by using a higher frequency for horizontal scanning and a consequent increase in the number of lines. However, it

was considered desirable to keep this frequency somewhat low so that the horizontal scanning circuits would require less power and would have lower dissipation.

(4) The vertical synchronizing pulses were made approximately one and a half lines long so that only one or at the most two horizontal synchronizing pulses would be lost during the vertical synchronizing period. It was considered unnecessary to add sufficient tubes and circuit elements to place slots in the vertical pulse as is done with the standard Radio Manufacturers Association synchronizing signal. The timing of the leading edge of the vertical synchronizing pulse was also made coincident with the leading edge of the vertical blanking pulse. This simplification of the synchronizing signals was made possible by the choice of sequential rather than interlaced scanning and resulted in a considerable saving in tubes and circuit elements.

(5) The overall video band width of the system was made 4.5 megacycles and both the transmitter and receiver were designed for double sideband operation. Obviously, single-sideband operation would have resulted in a more complex transmitter. The decision to make the receiver accept both side bands was based on the desirability of eliminating tuning controls in order to simplify receiver operation. The band width chosen was considered to be a fair compromise between the resolution capability of the 1848 iconoscope, normally somewhat better than 350 lines, and the number of video and intermediate-frequency amplifier stages required in the camera, transmitter and receiver in order to obtain adequate amplification. Field tests with earlier types of television equipment indicated that the resolution capability of the proposed system was entirely adequate for most of the military applications proposed.

(6) A transmitter frequency was chosen in the neighborhood of 100 megacycles. This made possible efficient tuned circuits with lumped constants of relatively small size and also an antenna structure not considered to be excessively large.

(7) Picture monitoring facilities were not included with the camera transmitter unit since unattended operation was expected. Instead a separate monitor unit employing a 7-inch kinescope was designed so that it could be connected to the camera transmitter with a single cable connection. When this was done, the $B+$ supply to the transmitter output tube was automatically transferred to the monitor circuit. When this connection was

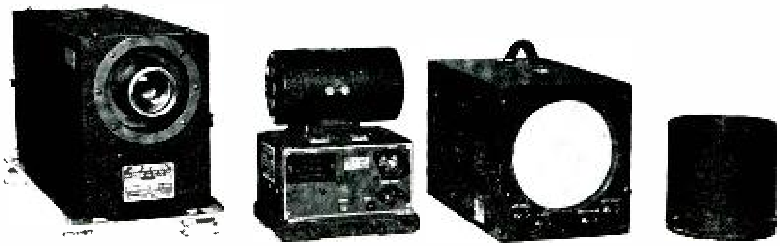

Fig. 1—Block I equipment at transmitting location.

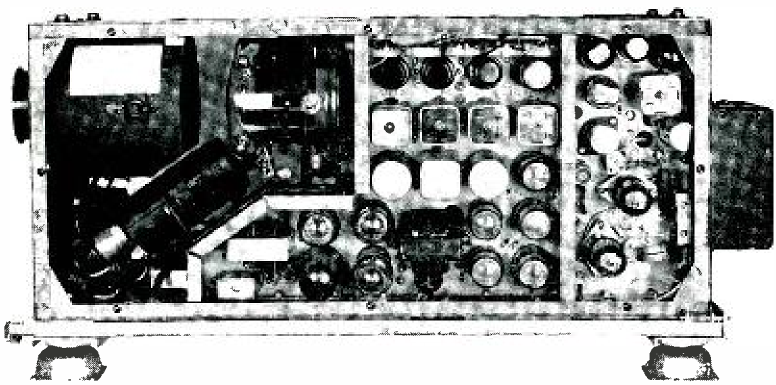

Fig. 2—Camera transmitter—tube side.

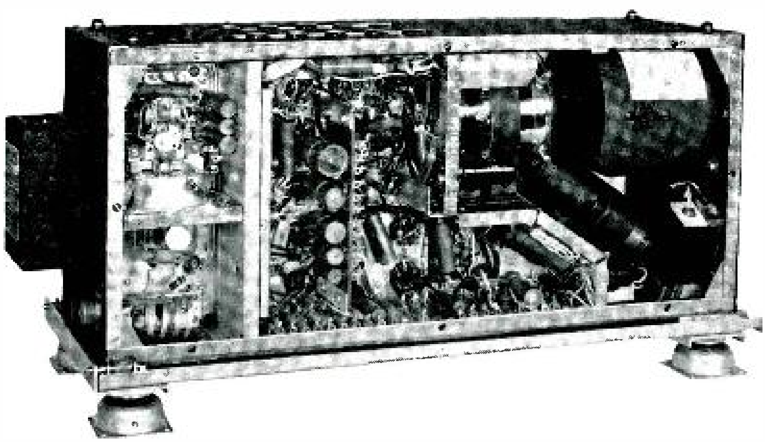

Fig. 3—Camera transmitter—circuit side.

made, the power drain on the dynamotor remained approximately the same as with the transmitter load.

Other design considerations involved in the development of this equipment will be discussed in the detailed descriptions of each unit which follow. The three units necessary at the transmitting location are shown in Figure 1. Figure 2 presents a side view of the camera-transmitter showing most of the tubes. As may be seen in the photograph the type of construction used resembles an "I" beam with closed ends and results in a light-weight but very rigid unit. The center chassis section on which most of the tubes are mounted is welded to the "wrap around." Both side covers of the unit, which are removable, are fastened in place by six "Dzus" type fasteners. Special spring contacts are placed on all the outside flanges of the case to insure a good ground between the covers and the case. The front compartment houses the lens mounting and iconoscope; the central portion of the case contains the video amplifier and deflecting circuits; and the transmitter is located in a narrow section at the rear of the unit.

Ten control knobs are recessed in the top of the case. These, together with the power switch on the back of the unit may be considered as "operating" controls. Their functions are as follows:

(1) Horizontal Sawtooth Shading
(2) Vertical Sawtooth Shading
(3) Horizontal Parabola Shading
(4) Vertical Parabola Shading
(5) Iconoscope Horizontal Centering
(6) Iconoscope Vertical Centering
(7) Iconoscope Size
(8) Iconoscope Focus
(9) Video Gain
(10) Iconoscope Bias

Five additional controls are accessible after removal of the right-hand cover. These controls require adjustment only when associated tubes are changed. The five controls are arranged along the partition separating the transmitter section from the rest of the unit. Reading from top to bottom they are:

(1) Synchronizing Amplitude
(2) Iconoscope Vertical Size
(3) Iconoscope Blanking
(4) Vertical Speed
(5) Horizontal Speed

The power from the dynamotor is brought in through a plug in the rear of the unit. A coaxial connection for a 75-ohm radio-frequency output line from the transmitter is also located at the rear of the unit.

Figure 3 shows the other side of the case where most of the circuit elements are placed. In spite of the large number of components most of them are readily accessible. Electrically there are four main groups of circuits making up this unit. These are:

(1) The master scanning oscillators and deflecting circuits
(2) The iconoscope with its associated video amplifier, shading, blanking and synchronization mixing circuits
(3) The transmitter circuits
(4) Power supply circuits

Considering first (1) above, this comprises that part of the schematic shown in Figure 4 (facing page 472) associated with tubes V6, V7, V8, V9, V10, V13, V14 and V15. Cathode-coupled multivibrators were chosen as master oscillators for both line and frame frequencies because they produce pulse wave shapes which can easily be transformed into synchronizing and blanking pulses. Across the common cathode resistor of this type of oscillator is found a positive pulse with steep vertical sides, a decreasing amplitude during the pulse itself and a constant voltage level between pulses. The wave shape of the voltage on the output plate of the oscillator is similar to that on the cathode but it has the opposite polarity. It may be noted that a relatively high value of cathode resistor is used in the multivibrator circuits in order to obtain sufficient pulse voltage on the cathodes for satisfactory operation of the mixer circuits. This requires a positive bias on the multivibrator grids.

An adjustment of this bias voltage on the output triode of the multivibrator provides a convenient speed control. The circuit elements of the multivibrators were so chosen that the resulting widths of the pulses appearing on the cathodes have a duration such that these pulses are suitable for use as kinescope blanking signals. Narrower pulses obtained by differentiating the cathode pulses are used in developing the synchronizing signal. The same cathode pulses are also fed to the usual type of discharge-tube sawtooth generating circuits in order to create the necessary scanning waveforms. The pulse appearing in the plate circuit of the vertical multivibrator is used as a source of blanking signal for the iconoscope grid.

Resistance mixing circuits are used to combine the horizontal and vertical pulses from the multivibrator cathodes in the grid circuit of the blanking amplifier (Section A of tube V6), and the resultant is added to the video signal in the common plate circuit of the video amplifier tube V5, Section A, and tube V6, Section B. The amplitude of these pulses on the grid of the tube V5, Section B, is sufficient to drive

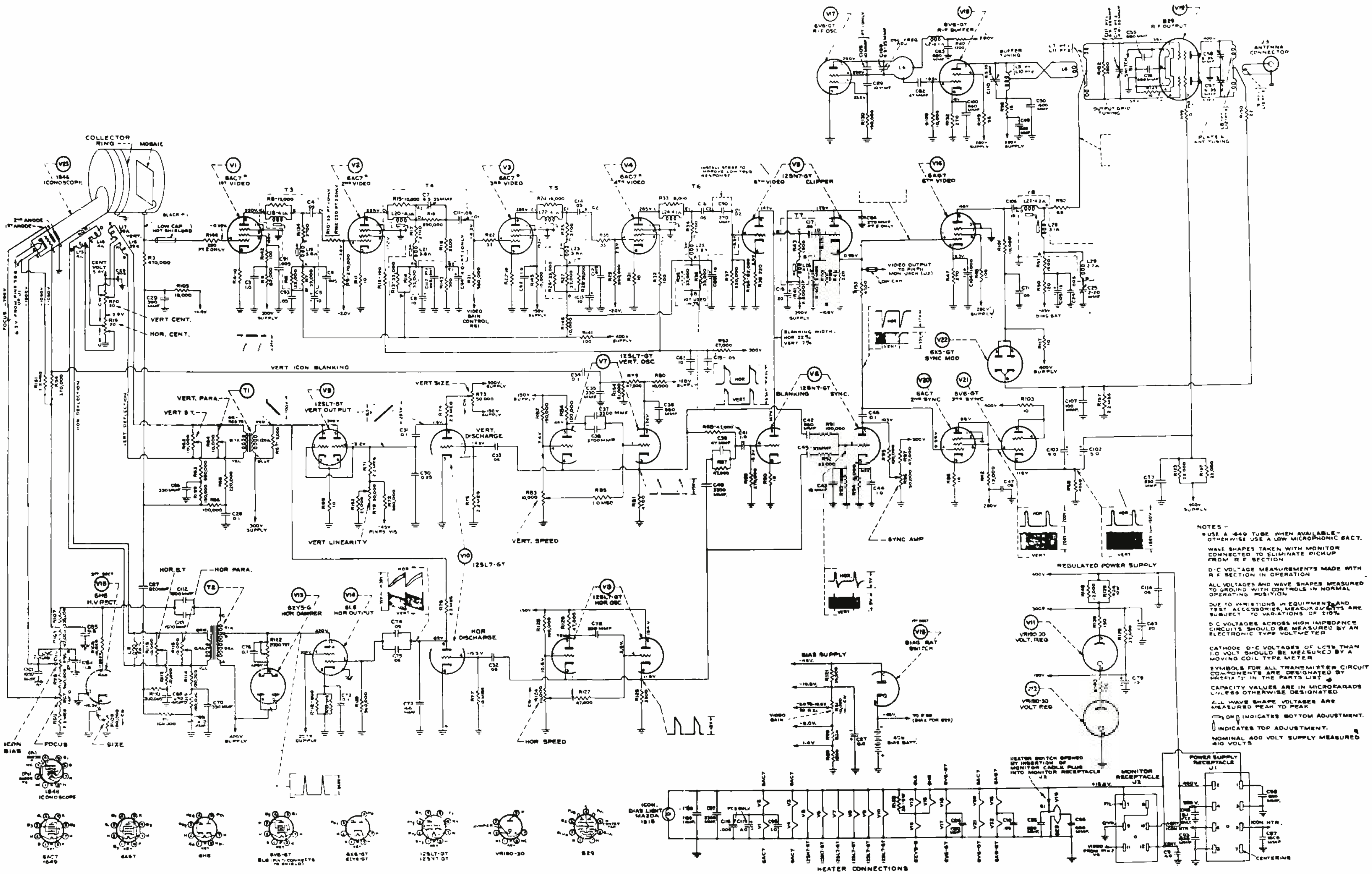

Fig. 4—Camera transmitter—schematic diagram.

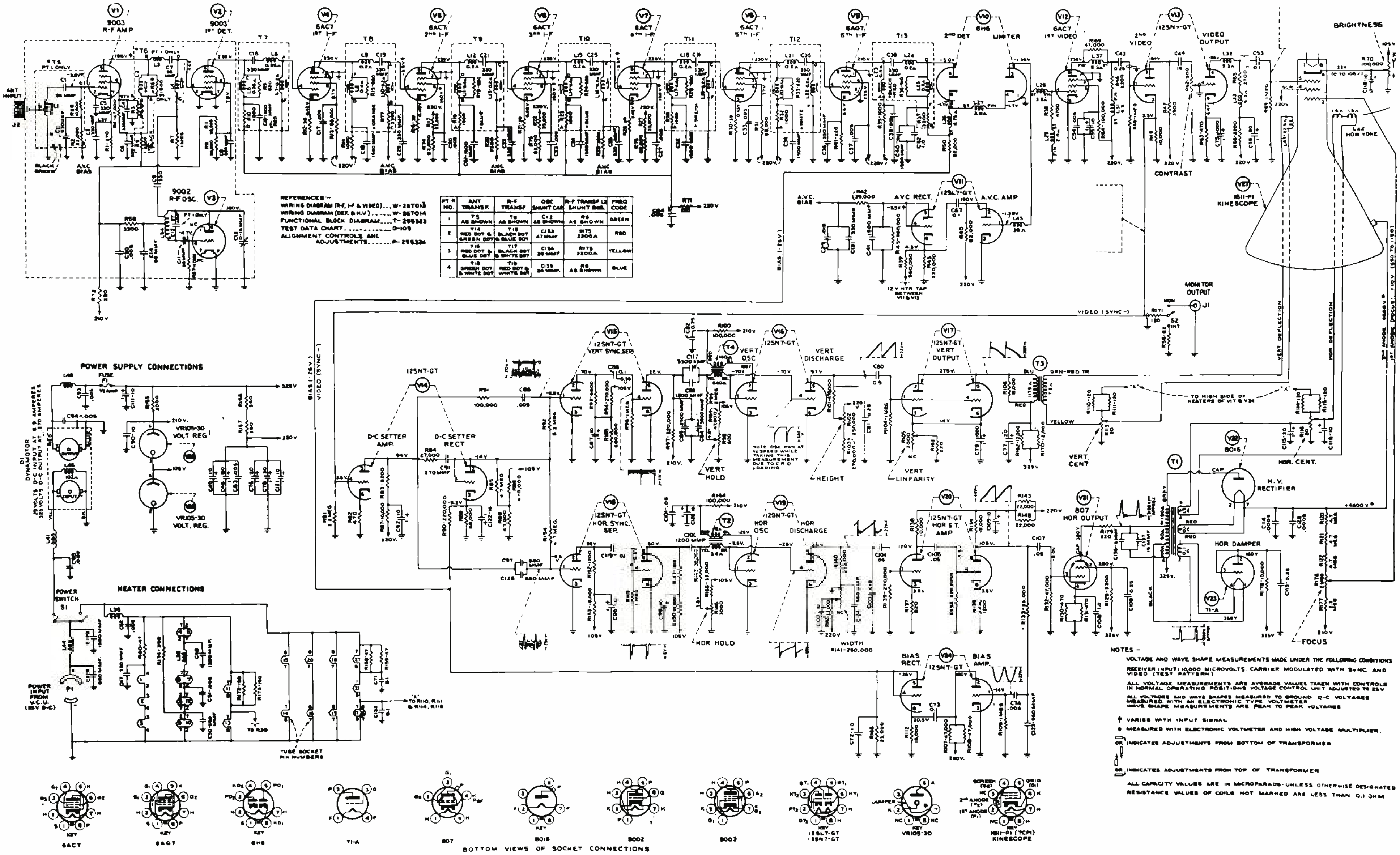

Fig. 10—Receiver—-schematic diagram.

the grid beyond cut-off. Thus no video signal is transmitted during the pulse time with the result that the kinescope screen of the receiver can be made black during this period and the retrace of the cathode ray beam will not be visible. The synchronizing pulses also obtained from the pulses appearing on the multivibrator cathodes, are much narrower than the corresponding blanking pulses because of the insertion of small coupling capacitors in the mixing circuit at the grid of tube V6, Section B, which pass only the steep wave front of the pulses. These short pulses appear in the plate circuit of the tube and are added to the combined video and blanking signal appearing at the cathode of the video amplifier tube, V5, Section B.

The function of the 18-micromicrofarad, additional shunt capacitance from grid to ground of the synchronizing pulse amplifier tube V6, is to delay slightly the leading edge of the horizontal synchronizing pulses with respect to the leading edge of the horizontal blanking pulses thereby producing a narrow but adequate "front porch" or delay period. The function of this delay is the same as that in the standard Radio Manufacturers Association signal—to prevent video signals near the leading edge of the blanking signal from changing the timing of the synchronizing signals. The same capacitor also causes a slight delay of the leading edge of the vertical synchronizing signal with respect to the vertical blanking signal with the same result.

As noted previously, the necessary saw-tooth wave forms for deflecting the iconoscope cathode-ray beam are derived from the pulses appearing on the cathodes of the two multivibrators by impressing them on the grids of the two discharge tube sections of tube V10. The vertical saw-tooth is amplified in tube V9 and fed through a step-down transformer to the vertical deflecting coils of the iconoscope yoke. A horizontal saw-tooth appears in the plate circuit of Section B of tube V10. The plate supply for this circuit comes from the vertical output tube V9. Thus the horizontal saw-tooth is modulated by a vertical saw-tooth thereby producing the keystone correction made necessary by the construction of the iconoscope. This requires a greater amplitude of horizontal scanning current when the beam strikes that part of the mosaic nearest the electron gun. A control is not required on the amplitude of keystone correction since the circuit constants of the horizontal saw-tooth generating circuit were chosen to give the exact amount of correction necessary. This keystone-corrected horizontal saw-tooth is amplified in the horizontal output tube (V14), and the horizontal damping tube V13 supplies the necessary damping to suppress horizontal frequency transients. A step-down transformer in the plate circuit of tube V14 feeds the deflecting signal to the horizontal deflection coils of the iconoscope yoke.

In order to obtain a direct-current second anode voltage of approximately -1000 volts for the iconoscope, the plate transformer of the tube V14 is also provided with a step-up winding which feeds enough horizontal signal to the high voltage rectifier tube V15 to obtain the necessary voltage.

Since a horizontal size control would also cause a variation in the second anode voltage which would, in turn, require a readjustment of vertical size, it was considered more desirable to keep the horizontal scanning current in the yoke at a fixed value and provide an adjustment of the second anode voltage supply.

This adjustment changes the deflection amplitude of horizontal and vertical scanning circuits simultaneously so that a quick check of scanning amplitudes can be made by decreasing the second anode voltage slightly which should normally bring the frame of the iconoscope mosaic uniformly into view at all four edges of the picture. The second anode voltage is made adjustable by inserting a variable resistor in the cathode lead of the rectifier tube V15.

The second major division of the camera transmitter mentioned in (2) above is the iconoscope and its associated video amplifier consisting of tubes V1, V2, V3, V4, and V5.

The iconoscope output resistor has a relatively high value in order to improve the signal-to-noise ratio at low frequencies, thereby reducing microphonics and ripple. This causes a drop in the high-frequency content of the video signal which is compensated for by the "high peaker" coupling circuit between tubes V2 and V3.

All the video stages except the one associated with V5 were made using both a series and a shunt inductance in the amplifier plate circuits in order to equalize the high frequency response. The single inductance in the plate circuit of the first triode section of V5 was adjusted to produce a broad resonance in the middle frequency range in the neighborhood of 2.5 to 3.0 megacycles in order to equalize the dip in response resulting from the other four circuits in cascade. In this manner an overall flat response was obtained to approximately five megacycles. The low-frequency phase and amplitude response, affected mainly by the choice of coupling capacitors, grid leaks, and plate-filter by-pass capacitors, is such that the amplifier is capable of passing a 40-cycle square wave with very little distortion. In order to eliminate some of the disturbing effects in the picture caused by microphonics especially at the lower frequencies, a small series coupling capacitor (270 micromicrofarads) is placed in series with the grid of the video amplifier V5. This is normally short-circuited with a short length of wire which can then be cut if noise and vibration conditions are especially severe. The consequent reduction in low-frequency response attenuates most of the usual microphonic frequencies but also results in some loss of picture intelligence. The most objectionable effect is the introduction of rather long horizontal light streaks after dark objects in the picture.

The potentiometers and associated circuits in the low-impedance side of the two deflection transformers provide shading signals which are mixed with the video signal in the iconoscope output circuit. These signals compensate to a certain extent for the spurious signal generated by the iconoscope itself (black spot signal). Four compensating signals are generated; horizontal and vertical saw-tooth wave forms, and horizontal and vertical parabola wave forms. Since the two deflecting output circuits are balanced to ground (by having the connection between pairs of deflecting coils grounded through the low-impedance centering circuits) the mid-position of the four potentiometers corresponds to zero shading signal. Moving the potentiometers to one side of this point provides one polarity of shading signal, moving them to the other side provides the opposite polarity.

The waveform of the signal appearing at the vertical transformer secondary is a reasonably linear saw-tooth so that a single differentiation circuit consisting of R65 (220,000 ohms) and C28 (0.1 microfarad) supplies a satisfactory vertical parabola. The 330-micromicrofarad capacitor (C68) is for the purpose of by-passing horizontal frequency pulses introduced into the vertical circuits through coupling in the deflecting yoke. Since the voltage waveforms appearing on the secondary side of the horizontal output transformer are pulses, it is necessary to use a single differentiating circuit to obtain saw-tooth waveforms and two of these circuits in series to obtain parabolic waveforms. The series resistors in the mixing circuit used to combine all four different waveforms are so chosen that the maximum voltages derived from all shading potentiometers are approximately the same.

The radio-frequency oscillator and buffer circuits are also shown in the schematic (Figure 4). V17 is the radio-frequency oscillator, the tank circuit of which consists of L4, C108, and C109. C109 may be removed for change of frequency and C108 is variable to permit tuning to the exact frequency desired within the range. L4 is tapped in two places, the center tap going to the plate supply voltage through L2 and R40, the other tap being the oscillator output and feeding through C82 to the buffer grid. The voltage supply is connected through the circuit of L4 to both the plate and the screen grid. This oscillator has the control grid and screen grid closely coupled together so that the tube oscillates as a low "mu" triode. The output is taken from L4 through C82 to the control grid of the buffer stage V18. This buffer stage is a radio-frequency amplifier with a tuned plate circuit consisting of L5 and C110. Connected to L5 is a link circuit which has a coupling coil L6 feeding to the radio-frequency output stage, V19.

The radio-frequency output and modulation circuits are also shown in Figure 4. The coupling coil L6 couples to L7 which is the grid tuning circuit for the radio-frequency output tube V19. This is a double section tube having essentially two beam-power structures within one envelope. L7 has a center-tap connection for grid-modulation and bias for V19. The bias supply, from a 45-volt dry battery, passes to L7 through the output circuit of the last video amplifier which contains the video signal, synchronization and blanking. Consequently, the grid voltages on V19 will be modulated with these various voltages. The plates of this tube are connected to their respective ends of L8. L8 is part of the output tank circuit which consists of L8, C58 and C57. C58 and C57 are tuning condensers ganged together and controllable by one adjustment. The heater circuit of V19 is supplied directly from the 12.5-volt power supply through a switch, shown as S4, which opens the heater circuit when the monitor cable plug is inserted. L9 is a radio-frequency output coil coupled to L8 and is connected to the antenna output terminal.

To insure that synchronizing pulses will always be transmitted at full power, further modulation is accomplished in the plate circuit of V19 for the synchronizing pulses only. The output from V20 consists of synchronizing pulses applied to the grid of V21. The plate current for V19 passes through R117, V22 and R150 to the mid-tap of L8. V22 is a half wave rectifier tube through which the plate current of V19 flows. Condenser C103 is connected from the plate circuit between V22 and R150 to ground through a 4,000-ohm resistor. This 4,000-ohm resistor is also in series with the cathode of V21. Between synchronizing pulses, V21 is not drawing plate current and consequently C103 will be charged to full plate voltage. When a synchronizing pulse arrives at the grid of V21 it will draw considerable plate current, causing current to flow through R68, making the lower end of C103 approximately 180 volts positive. This 180 volts added to the approximately 400 volts already on C103 will raise the plate voltage of V19 to about 580 volts. The condenser C103 supplies V19 with power during the short interval that the synchronizing pulse continues, as V22 will not allow a discharge of this condenser back through the power supply.

Consequently, during the period of a synchronizing pulse, V19 is driven completely to the limits of operation and definite modulation of synchronizing pulses is secured. In order to make the rise of plate voltage completely effective, C102 is placed in this circuit. This raises the screen grid voltage on V19 in much the same manner as the plate voltage was raised. Therefore, by having both grid and plate modulation for the synchronizing pulses, it is possible to hold the receiver in synchronism under severe conditions of noise and interference.

The power supply unit consists of a dynamotor, a power-control relay and two filter systems and is shown schematically in Figure 5. All voltages required for operation of the camera transmitter and monitor are obtained through connections to this unit. The dynamotor, which comprises the means for converting the 12-volt primary power into the high-voltage direct-current potential required for operation of the camera-transmitter and monitor, is mounted on the top of the unit, thus providing adequate ventilation and ease of access for routine service of the rotating equipment.

All cable connections required for normal operation of the equipment are completed by means of an 8-connector receptacle (camera-

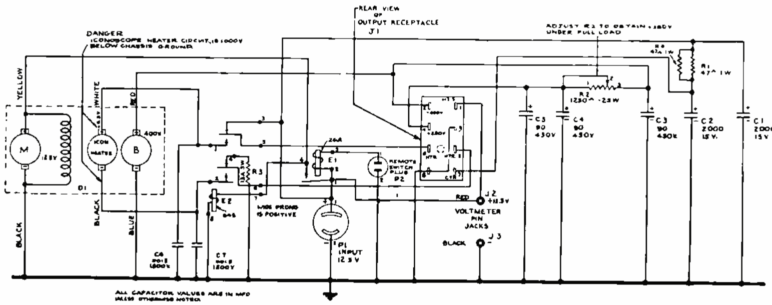

Fig. 5—Power supply—schematic diagram.

transmitter connections) and a 2-connector plug (power connection). A third receptacle is provided for connection of a remote power control switch if such is required.

The dynamotor is designed for operation from a 12.5 volts direct-current supply. Two secondary windings and commutators are provided; one insulated for 1000 volts delivering 6.3 volts direct-current to the filament of the iconoscope and the other delivering 400 volts direct-current to the camera-transmitter and to the input of a filter system. Approximately 280 volts is supplied to the camera-transmitter from the output of this filter. The iconoscope centering voltage is derived from the heater supply but is filtered to keep ripple on this supply from affecting the deflecting circuits.

In the camera transmitter unit itself are located two regulator tubes

V11 and V12 in series. These tubes provide a constant voltage supply to those circuits which are sensitive to slight changes in anode voltage. Such circuits include the scanning oscillator circuits, the screen grids of the video amplifier tubes, the blanking and synchronizing amplifier tubes and the clipper tube V5. The bias supply is derived from a 45-volt "C" battery. This was chosen primarily to supply the transmitter output tube with a low impedance bias source so that grid current would not change the bias voltage. A bleeder is used to provide the necessary bias voltages required for the video amplifier. This bleeder is supplied through one diode section of tube V15 so that the drain on the battery is automatically removed whenever the heater supply is removed. The second-anode supply for the iconoscope is derived from

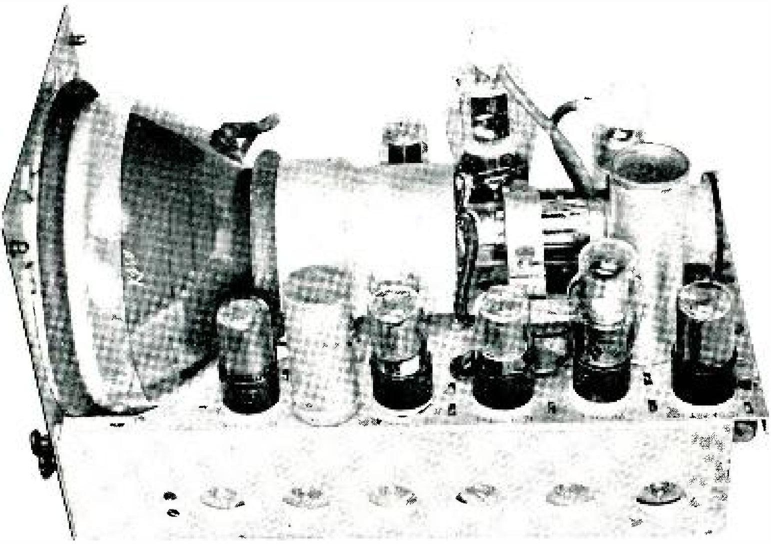

Fig. 6—Monitor—with cover removed.

the horizontal output transformer and was discussed in detail in the description of the scanning circuits.

The monitor designed for use with the camera transmitter is shown with the transmitting equipment in Figure 1. In Figure 6 it may be seen again with the cover removed. The monitor unit consists of a type 1811-P1, 7-inch kinescope, together with its associated deflection circuits and video amplifier. Connection of the monitor to the camera transmitter automatically opens the heater circuit of the Type 829 output tube in the transmitter (V19), thus compensating for the power required by the monitor unit, and maintaining normal $B+$ voltage.

Four controls required for normal operation of the monitor unit are located on the front panel, below the screen of the kinescope. These controls, progressing from left to right are, respectively, the "horizontal hold" control, the "focus" control, the "brightness" control and the "vertical hold" control. The "width", "height", "vertical linearity", "vertical centering" and "horizontal centering" controls, being used infrequently, are screwdriver adjustment. They are arranged from front to back along the right-hand side of the unit in the order named. The monitor is provided with a removable light shield.

To obtain access to tubes and circuit components, the chassis can be removed from the case after loosening the three "Dzus" fasteners on the front of the unit and also removing the power cable.

The monitor was designed specifically for operation from a com-

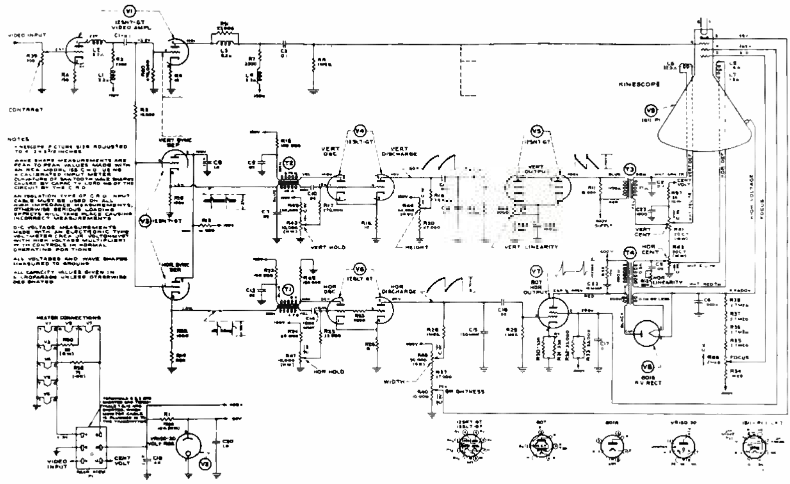

Fig. 7—Monitor—schematic diagram.

posite picture signal of the type used to feed the transmitter modulator and having approximately one volt peak-to-peak amplitude. It contains a video amplifier, separating circuits, line and frame frequency oscillators and their associated deflecting circuits. A schematic diagram is shown in Figure 7.

A 12SN7GT (V1) forms a two stage video amplifier which receives a video signal from the monitor output of the transmitter unit and feeds the signal to the kinescope grid. From the plate of the first amplifier stage the signal is fed to the synchronizing separator tube, a 12SL7GT (V3). There is sufficient synchronizing signal amplitude at this point so that the video signal drives the grids of this tube beyond cut-off and only synchronizing pulses appear on the two cathodes from which the vertical and horizontal oscillator (V4 and V6) are fed. Both horizontal and vertical oscillators are conventional blocking oscillators except that speed controls are provided by adjustments of the amount of positive bias on the oscillator grid leak ground returns.

A resistance-capacitance damping circuit is used across the horizontal deflection transformer in place of a damping tube in order to save space, although the circuit is somewhat less efficient.

This same transformer also has a step-up winding and a filament winding which supplies heater and plate voltage to a Type 8016 rectifier, V8. The rectified direct-current output (approximately 4500 volts) is applied to the second anode of the kinescope. A potentiometer in a bleeder circuit on this supply furnishes an adjustable first anode voltage. A VR-150-30 regulator tube (V2) supplies a substantially constant 150-volt potential to the video amplifier and the vertical and

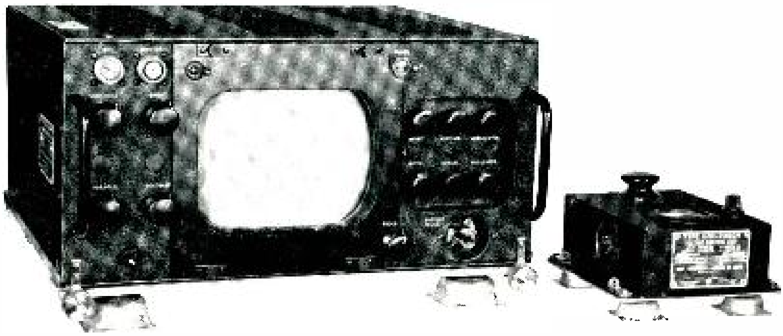

Fig. 8—Block I receiver and voltage control.

horizontal oscillators, so that gain and scanning oscillator frequency will be reasonably constant regardless of change in supply voltage.

Figure 8 shows the receiver and a voltage control box designed specifically for adjusting the input voltage to the receiver. In Figure 9 the top cover of the receiver has been removed to show the general construction and arrangement of parts. It may be seen that in design features it is somewhat similar to the camera-transmitter unit.

The receiver is completely self-contained and includes its own power supply, designed for operation from a 12- to 14-volt storage battery or the equivalent. The entire unit mounts, by means of thumb-screws, on a shock-mounting base supplied with the equipment. Plate power is supplied by an internal dynamotor developing 330 volts (direct-current), when connected to a 12.5-volt direct-current source. All cable connectors and controls necessary for set-up and operation of the unit are brought out to the front panel. Battery connections are terminated at a polarized plug, mounted at the lower right-hand side of the panel. The antenna connection is completed through a connector mounted at the upper left-hand side of the receiver. The antenna circuit is designed to work out of a 75-ohm coaxial line.

Controls requiring adjustment under normal operating conditions

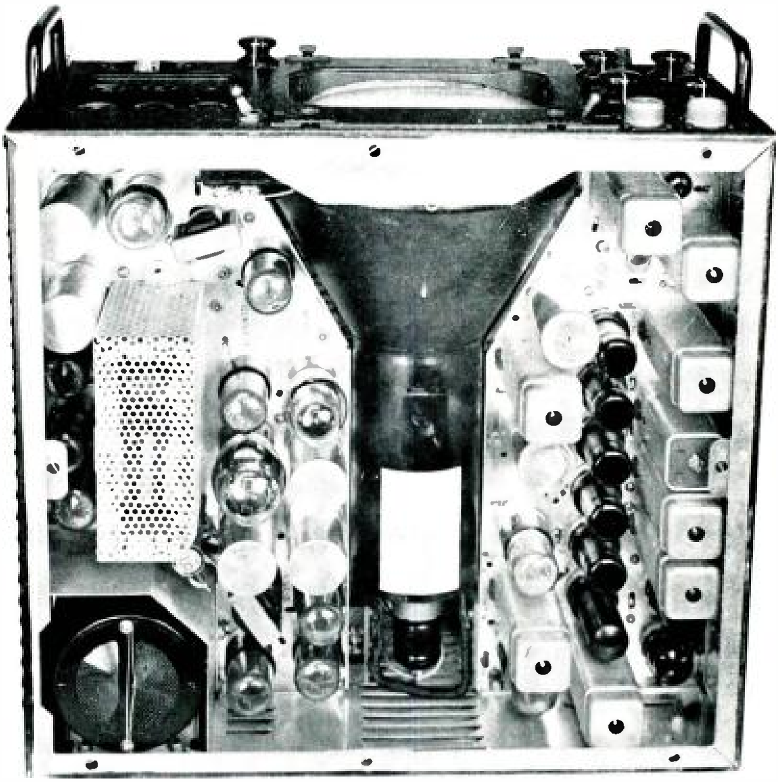

Fig. 9—Block I receiver—top cover removed.

are mounted on the front panel. Those controls requiring only infrequent adjustment are recessed below the level of the panel.

The controls mounted on the panel are the "vertical hold" control located on the left end of the upper row of controls, the "brightness" control at the right of the vertical hold control, the "horizontal hold" control at the left of the lower row of controls, the "contrast" control at the right of the horizontal hold control, and the "focus" control at the right and below the kinescope. The power switch is mounted directly above the focus control.

The controls which are recessed are (starting at the left end of the top row); the "vertical size" control, the "vertical linearity" control and the "vertical centering" control, the "horizontal linearity" control and the "horizontal centering" control.

The connector plug located at the right of the antenna plug is used when it is necessary to send a signal to a remote monitor. When it is desired to utilize the monitor signal, the small switch mounted next to the monitor output plug must be switched to the monitor ouput position; otherwise, the switch should always remain in the "internal" position.

Electrically, the receiver may be separated into four main divisions:

(1) The radio-frequency amplifier, intermediate-frequency amplifier and video amplifier chassis is located at the left of the kinescope when the receiver is viewed from the wiring side with the face of the kinescope tube at the top.

(2) The seven-inch kinescope, Type 1811-P1 is mounted in the middle of the receiver. The kinescope is housed in a mu-metal shield, within which is mounted the magnetic deflecting yoke.

(3) The deflection chassis, mounted to the right of the kinescope, contains the circuits which separate the synchronizing signal from the video signal, the blocking oscillators and discharge tubes, the saw-tooth voltage amplifiers and output tubes. These are utilized for generating the saw-tooth current wave in the magnetic deflecting yoke which deflects the electron beam in the kinescope. The horizontal output circuit which provides the high voltage power supply for the kinescope is also located in this section.

(4) The dynamotor, located at the right rear corner of the deflection chassis supplies all the low-voltage direct-current power required for operation of the receiver.

A schematic diagram of the receiver is shown in Figure 10 (facing page 473). The radio frequency amplifier, Type 9003 (V1), the first detector, Type 9003 (V2), and the fixed frequency radio-frequency oscillator, Type 9002 (V3), are mounted on a small chassis. This chassis is so designed that it can be removed from the radio-frequency and intermediate-frequency chassis of the receiver unit and replaced with a unit designed to receive another carrier frequency within reasonable limits if that becomes desirable. The radio frequency unit is followed by a six-stage intermediate frequency amplifier (tubes V4, V5, V6, V7, V8 and V9). The first five stages of the intermediate frequency amplifier utilize Type 6AC7 tubes and the sixth stage a Type 6AG7 tube.

The coupling circuits in the intermediate frequency amplifier circuit are "constant K" type filter sections with resistance loading on the input or plate side, and a full shunt arm on the grid or output side.

The sixth intermediate frequency amplifier stage drives the second detector, one section of a Type 6H6 tube (V10). The other section of V10 is used as a limiter tube to limit the noise peaks coming through the system.

The picture signal output from the second detector, after going through the limiter, is amplified in the video amplifier tube, Type 6AC7 (V12), and the two sections of the Type 12SN7GT output tube (V13). The first half of V13 furnishes at its cathode an output signal for the external monitor and the video signal for the synchronizing pulse separating system. Its plate circuit furnishes video signal to the contrast control. The second half of V13 is used as an output tube, the signal on its grid being obtained from the contrast control, the plate circuit driving the grid of the Type 1811-P1 (C7466) kinescope (V27).

The automatic volume control amplifier and rectifier tube, Type 12SL7GT (V11), is also mounted on this chassis. Its purpose is to hold the output of the set at a constant level over a wide range of signal input voltages.

Mounted on the deflection chassis is the video signal amplifier and direct-current setting tube, Type 12SN7GT (V14), the vertical synchronizing separator and clipper tube, Type 12SN7GT (V15), the vertical oscillator and discharge tube Type 12SN7GT (V16), and the vertical output tube, Type 12SN7GT (V17). Tube (V17) furnishes saw-tooth deflection current to the vertical coils in the deflection yoke through the vertical output transformer.

Contained also in this chassis are the horizontal synchronization separator and clipper tube, Type 12SN7GT (V18), the horizontal oscillator and discharge tube, Type 12SN7GT (V19), the horizontal saw-tooth voltage amplifier tube, Type 12SN7GT (V20), the horizontal output tube, Type 807 (V21) which furnishes a saw-tooth current wave to the horizontal deflection coils through the horizontal-output transformer, the controlled damper tube, Type 71A (V31), and the high-voltage rectifier tube, Type 8016 (V22). This tube rectifies the "kick-back" voltage generated in the horizontal output transformer during the return line time to provide high-voltage (direct-current) required by the second anode of the kinescope.

The bias amplifier and rectifier tube, Type 12SN7GT (V24) which amplifies and rectifies the horizontal saw-tooth voltage to provide a negative bias voltage, and the two voltage regulator tubes, V25 and

V26 (Type VR-105), which provide regulated voltage to the plates of the several oscillators (where stability of frequency is important) and to the synchronizing separator and clipper tubes, are also located on this chassis.

Antennas for use with this equipment are not discussed herein, but will be covered in a separate paper at a future date. However, one of the frequently-used antenna types was a quarter-wave vertical rod antenna working against two quarter-wave ground rods extending on opposite sides at the base of the vertical radiator. A matching unit was used to transform the impedance of the antenna to 75 ohms in order to match the transmission line.

Tests of the overall equipment have already been described in some detail.[2] These tests involved extensive work over a long period by both government and company engineers. Although a number of operating difficulties were experienced (see Part II) which were corrected in later equipment, it was found that the results obtained were in substantial agreement with the original design objectives.

More than 500 equipments of this type were built, most of them by the production engineering group in charge of A. Wright and K. A. Chittick. Later this group redesigned the Block I equipment in order to make it more suitable for quantity production. The new design was called Block III equipment and was used by the armed services under actual combat conditions.

PART II

TRANSMISSION PROBLEMS IN AIRBORNE TELEVISION SYSTEMS

Introduction

Early flight tests of Block 1 equipment indicated that the demand for satisfactory performance under conditions of actual operation in military aircraft imposed extremely severe requirements on the operating characteristics of the equipment. Normally after take-off of the transmitting plane it was impossible to readjust any controls to compensate for changes in power-supply voltage, light conditions, signal strength, interference, temperature, and vibration. This frequently resulted in inferior performance because of poor picture quality, poor synchronization or both.

Investigation showed that some of the principal defects were caused by the following conditions:

(1) Microphonics.

(2) Power supply voltage variations.

(3) Interference caused by ignition-type noise or by other carrier frequencies, particularly those having radar modulation.
(4) Unstable synchronization.
(5) Variation in light conditions on pickup tube.
(6) Insufficient signal at receiver.
(7) Multi-path transmission and frequency modulation of the transmitter master oscillator.

Considerable time and effort has been devoted to these problems and some progress has been made toward satisfactory solutions. There follows an account of the work done to date.

1. *Microphonics*

There are two basic methods for approaching this problem:

(a) by mechanically isolating the unit involved from noise and vibration; and
(b) by providing electrical means for reducing the resultant effects in the units themselves.

Where especially severe conditions are encountered, a combination of the two methods is probably necessary in order to obtain satisfactory performance. It is hoped that the improvements developed in electrical circuits will make it possible to operate under normal conditions without elaborate mechanical isolation.

The most troublesome microphonics originate in those tubes in the video amplifier which are followed by maximum low frequency gain. There has been some improvement in performance resulting from better tube construction and it is hoped that even more rugged tube types will be available for future equipment. A considerable improvement in performance was also obtained by applying the clamp circuit principle used in pre-war orthicon equipment to the video amplifier. This circuit effectively allows the gain at low frequencies in the video amplifier, where microphonics are normally most troublesome, to be reduced almost to zero but does not interfere with the reproduction of low-frequency signal components. Circuits of various types employing this fundamental principle were tried and flight-tested and a relatively simple circuit easily adaptable to Block equipment was finally evolved.

In Figure 11 are shown modifications of some of the circuits in a Block I camera-transmitter unit. One of these modifications is the addition of a clamp circuit in the video amplifier. The double diode V23 receives pulses from the horizontal output transformer in such a manner that positive pulses are impressed on the plate of one diode and negative pulses on the cathode of the other diode. The negative pulses are also fed to the iconoscope grid in order to provide a constant reference voltage from the iconoscope during the horizontal blanking period. A complete explanation of the operation of the clamp circuit may be found in an article by C. L. Townsend.[3]

More effective shock mounts than the original design, were developed by the Robinson Company and contributed to a marked reduction in noise signal originating in microphonic tubes. However, it was apparent that under some conditions microphonics originated from noise reaching the tubes through the walls of the unit itself. It was possible to reduce this effect considerably by enclosing the unit in a separate container lined with sound absorbing material. In such cases it was also necessary to provide forced ventilation in the unit to prevent overheating.

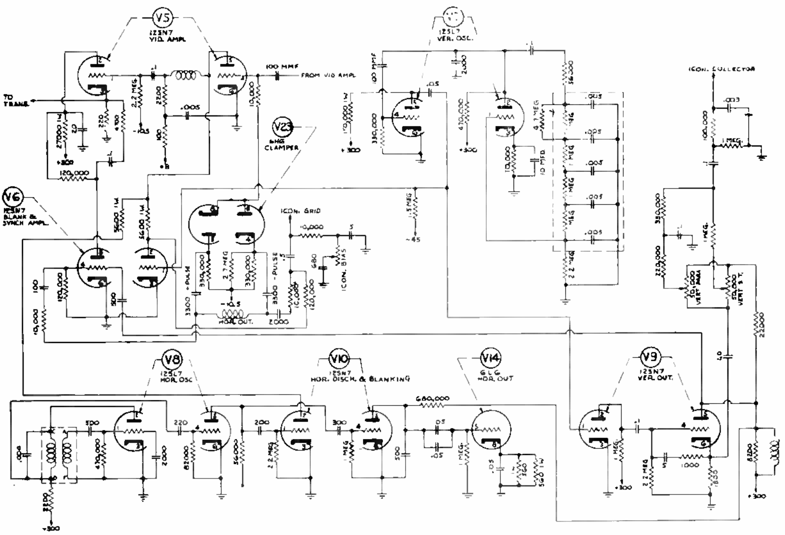

Fig. 11—Block I camera transmitter modifications.

2. *Power Supply Voltage Variations*

Occasionally the primary direct-current voltage source supplying the equipment was reasonably constant but frequently, especially in aircraft, the supply voltage was subject to considerable variation. This made the proper adjustment of equipment quite difficult since the voltage with the plane engine stopped or idling was usually very much lower than that obtained in flight. In order to overcome this difficulty some type of voltage regulator appeared to be essential. Automatic voltage regulator circuits, commonly used in other types of equipment

[3] C. L. Townsend, "The Clamp Circuit," *Broad. Eng. Jour.*, Vol. 12, No. 2, pp. 5-8, Feb., 1945; No. 3, pp. 5-7, March, 1945.

for the $B+$ supply, were adapted for use in Block equipment. In addition to providing a constant output voltage over a considerable range of primary supply voltage, this circuit has the advantage of providing a low impedance source for the various electrical circuits in the equipment, thereby eliminating to a large extent the necessity for large electrolytic by-pass capacitors normally used for preventing coupling at low frequencies between various circuits.

Some undesirable effects caused by variation in the primary voltage supplying heater power to the equipment still remained. It was possible to reduce these effects considerably by the use of devices such as ballast lamps, which provide a relatively constant current for a wide range of input voltage. This combination of a regulated $B+$ supply and ballast lamps on the low voltage supply resulted in satisfactory performance but the efficiency of these circuits was rather poor because of the power loss in the regulator tubes.

Recently it has been possible to obtain dynamotor power supplies which will supply a constant output voltage for the $B+$ supply over a considerable range of primary voltage change. These have the advantage of being much more efficient than the other devices described but it may still be desirable to use a regulated $B+$ supply in order to obtain a low impedance source for preventing crosstalk. Since this $B+$ regulator circuit does not need to be designed to accommodate a large variation of input voltage, the tubes can be designed to operate much more efficiently than in customary regulator service. Another possibility is the use of a circuit quite similar to the normal regulating circuit which will take out low frequency disturbances and provide a low impedance source but which will not regulate the direct-current voltage itself.

3. *Interference*

On some of the first flight tests it was discovered that radar equipment operating near the Block equipment carrier frequency would completely eliminate the picture signal appearing on the receiver. This would occur when sufficient interfering signal reached the automatic gain control circuit at the receiver to decrease the gain of the intermediate-frequency amplifier enough so that none of the desired signal was visible on the kinescope. A long series of experiments and flight tests were made in an effort to reduce this effect as much as possible.

For radar-type interference it was noticed that while the amplitude of the signal was large, the energy content was quite small. This characteristic was used to improve the operation of the automatic volume control. Under normal conditions the automatic volume control is controlled by the amplitude of the synchronizing pulses appearing at the output of the receiver second detector. Since the energy content of the synchronizing pulses is normally much greater than that of the radar pulses it was possible to alter the constants of the automatic volume control circuit in such a way that it operated to a large extent on signal energy rather than peak amplitude. This type of circuit is called the low impedance type automatic volume control circuit and is illustrated in Figure 12. It is capable of permitting satisfactory operation of the receiver under interference conditions from radar at least 100 times as severe as was formerly possible. Another type of "low impedance" automatic volume control circuit is illustrated in Figure 13. The input circuit to the automatic volume control diode is tuned to the line frequency in order to discriminate between the synchronizing pulses intended for rectification and the interfering radar pulses. This circuit, however, showed little improvement over the preceding one since shock excitation of the tuned circuit by radar or noise pulses also contributed to the automatic volume control bias developed.

A still further improvement was made possible by introducing pulses from the horizontal output circuit into the automatic volume control circuit in such a way that signals can reach the automatic volume control detector only during a short interval corresponding to slightly more than the period of the horizontal synchronizing pulse. This circuit is called a "keyed automatic volume control circuit" and prevents all the interference occurring during the interval in which picture signal is transmitted from affecting the operation of the automatic volume control. An improvement of approximately 10 to 1 was noted over the normal low impedance automatic volume control circuit. This type of circuit is shown in Figure 14.

Suitable limiter circuits on the signal output to the automatic volume control circuit are also helpful in improving performance. However, these circuits require rather careful adjustment so that the limiting level is always somewhat higher than the amplitude of the synchronizing signal itself. This also means that the automatic volume control characteristic must be quite flat so that with any reasonable value of signal input the synchronizing level will not exceed the limiter adjustment.

4. *Instability of Synchronization*

Another effect of noise and interference was to disturb seriously the synchronization of the picture at the receiver. In many cases this resulted in a loss of picture intelligence much greater than that caused by the presence of the interference in the picture signal itself. During

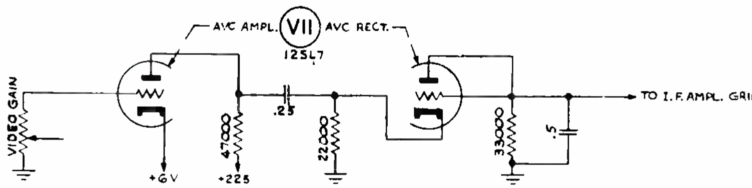

Fig. 12—Low impedance automatic volume control circuit.

flight tests it was frequently necessary for the operator at the receiver to adjust the hold controls in order to keep a stationary picture on the kinescope screen.

A fundamental difficulty with the first Block equipment lay in scanning oscillator instability. Both the line and frame frequency scanning oscillators were cathode-coupled multivibrators whose frequency was a function of a large number of variables including the supply voltage,

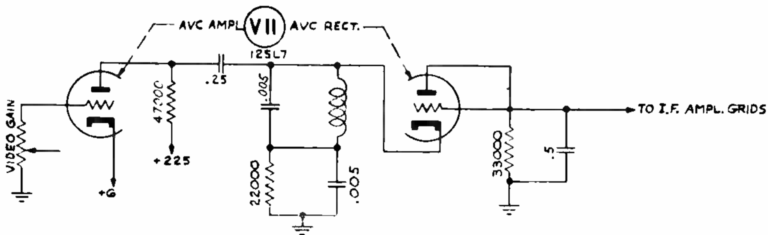

Fig. 13—Low impedance automatic volume control circuit with tuned circuit discrimination.

ambient temperature, and all the resistors and capacitors used in the feedback circuit. This type of oscillator has the advantage that it produces pulse wave shapes which can be used directly for blanking and synchronizing signals. Various other types of relaxation oscillators were tried in an effort to find some whose frequency stability was superior to those in use. However, none of them exhibited a frequency stability comparable to that obtained from a sine wave oscillator.

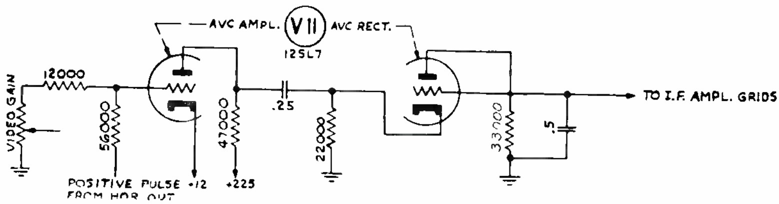

Fig. 14—Keyed low impedance automatic volume control circuit.

It was finally decided to use sine-wave type oscillators for both line and frame frequency scanning circuits and relatively simple shaping circuits were developed to obtain the necessary blanking and synchronizing pulses from the wave shape obtained from the sine wave oscillator. The horizontal oscillator was of the inductance-capacitance type with an adjustable powdered iron core for tuning the circuit to the proper frequency. The vertical oscillator was of the resistance-capacitance network type. This was used in place of a conventional inductance-capacitance oscillator because of the large size of the inductance required for a 40-cycle oscillator of the latter type. The stabilization of these scanning frequencies at the transmitter resulted in a considerable improvement in performance since the receiver synchronizing circuits did not have to compensate for any changes of frequency at the transmitter. It was also possible to adjust two different camera transmitter units to have exactly the same scanning oscillator frequencies so that a receiver could be tuned from one transmitter to another without requiring readjustment of the synchronizing hold controls. Figure 11 shows modifications made in the scanning oscillator circuits of a Block camera transmitter unit in order to permit the use of stable oscillators.

One-half of the 12SL7 (V8) is the horizontal frequency oscillator. The other half is an amplifier and clipper of the pulse-wave form appearing across the 2200-ohm resistor in the plate circuit of the oscillator. The plate circuit of the pulse amplifier is coupled to both triodes of the 12SN7 (V10). One triode is the horizontal discharge tube which develops a saw-tooth in its plate circuit for the horizontal output tube. The other triode is the horizontal blanking amplifier supplying blanking to the 12SN7 clipper tube (V5) in the video amplifier. One-half of the 12SL7 (V7) is the vertical scanning oscillator; the other half is an amplifier stage which amplifies a vertical pulse obtained by differentiating the plate voltage wave of the oscillator. One-half of the 12SN7 (V9) is the vertical discharge tube which converts the pulse to a saw-tooth wave form. The other half of (V9) is the vertical output tube which feeds the output transformer. Vertical shading voltages are also obtained from the plate and cathode voltages of this tube. Vertical blanking signal is obtained from one-half of 12SN7 (V6). In the plate circuit the horizontal and vertical blanking signals and the video signal are mixed together to be impressed on the clipper grid. The other half of (V6) is the synchronizing signal amplifier which obtains its grid voltage from differentiating circuits connected to the horizontal and vertical output circuits.

However, even with this improvement severe noise and interference conditions caused the picture on the receiver to be unstable. Radar interference of insufficient intensity to interfere with the picture signal was enough to cause the picture edges to be quite irregular so that a considerable amount of picture intelligence was lost from that cause. To improve this condition the automatic frequency control principle was applied to the scanning oscillators in the receiver. This circuit uses a phase detector which operates in such a way that it maintains the pulse from the synchronizing signal and the pulse from the horizontal output circuit within a very short time interval of each other. The output of the phase detector operates on a reactance tube which changes the frequency of the scanning oscillator in such a way that the proper phase relationship is maintained. This circuit has a sufficiently long time constant so that noise pulses of short duration have practically no effect on the bias supplied to the reactance tube.

In this way the average frequency of the scanning oscillator at the receiver is made the same as the average frequency of the scanning oscillator at the transmitter. If, for instance, both horizontal oscillators maintain this constant frequency over a period greater than one vertical frame then the picture edges will automatically be exactly vertical. The longer the time constant of this circuit the better noise immunity it will have for low frequency interference. However, the long time constant circuit requires an appreciably longer time to come into synchronism if the signal is temporarily lost. Another difficulty noticed with the long time constant circuit on the horizontal oscillator is the lateral movement of the picture caused by rapid changes in the path length of the signal reaching the receiver. This happens only when the receiver is obtaining its incoming carrier signal over two separate paths, one of them usually a direct path and the other a reflection from the ground or some large object. The best value of the time constant of the horizontal automatic frequency control is a matter of compromise. Experiments indicate that an optimum value will permit the horizontal receiver scanning oscillator to stay in synchronism with the transmitting scanning oscillator even though a synchronizing signal may be lost for one complete frame and still allow a reasonably fast pull-in after temporary loss of synchronizing signal. The time constant for the automatic frequency control circuit controlling vertical oscillator speed is more difficult to adjust. In order to obtain superior noise immunity over the standard lock-in circuit the time constant must be made long compared to one vertical frame. This makes the recovery time after a temporary loss of signal too slow to be satisfactory. What is necessary is a circuit which has a long time-constant while the vertical oscillator is locked in but permits a short recovery time when the oscillator is out of step. One outstanding advantage of this circuit arrangement for both horizontal and vertical oscillators is that it practically relieves the operator at the receiving location from the necessity of adjusting the receiver hold controls once they have been properly adjusted.

Figure 15 shows the horizontal and vertical lock-in circuits of a Block 1 receiver modified to include automatic frequency control of both scanning oscillators. Referring to Figure 15, one-half of the 12SL7 (V24) is the horizontal sine-wave oscillator. The 6AC7 (V20) is a reactance tube operating in a conventional circuit. A change in the direct-current bias of this tube causes it to change its gain. Since the plate current of the tube is almost 90 degrees out of phase with the plate voltage due to the resistance capacity network in its grid circuit, the change in gain causes a change in frequency because of the change in reactive current through this tube which is in shunt with the oscillator tank circuit. The bias which operates the reactance tube and thereby changes its frequency comes from a 6H6 discriminator tube (V19) which receives signal from a synchronizing amplifier and horizontal synchronizing separator (V14). The plate circuit of the separator has a transformer which supplies a synchronizing signal of opposite polarity to the two discriminator diodes. A horizontal pulse from the horizontal output is introduced into the discriminator circuit in such a way that the same polarity pulse appears on both diodes. When the phase relationship between the incoming synchronizing pulses and the pulses from the horizontal deflecting circuit changes, the direct-current output from the discriminator changes. If the phase shift is in one direction, the direct-current output is positive; if in the opposite direction, the direct-current output is negative. This bias applied to the reactance tube tends to shift the oscillator frequency just enough to keep the synchronizing pulse and the horizontal output pulse in proper phase relationship. The time-constant of the resistance-capacitance network in the output of the discriminator determines the speed of response of the horizontal automatic frequency control system. If this time constant is slow, the oscillator will hold its frequency over longer periods of absence of synchronizing signal than is possible with a short time constant. However, if the oscillator is out of synchronism either at the time the receiver is turned on, or because of excessive noise or lack of signal, the time of recovery is much slower with the long-time constant circuit than with the other.

It may be noted that the horizontal discriminator circuit is not balanced with respect to ground. Test indicated that it was highly desirable to have the discriminator output nearly zero when no input

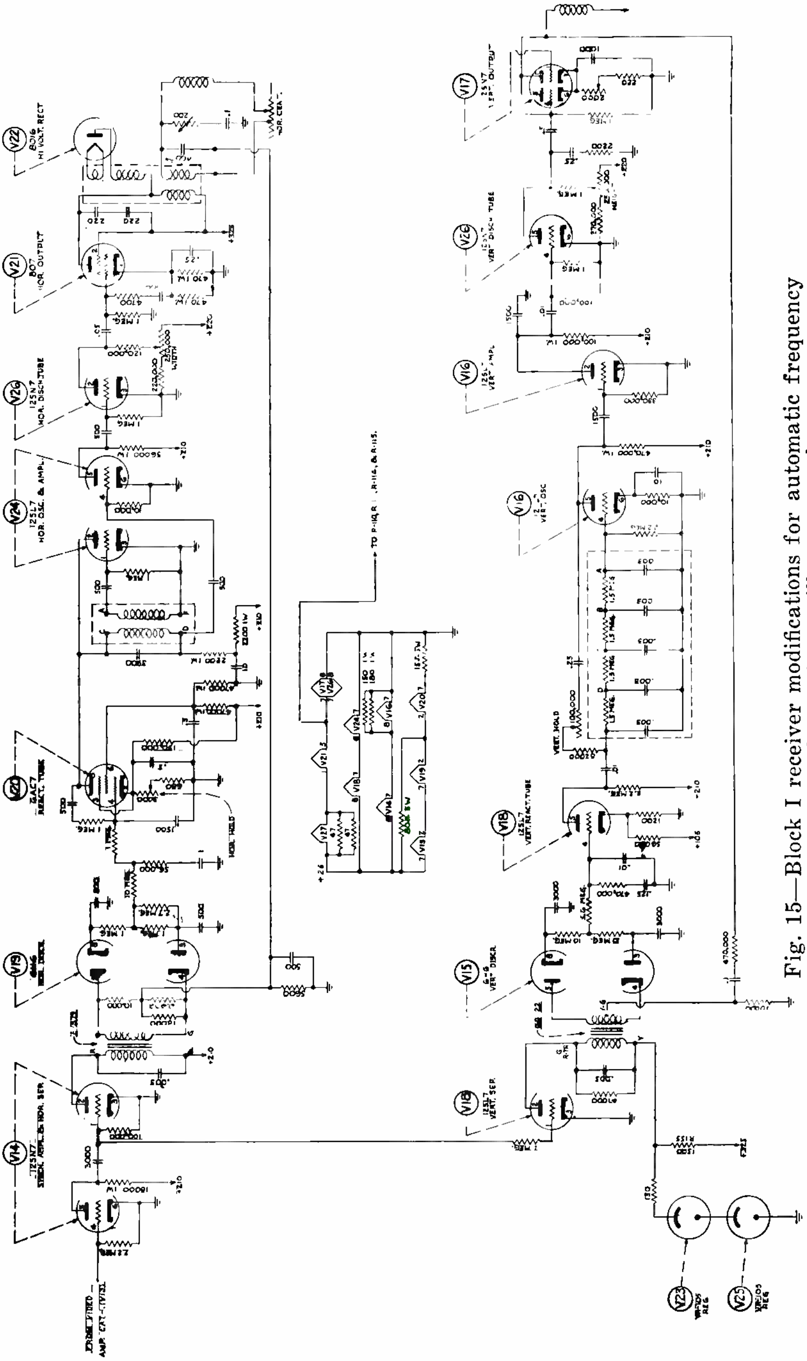

Fig. 15—Block I receiver modifications for automatic frequency control scanning oscillator control.

signal was present. Since a pulse is used from the horizontal output instead of a saw-tooth, the discriminator output will not be zero unless the circuit is unbalanced with respect to ground. If this condition is not fulfilled, the recovery or pull-in time is quite long because of the necessity for charging the one-microfarad capacitor through the 10-megohm resistor to a definite direct-current potential.

The speed control for the horizontal hold adjustment is a rheostat in the cathode circuit of the reactance tube (V20). A powdered iron plug in the oscillator coil acts as an auxiliary speed control. The cathode voltage adjustment also makes it possible to change the sensitivity of the automatic frequency control system. When the control is near the lowest value of cathode voltage possible, the sensitivity is greatest. With this adjustment the horizontal oscillator will lock in about 400 cycles above and below the 14,000 cycle normal frequency. At the highest value of cathode voltage, this range drops to about 100 cycles each way.

The automatic frequency control circuit for the vertical oscillator is quite similar to the circuit used for the horizontal oscillator. The oscillator is one-half of a 12SL7 (V16) with the resistance-capacitance feedback circuit mentioned before. Here the reactance tube changes the impedance to ground of one leg of the resistance-capacitance network. By changing the bias on the grid of one-half of the 12SL7 (V18) the plate impedance of the tube varies and therefore the impedance in series with the .01-microfarad capacitor to ground also changes, thereby changing the frequency. The vertical separator (V18) and discriminator (V15) have the same circuit arrangement as the equivalent horizontal circuits. Vertical synchronizing pulses appear on the two discriminator diodes of the 6H6 with opposite polarities and a vertical sawtooth from the vertical output has the same polarity on both diodes. The resultant direct-current voltage which is generated when the diodes are unbalanced tends to shift the oscillator frequency so that the correct phase relationship between the vertical synchronizing pulses and the vertical saw-tooth output is maintained. The time constant used here is again a compromise between the time required for the absence of synchronizing signal to cause the oscillator to fall out of synchronism and the time for recovery of synchronism once the oscillator is out of step. A choice is more difficult here than with the horizontal circuit because of the low frequency of vertical scanning. The time constant should normally be quite long compared to the time required for one picture so that the vertical automatic frequency control circuit, in effect, obtains its synchronizing information from a large number of vertical pulses. However, if this time constant is very long, and the vertical oscillator does lose synchronism so that the vertical blanking pulse appears near the center of the picture, a number of seconds may be required before the automatic frequency control circuit can reestablish normal synchronization.

The vertical hold control is a rheostat in a resistive element of the resistance-capacitance oscillator network. With a reasonably strong signal, the automatic frequency control circuit will keep the vertical oscillator in step over a range of plus or minus two cycles variation from the normal 40-cycle frame frequency.

5. *Variation in Light Conditions*

The difference in light conditions prevailing at even moderate altitudes compared to the light conditions on the ground is usually extremely great. The normal characteristics of a scene observed from an airplane during daylight include an extremely high light level and very low contrast, caused by either haze or smoke in the atmosphere. This complicates the problem of adjusting the equipment on the ground for satisfactory performance in the air. For the early tests of Block 1 equipment, a test bench was developed for setting up the camera transmitter unit on the ground and was arranged so that a slide with very little contrast could be projected on the iconoscope mosaic at a light level comparable to that obtained during flight. This, in general, improved performance considerably over that obtained when preliminary adjustments were made for a scene on the ground.

A number of experiments were made in an effort to improve the signal output from the iconoscope under high light level and low contrast conditions. It was discovered that many of the earlier iconoscopes would saturate at very-high light levels and would produce considerable noise output and very little signal. In many cases improved performance was obtained by a reduction in the aperture of the lens used in the camera. The effect of various types of filters was studied but the results were somewhat disappointing since the insertion of a filter which cut down the effect of haze also reduced the signal output from the iconoscope to the point where hiss noise from the first video amplifier tube became quite noticeable. In most cases it was found that a Wratten #25 orange filter would improve picture contrast without seriously increasing noise.

An investigation of the saturation phenomena in the iconoscope by the tube engineers resulted in a slight modification in the tube which permitted better operation at high light levels. A high-light test was also included in the iconoscope test specifications which would insure satisfactory operation.

Two methods for obtaining a more nearly constant signal output from the iconoscope were also investigated. One method was the development of an automatic volume control circuit which would change the gain of the video amplifier as a function of its signal output. This was somewhat difficult because of the spurious signal delivered by the iconoscope during the blanking intervals and the spurious signals commonly referred to as "dark spot" and "edge flare".

Two types of automatic volume control circuits were developed. In one circuit only the high frequency components of the video signal were used to operate the automatic volume control system. This operated satisfactorily as long as there was a sufficient amount of fine detail in the picture from which high frequency signals could be obtained. A circuit somewhat more satisfactory was made by keying the automatic volume control circuit in such a way that only the picture information in the central part of the picture was used to operate the automatic volume control. In actual flight tests these circuits showed very little, if any, advantage over the normal circuit because the usual light conditions normally required all of the video gain available consistent with a reasonably satisfactory signal to noise ratio. An automatic iris control was also suggested for this purpose and was actually developed, although primarily for use with the orthicon tube.

There were two other possibilities suggested for improving the performance of the iconoscope under flight conditions: (1) the possibility of introducing automatic shading circuits; and (2) the possibility of controlling automatically the beam current of the iconoscope itself. These investigations were discontinued because it was felt that the development of the small orthicon and the image orthicon would eliminate the necessity for circuits of this type.

6. *Low Signal Strength*

With a carrier power output of approximately 15 watts from the transmitter of Block I equipment the reliable operating range from aircraft to ground was approximately 10 miles. Operation beyond this range was possible but the signal would sometimes be lost because of the change in position of the plane. An increase in this range was considered very desirable but it was felt that considerable increase in transmitter power would be necessary for any appreciable improvement in performance. This would, of course, mean a much larger unit at the transmitter with a corresponding increase in weight and power drain. It was decided first to improve the signal-to-noise ratio of the receiver as much as possible before attempting to provide more transmitter power. It was also considered desirable to provide the receiver operator with a tuning adjustment so that interference effects from carrier frequencies close to the desired frequency could be minimized as much as possible. A new type of head-end tuner for the receiver was developed which had the following advantages:

(a) an improvement in signal-to-noise ratio of at least 2 to 1 over previous circuits;

(b) considerable improvement in the radio-frequency selectivity and prevention of excessive oscillator radiation by the inclusion of a radio-frequency stage; and

(c) better performance, as evidenced by flight tests of a receiver with the improved tuner, not only because of the improved sensitivity but because it was possible by slight changes of tuning to eliminate or greatly reduce interference near the edge of the channel. (This was impossible with the earlier fixed-tuned type receivers.)

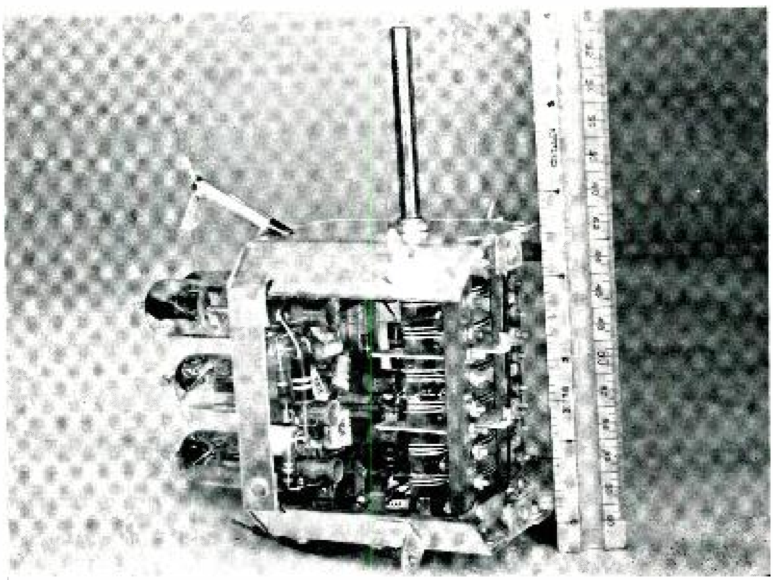

Fig. 16—Block I receiver tuning unit—top cover removed.

It is estimated that the addition of this tuner increased the effective operating range of the receiver to at least 25 miles. A photograph of one of these tuners is shown in Figure 16 and a schematic diagram is shown in Figure 17.

The general principle of this tuner, which has wide application, is that of connecting appreciable fixed inductance between the low potential side of the variable capacity and ground or virtual ground. Thus, inductance is added to the circuit as the variable capacitance is increased. This action occurs most effectively when the minimum capacitance is small compared to the distributed capacitance immediately external to the capacitor. Under this condition, the circuit current follows a shorter path when the variable capacitor is at minimum than when at maximum.

With ideal capacitor construction and with the largest useful auxiliary inductance, the maximum possible tuning ratio should increase from two to three by the addition of the inductance. In practice, of course, the maximum tuning range of 2 cannot be realized because of limited capacity range. Hence, the addition of inductance in the capacitor may increase the tuning range by more than the theoretical maximum factor of 3 to 2. In the actual application, the frequency

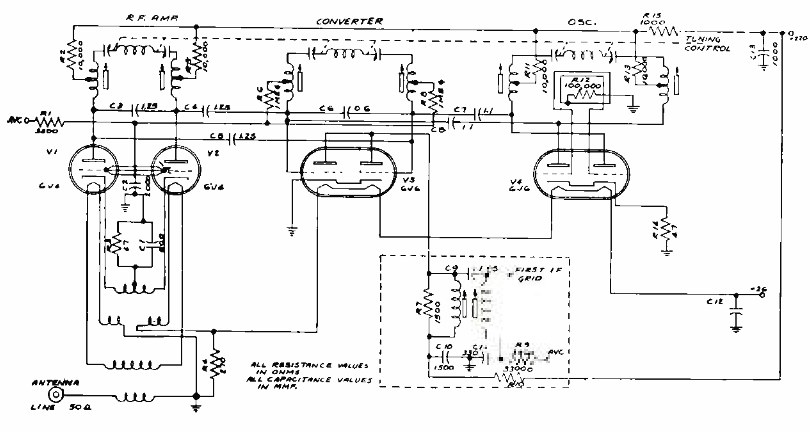

Fig. 17—Block I receiver tuning unit—schematic diagram.

coverage was doubled by the inductance. However, the tuning ratio has been limited to less than 2 in order to reduce the crowding at the low-frequency end of the dial when semi-circular plates are used. "Midline" or straight-line-frequency plates may also be used to improve dial linearity. Inasmuch as the tube capacitance at one end of the circuit is in series with the tuning capacitor at the other end, there is little to be gained by using a tuning capacitor with a greater maximum than two or three times the tube capacitance.

7. *Multipath Transmission*

One of the troublesome problems encountered in the transmission of television signals from one plane to another is the effect of two out-of-phase signals of nearly the same field strength reaching the receiver simultaneously. This frequently causes two pictures to appear simultaneously on the face of the picture tube. The one having lower intensity is sometimes called a ghost. Synchronization difficulty is also experienced because of the presence of two signals. Considerable improvement can be obtained by using directional antennas and by locating them in such a way that reflections from the ground are minimized.

The multipath problem becomes considerably more serious if frequency modulation is present on the transmitter carrier especially if the frequency modulation is caused by video signal. Peculiar interference patterns are caused by the combination of multipath transmission and transmitter frequency modulation which may be so strong as to destroy entirely the usefulness of the picture. When transmitting from plane to ground the difficulty usually arises because a second signal arrives at the receiving location, reflected from a large building or other similar object in the area near the receiving location. When the transmission path is from plane to plane the second signal at the receiver is the one arriving by virtue of a reflection from the ground. A similar effect can be obtained even though the receiver is at a ground location if the antenna is located high enough so that a ground-reflected signal will have a phase difference, compared to the direct signal, sufficient to cause trouble. How this may come about is explained briefly in the following paragraphs.

Consider the elementary propagation system consisting of a transmitter (T), a receiver (R), the propagation medium, and the earth's surface (E). In the simplest case, the signal is transmitted to the receiver, R, from the transmitter, T, over just two paths, one being the direct path TR and the other the indirect path, or path of reflection TER. The resultant signal at R is the vector sum of the two.

Obviously, the direct signal from T will arrive at R ahead of the indirect signal, the time difference, t, depending upon the heights of the transmitter and receiver as well as the distance between them. Thus, if a simple rectangular pulse of very short duration as compared to t is sent from the transmitter, two pulses will be received, spaced t seconds apart. Now let us assume that the transmitter is sending out a continuous signal of frequency f_1 and after a period of time the frequency is suddenly changed to f_2 for a time much less than t after which the frequency again returns to f_1. No amplitude changes occur at the transmitter during the transition from f_1 to f_2 to f_1. In t seconds after the pulse of frequency f_2 has left the transmitter two signals will be received, f_1 and f_2. Both will be demodulated at the receiver producing a beat note having a frequency $f_1 \pm f_2$ but only during the times when different frequencies arrive by the two paths. The frequency $f_1 + f_2$ is too high to be accepted by the receiver, but $f_1 - f_2$ may be a very low frequency falling within the video acceptance band. When this is the case, the beat note will appear superimposed on the picture signal.

In actual television practice, large repetitive changes in modulation occur mainly at the beginning and ending of the horizontal blanking pulses. This means that, when the reaction from power amplifier to oscillator is appreciable, there is a beat note of large amplitude during the horizontal return time, or fly-back, with consequent vertical bar patterns appearing in the picture. Often, the disturbance is further complicated by the presence of not just one but several paths of reflection so that a corresponding number of beat notes can appear simultaneously and for varying lengths of time.

Because of this difficulty it is extremely important to keep frequency modulation of the transmitter to an absolute minimum. A buffer stage between the transmitter oscillator and modulator is almost an absolute necessity. It is also important that mechanical vibration of the oscillator circuit elements be kept below a level which will cause excessive frequency modulation. In general, it seems that more than 1 per cent frequency modulation can cause noticeable effects in the picture.

* * *

The development of Block I television equipment and the subsequent investigations carried out in an effort to improve the performance were the collective work of a large group of engineers working under the direction of R. D. Kell, in charge of television research at RCA Laboratories Division, and G. L. Beers, Assistant Director of Engineering in charge of Advanced Development at RCA Victor Division. The development and field test of this equipment would have been at best extremely difficult without the whole-hearted interest and cooperation of NDRC, Military, and Naval personnel associated with this project. This paper covers work done in whole or in part under the following contracts: W535sc238, NOs-86775, PDRC-29, NXs3405-A, and OEMsr-441, all with Radio Corporation of America.

FUNCTIONAL SOUND ABSORBERS*

BY

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Summary—*A functional sound absorber is a sound absorbing means designed for the single purpose of absorbing sound as contrasted to a conventional sound absorbing material which serves a two-fold function, namely; as a building or wall material and as a sound absorber. Apparatus for the measurement of direct-current acoustic resistance, acoustic impedance and the sound absorption characteristic of functional sound absorbers is described. Functional sound absorbers of four different sizes and four different shapes have been built and tested. The absorption coefficient of the functional sound absorber is on the order of two times that of conventional materials. Installations of the functional sound absorber illustrate the use and the high absorbing efficiency.*

INTRODUCTION

CONVENTIONAL sound absorbing materials are designed to serve a two-fold function, namely; as a building or wall material and as a sound absorber. Because of this compromise the sound absorbing efficiency is low. There are certain applications where the principal problem is to absorb sound; there are some rooms where conventional materials cannot be applied to the ceiling and walls. For these applications the logical solution is the use of a functional sound absorber of relatively high absorbing efficiency. The amount of energy absorbed depends upon the sound pressure of the source and the acoustic impedance of the medium and the absorber. The acoustic impedance of the medium and the absorber are controlled by the design of the absorber. This suggests that it should be possible to obtain high absorbing efficiency by appropriate design. A study has been made of this problem and a functional sound absorber of high efficiency has been developed. It is the purpose of this paper to describe this functional sound absorber and discuss its high efficiency.

THEORY OF THE FUNCTIONAL SOUND ABSORBER

In an acoustical system, acoustic resistance is the element responsible for the dissipation of energy. Accordingly, the absorption of sound requires an acoustic resistance. An acoustic resistance in which the

* Decimal classification: R800 (534).

acoustical energy is transformed into heat energy seems to be the most direct. When air, either alternating or direct, is forced through small narrow passages, the acoustical energy is converted into heat energy due to the high viscosity of air when it is forced through such passages. Thus any material with narrow passages, such as wood, mineral or metal fibers, cloth, felt, laminations of paper, fiber or metal and the like, forms an acoustic resistance. In some of the materials additional acoustic resistance is introduced by bending, flexing, shearing and rubbing of the vibrating fibers due to the passage of sound. By appropriate processing, practically any desired value of acoustic resistance can be obtained from any of the above materials.

Conventional acoustic absorbing materials are employed as a wall

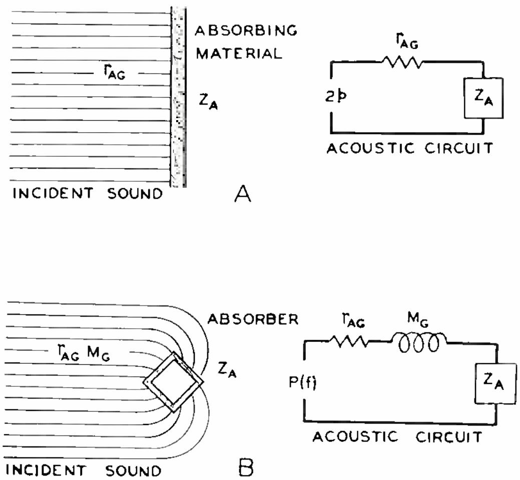

Fig. 1—(A) In conventional materials placed on the boundaries of the room, the incident sound wave is essentially plane. In the acoustic circuit: $p =$ free field sound pressure; $r_{AG} =$ characteristic acoustic resistance of air; and $Z_A =$ acoustic impedance of the material. (B) In the unit sound absorber, the incident sound wave is not plane but flows into the material. In the acoustic circuit: $P(f) =$ generator sound pressure (The generator sound pressure is a function of the frequency and the characteristics of the absorber); $r_{AG} =$ acoustic resistance of the generator; $M_G =$ inertance of the generator; $Z_{AG} = r_{AG} + j\omega M_G$; and $Z_A =$ acoustic impedance of the absorber.

covering on the boundaries of the room. The absorbing mechanism may be depicted by an acoustic network with lumped constants. A consideration of the acoustic circuit of Figure 1(A) shows that the maximum absorption occurs when the acoustic impedance, Z_A, of the material is an acoustic resistance equal to the characteristic acoustic resistance, r_{AG}, of air. Under these conditions, the absorption of sound is 100 per cent. In the case of the absorbing wall the maximum efficiency that can be obtained is 100 per cent because the ratio of the area of the wavefront to area of the wall is unity. In most practical materials, 100 per cent absorption is not attainable because a material with high absorption characteristics is not suitable as a wall material and the average absorption is usually about 70 per cent. To increase the absorption beyond 100 per cent, requires a reduction in the value of the generator impedance. This can be accomplished by the use of diffraction as shown in Figure 1(B). For this condition, the value of the source impedance, Z_{AG}, can be made very small. A consideration of the acoustic circuit shows that an appropriate value for the absorption impedance, Z_A, will yield an absorption coefficient which is more than unity. In the case of diffraction, the absorption may be more than 100 per cent because the ratio of the effective area of the wavefront absorbed to the area of the absorber may be more than unity. The above lumped constant theory applies in the region in which the dimensions of the absorber are small compared to the wave length.

Design of the Functional Sound Absorber

The preceding section indicates that the efficiency of sound absorption may be increased by the use of a unit type sound absorber. The problem is to develop a practical functional sound absorber. In this connection, low cost is one of the prime considerations. This in turn means a simple manufacturing process coupled with a minimum amount of raw material. The functional sound absorber may be built in several different ways as shown in Figure 2. In (A), a single shell of acoustic resistance, r_{AM}, encloses a volume constituting the acoustic capacitance C_{AC}. In (B), a shell of acoustic resistance r_{AM1} encloses a second shell of acoustic resistance r_{AM2}, the two resulting volumes being the acoustic capacitances C_{AC1} and C_{AC2}. In (C), the material is homogeneous and may be represented as a large number of acoustic resistances, $r_{AM1} - r_{AMN}$, and acoustic capacitances $C_{AC1} - C_{ACN}$. A consideration of the acoustic networks of Figure 2 shows that the material will be used at the maximum efficiency if the acoustic resistance is in the form of a single shell as shown in Figure 2(A). A photograph of a single shell type conical functional sound absorber is shown in Figure 3. The conical shape was chosen because it was comparatively easy to build and test. From the standpoint of absorption alone, a spherical shape would be slightly superior. The acoustic network of the conical functional sound absorber is shown in Figure 4. It will be noted that there

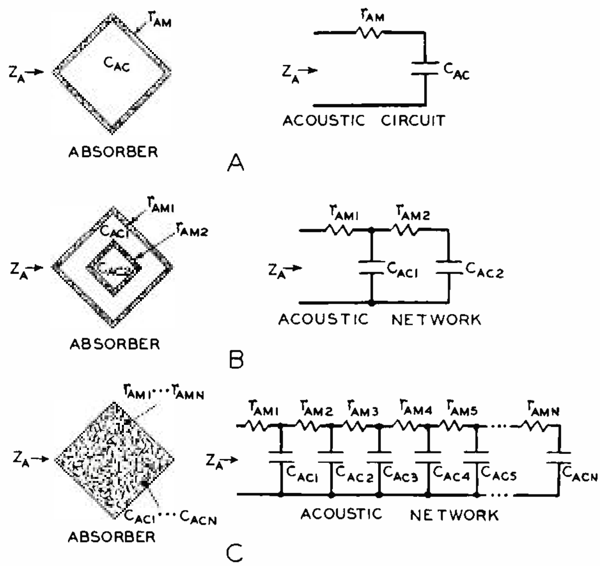

Fig. 2—Functional sound absorbers: (A) Single shell of acoustic material. In the acoustic circuit: Z_A = acoustic impedance of the absorber; r_{AM} = acoustic resistance of the absorber; and C_{AC} = acoustic capacitance of the volume within the shell. (B) Two shells of acoustic material, one within the other. In the acoustic network: r_{AM1} = acoustic resistance of the outer shell; r_{AM2} = acoustic resistance of the inner shell; C_{AC1} = acoustic capacitance of the volume between the shells; and C_{AC2} = acoustic capacitance of the volume of the inner shell. (C) A homogeneous material. In the acoustic network: $r_{AM1} — r_{AMN}$ = distributed acoustic resistance; and $C_{AC1} — C_{ACN}$ = distributed acoustic capacitance.

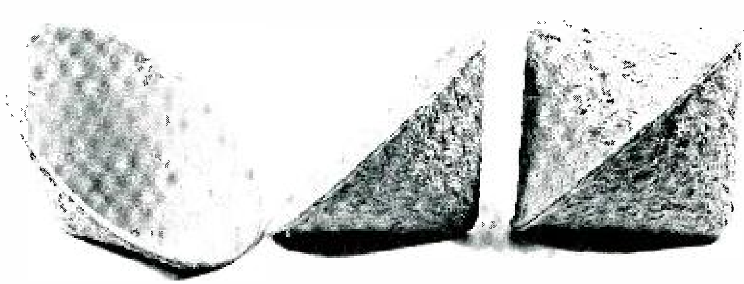

Fig. 3—Single shell functional sound absorber.

are two types of acoustic resistance, namely, viscous and mechanical. The acoustic resistance due to viscosity plays the most important role in absorption of sound in the functional sound absorber of the type shown in Figure 3.

MEASUREMENTS

One of the important characteristics of a sound absorbing material is the direct-current acoustic resistance. The direct-current acoustic resistance r_{ADC} can be obtained from the volume current and the pressure as follows:

$$r_{ADC} = \frac{p}{U} \tag{1}$$

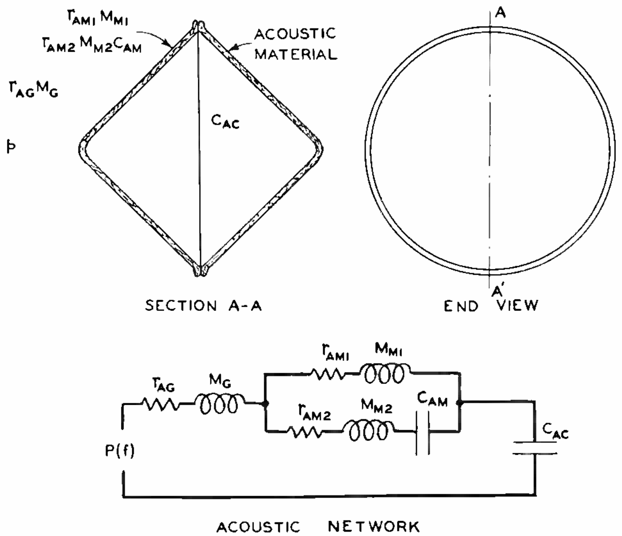

Fig. 4—Single shell functional sound absorber. In the acoustic network: p = free field sound pressure; $P(f)$ = generator sound pressure; r_{AG} = acoustic resistance of the generator; M_G = inertance of the generator; r_{AM1} = acoustic resistance due to flow through the material; M_{M1} = inertance due to flow through the material; r_{AM2} = acoustic resistance due to the mechanical resistance of the acoustic material; M_{M2} = inertance due to the mass of the acoustic material; C_{AM} = acoustic capacitance due to the compliance of the acoustic material; and C_{AC} = acoustic capacitance of the cavity within the shell of acoustic material.

where p = difference in pressure between the surfaces of the material, in dynes per square centimeter, and

U = volume current through the material, in cubic centimeters per second.

Apparatus[1] for measuring direct-current acoustic resistance is shown in Figure 5. The difference in pressure between the two sides of the acoustic material is measured by an inclined manometer. Water is syphoned out of the large flask at a constant rate. The volume current or rate of change of volume displacement can be obtained from the volume displacement and the time.

The acoustic impedance is another important characteristic of

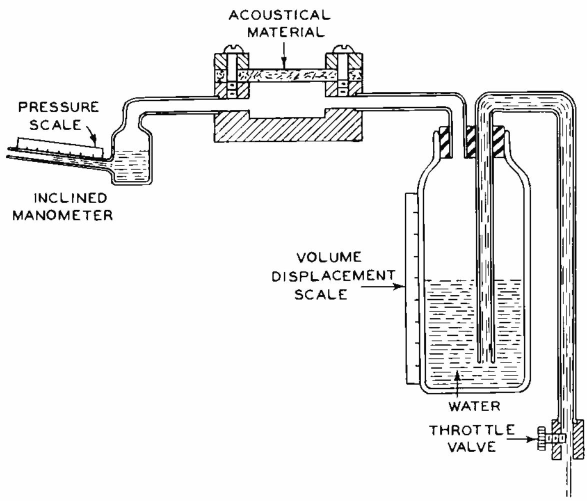

Fig. 5—Schematic of the apparatus for measuring the direct-current acoustic resistance of acoustic material.

sound absorbing material. The use of an acoustic tube for measuring the acoustic impedance of absorbing materials was first proposed by H. O. Taylor[2]. Modifications of this method have been made by many investigators[3]. A schematic of the system used for measuring acoustic

[1] L. L. Beranek, "Acoustic Impedance of Porous Materials", *Jour. Acous. Soc. of Amer.*, Vol. 13, No. 3, pp. 248-260, January, 1942.

[2] H. O. Taylor, "A Direct Method of Finding the Value of Materials as Sound Absorbers", *Phys. Rev.*, Vol. II, No. 4, pp. 270-287, October, 1913.

[3] H. J. Sabine, "Notes on Acoustic Impedance Measurement", *Jour. Acous. Soc. of Amer.*, Vol. 14, No. 2, pp. 143-150, October, 1942.

impedance in these experiments is shown in Figure 6. The absorption coefficient is given by

$$\alpha = \frac{4MN}{(M+N)^2} \tag{2}$$

$$\alpha = 1 - K^2 \tag{3}$$

where M and N = maximum and minimum sound pressure in the standing wave system, and

K = pressure reflection coefficient.

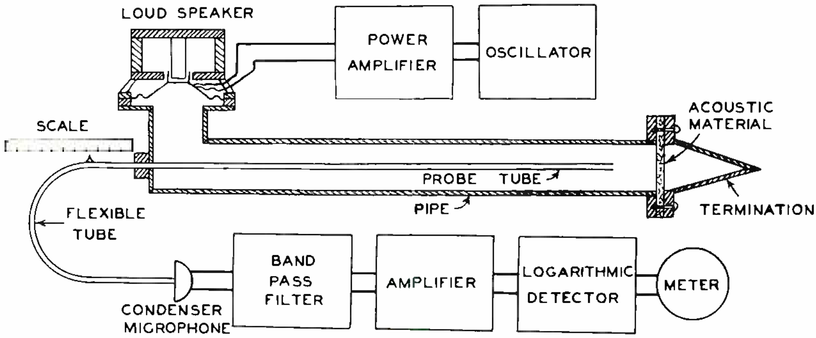

Fig. 6—Schematic of the apparatus for measuring the acoustic impedance of acoustic materials.

The acoustic resistance and reactance per unit area is given by,

$$r_A = \frac{1 - K^2}{1 + K^2 + 2K\cos\left(\dfrac{2\pi D_1}{D_2}\right)}\,\rho c \tag{4}$$

$$x_A = \frac{2K \sin \dfrac{2\pi D_1}{D_2}}{1 + K^2 + 2K \cos \dfrac{2\pi D_1}{D_2}}\,\rho c \tag{5}$$

where D_1 = distance from the surface of the material to the first pressure minimum, in centimeters,

D_2 = distance between first and second pressure minima, in centimeters,

c = velocity of sound in air, in centimeters per second, and

ρ = density of air, in grams per cubic centimeter.

Measurement of the absorption in a reverberation chamber is the classical method for obtaining the performance of a material under field conditions. A schematic of the apparatus for measuring the reverberation time is shown in Figure 7. The trace of sound decay is depicted on a cathode-ray tube with a persistence image screen. Decay is observed over a range of 48 decibels. A transparent time scale over the front of the tube is used to read the reverberation time. The decay sequence is repeated every nine seconds. The beat frequency oscillator is warbled to reduce the effects of standing wave systems and thereby obtain a smoother decay trace. As a further aid in smoothing the decay response, six loud speakers and six microphones are used. A photograph of the reverberation chamber is shown in Figure 8. The volume of the reverberation chamber is 3400 cubic feet.

The total absorption[4,5,a] in the reverberation chamber, in sabins, is given by

$$A = .050 \frac{V}{T} \tag{6}$$

where V = volume of the chamber, in cubic feet, and

T = time required for the sound level to decay to one-millionth of its original intensity, in seconds.

The total number of sabins of absorption of any acoustical material may be obtained by measuring the reverberation time before and after the introduction of the material in the reverberation chamber. The total absorption of the acoustical material is the difference between the total absorption after and before the introduction of the material.

The acoustic absorptivity (or absorption coefficient) of a surface is the ratio of the flow of sound energy into the surface on the side of

[4] W. C. Sabine, COLLECTED PAPERS IN ACOUSTICS, Harvard University Press, Cambridge, Mass., 1923.

[5] C. F. Eyring, "Reverberation Time in 'Dead' Rooms", *Jour. Acous. Soc. of Amer.*, Vol. 1, No. 2, pp. 217-241, January, 1930.

[a] Equation 6 holds when the effective absorption coefficient of the chamber is small.

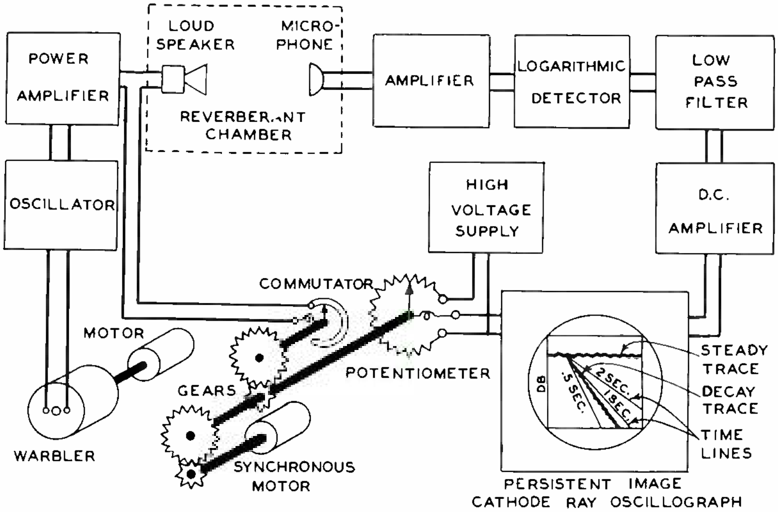

Fig. 7—Schematic of the apparatus for measuring the reverberation time of the reverberation chamber.

incidence to the incident rate of flow. The sabin is a unit of equivalent absorption and is equal to the equivalent absorption of one square foot

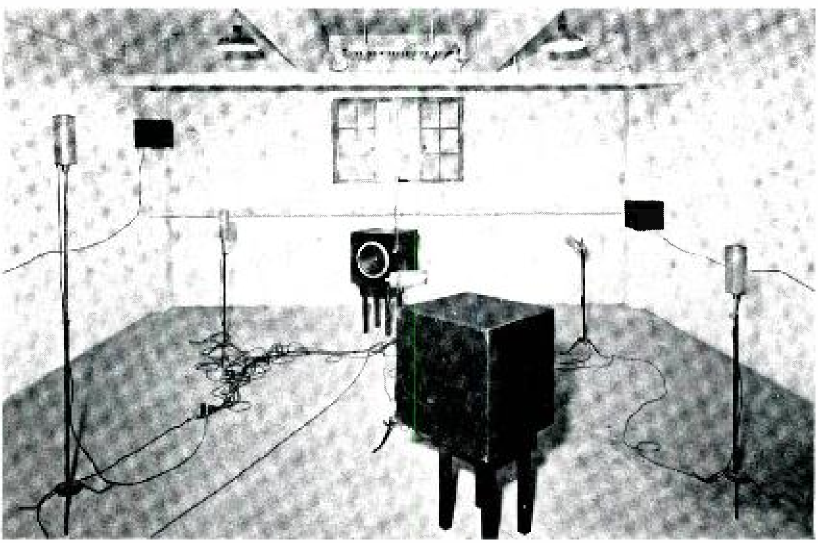

Fig. 8—Reverberation chamber.

of a surface with unit absorptivity.

The absorption coefficient of the functional sound absorber was obtained by dividing the measured absorption in sabins by the area in square feet.

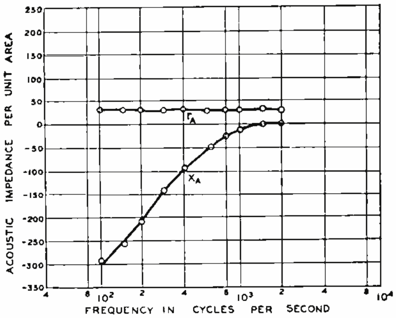

Fig. 9—Acoustic resistance, r_A, and reactance, X_A, of an acoustic material with a low acoustic resistance. Direct-current acoustic resistance = 30 acoustic ohms per unit area.

ABSORBING EFFICIENCY AS A FUNCTION OF THE ACOUSTIC IMPEDANCE

The acoustic impedance of three different combinations of shredded wood and a binder, designated as low, medium and high acoustic

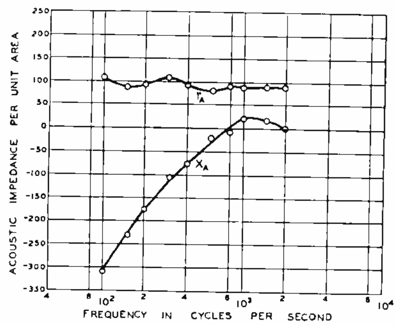

Fig. 10—Acoustic resistance, r_A, and reactance, X_A, of an acoustic material with a medium acoustic resistance. Direct-current acoustic resistance = 92 acoustic ohms.

impedance materials, are shown in Figures 9, 10 and 11. The termination for the material is a metal cone. This cone simulates the termination in the cone absorber at the low frequencies. The direct-current acoustic resistance component measured by means of the apparatus of

Figure 5 was 30, 96 and 320 acoustic ohms per unit area for the low, medium and high acoustic impedance samples. This compares with 30, 92 and 230 for the alternating-current acoustic resistance. The variation in acoustic resistance for the three materials covers a range

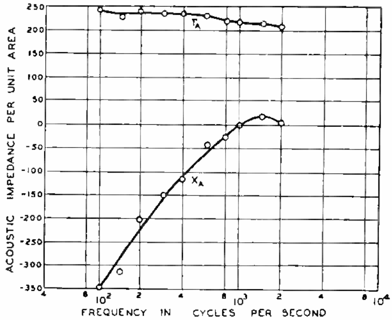

Fig. 11—Acoustic resistance, r_A, and reactance, X_A, of an acoustic material with a high acoustic resistance. Direct-current acoustic resistance = 320 acoustic ohms.

of about nine to one, in steps of three to one.

The absorption coefficient frequency characteristics of functional sound absorbers obtained from reverberation time measurements are shown in Figure 12. The diameter at the base of the cone of the

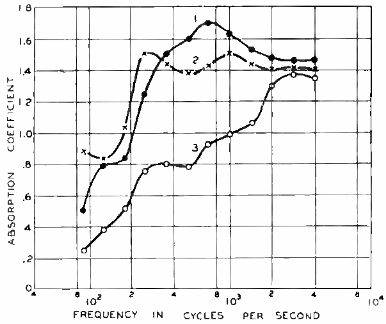

Fig. 12—Absorbing efficiency of the conical functional sound absorber as a function of the acoustic impedance. 1. Acoustic material with low acoustic resistance (Fig. 9). 2. Acoustic material with medium acoustic resistance (Fig. 10). 3. Acoustic material with high acoustic resistance (Fig. 11).

absorber is fourteen inches. In these tests twelve absorbers of each type were used. Two absorbers were placed on each boundary of the chamber. The absorption coefficient was obtained by dividing the absorption in sabins per absorber by the area of the absorber in square feet. As would be expected from a consideration of the acoustic circuit, the absorption efficiency for the material with a high acoustic resistance is inferior to the materials with medium and low acoustic impedances. Comparing the absorbing efficiency of materials with medium and low acoustic impedances there is not enough difference to be of any practical significance. However, it was felt that the absorption coefficient frequency characteristic of the material with medium acoustic impedance was the most suitable. Therefore, this material was used in the remainder of the tests.

Absorbing Efficiency as a Function of Size

The generator acoustic impedance and the absorber acoustic impedance depend upon the dimensions of the functional sound absorber. The absorption coefficient frequency characteristic, of four sizes of functional sound absorbers shown in Figure 13, were obtained from reverberation time measurements. The absorption coefficient frequency characteristics as a function of size are shown in Figure 14. These results are in general agreement with the results to be expected from a consideration of the acoustic circuit, namely; that the efficiency of absorption at the high frequencies will increase with a decrease in the dimensions of the absorber, and that the efficiency at the lower frequencies, where the acoustic reactance due to the acoustic capacitance plays an important part, will increase with an increase in the dimensions.

Fig. 13—Four different sizes of conical functional sound absorbers (7-, 10-, 14- and 21-inch cone bases.)

DIFFRACTION AND ABSORBING EFFICIENCY

It is apparent from the preceding tests that the high absorbing efficiency is due to diffraction. To test the effect of diffraction, the

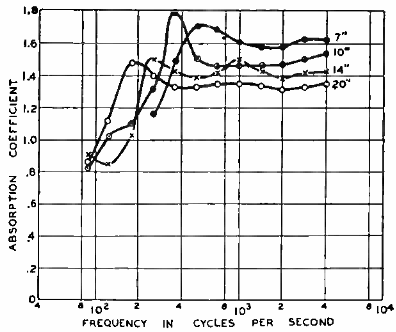

Fig. 14—The absorbing efficiency of the conical functional sound absorber as a function of size. The dimension refers to the diameter at the base of the cone.

material used in the absorber wall was nailed to strips which spaced the material two and one half inches from the floor. The area used was seventy-two square feet. In the next test forty-nine cones, fourteen inches in diameter at the base, were placed edge to edge on the floor in a pattern seven by seven. These results are practically the same as that of the flat material. Following this, twenty-four cones were spaced

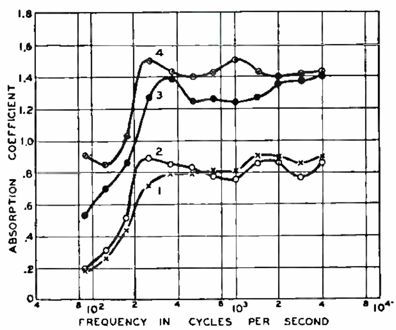

Fig. 15—Diffraction and absorbing efficiency. 1. Seventy-two square feet of absorber material in flat form spaced two and one half inches from floor. 2. Forty-nine fourteen-inch cones spaced edge to edge with the base resting on the floor. 3. Twenty-four fourteen-inch cones spaced three feet apart on the floor. 4. Twelve conical, fourteen-inch functional sound absorbers, two on each surface of the room.

three feet apart on the floor. The results of these three tests are compared to those obtained using twelve assembled absorbers in Figure 15. In the mid and high frequency ranges the absorbing efficiency is almost doubled due to diffraction. At the low frequencies the absorbing efficiency is tripled due to diffraction.

ABSORBING EFFICIENCY AS A FUNCTION OF SHAPE

The simple theory indicates that the difference in sound absorbing

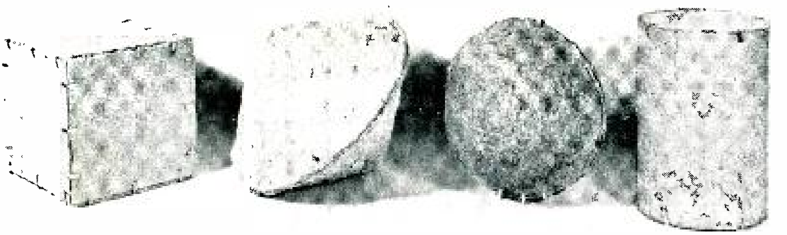

Fig. 16—A cubical, conical, spherical and cylindrical functional sound absorber all of the same volume.

efficiency between a cubical, conical, cylindrical or spherical absorber should not be very great. To test the effect of shape, a cube, cylinder and sphere each of the same volume as the fourteen inch conical absorber were formed by hand tools. These formed absorbers are shown in Figure 16. The results of absorption tests on the formed shapes are

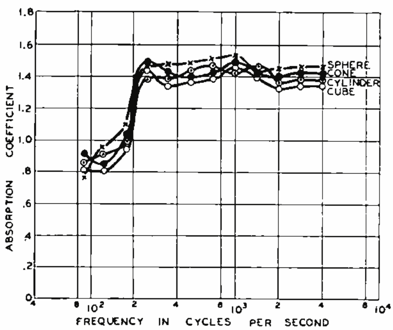

Fig. 17—Absorbing efficiency of a cubical, conical, spherical and cylindrical functional sound absorber all of the same material and same volume.

shown in Figure 17. The sphere exhibits the greatest efficiency. Of course, this is to be expected because the ratio of volume to surface area is the largest for the sphere.

ABSORBING EFFICIENCY AS A FUNCTION OF THE POSITION IN THE ROOM

Position in the room influences the pressure and acoustic impedance of the source and as a consequence affects the absorbing efficiency. Figure 18 shows the absorption coefficient frequency characteristic of twelve absorbers under three different conditions, as follows: two on each surface of the room; in three rows on the floor; and suspended three feet above the floor on two strings spaced two feet apart on the strings. Spacing from a reflecting surface reduces the efficiency in the midrange with no appreciable change at the low and high frequencies.

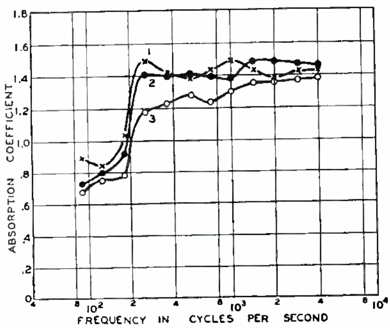

Fig. 18—Absorbing efficiency of a functional sound absorber as a function of the position in the room. 1. Twelve, two on each surface of the room. 2. Twelve, on the floor in three rows. 3. Twelve, on two strings two feet apart on the strings, strings three feet from floor.

ABSORBING EFFICIENCY AND WEIGHT OF MATERIAL

An indication of the efficiency of sound absorption may be obtained from the sound absorption in sabins per pound of material. The average absorption, in sabins per pound of material, for the frequencies 256, 512, 1024 and 2048 for the fourteen-inch conical functional sound absorber and a well-known conventional wall type high acoustic impedance absorbing material made from shredded wood and a binder is depicted in Figure 19. The ratio of sound absorption per pound of material of twelve-to-one in favor of the functional sound absorber indicates the high efficiency of sound absorption exhibited by this system.

Typical Installations

Four installations of the functional sound absorber have been made at the laboratories. These installations were made in rooms with truss type ceiling structures. Employing conventional acoustical materials under these conditions would be very costly. In addition, the use of conventional acoustical materials in the ceiling would impair the sky-lighting arrangements. One of the installations of the functional sound absorber was made in the cabinet shop. This shop has a truss roof with a ceiling height ranging from thirteen to eighteen feet. The functional sound absorbers were hung at a level of ten feet, with a

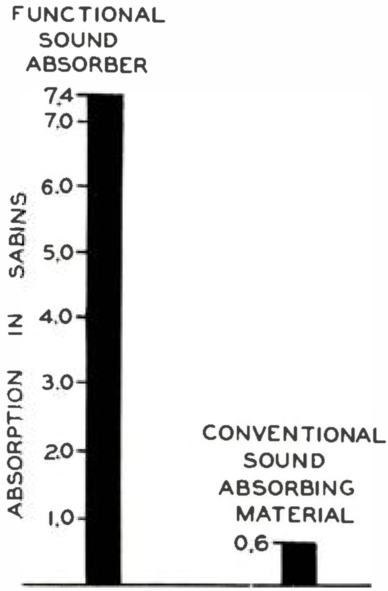

Fig. 19—Comparison of the absorption in sabins per pound of material for the functional sound absorber to a typical conventional sound absorbing material.

density of one in each five square feet, as shown in the photograph of Figure 20. The noise level was reduced more than 6 decibels by the installation of the functional sound absorber. Before treatment, the noise produced by rotary saws, band saws, hammers, planers, etc. would "hang on". In addition, it was impossible to localize sound. After treatment the hammering and sawing seemed to be softer and subdued and it was possible to localize the sound and thereby discriminate against the offending noise. These subjective observations were ob-

Fig. 20—Installation of the functional sound absorber in the cabinet shop.

tained from the personnel of the cabinet shop. Another installation of the functional sound absorber was made in the drafting room (Figure 21). This is a very long room which had very disturbing

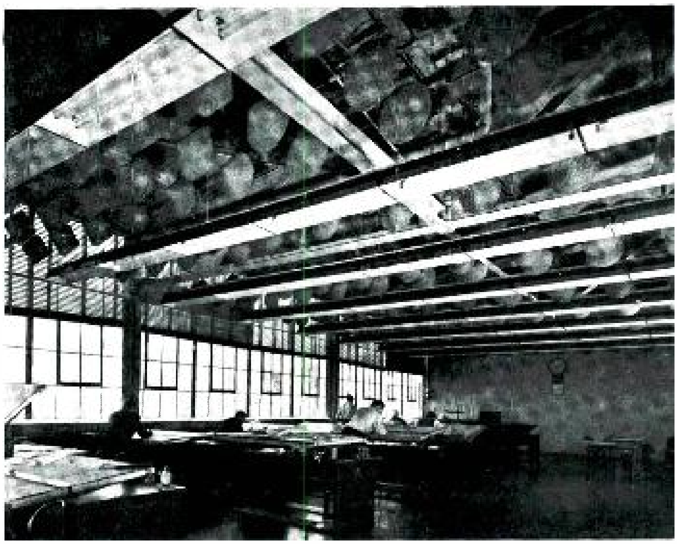

Fig. 21—Installation of the functional sound absorber in the drafting room.

echos and excessive reverberation before the installation of the functional sound absorbers. The drafting room has a truss roof with a ceiling height ranging from twelve to twenty feet. The functional sound absorbers were hung at a level of ten feet with a density of one in each five square feet. The echos and excessive reverberation were eliminated and the noise level was reduced about 6 decibels. Another installation was made in the blueprint room. The machinery noise in this location was very disturbing to the operators. The use of the functional sound absorber corrected the excessive noise condition. Another installation was made in the expeditors' office in the model shop. Being in the model shop this was a very noisy location. The principal complaints were difficulties in carrying on telephone conversations and conducting conferences. The application of the functional sound absorber corrected the noise condition and made it possible to carry on normal office functions without difficulty.

Another feature of importance in the application of sound absorbing material is the time and expense required for installing the material. The drafting room floor area is 1500 square feet. Three hundred functional sound absorbers were used. Three workmen, who had no previous experience in installing these absorbers, required only a little less than two days to make this installation. To apply conventional acoustical material would necessitate the construction of a ceiling supporting means below the steel truss structure. This would be a very costly procedure. In addition, the new ceiling would impair the skylighting and roof ventilating systems. The use of the functional sound absorber does not interfere with existing ventilating and lighting systems.

Discussion of Results

The experimental results outlined in this paper are in qualitative agreement with the performance deduced from theoretical considerations. The absorption is increased due to diffraction, the absorption coefficient increases at the higher frequencies with a decrease of the dimensions of the absorber, the absorption decreases for relatively large and small values of acoustic resistance, and in the case of various shapes the largest absorption coefficient is obtained when the ratio of surface area to volume is a minimum. The qualitative agreement with theory does not mean that the functional sound absorber is in quantitative agreement with theory. For example, the average experimental absorption coefficient as outlined in this paper is 1.4, whereas, from theoretical considerations, sound absorbers of this type should exhibit more than this measured value. Some of this discrepancy may be attributed to measurements. It appears that there is a wide-open field for improvement in the existing functional sound absorber, in spite of the fact that it is many times more efficient than existing materials. Conventional sound materials have been in the process of research and development for two decades and there is little hope for improvement in efficiency because the difference between theoretically possible values and the actual results are very small. On the other hand, the functional sound absorber is a new and revolutionary idea in the field of sound absorption and the possibilities of improvement in efficiency beyond that already obtained and discussed in this paper are tremendous.

FREQUENCY MODULATION DISTORTION CAUSED BY COMMON- AND ADJACENT-CHANNEL INTERFERENCE*

By

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Summary—*During frequency-modulated radio broadcasting the signal is liable to be badly distorted whenever multipath transmission occurs or when any other interfering signal is present on the same or an adjacent channel. During hot weather, or before a storm, long-distance reception has been observed from frequency modulation broadcast stations on the 42–50 megacycle band. When such a distant station was in the same channel as a desired station, it sometimes happened that for short intervals the undesired station became stronger than the desired one. When this happened there was a small amount of noise and the programs suddenly changed. This interchange often lasted for several seconds but sometimes was limited to a word or two or a few notes of music.*

Formulas are given for computing the amplitudes of the harmonics and cross-modulation frequencies produced by the interference. These enable the calculation of the effect of a de-emphasis network following the discriminator, of a low-pass audio filter, and of nonlinear phase shift in the amplifiers.

INTRODUCTION

FOR several years it has been evident that frequency-modulated radio broadcasting offers certain advantages in noise reduction when compared with the usual amplitude-modulation systems. Many papers describe and discuss frequency modulation systems and their noise-suppressing properties.[1–8] Extensive field tests showed[9] that

* Decimal Classification: R148.2 × R430.

[1] Edwin H. Armstrong, "A Method of Reducing Disturbances in Radio Signaling by a System of Frequency Modulation," *Proc. I.R.E.*, Vol. 24, No. 5, pp. 689–740; May, 1936.

[2] Murray G. Crosby, "Frequency Modulation Noise Characteristics," *Proc. I.R.E.*, Vol. 25, No. 4, pp. 472–514; April, 1937.

[3] H. Roder, "Noise in Frequency Modulation," *Electronics*, Vol. 10, No. 5, pp. 22–25, 60, 62, 64; May, 1937.

[4] E. H. Plump, "Störverminderung durch Frequenzmodulation," *Hochfrequenztechnik und Elektroakustik*, Vol. 52, pp. 73–80; September, 1938.

[5] Stanford Goldman, "F-M Noise and Interference," *Electronics*, Vol. 14, No. 8, pp. 37–42; August, 1941.

[6] Harold A. Wheeler, "Common-Channel Interference Between Two Frequency-Modulated Signals," *Proc. I.R.E.*, Vol. 30, No. 1, pp. 34–50; January, 1942.

(See opposite page for References 7, 8 and 9.)

when frequency modulation was used there was less interference produced by two stations operating at the same frequency than for the corresponding case of amplitude modulation, and that less power was required to cover a given area. It was also found that when the ratio of the carrier voltage to the noise voltage is high, the signal-to-noise ratio improvement due to frequency modulation is considerable. As the interfering noise voltage is increased with respect to the desired carrier-wave voltage, the improved noise suppression is obtained as long as the desired signal is several times as strong as the noise.

When a definite carrier-to-noise voltage ratio is reached (a ratio of 2 or 3 for wide-band frequency modulation) the amount of distortion in the audio output increases rapidly. When the noise voltage exceeds the signal voltage during all parts of the audio cycle, the noise eliminates the desired signal. This means that when frequency modulation is used the signal is either good or bad; there is only a small range for the ratio of carrier voltage to noise voltage that gives a noisy, but tolerable, signal.

Multipath transmission occurs when two or more interfering signals come from the same transmitter, but one is delayed with respect to the others because of a longer transmission path. Considerable distortion has been observed when multipath transmission occurs in frequency-modulated broadcasting and fairly complete discussions of this problem are available.[10-13] If the second wave comes from a different station than the desired wave, the result is common- or adjacent-channel interference according to whether the two carrier frequencies are nearly the same or are separated by the width of one channel.

There is not much information available on the amount of interference to be expected in the new frequency modulation band. The effects to be described were observed on the old 42–50 megacycle band and on the 30–42 megacycle police bands. The frequency of occurrence and the magnitude of these effects will not be known for the new 88–108 megacycle band until a reasonable number of transmitters with normal power and antenna gains are in operation. If such interference does occur, the analysis given here will be applicable.

[7] Herbert J. Reich, "Interference Suppression in A-M and F-M," *Communications*, Vol. 22, No. 8, pp. 7, 16, 19, 20; August, 1942.

[8] Robert N. Johnson, "Interference in F-M Receivers," *Electronics*, Vol. 18, No. 9, pp. 129–131; September, 1945.

[9] I. R. Weir, "Field Tests of Frequency- and Amplitude-Modulation With Ultra-High-Frequency Waves," *Gen. Elec. Rev.*, Vol. 42, Nos. 5 and 6, pp. 188–191, May, 1939; pp. 270–273, June, 1939.

[10] Murray G. Crosby, "Observations of Frequency-Modulation Propagation on 26 Megacycles," *Proc. I.R.E.*, Vol. 29, No. 7, pp. 398–403; July, 1941.

[11] A. D. Mayo and Charles W. Sumner, "F.M. Distortion in Mountainous Terrain," *Q.S.T.*, Vol. 28, No. 3, pp. 34–36; March, 1944.

[12] Murlan S. Corrington, "Frequency-Modulation Distortion Caused by Multipath Transmission," *Proc. I.R.E.*, Vol. 33, No. 12, pp. 878–891; Dec., 1945.

[13] S. T. Meyers, "Nonlinearity in frequency-modulation radio systems due to multipath propagation," *Proc. I.R.E.*, Vol. 34, No. 5, pp. 256–265; May, 1946.

Sometimes during hot weather, or before a storm, long-distance transmission has been observed from frequency modulation broadcast stations. During the summer of 1944, station WSM-FM in Nashville, Tenn. was heard often in Camden, New Jersey. During July it was very strong and free of noise for nine evenings in succession and it was heard several other evenings. Occasionally, long-distance reception from stations in all directions was observed. On July 7, 1944 nearly all the mid-western stations and several from other directions could be received in Camden, for about 2½ hours with a standard commercial receiver and indoors antenna. The following list of stations received was compiled that evening:

Call	*Station*	*Megacycles*
WWZR	Zenith Radio Corp., Chicago	45.1
WGNB	WGN, Inc., Chicago	45.9
WBBM-FM	Columbia Broadcasting System, Chicago	46.7
WDLM	Moody Bible Institute, Chicago	47.5
WSBF	South Bend Tribune, South Bend, Ind.	47.1
WMLL	Evansville on the Air, Evansville, Indiana	44.5
WENA	Evening News Assn., Detroit	44.5
WMFM	The Journal Company, Milwaukee, Wisconsin	45.4
WSM-FM	National Life & Accident Ins. Co., Nashville, Tenn.	44.7
WMIT	Gordon Gray, Winston-Salem, N. C.	44.1
WMTW	Yankee Network, Mt. Washington, N. H.	43.9
W2XMN	Edwin H. Armstrong, New York	43.1
WHNF	Marcus Loew Booking Agency, New York	46.3
WBAM	Bamberger Broadcasting Service, New York	47.1
WABC-FM	Columbia Broadcasting System, New York	46.7
WABF	Metropolitan Television, Inc., New York	47.5
WIP-FM	Pennsylvania Broadcasting Co., Philadelphia	44.9

Some interesting common-channel phenomena were observed. Stations WENA, Detroit, and WMLL, Evansville, were of nearly equal strength. First one, and then the other was received; they changed about every fifteen seconds. There would be a slight amount of noise and the programs would suddenly be interchanged. This continued for about one-half hour. Sometimes the carrier-wave voltage levels dropped below the level at which the limiter in the receiver operated and both programs could be heard simultaneously.

Stations WSBF, South Bend, and WBAM, New York, were also in a common channel. WSBF was stronger and was clear most of the time; WBAM would come in with sudden bursts of a word or two or a bit of music as station WSBF faded rapidly. These bursts occurred at intervals of about ten seconds.

Some of the state police frequency-modulation systems have reported serious skip interference on numerous occasions. In Missouri, on the talk-back frequency of 39.78 megacycles, the interfering signals are usually those of the New Jersey State Police and the North Carolina Highway Patrol Cars, although cars of the Ohio State Patrol and those of Rhode Island occasionally cause interference. The signal strengths of the undesired stations are greatest during May, June, and July and range from weak to strong. The strong signals are of sufficient intensity to swamp out the local cars and may be received for an hour or two or for the whole day, from about two hours after sunrise to an hour or so after sunset.

The Florida State Patrol have reported considerable interference on frequency modulation from stations in California, New Jersey, Connecticut, and Massachusetts, and they have made car-to-car contacts with Pittsfield, Massachusetts. The Michigan State Police reported that signals from the Alabama State Patrol stations were received by their patrol cars with signal levels at the input to the receiver as high as 300 microvolts, and these stations in Alabama have taken control of their receivers throughout Michigan for hours at a time.

The Indiana State Police have had their cars blocked out by stations in Virginia and Oklahoma for all cars more than three miles from the transmitter. During the hunt for escaped German war prisoners near Carlisle, Indiana, on June 10th, the interference was so bad they had considerable difficulty maintaining contact with their cars. On June 22nd, during a man-hunt and road blockade following a bank holdup at San Pierce, Indiana, cars were completely blocked out at various times by cars in Virginia and Massachusetts. Further disruption of service was caused many afternoons by the second harmonic of short-wave broadcast stations in Massachusetts and New York.

Recent observations by the Federal Communications Commission show that such bursts or sudden increases in strength of signals received beyond the line of sight occur regularly.[14-15] The long-distance transmission that occurs during such bursts can be interpreted as reflections from media of height comparable to the E layer, but lying at each side of the great-circle plane. It is assumed that when meteors pass through the upper atmosphere, the air is ionized and this causes the bursts.

If a local station is on the same channel as a distant one which is being received in bursts, interference may be expected to occur for intervals as

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[14] "Measurement of V-H-F Bursts," *Electronics*, Vol. 18, No. 1, p. 105; January, 1945.

[15] K. A. Norton and E. W. Allen, Jr., "Very-High-Frequency and Ultra-High-Frequency Signal Ranges as Limited by Noise and Co-Channel Interference," *Proc. I.R.E.*, Vol. 33, No. 1, p. 58; January, 1945.

long as several seconds. This might even cause the program to change suddenly from one station to the other during these short intervals.

Analysis of Fundamental Case

The most elementary case of frequency modulation interference is that produced when two unmodulated radio-frequency carriers, having nearly the same frequency, are added together. This gives the usual heterodyne envelope as the two voltages beat together. In addition there is a variation in the phase of the resultant which is equivalent to frequency modulation. If the difference in frequency of the two carriers is now varied sinusoidally by changing the frequency of one, keeping the two amplitudes constant, the result is common-channel interference or adjacent-channel interference, depending upon the way the one frequency is varied. It is thus evident that, if the most elementary case is properly analyzed, the frequency modulation interference is merely a generalization of the results.

Heterodyne Envelope

As shown in Appendix I, if two radio-frequency carriers $e_1 \sin \omega t$ and $e_2 \sin (\omega + 2\pi\mu)t$ are added, the heterodyne envelope is given by

$$\text{Envelope} = e_1\sqrt{1 + x^2 + 2x \cos 2\pi\mu t} \tag{1}$$

where

e_1 = amplitude of first carrier
e_2 = amplitude of second carrier
x = e_2/e_1
ω = angular frequency of first carrier, radians per second
μ = difference in frequency, cycles per second

This is the voltage that will be obtained if the resultant signal is sent through a linear rectifier and filtered. Figure 1 shows the variation of the envelope over one beat-note cycle as the ratio of the amplitudes of the two signals, x, is changed. For small values of x the envelope is approximately,

$$\text{Envelope} = e_1(1 + x \cos 2\pi\mu t) \qquad x \ll 1 \tag{2}$$

As the ratio x is increased gradually, the higher harmonics increase in amplitude; so the peaks become broader and the hole in the carrier becomes deeper and narrower. In the limit, as $x \rightarrow 1$, the envelope becomes a series of rectified cosine waves, or:

$$\text{Envelope} = 2e_1 \left| \cos \pi\mu t \right| \qquad x = 1 \tag{3}$$

Average Value of Envelope. If the resultant heterodyne voltage is sent through a linear rectifier, the direct-current voltage across the rectifier output increases gradually as x is increased. Figure 2 shows that this voltage increases 27.3 per cent when x changes from zero to one. As shown in Appendix I, this voltage is given by:

$$\text{Average voltage} = \frac{2(1+x)e_1}{\pi} \, E\left\{\frac{2\sqrt{x}}{1+x}\right\} \tag{4}$$

where $E\left\{\frac{2\sqrt{x}}{1+x}\right\}$ is a complete elliptic integral of the second kind with modulus $\frac{2\sqrt{x}}{1+x}$.

Root-Mean-Square Value of Envelope. If a square-law rectifier instead of a linear rectifier is used, the root-mean-square value of the rectified envelope can be read with an average-reading direct-current voltmeter. The root-mean-square voltage will increase more rapidly with x than the average

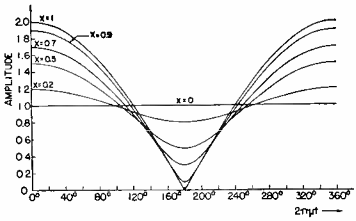

Fig. 1—The heterodyne envelope.

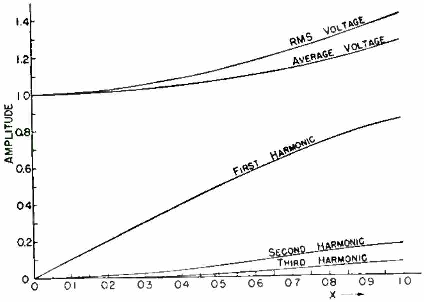

Fig. 2—Harmonic content of the heterodyne envelope.

voltage, as shown by Figure 2. The voltage is given by:

$$\text{Root-mean-square voltage} = e_1\sqrt{1+x^2} \tag{5}$$

and it increases 41.4 per cent when x increases from zero to one.

Fourier-Series Analysis of Envelope. If the heterodyne envelope is rectified with a linear rectifier, and the radio frequency is filtered out, the resultant audio signal (shown by Figure 1) can be expanded in a Fourier series. The coefficients of this series are given in Appendix I and the zero-frequency component is the same as the average value which is shown by Figure 2. The fundamental component increases almost linearly with increasing x to a maximum value of $2/3$ of the corresponding direct current voltage. The second harmonic increases slowly until it equals 20 per cent of the fundamental when $x = 1$, and the third harmonic has a maximum value of 8.6 per cent of the fundamental.

Phase-Angle Variations

The two signals $e_1 \sin \omega t$ and $e_2 \sin (\omega + 2\pi\mu)t$ are in phase when $t = 0$. Since the frequency of the second signal is higher than the frequency of the first signal, this means that a vector representing e_2 will rotate with respect to one representing e_1. If e_1 is a vector rotating at ω radians per second, then e_2 will rotate at $\omega + 2\pi\mu$ radians per second.

Figure 3 shows the variation of the phase angle φ which the resultant, R, of e_1 and e_2 makes at any given instant with the vector e_1. When $t = 0$, the two vectors are in phase and $\varphi = 0$. At a later time $2\pi\mu t =$ 90 degrees, so e_2 and e_1 are at right angles and $\tan \varphi = e_2/e_1 = x$. When $2\pi\mu t = 180$ degrees, φ is again zero. This process gives the variations in φ shown by Figure 4. The maximum value of φ is equal to $\sin^{-1} x$, as shown

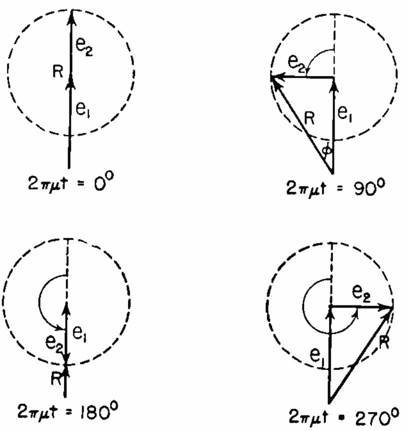

Fig. 3—Variations of the phase angle.

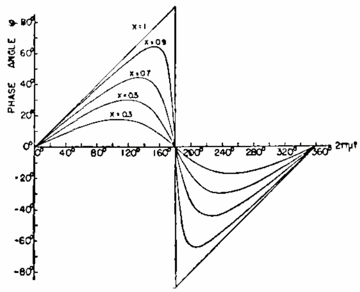

Fig. 4—Phase-angle variations.

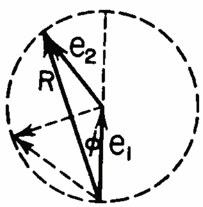

Fig. 6—Variations of φ.

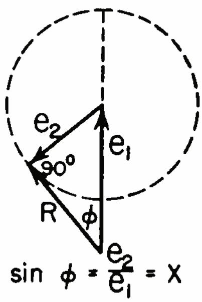

Fig. 5—Maximum value of φ.

by Figure 5. As x approaches one, the angle φ varies more and more rapidly near $2\pi\mu t = 180$ degrees. When $e_2 = e_1$ or $x = 1$, φ increases linearly from zero to 90 degrees as e_2 turns through 180 degrees.

As shown by Figure 6, φ is then an inscribed angle, and since an inscribed angle is measured by one-half its intercepted arc, φ increases linearly when e_2 turns uniformly. As e_2 approaches cancellation of e_1, R is an infinitesimal vector and $\varphi \rightarrow +90$ degrees. As e_2 swings past cancellation, the direction of R suddenly reverses so $\varphi = -90$ degrees; i. e., there is an instantaneous change of φ equal to 180 degrees. Beyond that point φ increases linearly toward 0 degrees, as shown by Figure 4.

Instantaneous Frequency

The output from a linear discriminator is proportional to the instantaneous frequency, where the instantaneous frequency is defined by:[16]

$$f = \frac{1}{2\pi}\frac{d}{dt} \text{ (argument of sine function).} \tag{6}$$

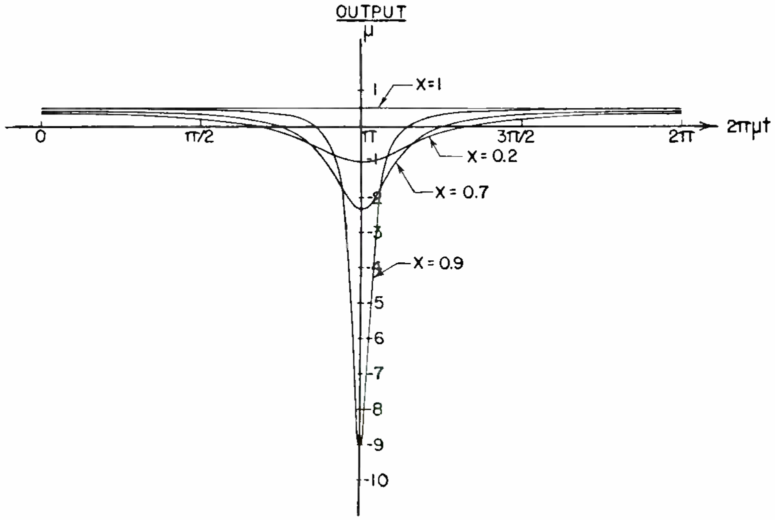

Fig. 7—Audio output, $x < 1$.

For a balanced linear discriminator, tuned to frequency ω, the output is proportional to the deviation in frequency from the center frequency ω. As shown in Appendix I, the output is given by

$$\text{Output} \propto \frac{\mu}{\dfrac{\cos 2\pi\mu t + 1/x}{\cos 2\pi\mu t + x} + 1} \tag{7}$$

Obviously this output is proportional to the slope of the curves of Figure 4, since it represents the first derivative with respect to time.

The curves of Figure 7 show the wave form in the audio output from

[16] J. R. Carson, "Notes on the Theory of Modulation," *Proc. I.R.E.*, Vol. 10, No. 2, p. 57; February, 1922.

a frequency modulation receiver, with perfect limiting and linear phase shift in the tuned circuits. As x approaches one, the output becomes more and more like an impulse, until at $x = 1$, the output has the constant value one-half except when $2\pi\mu t = \pi$; here the output becomes infinite. The area between the line one-half unit above the time axis and the curve for the instantaneous frequency over one cycle is constant for all values of x and equals $-\pi\mu$. This means that as $x \rightarrow 1$ the output is constant except at $2\pi\mu t = \pi$ and at that point is an impulse equal to $\pi\mu$ times a unit-impulse function.

When x becomes greater than one, the polarity of the impulse changes, but the shape is the same, as shown by Figure 8.

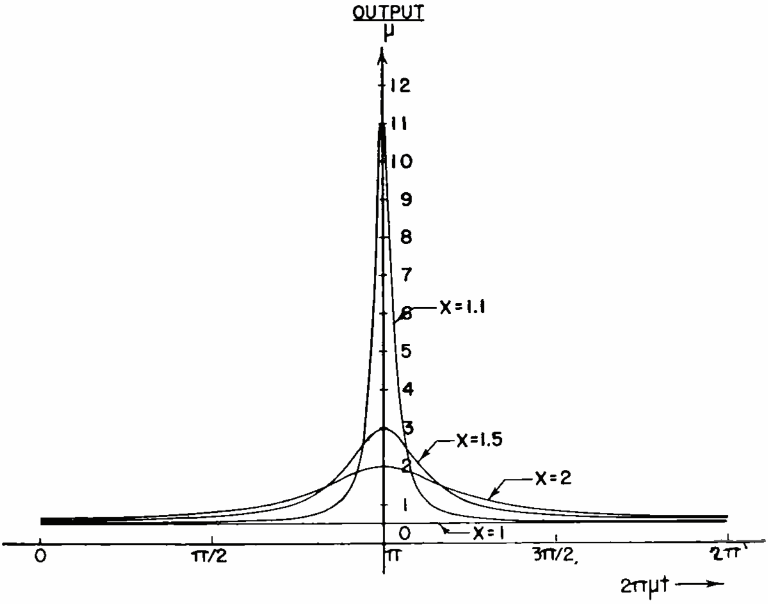

Fig. 8—Audio output, $x > 1$.

Average Value of Instantaneous Frequency. If the discriminator is tuned to the frequency ω, the average audio output is zero when $x < 1$. As shown in Appendix I, the average output is proportional to μ when $x > 1$. The curves of Figure 8 show this shift in average value when e_2 becomes stronger than e_1 and takes control.

Root-Mean-Square Value of Instantaneous Frequency. If the audio output from the discriminator is measured with an root-mean-square meter, the readings will vary as shown by Figure 9. The output increases uniformly from zero when $x = 0$ until it rapidly approaches infinity when $x = 1$. When $x > 1$ the output decreases uniformly to one as x becomes large.

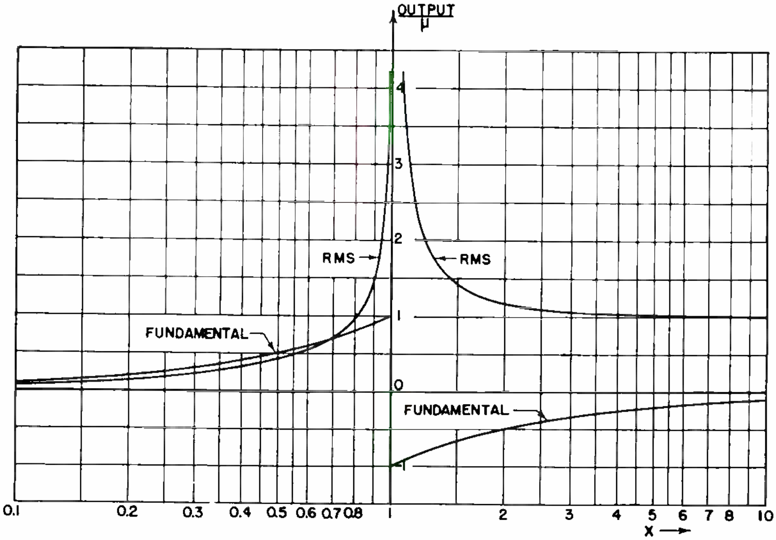

Fig 9—Root-mean-square voltage output.

As shown in Appendix I:

$$\text{Root-mean-square output} \propto \frac{x\mu}{\sqrt{2(1-x^2)}} \qquad \text{when } x<1 \tag{8}$$

$$= \mu \sqrt{\frac{2x^2-1}{2(x^2-1)}} \qquad \text{when } x>1. \tag{9}$$

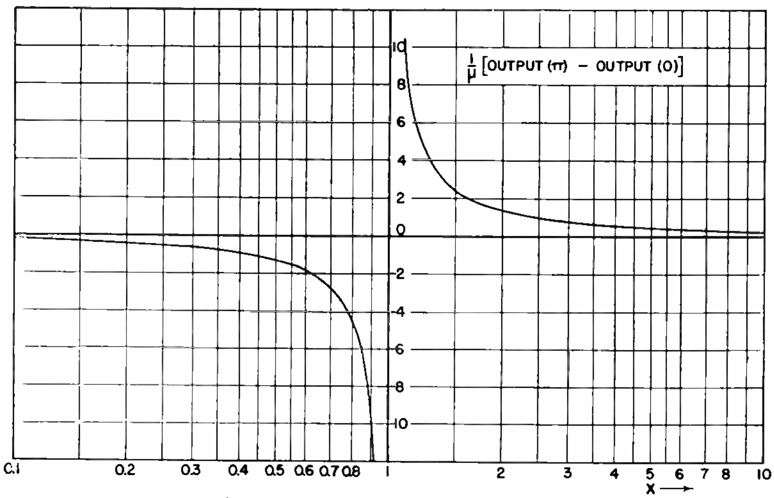

Fig. 10—Peak-to-peak audio output.

Peak-to-Peak Value of Instantaneous Frequency. The output when $2\pi\mu t = \pi$ minus the output at $2\pi\mu t = 0$ gives the peak-to-peak value of the instantaneous frequency. This is given by:

$$\text{Output } (\pi) - \text{Output } (0) = \frac{2x\mu}{x^2 - 1} \tag{10}$$

The curves of Figure 10 show how the peak-to-peak output varies as x increases. When $x = 1$, the peak-to-peak output becomes infinite, and it decreases uniformly beyond this point.

Harmonic Analysis of Instantaneous Frequency. If the harmonic content of the audio output is calculated by means of a Fourier-series analysis, the result can be expressed as:

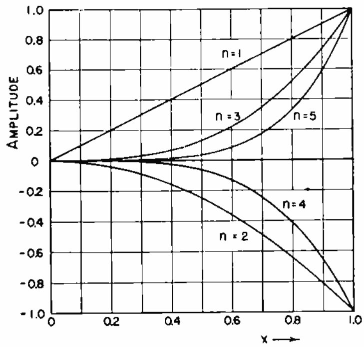

Fig. 11—Harmonic content of audio output.

$$\text{Output} \propto -\mu \sum_{n=1}^{\infty} (-x)^n \cos n(2\pi\mu t). \tag{11}$$

This means that the nth harmonic amplitude is proportional to μx^n. Figure 11 shows the increase of the harmonic amplitudes with increasing x for the first five harmonics. For small values of x, the higher harmonics are much smaller than the fundamental; but as x approaches one, the higher harmonics increase rapidly, until at $x = 1$ all harmonics are equal.

Effect of Limited Band Width. If the audio output from the discriminator is sent through a low-pass filter, having approximately linear phase-shift, the resultant wave form will depend upon how many harmonics are passed by the filter. In Figure 12, the case of $x = 0.9$ is shown for various low-pass filters. The case $n = 1$ means that only one harmonic, the fundamental, is passed by the filter. If two harmonics are passed, $n = 2$, the center begins to dip more because both harmonics are in phase at that point. The cases for $n = 3$ and $n = 5$ are also shown. The effect, there-

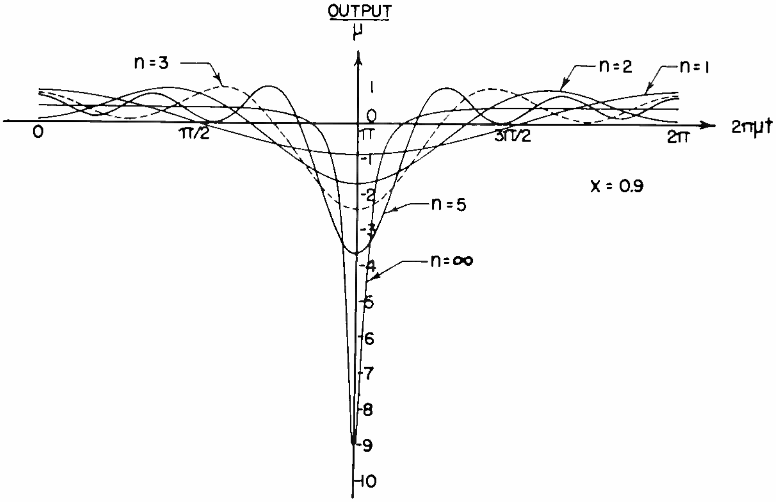

Fig. 12—Effect of low-pass filter.

fore, of limited band width is to reduce the output at $2\pi\mu t = \pi$ and to cause the resulting wave to oscillate about the curve that would be obtained with unlimited band width. For the case when $n = 5$, the peak output is reduced from 9.0 to 3.69, or the output becomes 41 per cent of that for unlimited band width. The curves of Figure 13 show the effect of limited band width. The variable on the axis of abscissas shows the number of harmonics passed by the low-pass filter, and the other axis shows the percent of peak amplitude compared to that for unlimited band width. Thus, if $x = 0.9$ and 10 harmonics are passed by the filter, the peak output

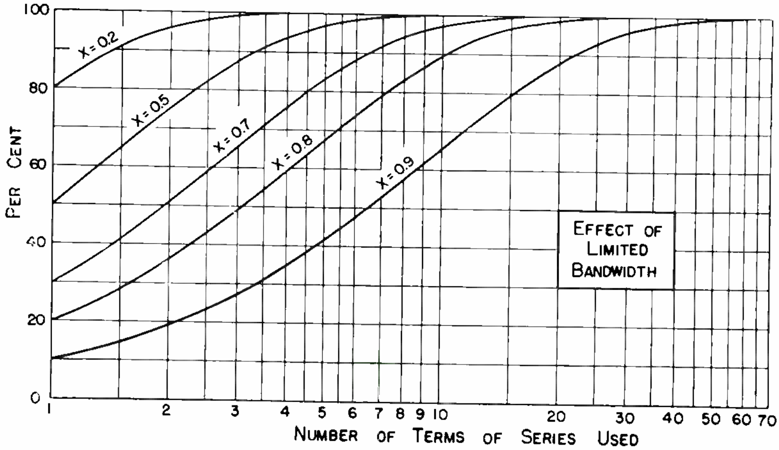

Fig. 13—Effect of limited band width.

will be approximately 65 per cent of what it would be if all harmonics were passed. If $x = 0.5$, it is evident that five or six harmonics will give nearly undistorted output.

As shown by Appendix I, this ratio of the peak output to the corresponding peak for unlimited band width is equal to $1 - x^n$ where n is the number of harmonics passed.

Common- and Adjacent-Channel Interference

The simplest case of frequency modulation interference (that of two

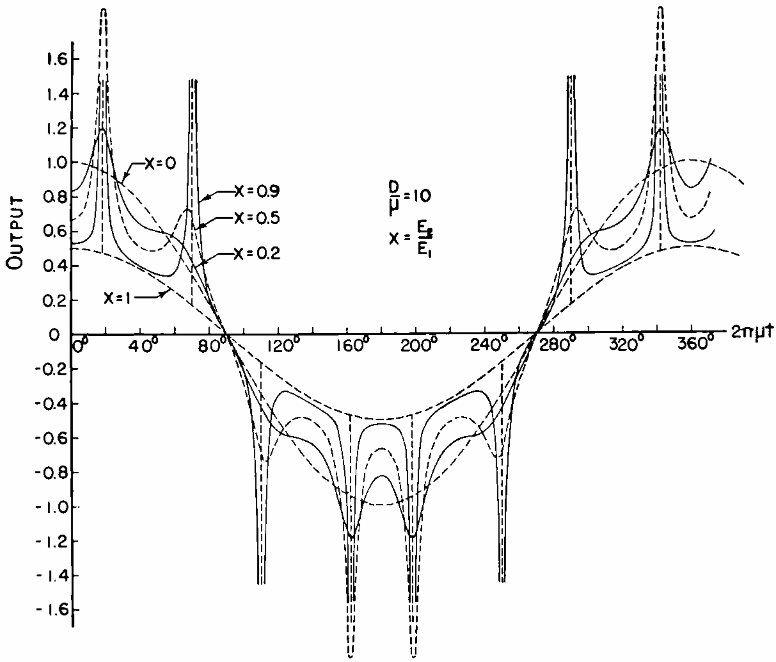

Fig. 14—Variation of distortion as interfering signal becomes stronger.

unmodulated carriers of slightly different frequency) has already been discussed. If now the amplitudes of the two waves are kept constant, but the frequency of one carrier is changed, the problem becomes one of common- or adjacent-channel interference depending upon what range of frequencies the swings of the modulated carrier cover. If the deviations of the one wave are about a mean frequency which coincides with the frequency of the second carrier, the result is common-channel interference. If the mean frequencies are separated by the width of one channel, the result will be adjacent-channel interference.

Common-Channel Interference, Interfering Signal Unmodulated. If a frequency-modulated signal and an unmodulated carrier produce the beat-note interference, the output from a frequency-modulation receiver with limiter will be as shown by Figure 14. This shows the wave form for the various ratios of the interfering signal voltage x. When $x = 0$ (i.e., no interference) the output is an undistorted cosine wave, as shown by the dotted line. As the interference increases, the peaks and dips increase in size, until finally, in the limit, they become very narrow pulses superimposed on a cosine wave of one-half the amplitude obtained with no interference.

As x becomes greater than one, the interfering signal takes control and

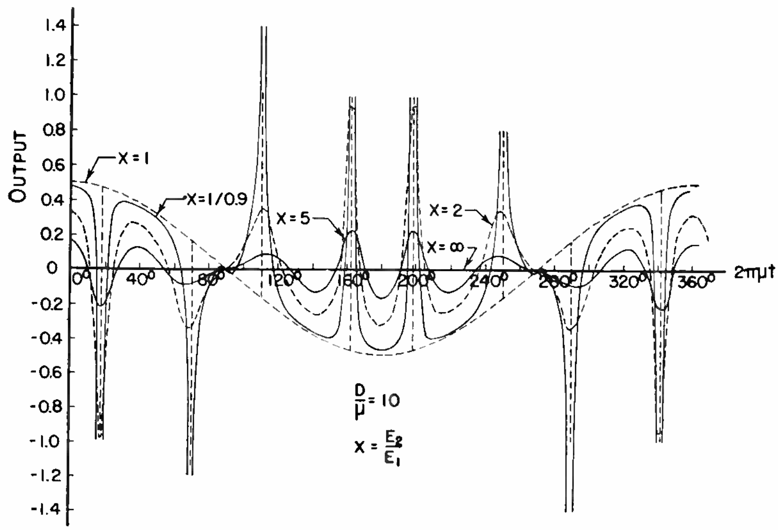

Fig. 15—Variation of distortion as interfering signal becomes stronger.

the modulation of the desired signal is suppressed. Figure 15 shows how the peaks and dips in output decrease when x increases from one to infinity. The envelope of the carrier amplitude corresponding to Figures 14 and 15 is shown by Figure 16. There is one cancellation or hole in the carrier amplitude corresponding to each peak or dip in the output, since the rapid phase change which occurs at cancellation produces the large frequency deviation. If the limiter is not able to maintain a constant voltage input to the discriminator, the amplitude variations of the carrier will cause a reduction in the peaks in the output.

Envelope of Beat-note Pattern. As shown in Appendix II, the beat-note produced in the output of a receiver with a perfect limiter during common-channel interference is a series of peaks and dips which are limited by the

Figure 19 shows how the beat-note then becomes unsymmetrical. At one end of the swing the two signals have nearly the same frequency and the beats come slowly. At the other end of the swing there is a considerable frequency difference and the beats are very much more rapid. The peaks and dips are limited by the two curves:

$$\text{Envelope} = \frac{D}{1+x}\cos 2\pi\mu t + \frac{\alpha}{2\pi}\,\frac{x}{x+1} \tag{12}$$

$$\text{and} \qquad \frac{D}{1-x}\cos 2\pi\mu t + \frac{\alpha}{2\pi}\,\frac{x}{x-1} \tag{13}$$

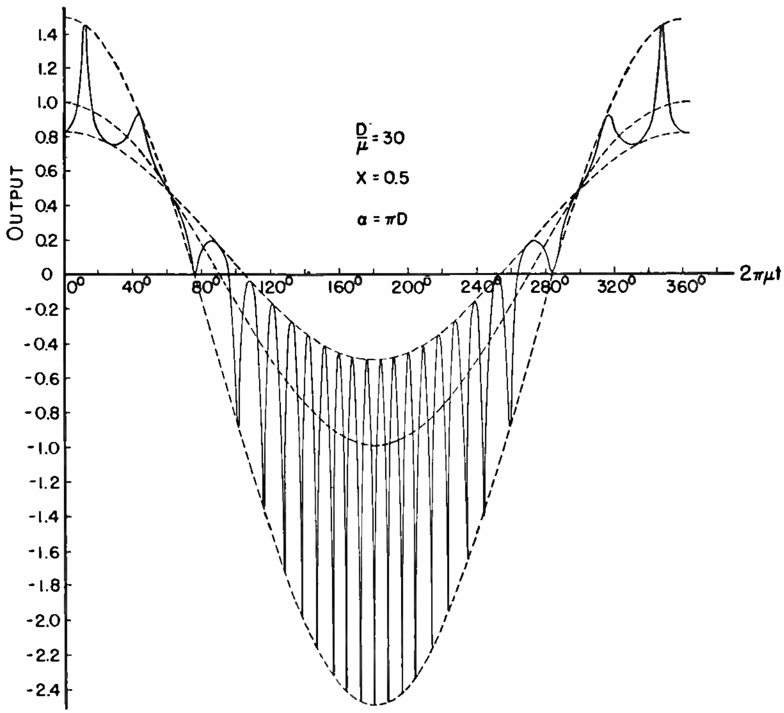

Fig. 19—Common-channel interference, interfering signal detuned.

Fourier-Series Analysis of Distorted Output. If the desired frequency-modulated signal is:

$$e_1 = E_1 \sin\left(\omega t + \frac{D}{\mu}\sin 2\pi\mu t\right) \tag{14}$$

and the interference is an unmodulated *r-f* carrier of angular frequency

$\omega + \alpha$, and phase angle θ, or:

$$e_2 = E_2 \sin\{(\omega + \alpha)t + \theta\} \tag{15}$$

then, as shown in Appendix II, the envelope of the resultant carrier is given by:

$$\text{Envelope} = E_1 \sqrt{1 + x^2 + 2x \cos\beta} \tag{16}$$

where: $\beta = \dfrac{D}{\mu} \sin 2\pi\mu t - \alpha t - \theta$

and the audio output is given by:

$$\text{Output} = D \cos 2\pi\mu t - \frac{D \cos 2\pi\mu t - \alpha/2\pi}{\dfrac{\cos\beta + 1/x}{\cos\beta + x} + 1} \tag{17}$$

When this is expanded in a Fourier series to determine the harmonic and cross-modulation distortion, the audio output is given by:

$$\text{Output} = D \cos 2\pi\mu t$$

$$+ \sum_{n=1}^{\infty} \sum_{r=-\infty}^{\infty} (-x)^n \left\{ \frac{\mu r}{n} - \frac{\alpha}{2\pi} \right\} J_r(nD/\mu) \cos(r\varepsilon - n\alpha t - n\theta) \tag{18}$$

where $\varepsilon = 2\pi\mu t$, and $x<1$.

This shows that the effect of the interfering signal is to produce cross modulation between the desired signal modulated with audio frequency μ and the interfering unmodulated carrier of angular frequency $\omega + \alpha$. The amplitude of each cross-modulation frequency can be computed with the help of a table of Bessel functions of the first kind.

When $\alpha = 0$, (i.e., common-channel interference) the output becomes:

$$\text{Output} = D \cos 2\pi\mu t$$

$$+ 2\mu \sum_{r=1}^{\infty} (2r-1)\, C(2r-1, D/\mu; x, \theta) \cos\{(2r-1)(2\pi\mu t)\}$$

$$+ 2\mu \sum_{r=1}^{\infty} (2r)\, S(2r, D/\mu; x, \theta) \sin\{(2r)(2\pi\mu t)\} \tag{19}$$

where the C- and S-functions are defined as follows:

$$C(m, n; x, \theta) = \sum_{s=1}^{\infty} \frac{(-x)^s}{s} J_m(sn) \cos s\theta \tag{20}$$

$$S(m, n; x, \theta) = \sum_{s=1}^{\infty} \frac{(-x)^s}{s} J_m(sn) \sin s\theta \tag{21}$$

$$x^2 \leqq 1$$

To find the amplitudes of the various harmonics produced during common-channel interference, compute the value of the desired $C-$ or $S-$ function from equations 20 and 21, and multiply by the proper factor, which is shown by the above equation 19 for the audio output. A special table of Bessel functions has been prepared for this purpose.[17]

The effect of a de-emphasis network following the discriminator, and of a low-pass audio filter, can be determined by computing the amplitude of each harmonic that falls within the working range, correcting each one for amplitude and phase changes in the audio amplifier and filters, and then recombining them by superposition.

If the signal-noise ratio is defined as the desired audio output with no interfering carrier present, divided by the peak noise (i.e., the maximum departure from the desired audio output when no interference is present), then, as shown by Figures 17 and 18, the signal-noise ratio is independent of the modulation index, but depends only on the ratio of the two voltages, x. This assumes a perfect limiter, adequate band width in the amplifiers and discriminator, and linear-phase-shift circuits.

If a de-emphasis network and a low-pass audio filter are used, many of the harmonics will be attentuated or removed, and the nonlinear phase shift will prevent the remaining harmonics from coming into phase all at the same time. The peaks of noise are therefore reduced considerably. When the modulation index, D/μ, is large, the noise beat-note peaks come very rapidly. This means that the harmonics will be of high order and they will be reduced or removed by the audio selectivity. This accounts for the observed noise reduction with wide-band frequency modulation and shows that it is very important to use a de-emphasis network and low-pass filter.

Common-Channel Interference, Both Signals Modulated. The preceding cases have described the interference produced by an unmodulated carrier on the same channel as the desired signal, and the effect of detuning the interfering carrier. This section is a discussion of the case when both the desired and undesired signals are modulated sinusoidally, and of the resultant distortion, which is even more complicated.

In order to illustrate this form of interference, assume the following conditions:

$D_1/\mu_1 = 10$, $D_2/\mu_2 = 5$, $D_1 = 4D_2$, $\mu_1 = 2\mu_2$, $x = E_2/E_1$
For example, $D_1 = 60$ kc, $\mu_1 = 6$ kc, $D_2 = 15$ kc, $\mu_2 = 3$ kc, $x = 0.5$ and 0.9 could be one set of numerical values.

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[17] Murlan S. Corrington and William Miehle, "Tables of Bessel Functions $J_n(x)$ for Large Arguments," *Jour. Math. Phys.*, Vol. 24, No. 1, pp. 30–50; Feb., 1945.

The beat-note envelope produced in this case is shown by Figure 20. The characteristic peaks and holes in the resultant carrier amplitude are present, but some of them are modified in shape because the two audio frequencies are present simultaneously.

Near 40 degrees and again near 130 degrees the two voltages start to go out-of-phase, but the two vectors then begin to reverse themselves and only a small decrease in amplitude occurs.

If this signal is sent through a receiver with a perfect limiter and linear discriminator, the resultant audio output will be as shown by Figure 21. Two cycles of the desired signal are shown as a dotted curve. This corresponds to one cycle of the undesired signal. As x increases toward one, the beat-note interference increases in amplitude until in the limit as $x \rightarrow 1$, the pulses become very narrow and long. If x becomes greater than one,

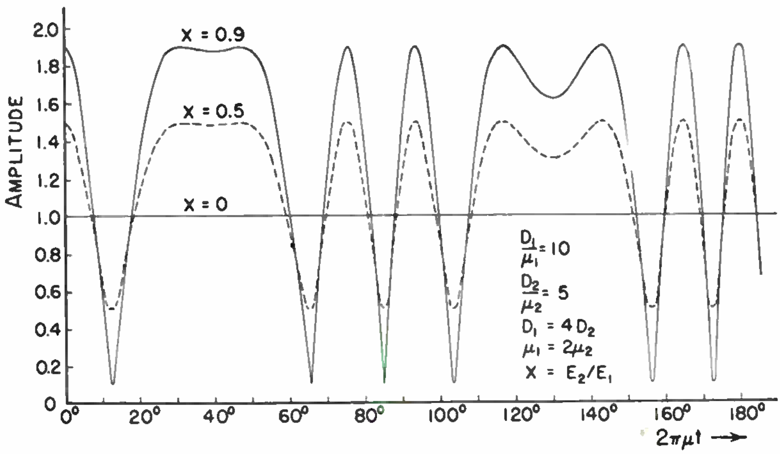

Fig. 20—Heterodyne envelope.

the polarity of the pulses is reversed (as shown by Figure 22) and this undesired signal gains control. When x becomes very large, only the undesired signal is received, as shown by the dotted cosine wave of unit amplitude.

The equations for the envelope and the beat-note interference are derived in Appendix III. If D_1 and D_2 are the two deviations and μ_1 and μ_2 are the corresponding audio frequencies, the envelope of the carrier is:

$$\text{Envelope} = E_1 \sqrt{1 + x^2 + 2x \cos \{D_1/\mu_1 \sin 2\pi\mu_1 t - D_2/\mu_2 \sin 2\pi\mu_2 t\}} \qquad (22)$$

and the audio output from a receiver with limiter and balanced discriminator is:

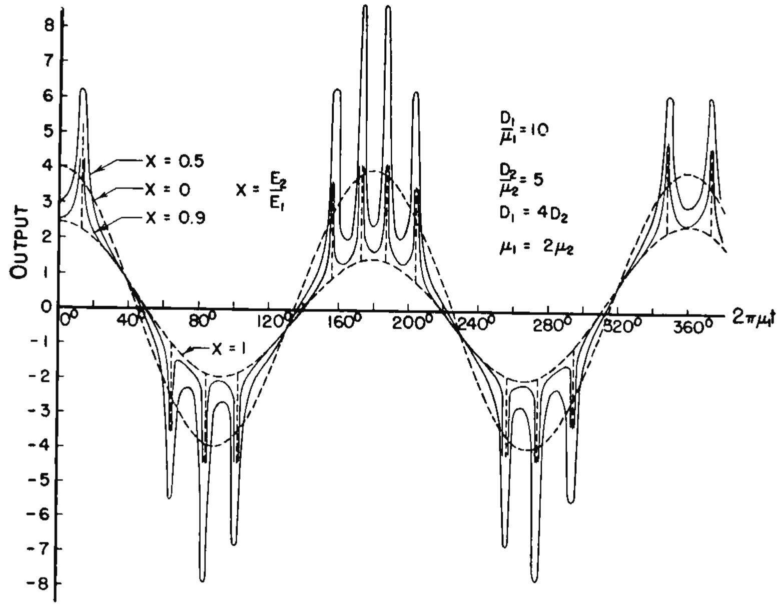

Fig. 21—Beat-note interference, $x < 1$.

$$\text{Output} \propto D_1 \cos 2\pi\mu_1 t - \frac{D_1 \cos 2\pi\mu_1 t - D_2 \cos 2\pi\mu_2 t}{\dfrac{\cos\theta + 1/x}{\cos\theta + x} + 1} \qquad (23)$$

where
$$\theta = \frac{D_1}{\mu_1} \sin 2\pi\mu_1 t - \frac{D_2}{\mu_2} \sin 2\pi\mu_2 t. \qquad (24)$$

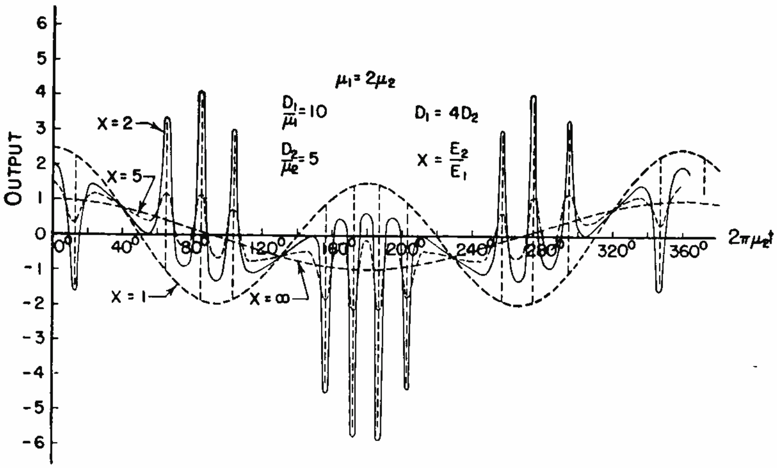

Fig. 22—Beat-note interference, $x > 1$.

The audio output is composed of a beat-note pattern which is limited by the two envelopes:

$$\text{Envelope} = \frac{D_1}{1+x}\cos 2\pi\mu_1 t + \frac{D_2 x}{1+x}\cos 2\pi\mu_2 t \tag{25}$$

and

$$\frac{D_1}{1-x}\cos 2\pi\mu_1 t + \frac{D_2 x}{x-1}\cos 2\pi\mu_2 t \tag{26}$$

This effect is shown by Figure 23 for the set of values given. In case of imperfect limiting, limited band width, or nonlinear phase shift in the amplifiers, these peaks will not be so long and narrow; the two envelopes shown represent the limits of the distortion.

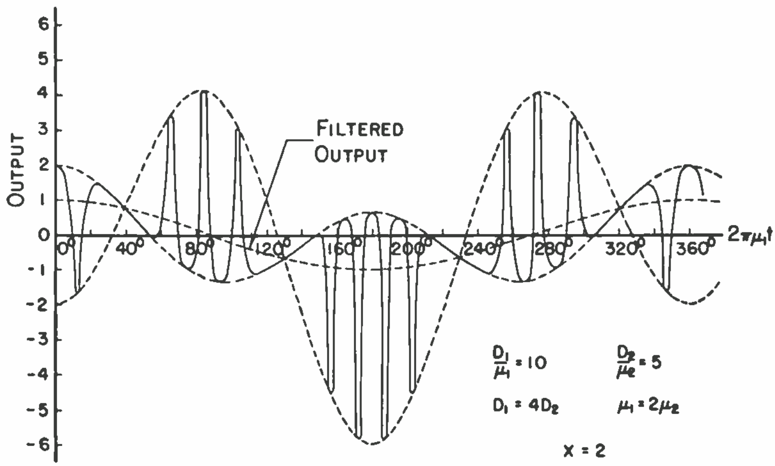

Fig. 23—Envelope of beat-note pattern.

The effect of low-pass filters or other audio selectivity can be determined from a study of the harmonic content of the distortion. As shown in Appendix III, the audio output can be expressed as a Fourier series which gives the cross modulation terms produced and their amplitudes.

Thus: Output $\propto D_1 \cos 2\pi\mu_1 t$

$$+ \sum_{r=-\infty}^{\infty} \sum_{s=-\infty}^{\infty} (r\mu_1 - s\mu_2)\, C(r, D_1/\mu_1; s, D_2/\mu_2 : x, 0) \cos (r\alpha - s\beta) \tag{27}$$

where $\alpha = 2\pi\mu_1 t$, $\beta = 2\pi\mu_2 t$ and the generalized C-function is defined as:

$$C(k, l; m, n : x, \theta) = \sum_{s=1}^{\infty} \frac{(-x)^s}{s} J_k(sl)\, J_m(sn) \cos s\theta. \tag{28}$$

The amplitude of any desired combination tone can be determined by choosing the appropriate values of r and s and by computing the desired C-function. Since the C-function cannot exceed unity for a given combination tone, it is evident that if $D_1 \gg \mu_1$ or μ_2 the distortion will be reduced with increasing modulation index.

Conclusions

Frequency-modulated radio broadcasting offers the advantage of improved noise reduction when compared with the usual amplitude-modulation systems. There is less interference between stations operating on the same frequency than for the corresponding case of amplitude modulation, and less power is required to cover a given area.

A difficulty arose occasionally in the 42–50 megacycle frequency modulation band because long-distance transmission could be observed from frequency modulation broadcast stations during hot weather or before a storm. It sometimes happened that such an interfering station became stronger than a desired station in the same channel for short intervals. When this happened there was a small amount of noise and the programs suddenly were interchanged. This change often lasted for several seconds but sometimes was limited to a word or two or a few notes of music. If the proposed new frequency modulation stations are all completed, this interference may occur again. When the interfering station has nearly the same carrier frequency as the desired station this effect is called common-channel interference. If the two carrier frequencies are separated by the width of one channel the result is called adjacent-channel interference.

The simplest case of frequency modulation interference occurs when two modulated carriers, having nearly the same frequency, beat together to produce a resultant signal. As the two voltages alternately reinforce and cancel each other, the result is a heterodyne envelope consisting of a series of broad peaks and sharp dips. Each time the two interfering voltages cancel each other to produce a hole in the envelope, there is a rapid phase shift of the resultant voltage. Since the audio output from a frequency-modulation receiver is proportional to the rate of change of the phase of this resultant, the rapid phase shift produces a distorted audio output, which becomes more and more like an impulse as the interfering carrier voltage becomes nearly equal to the desired carrier voltage.

When the two amplitudes of the interfering voltages are kept constant but the frequency of one is changed, the result is common- or adjacent-channel interference depending upon what range of frequencies the swings of the modulated carrier cover. The beat-note produced by this interference consists of a series of sharp peaks and dips of noise and is superimposed on the desired audio output. When the modulation index is increased, these peaks occur more and more rapidly, and the harmonics produced are redistributed to higher and higher orders. If the receiver has sufficient band width, a perfect limiter, and a wide-band audio system, the signal-noise ratio does not depend on the modulation index, but is determined solely by the ratio of the desired signal voltage to the undesired signal voltage.

Formulas are given for computing the amplitudes of the harmonics and cross-modulation frequencies produced by the interference. The effect of a de-emphasis network following the discriminator, of a low-pass audio filter, and of nonlinear phase-shift can be determined by computing the amplitude of each harmonic that falls within the working range. Each such harmonic is then corrected for amplitude and phase changes in the audio amplifier and filters, and they are then recombined by superposition to obtain the filtered audio output.

When the modulation index is large, the beat-notes of the noise come very rapidly, and since this means that the harmonics are then of high order, most of the distortion will be removed by the audio selectivity. This accounts for the observed noise reduction with wide-band frequency modulation and shows that it is very important that a de-emphasis network and low-pass filter be used to obtain maximum performance. In order to obtain the maximum signal-noise ratio, it is necessary to use some means for removing the variations in the amplitude of the resultant signal so that the discriminator responds to the variations in the instantaneous frequency, but is not affected by amplitude variations.

* * *

APPENDIX I.

Analysis of Fundamental Case

Let the two interfering signals be $e_1 \sin \omega t$ and $e_2 \sin (\omega + 2\pi\mu)t$ The resultant voltage is then:

$$e_1 \sin \omega t + e_2 \sin (\omega + 2\pi\mu)t$$

$$= e_1\sqrt{1 + x^2 + 2x \cos 2\pi\mu t} \sin (\omega t + \varphi) \qquad (29)$$

where $e_2/e_1 = x$ and $\tan \varphi = \dfrac{x \sin 2\pi\mu t}{1 + x \cos 2\pi\mu t}$.

The instantaneous frequency becomes:

$$f = \frac{1}{2\pi}\frac{d}{dt}(\omega t + \varphi) = \frac{\omega}{2\pi} + \frac{1}{2\pi}\frac{d}{dt}\tan^{-1}\frac{x \sin 2\pi\mu t}{1 + x\cos 2\pi\mu t}$$

$$= \frac{\omega}{2\pi} + \mu\frac{x\cos 2\pi\mu t + x^2}{1 + x^2 + 2x\cos 2\pi\mu t}$$

$$= \frac{\omega}{2\pi} + \frac{\mu}{\dfrac{\cos 2\pi\mu t + 1/x}{\cos 2\pi\mu t + x} + 1} \tag{30}$$

This is valid for all values of x.

For a balanced linear discriminator, the audio output is proportional to:

$$\text{Output} \propto \frac{\mu}{\dfrac{\cos 2\pi\mu t + 1/x}{\cos 2\pi\mu t + x} + 1} \tag{31}$$

When $x \ll 1$, this is, approximately,

$$\text{Output} \propto \mu x \cos 2\pi\mu t \tag{32}$$

The instantaneous frequency can be written:

$$f = \frac{\omega}{2\pi} + \mu - \mu\frac{\dfrac{1}{x}\cos 2\pi\mu t + \dfrac{1}{x^2}}{1 + \dfrac{1}{x^2} + \dfrac{2}{x}\cos 2\pi\mu t} \tag{33}$$

This means that as x goes from less than one to greater than one (i.e., if x is changed to $1/x$) there is an apparent change in frequency equal to μ and a reversal in polarity of the modulation. This means that e_2 becomes stronger than e_1 and takes control.

Average Voltage of Rectified Envelope

The average voltage of the carrier envelope is:

$$\text{Average voltage} = \frac{1}{\pi}\int_0^{\pi} e_1\sqrt{1 + x^2 + 2x\cos\theta}\,d\theta$$

$$= \frac{2(1+x)\,e_1}{\pi}\int_0^{\pi/2}\sqrt{1 - \frac{4x}{(1+x)^2}\sin^2\alpha}\,d\alpha$$

$$= \frac{2(1+x)e_1}{\pi}E\left\{\frac{2\sqrt{x}}{1+x}\right\} \tag{34}$$

where $E\left\{\frac{2\sqrt{x}}{1+x}\right\}$ is a complete elliptic integral of the second kind with modulus $\frac{2\sqrt{x}}{1+x}$.

Root-Mean-Square Voltage of Rectified Envelope

The Root-Mean-Square voltage of the rectified carrier envelope is:

$$\text{Root-Mean-Square voltage} = e_1\sqrt{\frac{1}{\pi}\int_0^\pi (1 + x^2 + 2x\cos\theta)\, d\theta}$$

$$= e_1\sqrt{\frac{1}{\pi}\Big[(1 + x^2)\theta + 2x\sin\theta\Big]_0^\pi}$$

$$= e_1\sqrt{1 + x^2} \tag{35}$$

Fourier-Series Analysis of Envelope

The envelope of the carrier is given by:

$$\text{Envelope} = e_1\sqrt{1 + x^2 + 2x\cos 2\pi\mu t} \qquad \text{where } x \leqq 1. \tag{36}$$

Consider the expression:

$$\sqrt{1 + x^2 + 2x\cos\beta} = (1 + xe^{i\beta})^{\frac{1}{2}}(1 + xe^{-i\beta})^{\frac{1}{2}}$$

$$= \left\{1 + \frac{1}{2}xe^{i\beta} - \frac{1(1)}{2(4)}x^2e^{2i\beta} + \frac{1(1)(3)}{2(4)(6)}x^3e^{3i\beta} - \frac{1(1)(3)(5)}{2(4)(6)(8)}x^4e^{4i\beta} + \dots\right\}$$

$$\times\left\{1 + \frac{1}{2}xe^{-i\beta} - \frac{1(1)}{2(4)}x^2e^{-2i\beta} + \frac{1(1)(3)}{2(4)(6)}x^3e^{-3i\beta} - \frac{1(1)(3)(5)}{2(4)(6)(8)}x^4e^{-4i\beta} + \dots\right\} \tag{37}$$

by the usual binomial series expansion.[18]

Multiply these factors together, term by term, then:

$$\sqrt{1 + x^2 + 2x\cos\beta} = a_0 + a_1\cos\beta + a_2\cos 2\beta + \dots \tag{38}$$

where:

$$a_0 = 1 + \frac{x^2}{4} + \frac{x^4}{64} + \frac{x^6}{256} + \frac{25x^8}{16384} + \dots \tag{39}$$

$$a_1 = x\left\{1 - \frac{x^2}{8} - \frac{x^4}{64} - \frac{5x^6}{1024} - \frac{35x^8}{16384} - \dots\right\} \tag{40}$$

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[18] Edwin P. Adams, SMITHSONIAN MATHEMATICAL FORMULAE AND TABLES OF ELLIPTIC FUNCTIONS, Smithsonian Institution, Washington, D. C., 1939, p. 117.

$$a_2 = -\frac{x^2}{4}\left\{1 - \frac{x^2}{4} - \frac{5x^4}{128} - \frac{7x^6}{512} - \frac{105x^8}{16384} - \ldots\right\} \tag{41}$$

$$a_3 = \frac{x^3}{8}\left\{1 - \frac{5x^2}{16} - \frac{7x^4}{128} - \frac{21x^6}{1024} - \frac{165x^8}{16384} - \ldots\right\} \tag{42}$$

$$a_4 = -\frac{5x^4}{64}\left\{1 - \frac{7x^2}{20} - \frac{21x^4}{320} - \frac{33x^6}{1280} - \frac{429x^8}{32768} - \ldots\right\} \tag{43}$$

$$a_n = 2(-1)^n\left\{\frac{1(3)\ldots(2n-1)}{n!}\right\}\frac{x^n}{2^n}\left\{\frac{-1}{2n-1} + \frac{1}{1(n+1)}\frac{x^2}{2^2}\right.$$

$$\left.+\sum_{k=2}^{\infty}\frac{1(3)\ldots(2k-3)}{k!\,2^{2k}}\frac{(2n+1)(2n+3)\ldots(2n+2k-3)}{(n+1)(n+2)\ldots(n+k)}x^{2k}\right\} \tag{44}$$

This expression for a_n was previously obtained by Vigoureux[19] and Moullin[20]

In the limit as $x \to 1$:

$$\text{Envelope} = \sqrt{2}\,e_1\sqrt{1+\cos 2\pi\mu t} = 2e_1\left|\cos \pi\mu t\right|$$

$$= \frac{4e_1}{\pi}\left\{1 + \frac{2}{3}\cos\theta - \frac{2}{15}\cos 2\theta\right.$$

$$\left. + \frac{2}{35}\cos 3\theta - \frac{2}{63}\cos 4\theta - \ldots\right\}$$

$$= \frac{4e_1}{\pi}\left\{1 - 2\sum_{n=1}^{\infty}\frac{(-1)^n\cos n\theta}{(2n)^2-1}\right\}$$

where $\theta = 2\pi\mu t$. (45)

Calculation of Average Value of Instantaneous Frequency

Consider the integral:

$$I = \frac{\mu}{\pi}\int_0^{\pi}\frac{x^2 + x\cos\varepsilon}{1+x^2+2x\cos\varepsilon}\,d\varepsilon \tag{46}$$

Make the transformation: $\cos\varepsilon = \dfrac{1-t^2}{1+t^2}$, $d\varepsilon = \dfrac{2dt}{1+t^2}$

[19] F. M. Colebrook, "A Note on the Frequency Analysis of the Heterodyne Envelope. Its Relation to Problems of Interference." *Wireless Engineer & Experimental Wireless*, Vol. 9, p. 200, April, 1932.

[20] E. B. Moullin, "The Detection by a Straight Line Rectifier of Modulated and Heterodyne Signals," *Wireless Engineer & Experimental Wireless*, Vol. 9, pp. 378–383; July, 1932.

Then.

$$I = \frac{2x\mu}{\pi}\int_0^\infty \frac{(1+x)-(1-x)t^2}{(1+x)^2+(1-x)^2t^2}\,\frac{dt}{1+t^2}$$

$$= \frac{(x^2-1)\mu}{\pi}\int_0^\infty \frac{dt}{(1+x)^2+(1-x)^2t^2} + \frac{\mu}{\pi}\int_0^\infty \frac{dt}{1+t^2}$$

$$= \frac{(x^2-1)\mu}{\pi}\left\{\frac{\pi}{2(1-x^2)}\right\} + \frac{\mu}{\pi}\left(\frac{\pi}{2}\right) = 0 \text{ when } x<1$$

$$= \mu \text{ when } x > 1. \qquad (47)$$

The average value of the instantaneous frequency therefore equals zero when $x<1$ and is proportional to μ when $x>1$.

Calculation of the Root-Mean-Square Value of Instantaneous Frequency

Consider the integral:

$$I = \frac{1}{\pi}\int_0^\pi \left\{\frac{x^2 + x\cos\varepsilon}{1+x^2+2x\cos\varepsilon}\right\}^2 d\varepsilon \qquad (48)$$

Make the transformation: $\cos\varepsilon = \dfrac{1-t^2}{1+t^2}$, $d\varepsilon = \dfrac{2\,dt}{1+t^2}$.

Then:

$$I = \frac{2x^2}{\pi}\int_0^\infty \left\{\frac{(1+x)-(1-x)t^2}{(1+x)^2+(1-x)^2t^2}\right\}^2 \frac{dt}{1+t^2}$$

$$= \frac{-2x(1+x)^2}{\pi}\int_0^\infty \frac{dt}{\{(1+x)^2+(1-x)^2t^2\}^2}$$

$$+ \frac{(3x-1)(1+x)}{2\pi}\int_0^\infty \frac{dt}{\{(1+x)^2+(1-x)^2t^2\}} + \frac{1}{2\pi}\int_0^\infty \frac{dt}{1+t^2} \qquad (49)$$

Consider the integral:

$$\int_0^\infty \frac{dt}{\{a^2+b^2t^2\}^2} = \frac{1}{2a^2}\left[\frac{t}{a^2+b^2t^2}\right]_0^\infty + \frac{1}{2a^2}\int_0^\infty \frac{dt}{a^2+b^2t^2}$$

$$= \frac{1}{2a^2}\int_0^\infty \frac{dt}{a^2+b^2t^2} = \frac{1}{2a^3b}\left[\tan^{-1}\frac{b}{a}t\right]_0^\infty = \frac{\pi}{4a^3b} \text{ when } x<1 \qquad (50)$$

$$= -\frac{\pi}{4a^3b} \text{ when } x > 1. \qquad (51)$$

Therefore I becomes:

$$I = \frac{-2x(1+x)^2}{\pi}\left\{\frac{\pi}{4(1+x)^3(1-x)}\right\}$$

$$+\frac{(3x-1)(1+x)}{2\pi}\left\{\frac{\pi}{2(1+x)(1-x)}\right\}+\frac{1}{2\pi}\left\{\frac{\pi}{2}\right\}$$

$$=\frac{-x}{2(1-x^2)}+\frac{3x-1}{4(1-x)}+\frac{1}{4}=\frac{x^2}{2(1-x^2)} \text{ when } x<1, \tag{52}$$

$$=\frac{1-2x^2}{2(1-x^2)} \text{ when } x>1. \tag{53}$$

The root-mean-square voltage is proportional to:

$$\mu\sqrt{I}=\frac{x\mu}{\sqrt{2(1-x^2)}} \text{ when } x<1 \tag{54}$$

$$=\sqrt{\frac{2x^2-1}{2(x^2-1)}}\,\mu \text{ when } x>1. \tag{55}$$

Calculation of the Area of One Cycle of the Instantaneous Frequency

The area bounded by one cycle of the variation of the instantaneous frequency and a line one-half unit above the time axis, as shown by Figure 7, will now be computed. From equation 2 the instantaneous frequency is given by:

$$f=\frac{\omega}{2\pi}+\frac{1}{2\pi}\frac{d}{dt}\tan^{-1}\frac{x\sin 2\pi\mu t}{1+x\cos 2\pi\mu t} \tag{56}$$

$$\text{Area}=2\int_0^{\pi}\left\{\frac{1}{2\pi}\frac{d}{dt}\tan^{-1}\frac{x\sin 2\pi\mu t}{1+x\cos 2\pi\mu t}-\frac{1}{2}\right\}d(2\pi\mu t)$$

$$=2\mu\int_0^{\pi}\left\{\frac{d}{d\theta}\tan^{-1}\frac{x\sin\theta}{1+x\cos\theta}-\frac{1}{2}\right\}d\theta$$

$$=2\mu\left[\tan^{-1}\frac{x\sin\theta}{1+x\cos\theta}-\frac{\theta}{2}\right]_0^{\pi}=-\pi\mu \tag{57}$$

for all values of x.

Fourier-Series Analysis of Instantaneous Frequency

The audio output is proportional to:

$$\mu\frac{x\cos 2\pi\mu t+x^2}{1+x^2+2x\cos 2\pi\mu t}=\frac{1}{2\pi}\frac{d}{dt}\tan^{-1}\frac{x\sin 2\pi\mu t}{1+x\cos 2\pi\mu t} \tag{58}$$

Let $2\pi\mu t = \beta$ and $\tan \alpha = \dfrac{x \sin \beta}{1 + x \cos \beta}$

Then $k \sin \alpha = x \sin \beta$ (59)

$k \cos \alpha = 1 + x \cos \beta$ (60)

where: $k = \sqrt{1 + x^2 + 2x \cos \beta}$ (61)

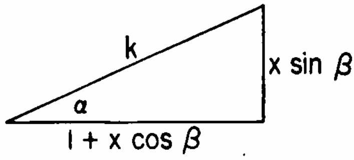

Fig. 24—Determination of k.

Multiply equation 59 by i and add equation 60.

Then: $1 + x \cos \beta + ix \sin \beta = k (\cos \alpha + i \sin \alpha)$ (62)

or: $1 + x\, e^{i\beta} = k\, e^{i\alpha}$. Take logarithms of both sides. Then:

$\log (1 + x\, e^{i\beta}) = \log k + i\alpha.$

Since:

$$\log (1 + x) = x - \frac{x^2}{2} + \frac{x^3}{3} - \frac{x^4}{4} + \ldots \qquad -1 < x \leqq 1$$

$$\log (1 + x\, e^{i\beta}) = xe^{i\beta} - \frac{x^2}{2} e^{2i\beta} + \frac{x^3}{3} e^{3i\beta} - \ldots$$

so:

$$\log k + i\alpha = x(\cos \beta + i \sin \beta) - \frac{x^2}{2} (\cos 2\beta + i \sin 2\beta)$$

$$+ \frac{x^3}{3} (\cos 3\beta + i \sin 3\beta) - \ldots \tag{63}$$

Equate imaginary terms:

$$\alpha = x \sin \beta - \frac{x^2}{2} \sin 2\beta + \frac{x^3}{3} \sin 3\beta - \ldots \tag{64}$$

Differentiate:

$$\frac{1}{2\pi} \frac{d\alpha}{dt} = \mu\, (x \cos \beta - x^2 \cos 2\beta + x^3 \cos 3\beta - \ldots) \tag{65}$$

The audio output is therefore proportional to:

$$\text{Output} \propto -\mu \sum_{n=1}^{\infty} (-x)^n \cos n\beta \qquad -1 < x \leq 1 \tag{66}$$

When $\beta = 0$,

$$\text{Output} \propto -\mu \sum_{n=1}^{\infty} (-x)^n = \frac{\mu x}{x+1} \tag{67}$$

When $\beta = \pi$, $$\text{Output} \propto -\mu \sum_{n=1}^{\infty} x^n = \frac{\mu x}{x-1} \tag{68}$$

Effect of Limited Band Width

To show the effect of a limited band width, consider the geometrical progression:

$$S = -\mu \sum_{n=1}^{n} x^n \tag{69}$$

By ordinary long division:

$$\frac{p^n - 1}{p - 1} = 1 + p + p^2 + \ldots + p^{n-1} \tag{70}$$

so,

$$S = -\mu x \frac{x^n - 1}{x - 1} \tag{71}$$

The ratio of the partial sum to the output at $\beta = \pi$ equals $1 - x^n$.

APPENDIX II.

Common- and Adjacent-Channel Interference

In order to show the effect of common- and adjacent-channel interference, let the desired frequency-modulated signal be:

$$e_1 = E_1 \sin (\omega t + \frac{D}{\mu} \sin 2\pi\mu t) \tag{72}$$

and let the interference be an unmodulated radio-frequency carrier at angular frequency $\omega + \alpha$, and phase angle θ, or

$$e_2 = E_2 \sin \{(\omega + \alpha)t + \theta\} \tag{73}$$

Then:

$$e_1 + e_2 = E_1\sqrt{1 + x^2 + 2x\cos\beta}\,\sin\left\{\omega t + \frac{D}{\mu}\sin 2\pi\mu t - \varphi\right\} \quad (74)$$

where $x = E_2/E_1$, $\beta = \frac{D}{\mu}\sin 2\pi\mu t - \alpha t - \theta$

and $\tan\varphi = \frac{x\sin\beta}{1 + x\cos\beta}$

The instantaneous frequency becomes:

$$\frac{\omega}{2\pi} + D\cos 2\pi\mu t - \frac{1}{2\pi}\frac{d}{dt}\tan^{-1}\frac{x\sin\beta}{1+x\cos\beta}$$

$$= \frac{\omega}{2\pi} + D\cos 2\pi\mu t - \frac{x\cos\beta + x^2}{1 + x^2 + 2x\cos\beta}\left\{D\cos 2\pi\mu t - \frac{\alpha}{2\pi}\right\}$$

$$= \frac{\omega}{2\pi} + D\cos 2\pi\mu t - \frac{D\cos 2\pi\mu t - \alpha/2\pi}{\dfrac{\cos\beta + 1/x}{\cos\beta + x} + 1} \quad (75)$$

Envelope of Beatnote Pattern

The beatnote produced in the output of a receiver with a perfect limiter is given by:

$$\text{Output} = D\cos 2\pi\mu t - \frac{D\cos 2\pi\mu t - \alpha/2\pi}{\dfrac{\cos\beta + 1/x}{\cos\beta + x} + 1} \quad (76)$$

where: $\beta = \frac{D}{\mu}\sin 2\pi\mu t - \alpha t - \theta$. The two envelopes of the maxima and minima of the beat-note pattern are obtained by setting $\beta = 2n\pi$ or $(2n + 1)\pi$ where n is an integer. This gives the two envelopes:

$$\text{Envelope} = \frac{D}{1 + x}\cos 2\pi\mu t + \frac{\alpha}{2\pi}\frac{x}{x + 1} \quad (77)$$

and:

$$\frac{D}{1 - x}\cos 2\pi\mu t + \frac{\alpha}{2\pi}\frac{x}{x - 1} \quad (78)$$

Fourier-Series Analysis of Instantaneous Frequency

In accordance with the analysis of Appendix I, equation 65:

$$\frac{d}{dt}\tan^{-1}\frac{x\sin\beta}{1+x\cos\beta}=-\sum_{n=1}^{\infty}(-x)^n\cos n\beta\,\frac{d\beta}{dt} \tag{79}$$

When: $\beta=\frac{D}{\mu}\sin 2\pi\mu t-\alpha t-\theta$ and $\varepsilon=2\pi\mu t$

the instantaneous frequency is:

$$f=\frac{\omega}{2\pi}+D\cos\varepsilon+\sum_{n=1}^{\infty}(-x)^n\cos n\left\{\frac{D}{\mu}\sin\varepsilon-\alpha t-\theta\right\}\left\{D\cos\varepsilon-\frac{\alpha}{2\pi}\right\} \tag{80}$$

Let: $\gamma=D/\mu$, then

$$f=\frac{\omega}{2\pi}+D\cos\varepsilon$$

$$+\frac{1}{4}\sum_{n=1}^{\infty}(-x)^n\left\{D(e^{i\varepsilon}+e^{-i\varepsilon})-\frac{\alpha}{\pi}\right\}\left\{e^{in\gamma\sin\varepsilon-in\alpha t-in\theta}+e^{-in\gamma\sin\varepsilon+in\alpha t+in\theta}\right\}$$

$$=\frac{\omega}{2\pi}+D\cos\varepsilon$$

$$+\frac{1}{4}\sum_{n=1}^{\infty}(-x)^n\Big\{D\,e^{i(\varepsilon-n\alpha t-n\theta)}\,e^{in\gamma\sin\varepsilon}+D\,e^{i(\varepsilon+n\alpha t+n\theta)}\,e^{-in\gamma\sin\varepsilon}$$

$$+D\,e^{-i(\varepsilon+n\alpha t+n\theta)}\,e^{in\gamma\sin\varepsilon}+D\,e^{-i(\varepsilon-n\alpha t-n\theta)}\,e^{-in\gamma\sin\varepsilon}$$

$$-\frac{\alpha}{\pi}e^{-i(n\alpha t+n\theta)}\,e^{in\gamma\sin\varepsilon}-\frac{\alpha}{\pi}e^{i(n\alpha t+n\theta)}\,e^{-in\gamma\sin\varepsilon}\Big\} \tag{81}$$

Using the identities:

$$e^{ix\sin\varepsilon}=\sum_{k=-\infty}^{\infty}J_k(x)\,e^{ik\varepsilon} \tag{82}$$

and $J_k(-x)=J_{-k}(x)$ where $J_k(x)$ is a Bessel function of the first kind of order k and argument x, the instantaneous frequency becomes:

$$f=\frac{\omega}{2\pi}+D\cos\varepsilon$$

$$+\frac{1}{4}\sum_{n=1}^{\infty}(-x)^n\Big\{D\,e^{i(\varepsilon-n\alpha t-n\theta)}\sum_{k=-\infty}^{\infty}J_k(n\gamma)\,e^{ik\varepsilon}$$

$$+ D\, e^{i(\varepsilon + n\alpha t + n\theta)} \sum_{k=-\infty}^{\infty} J_{-k}(n\gamma)\, e^{ik\varepsilon}$$

$$+ D\, e^{-i(\varepsilon + n\alpha t + n\theta)} \sum_{k=-\infty}^{\infty} J_{k}(n\gamma)\, e^{ik\varepsilon}$$

$$+ D\, e^{-i(\varepsilon - n\alpha t - n\theta)} \sum_{k=-\infty}^{\infty} J_{-k}(n\gamma)\, e^{ik\varepsilon}$$

$$- \frac{\alpha}{\pi} e^{-i(n\alpha t + n\theta)} \sum_{k=-\infty}^{\infty} J_{k}(n\gamma)\, e^{ik\varepsilon}$$

$$- \frac{\alpha}{\pi} e^{i(n\alpha t + n\theta)} \sum_{k=-\infty}^{\infty} J_{-k}(n\gamma)\, e^{ik\varepsilon} \Big\}$$

$$= \frac{\omega}{2\pi} + D \cos \varepsilon$$

$$+ \frac{1}{4} \sum_{n=1}^{\infty} (-x)^n \Big\{ D \sum_{k=-\infty}^{\infty} J_k(n\gamma)\, e^{i\{(k+1)\varepsilon - n\alpha t - n\theta\}}$$

$$+ D \sum_{k=-\infty}^{\infty} J_{-k}(n\gamma)\, e^{i\{(k+1)\varepsilon + n\alpha t + n\theta\}}$$

$$+ D \sum_{k=-\infty}^{\infty} J_{k}(n\gamma)\, e^{i\{(k-1)\varepsilon - n\alpha t - n\theta\}}$$

$$+ D \sum_{k=-\infty}^{\infty} J_{-k}(n\gamma)\, e^{i\{(k-1)\varepsilon + n\alpha t + n\theta\}}$$

$$- \frac{\alpha}{\pi} \sum_{k=-\infty}^{\infty} J_{k}(n\gamma)\, e^{i\{k\varepsilon - n\alpha t - n\theta\}}$$

$$- \frac{\alpha}{\pi} \sum_{k=-\infty}^{\infty} J_{-k}(n\gamma)\, e^{i\{k\varepsilon + n\alpha t + n\theta\}} \Big\} \qquad (83)$$

Make the substitutions:

$k + 1 = r$ in term 1	$k - 1 = -r$ in term 4
$k + 1 = -r$ in term 2	$k = r$ in term 5
$k - 1 = r$ in term 3	$k = -r$ in term 6

Then:

$$f = \frac{\omega}{2\pi} + D \cos \varepsilon$$

$$+\frac{1}{4}\sum_{n=1}^{\infty}(-x)^n\Big\{D\sum_{r=-\infty}^{\infty}\Big[J_{r-1}(n\gamma)+J_{r+1}(n\gamma)\Big]e^{i(r\varepsilon-n\alpha t-n\theta)}$$

$$+D\sum_{r=-\infty}^{\infty}\Big[J_{r-1}(n\gamma)+J_{r+1}(n\gamma)\Big]e^{-i(r\varepsilon-n\alpha t-n\theta)}$$

$$-\frac{\alpha}{\pi}\sum_{r=-\infty}^{\infty}J_r(n\gamma)\Big[e^{i(r\varepsilon-n\alpha t-n\theta)}+e^{-i(r\varepsilon-n\alpha t-n\theta)}\Big]\Big\} \tag{84}$$

Apply the relation:

$$\mathrm{J}_{r-1}(n\gamma)+J_{r+1}(n\gamma)=\frac{2r}{n\gamma}J_r(n\gamma) \tag{85}$$

Then:

$$f=\frac{\omega}{2\pi}+D\cos\varepsilon$$

$$+\sum_{n=1}^{\infty}(-x)^n\Big\{D\sum_{r=-\infty}^{\infty}\frac{r}{n\gamma}J_r(n\gamma)\cos(r\varepsilon-n\alpha t-n\theta)$$

$$-\frac{\alpha}{2\pi}\sum_{r=-\infty}^{\infty}J_r(n\gamma)\cos(r\varepsilon-n\alpha t-n\theta)\Big\} \tag{86}$$

$$=\frac{\omega}{2\pi}+D\cos 2\pi\mu t$$

$$+\sum_{n=1}^{\infty}\sum_{r=-\infty}^{\infty}(-x)^n\Big\{\frac{\mu r}{n}-\frac{\alpha}{2\pi}\Big\}J_r\Big(\frac{nD}{\mu}\Big)\cos(2\pi r\mu t-n\alpha t-n\theta) \tag{87}$$

where: $x<1$.

This shows that the effect of the interfering signal is to produce cross modulation between the desired audio signal of frequency μ and the difference angular frequency α. The amplitude of each frequency can be computed from this equation. When $\alpha=0$, i.e., common-channel interference, this reduces to:

$$f=\frac{\omega}{2\pi}+D\cos 2\pi\mu t$$

$$+\sum_{n=1}^{\infty}\sum_{r=-\infty}^{\infty}(-x)^n\frac{\mu r}{n}J_r\Big(\frac{nD}{\mu}\Big)\Big\{\cos 2\pi r\mu t\cos n\theta+\sin 2\pi r\mu t\sin n\theta\Big\}$$

$$=\frac{\omega}{2\pi}+D\cos 2\pi\mu t$$

$$+ 2\mu \sum_{r=1}^{\infty} (2r - 1)\, C(2r - 1, \frac{D}{\mu}; x, \theta) \cos \{ (2r - 1)(2\pi\mu t)\}$$

$$+ 2\mu \sum_{r=1}^{\infty} (2r)\, S(2r, \frac{D}{\mu}; x, \theta) \sin \{(2r)(2\pi\mu t)\} \tag{88}$$

where the C- and S- functions are defined as follows:

$$C(m, n; x, \theta) = \sum_{s=1}^{\infty} \frac{(-x)^s}{s} J_n(ms) \cos s\theta \tag{89}$$

$$S(m, n; x, \theta) = \sum_{s=1}^{\infty} \frac{(-x)^s}{s} J_n(ms) \sin s\theta \tag{90}$$

$$x^2 \leqq 1.$$

APPENDIX III.

Common-Channel Interference, Both Signals Modulated

When two frequency-modulated signals with a common carrier frequency produce interference, the effect is similar to that which occurs when one wave is not modulated. The exact relations can be obtained in the following way:

Let: $$e_1 = E_1 \sin (\omega t + \frac{D_1}{\mu_1} \sin 2\pi\mu_1 t) \tag{91}$$

and: $$e_2 = E_2 \sin (\omega t + \frac{D_2}{\mu_2} \sin 2\pi\mu_2 t) \tag{92}$$

be the two interfering waves. Then:

$$e_1 + e_2 = \sqrt{E_1^2 + E_2^2 + 2E_1E_2 \cos \psi}\ \sin (\omega t + \frac{D_1}{\mu_1} \sin 2\pi\mu_1 t - \varphi) \tag{93}$$

where: $$\tan \varphi = \frac{x \sin \psi}{1 + x \cos \psi}$$

and: $$\psi = \frac{D_1}{\mu_1} \sin 2\pi\mu_1 t - \frac{D_2}{\mu_2} \sin 2\pi\mu_2 t$$

The instantaneous frequency becomes:

$$f = \frac{\omega}{2\pi} + D_1 \cos 2\pi\mu_1 t - \frac{D_1 \cos 2\pi\mu_1 t - D_2 \cos 2\pi\mu_2 t}{\dfrac{\cos\psi + 1/x}{\cos\psi + x} + 1} \tag{94}$$

Envelope of Beat-note Pattern

The beat-note produced in the output of a receiver with a perfect limiter and balanced discriminator is given by:

$$\text{Output} = D_1 \cos 2\pi\mu_1 t - \frac{D_1 \cos 2\pi\mu_1 t - D_2 \cos 2\pi\mu_2 t}{\dfrac{\cos\psi + 1/x}{\cos\psi + x} + 1} \tag{95}$$

The two envelopes of the maxima and minima of the beat-note pattern are obtained by setting $\psi = 2n\pi$ or $\psi = (2n+1)\pi$, where n is an integer. This gives the result:

$$\text{Envelope} = \frac{D_1}{1+x} \cos 2\pi\mu_1 t + \frac{D_2 x}{x+1} \cos 2\pi\mu_2 t \tag{96}$$

and:

$$\frac{D_1}{1-x} \cos 2\pi\mu_1 t + \frac{D_2 x}{x-1} \cos 2\pi\mu_2 t \tag{97}$$

Fourier-Series Analysis of Instantaneous Frequency

The distortion present in the instantaneous frequency is given by

$$-\frac{d}{dt}\tan^{-1}\frac{x\sin\psi}{1+x\cos\psi} = \sum_{n=1}^{\infty}(-x)^n \cos n\psi \frac{d\psi}{dt} \tag{98}$$

Consider the expression:

$$\cos n\psi \frac{d\psi}{dt} = 2\pi \cos n\psi\,(D_1 \cos 2\pi\mu_1 t - D_2 \cos 2\pi\mu_2 t) \tag{99}$$

Make the substitutions:

$$\alpha = 2\pi\mu_1 t \qquad \gamma = n\frac{D_1}{\mu_1}$$

$$\beta = 2\pi\mu_2 t \qquad \delta = n\frac{D_2}{\mu_2}$$

Then the first term becomes:

$$D_1 \cos\alpha \cos\{\gamma\sin\alpha - \delta\sin\beta\}$$

$$= \frac{D_1}{4}\left\{e^{i\alpha} + e^{-i\alpha}\right\}\left\{e^{i\gamma\sin\alpha}e^{-i\delta\sin\beta} + e^{-i\gamma\sin\alpha}e^{i\delta\sin\beta}\right\}$$

$$= \frac{D_1}{4}\left\{e^{i\alpha}\sum_{r=-\infty}^{\infty} J_r(\gamma)\,e^{ir\alpha}\sum_{s=-\infty}^{\infty} J_{-s}(\delta)\,e^{is\beta}\right.$$

$$+ e^{-i\alpha} \sum_{r=-\infty}^{\infty} J_r(\gamma)\, e^{ir\alpha} \sum_{s=-\infty}^{\infty} J_{-s}(\delta)\, e^{is\beta}$$

$$+ e^{i\alpha} \sum_{r=-\infty}^{\infty} J_{-r}(\gamma)\, e^{ir\alpha} \sum_{s=-\infty}^{\infty} J_s(\delta)\, e^{is\beta}$$

$$+ e^{-i\alpha} \sum_{r=-\infty}^{\infty} J_{-r}(\gamma)\, e^{ir\alpha} \sum_{s=-\infty}^{\infty} J_s(\delta)\, e^{is\beta} \Big\}$$

$$= \frac{D_1}{4} \Big\{ \sum_{r=-\infty}^{\infty} \left[J_r(\gamma)\, e^{i(r+1)\alpha} + J_r(\gamma)\, e^{i(r-1)\alpha} \right] \sum_{s=-\infty}^{\infty} J_{-s}(\delta)\, e^{is\beta}$$

$$+ \sum_{r=-\infty}^{\infty} \left[J_{-r}(\gamma)\, e^{i(r+1)\alpha} + J_{-r}(\gamma)\, e^{i(r-1)\alpha} \right] \sum_{s=-\infty}^{\infty} J_s(\delta)\, e^{is\beta} \Big\} \tag{100}$$

Make the following substitutions:

$r + 1 = k$ in the first expression in the first bracket
$r - 1 = k$ in the second expression in the first bracket
$r + 1 = -\,k$ in the first expression in the second bracket
$r - 1 = -\,k$ in the second expression in the second bracket

and apply the identity:

$$J_{k-1}(\gamma) + J_{k+1}(\gamma) = \frac{2k}{\gamma} J_k(\gamma). \tag{101}$$

This gives:

$$D_1 \cos \alpha \cos \{\gamma \sin \alpha - \delta \sin \beta\}$$

$$= \frac{D_1}{4} \Big\{ \sum_{k=-\infty}^{\infty} \left[J_{k-1}(\gamma) + J_{k+1}(\gamma) \right] e^{ik\alpha} \sum_{s=-\infty}^{\infty} J_s(\delta)\, e^{-is\beta}$$

$$+ \sum_{k=-\infty}^{\infty} \left[J_{k-1}(\gamma) + J_{k+1}(\gamma) \right] e^{-ik\alpha} \sum_{s=-\infty}^{\infty} J_s(\delta)\, e^{is\beta} \Big\}$$

$$= \frac{D_1}{2} \Big\{ \sum_{k=-\infty}^{\infty} \frac{k}{\gamma} J_k(\gamma)\, e^{ik\alpha} \sum_{s=-\infty}^{\infty} J_s(\delta)\, e^{-is\beta}$$

$$+ \sum_{k=-\infty}^{\infty} \frac{k}{\gamma} J_k(\gamma)\, e^{-ik\alpha} \sum_{s=-\infty}^{\infty} J_s(\delta)\, e^{is\beta} \Big\}$$

$$= D_1 \sum_{r=-\infty}^{\infty} \sum_{s=-\infty}^{\infty} \frac{r}{\gamma} J_r(\gamma)\, J_s(\delta) \cos (r\alpha - s\beta) \tag{102}$$

By the same process:

$$D_2 \cos \beta \cos \{\gamma \sin \alpha - \delta \sin \beta\}$$

$$= D_2 \sum_{r=-\infty}^{\infty} \sum_{s=-\infty}^{\infty} \frac{s}{\delta} J_r(\gamma)\, J_s(\delta) \cos (r\alpha - s\beta) \tag{103}$$

These two results give the expression:

$$\cos n\psi \frac{d\psi}{dt} = 2\pi \sum_{r=-\infty}^{\infty} \sum_{s=-\infty}^{\infty} \left\{ \frac{rD_1}{\gamma} - \frac{sD_2}{\gamma} \right\} J_r(\gamma)\, J_s(\delta) \cos (r\alpha - s\beta)$$

$$= \frac{2\pi}{n} \sum_{r=-\infty}^{\infty} \sum_{s=-\infty}^{\infty} (r\mu_1 - s\mu_2)\, J_r(\gamma)\, J_s(\delta) \cos (r\alpha - s\beta) \tag{104}$$

The audio output from a balanced discriminator thus becomes:

Output $\propto D_1 \cos 2\pi\mu_1 t$

$$\sum_{n=1}^{\infty} \sum_{r=-\infty}^{\infty} \sum_{s=-\infty}^{\infty} \frac{(-x)^n}{n} (r\mu_1 - s\mu_2)\, J_r\left(\frac{nD_1}{\mu_1}\right) J_s\left(\frac{nD_2}{\mu_2}\right) \cos (r\alpha - s\beta)$$

$$= D_1 \cos 2\pi\mu_1 t + \sum_{r=-\infty}^{\infty} \sum_{s=-\infty}^{\infty} (r\mu_1 - s\mu_2)\, C\left(r, \frac{D_1}{\mu_1}; s, \frac{D_2}{\mu_2}: x, 0\right)$$

$$\cos (r\alpha - s\beta) \tag{105}$$

where the generalized C-function is defined as follows:

$$C(k, l; m, n: x, \theta) = \sum_{s=1}^{\infty} \frac{(-x)^s}{s} J_k(s\,l)\, J_m(sn) \cos s\theta, \tag{106}$$

$\alpha = 2\pi\mu_1 t$ and $\beta = 2\pi\mu_2 t.$

PULSE TIME DIVISION RADIO RELAY*†

By

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Summary—*This paper describes a radio relay set which provides eight telephone circuits and operates in the 1350 to 1500-megacycle band. The set uses a time division multiplex system and pulse position modulation. An explanation of the system is given. The signal-to-noise ratio at threshold signal condition is derived and compared with the measured signal-to-noise ratio. This radio set was developed for the U. S. Army during the war and is designated the AN/TRC-5.*

Introduction

THE radio relay set herein described consists of a radio transmitter and receiver along with its associated multiplex apparatus. Two of these sets provide eight two-way telephone circuits of high quality over a line of sight path up to 50 or 100 miles long. The performance of each radio set permits operation of many radio links in tandem so that large distances may be covered. The receiving and transmitting circuits of one set use separate radio frequencies in the range of 1350 to 1500 megacycles. The radio frequency energy is transmitted in short pulses and multiplexing is accomplished on a time division basis. Modulation of a particular channel pulse is effected by displacing it, from its normal unmodulated position, by a time interval proportional to the amplitude of the modulating signal. This is known as pulse position modulation. This radio set was developed for the U. S. Army Signal Corps during the war, and was designated AN/TRC-5. An overall view of the set is shown in Figure 1.

The total power drain for one radio set is 1500 watts from a 115 volt, 50/60-cycle source. The peak radiated power is 400 watts, which corresponds with an average power of 20 watts. An input level of one milliwatt will modulate any channel 100 per cent and the output of this channel will be one milliwatt. The cross talk between adjacent channels when one channel is modulated 100 per cent is 60 decibels below the level corresponding to 100 per cent modulation.

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* Decimal Classification: R148.6 × R480.

† This paper and "A Microwave Relay Communication System", which follows, comprise a detailed presentation of specific equipment referred to in C. W. Hansell's general paper on radio relays in the September 1946 issue of *RCA REVIEW*.

DESCRIPTION OF THE MULTIPLEX SYSTEM

The pulses of a single audio channel may be considered as a phase modulated subcarrier. This subcarrier must have a frequency greater than twice the highest modulating frequency (with small phase deviations) in order that the lower sideband may be separated from the highest modulating frequency. In this system the channel pulse repetition rate was set at 10,000 per second, giving an adequate margin for modulation frequencies up to 3000 cycles per second. Since the basic channel period is 100 microseconds it is necessary to fit eight channel pulses and a synchronizing or marker pulse into this time interval. Accordingly, this interval is divided into nine equal periods of 11.1 microseconds with a pulse assigned to each period. A marker pulse

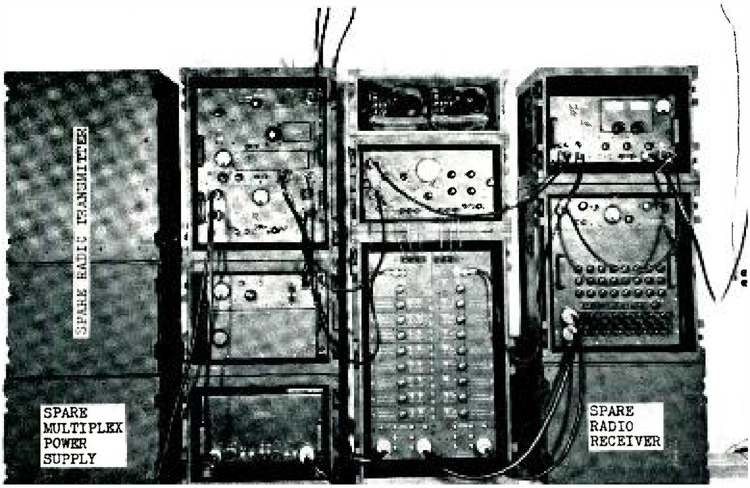

Fig. 1—Overall view of the AN/TRC-5.

serves as a time base and reference point to allow the respective pulses to be switched to their proper channels at the receiving end of the circuit.

One basic time period is shown in Figure 2 in which S represents the 2 microsecond marker pulse, and the 0.4-microsecond channel pulses are numbered 1-8. In order to avoid interference and cross channel modulation, the maximum time displacement of any channel pulse is limited to $\pm$ 3.5 microseconds, which allows an adequate guard space between adjacent pulses when each is fully modulated. The extra width of the marker pulse allows it to be isolated from the channel pulses at the receiver to recreate the original basic time period. The limits of modulation and the guard space are shown by the dotted lines for channels 1 and 2.

A modulating voltage displaces its channel pulse from the normal unmodulated position by an amount corresponding to the instantaneous amplitude of the modulating voltage at the occurrence time of that pulse. Thus the pulse is displaced at a rate corresponding to the frequency of the modulating voltage. The displacement corresponding to 100 per cent modulation is set at $\pm$ 3.5 microseconds. Limiting is used to prevent greater displacements to avoid excessive cross talk between channels. Figure 3 shows ten consecutive traces of a channel pulse with 100 per cent modulation by a 1000-cycle tone. Each horizontal line is that part of the basic time period allotted to this channel.

MULTIPLEX EQUIPMENT

The transmitting and receiving multiplex equipments are housed in one carrying case which is divided in two sections. Each equipment consists of a removable common unit with its eight associated identical

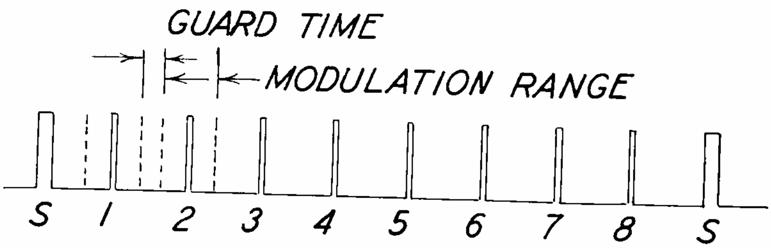

Fig. 2—One basic time period.

channel units. Each channel unit is removable without interfering with the operation of the other channel units. Thus the whole multiplex equipment is comprised of two common units and sixteen channel units. The removable feature of these units allows complete accessibility to all components for maintenance and repair. Figure 4 is a block diagram of the AN/TRC-5 set.

The transmitting multiplex equipment generates a signal as shown in Figure 2, which is used to drive the radio transmitter. The receiving multiplex equipment receives the pulsed signal from the radio receiver and delivers eight voice channel outputs. The transmitting and receiving multiplex equipments are entirely independent except for a common power supply.

The transmitting common unit functions are:

(1) To allot equal consecutive 11.1 microsecond time intervals to the eight channel units.

(2) To generate a marker pulse and locate it properly in its assigned 11.1 microsecond interval.
(3) To combine, shape, and amplify the output pulses from all eight channel units.

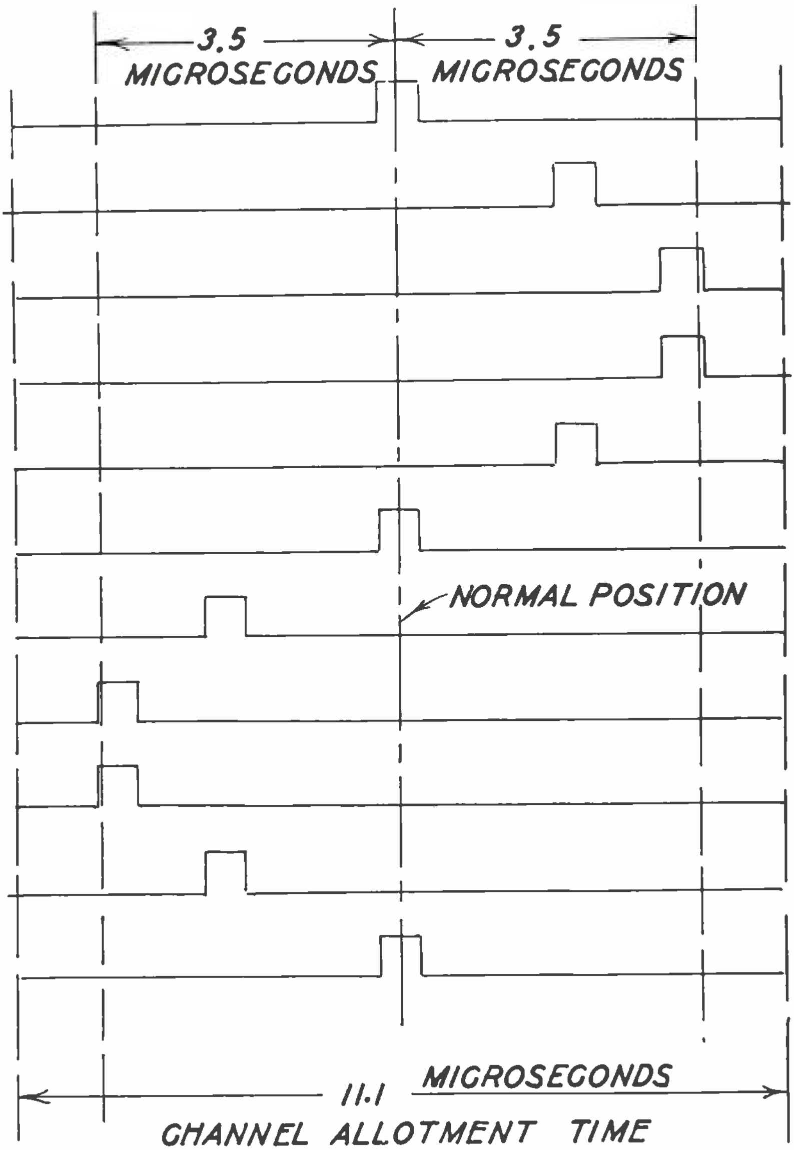

Fig. 3—Consecutive traces of one channel pulse modulated with 1,000 cycle tone.

The transmitting channel unit functions are:

(1) To generate a pulse properly located in the time interval assigned by the common unit.

(2) To vary the occurrence time of the generated pulse in accordance with the audio input to the channel.

(3) To advance the pulse to the leading edge of its channel interval and keep it there for the duration of applied ringing voltage from the audio input terminals.

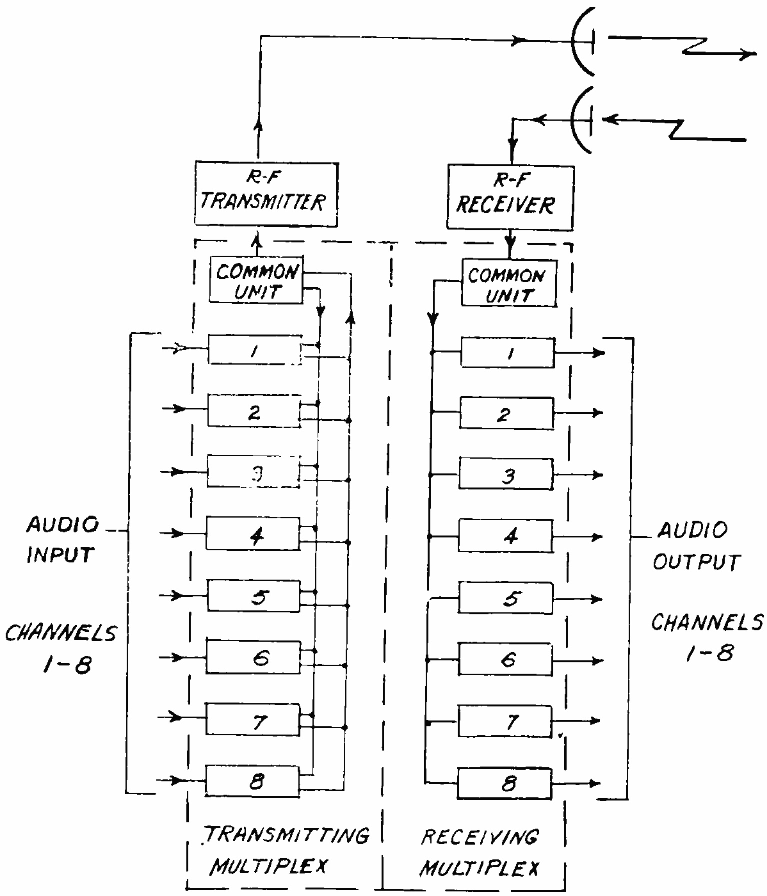

Fig. 4—Block diagram of the AN/TRC-5.

The transmitting common unit consists of: (1) a 90-kilocycle crystal-controlled oscillator which is used to lock a 90-kilocycle pulse oscillator; (2) a counter circuit which produces a step wave consisting of nine steps each 11.1 microseconds long (the step wave is coupled to the channel units and allows 11.1-microsecond time intervals to be determined on an amplitude basis); (3) a marker pulse generator which produces a 2-microsecond pulse and locates it in the period immediately succeeding the step wave discharge; (4) an amplifier stage which shapes and amplifies the pulses from the channel units; and (5) an output stage which combines the channel and marker pulses and couples them to the radio transmitter.

A transmitting channel unit consists of: (1) a channel selector which allows the channel unit to operate in a desired time interval; (2) a saw-tooth generator which produces a saw-tooth wave whose length is adjustable and is made equal to the modulation range (about 7 microseconds); (3) a pulse generator and modulator which produces a pulse whose occurrence time is made variable over the linear rise

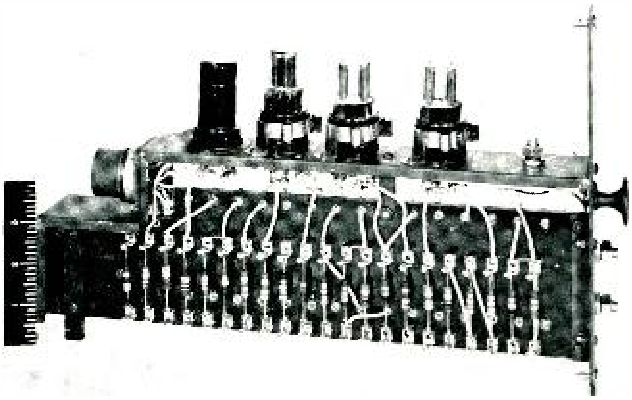

Fig. 5—The transmitting multiplex channel unit. (Capacitors are on the back of the terminal board).

time of the saw-tooth waves; and (4) a circuit which rectifies the input ringing voltage, causing the pulse to remain in an advanced position for the duration of the ringing signal. Figure 5 shows the transmitting channel unit.

The receiving common unit functions are:

(1) To separate the marker pulse from the radio receiver output.

(2) To derive from the separated marker pulse a new set of pulses having a 90-kilocycle repetition rate which in turn is used to generate a step wave having nine steps. (This step wave is used to allot equal consecutive 11.1-microsecond time intervals to the eight channel units.)

The receiving channel unit functions are:

(1) To isolate a particular channel pulse and convert the time displacement modulation into pulse width modulation.
(2) To detect the width modulation by means of a low-pass filter.
(3) To generate a direct-current voltage which operates the ringing relay when the channel pulse is held at its most advanced position.

The receiving common unit consists of: (1) a marker pulse separator circuit; (2) a circuit to generate pulses having a 90-kilocycle repetition rate from the separated marker pulses; (3) a circuit to allow an adjustment of the phase of the 90-kilocycle pulse rate; and (4) a

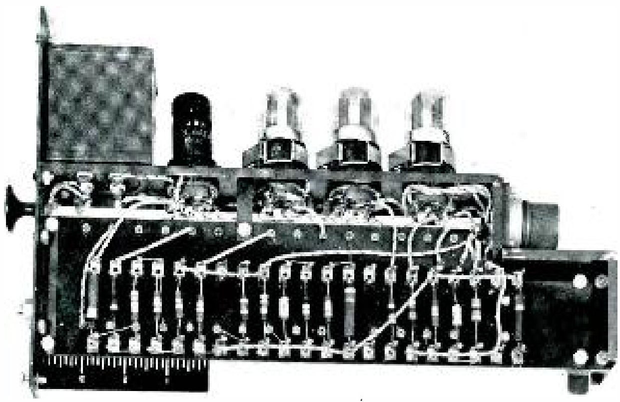

Fig. 6—The receiving multiplex channel unit. (The 3,000-cycle low-pass filter is at the upper left. The capacitors are on the back of the terminal board).

step wave generator to produce a step wave having nine steps. The interval between risers is 11.1 microseconds. The step wave is coupled to the channel units and allows 11.1-microsecond time intervals to be determined on an amplitude basis as was done in the transmitting multiplex equipment.

A receiving channel unit consists of: (1) a channel selector which allows the channel unit to operate in a desired time interval; (2) a multivibrator type circuit and associated gate to select the proper channel pulse and convert the pulse position modulation into a pulse width modulation; (3) a low pass filter and audio amplifier; and (4) a circuit to generate a negative direct-current voltage when the channel pulse is held at its most advanced position. This voltage is applied to the grid of the audio amplifier tube to provide a cut-off bias which operates the ringing relay in the plate circuit of the audio tube. The receiving channel unit is shown in Figure 6.

As an aid in operating and maintaining the multiplex equipment a suitable cathode-ray oscilloscope is provided. The horizontal linear sweep circuits are driven from either the transmitting or receiving multiplex units. This arrangement gives a permanently synchronized horizontal sweep of the correct frequency.

In order that this radio set be capable of operating into a telephone switchboard, or any set of two wire pairs, hybrid circuits are provided to convert the eight incoming and eight outgoing audio pairs into eight two-wire lines. These hybrid transformers are located in a separate terminal unit which also contains the terminals for the audio lines, ringing relays, 20-cycle ringing generator, metering facilities, 1000-cycle test oscillator, and hybrid switching arrangements. The switching circuits allow the operator to set up any channel on a two-wire, two-way, or four-wire, two-way, basis. The 1000-cycle oscillator is supplied for setting line levels properly.

Radio Transmitter

The transmitter consists of a magnetron oscillator and its associated equipment, as shown in Figure 7. The magnetron has an oxide cathode, eight anodes, and eight resonant cavities. Alternate anodes are tied together with two strapping rings. The magnetron operating frequency is adjustable from 1350 to 1500 megacycles by means of a tunable section of coaxial line coupled to the cavities. The adjustable short on the coaxial line is outside the evacuated envelope. Output is obtained by a loop coupled into one cavity, as is commonly done with this type of magnetron.

The magnetic field for the magnetron is supplied by an adjustable permanent magnet. Normal field for operation of the magnetron is about 900 oersteds.

A matching circuit adjusts the impedance of the antenna to load the magnetron most favorably. The matching circuit has a high Q so as to attenuate the unwanted side frequencies resulting from the abrupt starting time of the magnetron.

The radio frequencies are measured by a quarter wave long coaxial line operating as an absorber type wavemeter.

The pulse drive for the magnetron is obtained from a three-stage transformer coupled limiter-amplifier. The first two stages use 6V6 type tubes. The output stage uses two 813 type tubes and is coupled through a blocking condenser to the magnetron cathode. The peak amplitude of the pulse voltage input to the magnetron is about 3500 volts and the peak current is about one ampere.

Operation is made independent of the 60-cycle supply line voltage over a limited range by using a current regulator to maintain the average current of the magnetron constant. Since the per cent mark or duty factor of the system does not vary, maintaining the average current constant is equivalent to maintaining the peak current constant.

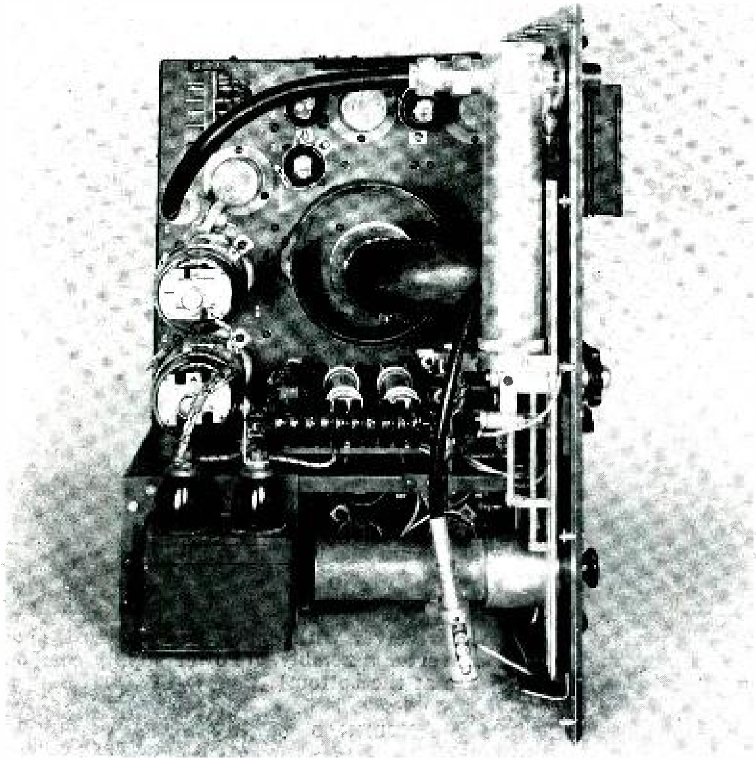

Fig. 7—Top view of the radio transmitter showing the magnetron and magnet. (The 813 driver tubes are to the left, antenna matching unit to the right.)

Radio Receiver

The superheterodyne radio receiver (see Figure 8) uses a 1N21 type crystal converter with a 2C40 triode local oscillator which can be tuned to either side of the signal frequency. The 16-megacycle intermediate frequency amplifier using 6AC7 tubes is single-tuned having a bandwidth of 3 megacycles. This is followed by a multistage video limiter and clipper and an output cathode follower stage. The receiver noise factor referred to a KTΔf base is 10 to 12 decibels, depending upon the choice of crystal used in the converter. An image ratio of better than 28 decibels is maintained over the tuning range by two ganged, quarter-wave, resonant selective circuits preceding the crystal converter.

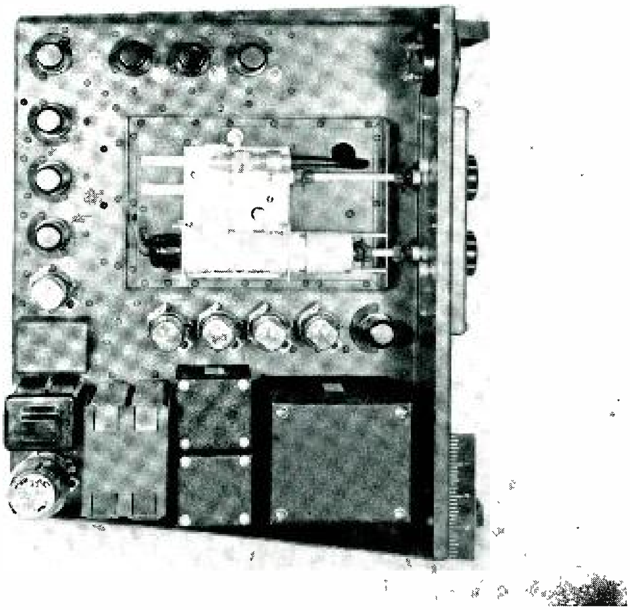

Fig. 8—Top view of the radio receiver with dust cover removed, showing local oscillator and crystal converter assembly. (Power supply components are at the bottom).

ANTENNAS

The transmitting and receiving antennas are identical and supported side by side on a 50 foot pipe mast at each radio set. Figure 9 is a photograph of such an installation. Each consists of a parabolic reflector, 50 inches in diameter, which directs the radiation from a half-wave dipole. The directivity of the antenna is further increased by a parasitic half-wave dipole located in front of the fed dipole. The reflectors consist of a tubular framework supporting a galvanized wire mesh with openings about three-eighths of an inch square. For ease of carrying, each reflector is built in two sections and held together in operation by four draw clamps. The power gain of each antenna over a half-wave dipole is about 19 decibels. The total beam width at half power is 15 degrees. Coupling between antennas, both horizontally polarized, on the same mast measures 70 decibles at the coaxial feed line inputs. The 65-foot coaxial cable feed line loss of 5 decibels combined with the antenna gain results in a net gain of 14 decibels.

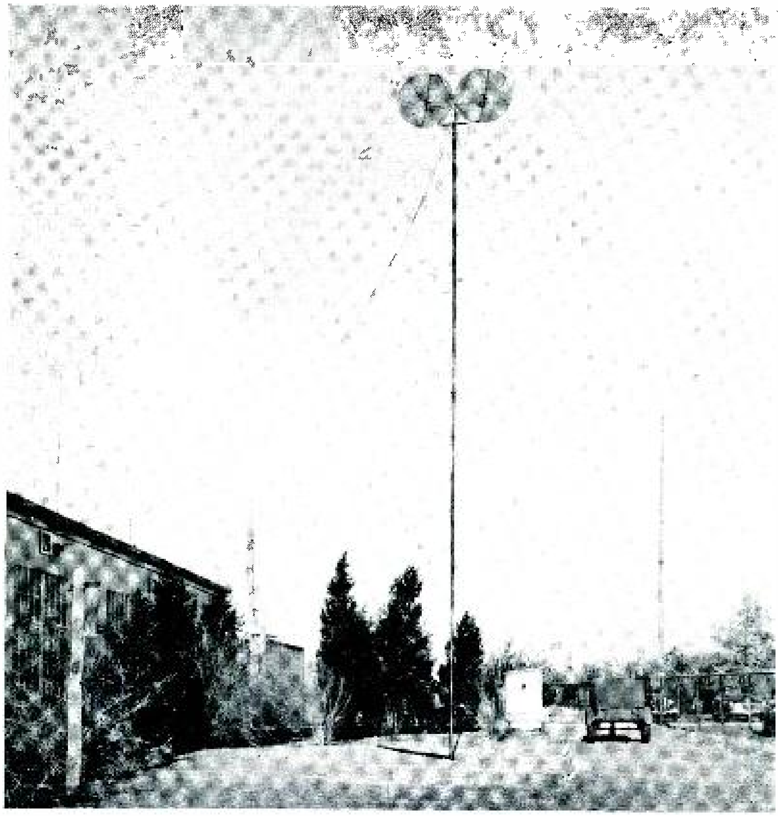

Fig. 9—Antenna assembly at the top of 50-foot mast. (The mast rotation bar is near the bottom).

Pulse Modulation Considerations

Time division multiplexing with pulse position modulation offers several advantages over the frequency division method of channeling. Perhaps the most important, from the military point-of-view, is the elimination of heavy filters, the use of which involves quite linear amplifiers to avoid excessive cross talk between channels. Another advantage is the use of readily obtainable standard capacitors and resistors with tolerances no better than ± 10 per cent. Pulses are handled entirely with non-linear amplifiers, which gives an easily-managed system free from the usual cross talk problems. At the time of this development the pulsed magnetron was the most satisfactory transmitting tube which could be made available to give 20 or 30 watts average power with relatively high efficiency. This fact further contributed toward the choice of a pulse system.

A system using pulse position modulation provides a reduction in output noise as the occupied frequency band is increased. This noise reduction is analogous to that obtained in continuous-wave phase and frequency modulation systems. This characteristic results in a fairly large audio signal-to-noise ratio for all signal levels down to the threshold level. The threshold signal occurs when the peak pulse amplitude is just 6 decibels above the peak noise amplitude in the frequency band required to pass the pulses. In the following it will be assumed that transmitted signal has a constant average power, a constant pulse repetition frequency, and the minimum pulse length that the receiver band will pass. The transmitted pulse must always have a rise time at least as short as the reciprocal of the receiver pass band. Under these conditions the peak pulse power and the peak noise power both increase linearly with increase in bandwidth. A gain in the audio signal-to-noise ratio is obtained by decreasing the pulse length and increasing the bandwidth to correspond, but this does not result in a change of the threshold signal level at the receiver and so does not allow any greater distances to be covered despite the higher peak power. More average power is required to cover a greater distance.

To a close approximation the audio signal-to-noise ratio at the threshold signal level is:

$$\frac{S}{N} = 6.32\, B_r t_m \sqrt{\frac{f_p}{B_m}} \text{ (root-mean-square) for random noise} \quad (1)$$

where B_r = the effective radio-frequency bandwidth for noise purposes of the receiver in cycles per second (usually the intermediate frequency amplifier bandwidth).

t_m = peak time displacement of pulse due to modulation in seconds.

B_m = the effective bandwidth for noise purposes of the audio system in cycles per second.

f_p = pulse repetition frequency in cycles per second.

Equation (1) is derived as follows: Figure 10 shows the time position modulation of the edges of the signal pulse due to noise. The time of rise of the leading edge of the pulse is equal to $(B_r)^{-1}$. This will be the case for the assumed condition that the transmitted pulse has a build-up time as fast or faster than this value. For simplicity the noise is represented as having a steady peak amplitude equal to about one-half the peak amplitude of the signal pulse. The solid lines repre-

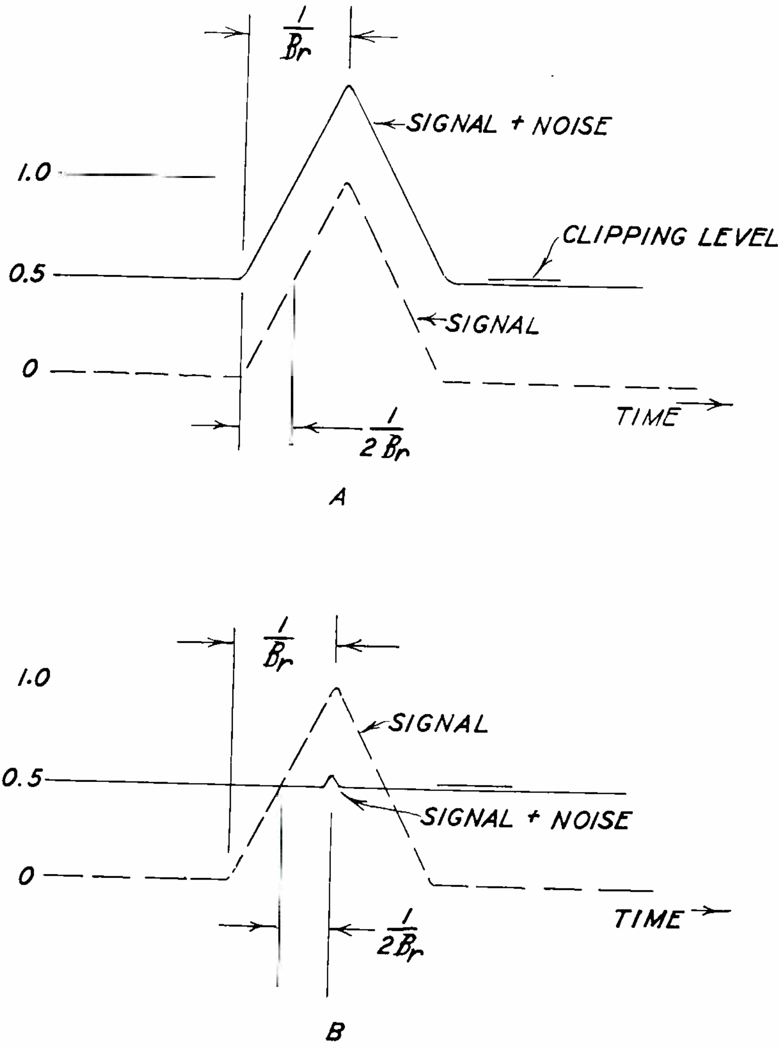

Fig. 10—Time position modulation of signal by noise.

sent the resultant of the noise and the signal pulses after linear rectification without restriction of bandwidth. If the phase of the noise voltage is such as to add to the signal, the leading edge of the resultant will be advanced $\frac{1}{2B_r}$ seconds as shown in Figure 10(A). If the noise voltage is such as to oppose the signal at the proper time the leading edge of the resultant will be retarded $\frac{1}{2B_r}$ seconds as shown in Figure 10(B). In order to realize the wide band gain of the system it is necessary to limit and clip the pulses so that no noise is present other than that which causes a time displacement of the leading and trailing edges of the pulses. This is equivalent to amplifying a thin slice of the resultant signal at a level just above the 0.5 amplitude. The resulting signal will consist of pulses of constant amplitude and variable width.

The peak deviation of the leading edge of the pulse due to random noise is:

$$t_N = \frac{1}{2B_r} \text{ seconds, at threshold.}$$

The ratio of the peak deviation due to the audio signal to the peak deviation due to the noise is:

$$\frac{t_m}{t_N} = 2B_r t_m \quad \text{(peak)}.$$

Since only part of the noise is passed by the audio amplifier, the signal-to-noise ratio at the audio terminals is:

$$\frac{t_m}{t_n} = 2B_r t_m \sqrt{\frac{f_p}{B_m}} \quad \text{(peak)}.$$

The ratio of the root-mean-square voltage to the peak voltage for a sine wave is 0.707. The ratio of the root-mean-square voltage to the peak voltage for thermal agitation noise is 0.224. Therefore, for a threshold signal the root-mean-square signal-to-noise ratio at the audio terminals is:

$$\frac{S}{N} = \frac{.707}{.224} 2B_r t_m \sqrt{\frac{f_p}{B_m}}$$

$$= 6.32\, B_r t_m \sqrt{\frac{f_p}{B_m}} \quad \text{(root-mean-square)}. \qquad (1)$$

The above analysis assumes a pulse having a linear rise with time, which is not actually the case. Experience has shown, however, that equation (1) is sufficiently accurate for all practical purposes.

Substituting in equation (1) values corresponding with the factors in the AN/TRC-5 radio set, the calculated threshold signal-to-noise ratio is 41.6 decibels. This calculation does not take into account noise introduced in all channels by the marker pulse. The audio output is determined by the phase relation between the marker pulse and a channel pulse. Since the marker pulse is also position modulated by noise and contributes an amount of noise power to the output terminals equal to that due to each channel pulse, the above calculated signal-to-noise ratio should be decreased by 3 decibels. Actual measurements show the marker pulse reduces the signal-to-noise ratio only 2.3 decibels. The selectivity of the 90-kilocycle circuit which is excited by the marker pulse reduces the noise contributed by the marker pulse somewhat. The measured signal-to-noise ratio at threshold for one channel in the AN/TRC-5 is + 39.3 decibels.

The work described in this paper was performed under War Department contract W3434-sc-279 with Radio Corporation of America.

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A MICROWAVE RELAY COMMUNICATION SYSTEM*†

BY

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Summary—*This paper reviews the experimental results obtained with a 4000-megacycle multichannel relay system connecting New York and Philadelphia. The system employs a frequency-modulated subcarrier which in turn is used to frequency-modulate the final carrier. Demodulation to the subcarrier frequency is effected at relay stations. The microwave relay equipment, which resulted from this experimental work and which will be installed by the Western Union Telegraph Company in a circuit connecting New York, Washington and Pittsburgh, is described.*

INTRODUCTION

UNTIL recently, multichannel radio circuits were employed as trunk lines in telephone and telegraph systems only where the installation of wire lines was impracticable. Such radio circuits usually operated at ultra-high frequencies and were limited to single hops over short distances[1]. This limited use of radio was not surprising in view of the technical difficulties involved in broad-band, multichannel operation in the ultra-high frequency region.

The rapid development of microwave techniques during the wartime period, however, provided many new tools for the radio communications engineer, and economical operation of long-distance, multichannel relay systems in the microwave region can now be realized[2,3,4].

At microwave frequencies, the wide modulating bands employed in multiplexed telephony and telegraphy represent a small percentage

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* Decimal Classification: R480.

† Presented at the National Electronics Conference, Chicago, Illinois, on October 4, 1946.

1 N. F. Schlaak and A. C. Dickieson: "Cape Charles-Norfolk Ultra-Short-Wave Multiplex System," *Proc. I.R.E.*, Vol. 33, No. 2, pp. 78-83, February, 1945.

2 C. W. Hansell, "Development of Radio Relay Systems", *RCA REVIEW*, Vol. VII, No. 3, pp. 367-384, Sept., 1946.

3 Colonel Julian Z. Millar, "A Preview of the Western Union System of Radio Beam Telegraphy," *Jour. Frank. Inst.*, Vol. 241, No. 12, pp. 397-413, June 1946; and Vol. 242, No. 1, pp. 23-40, July, 1946.

4 John J. Kelleher, "Pulse-Modulated Radio Relay Equipment", *Electronics*, Vol. 19, No. 5, pp. 124-129, May, 1946.

of the final carrier frequency, and the design of economical equipment is, therefore, greatly simplified. Highly directional antenna systems at these frequencies are both practicable and economical, and low-power transmitters may be employed. Since transmission is confined to narrow beams over line-of-sight paths, interference between adjacent systems is minimized and the frequency spectrum utilization is efficient.

The relaying of multiplexed signals through a large number of repeater stations, however, still imposes serious problems in the design of modulators, demodulators, and repeaters. Each repeater station will introduce some noise, distortion, and cross-modulation, and the cumulative effects of these signal degradations must be held to tolerable limits.

Early in 1945 RCA installed an experimental microwave relay system connecting New York and Philadelphia for the purpose of observing propagation effects, and for the field-testing of developmental equipment. These studies were made with the cooperation of the Western Union Telegraph Company. The early experimental results were encouraging, and a design program was started which resulted in the equipment described in this paper.

The Western Union Telegraph Company is now installing equipment of this type in a microwave relay system connecting Washington, Pittsburgh, and New York.[3]

Methods of Operation

In designing a multichannel radio system suitable for public-carrier telephone and telegraph trunk-line operation, it was considered highly desirable to provide for straightforward interconnection with existing land lines, and with the conventional frequency-division multiplexing equipments employed in land-line systems.[3,5,6,7] It was also considered a necessity to provide voice communication facilities (service channel) between relay stations and terminal stations as an aid to test and maintenance personnel. Another requirement, imposed by the fact that relay stations are normally unattended, was the provision of fault-locating means capable of localizing equipment failures.

A modulation method which satisfies these requirements was devel-

[5] B. W. Kendall and H. A. Affel, "A Twelve-Channel Carrier Telephone System for Open-Wire Lines," *Bell Sys. Tech. Jour.*, Vol. 18, No. 1, pp. 119-142, January, 1939.

[6] F. B. Bramhall and J. E. Boughtwood, "Frequency-Modulated Carrier Telegraph System," *Elec. Eng.*, Vol. 61, No. 1, pp. 36-39, January, 1942.

[7] F. B. Bramhall, "Carrier Telegraph Systems", *Elec. Eng.*, Vol. 63, No. 8, pp. 288-286, August, 1944.

oped by L. E. Thompson. This method is illustrated in the block diagram shown in Figure 1 which, for convenience, includes only one repeater station. At the transmitting terminal, multiplexed signals, in a band of 30 to 150,000 cycles, frequency-modulate a 1.0-megacycle subcarrier, which, in turn, frequency-modulates the final microwave carrier. At relay stations, this frequency-modulated subcarrier is recovered by demodulation of the received signal, and the subcarrier is then used to frequency-modulate the relaying transmitter. Demodulation of the subcarrier does not occur except at the receiving terminal of the relay chain.

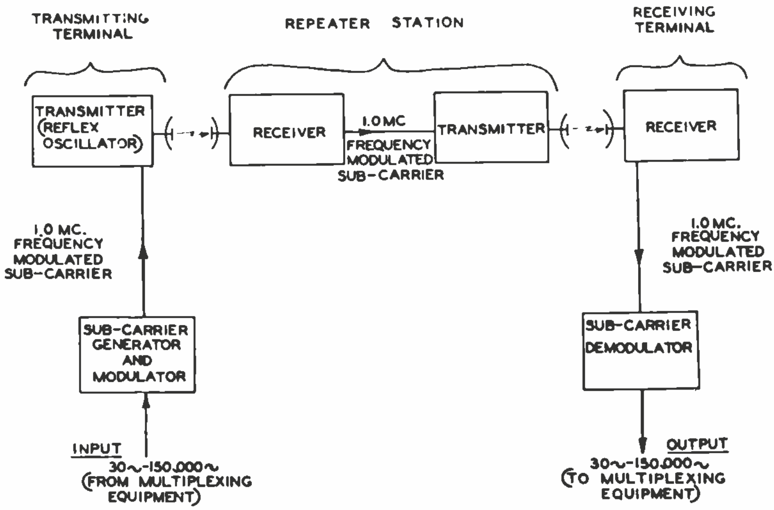

Fig. 1—Block diagram of basic system.

This method greatly simplifies the design problems, since only the modulator and demodulator of the subcarrier need be linear, and these are employed only at the terminal stations. The equipment required at relay stations becomes relatively simple; here, nonlinear demodulators and modulators may be employed without causing degradation of the multiplexed signals, since waveform distortion of the frequency-modulated subcarrier wave is not harmful.

Figure 2 illustrates, in block diagram form, the double frequency-modulation process performed at the transmitting terminal of a relay chain. Signals, in the intelligence band of 30 cycles to 150,000 cycles, which are usually provided by conventional frequency-division multiplexing equipments, are applied through a pre-emphasis network to two reactance-tube modulators, operating in push-pull, which frequency-modulate two oscillators—one operating at 9.0 megacycles and the other at 9.33 megacycles. The frequencies of these oscillators are tripled, and the outputs of the tripler stages are combined in a mixer. The difference frequency of 1.0 megacycle is utilized as the subcarrier. This subcarrier, which has a peak deviation of plus and minus 400 kilocycles, is applied to the repeller electrode of a reflex oscillator tube[8] operating at about 4000 megacycles. The deviation of the final carrier is approximately plus and minus two megacycles. For service channel purposes, voice signals or audio tones may be applied to the repeller electrode, simultaneously with the 1.0-megacycle subcarrier, to frequency-modulate the final carrier directly, with a maximum deviation of approximately plus and minus 75 kilocycles.

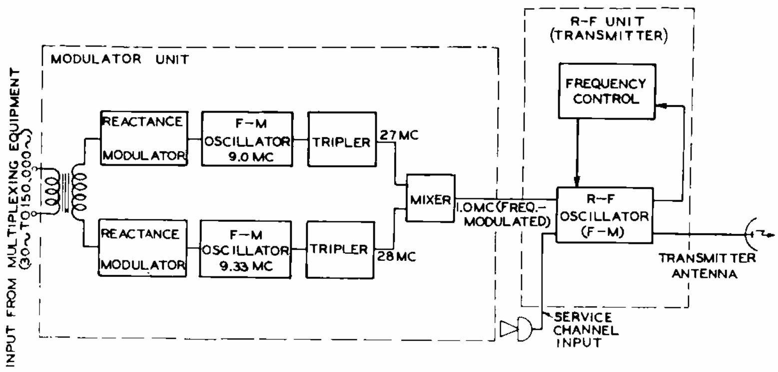

Fig. 2—Block diagram of transmitting terminal station.

Operation at a relay station is illustrated in Figure 3. The received signal is combined in a crystal mixer with the output of a local reflex oscillator. Amplification and limiting are performed at an intermediate frequency of 32 megacycles, and the 1.0-megacycle subcarrier is recovered at a 32-megacycle discriminator. The subcarrier, after amplification and limiting, frequency-modulates the reflex oscillator of the outbound relaying transmitter. The service channel signals are also derived at the 32-megacycle discriminator for operation of headphones and the relaying transmitter may be modulated directly by the output of a microphone. In a two-way relay circuit, these service channel facilities at terminal and relay stations provide a system-wide "party line."

[8] J. R. Pierce, "Reflex Oscillators", *Proc. I. R. E.*, Vol. 33, No. 2, pp. 112-118, February, 1945.

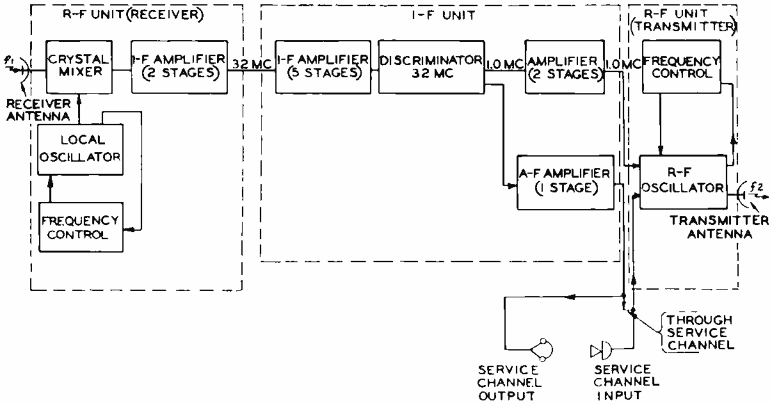

Fig. 3—Block diagram of a one-way repeater station.

Figure 4 illustrates the receiving terminal. The operation of a receiver terminal is identical with that of a relay station up to the subcarrier output circuit of the intermediate-frequency unit. Instead of being used to frequency-modulate a relaying transmitter, however, the subcarrier is fed to a demodulator circuit. In this demodulator, the 1.0-megacycle subcarrier is first converted to 13 megacycles in a balanced mixer circuit, and is then amplified and limited before demodulation by the 13-megacycle discriminator. This conversion to 13

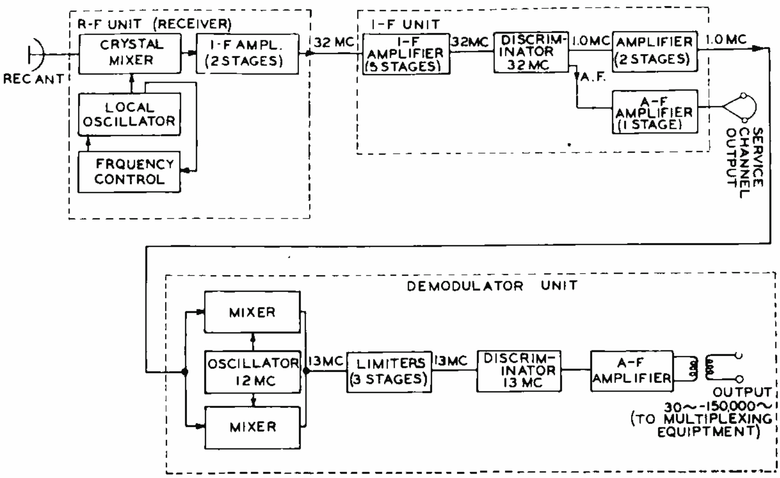

Fig. 4—Block diagram of receiving terminal station.

megacycles is performed in order to simplify the design of the subcarrier discriminator, which requires a high degree of linearity. The intelligence-bearing signals derived from the 13-megacycle discriminator are amplified and fed through a de-emphasis network to conventional frequency-division multiplexing equipment for conversion to individual voice bands or telegraph channels.

Figure 5 illustrates the method of diversity operation employed

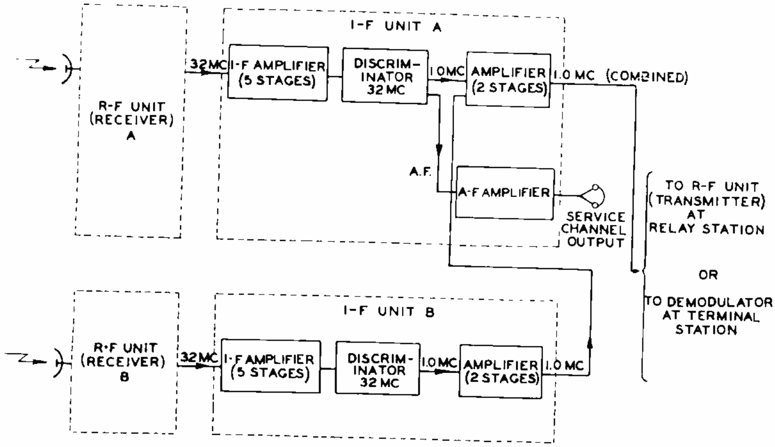

Fig. 5—Block diagram of receivers connected for diversity operation.

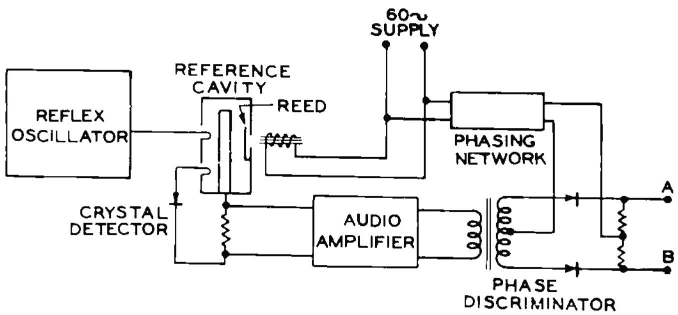

Fig. 6—Functional diagram of automatic frequency control used in transmitter and receiver.

for transmission hops greater than about 15 miles. The two receivers, separated vertically by approximately 25 feet, are identical. The subcarrier signals of the two receivers are mixed in the limiter stages of one of the intermediate-frequency units. At a relay station the combined 1.0-megacycle subcarrier modulates the outgoing transmitter; at a receiving terminal the combined subcarrier is demodulated.

The automatic frequency control circuit shown in Figure 6 operates by comparing the oscillator frequency with the resonant frequency of a stabilized coaxial cavity having a Q of approximately 800. A small reed in the cavity is driven magnetically at 60 cycles per second, and the vibration of this reed varies the resonant frequency of the cavity by approximately plus and minus 1.0 megacycle at the rate of 60 cycles per second. Coupled to the cavity is a crystal detector. Audio currents from this detector are amplified and fed to a balanced phase-discriminator circuit.

When the frequency of the oscillator corresponds to the midpoint of the frequency range swept by the cavity, the alternating-current component of the crystal current has a frequency twice that of the driving frequency, or 120 cycles. This 120-cycle current produces no unbalance in the discriminator circuit and the voltage between A and B is zero. However, if the frequency of the oscillator drifts from the midpoint of the cavity sweep range, the crystal current will contain a 60-cycle component. The phase of this 60-cycle component, when the oscillator frequency drifts higher in frequency, is 180 degrees different from its phase when the oscillator drifts lower in frequency. These 60-cycle currents produce a differential voltage between A and B, the polarity of which depends on whether the oscillator has drifted higher or lower than the midpoint of the sweep range of the reference cavity. This differential voltage is applied to the grid of a control tube which determines the potential of the repeller electrode of the reflex oscillator, and effects, in the repeller electrode potential, a change sufficient to return the oscillator frequency close to the midpoint of the cavity sweep range.

A loop-test method is employed for locating trouble in a two-way relay circuit. Each relay station is assigned a particular identifying audio-frequency tone and is equipped with means for looping this identifying tone back to the terminal station from which it was transmitted. In case of circuit failure, the operator at a terminal station performs loop tests, through successive stations, by applying the appropriate identifying tones to the service channel input of his transmitter, and observing the returned tones. The operator locates the failure, approximately, by noting the first relay station which fails to return its test tone. Further tests enable him to locate exactly the station at which the failure has occurred.

The fault-locating equipment which provides the loop circuit at relay stations is illustrated in Figure 7. A narrow pass-band filter, accepting only the correct identifying tone, loops the service channel output of the receiver in the eastbound circuit to the service channel input of the westbound transmitter. A similar loop is made between the westbound and eastbound circuits. Included in this loop circuit are a gate stage and a balanced modulator. The gate stage is cut off upon failure of the 1.0-megacycle output of either receiver. Thus, if a failure occurs in receivers A or B, or in the westbound transmitter, T_1, the test loop through the station is broken. The identifying test tone normally passed through the loop circuit is modulated at the balanced modulator by the 60-cycle and 120-cycle voltages originating in the automatic frequency control circuits of the outgoing transmitter, T_2. If operation of T_2 is normal, the 120-cycle voltage is high and the

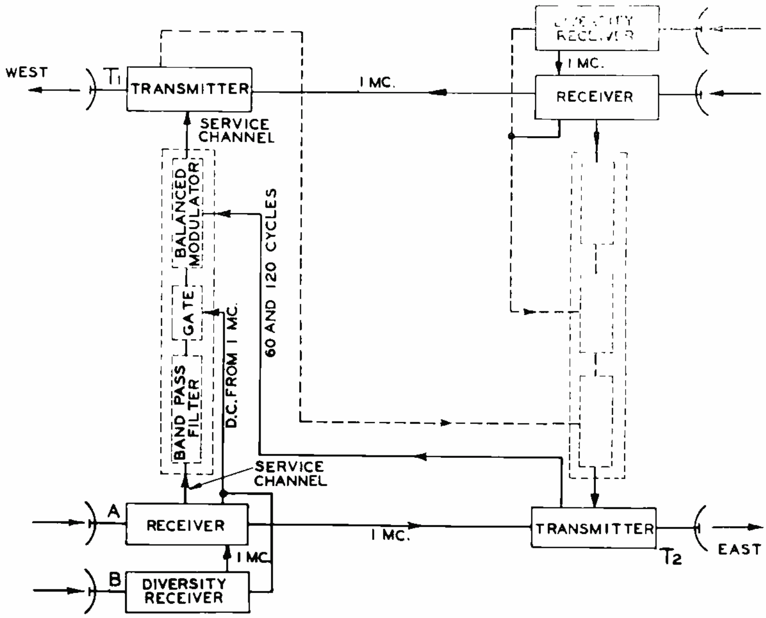

Fig. 7—Block diagram of two-way relay station showing method employed for location of relay circuit failure.

60-cycle voltage is low. If the carrier frequency of T_2 has drifted abnormally, the 60-cycle voltage is high and the 120-cycle voltage is low. Neither the 60-cycle voltage nor the 120-cycle voltage is present if T_2 fails entirely. Thus, the modulation of the identifying tones at a relay station by these 60-cycle and 120-cycle voltages is an indication of the operation of the outbound transmitter, T_2.

The fault-tone generator and fault-tone receiver provided at terminal stations for loop testing are illustrated in Figure 8. The fault-tone receiver which permits the operator to analyze the returned tones, demodulates these tones and provides a comparison of the relative amplitudes of the 60-cycle and 120-cycle voltages which modulate the identifying tone at the appropriate relay station.

When some types of failure occur (such as failure of an outbound transmitter) the operator at a terminal station can readily trace the trouble to a particular station. For other types of trouble, such as failure of a receiver on the inbound circuit, the operator can trace the trouble to only two possibilities: the first station which fails to return the tone, or the one preceding it. However, the terminal station operators at opposite ends of the circuit are usually in communication

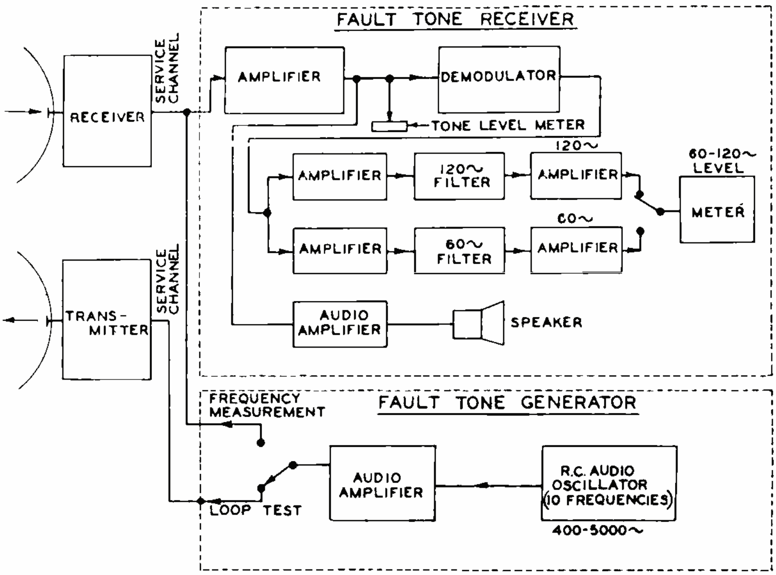

Fig. 8—Block diagram of fault-locating equipment employed at terminals of two-way relay circuits.

through the duplicate fall-back equipment, or alternate-route circuits, generally employed in telephone and telegraph systems. By working cooperatively, these operators can trace a failure to a particular station regardless of the type of failure.

Each relay station is also provided with an audio oscillator the frequency of which is the same as the identifying tone for that station. This oscillator, which is connected to the service channel, is normally inoperative but may be switched on automatically and keyed in certain code sequences to indicate abnormal conditions at the relay stations, for example—operation of emergency power supplies, failure of auxiliary equipment, illegal entry. The output of this oscillator is applied to the service channels of both eastbound and westbound circuits. Upon hearing these keyed tones in the loudspeaker of the tone receiver, the terminal operator can determine the frequency of the incoming tone by switching the output of his tone generator to the input of the fault-tone receiver, and then adjusting for zero-beat with the incoming tone. The frequency of the incoming tone identifies the relay station at which the tone originated.

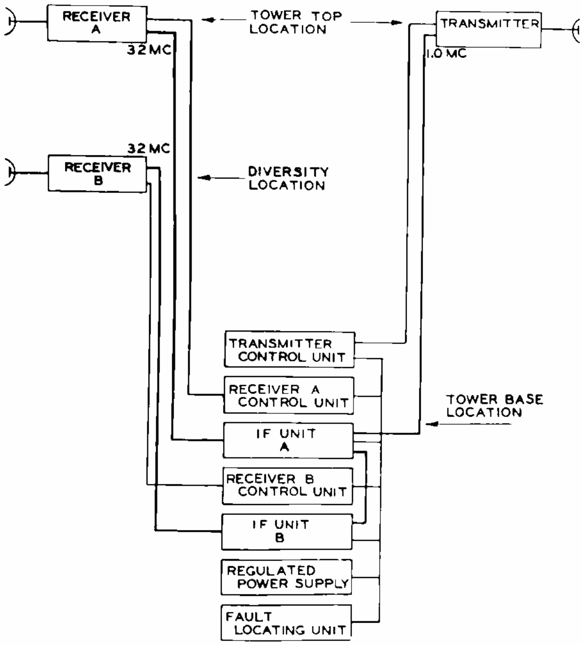

Fig. 9—Equipment arrangement at a relay station for one direction of transmission. (The equipment shown is duplicated for the opposite direction of transmission, except for the fault-locating unit which is common to both sides of a two-way circuit.)

Equipment

The equipment has been designed to permit installation of the radio-frequency head-end units adjacent to their associated antennas on towers or roofs of buildings, with the bulkier, lower-frequency equipment installed in radio rooms some distance removed from the head-ends and antennas. Figure 9 shows a typical relay station arrangement for one side of a two-way circuit. The equipment illustrated is duplicated for the other direction of transmission, with the exception of the fault-locating unit which is common to both sides of a two-way circuit. Connected to each transmitter and receiver head-end is a control unit which provides a small degree of frequency adjustment of the remote radio-frequency head-end unit, and which permits metering of the remote circuits. These control units also contain the low frequency portions of the automatic frequency control circuit. Service

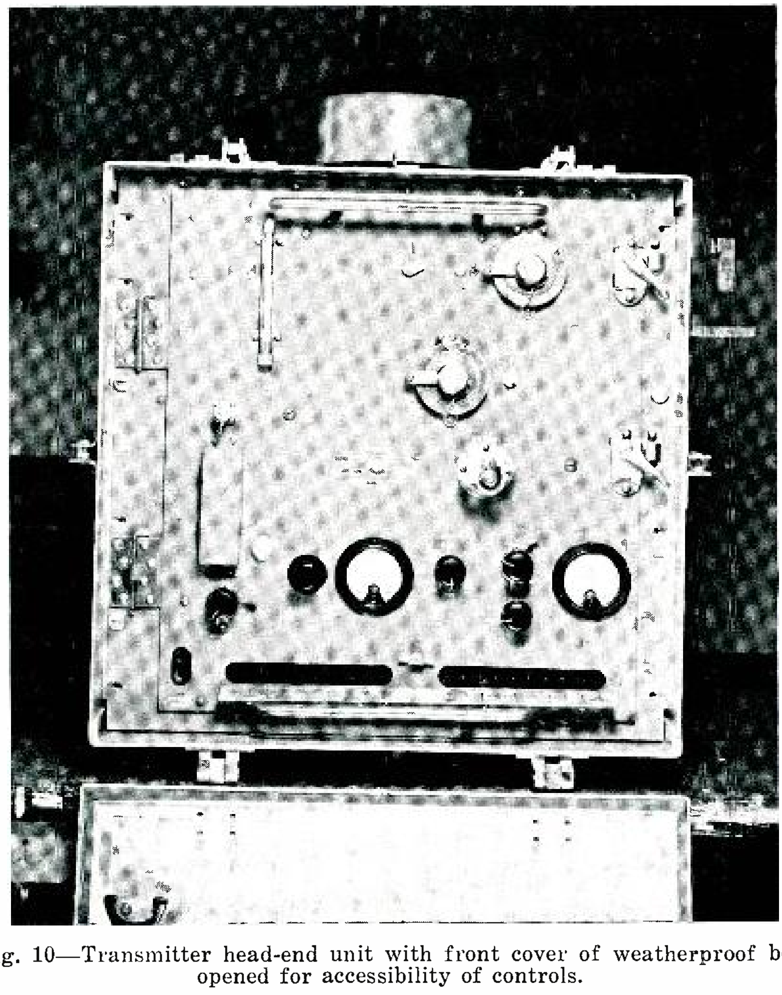

Fig. 10—Transmitter head-end unit with front cover of weatherproof box opened for accessibility of controls.

channel connections are provided in the transmitter control unit. Coaxial lines carry the 32-megacycle intermediate-frequency currents from receiver head-end units to intermediate-frequency units, and 1.0-megacycle currents from intermediate-frequency units to transmitter head-end units.

Figure 10 is a photograph of the transmitter head-end unit with the cover of its weatherproof box opened for accessibility of controls. Controls are provided on the front panel for such adjustments as that of the internal cavity of the reflex oscillator tube, reference cavity employed in the automatic frequency control circuit, coupling loop, and repeller voltage of the reflex oscillator. Metering circuits provide means for testing operation of the unit and also provide essential information on the performance of the remote lower-frequency equipments. Each transmitter head-end unit is equipped for connection of a telephone handset when service-channel communication is desired.

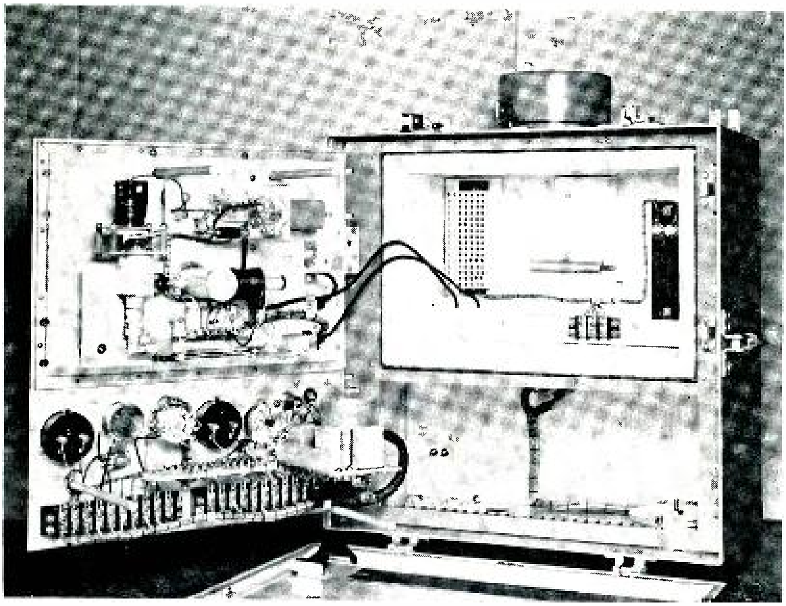

Fig. 11—Transmitter head-end with hinged front panel opened for accessibility of components.

Figure 11 shows the transmitter unit with the hinged panel swung open for accessibility of components. The type 2K56 reflex oscillator tube appears in the upper lefthand corner. This tube provides approximately 100 milliwatts output power. Its 70-ohm output circuit is coupled through the tapered line section (shown just under the tube) to a 50-ohm double-slug tuner (shown at the bottom of the plumbing assembly), to which is connected the coaxial line feeding the antenna. The second coaxial line (seen at the bottom of the box) carries the 1.0-megacycle subcarrier from the distant intermediate-frequency unit or modulator unit. Very little voltage is required for modulating the

2K56 tube; a repeller electrode voltage change of only one volt results in a frequency change of one megacycle. The coaxial cavity which establishes the reference frequency in the automatic frequency control circuit may be seen in the center of the upper assembly. When the

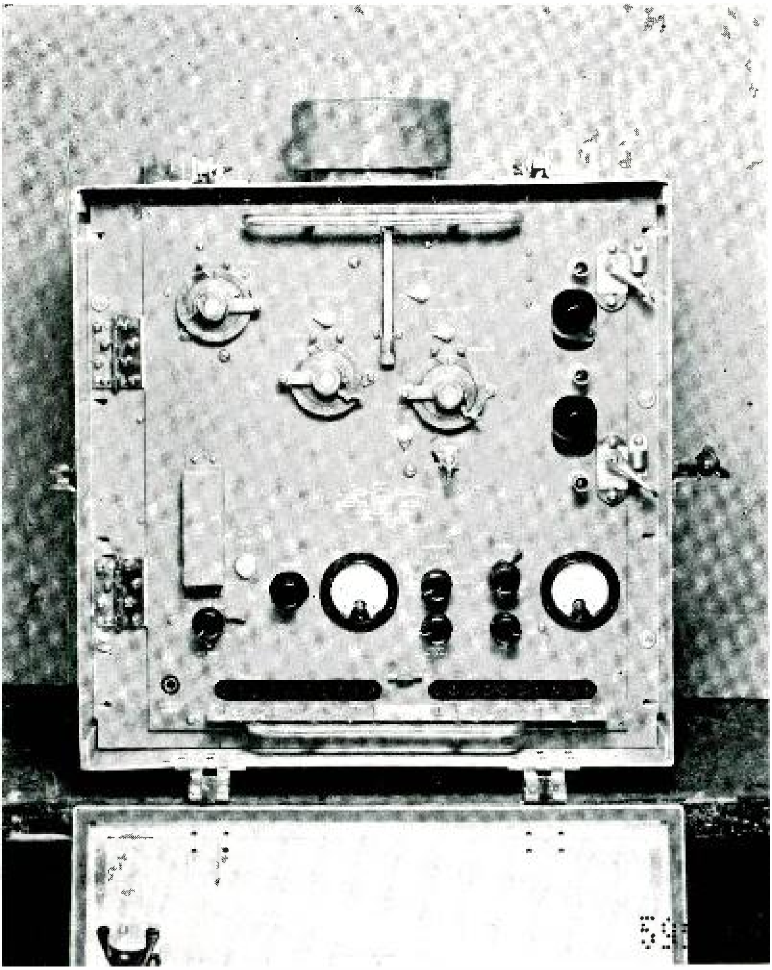

Fig. 12—Receiver head-end unit with front cover of weatherproof box opened for accessibility of controls.

panel is closed, the plumbing assembly swings into the temperature-controlled oven supported by the box.

The mechanical arrangement of the receiver head-end unit shown in Figure 12 is quite similar to that of the transmitter. A 2K56 reflex oscillator tube is also used for the local oscillator, which is controlled by an automatic frequency control circuit identical with that employed in the transmitter. The radio-frequency circuits include a preselector cavity to reduce image response and local oscillator radiation.

Figure 13 shows the ground-level equipment required at a relay station of a two-way circuit. Each rack contains, starting at the top,

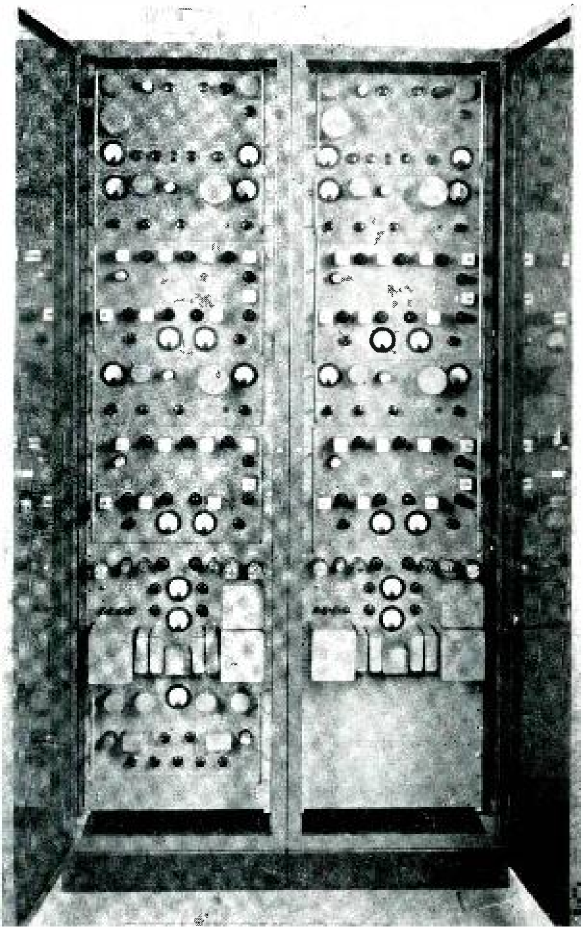

Fig. 13—Lower frequency equipment for a two-way relay station.

a transmitter control unit, first receiver control unit, first intermediate-frequency unit, diversity receiver control unit, diversity receiver intermediate-frequency unit, and regulated power supply. The fault-locating unit is shown at the bottom of the left-hand rack.

Figure 14 shows similar equipment for the terminal station of a two-way relay circuit. The transmitting and receiving equipment contained in the left-hand rack is identical with that employed for a one-way relay station. The right-hand rack contains the subcarrier

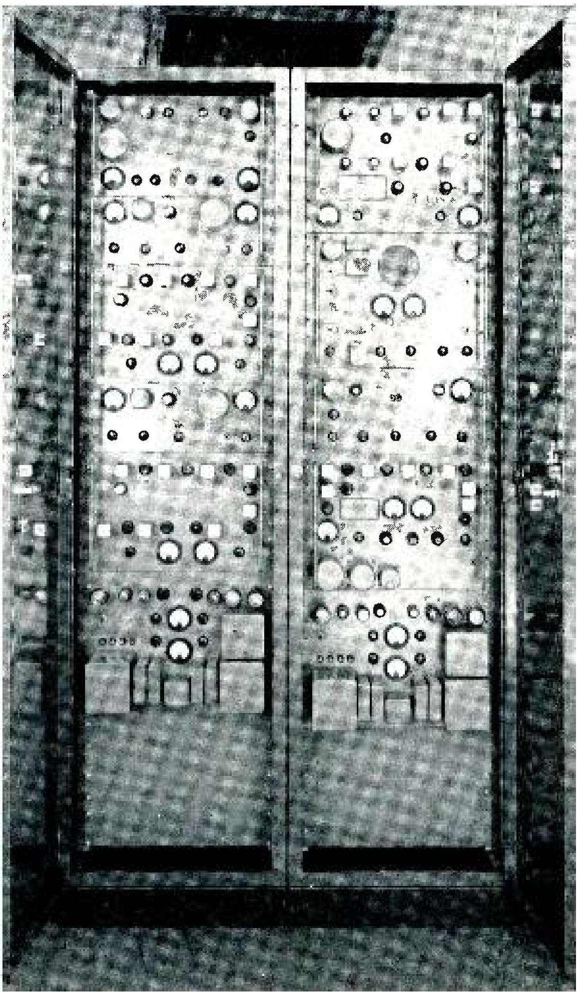

Fig. 14—Lower frequency equipment for a two-way terminal station.

modulator unit, fault-tone receiver, fault-tone generator, subcarrier demodulator unit, and regulated power supply. A rear view of this terminal station equipment, with dust covers removed, is shown in Figure 15.

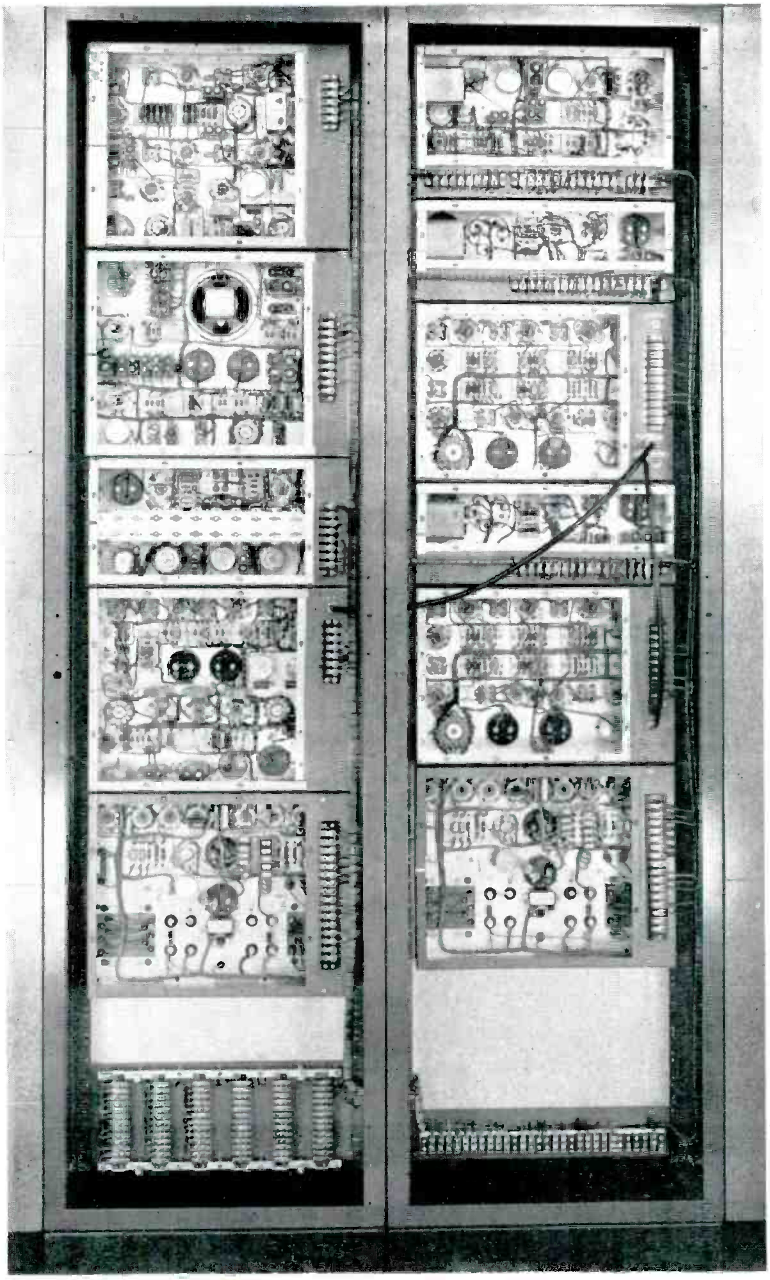

Fig. 15—Rear view of the lower frequency equipment at a two-way terminal station. (Dust covers removed.)

The performance characteristics of the equipment are summarized in Table I.

Table 1

SPECIFICATIONS

Frequency range	3900-4400 megacycles
Transmitter power	0.1 watt—approximately
Transmitter frequency stability	± 0.02 per cent—approximately
Receiver local-oscillator frequency stability	± 0.02 per cent—approximately
Receiver bandwidth	4.0 megacycles
Modulating frequency range	30 cycles to 150,000 cycles
Overall system frequency response...	within ± 1.0 decibel over range of 30 to 150,000 cycles
Antenna gain	32 decibels—approximately
Antenna beam width	6 degrees, approximately, between half-power points
Power consumption—2-way relay equipment	1400 watts
Power consumption—2-way terminal equipment	1200 watts

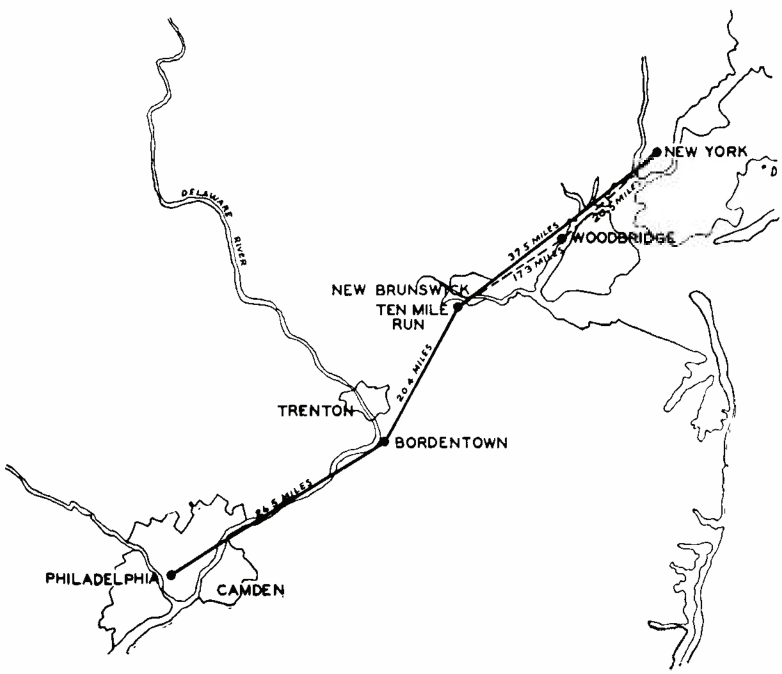

Fig. 16—Map showing locations of stations in the experimental microwave relay circuit. (The route between New York and Ten Mile Run, after installation of the Woodbridge station, is indicated by the dotted lines.)

Experimental Relay System

Figure 16 shows the station locations for an 85-mile, three-hop, two-way relay system connecting New York and Philadelphia. This system was placed in operation, using experimental models of 3300-megacycle equipment, during April, 1945, and was used for a study of propagation characteristics, and for traffic-handling tests, until February, 1946. At this time the early experimental equipments were removed and the newer types of 4000-megacycle equipment described in this paper were installed. Operation was resumed during May, 1946. A third relay station is now being constructed at Woodbridge, N. J., since earlier tests had shown transmission between Ten Mile Run and

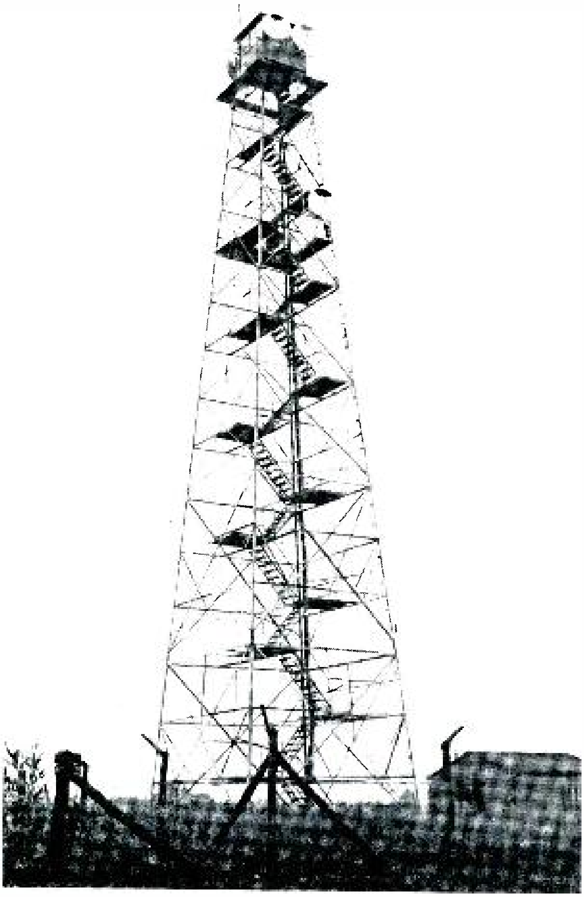

Fig. 17—Two-way relay station at Bordentown, N. J.

New York City to be affected by obstructions near the center of the path.

Figure 17 shows the tower installation at Bordentown, N. J. Installations at other relay stations are very similar. The tower, which is a standard forestry fire-watcher's type, is 100 feet high. The upper antennas and parabolic reflectors, which are four feet in diameter, are mounted on the side of the tower cabin. Two transmitter and two receiver head-end units are installed inside the cabin just behind their associated antennas. The diversity receiver head-end units and antennas are located about twenty-five feet lower on the tower.

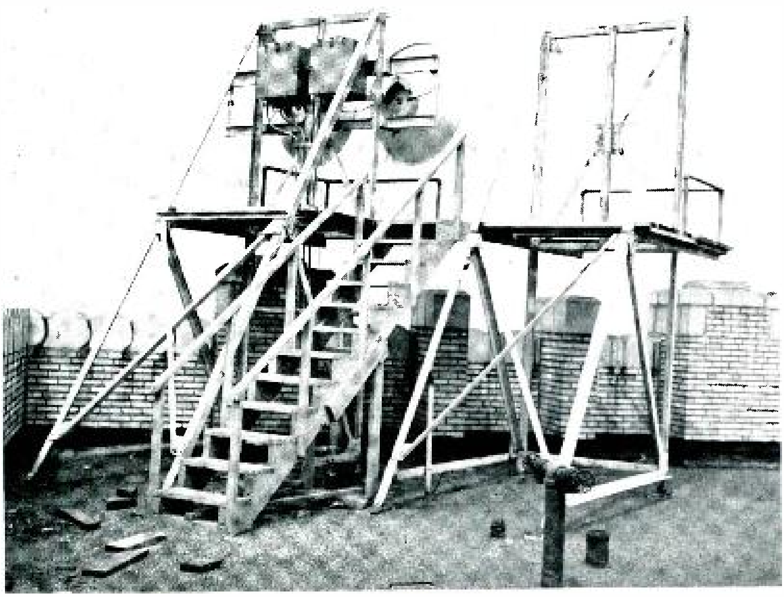

Fig. 18—Roof-top installation at Philadelphia, Pa. (The frame at the right will be used for a second New York-Philadelphia circuit.)

Figure 18 is a photograph of the roof-top installation at the Philadelphia terminal. The transmitter and receiver head-end units are mounted directly behind the parabolic reflectors. The diversity receiver and the lower-frequency equipment are installed in a small room just under the antenna supporting structure shown. This Philadelphia terminal station is connected by underground cable pairs with the Western Union office approximately a half-mile away, and multiplexed telegraph signals derived from standard Western Union frequency-division channelizing equipments are carried to and from the terminal station on these cable pairs.

The New York terminal station is installed at the Western Union building at 60 Hudson Street. The upper level receiving and transmitting antennas and head-end units are located on the roof. The diversity equipment and the lower frequency units are installed in a room on the 24th floor.

System Performance

The high front-to-back power ratios of the antenna systems have permitted the entire circuit to be operated on only two carrier frequencies. During the second test period, the transmitters at Philadelphia and Ten Mile Run operated at 3970 megacycles; the transmitters at New York and Bordentown operated at 4120 megacycles.

Continuous graph recorders were used to provide a relative indication of received signal strength. These recorders traced the variation in the grid current of the first 32-megacycle limiter tube.

Several types of fading were observed. One was a relatively slow variation, usually occurring during the night and early morning hours, which continued over a period of one to several hours. The decrease or increase from the normal signal depended on the clearance of the path above terrain. For example, the signal variations were relatively small between Bordentown and Philadelphia where the calculated line-of-sight, based on actual earth's curvature, was 100 feet above grazing. Here, the decrease from normal signal during these slow-fade periods rarely exceeded 10 decibels. Between New York and Ten Mile Run the calculated path was 30 feet above the earth's surface, but line-of-sight did not obtain because of mid-path surface obstructions such as trees and buildings. These obstructions resulted in a normal signal about 6 decibels below the calculated free-space value. Here, signal variations were large; occasional fades greater than 25 decibels from normal signal were noted, and the signal occasionally increased to the calculated free-space value. This type of fading appears to be quite similar to that observed in the ultra-high-frequency region.[9,10]

The signal over the obstructed path between Ten Mile Run and New York occasionally faded out during these periods of slow fading. Since

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[9] C. R. Englund, A. B. Crawford, and W. W. Mumford, "Ultra-Short-Wave Transmission over a 39-Mile 'Optical' Path," *Proc. I.R.E.*, Vol. 28, No. 8, pp. 360-369, August, 1940.

[10] Albert W. Friend, "A Summary and Interpretation of Ultra-High-Frequency-Propagation Data Collected by the late Ross A. Hull," *Proc. I.R.E.*, Vol. 33, No. 6, pp. 358-373, June, 1945.

it was not practicable to erect a tower at Ten Mile Run tall enough to provide the necessary line-of-sight clearance, this unreliable hop is being eliminated and a third relay station is being installed at Woodbridge, N. J.

A second type of fading, the practical effect of which was much more serious than the first, was also experienced, especially during the summer months. This was a relatively rapid fading; the signal frequently varied, over a period of several minutes, from values below the thermal-noise level of the receiver to values greater than normal.

This type of fading is illustrated in Figure 19 which shows fading of signals received at Philadelphia from June 22 to June 29, 1945. The rapidity and violence of this fading indicated multipath transmission

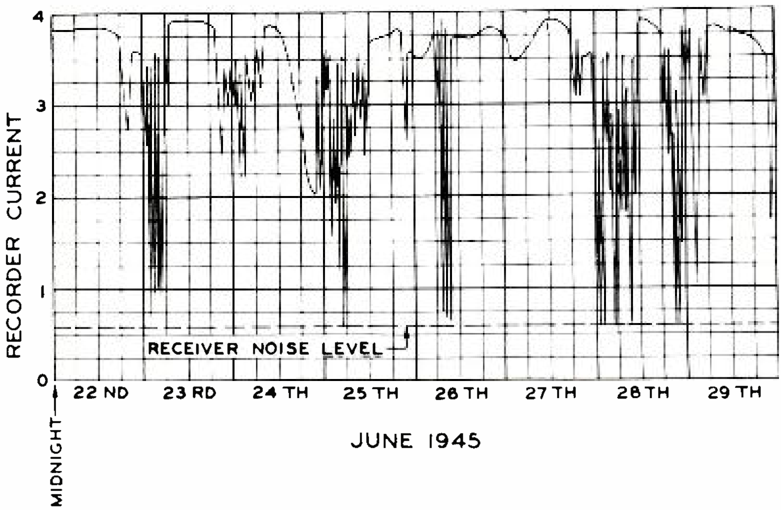

Fig. 19—Chart showing fading at 3475 megacycles on the 26.5-mile circuit between Bordentown and Philadelphia during June 1945. (Graph shows maximum and minimum signals for each hour during rapid fading periods, and average signals for each hour during the slower fading periods.)

with reinforcement and cancellation occurring between signals arriving over different paths, varying in length by multiples of one-half wavelength. Increased transmitter power was not helpful in eliminating outages caused by this type of fading. Tests between Bordentown and Philadelphia were made using an antenna input power of 12 watts, and, in these tests, fading of greater than 50 decibels from normal signal was observed.

To eliminate circuit interruptions resulting from this supposedly multipath type of fading, the system was modified to provide the diversity reception operation described earlier in this paper. A vertical separation of the antennas by about 25 feet was sufficient for this purpose since very short wavelengths (7.5 centimeters) are involved.

Figure 20 shows signal variations occurring, in the two receivers employed for diversity operation at Philadelphia, during a severe fading period. In the period covered by these records the signal in the

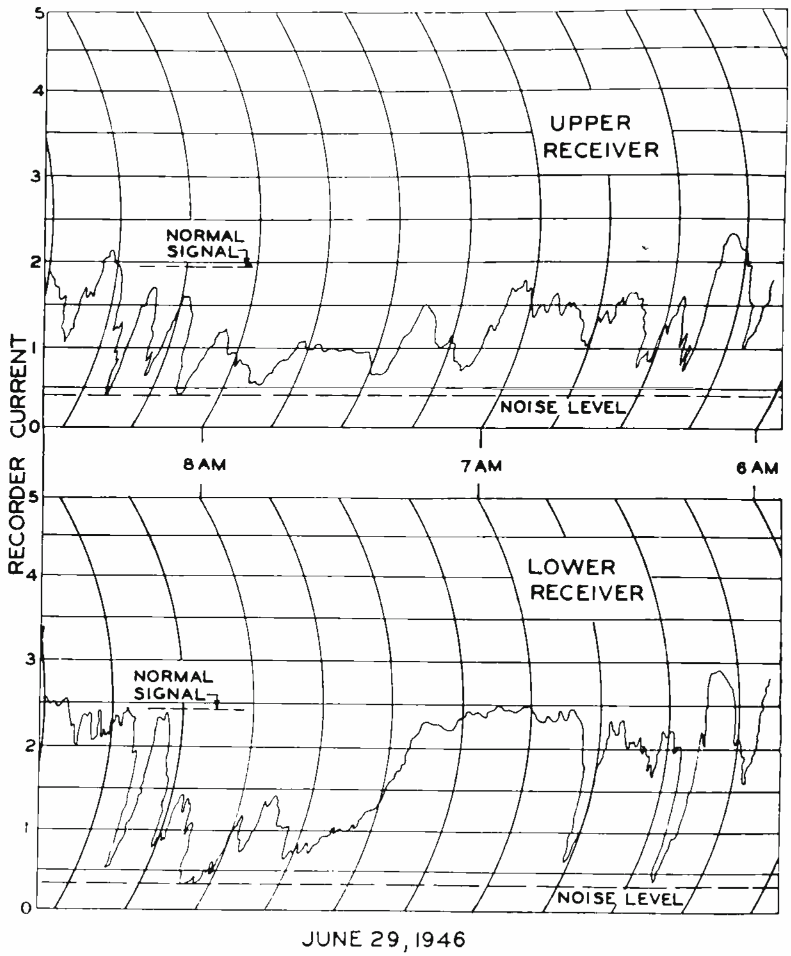

Fig. 20—Reproduction of original charts showing 4120-megacycle signal variation at the diversity-operated receivers at Philadelphia during a severe fading period.

upper receiver twice faded to the noise level, as did the signal in the lower receiver. The fades at the two receivers, however, did not occur simultaneously, and circuit operation was not interrupted. Since the time of the installation of equipment employing this type of diversity reception, no circuit interruptions caused by multipath fading have been experienced, although frequent signal fades to thermal-noise level have been observed in the individual receivers.

The performance of the relay system was entirely satisfactory when used as a telegraph system trunk line handling the New York-Washington telegraph traffic normally handled by Western Union's wireline carrier system. During these tests, the radio circuit was substituted for the regular cable pair between New York and Philadelphia. Separation filters at Philadelphia were used to connect the radio circuit (which corresponds to a 4-wire system) to the normal 2-wire land-line system connecting Philadelphia and Washington.

Western Union's Type G channeling equipment,[3,7] employed at New York and at Washington, supplied four multiplexed channels, each approximately 3000 cycles wide, in each direction. The four multiplexed channels from New York to Washington occupied the frequency band from 0.5 to 13.9 kilocycles; the four multiplexed channels from Washington to New York occupied the frequency band from 17.3 to 30.7 kilocycles. Each of the 3000-cycle width channels was divided by Western Union's Type 15 terminal equipment at the New York and Washington terminals into nine or ten telegraph channels, each 160 cycles wide. Thus, approximately 36 telegraph channels were carried in each direction. Since Western Union's mechanical time-division multiplex system[3] permits operation of four telegraph-printer circuits in each of these 160-cycle-width telegraph channels, the test setup provided facilities for the simultaneous transmission of approximately 144 messages in each direction. During the tests the actual traffic reached a peak load of 109 messages.

The signal-to-noise values shown in Table 2 were measured in the various types of channels provided by the Western Union terminal equipment under conditions simulating the traffic loads mentioned above. The test-tone level for each channel corresponded to the combined levels of the multiplexed telegraph signals normally contained in that channel.

Table 2

Channel or Block Width (Cycles)	Test Tone Level at Output of Channel (db)	Noise Level at Output of Channel (db)	Signal/Noise (db)
15,000	+ 29	— 29	58
3,000	+ 19	— 34	53
160	+ 1	— 45	46

The interchannel crosstalk occurring in Western Union's Type G multiplexing equipment is approximately 50 decibels. This crosstalk level in the multiplexing equipment was too high to permit any precise measurements of the crosstalk contributed by the microwave relay system; however, no increase in this crosstalk level was observed when the relay circuit was included in the system.

During extended test periods in which Western Union's New York-Washington traffic was handled, the performance of the radio circuit was indistinguishable from that of a normal wire-line carrier circuit. The reliability of the radio system has also compared very favorably with that of land-line facilities.

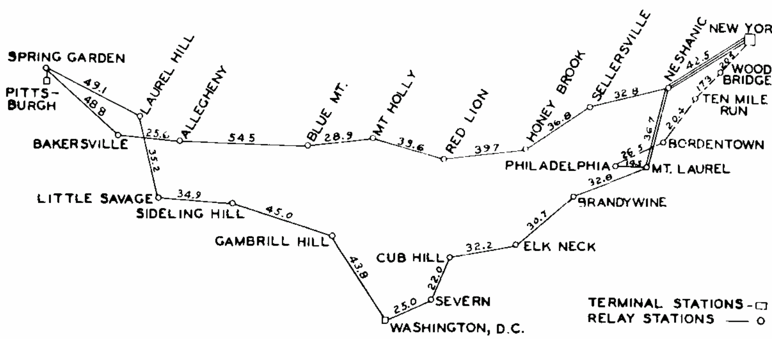

Fig. 21—Station arrangement for the Western Union Telegraph Company's microwave relay system connecting New York, Washington, and Pittsburgh. (The figures between stations indicate distance in miles.)

System Application

A microwave relay trunk-line system of the type described is now being installed by the Western Union Telegraph Company to connect New York, Washington, and Pittsburgh.[3] The station locations of this system are shown in Figure 21. A second circuit connecting New York and Philadelphia is included in this system.

The triangular arrangement of the New York, Washington, Pittsburgh circuits will provide the "alternate route" type of fall-back facilities normally employed in Western Union's land-line systems. If a failure occurs in one side of the triangle, traffic is switched to spare multiplexed channels normally provided in the other two sides of the triangle. The present experimental circuit will be used as a fall-back circuit for the separate New York-Philadelphia system.

The sites for this system were selected to provide a minimum clearance of 50 feet above obstructions such as trees and buildings with calculated profiles based on actual earth's curvature.

A total of eight radio frequency channels will be required for the four transmitting terminals and four receiving terminals located at New York. It is expected that, with the low-power transmitters, and the frequency-modulation method employed, these same channels can be used for all the anticipated future circuits which will enter and leave the New York station, provided the paths of circuits operating in the same channels can be separated by an azimuth angle of at least 45 degrees.

This New York-Washington-Pittsburgh circuit will employ Western Union's newly developed 32-voice-channel multiplexing equipment[3] which will permit the simultaneous transmission of approximately 1100 telegraph messages.

ACKNOWLEDGMENTS

Credit is due the many engineers of the Western Union Telegraph Company and of RCA Victor Division who participated in the planning, development, and testing of the equipment and system. Particular credit is due L. E. Thompson for his major contribution to the basic methods employed, and to J. B. Coleman and D. S. Bond under whose supervision the work was performed. The support and encouragement received throughout the project from Mr. F. E. d'Humy, Vice-president and Chief Engineer of the Western Union Telegraph Company, is gratefully acknowledged.

TELERAN*†

Air Navigation and Traffic Control by Means of Television and Radar

By

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Summary—*Wartime development of radar techniques offer a new approach to the problem of improving air navigation and traffic control, two fields in which existing equipment is obsolescent. Since no one military equipment appears to be ideally suited for solving the many problems, a re-investigation of the requirements necessary to handle very heavy traffic is first made.*

After analyzing the requirements, a system is described which appears to fulfill these requirements and offers unique advantages. In this system (Teleran), aircraft position information is presented to ground observers and controllers on a series of plan position indicators. One indicator is used for each altitude layer. Information on the position of aircraft in a given altitude layer is superimposed on a map of the region covered by the ground radar; this, together with weather, traffic control, and other desired information, is transmitted by television to each aircraft in the region. Each cooperating aircraft is equipped with a transponder beacon which serves not only to reinforce the radar echo but also to provide an altitude-dependent reply which allows the ground station to differentiate among aircraft by altitude.

The application of this system to concrete problems of navigation and traffic control is also discussed. Included are the problems of enroute navigation, approach and landing procedures, automatic flight and landing, collision prevention, personal identification and communication, and air traffic control.

The Air Navigation Problem

It is the purpose of this paper to describe a system for air navigation and traffic control which is built around pictorial presentation of information to both the pilot and the ground controller. It is made possible—and practical—by the combination of television and radar techniques. This system, based upon certain proposals[1] by L. F. Jones, was later amplified essentially to the form to be described in this paper with the collaboration of P. J. Herbst and Irving Wolff. Such a system is urgently needed because of the enormous increase in traffic both on and off the airways, an increase which most probably cannot

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* Decimal Classification: 629.132.5 x R583 x 537.

† Presented at the National Electronics Conference, Chicago, Illinois, on October 4, 1946.

[1] P. J. Herbst, I. Wolff, Douglas H. Ewing, and L. F. Jones, "The Teleran Proposal", *Electronics*, Vol. 19, No. 2, pp. 124-127, Feb., 1946.

be handled by existing equipment or methods of operation. It is no longer possible to add another piece of equipment in the airplane or on the ground to provide a new facility, nor simply to shift frequency to provide more reliable operation. A new approach to the over-all problem is necessary, aimed at simplification of equipment by integration of the parts into a well coordinated system. A reduction in the burden on the pilot and the ground controller is likewise necessary.

The required features of a new system will first be outlined. The techniques available for providing these features will then be examined, after which the Teleran system will be described. The requirements are:

(1) The information for navigation, traffic control, collision prevention, landing and taxiing must be presented to the pilot in a simple, natural, thoroughly understandable manner—a manner so obvious that little skill will be required for its use.

(2) All weather operation must not be restricted by the navigation system, even if it continues to be limited by the structural strength of airframes.

(3) The traffic controllers along the airways and at the airports must be provided with accurate information as to the present and future positions of all aircraft so that a smooth flow of traffic may be maintained. Efficient use of the airspace, emergency control, simplification of the controllers' work, and flexibility of routing must also be provided.

(4) In view of international language differences and the presently overloaded state of communication facilities, a new system should sharply reduce dependence on voice communications.

These four requirements dictate the use of radar. The type of radar and the amount and type of auxiliary equipment are further defined by the following additional needs:

(5) The weight and complication of the equipment carried in the aircraft must be minimized.

(6) The system should be capable of extension to provide automatic flight and automatic landing for airplanes requiring those services.

(7) The system should be capable of providing supplementary information such as weather data.

(8) Identification of individual aircraft by ground personnel should be automatic and practically instantaneous, or, better still, should be unnecessary.

(9) The system must be capable of expansion to meet the demands of continually increasing traffic density, and saturation in this respect must be as nearly non-existent as possible.

(10) It should be incapable of giving incorrect navigation information or should provide instant and positive indication of any failure in the system.

(11) When used for general navigation purposes, it should not require the constant attention of ground personnel.

(12) The system should allow pilot participation in traffic control, and provide him sufficient data to avoid collisions.

(13) The system should provide assistance in locating distressed airplanes.

One remaining requirement, not strictly technical, is nevertheless extremely important. The system must be such that its adoption will not create a chaotic condition in the aviation industry. The transition period must be orderly regardless of the facilities then in use. At no time during the transition to the new system can there be a reduction in the services available to the pilot.

Analysis of the requirements delineated above shows that radar alone is not suitable for a comprehensive system. Airborne radar equipment is relatively heavy and bulky, and requires a high degree of skill and experience for operation. Limitations on antenna size make it difficult to produce information of sufficiently high definition. Pilot familiarity with terrain and routes would still be required, even if high definition pictures were obtained at the expense of payload and drag. Most important of all, the data from airborne radar is completely con-contained within the airplane and is therefore unavailable to traffic control personnel on the ground.

Ground radar also has shortcomings. The definition and coverage can, of course, be as complete as necessary without having to consider weight or drag. The information, however, is not available to the pilot. Unless other means are provided he has no collision warning, nor can he locate himself quickly or check drift, particularly in bad weather. The solution is to send a radar picture to the airplane which has been obtained on the ground and altered in several important ways. Television appears to be uniquely suitable for sending such a picture aloft and fortunately, television know-how is available to make such transmission practical.

The Teleran System

General Description

Essentially, Teleran employs a ground search radar which surveys the air space of interest and displays on a cathode ray tube the information thus received. This radar presentation is viewed by a television camera, a map of the area is superimposed, and the combination picture is broadcast by a television transmitter. The picture is reproduced by a television receiver in the airplane and the pilot sees his plane as a spot of light moving across a map; other planes appear as different spots of light, each moving according to its actual course.

Since the received picture would be confusing if all radar echoes were displayed in the aircraft and since each pilot is primarily interested only in those aircraft at approximately his own altitude, the Teleran system includes a method of separating the radar echoes according to altitude and transmitting a separate picture (by time sharing methods) for each altitude level. This is accomplished by having each aircraft carry a transponder, which consists of a receiver

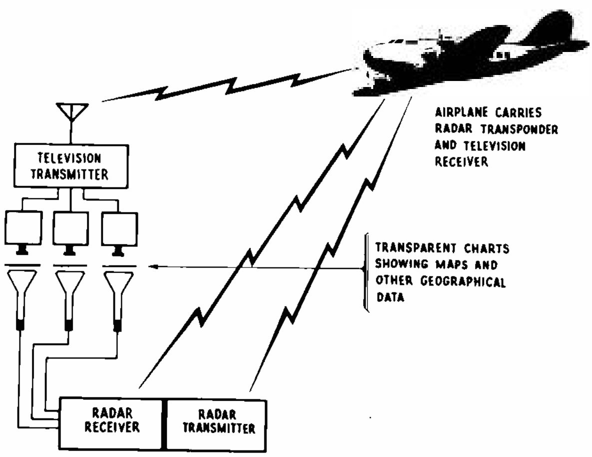

Fig. 1—Functional diagram of basic Teleran system.

and transmitter connected together so that the transmitter emits one or more pulses when the receiver picks up a pulse from the ground radar. If the transponder emits two pulses separated by a time interval which depends on the airplane's altitude, a discriminator at the ground station can be made to sort out automatically the responses according to altitude. Other methods of coding are of course possible, but the foregoing description illustrates the principles involved. These principles were first proposed by RCA in 1941 and have since been adopted in other navigation systems. The transponder also increases the range of the radar equipment, and by use of a reply frequency different from that of the ground radar transmitter, confusing radar echoes from ground objects are eliminated. Aircraft flying at low altitudes may thus be easily distinguished. The various components described above are illustrated in Figure 1.

The combination picture of map and radar indications can be produced in several ways. Probably the most satisfactory method is as shown in Figure 1, where a transparent sheet carrying the proper lines and symbols is placed between the radar tube and the television camera.

Figure 2 shows the picture seen in an airplane flying between 10,000 and 15,000 feet. Airplanes are shown by their radar pips, and the pilot's own plane is identified by a radial line passing through

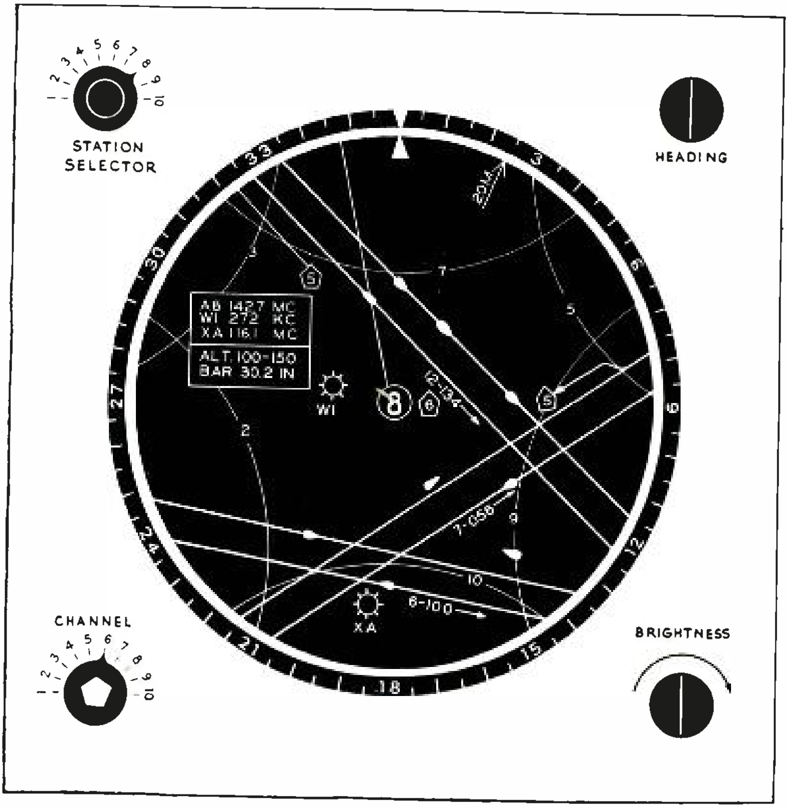

Fig. 2—Typical picture received at high altitude. (Data in large block shows altitude, barometric setting, and communication frequencies for airways station (A8), Wilkes-Barre (WI) and Allentown (XA). Large "8" in center identifies area covered by this Teleran station—overlapping Teleran areas are denoted by numbered arcs. Numbers in pentagons represent altitude layers. Central "6" indicates picture is for 10,000-15,000 feet. The "5" 's show where descending planes will enter next lower altitude layer. Wind is shown by arrow as being 20 miles per hour at 25°.)

the proper pip. Each pilot sees a different line which passes through the pip corresponding to his plane. The flight altitude, wind direction and velocity are shown, as well as the television receiver channel assigned to this particular area. The arcs near the edge of the picture show the overlapping coverage of adjacent ground radar and television stations with their associated receiver tuning. In each case the number 6 indicates the 10,000-15,000 foot altitude level. Other data can of course be added. Red lines will show the heading of the plane; they are engraved on a transparent disk in front of the television tube and rotated by a servo link with the stabilized gyro compass. The various features of the picture will be discussed in greater detail later in connection with enroute navigation.

It is proposed that the airspace up to 30,000 feet be divided into altitude levels approximately as follows:

0- 2,000 feet
2,000- 4,000
4,000- 6,000
6,000- 8,000
8,000-10,000
10,000-15,000
15,000-20,000
20,000-30,000

Other choices of altitude division can be made, but it seems desirable to provide more layers where the traffic is more dense. At least 500 feet of overlap will be available on each side of the nominal layer boundary, so that a pilot of a climbing or descending plane will not see an abrupt picture change. For example, an airplane at 2,400 feet will appear on both the 0-2,000 and 2,000-4,000 foot pictures.

It has been mentioned previously that the method of altitude layer separation used in the Teleran system is that of sorting out the transponder replies on the ground according to altitude and displaying them on appropriate scopes. This permits placing most of the pulse separation or decoding equipment on the ground, and an attendant saving of weight in the airplane. Another reason for using this method is that the pilot of an airplane may look at any other altitude layer without having to be in or near it, thereby reducing traffic hazards.

The basic system therefore requires a ground search radar, ground selection equipment to separate the transponder responses according to altitude levels, television cameras for picking up radar presentations and maps, television transmitters for sending the pictures aloft, plus a television receiver and transponder in the airplane.

This system is also useful for landing aid and for airport traffic control when suitable radars are employed. The radar stations will hereafter be called airways search radar, airport search radar, and landing radar. They will be described separately in some detail, together with the part each plays in the solution of the overall air navigation and traffic control problem. Figure 3 shows pictorially the relationships between the various radar units.

The traffic density in the Washington-New York-Boston area is the highest in the United States. In such a region the number of radar (and television) stations necessary to provide airway coverage is

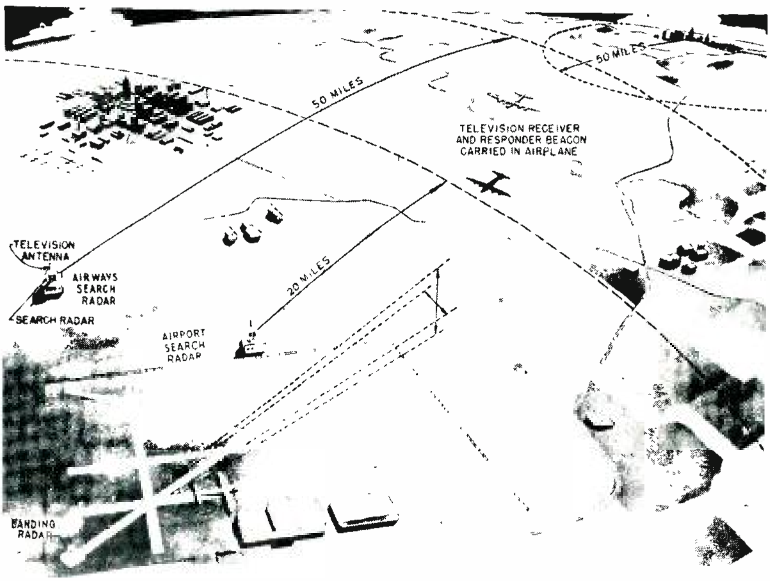

Fig. 3—Sketch depicting complete Teleran installation.

virtually the same as that required for complete coverage. Major airports are also numerous so that a large number of airport installations is justified. In less congested areas, desired air lanes can be covered by a number of stations smaller than that required for complete geographic coverage.

In addition to traffic density, the required number of television stations in an area is influenced by the desired minimum altitude coverage and terrain contours and to a lesser extent by radio frequency, polarization and soil conductivity. These factors also affect the number of frequency channels required to prevent interference. A study of the propagation characteristics of frequencies which might be used indicates that a maximum of thirteen channels would be necessary for the high-power airway television transmitters when complete geographical coverage is desired.

The airport television transmitters are not required to cover an extensive area either in distance or altitude. About eight channels should be sufficient for nearly any combination of airport spacings where large aircraft are landed.

The coverage of any one airway radar station in the Teleran system has been limited to 50 miles in radius. This might appear to be unnecessarily conservative. Even including normal refraction, however, line of sight at 75 miles requires an altitude of 2,800 feet. Low angle coverage would be rather poor, even if all planes were equipped with transponders. Fifty mile radius service area appears to be a realistic value to use in predicting coverage.

Application to Enroute Navigation

A typical picture received in an airplane flying between ten and fifteen thousand feet has been shown in Figure 2. At this altitude the topography of the ground is of little interest, and only the major features are included. The airways and their headings are displayed, together with the frequency channels to be used, the wind data and the barometric correction. Other pertinent data such as the existence of a line squall, anticipated icing regions, etc. could be added.

Besides seeing the position of all aircraft near his altitude, the pilot is able to see the direction in which they are traveling. This is due to the radar cathode ray phosphor material which causes a "trail" to appear on each signal, like a tail on a comet. This trail represents the ground track made by each aircraft for a short period preceding each observation. Other and possibly superior means for achieving this presentation are being investigated.

The radial self-identification line is obtained by televising a mechanical marker which rotates in phase with the ground radar antenna. It is normally invisible, appearing in the television picture only when the transponder is active. Since the transponder is active only when the radar is pointing at it, the radial line necessarily passes through the proper pip, and each pilot sees a different radial line which indicates his own plane. Simple circuit means are provided to insure that the line appears only in relation to the proper radar station, so that no confusion can result when the transponder is interrogated by two radars in overlapping regions.

When an airplane is flying at a low altitude, the pilot is interested in more topographical detail; he should also be informed of current airport approach paths and he requires more traffic control instructions. Figure 4 represents the Scranton-Allentown sector as shown to pilots flying in the 2,000-4,000 foot altitude layer. Approach patterns for Scranton, Wilkes-Barre and Allentown are shown for the existing wind condition. Terrain features too low to be of interest to high-flying planes but of vital interest to low-flying ones are also shown. Other

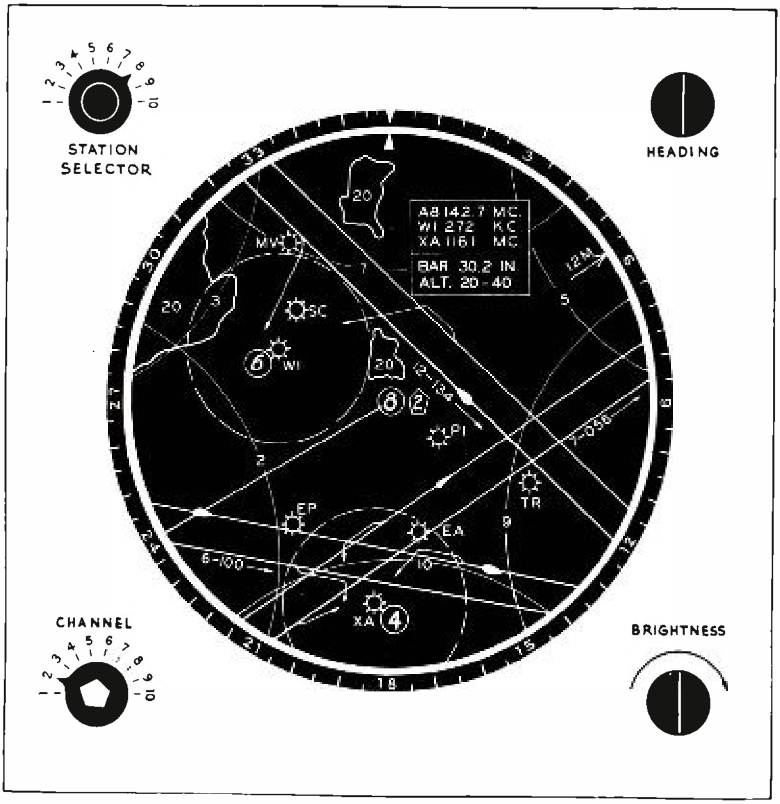

Fig. 4—Typical picture received at 2,000-4,000 feet. (Wilkes-Barre and Allentown Teleran approach control zones are shown in dashed circles, with let-down paths into those areas. Contour lines are shown for 2,000 feet.)

data may of course be added whenever the need arises, to be erased or removed when no longer necessary.

Automatic Flight by Teleran

Although the need for automatic flight is less urgent than for some other types of service, Teleran would be much less useful if this service were not easily available for both on and off airway flight. The data needed are range and azimuth of the aircraft from the station. These data, together with the coordinates of the destination with respect to the station, are fed to a computer, whose output goes to the automatic pilot.

In Teleran, the azimuth information is obtained by transmitting an omnidirectional pulse from the television transmitter at the instant the radar antenna is pointing north. Azimuth is measured by electronically counting in the airplane the time interval between reception of the North pulse on the television channel and that of the direct interrogation pulse from the radar beam. The azimuth information so

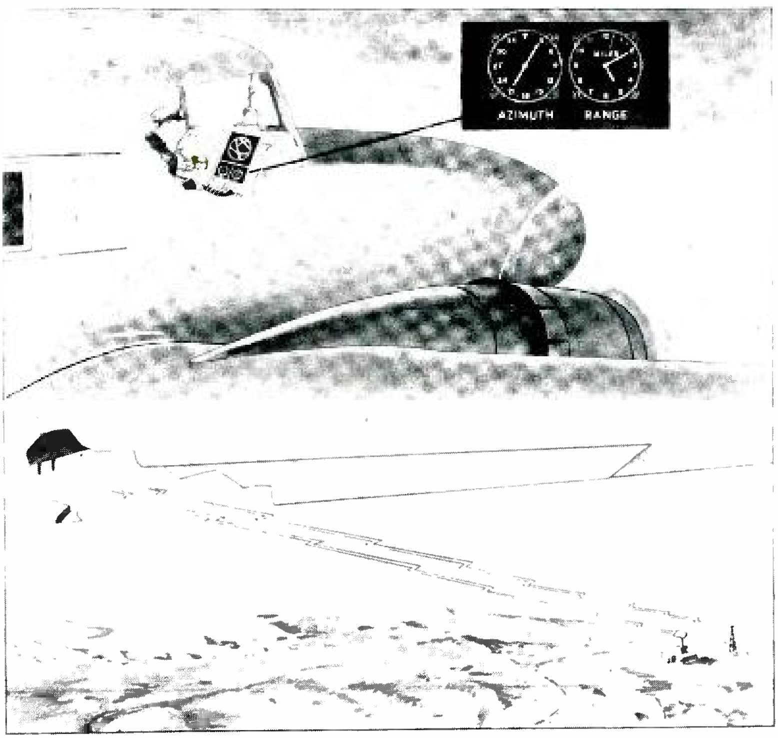

Fig. 5—Azimuth and distance meters. (No additional frequency channels or airborne receivers are required.)

obtained supplements that obtained directly from the television picture, but is in more usable form for automatic flight and can be put on an azimuth meter, if desired.

Distance information is obtained by using the airplane's transponder as an interrogator a short time after it replies to the ground station. The interrogating pulses are received on the ground and sent back on the television channel affording a measurement of distance to the station. Figure 5 illustrates these functions.

Airways Traffic Control

The need for revision in the method of controlling air traffic is caused by the constantly increasing number of airplanes. Control equipment, however, is not easy to develop because the problems to be solved are concerned with characteristics of airplanes and their operation, which are subject to considerable change.

Although numerous and widely different methods of accomplishing traffic control have been proposed, the same essentials appear in all of

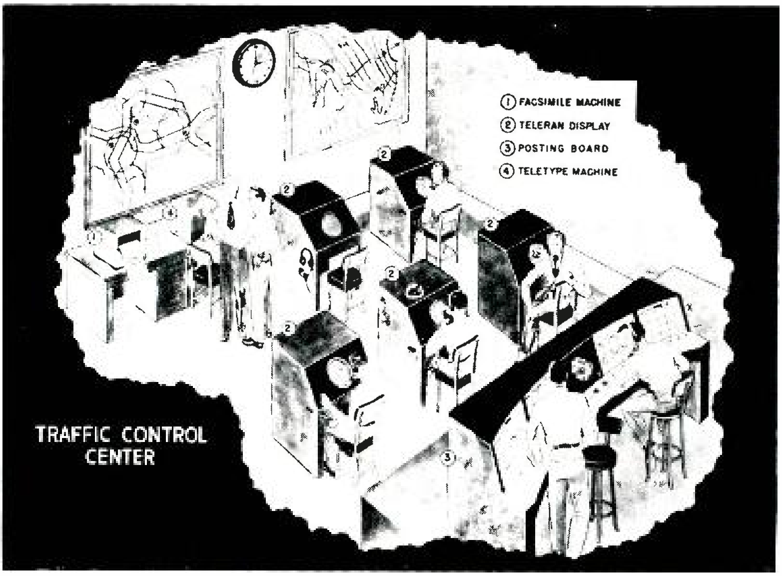

Fig. 6—Traffic control center with a number of Teleran indicators. (Each operator can switch to picture of other altitude layers or adjacent Teleran areas. Other arrangements of personnel and equipment are feasible, and would be determined by traffic density. For areas having dense traffic, computers and other accessory equipment can be added.)

them. Each system includes elements to gather information continuously as to the exact location of airplanes, usually in terms of azimuth and distance with respect to the control station. Each system also provides some means of analyzing the data which has been received to determine whether conflicts will occur. Posting the data helps in the analysis, and may be done automatically. While exact identification of each aircraft is not necessary, it is useful to tag each one in some way for reference purposes. It is also helpful to know the intended destination of each aircraft.

In the Teleran system, the location of all airplanes in the area is obtained by ground radar. This method of obtaining the information is inherently safe, because the airplane's presence will be taken into account even if the airborne equipment fails. Also, planes without transponders may be included in the traffic pattern.

With the coordinates of each airplane known, the progress may be checked visually to avoid collisions. In dense traffic areas, where tight control is necessary, the individual airplanes may be tracked and the data fed to a computer. Figure 6 shows an arrangement of an area traffic control center in a dense traffic region.

In one type of traffic control system an airplane progressing cross-country or along an airway may be considered as moving through successive blocks of airspace; in another, it is considered to be surrounded by a moving block of space. Teleran is applicable to either of these divisions because both pilot and traffic controller see the limits of the block, and failure to keep inside a block is immediately apparent to them both. If the blocks are fixed they would probably be drawn on a map of the region. If the block is to move with the airplane, the block limits can be made to move by means of the cursors on the traffic control PPI (plan position indicator) scope. The block may be any shape and can be made larger or smaller without difficulty, thus giving high-speed airplanes proportionately greater latitude.

The "Command" function of Teleran may be used to control an aircraft in an emergency, to provide special information to a particular airplane, and for identification purposes. The ground equipment consists of coder and narrow beam radar equipment of the "searchlight" type and a relatively simple decoder. When a controller wishes to "Command" an aircraft, he adjusts a pair of crosshairs to intersect on the appropriate radar pip. This action rotates the antenna beam in azimuth and effects a very precise range gate. Control is therefore limited to the airplane in question. The controller may then obtain identity by pushing a button causing the command interrogation to be coded so that this aircraft replies automatically with an identification signal. It is also possible to have the airplane execute one of several maneuvers by similar coding methods, which show in the cockpit as colored lights or illumination behind small windows on which are printed the maneuver to be performed.

It is anticipated that the "Command" feature will be helpful in routine traffic surveillance. As each airplane reaches the overlap region of the next station ahead, it can be interrogated, identified, and accounted for in the traffic pattern.

In addition to the features outlined above Teleran is capable of accomplishing variations in patterns and transmission of information not possible in other systems. Any number of parallel courses can be established at will and flown in safety because the position of all aircraft can be monitored both in the plane and on the ground. Since a new course can be established merely by drawing it on a map, the

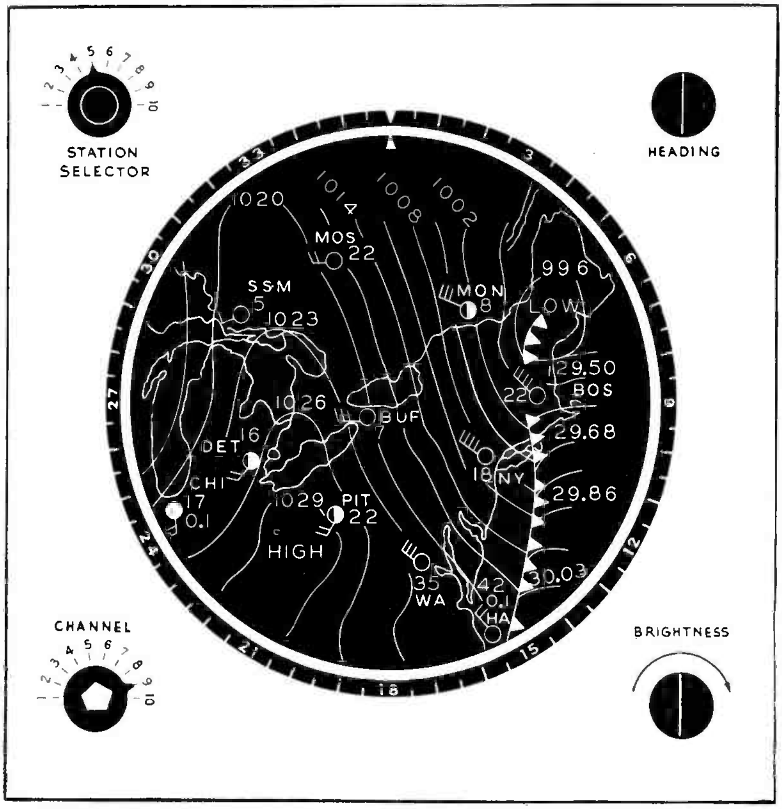

Fig. 7—Typical weather map, constantly available to pilot by switching to weather channel. (Area covered includes several Teleran zones, thus providing advance information.)

system provides a maximum of flexibility to accommodate future conditions.

The prompt transmission of complete weather information is a unique feature of Teleran. Figure 7 is an example of the information which can be included. By a time scheduling arrangement, the weather in other areas of interest may be substituted for the region in which the plane is flying. The information is of interest not only for safety reasons, but as a definite aid to traffic control. Advance weather information available in the air will reduce congestion at airports where the weather is changing. It should also be of value in avoiding unsafe areas such as line squalls or icing conditions, when such regions are accurately located.

It will be appreciated that there will exist regions over which adequate coverage of 50-mile radius will be impossible because of terrain. In these cases additional equipments must be sited so as to fill the gaps.

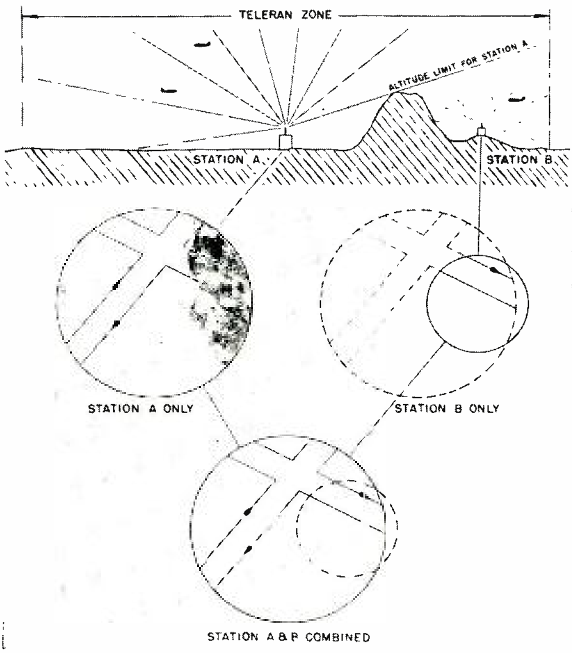

Fig. 8—Diagram illustrating how shadow areas may be filled in by combining displays from one or more auxiliary radars. (This is most easily accomplished by television. Similar techniques are useful in combining displays in traffic control centers.)

Data from all the equipments in a given area can be easily combined into a single picture which can then be used by control personnel and aircrew alike. Figure 8 depicts a situation of this sort.

Application To Airport Approach

The radar-television methods described above are ideally suited to the control of traffic approaching or departing from an airport. For this purpose, the radar range may be greatly reduced since the airport surveillance need not extend beyond the distance necessary to maneuver the planes into the final approach pattern. As previously suggested, a radius of about twenty miles would probably be adequate for airport approach control. The presentation transmitted aloft is illustrated in Figure 9. Additional information has been added to that in other pictures to allow instrument approaches under most conditions.

As in the case of airways traffic control, the airport controller is

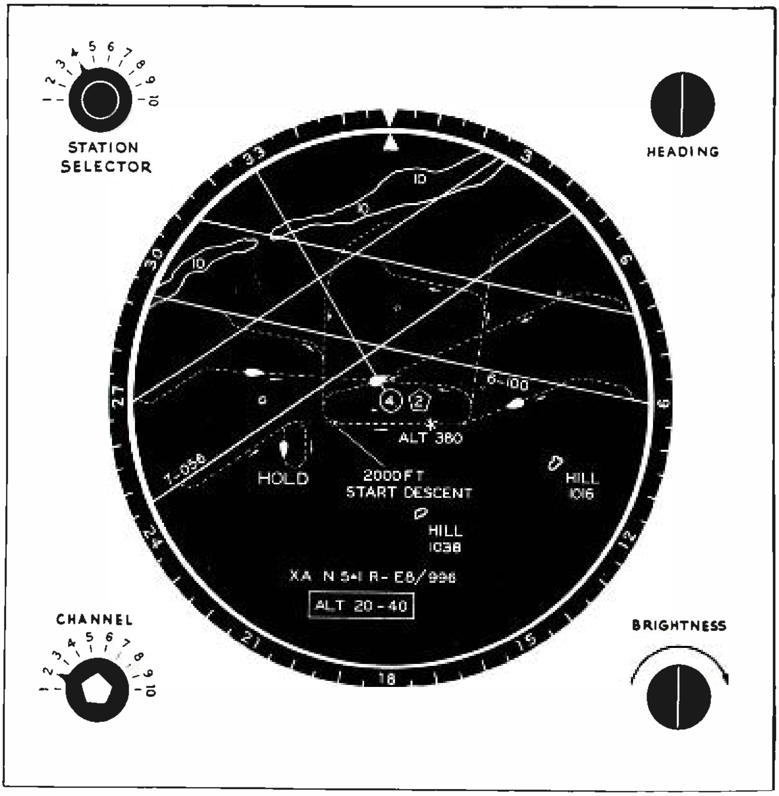

Fig. 9—Typical approach zone Teleran picture. (The approach paths will accommodate relatively dense traffic. Holding courses or new air paths can be added and closely monitored by ground controllers as need arises. Techniques available permit information to be added as quickly as it can be pencilled on the chart, and subsequently erased if desired.)

provided with enough information and flexible communication means to control traffic adequately. For instance, if the controller finds that two airplanes would arrive at the airport simultaneously, he can establish a holding course for one of them simply by drawing a line on the chart. The line may be erased when it has served its purpose. By procedures of this type he is able to establish an orderly flow of traffic within the airport control zone, and congestion can largely be eliminated. Figure 10 shows an operational set up in an airport control tower.

Traffic instructions can be given much more rapidly and with less chance of misunderstanding by means of the television picture than by voice communication. Data of interest to all pilots and now requiring individual voice communications—wind, ceiling, visibility and runway in use—are transmitted as a part of the Teleran picture. Special messages can be sent to individual planes by pointing to the proper pip on the indicator and transmitting the message either verbally over the communication channel or graphically over the television channel, or via the "Command" function.

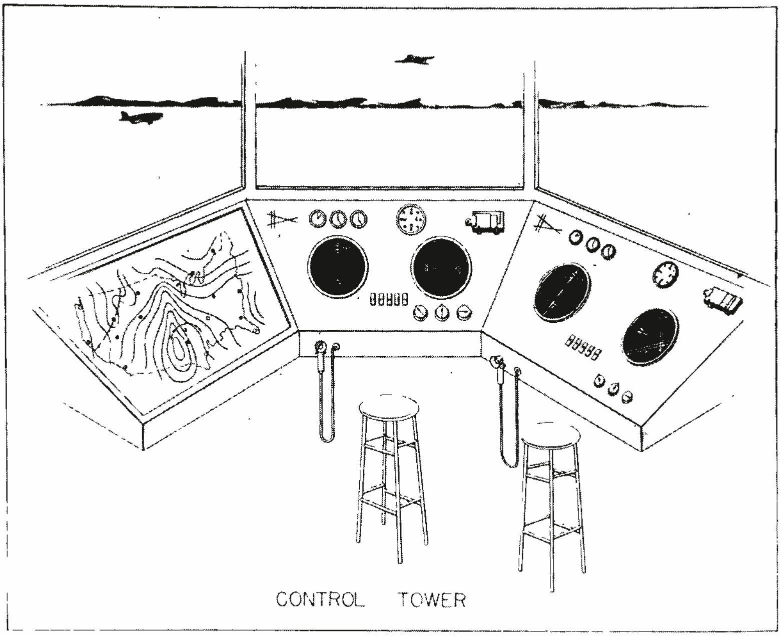

Fig. 10—Control tower installation of Teleran showing approach and landing displays. (Simplicity of display enables control tower operator to provide "talk down" landing service.)

Effective coordination between airways traffic and airport traffic controllers can be provided by the use of relay links between them, thus giving advance notice of a possible emergency. In the event traffic flow toward a particular airport became great enough to warrant holding or stacking procedures, they could be instituted with minimum inconvenience to controllers and pilots. Should this traffic flow exceed the handling capacities of the airport, the airways traffic controller can direct planes quickly and safely to other airports by drawing temporary courses on the chart. These courses can be cross-country and by the most convenient route.

Since all information relative to the airport approach pattern is given pictorially it would normally not be necessary for pilots to use airport facilities charts when entering unfamiliar areas, nor to use airways maps even for extended flights. The prompt dispersal of aircraft after landing, and the general policing of the ground area adjacent to the runways is a necessity for efficient traffic handling. There is reason to believe that radar may be developed which is capable of performing this task, although it is admittedly one of the most difficult to design. When such equipment becomes available, be it radar or of some other type, its integration into Teleran is simple and logical, as the presentation may be made available to the pilot as well as the control tower operator.

Instrument Approach and Landings

The Teleran system includes means of allowing aircraft to complete their landings on instruments. The type of display is rather unique, as shown in Figure 11. A vertical line representing an extension of the runway to be used shows the pilot when he is on course, and mileage marks along the line indicate the distance to the airport. A horizontal line shows the plane's position with respect to the correct glide path; that is, when the plane is at the correct altitude the line passes through the pip. If the plane is too high the pip is above the line, and vice versa, as the line and the pip move toward the airport. This line is produced automatically, without attention from ground or air personnel. Course softening, a feature recently added to beam-type landing systems, can be included in Teleran to aid in keeping control of the aircraft during the last few seconds before touchdown. Unlike any other instrument approach display, the Teleran display presents all aircraft which are on the final approach path.

This type of presentation is similar to the well-known "crossed pointer" indication, but has several advantages over it. In Teleran the distance information is continuously and automatically displayed, eliminating the need to glance away from the picture to read a distance meter. The indication for the glide path has the same sensing as an artificial horizon, so that a natural presentation is produced. The size of the picture is large and has negligible parallax, eliminating a need for two crossed pointer meters. The pilot sees the positions of all other planes on the landing path as well as his own, and can therefore make his let-down with complete confidence. The lines representing the plane's heading refer to the directional gyro. In a crosswind they help establish the correct heading during descent and allow a direct check of the amount of correction to make at the moment of touchdown. This

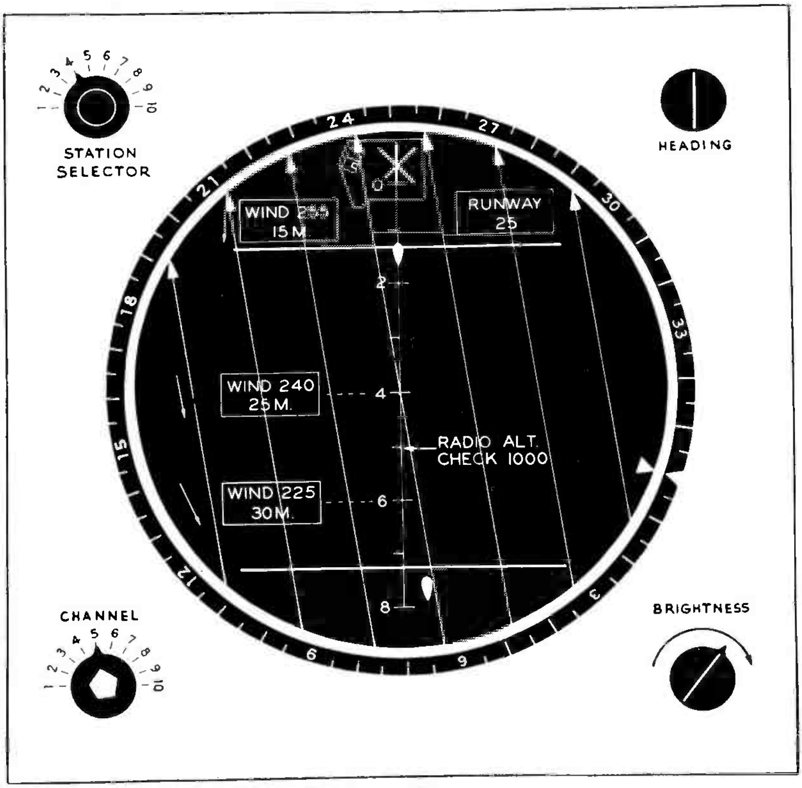

Fig. 11—Picture received during final approach. (Position with respect to proper let-down path and the distance to airport are shown continuously and in convenient manner. The pilot does not change his mental concept; his aircraft is still a spot moving on a map. He can also see planes ahead of and behind him.)

is important because the amount of crosswind usually varies during the letdown procedure. The landing radar used to provide this data would be similar to the azimuth and elevation sector scan radars of the Ground Controlled Approach (GCA) system. Since the search radar and communications gear of the GCA are not necessary for Teleran landings, the size and complexity of the ground equipment is greatly reduced. Furthermore, no operating personnel would be required. The Teleran landing radar transmitters, receivers and antennas would be installed at the end of one or more runways and data relayed to the control tower. In the event that a plane is not equipped with a television receiver, the equipment can be used as a "talk down" system. However, the presentation is so simple that a single observer can control an airplane during the letdown.

It is expected that automatic landings will ultimately be a requirement of any complete air navigation system, and provision for them is possible by adding a relatively simple modification kit to the Teleran equipment. The addition will not interfere with the television picture transmitted for manual landing. Even if other means were to be used

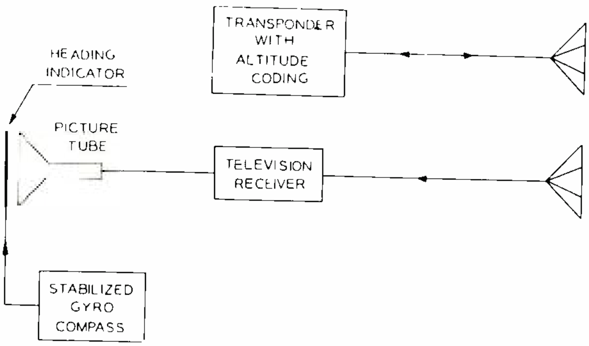

Fig. 12—Block diagram of airborne equipment for Teleran. (Weight is approximately 90 pounds, and undoubtedly will be less after further development.)

ultimately for automatic landing, Teleran provides a valuable monitor on the operation of the system, showing a pilot his progress and indicating the positions of other aircraft on the same landing path.

Airborne Equipment

Only three main controls are normally needed to operate the equipment. These are the station selector, channel selector, and the brightness control. Focus, centering, synchronization, and similar controls which need be set only at long intervals, and then only during pre-flight check, are mounted on a covered countersunk panel at the bottom of the receiver. Tubes capable of high brilliance and contrast make cockpit installation practical as used in television presentation.

A block diagram of the necessary airborne equipment is shown in Figure 12. It is estimated from experience with radar and television equipment designed for the armed services during the war that the complete airborne equipment can be designed to weigh under one hundred pounds. Further development will, of course, reduce this weight.

Planes Without Transponders

In the case of aircraft not equipped with transponders, or in cases in which the transponder has failed, the ground search radar still is able to observe the position and movement of the plane by means of the radar echo instead of the transponder response. Provision is made in Teleran for separating such echoes from the transponder responses of other aircraft and displaying them exclusively on a separate indicator. Since the radar echoes do not indicate the altitudes of the corresponding planes, echoes from non-beacon aircraft at all altitudes will be displayed on the same indicator. Depending on the menace that such aircraft might present to other aircraft, this data could be transmitted to some or all altitude levels automatically or it could simply be used by the controller to warn other aircraft by voice or written message. In situations in which the number of aircraft without transponders is large, an auxiliary height finder radar could be provided.

Small Aircraft

While the services in the Teleran system are comprehensive enough for large transport plane use, the needs of smaller aircraft have not been overlooked. Since the Teleran receiver and transponder will probably weigh less than 100 pounds, it will find use in small executive transports. Small single-engined craft will at least temporarily be denied the use of all of Teleran's many services.

A little reflection will show that this apparent shortcoming is by no means an objection to the acceptance of Teleran as a standard system of air navigation. Light airplanes may appear in the near future which are adequate for instrument navigation. In the meantime, it is possible to use portions of the Teleran system to supply basic information such as azimuth and distance from a station. Simple equipment for this purpose can be built which is light enough for even the smallest light airplane and without adding to the ground equipment which is already contemplated for Teleran. Light aircraft would also have use of the auxiliary "talk down" low approach system which Teleran provides, and their presence in a traffic pattern would not create a hazard in spite of the fact that they were not carrying a transponder or other special equipment. Voice communications would be carried on by the usual means.

CONCLUSION

By combining the information gathering capabilities of radar with the tremendous intelligence transmitting capacities of television, Teleran will provide information for the safe and rapid navigation, traffic control, and instrument approach of large numbers and types of aircraft. Teleran may be called complex. It does use considerable equipment, but its complexity is almost negligible compared to that of the problems which it will solve. Its ultimate cost can only be measured in terms of the service it offers.

MICA WINDOWS AS ELEMENTS IN MICROWAVE SYSTEMS*†

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***Summary**—The design of a virtually reflectionless, vacuum-tight window made of mica for use in a waveguide system is described. The technique of manufacture and the experimental results with a number of models are given. Such mica windows have many applications but are particularly useful for the transmission of microwave power or electro-magnetic radiation in particular portions of the spectrum.*

A VACUUM OR AIRTIGHT window is often needed in the design of microwave systems and components. Examples are the windows of waveguide-output magnetrons, T-R (transmit-receive) boxes, and pressurized waveguide systems. In the past, it has been a common practice to make such windows of glass, sealed to a suitable metal. In some cases, however, particularly where high power operation is involved, the radio-frequency losses are excessive, due to the types of glass which can be employed. Consequently, an alternate means for the transmission of power through a window is desirable. A solution to this problem was found by using mica as a window material and sealing it to metal or glass by means of a glass bonding material. This method greatly reduces the power transmission difficulties and at the same time, in certain cases, results in constructional simplifications.

There are many applications of this technique. Among them are:

1) Windows for waveguide-output magnetrons
2) Windows for T-R and anti-T-R boxes
3) Vacuum or pressure seals in coaxial lines
4) Vacuum or pressure seals in waveguide systems
5) Windows for optical systems or phototubes requiring special spectral transmission characteristics
6) Windows for X-ray tubes.

This paper deals primarily with the theory and techniques con-

* Decimal Classification: R310.

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cerned with the use of windows for waveguide-output magnetrons, although the methods described here can be adapted very readily to other applications such as those listed above.

Waveguide-Output Magnetrons

With magnetrons operating at short wavelengths, the use of a waveguide-output necessitates that a low-loss vacuum-tight material be placed somewhere between the output cavity of the magnetron and the external plumbing. In practice this material has been a disc of Corning 707 low-loss glass sealed to a kovar cup soldered to the waveguide. (See Figure 1) In general, this has constituted a satisfactory method for the transfer of radio-frequency power from the magnetron, but several difficulties and disadvantages inherent in the glass windows

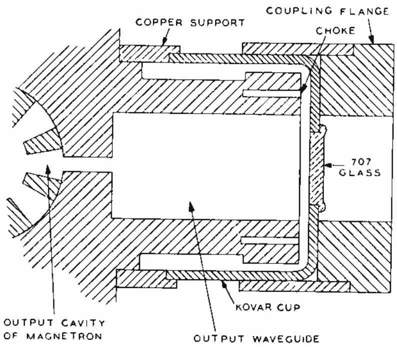

Fig. 1—Cross-sectional view of a magnetron employing output waveguide.

make it desirable to seek some other solution. The first (and perhaps least important) disadvantage associated with glass windows resides in the mechanical problems involved in making glass discs and sealing them in place so that the reflections from the window are not excessive. It is the practice to set an upper limit, usually 1.1, upon the voltage standing-wave ratio introduced by the window in a matched-waveguide system. A more serious disadvantage is that, on occasion, windows "suck-in" during operation at high-output powers as a result of dielectric losses. Since no type of glass is available at present with losses substantially lower than that of Corning 707, mica was considered as a possible alternate material. As the losses in ruby mica in the microwave region are only about 20 per cent of the losses in 707 glass, the use of mica should result in considerably greater power outputs in waveguide-output tubes. Further gain in power handling is possible because the mica windows can be much thinner than glass. As will be described below, considerably improved broad-band characteristics are also obtained. Because the possibility of making vacuum-tight seals of mica to metal[1] had previously been demonstrated, it was only necessary to determine whether the existing technique could be applied to obtaining a solution that was electrically satisfactory.

THEORETICAL CONSIDERATIONS

An understanding of the principles underlying mica-window design can best be arrived at on the basis of impedance considerations involving the use of the circular impedance chart. The mica window is, in essence, an assembly inserted into a waveguide of characteristic impedance Z_0. It is desired that the impedance looking toward the mica

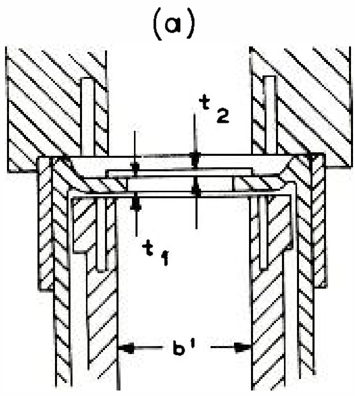

Fig. 2(a)—Single-cup assembly.

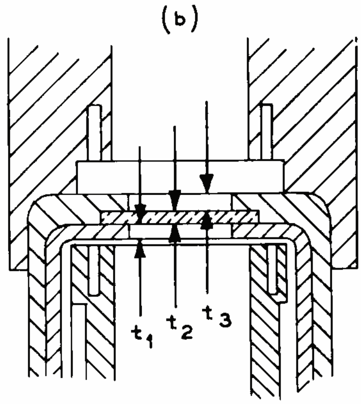

Fig. 2(b)—Double-cup assembly.

window in either direction of the system be Z_0. (It is assumed that the waveguide system is matched before the introduction of the window.) The introduction of the mica window (together with any necessary matching device) will cause no impedance change at any point in the system, provided the window produces no reflections.

Two types of window assemblies were studied. These types are referred to as the "single-cup" window illustrated in Figure 2(a), and the "double-cup" window illustrated in Figure 2(b). While the considerations of this paper are limited for mechanical reasons to circular holes, at times it would be more desirable to use apertures with other shapes.

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[1] J. S. Donal, Jr., "Sealing Mica to Glass or Metal to Form a Vacuum-Tight Joint," *Rev. Sci. Instr.*, Vol. 13, No. 6, pp. 266-269, June, 1942.

SINGLE-CUP ASSEMBLY

The single-cup assembly, Figure 2(a), is considered first. Figure 2(c) is a projection showing the dimensions of the main waveguide (major dimension b, minor dimension a) and the diameter (b') of the cup hole. The thicknesses of the cup and mica are indicated by t_1 and t_2, respectively. The window assembly is essentially a complicated combination of series and shunt elements connected in a waveguide of uniform cross-section. The window assembly may thus be considered as consisting of two short waveguide sections in series, the first having a vacuum dielectric and the second a mica dielectric. The vacuum section will be designated as section 1 and the mica section as 2. Section 1 has a height a and a width which, though not uniform, is, in practice, so close to b' (the hole diameter) that b' may be set

(c)

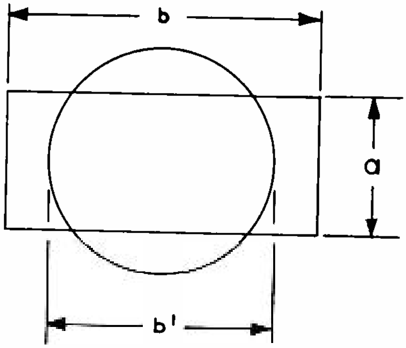

Fig. 2(c)—Waveguide dimensions.

for its magnitude. This value implies that the cup-hole diameter b' is always greater than a, which was the case in this experiment. While the dimensions of section 2 are less clearly defined, it is assumed that the waveguide choke will make the effective dimensions of section 2 the same as section 1. These approximations may seem rather liberal, but since "end effects" are completely neglected, considerable liberties in other respects may well be tolerated. In any event, theory alone was not expected to furnish a complete window design, but was employed to indicate trends and to explain qualitatively the actual experimental results.

Figure 3 is a schematic representation of the single-cup assembly. Z_1 and Z_2 are the characteristic impedances of sections 1 and 2, respectively. Since $b' < b$, it follows that $Z_1 > Z_0$. Furthermore, although this same condition holds in section 2, the presence of the mica results in $Z_2 < Z_o$ for values of b lying between $\frac{\lambda_o}{2\sqrt{\epsilon}}$ and $\frac{\lambda_o}{\sqrt{\epsilon}} = \lambda_m$ where λ_o is the free-space wavelength of the oscillation, ϵ is the dielectric constant of the mica, and λ_m is the wavelength in "bulk" mica. This follows from the equation[2]

$$Z_{eq} = \sqrt{\frac{\mu}{\epsilon}} \frac{1}{\sqrt{1-(\lambda/2a)^2}} \frac{b}{a}.$$

In the circular impedance chart of Figure 4, let point A be the characteristic impedance Z_o of the main waveguide normalized with respect to Z_1. It is thus the "normalized" impedance (with respect to Z_1) which is seen when one looks to the left of a plane through α of

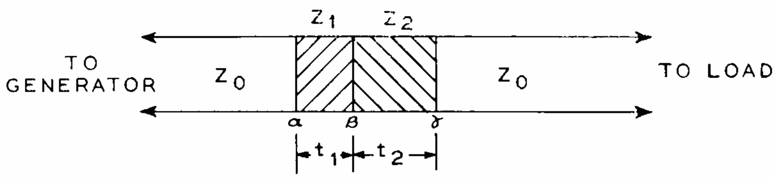

Fig. 3—Characteristic impedances—single-cup assembly.

Figure 3. The normalized impedance looking to the left of a plane through β is point B of Figure 4. The phase shift θ_1 in going from α to β is $4\pi \frac{t_1}{\lambda_o}$. The value of B is then normalized with respect to Z_2 (assume $Z_2 < Z_o$) and appears as point B' on the same figure. By a proper choice of mica thickness, the impedance at γ can be made to be purely resistive and be given by point C'. The mica thickness required will be that for which $\theta_2 = 4\pi \frac{t_2}{\lambda_m}$. If C' is now normalized with respect to Z_o, a value Z_c is obtained which is the impedance seen when one looks out of the main waveguide into the window assembly at plane γ. It is obvious that although Z_c can be made real, it can never be made exactly equal to Z_o, so that a system of this sort will always set up a standing wave in a matched system. For this case, the voltage standing-wave ratio is $\frac{Z_c}{Z_o}$. However, by designing the system so that

[2] J. C. Slater, MICROWAVE TRANSMISSION, McGraw-Hill Book Co., New York, N. Y., 1942 (p. 185, Equation 24.2).

θ_1 (and correspondingly θ_2) is made very small, Z_c can be made to approach Z_0 with a consequent reduction in voltage standing-wave ratio to an acceptable value. This can be accomplished by making t_1 (and correspondingly t_2) very small, or by letting b approach b'. As b approaches b', Z_1 approaches Z_0 until no appreciable "phase shift" occurs in section 1. As a consequence, little phase shift is required in section 2. This shift is accomplished physically by using a thinner section of mica. In certain cases, however, the minimum voltage standing-wave ratio that can be obtained is still excessively high. In

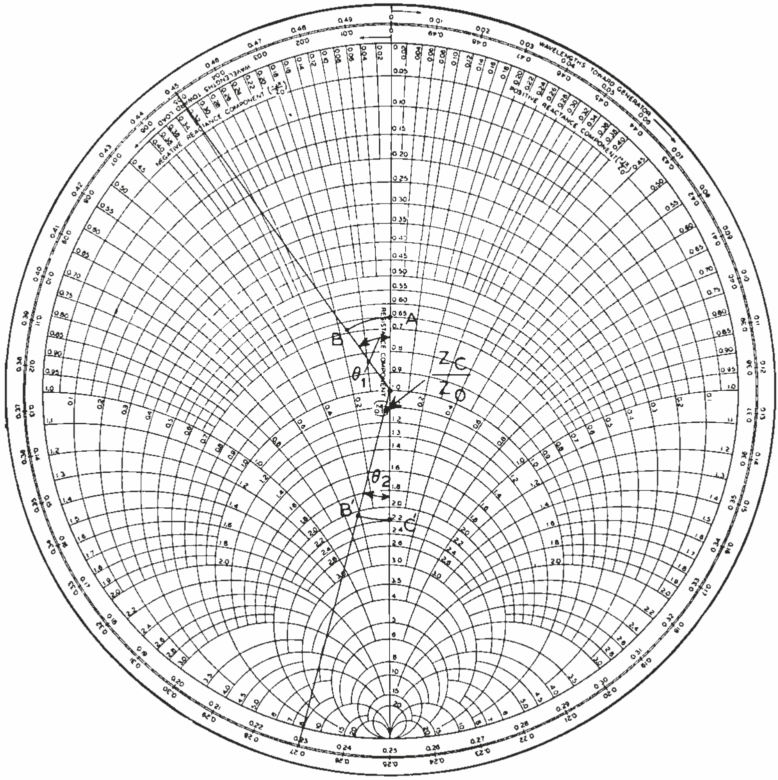

Fig. 4—Impedance chart for single-cup assembly.

these cases, perfect match can be obtained by matching the mica window system into the standard waveguide by means of a section of quarter-wave transformer or by a tapered section. Since these matching sections differ only slightly in dimensions from a standard waveguide, a broad-band match is obtained. The theoretical conclusions for the single-cup assembly may be summed up as follows:

1) A perfect match can be attained only by the use of a matching section.
2) For a given cup-hole diameter, improved match (lower voltage standing-wave ratio) can be obtained by using a thinner cup and thinner mica.
3) For a given thickness, improved match can be obtained by increasing the hole diameter and by using thinner mica.

As will be seen, these conclusions are borne out by experiment.

In view of the thinness of the window, it appears that a treatment of the problem in which the iris is considered as a shunt circuit might be more suitable. In this case, the desideratum would be to design the window for resonance at the middle of the operating band. An actual treatment along these lines was carried out and yielded the same qualitative conclusions as the series treatment.

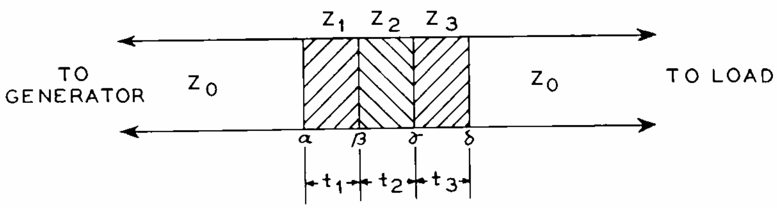

Fig. 5—Characteristic impedances—double-cup assembly.

Double-Cup Assembly

Figure 5 is a schematic representation of the "double-cup" assembly shown in Figure 2(b).

Although the double-cup assembly can be designed in any desired form, it is convenient to make $t_1 = t_3$ and to use the same hole diameter in both cups. The theoretical treatment will be worked out for this special case. Its extension to the general case, however, is quite similar.

The projected view of the essential dimensions of the double-cup assembly is shown in Figure 2(c). As before, the mica section is referred to as section 2, the other two sections, one in vacuum and the other in air, are referred to as sections 1 and 3 respectively. Once again $Z_1 > Z_o$, $Z_2 < Z_o$, $Z_3 > Z_o$. In Figure 6, point A is the value of Z_o normalized with respect to Z_1. Point B is the value of the impedance to the left of β normalized with respect to Z_1. Point B is then normalized with respect to Z_2 and appears as point B' on the same figure. By a proper choice of mica thickness, the impedance at γ can be made equal to that at β, with the reactive term of opposite sign. This reactive term is shown as point C'. Point C' is now normalized with respect to $Z_3 = Z_1$ and is shown as point C''. If $t_3 = t_1$, the impedance at δ (normalized with respect to Z_1) is equal to point A. The final impedance (looking to the left) at δ, normalized with respect to Z_o, is Z_o, and the system is therefore undisturbed by the introduction of the window assembly. Thus, these studies indicate that the double-cup assembly is essentially superior to the single-cup type because a "perfect" match can be achieved without the use of an additional matching

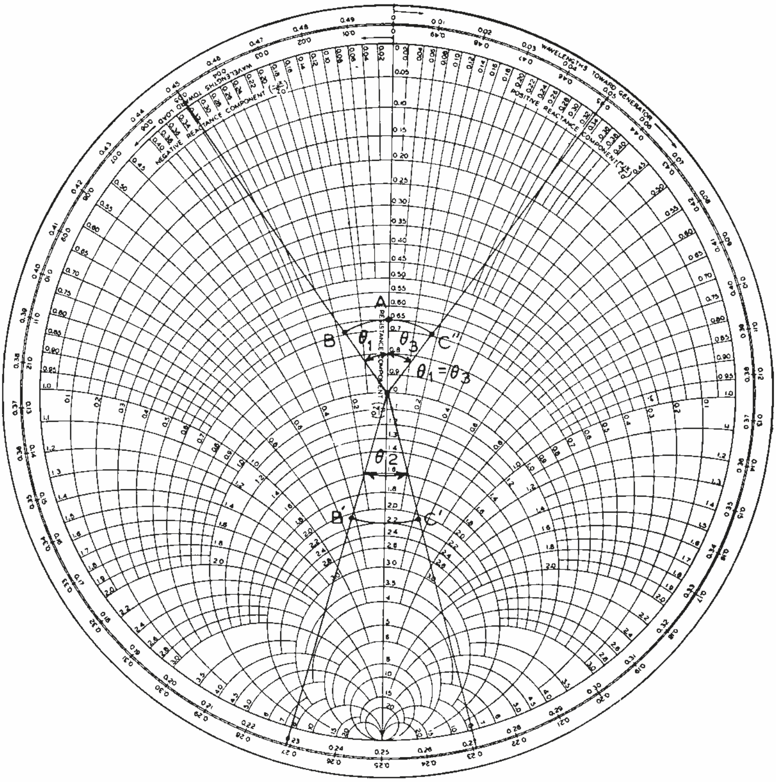

Fig. 6—Impedance chart for double-cup assembly.

section. The mica thickness required is proportional to the thickness of the cup and inversely proportional to the hole diameter. In this respect both types of assembly are similar.

Frequency Dependence

A theoretical study was made of the frequency dependence of both single- and double-cup assemblies. A single-cup design which gave a voltage standing-wave ratio of 1.02 had a voltage standing-wave ratio of 1.06 at frequency deviations of ± 20 per cent. A double-cup assembly of the same cup thickness and opening, with mica thickness designed to yield match, had a voltage standing-wave ratio of 1.09 at − 20 per cent frequency and a voltage standing-wave ratio of 1.10 at + 20 per cent frequency. This study indicates, therefore, that the single-cup assembly is superior with respect to frequency dependence.

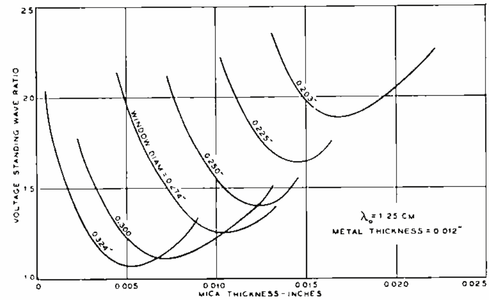

Fig. 7—Variation of voltage standing wave ratio with mica thickness for several window diameters (single-cup).

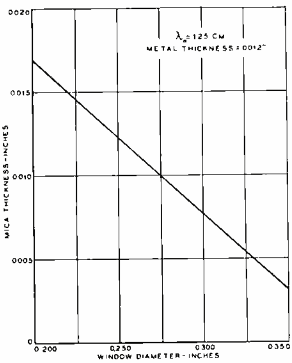

Fig. 8—Variation of mica thickness at minimum voltage standing wave ratio with window diameter (single-cup).

EXPERIMENTAL CHECK OF THEORETICAL CONCLUSIONS

Measurements were made at $\lambda_0 = 1.25$ centimeters on both single- and double-cup assemblies with results shown in Figures 7, 8, 9 and 10. From these it can be seen that the previously discussed theoretical results are supported qualitatively. The double-cup assembly yields much better matches than the single-cup type. Furthermore, in both cases, for minimum voltage standing-wave ratio, the mica thickness is an inverse function of window openings. In the case of the single-cup assembly without a matching section, large window openings must be used if low values of voltage standing-wave ratio are to be obtained. Attempts to correlate the results of the structures of Figures 2(a) and 2(b) with the theory yield appreciable numerical divergence. This indicates that the simplifications in the theory are too extreme and that more accurate results can be obtained experimentally.

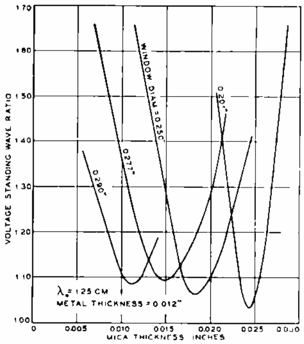

Fig. 9—Variation of voltage standing wave ratio with mica thickness for several window diameters (double-cup).

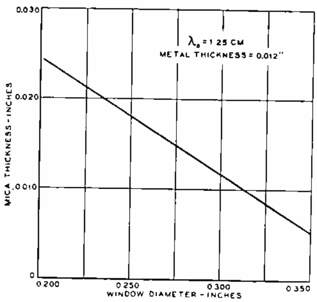

Fig. 10—Variation of mica thickness at minimum voltage standing wave ratio with window diameter (double-cup).

In an attempt to design a single-cup mica window for operation with a tunable waveguide-output magnetron at 3.2 centimeters, the minimum voltage standing-wave ratio obtainable was 1.16. By the use of a quarter-wave matching section, however, the following voltage standing-wave ratio values were obtained: 1.04 at 3.0 centimeters, 1.03 at 3.2 centimeters, and 1.08 at 3.4 centimeters. The design finally chosen for this assembly is shown in Figure 11.

Mica Window Design and Development

The technique developed by J. S. Donal, Jr. to secure vacuum-tight mica-to-metal seals was employed. In this technique the mica is bonded to the metal by means of a low-melting point glass which wets both the mica and metal. It is also essential that the coefficients of expansion of the three materials match fairly well. The glass and metal are chosen to match the mica whose coefficient of expansion is about 100 x 10^{-7} per degree Centigrade. The metal employed, known as Sylvania #4 Alloy (52 per cent Fe, 42 per cent Ni, 6 per cent Cr) has a coefficient of expansion of about 95 x 10^{-7} per degree Centigrade. The

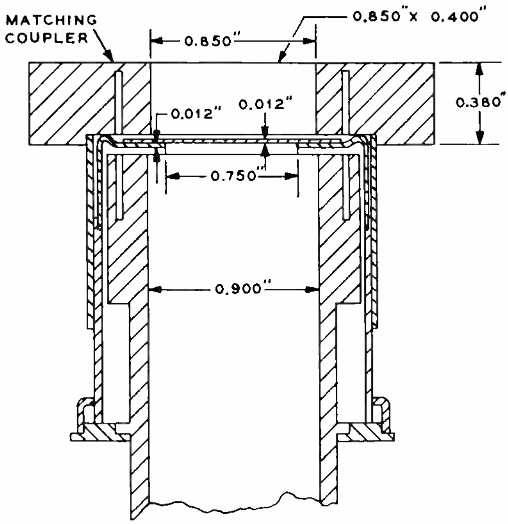

Fig. 11—Single-cup assembly—final design for 3.2 centimeter tube.

softening point of the glass is sufficiently lower than the temperature at which mica begins to disintegrate so that excellent seals could be achieved by baking assemblies at 600 — 650 degrees Centigrade. The mica used was high-quality India mica of the type employed in television pickup tubes.

Initial experiments were based on the double-cup design of Figure 2(b). However, it was found difficult to obtain uniformly vacuum-tight seals with this construction and as a consequence, the single-cup design of Figure 2(a) was adopted. This design has the advantage that the seal can be examined after forming. It is thus possible to determine visually whether or not a seal is satisfactory.

In practice the #4 Alloy cup is first cleaned by firing in dry hydrogen at 1100 degrees Centigrade for 15 minutes, and then oxidized by firing in "line" hydrogen at the same temperature for the same length of time. This oxidation is essential if the glass is to wet the metal. The mica is then put in place and a water paste of the special glass in finely powdered form is painted around its edges. A steatite disc is fastened in place so that it bears down on the center of the mica disc. The assembly is heated in an oxidizing or neutral atmosphere at 600 degrees — 650 degrees Centigrade for 15 minutes. (A reducing atmosphere would reduce the lead oxides in the glass.) If on examination the seal appears defective, additional glass may be painted on and the assembly reheated. The cup is then brazed in place on the tube. Since #4 Alloy does not braze readily, it is first plated with nickel, after which any of the standard high-temperature solders wet it well. Leaky windows can be healed even after complete assembly of the tube. This constitutes a decided advantage of the mica window over the glass type.

Question has been raised as to the ability of the windows to stand up under the pressures to which they may be subjected. In this connection the experiments of J. S. Donal, Jr.[1] indicate that this need be of no concern. Scores of tubes employing mica windows were built for operation at 1.25 and 3.2 centimeters. Pulsed power outputs in excess of 50 kilowatts at 1.25 centimeters and continuous-wave outputs in excess of 500 watts at 3.2 centimeters were "pumped" through the window. In only one case was mica disintegration observed, and this might have been due to impurities in the particular sample of mica involved.

ACKNOWLEDGMENT

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RECORDING STUDIO 3A*

By

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Summary—This article discusses the acoustical design problems in remodeling a studio for broadcast transcription and recording usage. The use of adjustable acoustical elements to provide a change in reverberation time of about 2:1 is covered and the description of the application of diffusely reflective surfaces in combination with absorbent areas for optimum acoustical results is included.

RECORDING Studio 3A, to be used jointly by the NBC (Radio Recording Division) and RCA Victor Division, designed for recording both program transcriptions and records for home use,

Fig. 1—Overall view of Studio 3A.

was placed in service in August 1946. It represents a modification of the former broadcast Studio 3A designed in 1933 when NBC first began

* Decimal Classification: R800 (681.843).

operations in Radio City. The studio has a volume of approximately 68,000 cubic feet, a length of 80 feet, a width of 50 feet, and a height of 17 feet. Two views of the studio are shown, the studio overall (Figure 1) and the platform end (Figure 2).

The acoustical problem in redesigning old Studio 3A for recording use was to provide as nearly as possible ideal acoustical conditions for performances of any type of program which was to be recorded and in which one to fifty performers would be participating. The ideals to be achieved for transcription type recording have been established by experience gained in broadcasting work wherein a transcription broadcast is simply a delayed transmission to the listener and must have the full qualities of a live program. Acoustical requirements for producing

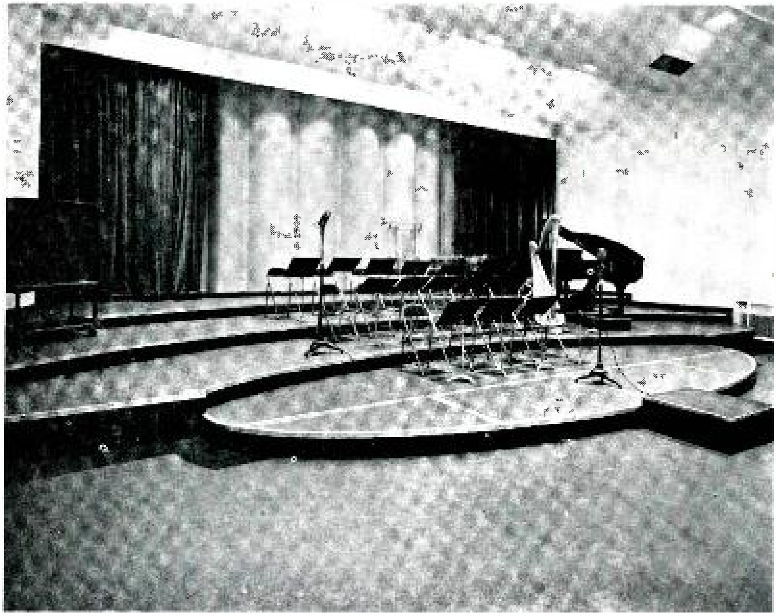

Fig. 2—Platform end of Studio 3A.

records for use in the home have also been determined through experience over a period of years.

The acoustical criteria for a studio designed for transcription purposes and one designed for recording records for home use are somewhat different, largely as to the frequency-reverberation time characteristics. Transcriptions are designed to simulate broadcasts and for that reason acoustical conditions should be substantially the same as those employed in broadcast studios. A single broadcast program usually involves an appreciable percentage of speech—either announcements, talks or dramatic presentations—which require a somewhat non-reverberant acoustical condition to prevent an excessive amount of reflected sound from reaching the microphone and creating an impression of room size. On the other hand, the studio should be fairly reverberant for the proper quality of music or other program material. In the broadcast studio these two requirements must be satisfied rather than compromised, since the latter course would result in unfavorable effects for both conditions.

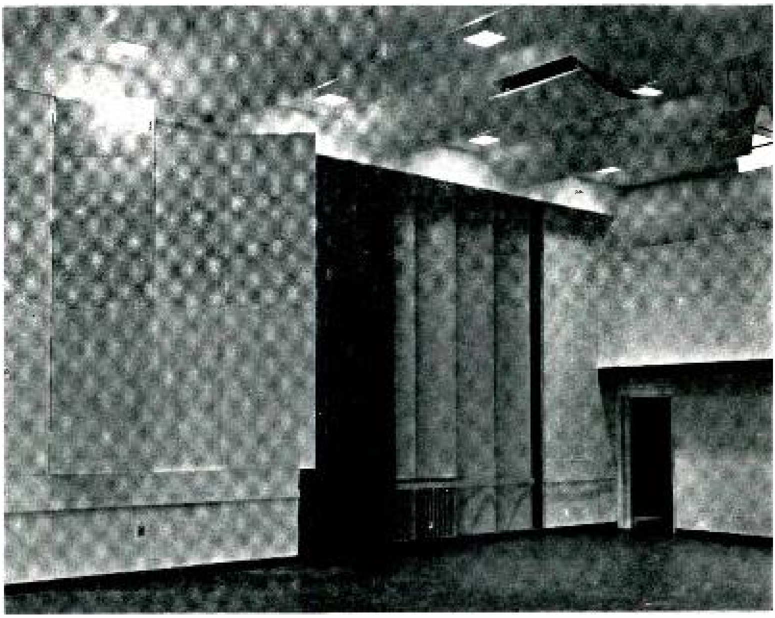

Fig. 3—Hinged acoustical panels.

The recording studio is concerned almost exclusively with music and for that reason the problem is to provide the proper reverberation characteristic for the type and size of the performing group. However, when speech is recorded, control may be exercised by the use of sound-absorbent flats. In a recording studio, the use of the flats presents no problem of obstructing visibility since no audience is present. In a broadcast studio, audiences frequently witness the programs and the control of acoustical conditions must be handled in a different way.

Studio 3A was designed so that the acoustical conditions could be altered to match the program. A consideration of several methods indicated that the most practical solution lay in the use of heavy lined and interlined draperies, together with hinged acoustical panels as shown in Figure 3. The change in reverberation time was almost two to one, that is, at 1000 cycles, a change from 0.9 seconds to about 1.7 seconds. The use of heavy draperies hung some distance from the wall insured that the change would be effective even at the lower frequencies (see Figure 4).

The shape of the reverberation time frequency characteristic, selected as a design objective, is the one employed by NBC and RCA in studio design in which the reverberation time below 500 cycles increases as the frequency decreases (see Figure 5). Experience has

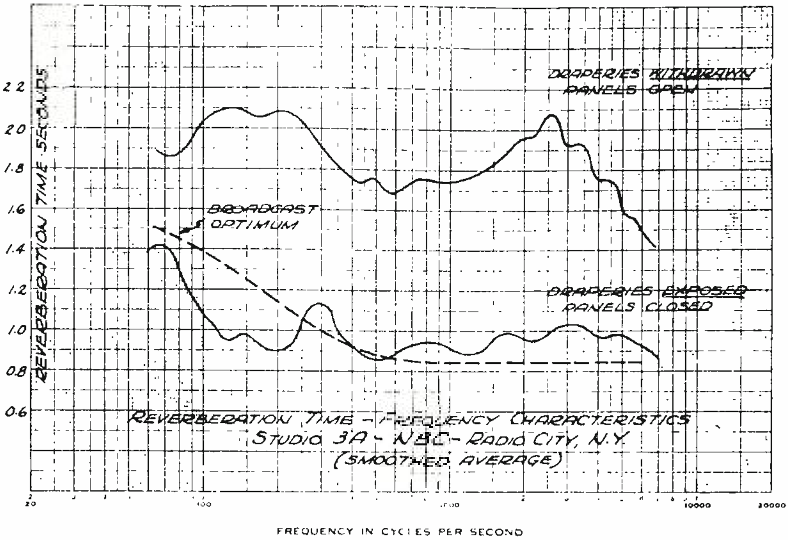

Fig. 4—Reverberation time-frequency characteristics.

shown that this characteristic insures a proper tonal balance for both speech and music.

It was desired further that the reverberation time characteristic extend to as high a frequency as possible before falling off to insure the highest possible fidelity of recording. It is a well-known fact that the absorption of the air prevents appreciable control above 8,000 cycles and, with a continued increase in the frequency, it becomes increasingly a larger factor than the absorption provided by the walls, floor and ceilings.

The selection of the reverberation time characteristics determines the total amount of absorbing material required but does not indicate its distribution. Distributed acoustical treatment combined with the use of diffusely reflective surfaces produces a more diffuse sound field than is obtained in rooms with plane boundary surfaces with concentrated areas of acoustical treatment. Under the former set of conditions the exact location of the microphone has been found to be less critical for obtaining a good balance thereby facilitating the obtaining of the highest quality of recording.

The emphasis in the design has been toward obtaining a very diffuse sound field. Where this condition is realized, the importance of reverberation time is lessened, i.e., in a plane rectangular room experience has shown relatively narrow limits of tolerance from an

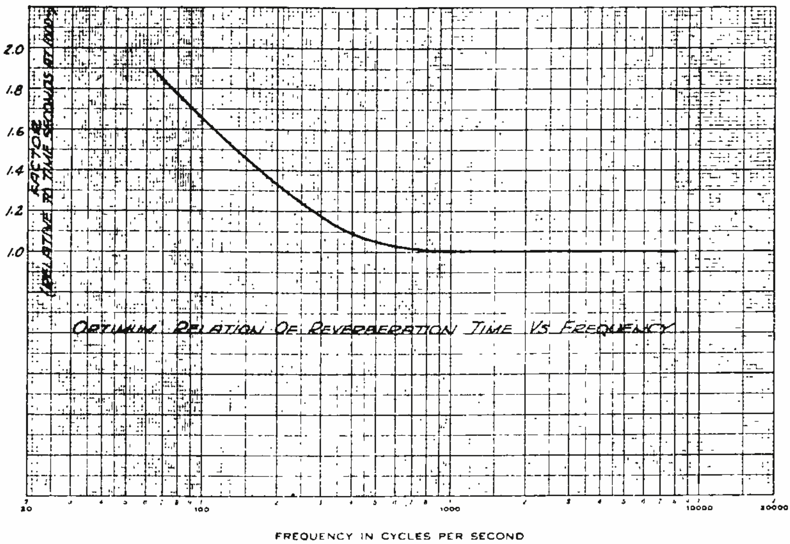

Fig. 5—Reverberation time-frequency ratio.

optimum reverberation time. Where the sound field is more diffuse, the upper limit of this tolerance may be raised appreciably with beneficial rather than objectionable effects.

The use of diffusing surfaces with interposed absorbing surfaces results in a very substantial increase in their absorbing efficiency which cannot be neglected in calculations. This is due partially to diffraction effects and to the better coupling of the absorbent medium to the air.

It was originally intended to employ a large amount of plywood on the wall and ceiling areas. Curved plaster surfaces were selected, however, in preference to wood in this case because of the necessity of not using critical materials. Experience thus far indicates that the diffusion is the major factor, and its manner of achievement is of lesser importance. It is not implied that plaster is a superior material to plywood but rather that the shape of the diffusely reflective surface is of more importance than the material employed in forming that shape.

The major acoustical adjustment is obtained by means of draperies which are usually considered to be deficient in absorption at the lower frequencies (below several hundred cycles). The draperies employed were quite heavy, lined and interlined, being about 100 per cent full and placed about one foot from a diffusely reflective wall. This procedure tends to insure a more uniform absorption characteristic with the use of these draperies than the characteristic absorption curve of lighter weight drapery in which the absorption increases with frequency up to one or two thousand cycles and then tends to decrease. This uniformity of the change in reverberation time-frequency characteristic (Figure 4) between the most reverberant and the least reverberant condition can be easily seen. The measured reverberation time frequency characteristic also is shown in Figure 4 together with the optimum curve for broadcasting. It will be noted that in the most reverberant condition, the curve is considerably above the broadcast optimum. Further, as the draperies and hinged absorbent panels are exposed, the curve tends to be reduced to the same general extent at all frequencies thereby maintaining a proper balance between low, medium and high frequencies almost irrespective of the reverberation condition selected.

In this studio the floor is of concrete covered by battleship linoleum and is highly reflective to sound. At the platform end, there are a series of steps, concave in plan, to permit arrangement of the performing group for usual microphone technique. These steps are augmented by wooden platforms as required for individual or groups of performers in an orchestral or choral performance.

The ceiling above the platform, for a distance of 25 feet from the rear wall, is treated with a series of highly reflective, curved plaster, quasi-elliptical shapes to reflect the sound diffusely and to deflect it generally outward into the studio. The remainder of the ceiling is treated with large semi-cylindrical plaster forms extending the full width of the studio, interspersed with 3-foot strips of 1-inch rock wool blanket covered with perforated transite.

The rear wall at the platform end consists of a number of vertical semi-cylinders of different diameters. These may be covered by a heavy drapery so that wall is changed from a highly diffusely-reflective surface to a highly absorbent one.

The side walls at the platform end diverge outward from the rear wall with a series of large, vertical, convex plaster shapes which tend to reflect diffusely the sound outward. The remainder of the side walls are, in general, of curved plaster arranged to be covered by draperies. These draperies are arranged to fit into pockets formed by the curved sections of the wall itself or behind the flat pilasters on which are mounted the hinged acoustical panels.

It is planned to study the comments of users of the studio and listeners of records made in this studio with a view of establishing the optimum acoustical criteria for the recording of transcriptions and records for home use. These studies may indicate the desirability of some physical changes in the studio for certain types of recordings. Many coordinated opinions are necessary to support any statement that the collective judgment of listeners prefer one condition instead of another. There is no question, however, that the studio represents a distinct advance over former recording studio design and, it is believed further usage will provide additional confirmation of that fact.

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TELEVISION

A Bibliography of Technical Papers

by RCA Authors

1929 — 1946

This listing includes some 275 technical papers on TELEVISION and closely related subjects, selected from those written by RCA Authors and published during the period 1929-1946.

Papers are listed chronologically except in cases of multiple publication. Papers which have appeared in more than one journal are listed once, with additional publication data appended.

Abbreviations used in listing the various journals are given on the following page.

ABBREVIATIONS

An. Amer. Acad. Polit. Soc. Sci.	ANNALS OF THE AMERICAN ACADEMY OF POLITICAL AND SOCIAL SCIENCES
Broadcast News	BROADCAST NEWS
Broad. Eng. Jour.	BROADCAST ENGINEERS JOURNAL
Communications	COMMUNICATIONS
Elec. Eng.	ELECTRICAL ENGINEERING (TRANSACTION A.I.E.E.)
Electronics	ELECTRONICS
Electronic Ind.	ELECTRONIC INDUSTRIES
FM and Tele.	FM AND TELEVISION
FM Business	FM BUSINESS
Inter. Project	INTERNATIONAL PROJECTIONIST
Jour. Appl. Phys.	JOURNAL OF APPLIED PHYSICS
Jour. Frank. Inst.	JOURNAL OF THE FRANKLIN INSTITUTE
Jour. Opt. Soc. Amer.	JOURNAL OF THE OPTICAL SOCIETY OF AMERICA
Jour. Soc. Mot. Pic. Eng.	JOURNAL OF THE SOCIETY OF MOTION PICTURE ENGINEERS
Jour. Tele. Soc.	JOURNAL OF THE TELEVISION SOCIETY
Phys. Rev.	PHYSICAL REVIEW
Proc. I.R.E.	PROCEEDINGS OF THE INSTITUTE OF RADIO ENGINEERS
QST	QST (A.R.R.L.)
Radio and Tele.	RADIO AND TELEVISION
Radio Craft	RADIO CRAFT
Radio Eng.	RADIO ENGINEERING
Radio News	RADIO NEWS
Radio Tech. Digest	RADIO TECHNICAL DIGEST
RCA Rad. Serv. News	RCA RADIO SERVICE NEWS
RCA REVIEW	RCA REVIEW
RMA Eng.	RMA ENGINEER
Short Wave and Tele.	SHORT WAVE AND TELEVISION
TBA Annual	ANNUAL OF THE TELEVISION BROADCASTERS ASSOCIATION
Tele. News	TELEVISION NEWS
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Television	TELEVISION

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NOTE—Omissions or errors in these listings will be corrected in the yearly Index.

Corrections:

On page 366 of the September 1946 issue, the following sentence should have been included under ACKNOWLEDGMENTS: "This paper is based in whole or in part on work done for the Office of Scientific Research and Development under Contract OEMsr 441 with Radio Corporation of America."

In Figure 1 on page 180 of the June 1946 issue the indicated scale is in error. The actual scale used in the map is approximately double that indicated in the legend.

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GERALD G. GERLACH studied electrical engineering at Brooklyn Polytechnic Institute. From 1929 to 1935 he was associated with the Engineering Departments of F. A. D. Andrea, Inc., Hygrade Sylvania, Inc., and Wired Radio, Inc., specializing in receiver development and design. He joined Radio Corporation of America in October, 1935, and is now in the Communication Equipment Section of the Engineering Products Department, RCA Victor Division, Camden, N. J. Mr. Gerlach is an Associate Member of the Institute of Radio Engineers.

WILLIAM D. HOUGHTON graduated from the General Engineering Course at the New York school of RCA Institutes, Inc. in 1939. In 1940 he joined the Transmitter Research and Development Laboratory of RCA Communications, Inc. He has been a member of the Radio Systems Research Laboratory of RCA Laboratories Division at Rocky Point, L. I., N. Y. since 1942. Mr. Houghton is an Associate Member of the Institute of Radio Engineers.

ROBERT L. JEPSEN, following his graduation with a B.S. degree in Electrical Engineering from the State College of Washington in 1944, joined the Special Development Division of the RCA Victor Division at Lancaster, Pa. He is currently employed at the Columbia Radiation Laboratory, and is studying graduate physics at Columbia University.

LEWIS MALTER received the B.S. degree from the College of the City of New York in 1926 and the M.A. and Ph.D. degrees in physics from Cornell University in 1931 and 1936, respectively. He taught physics at the College of the City of New York from 1926 to 1928. He was with the Acoustic Research and Photophone Divisions of Radio Corporation of America from 1928 to 1930, and during the summer of 1931. From 1933 to 1942, he was with RCA Manufacturing Company at Camden, N. J., and from 1942 to 1943 with RCA Laboratories Division at Princeton, N. J. From 1943 to 1946, he was in charge of the Special Development Division of RCA Manufacturing Company at Lancaster, Pennsylvania. Since May, 1946, he has been in charge of the Vacuum Tube Research Section of the Naval Research Laboratory, Washington, D. C. Dr. Malter is a Fellow of the American Physical Society, a Senior Member of the Institute of Radio Engineers, a Member of the American Association for the Advancement of Science, and a Member of Sigma Xi.

GEORGE M. NIXON received his schooling at Pratt Institute, New York University and the American Institute of Electrical Engineering. He has been a member of the Engineering Department of NBC since 1928, in the positions of Student Engineer, Engineer-Development Group, Assistant Development Engineer and Assistant Director of Technical Development, which position he now holds. He is concerned with engineering development work associated with sound and television broadcasting and recording, acoustical design of studios, sound control, sound isolation, reverberation and attenuation measurements, development of parabolic reflector microphones, loudspeakers and amplifiers, and the standardization of operating practices.

HARRY F. OLSON received the B.S. degree in 1924, the M.S. degree in 1925, the Ph.D. degree in 1928 and the E.E. degree in 1932 from the University of Iowa. From 1928 to 1930 he was in the research department of Radio Corporation of America; from 1930 to 1932, in the engineering department of RCA Photophone; from 1932 to 1941, in the research division of RCA Manufacturing Company; since 1941, with RCA Laboratories Division. Dr. Olson is a member of Tau Beta Pi, Sigma Xi and the American Physical Society, an Associate Member of the Institute of Radio Engineers and a Fellow of the Acoustical Society of America.

WALDEMAR J. POCH was graduated from the University of Michigan in 1928 with the B.S. degree in Electrical Engineering. After a year under instruction with General Electric Company in Schenectady he entered the Radio Department and was transferred in 1930 to Radio Corporation of America in Camden, N. J. In 1936 he obtained his M.S. degree from the University of Pennsylvania. He has been continually associated with television development work since 1931; before and during the war he was concerned mainly with the development of portable field equipment and light-weight television equipment for use in military aircraft. He is now in charge of television systems and relay equipment design for RCA Victor Division. Mr. Poch is a member of Tau Beta Pi, Sigma Xi, and an associate member of the Institute of Radio Engineers.

ROBERT W. K. SMITH received his B.S. degree in Physics from Oregon State College in 1933. He was with Warner Bros. as sound technician and junior engineer for seven years. Following a year with Bendix Aviation as radio engineer, he joined Sperry Gyroscope Company in 1941 as an engineer on microwave applications for air navigation and communications. He is now with RCA Victor Division in Camden, N. J., as a development engineer.

MERRILL TRAINER was graduated from Drexel Institute of Technology, Philadelphia, with a B.S. in Electrical Engineering. He was associated with Dr. E. F. W. Alexanderson in television research from 1927 to 1930, and in the latter year joined Radio Corporation of America. Since that time he has been actively engaged in television systems development with particular reference to studio equipment and theater projection equipment design. During the war Mr. Trainer was engaged in supervising the development of television equipment for airborne use. He is now in charge of television terminal equipment development for RCA Victor Division in Camden, N. J. Mr. Trainer is a member of Tau Beta Pi.

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