Proceedings of the I·R·E

MAY 1942

VOLUME 30 NUMBER 5

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Published Monthly by
The Institute of Radio Engineers, Inc.

VOLUME 30 — *May, 1942* — NUMBER 5

Entered as second-class matter October 26, 1927, at the post office at Menasha, Wisconsin, under the Act of February 28, 1925, embodied in Paragraph 4, Section 538 of the Postal Laws and Regulations. Publication office, 450 Ahnaip Street, Menasha, Wisconsin. Editorial and advertising offices, 330 West 42nd St., New York, N. Y. Subscription, $10.00 per year; foreign, $11.00.

THE INSTITUTE OF RADIO ENGINEERS

INCORPORATED

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

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June 19	May 28	June 19	June 16

PHILADELPHIA	PITTSBURGH	WASHINGTON
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Experimental Polyphase Broadcasting*

PAUL LOYET†, MEMBER, I.R.E

Summary—*An experimental installation of polyphase broadcasting is described using 1000 watts carrier power on a frequency of 1000 kilocycles. Fidelity measurements made of the modulated field intensity radiated from this installation indicate that comparable performance to other present-day amplitude-modulated broadcast transmitting equipment can be obtained readily. It is demonstrated that transmitting equipment of this type need only have a peak power capability of the output tubes approximately $1\frac{1}{2}$ times carrier power.*

Introduction

ENGINEERING developments of the past few years have resulted in substantial increases in the over-all efficiency of broadcast transmitting equipment (550 to 1600 kilocycles), particularly in the higher-power ranges. These developments include the use of the Doherty amplifier and the use of high-level class B modulation. High-efficiency equipments of these types have resulted in over-all efficiencies of the order of 40 per cent being obtained for transmitters having a power output of 5 kilowatts and more. It is the purpose of this paper to describe a transmitting system which results in a further increase in the over-all efficiency by reducing the number of high-power vacuum tubes required. For high-efficiency equipment of the above types, nearly 30 per cent of the rated carrier power of the transmitter is required for filament heating of the final amplifier tubes alone.

For all conventional transmitting equipment for this service, the radio-frequency power output of the transmitter varies from zero to four times the carrier power during each cycle of the audio-frequency modulating voltage for 100 per cent modulation. It is thus necessary that the final power-amplifier tubes be capable of delivering a peak power equal to four times the carrier power of the transmitter. A unique feature of the system to be described lies in the fact that the radio-frequency power output of the transmitter is constant throughout the audio-frequency modulation cycle. This system of polyphase broadcasting has been described in detail by Byrne.[1]

Recognizing the possibilities of polyphase broadcasting, the Central Broadcasting Company in conjunction with the engineers of the Collins Radio Company undertook a typical test of the system at broadcast frequencies. As most of the characteristics of the system could equally well be studied on a low-power basis, an experimental license was obtained for use of 1000 watts operating on WHO's regularly assigned frequency of 1000 kilocycles. The results here reported were obtained at the WHO transmitter site, where the antenna system was modified for polyphase operation.

* Decimal classification: R550. Original manuscript received by the Institute September 4, 1941; revised manuscript received, April 6, 1942. Presented, Broadcast Engineering Conference, Ohio State University, Columbus, Ohio, February 13, 1941.

† Collins Radio Company, Cedar Rapids, Iowa.

[1] John F. Byrne, "Polpyphase broadcasting," *Elec. Eng.*, 58, 347–350; July, 1939.

These results show that the system is entirely practical and capable of high-fidelity broadcast service according to all present-day standards for amplitude-modulation equipment. Before presenting the data obtained, the theory of polyphase broadcasting will be reviewed and the equipment described under the following subdivisions: (1) The Antenna System, (2) The Transmitter, (3) The Monitoring System, (4) The Measuring Equipment.

Review of Polyphase Broadcasting

In the conventional transmitter, the antenna current varies from zero to twice carrier value when 100 per cent amplitude modulated, and the field intensity produced varies in a similar manner. Also the field intensity varies in this manner in all directions from the radiating system simultaneously. That is, the variation of the field intensity due to modulation at any point is in exact phase with the variation of the field intensity at any other point an equal distance away.

In the system of polyphase broadcasting, the variation of the field intensity due to modulation at a point in one direction from the antenna is *not* in phase with the variation in field intensity at points equally distant in other directions. For example, a modulation peak occurs north of the antenna system at the same time that a modulation null occurs south. One quarter of the modulation cycle later, the modulation peak occurs east of the antenna system and the modulation null occurs west. As the modulation progresses through the audio-frequency cycle, this condition rotates about the radiating system. Thus at 100 per cent modulation, the instantaneous power output of the system does not vary although the field intensity at any fixed point varies in accordance with the modulation from zero to twice carrier value. This rotating modulation field is obtained by means of a 5-element vertical antenna array and associated transmitting and coupling equipment.

In the antenna system, four identical radiators are arranged at the corners of a square with the fifth antenna in the center. The central antenna is supplied with unmodulated radio-frequency power and is referred to as the carrier antenna. The antennas at opposite ends of one diagonal constitute a directional pair and are supplied with suppressed-carrier modulated radio-frequency power. The two antennas at the ends of the other diagonal also constitute a directional pair and are fed with suppressed-carrier modulated radio-frequency power. The two antennas of each sideband pair are fed in series and carry currents of opposite phase so that the directional patterns of the sideband pairs are figures of eight at right angles. Further, the resultant phase of the suppressed-carrier currents in the sideband pairs differs from the phase of the current in the carrier antenna by an angle of 90 degrees. If, in addition, the modulating voltages to the suppressed-carrier modulators are 90 degrees in phase, the required conditions for a rotating modulation field are fulfilled.

The field-intensity patterns existing at three instants 45 degrees apart in the audio-frequency cycle of modulation are shown in Fig. 1. In column A, two sine waves are shown displaced 90 degrees in phase. These algebraic sum of these two patterns. This algebraic sum is shown graphically by the third figure in column A which is a cardioid pattern having twice carrier intensity to the west, zero intensity to the east, and carrier values north and south.

In column B is shown the conditions one eighth of the modulation cycle later. At this instant the modulating voltage associated with the east-west antenna pair has been reduced to 70 per cent of its maximum value, and the modulating voltage associated with the north-south pair has reached 70 per cent of its maximum value. At this instant both the east-west and north-south pairs are operating producing the fields shown in the second figure of column B. The third figure of column B is the algebraic sum of the field intensities and shows that the cardioid field pattern has rotated counterclockwise 45 degrees.

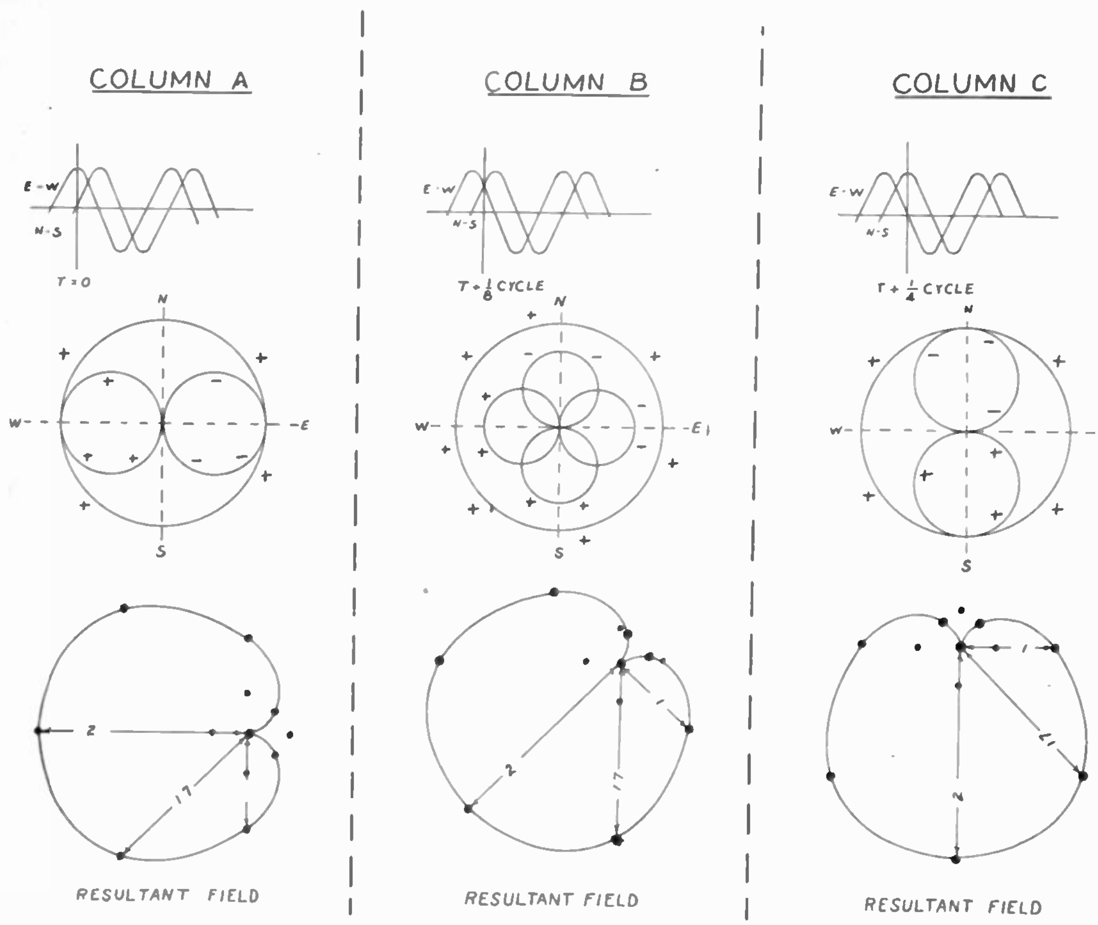

Fig. 1—Horizontal field of pattern development.

sine waves represent the magnitude of the audio modulating voltages applied to the suppressed-carrier modulators of the transmitter. At the particular instant shown $T=0$, the modulating voltage associated with the north-south antenna pair is zero and the modulating voltage associated with the east-west antenna pair is at its maximum. Immediately below this diagram is a diagram of the field intensities produced by the antenna system. The circle represents the field intensity due to the unmodulated-carrier antenna. The horizontal figure of eight represents the field intensity due to the east-west antenna pair at this instant. Further, the phase relation of the field produced by the east-west antenna is such that the left-hand half of the figure of eight is in phase with the carrier field and the right-hand half is 180 degrees out of phase with the carrier field. The field intensity in any direction is the

In column C, conditions one quarter of the modulation cycle later are shown. At this instant the east-west modulating voltage has been reduced to zero, while the north-south modulating voltage has reached its maximum. The instantaneous fields due to the antenna system are shown in the second figure and the algebraic sum of the fields is shown in the third figure of column C. At this instant the directional pattern of the array has rotated counterclockwise 90 degrees. In this manner, as the modulation progresses through a complete cycle, the field-intensity pattern rotates counterclockwise one revolution. The root-mean-square value of the field-intensity pattern does not change from instant to instant in the cycle and thus the total power output of the system is constant throughout the cycle. Calculation shows that at this 100 per cent modulated condition the root-mean-square value of the field has been increased $22\frac{1}{2}$ per cent and the radiated power has increased 50 per cent over the unmodulated carrier value.

The Antenna System

Fig. 2 is a photograph of the radiating system installed at WHO. Fig. 3 is an elevation drawing of the antenna system. A spacing of 110 feet was first used for the sideband antenna pairs and this was later increased to 150 feet in order to increase their radiation resistance. The four insulated guy wires were fastened at the top of the tower and anchored at a point approximately 350 feet from the base of the tower. At a point along the guy wire 140 feet from the top end, vertical radiators were suspended constructed of seven-strand No. 18 phosphor-bronze cable. These antennas were dropped vertically to a point 75 feet from the base of the tower. In order to increase the effective length of the sideband radiators, 110 feet of the guy wire was used to top load the vertical radiators. The radiating part of the system is indicated by heavy lines on Fig. 3 and the supporting guy wires and tower structure by light lines. Each sideband antenna pair was fed in series to produce the desired figure-of-eight pattern.

Three important points characterize the antenna system arrangement:

Fig. 2—WHO transmitter building and antennas.

1. The presence of current in one of the sideband pairs induces no voltage in the other sideband pair nor in the carrier antenna inasmuch as the currents are equal and opposite in phase and produce equal and opposite effects on the radiators in question.

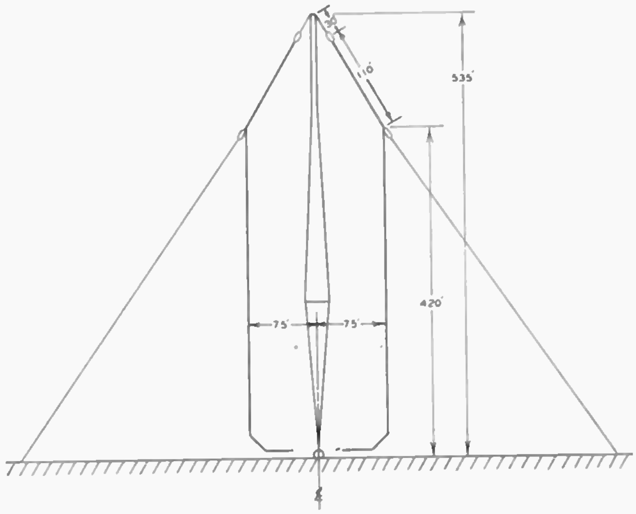

Fig. 3—Antenna elevation, polyphase experiment, WHO.

2. The carrier radiator when carrying current induces equal voltages in each of the sideband antennas. Inasmuch as the sideband antenna pairs are fed in series, the induced voltages are in phase opposition and thus produce no current.

3. Due to the fact that the mutual coupling between the carrier antenna and the sideband antennas, and between the sideband antennas themselves, is zero, all adjustments of tuning and phasing of the antenna system are noninterlocking, thus greatly simplifying the adjustment of the system.

The Transmitter

The transmitter used was designed and manufactured by the Collins Radio Company. Fig. 4 is a block diagram of the transmitter arrangement. Since the transmitter consists of three individual radio-frequency amplifier channels, the general configuration is somewhat more complicated than that of conventional broadcast transmitters. It includes a carrier amplifier channel and two sideband amplifier channels.

The carrier amplifier channel employs two type 833 tubes in the final stage. This section of the transmitter contains no unusual features, as it supplies unmodulated radio-frequency power to the carrier antenna. All stages are operated as class C amplifiers and all tubes may be operated according to class C telegraph ratings.

A unique feature of the transmitter equipment is that while double-sideband currents (suppressed carrier) are supplied to the two sideband antenna pairs, the power amplifiers employed in the sideband amplifier channels amplify only single-sideband currents. Each sideband amplifier channel employs a type HF-300 tube in the final stage operated as a linear class B

amplifier. At no modulation these tubes have no output, and at 100 per cent modulation level these tubes are operating at near saturation level.

The audio input to the transmitter is supplied to two audio-frequency phase-shift networks whose inputs are connected in parallel. These phase-shift networks are of the all-pass type with constants so chosen that the resulting phase difference of the audio-frequency output voltage is substantially 90 degrees throughout the usable range of 30 to 10,000 cycles per second. These quadrature audio-frequency voltages are used amplifier channels is similar to that of the balanced modulators. In this manner suppressed-carrier double-sideband currents as from balanced modulator *A* may be supplied to one sideband antenna pair and suppressed-carrier double-sideband currents as from balanced modulator *B* may be supplied to the other sideband antenna pair. A radio-frequency phase-shift network is added in one double-sideband output channel providing 90-degree phase shift. In this manner the phase of the suppressed carrier is the same for the double-sideband currents supplied to each sideband

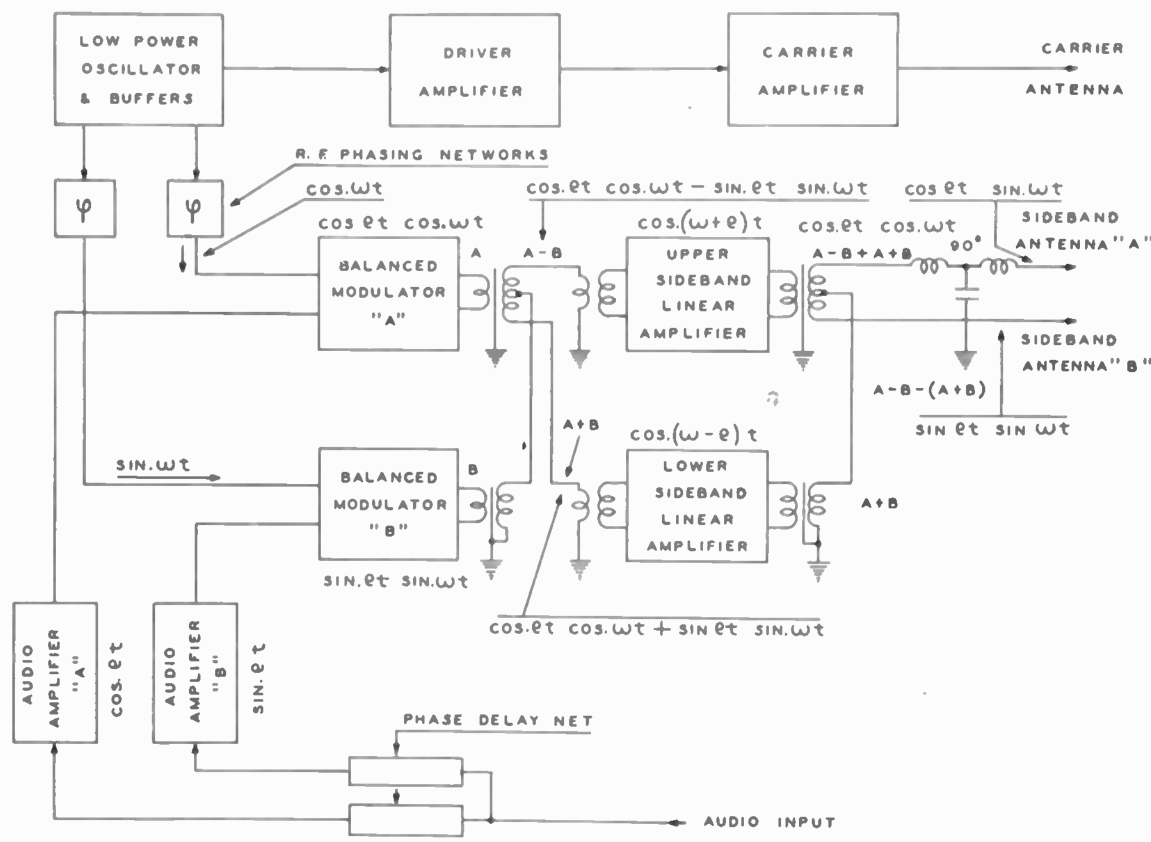

Fig. 4—Block diagram, polyphase transmitter.

to produce suppressed-carrier double-sideband voltages in balanced modulators *A* and *B*. The balanced modulators each consist of a pair of type 809 tubes. The radio-frequency excitation voltages supplied to the balanced modulators have a phase difference of 90 degrees.

The output of balanced modulator B is fed to the center-tapped winding in the output of balanced modulator *A*. Thus the sum of the output voltages appears in one sideband amplifier channel and the difference of the output voltages appears in the other. Since the output of balanced modulator *A* may be represented by cos pt cos ωt and the output of balanced modulator *B* may be represented by sin Pt sin ωt, the sum is the lower sideband and the difference the upper sideband. In these expressions, p is 2π times the audio frequency and ω is 2π times the radio frequency.

The output-circuit configuration of the sideband antenna pair. A result of this circuit arrangement is that the output of balanced modulator *A* retains its identity and completely controls the input to sideband antenna pair *A*. Similarly, the output of balanced modulator *B* affects only sideband antenna pair *B*.

It has been pointed out above that the total power output is constant and equal to 1$\frac{1}{2}$ times the carrier power of the transmitter for 100 per cent tone modulation. If the sideband linear amplifiers operate single sideband, these amplifier tubes or channels will operate at constant amplitude for a given percentage modulation at a single-frequency tone. Each amplifier operating at constant amplitude supplies a power equal to one fourth of the carrier power of the transmitter for 100 per cent tone modulation. The result is that the total power capability of the carrier amplifier and the two sideband amplifiers need only be equal to 1$\frac{1}{2}$ times the rated carrier power of the equipment. This is the

major point of the economy of the polyphase-broadcasting system. The total final-amplifier peak-power capability may be reduced from 4 to 1.5 units. It is readily shown that if the linear amplifiers were operated as suppressed-carrier double-sideband amplifiers, the total peak-power capability would be twice the carrier power, for under this condition of operation each amplifier is idle twice every cycle. Each amplifier must individually be capable of 50 per cent of the carrier power of the transmitter for this condition. In addition, the suppressed-carrier double-sideband amplifier would not operate as efficiently as the single-sideband amplifier. Inasmuch as single-sideband operation may be so easily obtained, this feature was included in the design of this experimental transmitter.

Fig. 5—Front view of 1000-watt polyphase transmitter.

Figs. 5 and 6 are photographs of the experimental 1000-watt transmitter. The center unit is the carrier amplifier and those adjacent, the two single-sideband amplifying units. Fig. 7 is a photograph showing the coupling house at WHO where the experimental transmitter was installed. The coupling units for matching to the sideband antenna pairs were temporarily supported at this time as a large ground screen was being installed around the base of the main radiator.

The Monitoring System

Another feature of the polyphase broadcasting system that is unique is the monitoring system. Inasmuch as the various components of the amplitude-modulated wave are supplied from independent radiators, not to mention independent amplifiers, the phase of these various radio-frequency components is quite arbitrary and dependent upon phase shifts in the transmission lines, coupling system, and antenna system. If these phase shifts were accurately known, a phase meter could be connected to the output of the carrier and sideband amplifiers and used for monitoring.

Fig. 6—Rear view of 1000-watt polyphase transmitter.

A more satisfactory method of maintaining the proper phase relation between the various components was found. A monitor receiver was installed at a distance from the antenna system and on a line 45 degrees from the two sideband antenna pairs. The intermediate-frequency output of this monitor receiver is transmitted back over a transmission line to the transmitter house and a cathode-ray oscilloscope used for an indicator. When the various components of the radio-frequency wave have the proper phase relations, it is possible to accomplish complete amplitude modulation. An adjustment procedure is to provide a modulating voltage, and adjust the phase of the radio-frequency input to the balanced modulators A and B until the maximum depth of modulation is obtained.

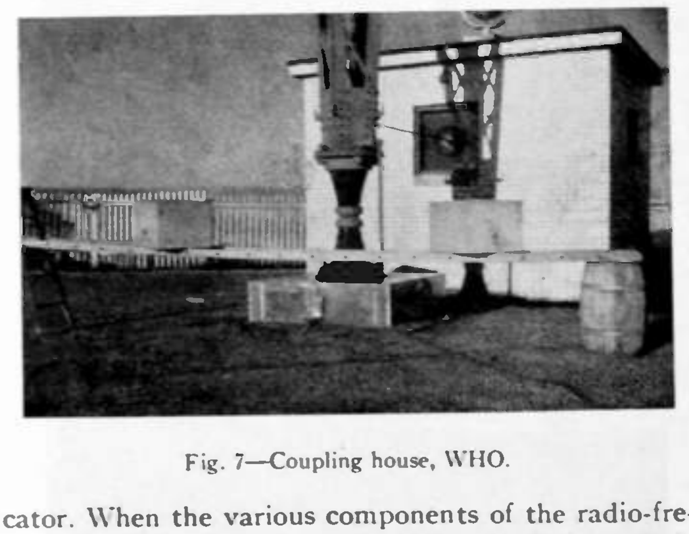

Fig. 7—Coupling house, WHO.

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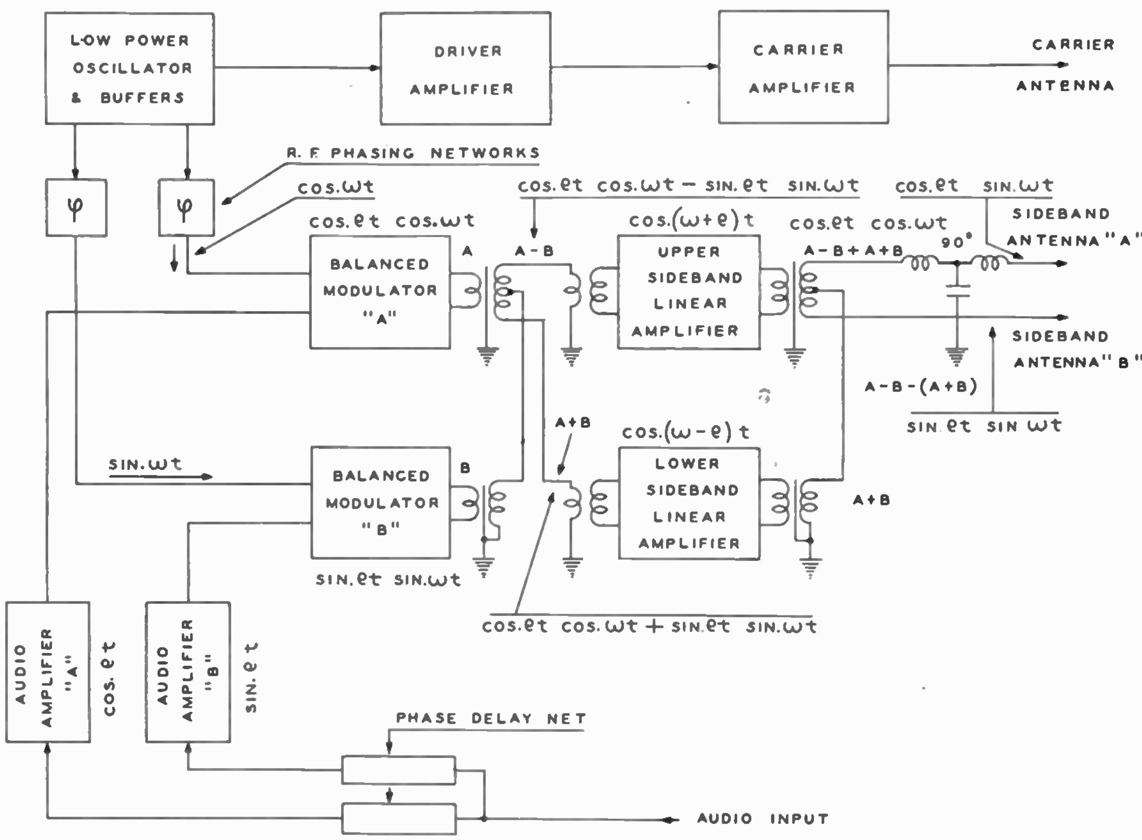

Fig. 4—Block diagram, polyphase transmitter.

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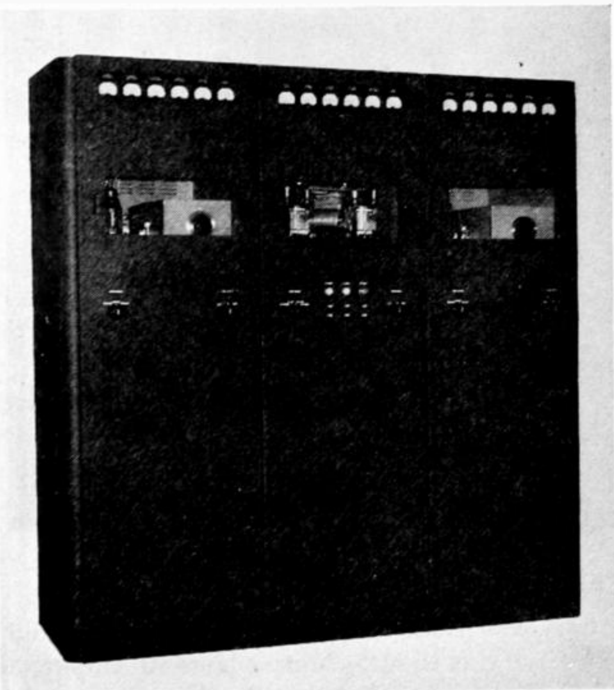

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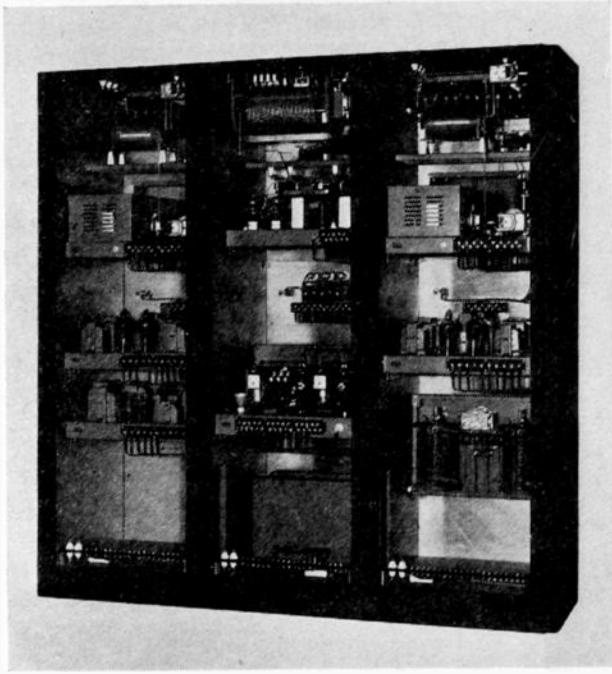

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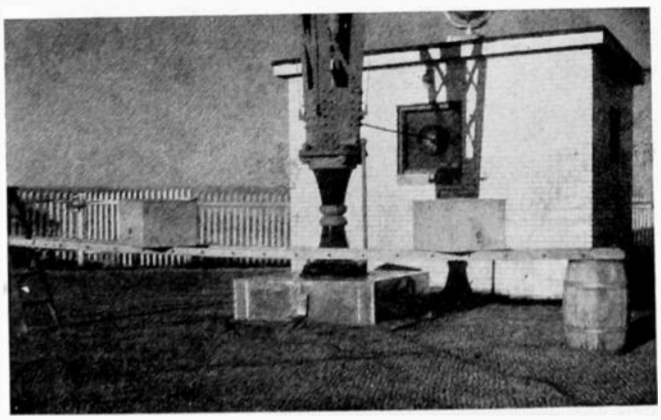

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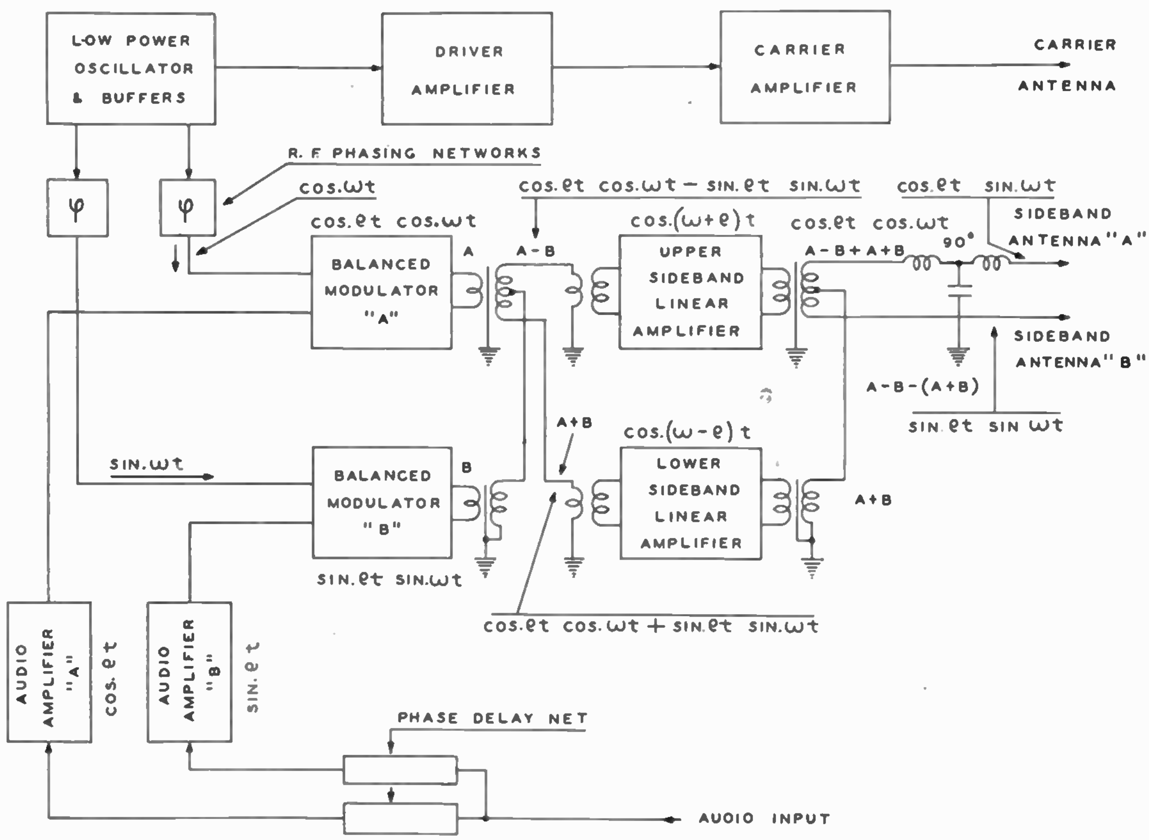

Fig. 4—Block diagram, polyphase transmitter.

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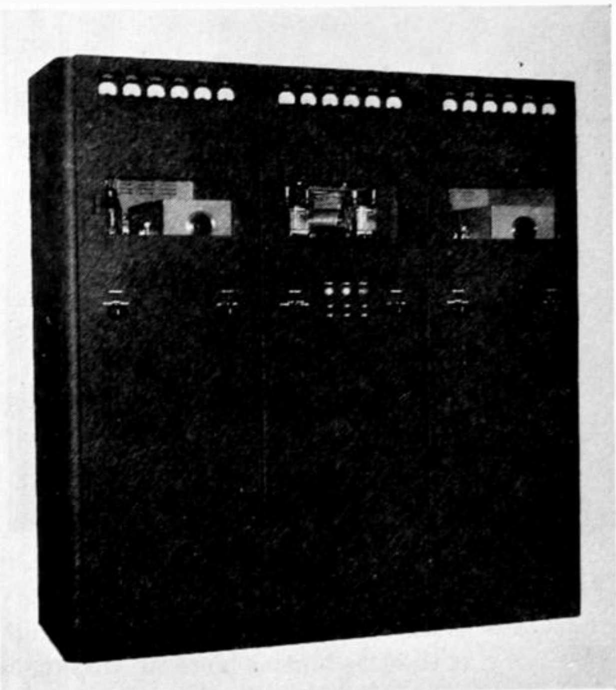

Fig. 5—Front view of 1000-watt polyphase transmitter.

Figs. 5 and 6 are photographs of the experimental 1000-watt transmitter. The center unit is the carrier amplifier and those adjacent, the two single-sideband amplifying units. Fig. 7 is a photograph showing the coupling house at WHO where the experimental transmitter was installed. The coupling units for matching to the sideband antenna pairs were temporarily supported at this time as a large ground screen was being installed around the base of the main radiator.

The Monitoring System

Another feature of the polyphase broadcasting system that is unique is the monitoring system. Inasmuch as the various components of the amplitude-modulated wave are supplied from independent radiators, not to mention independent amplifiers, the phase of these various radio-frequency components is quite arbitrary and dependent upon phase shifts in the transmission lines, coupling system, and antenna system. If these phase shifts were accurately known, a phase meter could be connected to the output of the carrier and sideband amplifiers and used for monitoring.

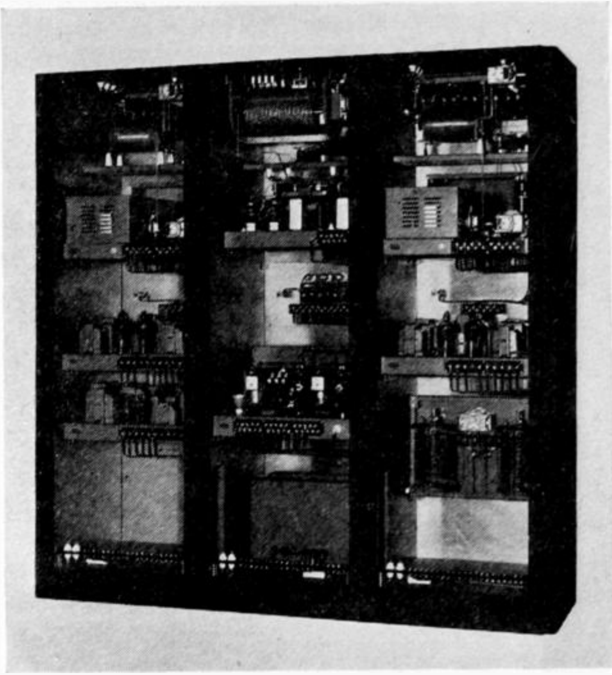

Fig. 6—Rear view of 1000-watt polyphase transmitter.

A more satisfactory method of maintaining the proper phase relation between the various components was found. A monitor receiver was installed at a distance from the antenna system and on a line 45 degrees from the two sideband antenna pairs. The intermediate-frequency output of this monitor receiver is transmitted back over a transmission line to the transmitter house and a cathode-ray oscilloscope used for an indicator. When the various components of the radio-frequency wave have the proper phase relations, it is possible to accomplish complete amplitude modulation. An adjustment procedure is to provide a modulating voltage, and adjust the phase of the radio-frequency input to the balanced modulators *A* and *B* until the maximum depth of modulation is obtained.

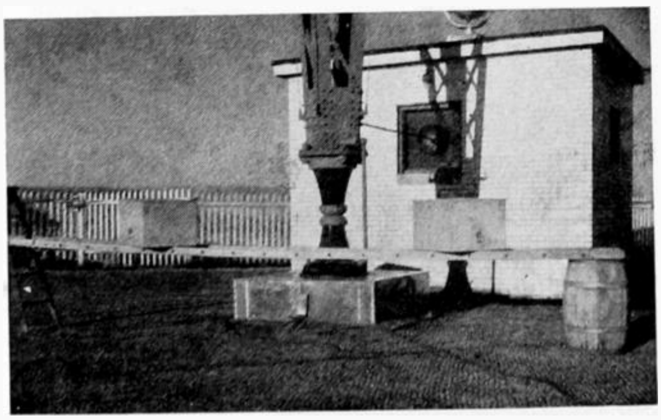

Fig. 7—Coupling house, WHO.

After a little experience, it was found possible to make this adjustment quite satisfactorily with regular program modulation in place of a single frequency tone.

The monitor-receiver location was approximately 1 mile from the antenna system, at which distance

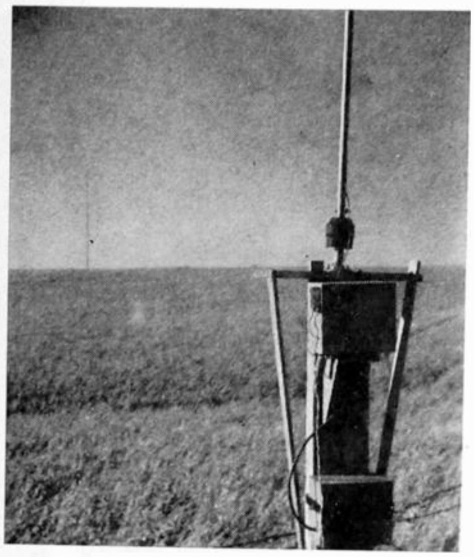

Fig. 8—Monitor receiver.

parallax due to physical dimensions of the antenna system is negligible. Fig. 8 is a photograph of the monitor receiver with the WHO antenna in the background. This monitor receiver was battery-operated and its output at intermediate frequency was transmitted to the transmitter house over a twisted-pair radio-frequency transmission line. To simplify transmission difficulties, an intermediate frequency of 195 kilocycles was used instead of the usual 456 kilocycles and it was also found necessary to insert at three points 1000-kilocycle traps in this line to prevent unwanted 1000-kilocycle pickup. The receiver employed a crystal oscillator, a converter tube and an untuned amplifier stage transformer-coupled to the transmission line. The transmission-line loss was high and it was necessary at the transmitter house to employ a 2-stage amplifier to raise the level sufficiently to operate a small cathode-ray tube. All data included in the latter part of this paper were obtained by visually phasing the antenna system by means of the monitor receiver, associated amplifier, and the cathode-ray tube.

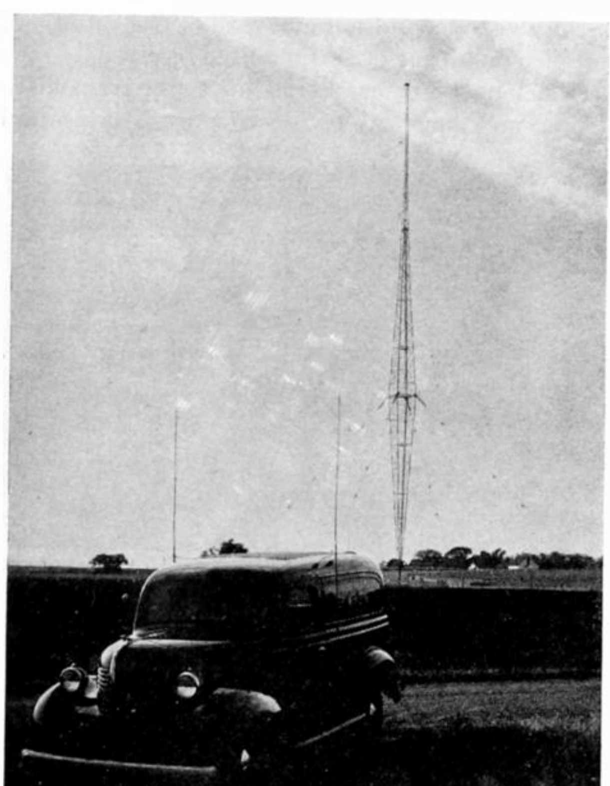

Fig. 9—Field car, housing measuring equipment.

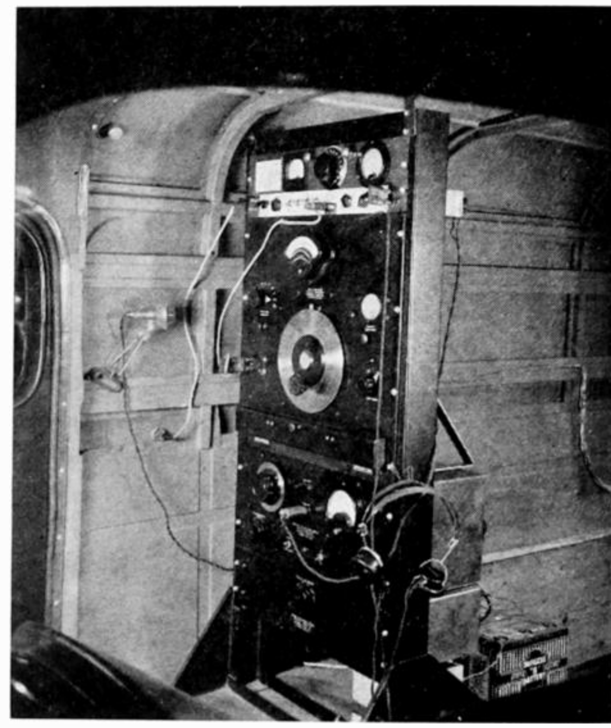

Fig. 10—Measuring equipment inside field car.

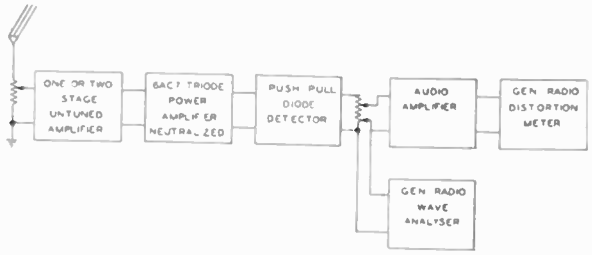

Fig. 11—Block diagram of measuring equipment.

The Measuring Equipment

In addition to the monitoring equipment, it was necessary to develop some special receiving apparatus for performance measurements of the transmitter. This apparatus was mounted in the truck shown in the photograph Fig. 9. Fig. 10 is a photograph of the equipment rack in the truck. A block diagram of

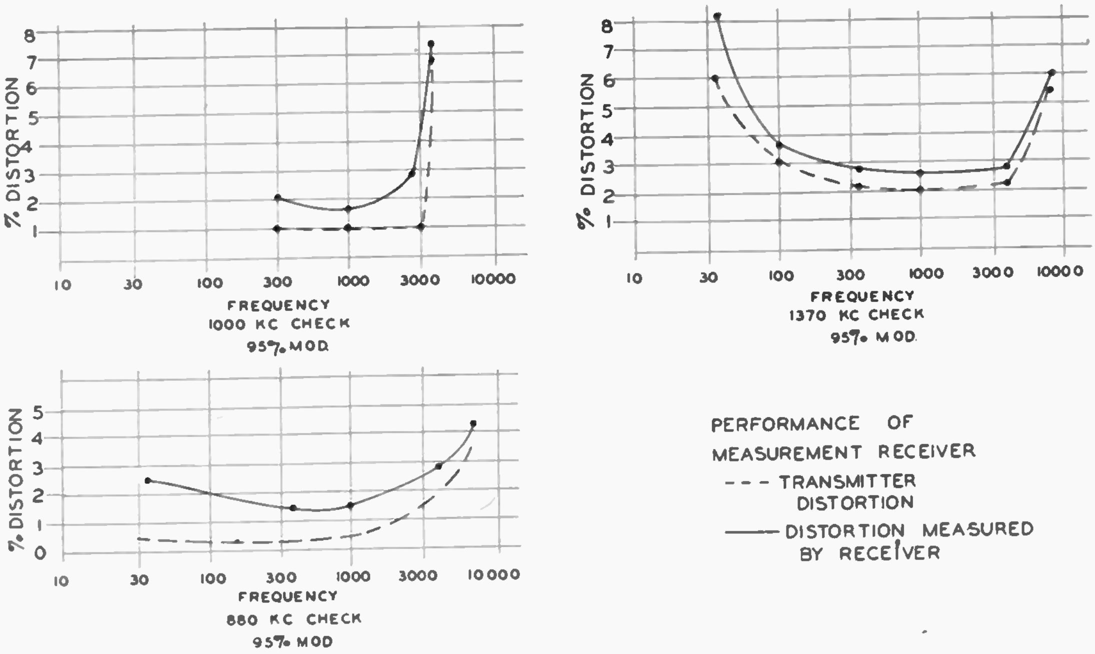

Fig. 12

the equipment installed for distortion-measuring purposes is shown in Fig. 11. Since very low orders of distortion were to be measured, the requirements for the receiver were quite severe, and one was built for the special purpose. It was arranged to operate either a General Radio wave analyzer or a General Radio distortion and noise meter and also was calibrated to indicate percentage modulation. A 12-foot telescoping vertical antenna was used working against the truck frame for the ground and terminated in a 10,000 ohm radio-frequency gain control. This method of gain control was used to avoid distortion at high signal inputs. A switch selects either one or two stages of reactance-coupled radio-frequency amplification designed for flat response to around 1200 kilocycles. A 6AC7 tube connected as a triode is coupled to a low-Q balanced and neutralized tank circuit which in turn is coupled to a push-pull diode detector. The diode load of 250,000 ohms is divided between a fixed resistor and a tapped attenuator supplying the wave analyzer and a single audio-frequency amplifier stage. A microammeter in the diode load circuit serves to indicate the carrier strength for a fixed radio-frequency gain setting and further assures that the triode stage is not being overloaded. While taking the measurements, the gain of the receiver was adjusted in each case so as to operate with a constant current in the linear detector. For this constant current, the volume indicator in the audio output was calibrated to indicate percentage modulation. The audio-frequency amplifier stage operates without a bypass condenser across the cathode resistor and the negative feedback thus present, results in reduced distortion. The complete equipment was battery-operated.

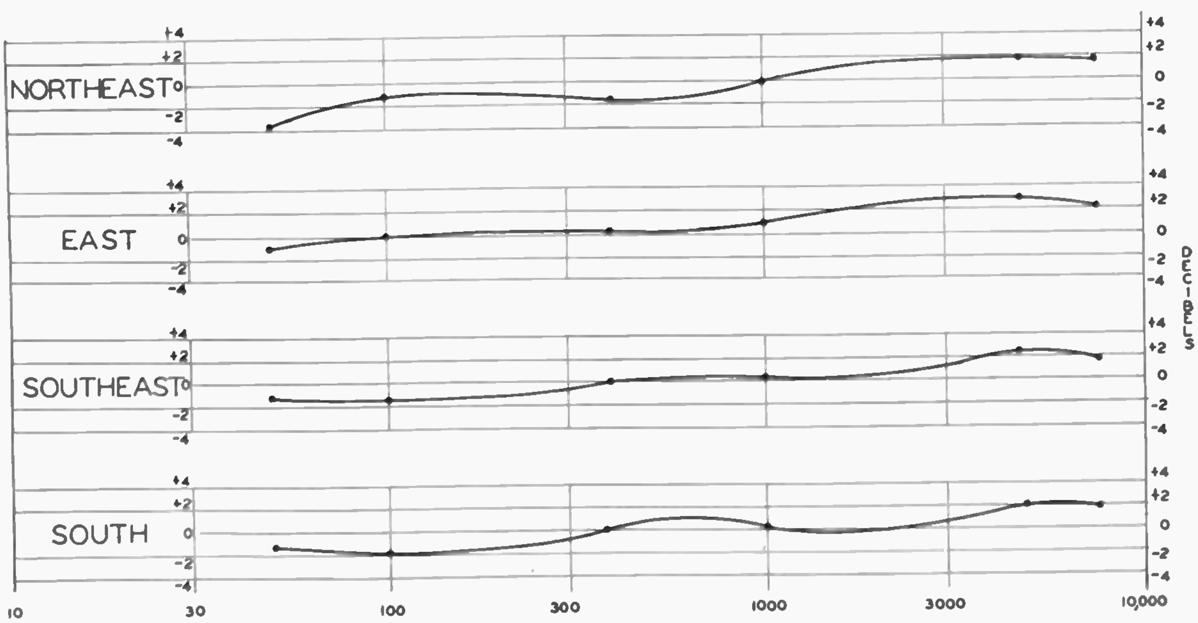

Fig. 13—Frequency response, polyphase experiment at WHO.

In order to facilitate the measurements, an ultra-high-frequency transmitter was installed on the truck and two-way contact was maintained during the course of the measurements.

The performance of the measuring equipment was checked by measuring response and distortion of several operating broadcast stations as well as a test oscillator and the results of these tests are shown in Fig. 12. Here the distortion figures as measured at the transmitter are compared with the measurements made in the field with the mobile unit. In general, while the readings made with the field equipment were somewhat pessimistic, the agreement is quite satisfactory.

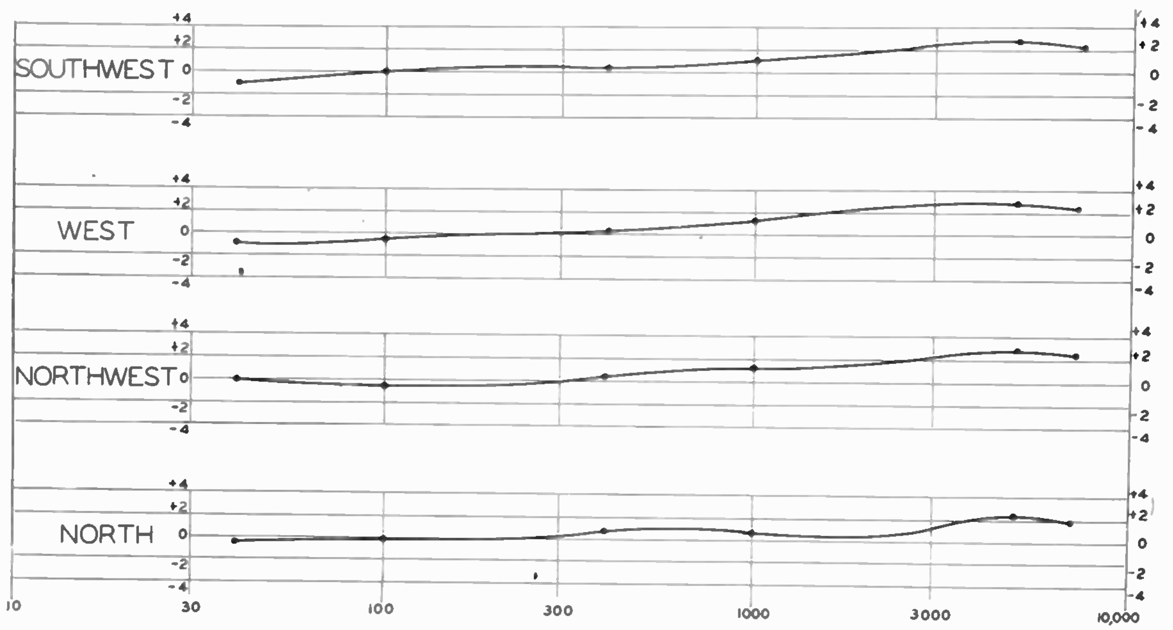

Fig. 14—Frequency response, polyphase experiment at WHO.

Performance Data

Measurements were taken of the frequency response and distortion characteristics of the polyphase-broadcasting system in various directions and at various distances from the antenna system. Measuring points were selected at distances of 1 to 3 miles from the antenna system and both in line with the sideband antenna pairs and along diagonal lines. At each measuring point, a complete test run was made including audio-frequency-response characteristics and amplitude-distortion measurements for modulating frequencies of 50, 100, 400, 1000, 5000, and 7500 cycles per second for various percentages of modulation.

The frequency-response characteristics in the eight directions are shown in Figs. 13 and 14. These response characteristics all show a slightly rising characteristic due to the fact that the audio-frequency phase-shift networks introduce somewhat less loss as the frequency is increased (Fig. 15). A corrective network placed ahead of the audio-frequency phase-shift networks would correct the response to within plus or minus $1\frac{1}{2}$ decibels. Further, these curves show a slight variation in response correlating with the phase-difference characteristic of the two phase-shift networks. Since these response characteristics were all taken at a constant input level at the transmitter, they show the percentage modulation to be essentially the same in all directions from the radiating system.

Fig. 16 shows the results of the measurements of the amplitude distortion. For each percentage of modulation, the root-sum-square amplitude-distortion measurements were averaged for the eight directions. The results of these measurements indicate that high-fidelity performance can be obtained with polyphase broad-

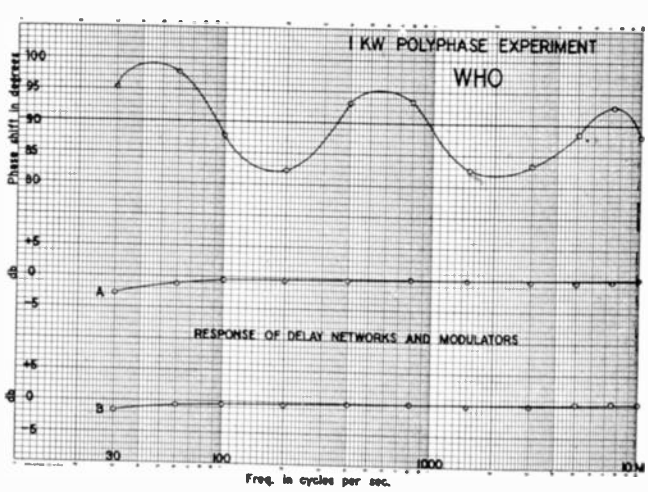

Fig. 15—Response of delay networks and modulators.

casting. These curves which represent the over-all contributions of the transmitter, tuning elements, and measuring receiver show performance quite comparable to conventional amplitude-modulation broadcast transmitting equipment. It is interesting to note that by adjusting the phase of the radio-frequency excitation to balanced modulators A and B, the distortion could be reduced to a value below the error in the measuring equipment for any single modulating frequency from 10 to 10,000 cycles per second. When the adjustment was such as to give lowest distortion at the low modulation frequencies, the distortion was progressively greater at higher modulation frequencies. By adjusting for minimum distortion at 10,000 cycles per second, the distortion was high at the low-frequency end. The measurements were taken for a setting which resulted in minimum distortion at a frequency near the middle of the audible range.

ment. The experience gained in operating equipment at this power level with an antenna system approximating that which might be used for a permanent installation indicates some of the points which could be modified to advantage in a final installation.

A modification of the monitoring system would be considered in an actual high-power installation. The monitoring locations would for a high-power transmitter receive enough signal intensity to permit transmission of the receiver energy back to the control room by means of a small concentric transmission line. Further, it seems desirable to use two monitoring locations,

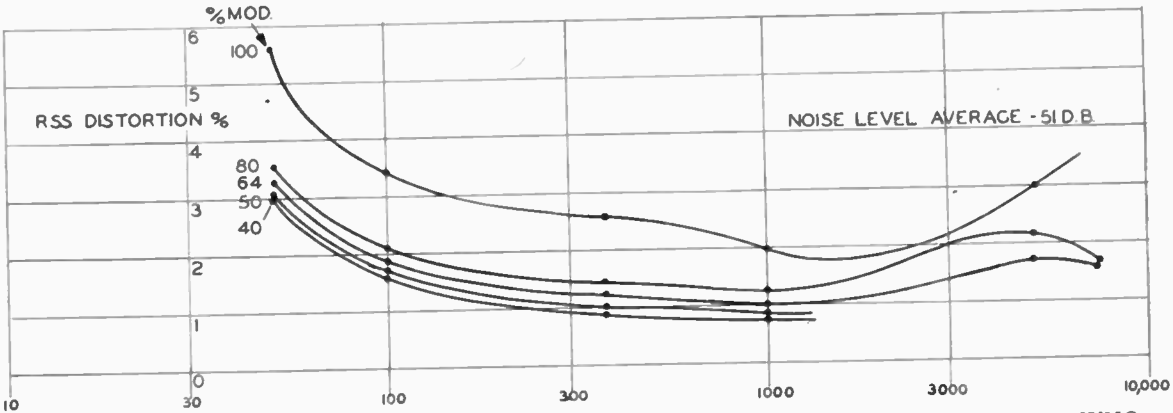

Fig. 16—Average root-mean-square distortion measured in eight directions. Polyphase experiment at WHO.

The power input to the sideband antennas was measured at 100 per cent modulation and found to be somewhat in excess of the theoretical figure, namely, 50 per cent of the carrier power. This difference between the actual and the theoretical power is almost entirely accounted for by the radio-frequency loss resistance of the sideband antenna pairs and in the sideband antenna tuning units. The antenna conductors were 3/16-inch phosphor-bronze and due to the fact that the radiation resistance of a sideband antenna pair is rather low, the radio-frequency resistance of these conductors becomes of more importance. It was necessary to provide a total sideband power equal to approximately 80 per cent of the carrier power in order to produce 100 per cent modulation. Of the excess 30 per cent, 17 per cent can be accounted for by the estimated radio-frequency resistance loss in the sideband antennas, 8 per cent in coupling networks to the sideband antennas, and the remaining 5 per cent is probably due to the fact that the sideband antennas are somewhat shorter than the main radiator.

Conclusions

To summarize the test results, it was definitely shown that the system of polyphase broadcasting is capable of high-fidelity operation, and the power measurements indicate that the expected gains in tube and power economy can be realized. During the experimental period, musical programs were broadcast at different times. Listening tests did not detect any lack of fidelity of the over-all performance of the equip-

one for each sideband pair. Using two monitoring locations, the signal from each monitoring location could be used to operate a modulation monitor to observe the action of both sideband pairs simultaneously.

The sideband antenna pairs used for this experimental work had a Q of about 50. This was reduced slightly during the work by widening the spacing. The resistance of a sideband antenna pair was 73 ohms at 1000 kilocycles and the resistance had dropped to about 31 ohms at 1010 kilocycles. This made matching impossible at all sideband frequencies. This mismatch lowered the efficiency of the coupling units to the sideband antenna pairs at the higher modulation frequencies and made it impossible to maintain correct phase relation of the suppressed carrier of the sideband currents to the carrier current for all sideband frequencies. The above difficulties can be minimized by using a slightly wider spacing for the sideband antenna pairs and by using separate tower sideband antennas or otherwise reducing their loss resistance and decreasing their characteristic impedance by increasing their cross section. From results obtained from the two spacings used for the sideband pairs, and from approximate mathematical calculations, a spacing up to 30 electrical degrees and the use of a cage or tower construction for each element of the sideband pairs would make it possible to maintain a low value of amplitude distortion at all modulating frequencies.

An improved design of the audio-frequency phase-shift networks has been completed which would result in maintaining the phase difference within 3 degrees of the 90-degree value over the range of 20 to 12,000 cycles per second. The use of this network with proper equalization to compensate for the slightly higher attenuation of the low audio frequencies would result in a further improvement in performance. Several other refinements dealing with the performance and efficiency of individual units have been given preliminary test in the laboratory and can be applied with no change in the general system here described.

ACKNOWLEDGMENT

The author wishes to express appreciation to the technical staff, and in particular Franklin W. Pierce, of the Central Broadcasting Company; Messrs. John Byrne, L. M. Craft, and W. H. Wirkler of the Collins Radio Company, Cedar Rapids, Iowa.

Short-Wave Spread Bands in Automobile and Home Receivers*

DUDLEY E. FOSTER†, MEMBER, I.R.E., AND GARRARD MOUNTJOY‡, MEMBER, I.R.E.

Summary—*The use of spread bands in both automobile and home receivers has increased greatly of late. This paper discusses the gain and selectivity characteristics of simple signal-frequency circuits having good performance when used with a short antenna. Oscillator circuits employing few switching elements are considered also. It is pointed out that in the majority of designs continuous tuning of the oscillator circuit only is required.*

The inclusion of short-wave bands in home receivers has been common for years, and has recently extended to automobiles. The increase in power of domestic and foreign stations, the increase in program schedules, and the war situation have created more possibilities and greater interest in having short-wave reception in automobiles. Information has recently been published on the quality of the service, in respect to field strengths, noise, fading, etc., over the country,[1] but it seems likely that at least several manufacturers will supply receivers, from which field experience will be accumulated. It is the object of this paper to describe spread-band circuits useful in automobile receivers, in which they are required in order to have sufficient ease of tuning. They will be useful in home receivers also, where their inclusion is desirable although not essential.

The bands usually of interest are

Band megacycles	Band Extent kilocycles
6.0– 6.2	200
9.5– 9.7	200
11.7–11.9	200
15.1–15.35	250
17.7–18.00	300

There are also a few stations assigned 7.2 to 7.3 megacycles, which range may be added if desired.

These bands are only 20 per cent of the width of the standard broadcast band so that a selector system in which each short-wave band is spread over the entire dial scale becomes easier to tune than the standard broadcast range.

PRESELECTOR CIRCUITS

Since the extent of each short-wave band is small compared with the center frequency, it is rarely worth while to provide continuous tuning for the signal-frequency circuits. Each such circuit may then be fixedly adjusted to the center of the respective bands and the entire station selection within the band be accomplished by tuning of the oscillator only.

A preselector circuit for use on short-wave bands of automobile receivers should possess certain characteristics; namely, 1. discrimination against spark noise, 2. reasonable gain when used with a low-capacitance antenna, 3. freedom from microphonism, and 4. freedom from the effect of slight changes in antenna capacitance (since some antennas are variable in length) on the oscillator frequency or on preselector gain.

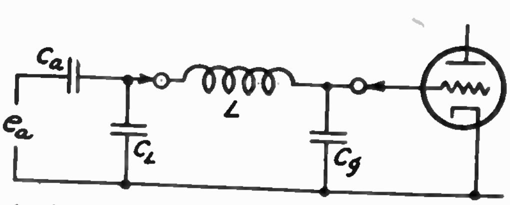

Fig. 1—Antenna input circuit utilizing one tuned circuit.

The same requirements, except for spark noise, also apply to household receivers. Since, in general, the automobile requirements are the more stringent, antenna constants typical of that service have been chosen for illustration. Suitable modification of circuit constants may be made readily when the circuits described are used on home receivers.

One type of preselector which fulfills these requirements to an appreciable extent is shown in Fig. 1. Where Ca = capacitance of that portion of the antenna exposed to signal pickup

* Decimal classification: R361. Original manuscript received by the Institute, August 28, 1941. Presented, Summer Convention, Detroit, Michigan, June 23, 1941.
† Formerly, RCA License Laboratory, New York, N. Y.; now, Rogers-Majestic (1941), Ltd., Toronto, Ont., Canada.
‡ RCA License Laboratory, New York, N. Y.
[1] J. H. Little and F. X. Rettenmeyer, "A five-band receiver for automobile service," PROC. I.R.E., vol. 29, pp. 151–166; April, 1941.

C_L = capacitance of lead-in and associated components
L = resonating inductor
C_g = input capacitance of the 1st tube and any desired added capacitance in shunt therewith
e_a = induced voltage in the antenna, and
Q = figure of merit for entire circuit,

the gain at resonance for this system may be readily derived as

$$g_0 = \frac{C_a Q}{C_a + C_L + C_g}. \tag{1}$$

The effective circuit Q will be influenced to a large extent by the damping of the lead-in and antenna resistance. If the inductor L is made large and C_g correspondingly made small the effects of lead-in and antenna damping will be minimized, since their combined reactance is thus made small compared to the reactance of the inductor and C_g. A high figure of merit may readily be obtained for L and C_g, and the over-all circuit Q may thus be increased.

By this same procedure of increasing L, the relative importance of changes in antenna capacitance value are minimized and detuning from resonance with attending loss in preselector gain and interlock (detuning) of the oscillator circuit are reduced. While in many applications C_a may be expected not to vary, there are other installations in which antenna length is made adjustable by the user.

The gain at center frequency of (1) becomes

$$g_0 = 0.3Q$$

when

$$C_a = 25 \text{ micromicrofarads}$$
$$C_L = 35 \text{ micromicrofarads}$$
$$C_g = 25 \text{ micromicrofarads.}$$

This will be true for all bands. A plot of gain (calculated) versus frequency is shown in Fig. 2 for three choices of circuit Q, namely, 100, 50, and 25. Response throughout each band is indicated.

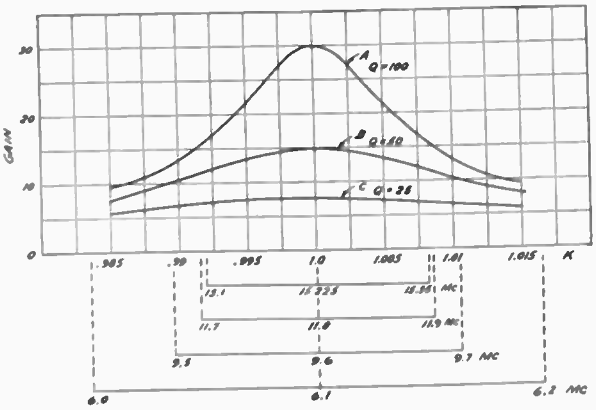

Fig. 2—Single-circuit selectivity for the four principal short-wave broadcast bands. Three values of Q.

When two tuned coupled circuits are used, a higher and more constant-gain-versus frequency response is obtained since the additional circuit may be made one of lower capacitance and high Q. The switching arrangement need not be more complicated than that for the single-tuned circuit. One system is shown in

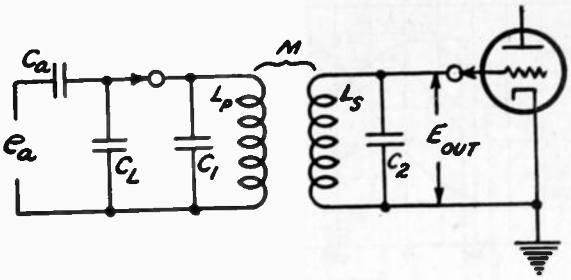

Fig. 3—Antenna input circuit consisting of two coupled tuned circuits.

Fig. 3. In this the gain at center frequency is

$$g_0 = \frac{C_a\sqrt{Q_1Q_2}}{\sqrt{C_2(C_a + C_L + C_1)}} \times \frac{S}{1 + S^2} \tag{2}$$

where the symbols and their arbitrarily assigned values are

C_a = internal capacitance of the antenna, 25 micromicrofarads
C_L = lead-in capacitance, 35 micromicrofarads
C_1 = 10 micromicrofarads
C_2 = 25 micromicrofarads
L_p resonates with $C_1 + C_L + C_a$ at center frequency
L_s resonates with C_2 at center frequency
S = ratio of coupling used to optimum coupling where optimum coupling is

$$M = \sqrt{\frac{L_pL_s}{Q_1Q_2}}$$

Q_1 = effective figure of merit of all primary components
Q_2 = effective figure of merit of all secondary components.

Attenuation for frequencies very near center frequency is

$$\frac{g_0}{g} = 1 - \frac{Q_1Q_2{}^{(k-1/k)^2}}{1 + S^2} + \frac{J(Q_1 + Q_2)^{(k-1/k)}}{1 + S^2} \tag{3}$$

where

$$k = \frac{\text{frequency considered}}{\text{center frequency}}.$$

A plot of gain versus frequency is shown in Fig. 4 for the case of

$$Q_1 = Q_2 = 100.$$

Several magnitudes of coupling are considered, namely, $S = 1$ (optimum), $S = \sqrt{3}$, and $S = 2.5$. The increase in performance over that indicated in Fig. 2 is quite marked. The degree of coupling may be chosen for each band to give the best average gain.

Another plot is shown in Fig. 5 for lower values of Q_1, where

$$Q_1 = 25$$
$$Q_2 = 150.$$

Again values of coupling of $S=1$ (optimum), $S=\sqrt{3}$, and $S=2.5$ are taken. These results may be compared to those of Fig. 2(c) for the case of the single tuned circuit with Q of 25. Approximately twice the gain of

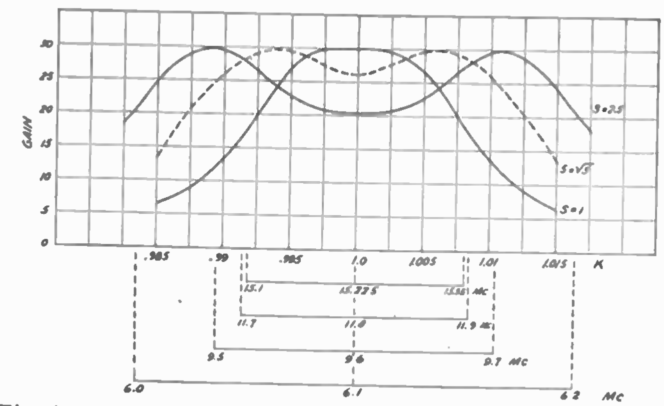

Fig. 4—Coupled-circuit selectivity for the four principal short-wave broadcast bands. Three degrees of coupling.

the single tuned circuit is obtained by two circuits coupled at $S=\sqrt{3}$, and $Q_1=25$ in both cases.

The use of two resonant circuits greatly reduces any interlocking effect between antenna and oscillator for changes in antenna capacitance. Elimination of interlock is particularly important when an antenna of variable length is used.

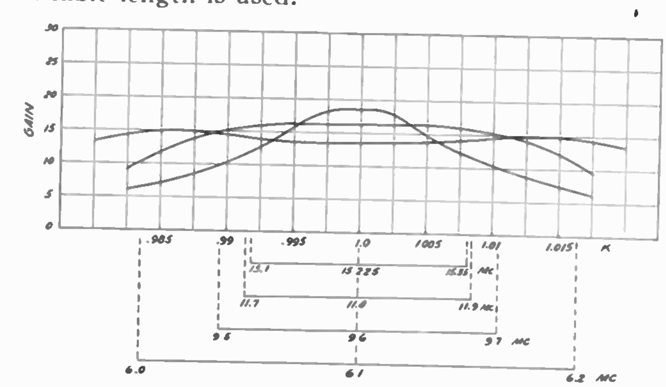

Fig. 5—Coupled-circuit selectivity for the four principal short-wave broadcast bands. Three degrees of coupling.

Radio-Frequency Stage

In the case of a tuned impedance-coupled radio-frequency stage, as shown in Fig. 6, the gain may be expressed for center frequency as

$$g_0 = \frac{g_m Q}{\omega C}$$

where C = total capacitance

Q = effective circuit Q, all damping effects considered.

Any usually desired gain may be obtained with this type of circuit. For example at 6.1 megacycles a gain of 33 may be realized with a g_m of 2000 μmhos, C of 40 micromicrofarads, and Q of 25. This gain should satisfy most design requirements. Similar gains may be had at the higher-frequency bands by increasing Q. Selectivity will be the same, near center frequency, as that shown in Fig. 2 for the single antenna circuit.

Gain and band coverage may be controlled by circuit damping. The radio-frequency stage is instrumental in isolating the antenna from the oscillator circuit with resulting elimination of interlock. If a radio-frequency stage is used, the antenna circuit may be of the single-tuned type unless the extra gain obtainable from the two-resonant-circuit system is desired.

There seems to be no advantage from the standpoint of gain in using two tuned circuits in the radio-frequency stage since good gain may be obtained with the single circuit.

When no radio-frequency stage is used the choice of a preselector circuit becomes increasingly important.

The double-tuned circuit, in addition to reducing interlock between oscillator and antenna circuit, will

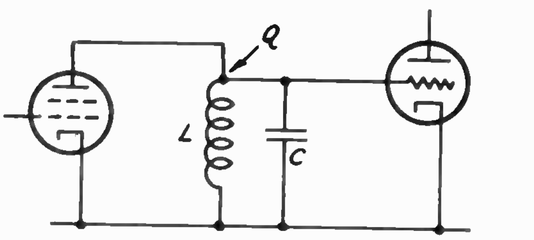

Fig. 6—Radio-frequency stage with single tuned output circuit.

maintain a higher gain for changes in antenna capacitance than will the single-circuit type of preselector.

When changes in antenna capacitance are expected, the degree of coupling between the two tuned circuits may be made large. Not much loss in average gain throughout the band will result and the improvement in performance over the single-tuned circuit for, say, a 50 per cent change in antenna capacitance, is large.

In the 15-megacycle band, this assumed antenna capacitance change would shift the single-resonant circuit of Fig. 1 an amount corresponding to the intermediate frequency, 450 kilocycles, approximately. If the oscillator frequency were higher than the signal, the interlock would be maximum.

In the case of the two-resonant-circuit system, the frequency shift of the tuned grid circuit produced by the 10-micromicrofarad change in antenna capacitance is relatively small. Detuning of the grid circuit will be a maximum when the change in antenna capacitance produces a series-tuned reactance in the first tuned circuit equal to the series-resistive component of the first circuit. For a system where the coupling factor S is less than 3, the grid-circuit resonant frequency will not be shifted in either direction by a frequency increment in the antenna branch as large as the intermediate frequency. Thus the worst condition of interlock, i.e., converter grid-tuned frequency equal to the oscillator frequency, may be prevented by the use of the two-circuit system.

Oscillator Tuning

From the idealistic viewpoint, it is desirable to use the full dial scale for each spread band. This can be done providing oscillator frequency drift is substantially eliminated. As the bands are narrow and the oscillator frequency is high, it may be difficult and costly to restrict frequency drift to a sufficiently low level to enable accurate calibration and complete spreading of the band.

Some convenience and economy may result in providing a tuning range greater than the bandwidth, so that all stations will still be received in spite of a moderate amount of oscillator drift.

There are numerous oscillator-tuning systems which may be considered for use. Several of these methods are discussed herein.

If the receiver is designed with capacitance tuning in the broadcast band, a somewhat expensive but good method would be to include a separate small oscillator section in the gang condenser of one stator and one

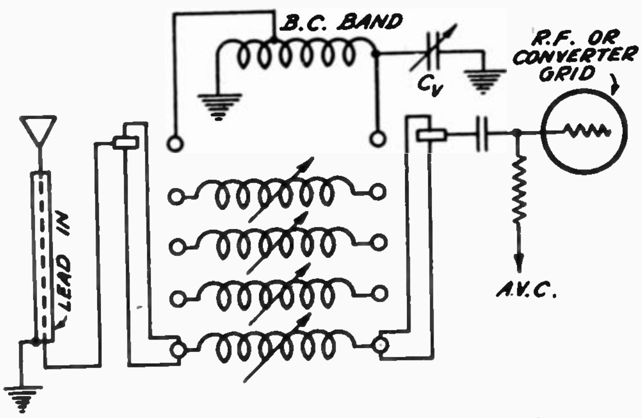

Fig. 7—Antenna circuit (single tuned stage).

rotor plate. A possible simplification might be made by the use of one stator plate adjacent to the outside rotor plate of the broadcast-oscillator section. One of the chief advantages of this method of tuning is the even distribution of stations over the dial.

The use of a padding capacitor in series with the broadcast-tuning section will provide the necessary restricted tuning range, but will result in some station crowding at the high-frequency end of the dial. The amount of crowding may not prove too detrimental since separation between channels on the dial will still be appreciable.

If the receiver is designed with inductance tuning in the broadcast band, an extra oscillator-tuning inductor of small inductance value and small percentage change may be used in conjunction with the main tuning mechanism and tuning knob. This tuning inductor will then be used in shunt with the oscillator circuits for each spread band.

The spread bands vary in percentage change in frequency. For example the lower band, 6.0 to 6.2 megacycles, undergoes a change of approximately 3 per cent in oscillator frequency, while the 15.1- to 15.35-megacycle band varies only 1.6 per cent.

If these bands are tuned by the same variable reactive element, one will occupy more dial space than the other. If it is desired to equalize the band spreading more nearly in relation to the dial coverage, several means suggest themselves.

For example, reactances may be put in series with the tuning element to control the dial coverage of each

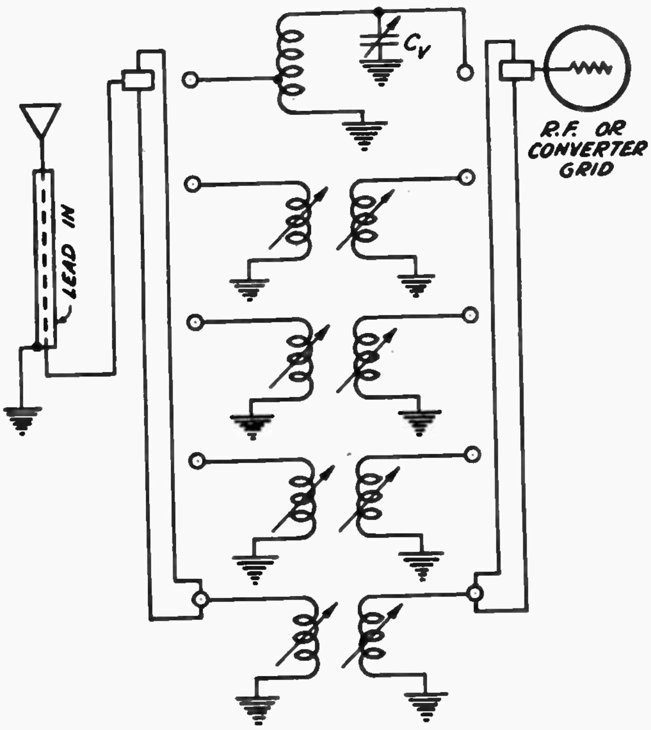

Fig. 8—Antenna circuit (two coupled circuits).

band; again, the total fixed circuit capacitance may be different for each band to provide similar dial travels.

Applied Circuits

Circuit requirements vary for each design problem, and any recommendation of specific circuits, complete with tuning and band-switching means, is of course outside the scope of this paper. Figs. 7 to 11, inclusive,

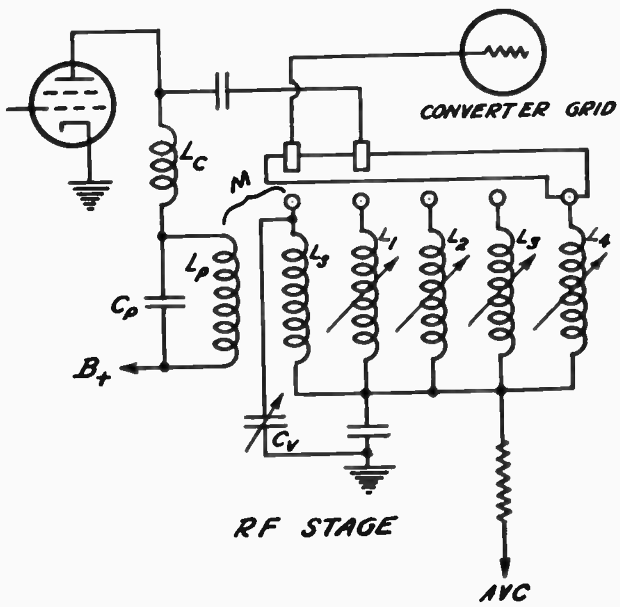

Fig. 9—Radio-frequency stage.

are offered as a suggestion of five out of an infinite possibility of combinations.

Figs. 7 and 8 show antenna networks of one and two resonant circuits per spread band, respectively. Either of the switching arrangements shown permits the use

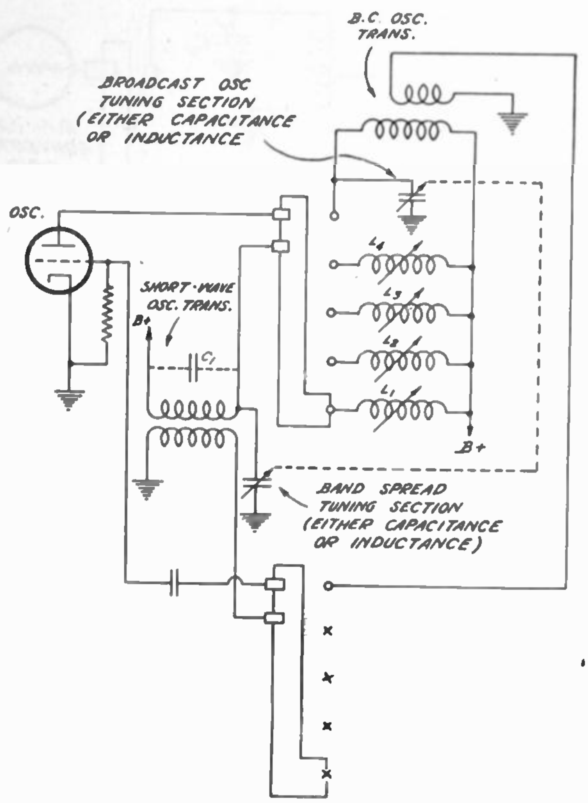

Fig. 10—Oscillator circuit with band-spread tuning element.

of any broadcast-band preselector circuit such as the one shown, or for example one using a primary coupled to the grid tuning inductor. Spread-band resonating inductors are indicated as the adjustable elements since this makes for a compact design. Trimming capacitors may be used instead, if desired, with an increased space requirement.

A suggested radio-frequency-stage circuit is shown in Fig. 9. It is one which requires only one switching section. The use of the radio-frequency choke L_c and primary capacitor C_p may not be found necessary or desirable. If not, this system is one requiring very few components.

Fig. 10 shows an oscillator circuit using a band-spread tuning element (either capacitance or inductance to work in conjunction with the type of broadcast tuning used). More or less spread may be obtained on each band by the choice of oscillator transformer and associated resonating capacitance C_1. Equalization of band spread for all bands may be accomplished by the use of capacitors across the trimming inductors L_1, L_2, L_3, and L_4 for applications wherein this extra cost would be warranted.

Fig. 11 shows an oscillator circuit similar to Fig. 10 except for the use of a padding capacitor in series with the variable tuning capacitor. Performance of this circuit is quite similar to that of Fig. 10.

Inductance Tuning for Spread Bands

Variable-inductance tuning, where indicated for spread-band service, may be accomplished by any of several methods, one of which is the movable-copper-core method. The small changes in inductance required to cover the spread bands may be accomplished without introduction of appreciable damping. In general, the losses introduced by a copper core are relatively less at high frequencies (15 megacycles) than at the lower frequencies (6 megacycles). Fortunately, the selectivity requirements are least in the 6-megacycle band since a greater percentage of frequency coverage relative to center frequency is necessary there. Damping effects due to the copper core will result in less than

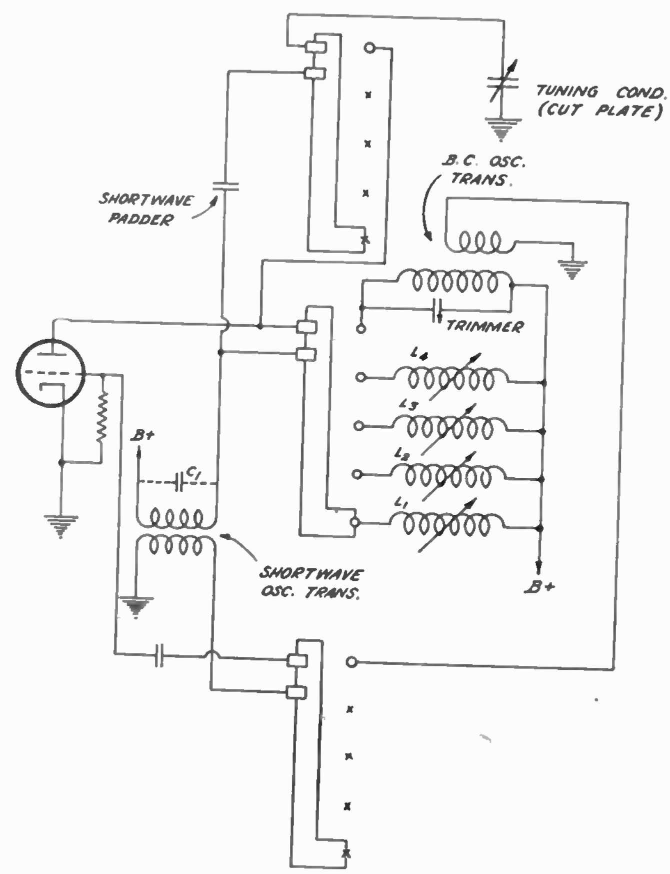

Fig. 11—Oscillator circuit with band-spread padding capacitor.

10 per cent loss in Q for any of the band-spread cases discussed herein.

Conclusions

Preselector circuits have been discussed which will meet the requirements of good receiver sensitivity and provide sufficiently uniform gain throughout each spread band. The systems indicated are only a few of many possible circuits, but the calculated results may be useful as a check of the performance of these or any other systems tried. The descriptions have been confined to fixed tuned preselectors as it is felt that adequate performance may be obtained without the use of variable preselector tuning. However, there may be applications where variable preselector tuning is desirable.

When the receiver is used in conjunction with an antenna of adjustable length and consequent variable capacitance, the use of a radio-frequency stage to prevent interlock between oscillator and preselector is advantageous. When no radio-frequency stage is used, the incorporation of a two-resonant-circuit preselector has the advantage of freedom from interlock in addition to increased antenna gain over the single-resonant antenna-circuit type.

Oscillator circuits are discussed and examples of band-selecting systems are shown.

While the subject of spread bands is treated herein for automobile receivers specifically, it is felt that other applications of spread-band tuning have as much or more utility.

Horizontal-Polar-Pattern Tracer for Directional Broadcast Antennas*

F. ALTON EVEREST†, ASSOCIATE, I.R.E., AND WILSON S. PRITCHETT‡, STUDENT, I.R.E.

Summary—*Making space-pattern calculations for the 3-element array is very much more difficult than for the more common 2-element array. Many problems of broadcast coverage or interference have their solution only in the nonsymmetrical patterns of 3-element directional antennas. This paper includes the derivation of the equations for the relative field strength in a horizontal plane of such an array. A mechanical tracer is described which automatically plots the horizontal pattern for 2- or 3-element arrays once the adjustments of tower configuration and current magnitudes and phasings are made. Patterns traced by this machine are shown to have an accuracy usually within the width of the line of the recording pen. Its simplicity and rapidity of operation adapt it particularly to preliminary exploration to find a pattern which meets certain coverage or interference problems.*

I. Introduction

THE USE of directional antennas by radio broadcasting stations is increasing rapidly. "By September of 1940, there were 116 directional antennas in use by United States stations in the standard broadcast band, and this figure probably approximates 140 as of the end of the year. At least 50 additional directional antennas will be required under the terms of the Havana Treaty of December, 1937. The majority of these directional antennas are employed at regional stations and are designed so that the station's nighttime power can be increased from 1000 to 5000 watts without increasing the interference to other stations operating on the same frequency. However, at least 11 transmitters of 50,000 watts are now using directional antennas. It is interesting to note that many of these antennas involve elaborate systems utilizing three or four radiating elements and that vertical as well as horizontal directivity is considered."[1]

This definite trend toward more complicated radiating systems cannot be denied, nor can the increased labor for the engineer called upon to design such a system. The calculations involved in the simple 2-element array are quite elementary and no small degree of assistance is obtained from the directive diagrams which have been published.[2] In going from a 2- to a 3-element array, however, the minimum number of variables required to describe the horizontal directivity increases from three to seven. As a result, the mathematical statements become unwieldly and their solution quite laborious. It is felt that a mechanical calculator and pattern tracer such as described in this paper or a large collection of 3-element patterns derived therefrom will be very valuable to the antenna designer.

II. Derivation of Equation

At any point P on a horizontal plane passing through the base of a vertical tower antenna, the field strength is directly proportional to the current in the antenna. With an array of identical towers, the field strength is the vector sum of the contributions from each of the individual elements. These voltage vectors have lengths proportional to the currents in their respective elements and angular relationships determined by the relative phasings of the currents in the elements and the difference in path length.

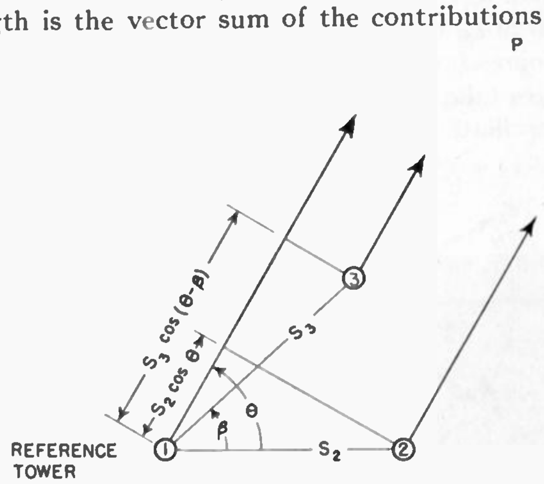

Fig. 1—Geometry and nomenclature for the 3-element array.

* Decimal classification: R125. Original manuscript received by the Institute, September 3, 1941. Presented, Pacific Coast Convention, Seattle, Washington, August 2, 1941.

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[1] "Radio progress during 1940, Part II—radio transmitters and transmitting antennas," Proc. I.R.E., vol. 29, pp. 94–95; March, 1941.

[2] R. M. Foster, "Directive diagrams of antenna arrays," *Bell Sys. Tech. Jour.*, vol. 5, pp. 292–307; April, 1926.

Assume that point P is at a distance from the array great compared to the element dimensions so that signals from each of the three towers of Fig. 1 to point P will follow essentially parallel paths. Let us consider only the directive characteristics of the array, i.e., neglect all propagation effects. Referring to Fig. 1, let us investigate the voltage received at point P from the individual elements. Let the current in tower 2 lead the current in tower 1 by the angle ϕ_2 and the current in tower 3 lead the current in tower 1 by the angle ϕ_3, then the relative signal e_2 from tower 2 will make the angle $S_2 \cos\theta+\phi_2$ with reference to the vector $e_1 \underline{|\alpha_1}$, (note that e_1 alone is not a vector) and the signal e_3 from tower 3 will make an angle $S_3\cos(\theta-\beta)+\phi_3$ with reference to the vector $e_1\underline{|\alpha_1}$. Summarized,

$$
\begin{aligned}
E_1 &= \text{relative signal from tower 1 at point } P = e_1 \underline{|\alpha_1} \\
E_2 &= \text{relative signal from tower 2 at point } P = e_2 \underline{|\alpha_2} \qquad (1)\\
E_3 &= \text{relative signal from tower 3 at point } P = e_3 \underline{|\alpha_3}
\end{aligned}
$$

where

$$
\begin{aligned}
\alpha_1 &= 0 \text{ degrees} \\
\alpha_2 &= (S_2\cos\theta) + \phi_2, \text{ degrees} \\
\alpha_3 &= S_3\cos(\theta-\beta) + \phi_3, \text{ degrees.}
\end{aligned}
$$

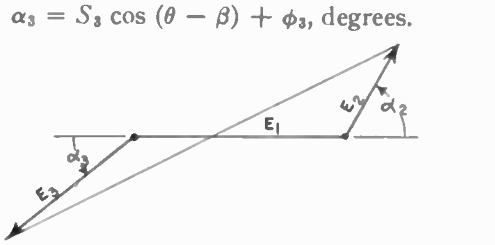

Fig. 2—Method of adding the contributions from the 3 array elements upon which the instrument described is based.

The vector diagram of Fig. 2 illustrates the relationship of these three vectors in a manner which will be useful later. The current ratio I_2/I_1 determines the length of E_2 and I_3/I_1 determines the length of E_3. As the point of observation P is moved around the array, changes take place in the angles α_2 and α_3. Vector E_2 will oscillate between the limits E_2' and E_2'' the extent of this oscillation being determined by the spacing S_2 as illustrated in Fig. 3. The mean point about which E_2 oscillates is determined by the current phasing ϕ_2. In a similar manner E_3 oscillates as P sweeps around the array, the extremities of the oscillation being determined by S_3 and the mean point by ϕ_3. The vectors E_2 and E_3 will each complete a cycle as the observation point P moves once around the array (i.e., θ describes 360 degrees). The resultant signal received at point P is always proportional to the length of the line connecting the end of E_2 with the end of E_3. A polar plot of this resultant against the angle θ yields the familiar directivity diagram for a given set of conditions of current ratios, phasings, and spacings.

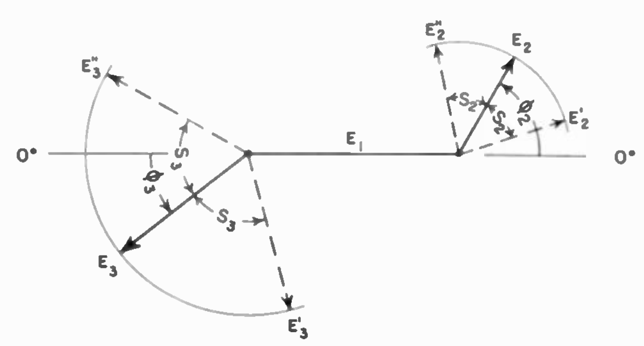

Fig. 3—Movements of vectors E_2 and E_3 as the horizontal angle is varied. This figure suggests the possibility of a calculator based upon the combination of phase-modulated waves.

An algebraic solution for the resultant signal received at point P may be obtained by expressing the three vectors of (1) in the rectangular form preparing them for addition.

$$
\begin{aligned}
&E_1 = e_1 + j0 \\
&E_2 = e_2\cos[(S_2\cos\theta)+\phi_2] + je_2\sin[(S_2\cos\theta)+\phi_2] \qquad (2)\\
&E_3 = e_3\cos\{[S_3\cos(\theta-\beta)]+\phi_3\} \\
&\qquad + je_3\sin\{[S_3\cos(\theta-\beta)]+\phi_3\}.
\end{aligned}
$$

The resultant voltage E_t becomes

$$E_t = E_1 + E_2 + E_3 \qquad (3)$$

$$
\begin{aligned}
A &= E_t\,(\text{real part}) = e_1 + e_2\cos[(S_2\cos\theta)+\phi_2] \\
&\quad + e_3\cos\{[S_3\cos(\theta-\beta)]+\phi_3\} \qquad (3a)\\
B &= E_t\,(\text{imaginary part}) = je_2\sin[(S_2\cos\theta)+\phi_2] \\
&\quad + je_3\sin\{[S_3\cos(\theta-\beta)]+\phi_3\}. \qquad (3b)
\end{aligned}
$$

The resultant signal at point P due to contributions of 3 elements is

$$e_t = \sqrt{A^2+B^2}. \qquad (4)$$

The instrument to be described gives a true solution of (4), which is the conventional form for computation. The principle of operation, however, is more clearly seen in (1).

III. The Instrument

Referring to (1), it is apparent that an instrument which includes two rotatable vector arms of adjustable length and one fixed arm of constant length, after the fashion of Fig. 3, can solve (4). A mechanism such as shown in Fig. 4 can evaluate the product $(S\cos\theta)$. The sliding plate moves freely in slotted guides under the influence of the disk rotating with the horizontal angle θ. The adjustable cam transfers the longitudinal component of motion to the sliding plate and thence to

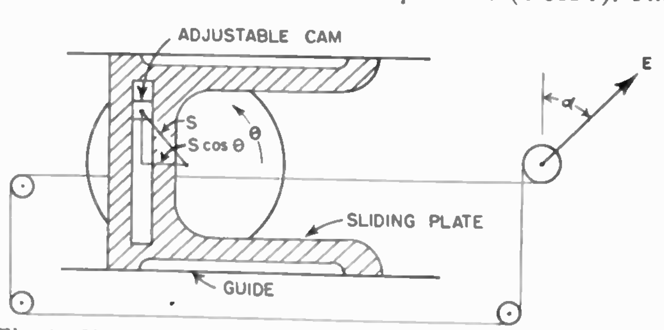

Fig. 4—Sketch showing the basic operating principle of the array calculator. With a fixed vector this constitutes a simple 2-element calculator.

the string-pulley system which rotates the vector E. The distance from the center of the θ disk to the center of the cam is proportional to S, the spacing between the tower in question and the reference tower. The longitudinal component of motion is then ($S \cos \theta$). The phasing angle ϕ is constant and may be introduced simply by loosening the string and rotating E plus or minus ϕ degrees, as the case may be, without moving any of the rest of the mechanism. For the extreme case of $S=0$, the tower in question would coincide with the reference tower and the center of the cam would be directly over the center of the θ disk causing the vector E to remain motionless. This would give a resultant of constant length and hence a circular pattern. In reality then, we see that the structure of Fig. 4 is basically a 2-element calculator if a constant reference vector and some means of measuring the resultant are provided.

Fig. 5 shows a photograph of the tracer designed for any number of array elements up to and including 3. Two complete mechanisms such as shown in Fig. 4 are included, one above the other. Disks 2 and 3 are clamped so that they rotate together, although an angle β may be inserted between them. The top disk

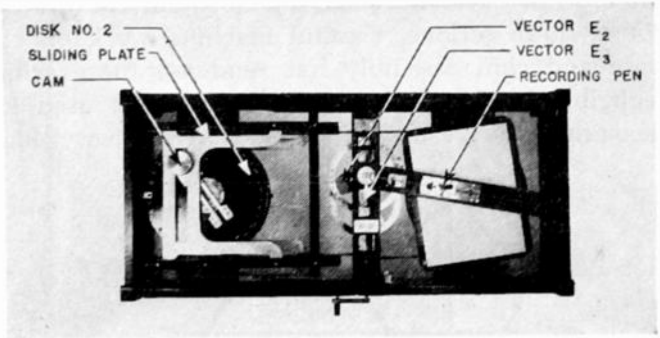

Fig. 5—Top view of pattern tracer.

drives a sliding plate which causes vector E_2 to move through the angle α_2. The lower disk drives vector E_3 through angle α_3. The reference vector E_1 is fixed both in position and length and it is incorporated in this instrument as the offset distance between the centers of vectors E_2 and E_3. This has been arbitrarily selected as 2 inches in this instrument. The resultant of the three vectors is obtained by means of a thread running through highly polished 0.04-inch holes connecting, in effect, the ends of vectors E_2 and E_3. The length of this resultant is then used to drive the recording pen against a restoring spring. A 2-inch change in length of the resultant causes a 1-inch movement of the recording pen due to the 2:1 reduction caused by the recorder pulley.

Fig. 6 shows in detail the lower driving disk (disk 3) and the fact that it is identical to disk 2 and rotates with it although the constant angle β may be introduced. The lengths of vectors E_2 and E_3 are determined solely by the current ratios. The manner of adjusting the length of vectors and the manner in which the thread measures the resultant are shown in Figs. 7 and

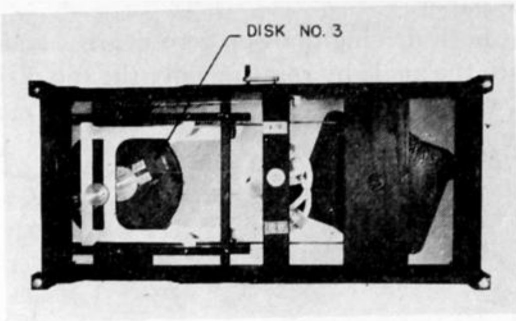

Fig. 6—Bottom view of pattern tracer.

8. The angles representing phasing and spacing are read on the protractors shown in Fig. 7, although the

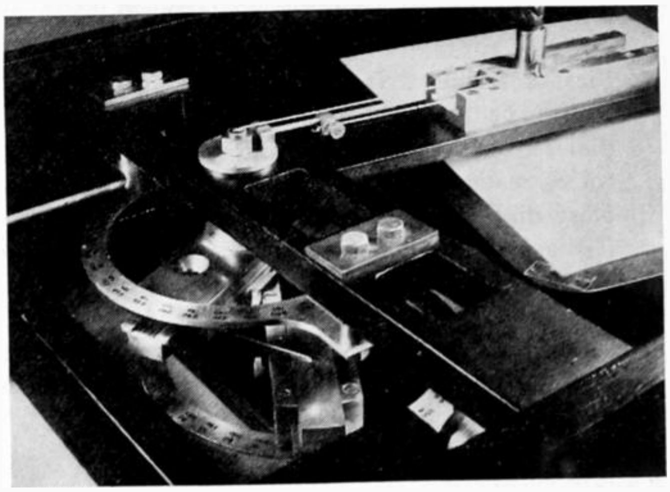

Fig. 7—Vector arms E_2 and E_3 with their associated protractors.

spacing adjustment is actually made by regulating the distance between the cam and the disk center.

Fig. 9 gives the necessary information for the tracing of a horizontal directivity pattern. Using it for an example, the steps in operating the calculator will be enumerated briefly:

1. Set current ratios by adjusting length of vector arms E_2 and E_3.

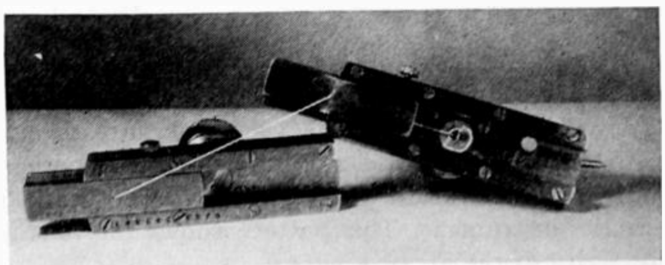

Fig. 8—Vector arms removed from their frame to show the method of adjusting their length (current ratio) and of measuring the resultant with a thread. The string shown is much larger than the thread actually used in order to render it visible in the photograph.

2. Set vectors E_2 and E_3 on zero degrees which places all vectors in line with E_1 and causes the resultant

to have maximum length. Set the recording pen to be $1+0.7+0.5=2.2$ units from the center of the paper.

3. Set both driving dials on zero degrees and introduce the angle by rotating only the top disk (disk 2), then lock disks 2 and 3 together by means of the lock ring.

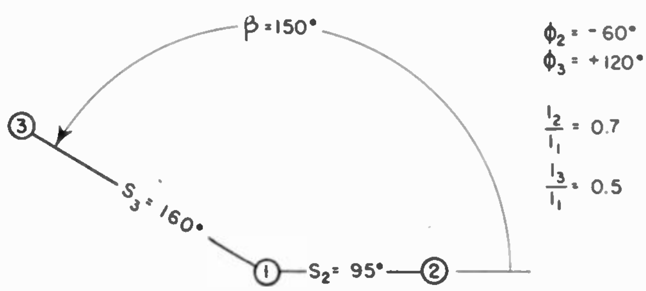

Fig. 9—The specifications of a 3-element array whose horizontal pattern is shown in Fig. 10.

4. First set dial 2 (top dial) on 90 degrees and restore vector E_2 to zero manually. Next rotate dial 2 to zero degrees, and adjust cam until vector E_2 reads +95 degrees which is the spacing S_2.
5. Rest dial 2 to 90 degrees which brings vector E_2 back to zero. Set vector E_2 on −60 degrees manually which introduces the phasing ϕ_2.
6. Repeat operations 4 and 5 for the lower dial (dial 3) and vector E_3. This introduces S_3 and ϕ_3.
7. One rotation of the dials traces a pattern such as shown in Fig. 10.

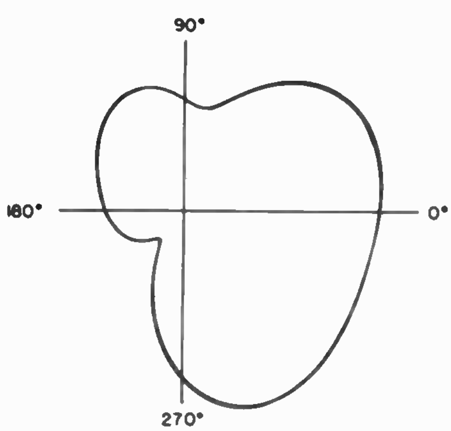

Fig. 10—Horizontal polar pattern of the array of Fig. 9 traced on the instrument described.

Actually introducing the seven basic adjustments on the machine is little more time-consuming than reading the above list of operations to one familiar with the instrument. This pattern shows the *relative* directivity of the array.

To check the accuracy with which the machine traces the patterns, calculations were made for several directional antenna systems. Fig. 11 shows a comparison between the calculated points and the pattern traced directly by the machine for a 2-element array spaced 135 degrees, phased 50 degrees, and having a current ratio of 0.7. The agreement is seen to be very satisfactory. The errors are largely a function of the type of pattern being traced. The errors in the 3-element pattern of Fig. 12 are seen to be somewhat greater, due to the greater stresses set up for this particularly combination of vectors.

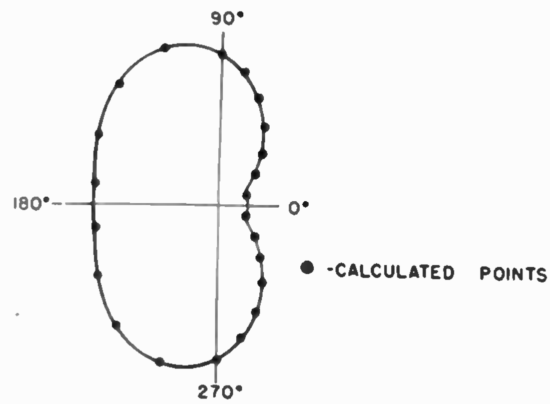

Fig. 11—Pattern traced for a 2-element array having the following characteristics: phasing 50 degrees, spacing 135 degrees, current ratio 0.7. Calculated points are shown for comparison.

The major sources of error can be traced to elasticity in the driving cables and the resultant-measuring thread. Freedom of movement of the sliding plate in its guides will contribute little to the over-all error, but any looseness between the cam and the driving slot is quite serious. Careful machine work on the plate and cam assembly has rendered these errors negligible in this instrument. The thread used for measuring the resultant should have a very high

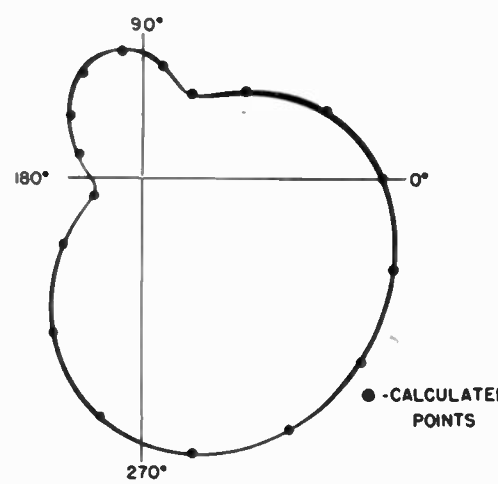

Fig. 12—Pattern traced for a 3-element array having the following characteristics: $S_2=100$ degrees, $S_3=100$ degrees, $\phi_2=-50$ degrees, $\phi_3=100$ degrees, $\beta=100$ degrees, $I_2=I_3=I_1=1.0$. Calculated points are shown for comparison.

modulus of elasticity to minimize error, but on the other hand it must be sufficiently flexible to turn on its own radius. The recording pen must be free to follow small vertical irregularities of the rotating table and yet it must not tip.

Fig. 13, which was traced exactly as illustrated in just a few minutes, shows how the variation of a single parameter will affect the operation of a directional antenna. The array of Fig. 11 is the case for 135-degree spacing. Fig. 14 shows the effect of

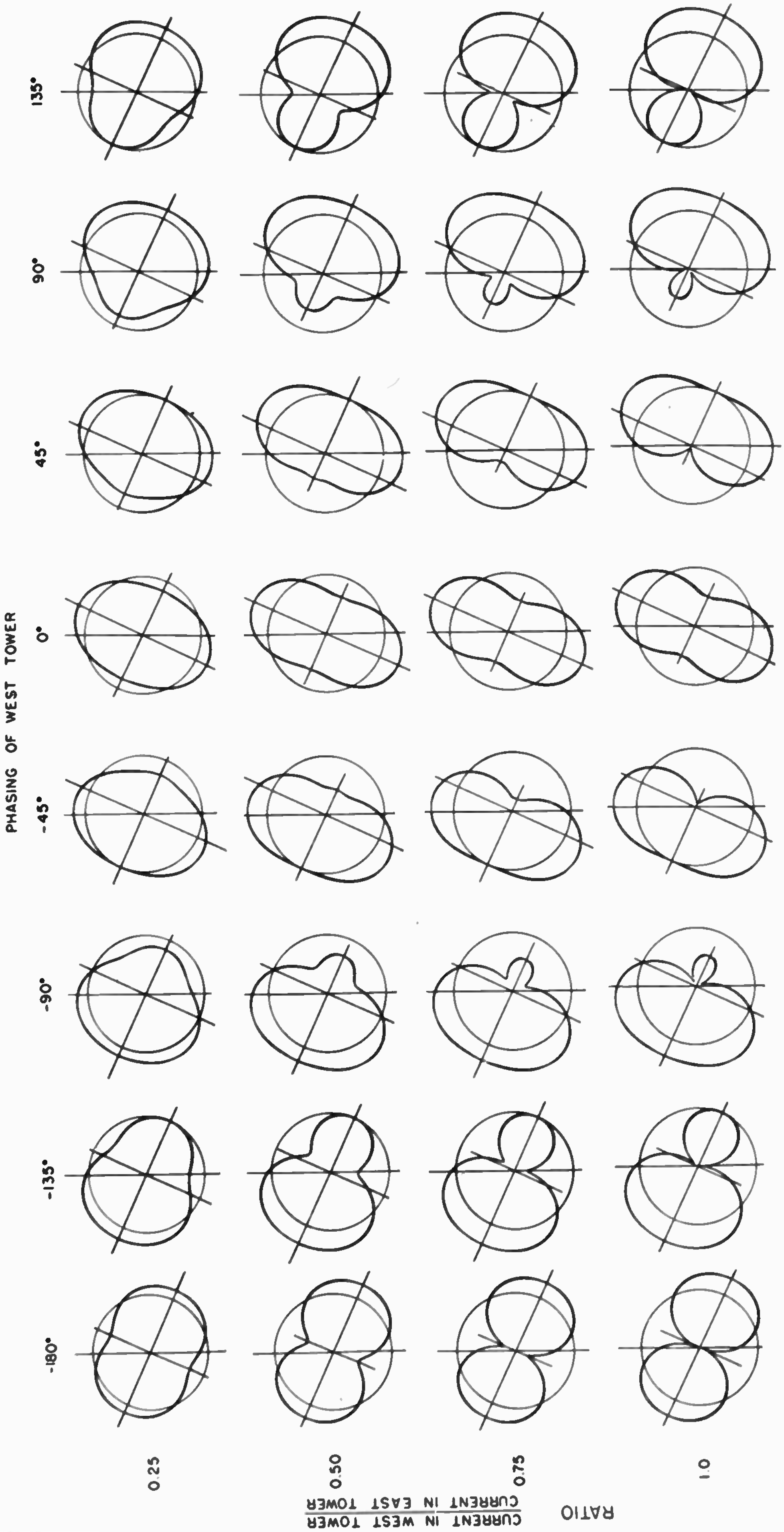

Fig. 15—The horizontal radiation patterns available from the 2-element directional antenna of radio station KOAC, Corvallis, Oregon. North is toward the top of the figure in the usual manner.

variation of the phasing of element 2 from +20 to −70 degrees for the array of Fig. 9. The use of this tracer in exploring the variety of horizontal patterns available from an array of given physical dimensions is illustrated in Fig. 15. Investigation of the operation of an array at frequencies other than that for which it was designed is another obvious use of this instrument.

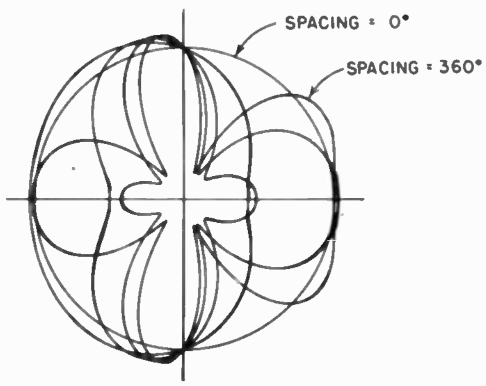

Fig. 13—Composite figure showing the 2-element array of Fig. 11 with spacing varied from 0 to 360 degrees in 90-degree steps.

The calculator as described is based on the premise of 2 or 3 elements of identical heights. Realizing the fact that any change in height changing the horizontal signal can be compensated by readjusting the current ratios, patterns for arrays having towers of unequal height can be drawn. For instance, if tower 1 has a horizontal figure of merit of 200 millivolts per meter at 1 mile for a given power input and the towers 2 and 3 each have a horizontal figure of merit of only 100, this decreased effectiveness of towers 2 and 3 can be simulated by putting only half the normal current in them. Changes in vertical directivity, which are not considered in this paper, will occur, but those experienced with antennas can estimate the degree of change quite closely.

While basically, this instrument is designed for tracing 3-element patterns, it can be used with full effectiveness for 2-element antennas by the expedient of reducing the length of vector E_3 to zero, i.e., reduce its current to zero. Only three variables exist for this case, therefore the operation is much simpler.

IV. Conclusions

While precision results are not claimed for the machine which has been constructed and described, there is apparently no reason why a precision machine could not be built on the general principles outlined.

The advantage of this calculator appears to be twofold: (1) exploration for a suitable pattern to meet given interference and coverage problems, and (2) investigating a given array to determine the effect of varying any parameter.

V. Acknowledgments

The aid of the Engineering Experiment Station of Oregon State College through S. H. Graf, director of engineering research, has made possible the construction of this calculator. W. H. Huggins suggested the method of measuring the resultant of the vectors.

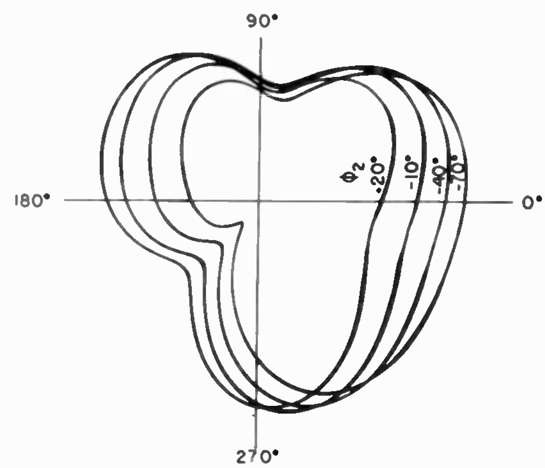

Fig. 14—Showing the effect of varying the phasing of element 2 of the 2-element array of Fig. 9 from +20 to −70 degrees.

A. C. Harwood, C. C. Dusek and C. S. Freed are responsible for the constructional work. The encouragement and interest of Professor F. O. McMillan, head of the department of electrical engineering, have been invaluable.

A Mechanical Calculator for Directional Antenna Patterns*

WILLIAM G. HUTTON† ASSOCIATE, I.R.E., AND R. MORRIS PIERCE†, ASSOCIATE, I.R.E.

Summary—In the design of a directional array the engineer is confronted with the solution of a problem that is laborious because of the many parameters to be considered. The variables that enter the problem are spacing, phasing, radiated field, and orientation for each antenna involved. In particular, for an array of 3 elements there are 9 parameters when one of the elements is considered to be at the origin. This paper describes a mechanical calculator that eliminates a major part of the tedious work involved in the calculation of the horizontal pattern of a 2- or 3-element array. The calculator is of equal use in the determination of the vertical patterns when the antenna heights are equal. An analysis is made to show that the results given by the calculator are solutions to the standard equation. The accuracy of the results are also considered.

Introduction

The computation of the field pattern for any given combination of antennas in an array is a straightforward but lengthy process. Because of the number of independent parameters that must be selected by the designer, the optimum condition is difficult to obtain. For example, in a 3-element array there are 9 variables to consider in order to fix the shape of the associated pattern.

The design procedure for a directional antenna array usually involves the calculation of several trial patterns before the desired pattern is obtained. A fairly complete representative list of patterns for 2-element arrays has been published,[1,2] and those patterns serve the purpose of pointing out a good starting place for the solution of a design problem where 2 antennas only are required. When 3 antennas are required to give the desired protection a much greater amount of work is necessary to calculate a pattern and more trial patterns are normally made in the hunt for the desired pattern.

Three-element symmetrical antenna arrays offer a much greater choice of arrangements than 2 antennas but as yet published patterns for them have not been systematized. The work associated with 3-element nonsymmetrical arrays is even more laborious than for 3 antennas in a straight line and in addition there is a much wider scope of possible configurations. A little investigation will show that many thousands of patterns would be necessary to give a good representation of the possible 3-element patterns, having complete nulls, for spacings up to 1 wavelength, in steps of 45-degree spacing, 45-degree phasing, and 45-degree orientation with current ratios selected to give the complete nulls.

It becomes evident after computing a few 3-element nonsymmetrical patterns that some aid to facilitate and to speed up the computations is almost vital to good design progress. Such an aid would not only be helpful in finding a pattern near the desired shape but would be most appreciated in exploring that near neighborhood for the pattern that would exactly meet the protection requirements.

This paper describes a 3-element, mechanical, directional-antenna-pattern calculator to be used when all the factors are known that completely determine the pattern. The device may be defined as a machine that can be adjusted for any or all the factors that determine the pattern and which, on operation, will automatically rotate the vectors, which represent the relative antenna fields, in such a way with respect to each other that the resultant vector field may be read directly for any direction from the antenna array. The mathematics used in the following analysis conforms with the Federal Communication Commission's "Standards of Good Engineering Practice" in regard to positive direction of measurement for angles θ and ϕ. That is to say, the horizontal direction variable ϕ is measured clockwise for positive direction, and the elevation angle θ is measured from horizontal (0) to vertical (90 degrees).

Calculation of Horizontal Patterns

A brief review as to the theory of directional arrays follows. Fig. 1 is a plan view of a 3-element antenna

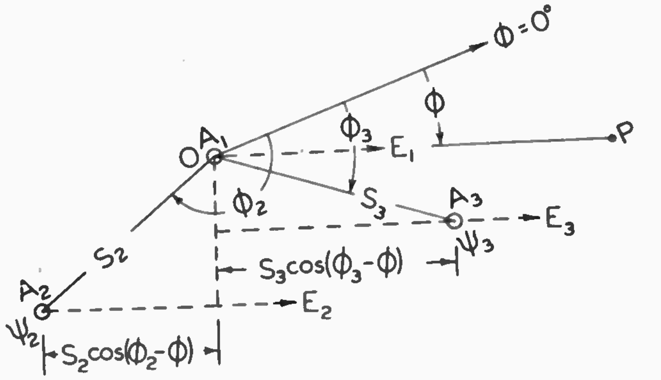

Fig. 1—Plan view of 3-element array.

array and the following discussion applies to the horizontal pattern. The point O serves as reference point or the origin and $\phi=0$ degrees is considered reference direction. We assume the observer to be at point P in the direction ϕ degrees with respect to reference direction, and to be sufficiently far removed from the array that, for all practical purposes, lines drawn from the different antennas to the point P are parallel. Antenna A_1 is located at the origin O. Antenna A_2 is located at

* Decimal classification: R325. Original manuscript received by the Institute, September 15, 1941. Presented, Summer Convention, Detroit, Michigan, June 24, 1941.

† Station WGAR, Hotel Statler, Cleveland, Ohio.

[1] G. C. Southworth, "Certain factors affecting the gain of directive antennas," Proc. I.R.E., vol. 18, pp. 1502–1536; September, 1930.

[2] A. James Ebel, *Electronics*, vol. 9, pp. 29–30; April, 1936.

a distance of S_2 degrees from A_1 and at an angle of ϕ_2 degrees from the reference direction, while antenna A_3 is located a distance S_3 degrees from A_1 in the direction ϕ_3 degrees. The current in antenna A_2 is assumed to lead the current in A_1 by ψ_2 degrees and the current in A_3 leads the current in A_1 by ψ_3 degrees.

It will be seen from Fig. 1 that to the observer at the point P, antenna A_3 appears to be closer than antenna A_1 by $S_3\cos(\phi_3-\phi)$ degrees, thus the signal from antenna A_3 will arrive at P ahead of that from A_1 by that many degrees due to spacing alone. The signal from A_3 leads that from A_1 by ψ_3 degrees due to time phasing in addition to its lead due to spacing so that the net phase relation on arrival at P is a lead of $S_3\cos(\phi_3-\phi)+\psi_3$ degrees. By a similar argument the signal from antenna A_2 can be shown to lead that from A_1 by $S_2\cos(\phi_2-\phi)+\psi_2$ degrees. For simplicity of expression these angles of lead will be referred to as β_3 and β_2, respectively. Thus,

$$\begin{aligned}\beta_2 &= S_2 \cos(\phi_2 - \phi) + \psi_2 \\ \beta_3 &= S_3 \cos(\phi_3 - \phi) + \psi_3.\end{aligned} \tag{1}$$

In Fig. 1 the root-mean-square value of fields E_1, E_2, and E_3 radiated from antennas A_1, A_2, and A_3, respectively, may be represented by signal vectors V_1, V_2, and V_3.

$$E_1 = KV_1, \quad E_2 = KV_2, \quad E_3 = KV_3. \tag{2}$$

K is a constant depending on the root-mean-square value of the total field radiated by the directional antenna system and the scale used for the vectors such that

$$E_{rms} = KV_{rms}. \tag{3}$$

V_{rms} is the root-mean-square value of the resultant vector V for the complete directional pattern.

Fig. 2 is a vector diagram showing the relationship of the signal vectors V_1, V_2, and V_3 as well as the resultant vector V. The angle β may be thought of as the phase angle of the resultant vector field at the point P.

Fig. 3 is a vector diagram obtained from Fig. 2 by resolving the three vectors V_1, V_2, and V_3 to component form.

From Fig. 3, (4) can be written directly for magnitude of V.

the addition of the vectors in actual operation.

With reference to Fig. 2, for fixed magnitudes of V_1, V_2, and V_3, the value of V depends only upon the values of β_2 and β_3. In other words it might be said that the description of the operation of the machine

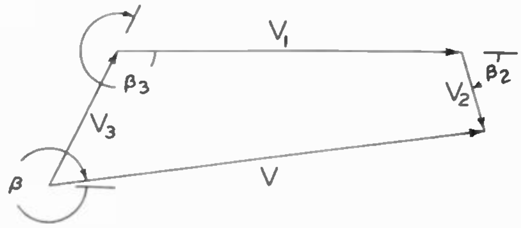

Fig. 2—Signal vector diagram.

reduces to a description of the manner in which the angles β_2 and β_3 are caused to change as the machine is rotated to indicate the various directions in which the resultant vector V is being determined. It can be seen that in the expressions for β_2 and β_3 the only variable

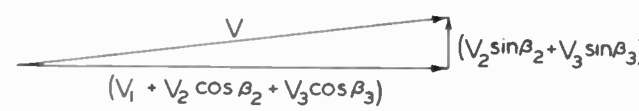

Fig. 3—Component vector diagram.

is ϕ for the particular pattern being determined and ϕ may be spoken of as the direction variable.

Fig. 4 is a schematic diagram for a 2-element machine and shows the fundamental operating principles employed in the 3-element machine as well as the 2. To the right we have the vector assembly, to the left we have the assembly representing the antennas, and at the top we have the translatory mechanism which serves the purpose of transferring the antenna-spacing relationship at the left to the vector angular relationship at the right.

It might be said that there are two methods by which the directional pattern from an antenna array may be determined; one would be to have the observer move around the array and measure the field as he moves, the other would be to have the observer remain at one position and measure the field while the antenna array is being rotated bodily. The latter method could be carried out by use of models and is the one made use of in this machine. Antenna A_1 is located at the

$$V = \sqrt{(V_1 + V_2\cos\beta_2 + V_3\cos\beta_3)^2 + (V_2\sin\beta_2 + V_3\sin\beta_3)^2}. \tag{4}$$

If we introduce the constant K, (4) becomes (5) which gives the field intensity.

center of rotation or origin, while the observer is considered to be in the direction marked 0 degrees, at a

$$E = K\sqrt{(V_1 + V_2\cos\beta_2 + V_3\cos\beta_3)^2 + (V_2\sin\beta_2 + V_3\sin\beta_3)^2}. \tag{5}$$

This equation can be recognized as the generally accepted standard rectangular form for the horizontal pattern.

The vector diagram in Fig. 2 is, in reality, a picture showing the manner in which the machine provides for

distant point P. Antenna A_2 is spaced S_2 degrees from A_1 but can be rotated about A_1.

From Fig. 4 it can be seen that as A_2 is rotated about A_1 the translatory mechanism will move to the left and right in accordance with a cosine function, and that

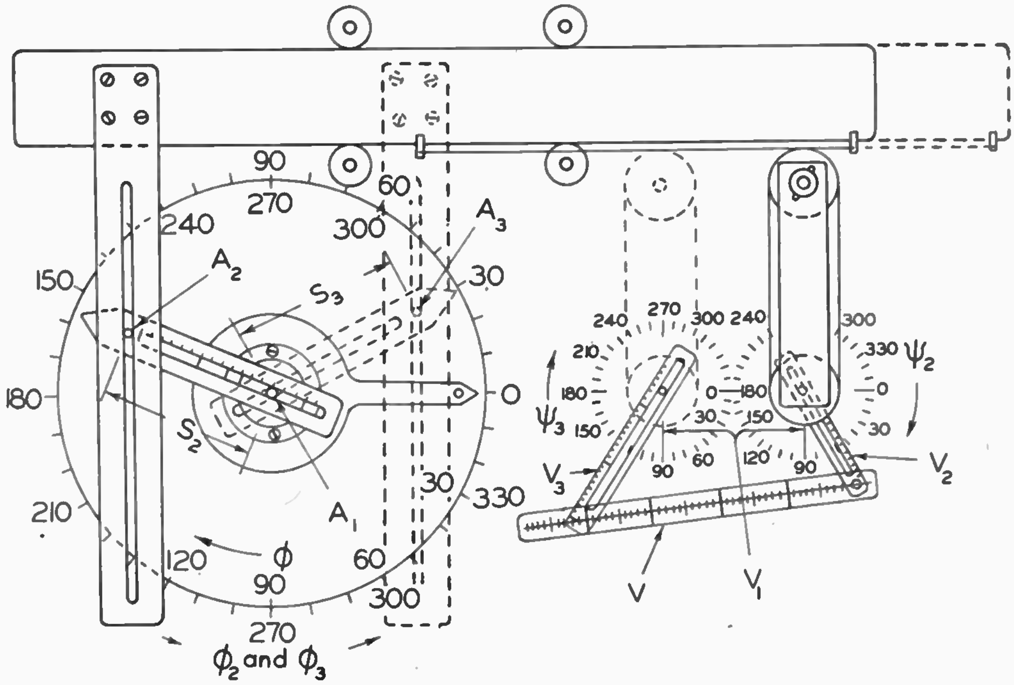

Fig. 5—Schematic diagram for 3-element machine.

the left and right motion of the translatory mechanism will be transferred to the rotation of vector V_2 by means of a cable and pulleys. Vector V_2 is so related to the fixed vector V_1 that the resultant vector V may be read at any position of V_2 on a scale provided for that purpose. Let it suffice to say that scales are so arranged for the various adjustments that the pointer I will indicate the direction in which the vector field V is being measured or determined.

Fig. 5 is a schematic diagram of the 3-element pattern calculator that has been in operation for some time. The third antenna A_3 and associated translatory equipment is mounted on the back side of the base plate as indicated by the dotted lines. The base plate serves the purpose of a frame for the machine. The vector V_3 is driven by a shaft that extends through the base plate from the rear side to the front. The resultant vector V is measured from the movable end of V_3 to the movable end of V_2 and a scale is provided to make that reading possible for any position of either V_3 or V_2.

There are nine possible adjustments on the machine to vary the parameters that describe a 3-element antenna array. The parameters are: the signal vector magnitudes V_1, V_2, and V_3 and the associated phase angles ψ_2 and ψ_3, the antenna spacings S_2 and S_3 together with the antenna orientation angles ϕ_2 and ϕ_3. Since antenna A_1 is located at the origin the vector V_1 is assumed to have an angle of 0 degrees and its direction is called reference direction for measuring phase angles ψ_2 and ψ_3.

Three linear scales, in arbitrary units, are provided for adjusting the 3 signal vector magnitudes to represent various fields radiated. A protractor scale is provided for adjusting each of the phase angles ψ_2 and ψ_3.

A linear scale, calibrated in degrees, is provided for adjusting the spacing between A_1 and A_2 and another for the spacing between A_1 and A_3. One protractor, scaled counterclockwise serves to adjust the orientation of either ϕ_2 or ϕ_3.

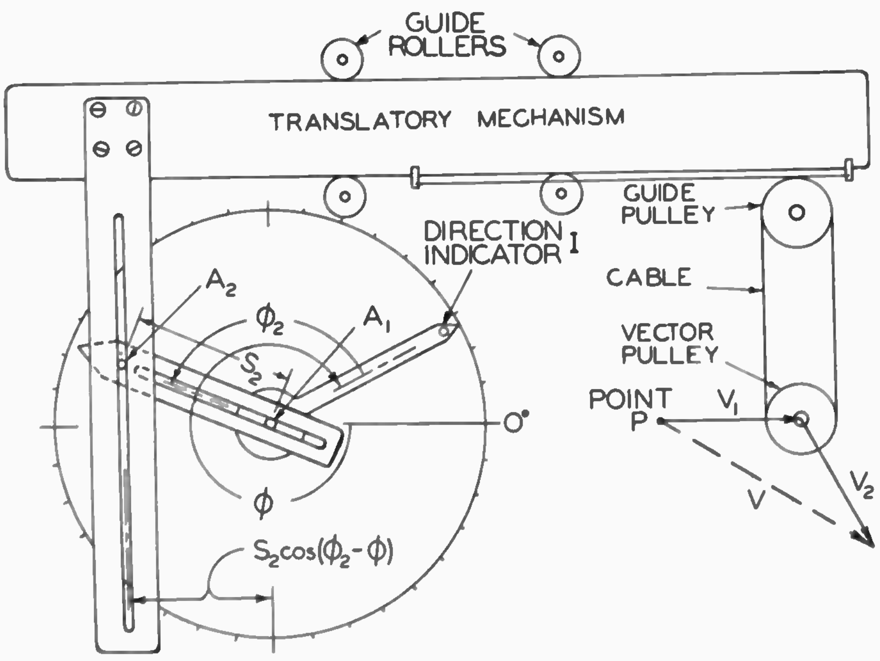

Fig. 4—Schematic diagram for 2-element machine.

A protractor, scaled clockwise, is provided for setting the indicator I in any desired direction. The machine operates in such a way that the value of the resultant vector V gives the relative magnitude of the field for the direction indicated by the indicator I.

Calculation of Vertical Patterns

The machine is very helpful in computing the vertical patterns for an array when the antennas are equal in height. For 3 antennas of equal height (6) holds true.

$$E = K \frac{\cos(h \sin \theta) - \cos h}{(1 - \cos h) \cos \theta} V \qquad (6)$$

where E = field intensity
h = height of antennas in degrees
θ = elevation angle

$$K = E_{rms}/V_{rms}$$

$$\frac{\cos(h \sin\theta) - \cos h}{(1 - \cos h)\cos\theta} = \text{vertical form factor}^{3}$$

into the same straight line with indicator I and I is rotated to 0 degrees, (3) the antenna spacings are then adjusted so that β_2 and β_3 are the same as noted in step (1). The machine will then be adjusted for the

$$V = \sqrt{\begin{array}{l}\{V_1 + V_2 \cos[S_2 \cos(\phi_2 - \phi)\cos\theta + \psi_2] + V_3 \cos[S_3 \cos(\phi_3 - \phi)\cos\theta + \psi_3]\}^2 \\ + \{V_2 \sin[S_2 \cos(\phi_2 - \phi)\cos\theta + \psi_2] + V_3 \sin[S_3 \cos(\phi_3 - \phi)\cos\theta + \psi_3]\}^2.\end{array}}$$

The machine gives the value of the radical V for different elevation angles θ in any one horizontal direction ϕ after the horizontal pattern has been computed. Readjustments are necessary for the determination of each vertical pattern but concern only the antenna orientations and spacings. In general the step procedure is as follows: (1) the vector angles β_2 and β_3 for the direction ϕ are noted when the machine is adjusted to run off the horizontal pattern (β_2 and β_3 are obtained from the scales which indicate ψ_2 and ψ_3, respectively, see Fig. 5), (2) all antennas are then thrown

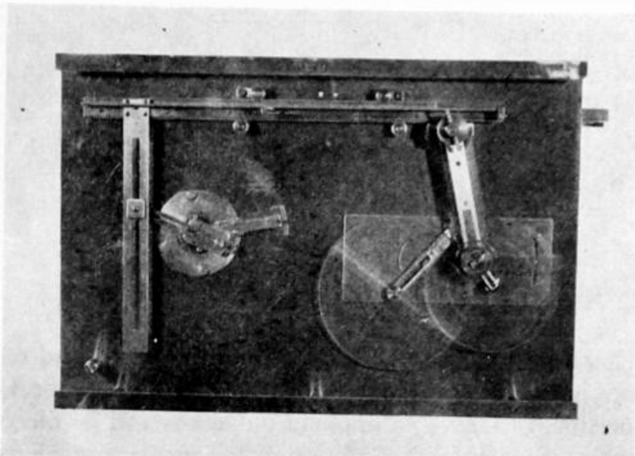

Fig. 6—Front view of calculator.

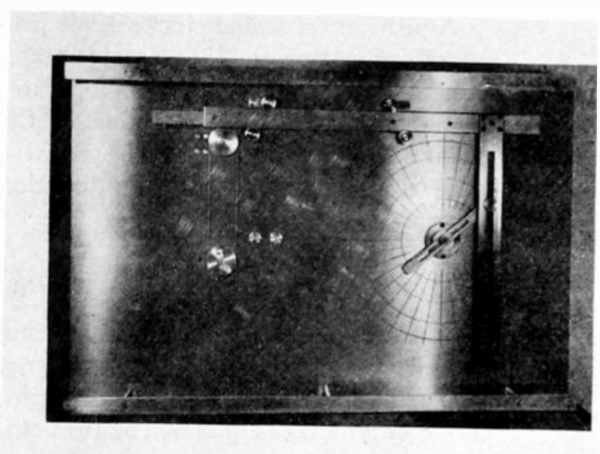

Fig. 7—Rear view of calculator.

[3] Derivation given by G. H. Brown, "Directional antennas," Proc. I.R.E., vol. 25, pp. 78–145; January, 1937. The form factor mentioned here is shown in the brackets of equation (182) on page 142 except for the denominator which is rewritten so that for $\theta = 0°$ the form factor is unity, G is equal to h, and θ is zenith angle rather than elevation angle.

determination of the vertical pattern in the direction ϕ and rotation of the indicator from 0 to 90 degrees will give the necessary values of the radical V. Substitution of these values of V in (6) will give the vertical pattern for the direction ϕ.

The Working Model

Fig. 6 is a photograph of the front or working side of the machine described in this paper; Fig. 7 is a photograph showing the rear (or under) side of the machine; and Fig. 8 is a photograph showing the calculator as set up for operation.

The machine can be comfortably operated in a sitting position. Normally it is necessary to lift the machine for one adjustment on the back (or under) side and that is for spacing S_3. The orientation adjustment ϕ_3 is made on the back side but A_3 is normally set to coincide with the direction indicator I for convenience of operation, since it is not visible from the front.

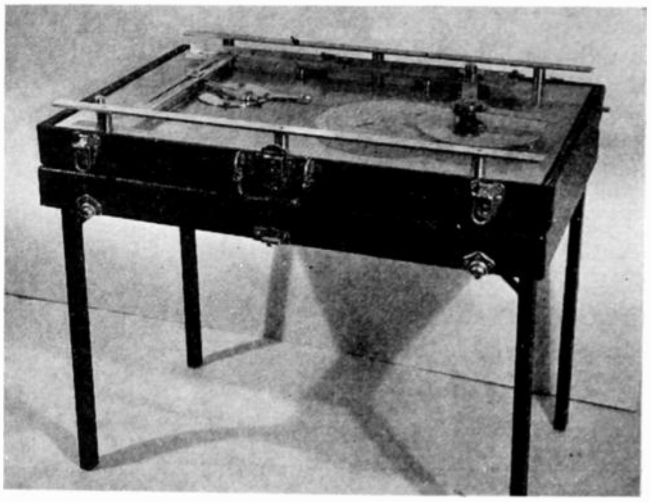

Fig. 8—Photograph of calculator set up for operation.

Conclusions

The mechanical calculator for directional antennas is a substantial aid in speeding up the work of drawing horizontal patterns and eliminates the laborious and tedious calculations by formula. It eliminates the bulk of the work in connection with drawing the vertical patterns when the antennas are of equal height.

The machine gives a concrete picture of the vector diagram made up of the vector fields from the antennas for any direction from the array.

The accuracy is sufficiently good to make the machine entirely practical since experience has shown it to be better than can be obtained by the use of a slide rule. The results obtained by the use of this machine are well within the accuracy of commercially available field-strength-measuring equipment.

The machine is portable and complete in one unit that fits into a carrying case. The carrying case unfolds to form a worktable on which the machine can be operated. As a result the device is useful either in the office for design work or on location as a valuable aid where an array is being tuned and finally adjusted for operation in service.

Acknowledgment

We wish to credit Mr. Windsor Atwater, construction and maintenance engineer at WGAR, with the complete construction of the machine described here. His excellent workmanship, ingenuity, and care in layout made the first model entirely practical.

Charts for the Determination of the Root-Mean-Square Value of the Horizontal Radiation Pattern of Two-Element Broadcast Antenna Arrays*

KARL SPANGENBERG†, ASSOCIATE, I.R.E.

Summary—*A set of two charts is presented from which the root-mean-square value of the horizontal radiation patterns of two-element antenna arrays may be determined quickly without calculation. These charts are based upon the formula* $E_{rms}=E_1\sqrt{1+M^2+2M\,J_0(2\pi S/360)\cdot\cos\psi}$. *Observations on the fact that the root-mean-square value is independent of phasing for 138 degrees spacing are made.*

In the design of antenna arrays it is desirable to know the root-mean-square value of the horizontal radiation pattern. This value gives an indication of the average power radiated. When compared with values of field strength in the directions of maximum radiation it indicates the relative power radiated in those directions. A determination of the root-mean-square value of the horizontal pattern is required in applications for broadcast-station licenses submitted to the Federal Communications Commission.

The horizontal patterns of two-element arrays for various spacings and for various phase differences in antenna currents are available.[1-4] Some of the published patterns show the root-mean-square value of the pattern. In no case are the data on this item complete, nor are they given for ratios of antenna currents other than one. It is the purpose of this paper to present a set of two charts from which the root-mean-square value of the horizontal radiation pattern may be determined quickly without calculation.

The root-mean-square value of the horizontal radiation pattern of a two-element antenna array consisting of vertical grounded antennas of the same height is given by

$$E_{rms} = E_1\sqrt{1 + M^2 + 2M\,J_0(2\pi S/360)\cos\psi} \qquad (1)$$

in which

E_{rms} = the root-mean-square value of the field strength of the array in the horizontal plane in millivolts per meter at 1 mile

E_1 = the effective value of the field strength of the antenna carrying the larger current when radiating alone in millivolts per meter at 1 mile

$E_2=ME_1$ = the effective value of the field strength of the antenna carrying the smaller current when radiating alone in millivolts per meter at 1 mile

M = ratio of the current in the two antennas, taken as the ratio of the smaller current to the larger current, thus making M less than unity

ψ = the phase angle by which the current in the second antenna (smaller current) lags the current in the first antenna (larger current)

S = the antenna spacing in electrical degrees. Equal to $360d/\lambda$ where d is the distance between antennas and λ is the wavelength

J_0 = the zero-order Bessel function of the antenna spacing in radians.

The derivation of the above expression is given in Appendix I.

It will be observed that the root-mean-square value of an antenna radiation pattern depends upon three

* Decimal classification: R125. Original manuscript received by the Institute, December 18, 1941.

† Stanford University, California.

[1] R. M. Foster, "Directive diagrams of antenna arrays," *Bell Sys. Tech. Jour.*, vol. 5, pp. 292–307; April, 1926.

[2] G. C. Southworth, "Factors affecting the gain of directive antenna arrays," *Bell Sys. Tech. Jour.*, vol. 10, pp. 63–95; January, 1931.

[3] G. H. Brown, "Directional antennas," Proc. I.R.E., vol. 25, pp. 78–145; January, 1935. (Root-mean-square patterns shown.)

[4] National Association of Broadcasters, "Engineering Handbook," Denver, Colorado, 1938. Section F (root-mean-square value of patterns shown.)

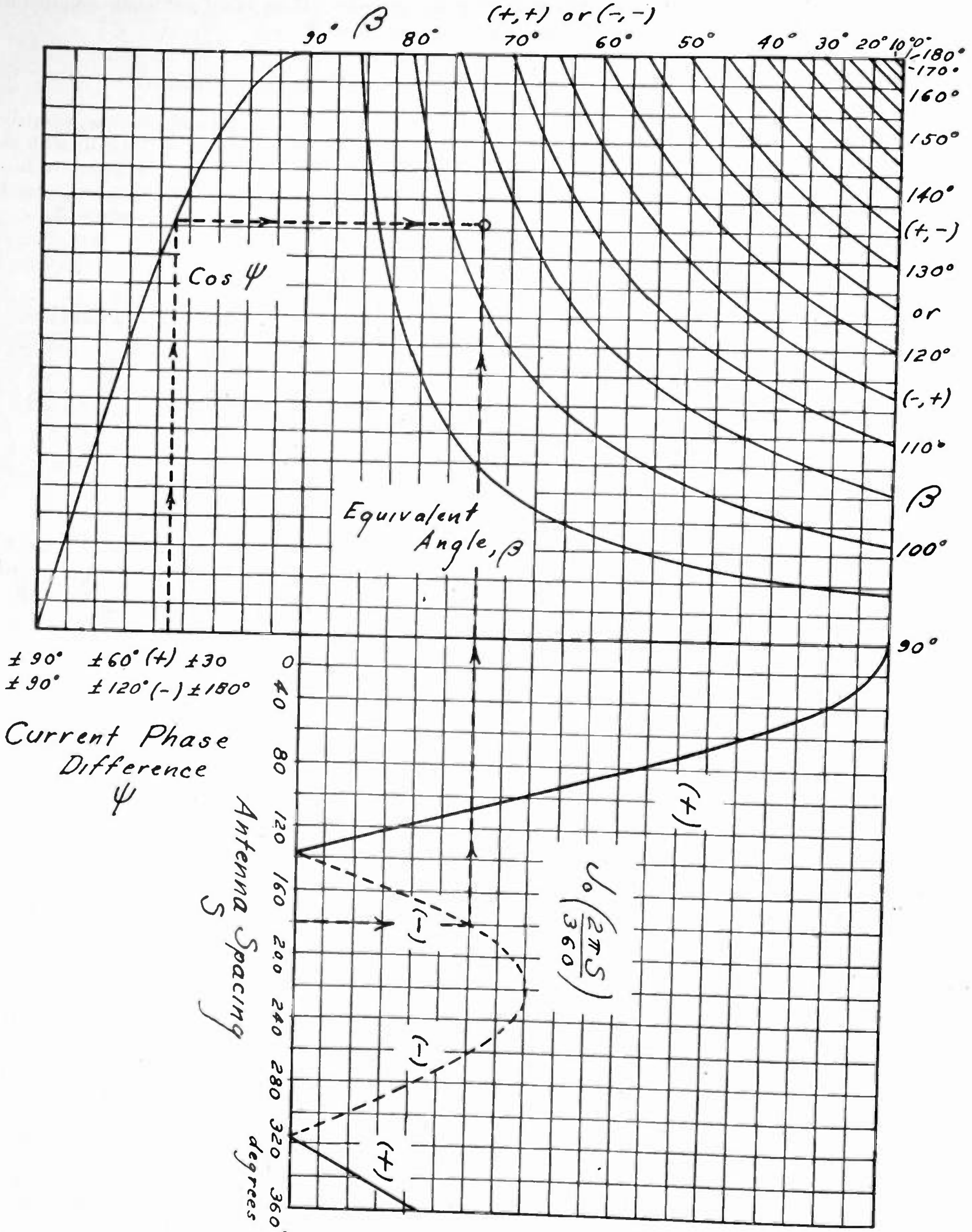

Fig. 1—Determination of equivalent angle from value of antenna spacing and current phasing.

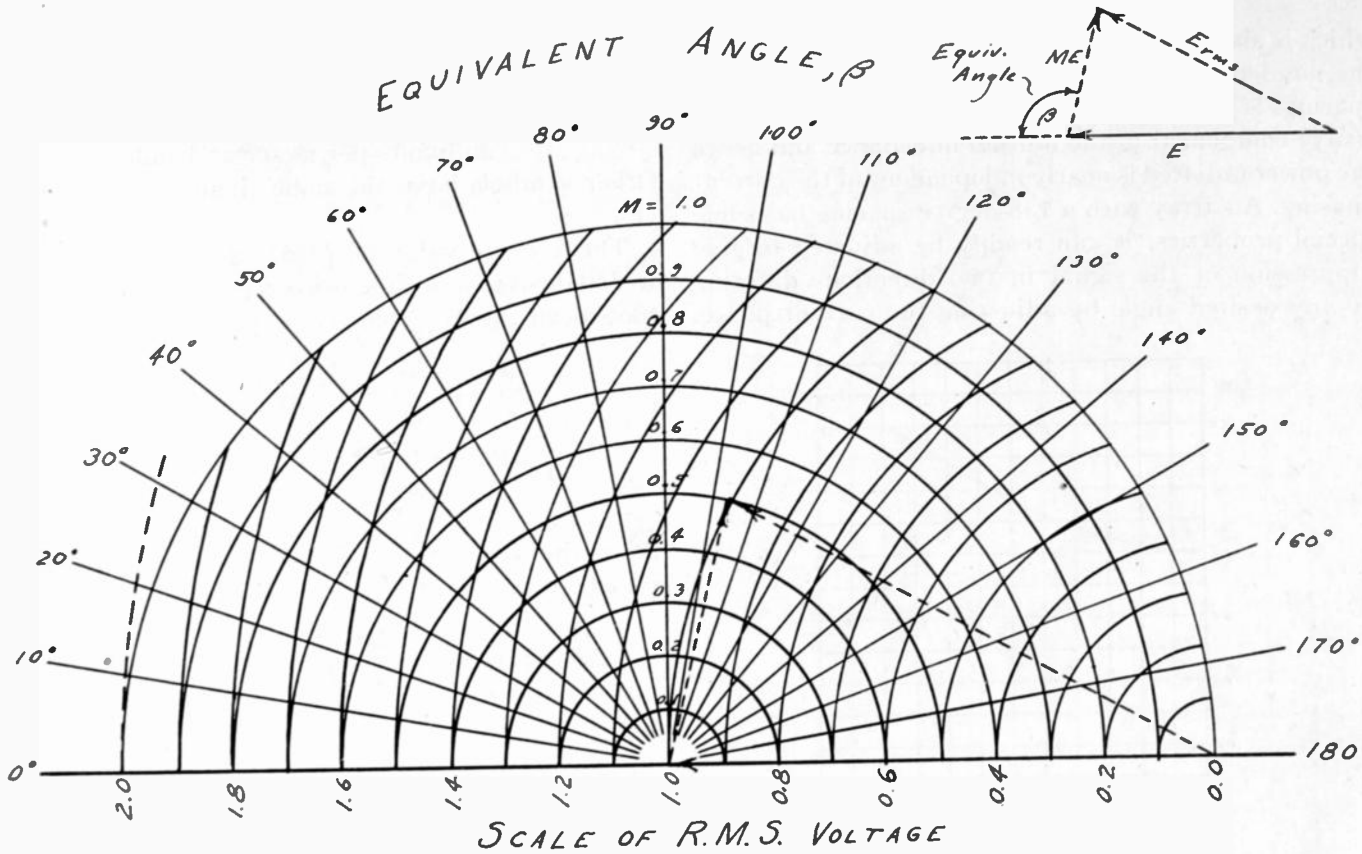

Fig. 2—Determination of root-mean-square value of field strength from equivalent angle and current ratio.

quantities; antenna-current ratio, antenna spacing, and phase difference of antenna currents indicated by the symbols M, S, and ψ, respectively. Thus, it is not easy to draw a set of curves covering all possible values of the three parameters. However, by means of the two charts given here it is possible to obtain the root-mean-square value of the horizontal pattern for all conditions. The charts are shown in Figs. 1 and 2.

In Fig. 1, a given spacing and phase difference angle of antenna currents determine an equivalent angle β. The equivalent angle β is transferred to Fig. 2 on which the simple triangular construction indicated gives the root-mean-square value of the pattern. In the example illustrated, for a current phase difference of 45 degrees and an antenna spacing of 180 degrees the value of the equivalent angle β is 103 degrees from the chart of Fig. 1. Application of this value to the chart of Fig. 2 shows that for a current ratio of 0.5, the root-mean-square value of the field strength is 1.01 E_1 where E_1 is the field that would have been produced by the antenna with the larger current radiating alone.

Some curves showing the variation of the root-mean-square value of the horizontal patterns for equal currents but for various values of phase and spacing are shown in Fig. 3. These values were obtained from the charts of Figs. 1 and 2.

It is observed from Figs. 1 and 2 that when the phase difference of current in the two antennas is ± 90 degrees, the root-mean-square value of the pattern is independent of the antenna spacing and is given by

$$E_{rms} = E_1\sqrt{1 + M^2}. \tag{2}$$

A curve of E_{rms}/E_1 against M for this condition is given in Fig. 4. The significance of the nondependence upon antenna spacing is that when the antenna currents are 90 degrees out of phase, there is no inter-

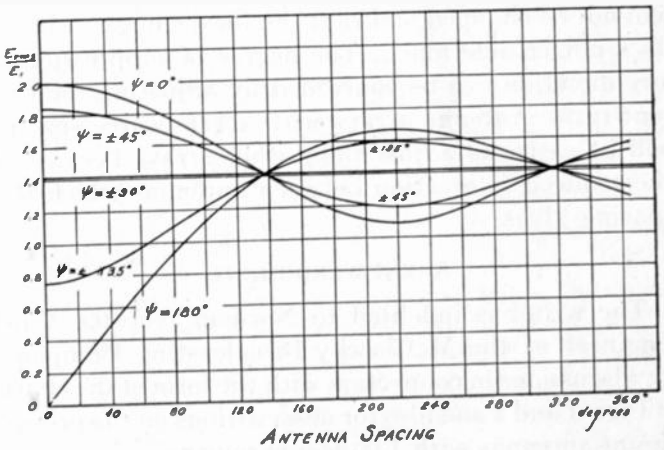

Fig. 3—Root-mean-square value of field strength as a function of current phasing and antenna spacing.

change of energy between the two antennas and they radiate independently of one another.

Also it will be observed that when the antenna spacing is 138, 316.5, or 496 degrees (corresponding to the first three roots of the zero-order Bessel function), then the root-mean-square value is independent of the phasing and is again given by

$$E_{rms} = E_1\sqrt{1 + M^2}$$

which is also represented by Fig. 4. The significance of the nondependence upon phasing for these particular spacings is that they give a nearly zero value of the resistive component of the mutual impedance and hence the power radiated is nearly independent of the current phasing. An array with a 138-degree spacing has some special properties. It can readily be adjusted to give suppression of the signal in two directions differing by any desired angle by adjusting the current phase.

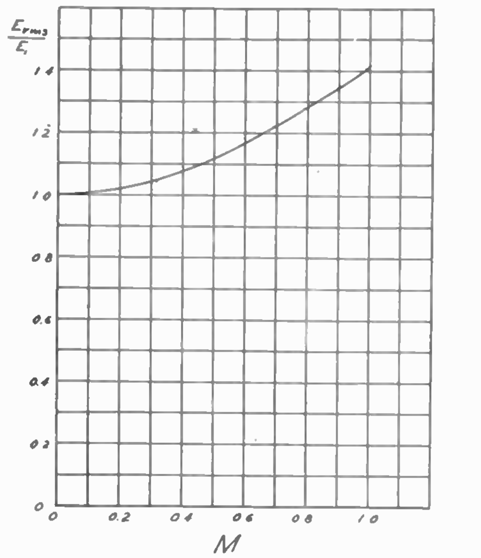

Fig. 4—Root-mean-square value of field strength as a function of current ratio when the current phasing is 90 degrees, or when spacing is 138.0, 316.5, or 496. degrees.

In the installation of such an array, adjustments of the current phase may be made with the assurance that the root-mean-square value of the horizontal pattern will not be changed and that the total energy radiated does not change much. The degree of suppression in any direction can be controlled by adjusting the current ratio. Antenna arrays with a 138-degree spacing will be easier to adjust than other arrays because of the reduced interaction between antennas which this spacing gives.

Acknowledgment

The writer is indebted to Norman Webster, Chief Engineer of the McClatchy Broadcasting Company, for discussions in connection with the form of the charts of Figs. 1 and 2 and also for observations on the properties of antennas with 138-degree spacing.

Appendix I

Derivation of Formula for Root-Mean-Square Value of Horizontal Radiation Pattern

The pattern in a horizontal plane of a two-element array is given by

$$E(\phi)\sin\omega t = E_1\sin\omega t + ME_1\sin(\omega t+\psi-S_r\cos\phi) \qquad (3)$$

where

S_r = the antenna spacing in radians, or $2\pi d/\lambda$

ϕ = the azimuth angle in degrees, measured clockwise from true north

$E(\phi)$ = the field strength in the direction ϕ in millivolts per meter at 1 mile

Other symbols have the same significance as used in (1).

The average value of $E^2(\phi)$ as ϕ is varied from 0 to 360 degrees and time is averaged over a cycle is the root-mean-square value given by

$$\frac{(E_{rms})^2}{2} = \frac{E_1^2}{2} + \frac{M^2E_1^2}{2} + \frac{2ME^2}{2}\cdot\frac{1}{2\pi}\int_{-\pi}^{\pi}\cos(\psi - S_r\cos\phi)d\phi. \qquad (4)$$

Since the average of $\sin^2\omega t$ is $\frac{1}{2}$ and the average of $\sin\omega t\cos\omega t$ is zero, the above can be written

$$(E_{rms})^2 = E_1\left[1+M+\frac{2M}{2\pi}\int_{-\pi}^{\pi}\cos(\psi-S_r\cos\phi)d\phi\right]. \qquad (5)$$

But

$$\int_{-\pi}^{\pi}\cos(\psi-S_r\cos\phi)d\phi = \int_{-\pi}^{\pi}\cos\psi\cos(S_r\cos\phi)d\phi + \int_{-\pi}^{\pi}\sin\psi\sin(S_r\cos\phi)d\phi. \qquad (6)$$

In the above the second integral is zero because the integrand is an odd function; i.e., values for corresponding positive and negative angles have opposite signs.

The nth order Bessel function is given by[5,6]

$$J_n(S) = \frac{1}{2\pi}\int_{-\pi}^{\pi}\cos(S\sin\phi - n\phi)d\phi. \qquad (7)$$

For $n=0$

$$J_0(S_r) = \frac{1}{2\pi}\int_{-\pi}^{\pi}\cos(S_r\sin\phi)d\phi \qquad (8)$$

which is the same as

$$J_0(S_r) = \frac{1}{2\pi}\int_{-\pi}^{\pi}\cos(S_r\cos\phi)d\phi. \qquad (9)$$

Hence

$$(E_{rms})^2 = E_1^2[1 + M^2 + 2M\cos\psi\cdot J_0(S_r)] \qquad (10)$$

or

$$E_{rms} = \sqrt{1 + M^2 + 2M\cos\psi\cdot J_0(S_r)}. \qquad (11)$$

When the spacing is expressed in degrees, this becomes

$$E_{rms} = E_1\sqrt{1 + M^2 + 2M\cos\psi\cdot J_0\left(\frac{360S}{2\pi}\right)}. \qquad (12)$$

[5] E. Jahnke and F. Emde, "Table of Functions," B. G. Teubner, Berlin, Germany, 1933, p. 218.

[6] R. S. Burrington and C. C. Torrance, "Higher Mathematics," McGraw-Hill Book Company, New York, N. Y., 1939, p. 440.

The Inclined Rhombic Antenna*

CHARLES W. HARRISON, JR.†, ASSOCIATE I.R.E.

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

THE horizontal rhombic antenna for receiving has been treated by Bruce, Beck, and Lowry.[1,2] Fading reduction by steering a horizontal rhombic antenna has been elaborated upon by Bruce and Beck.[3] Foster[4] has shown that the equations of Bruce, Beck, and Lowry are valid for the receiving case when the waves are polarized with the electric vector normal to the vertical plane containing the major axis of the antenna. The direction of maximum response for a receiving rhombic antenna is ordinarily in the vertical plane containing the major axis, and hence is of most interest.

Though other antenna types are now being shown greater favor, rhombic antennas are quite useful for receiving when a favorable front-to-undesired response is necessary. The horizontal rhombic antenna is usually designed to give maximum response at one vertical angle. Such an antenna is very effective only so long as the vertical angle of arrival of the incoming wave corresponds to the angle at which maximum response is obtained from the antenna. During periods of weak signals or fluctuations in the angle of arrival of the incoming wave, such an antenna may give results inferior to those obtainable from a simple doublet.

Some measure of control of the response at various angles of elevation can be obtained by inclining the plane of the antenna from the horizontal. The direct wave and the ground-reflected wave then add in a different manner from the case in which the plane of the antenna is horizontal. Analysis shows that an inclination of the antenna improves the response at both the low and the high angles of wave incidence simultaneously. Varying the side apex angle, the leg length, or height of a horizontal rhombic antenna merely improves the response in one direction, either to lower or to higher vertical angles.

In the formulas given by Bruce, Beck, and Lowry,[5] the phase reference is taken at the point of transmission-line attachment. When the antenna is inclined, it is desirable to shift the phase reference to the side apexes, about which the phase is symmetrical, before developing the formula for an inclined antenna. In (9) in the Appendix the formula for the free-space directivity of a rhombic antenna in the vertical plane containing the principal axis is developed using the side apexes as the phase reference.

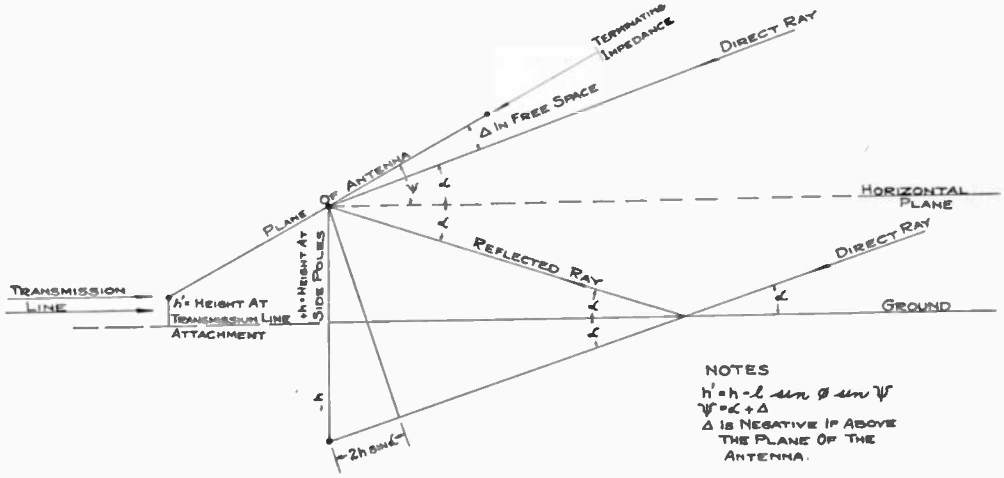

Fig. 1

This formula is

$$I_R = \frac{-j2E\lambda}{\pi Z_0} e^{-j2\pi l/\lambda} \left[\frac{\cos \phi}{1 - \sin \phi \cos \Delta} \right] \sin^2 \frac{\pi l}{\lambda} (1 - \sin \phi \cos \Delta) \qquad (1)$$

where E = the field strength
λ = the wavelength of the received signal
Z_0 = the characteristic impedance of the antenna
ϕ = the tilt angle (side semiangle)
l = the antenna leg length
Δ = the elevation angle from which the wave arrives, referred to the plane of the antenna

An inspection of equation (2)[6] given in the paper by Bruce, Beck, and Lowry reveals that the last two terms

* Decimal classification: R125. Original manuscript received by the Institute, April 19, 1941; revised manuscript received, October 14, 1941.

† Cruft Laboratory, Harvard University, Cambridge, Massachusetts.

[1] E. Bruce, "Developments in short-wave directive antennas," Proc. I.R.E., vol. 19, pp. 1406–1433; August, 1931.

[2] E. Bruce, A. C. Beck, and L. R. Lowry, "Horizontal rhombic antennas," Proc. I.R.E., vol. 23, pp. 24–46; January, 1935.

[3] E. Bruce and A. C. Beck, "Experiments with directivity steering for fading reduction," Proc. I.R.E., vol. 23, pp. 357–371; April, 1935.

[4] Donald Foster, "Radiation from rhombic antennas," Proc. I.R.E., vol. 25, pp. 1327–1353; October, 1937.

[5] See p. 46 of footnote 2.

[6] See p. 28 of footnote 2.

are identical to the last two terms of (1) above. These terms are of interest in this analysis.

In developing the equation of the receiver current for an antenna inclined with respect to ground, the direct ray must be added vectorially to the ground reflected ray. For this purpose the antenna is considered as being concentrated at its average height. The subject of ground reflection of horizontally polarized waves has been treated by numerous writers including Bruce, Beck, and Lowry.[2] The reader is referred to Fig. 2 of their paper, where it is shown that the as-

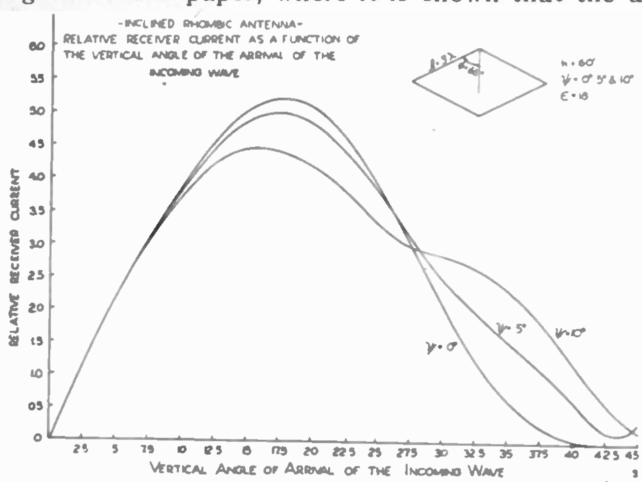

Fig. 2

sumption of a perfectly (conducting) ground (rather than a ground with finite constants) for the usual incident wave angles does not seriously affect the results. In general, it will be found that the resistivity of the earth may be neglected with negligible resulting error for antennas located over average soil, but it is desirable to take into account the fact that the magnitude of the reflected wave decreases as the angle of elevation in the plane of interest increases. The reflected wave, therefore, must be multiplied by a factor[7] given by

$$K = \frac{\sin\alpha - \sqrt{\epsilon - \cos^2\alpha}}{\sin\alpha + \sqrt{\epsilon - \cos^2\alpha}} \tag{2}$$

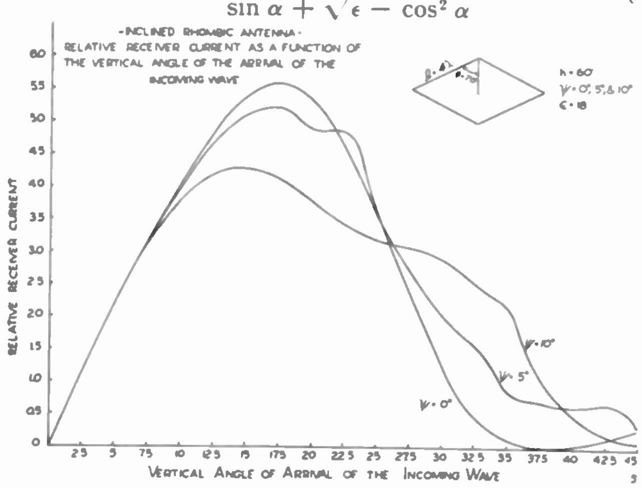

Fig. 3

[7] See, for instance, P. S. Carter, C. W. Hansell, and N. E. Lindenblad, "Development of directive transmitting antennas by R. C. A. Communications, Inc.," Proc. I.R.E., vol. 19, p. 1838, equation (91); October, 1931.

where ϵ = the dielectric constant of the ground

α = the elevation angle at which the relative response is desired, referred to the horizontal plane

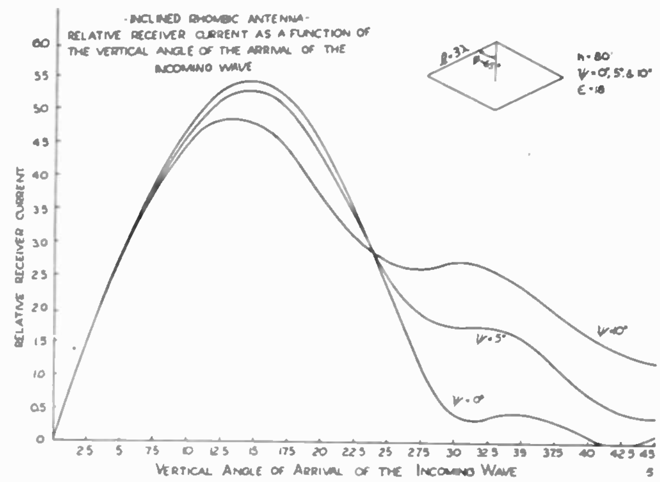

Fig. 4

Fig. 1 illustrates the method of taking into account the effect of the ground.

Here ψ = the angle of incline of the antenna, referred to the horizontal plane

h^1 = the height of the antenna at the point of transmission-line attachment

h = the height of the antenna at the side poles

It is seen that there is a path difference of $2h\sin\alpha$ between the direct and reflected rays.

The expression for the ground factor, taking into account the above remarks, is

$$Ke^{-j(4\pi h/\lambda)\sin\alpha}. \tag{3}$$

It is apparent from Fig. 1 that the angle made by the direct wave with the plane of the antenna is not equal to the angle made by the reflected wave with the plane of the antenna. These angles are equal only when the antenna is oriented horizontally.

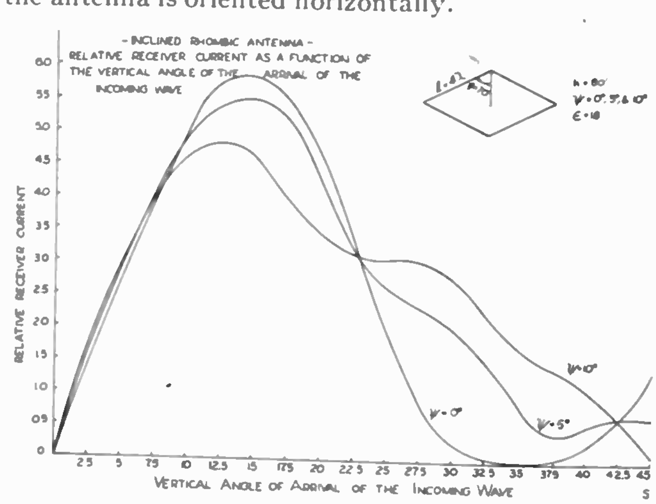

Fig. 5

One may now write the equation for the receiver current of an inclined rhombic antenna when the side poles are the reference for phases. It is of interest to note that in Fig. 1, $\Delta=\psi-\alpha$ and $\Delta+2\alpha=\psi+\alpha$. The final equation, using this notation, is

$$I_R = -j\frac{2E\lambda}{\pi Z_0}e^{-j2\pi l}\left\{\frac{\cos\phi}{1-\sin\phi\cos(\psi-\alpha)}\cdot\sin^2\frac{\pi l}{\lambda}(1-\sin\phi\cos(\psi-\alpha))\right.$$
$$\left.+Ke^{-j(4\pi h/\lambda)\sin\alpha}\cdot\frac{\cos\phi}{1-\sin\phi\cos(\psi+\alpha)}\cdot\sin^2\frac{\pi l}{\lambda}(1-\sin\phi\cos(\psi+\alpha))\right\}. \quad (4)$$

To determine the optimum angle of incline for a particular rhombic antenna is a matter of developing a series of patterns using (4).

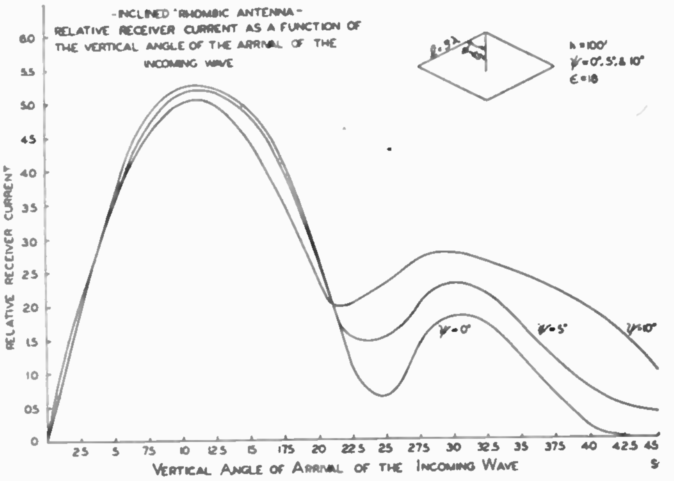

Fig. 6

In the curves which follow, relative receiver current is plotted as a function of the vertical angle of arrival of the incoming wave for two antennas. The relative receiver current at three incline angles is compared at three antenna heights for each antenna. The dielectric constant of the ground equals 18.

It is observed from these curves that the more a rhombic antenna is inclined, the greater the loss will be at the angle which would yield maximum response if the antenna were in a horizontal position; but at the same time, there is a decided improvement in response in the direction of higher elevation angles, as well as some improvement in response toward lower angles.

The principal effects obtained by inclining the rhombic antenna, therefore, are those of (1) increasing the range of vertical arrival angles to which the antenna responds and (2) making more uniform the antenna response over this range.

A large number of patterns have been developed which strongly indicate that the optimum incline angle lies within the range of 6 to 10 degrees.

The patterns obtained from (4) should be used on a relative basis. As such they will serve as a satisfactory criterion for the determination of the optimum angle at which a rhombic antenna should be inclined.

Acknowledgment

Lieutenant L. J. McKesson of the U. S. Navy Department suggested that an investigation be made of the theoretical performance of inclined rhombic antennas. Comments on the original manuscript by Mr. P. S. Carter of R.C.A. Communications, Inc., and by Dr. Ronold King of Harvard University are acknowledged.

Appendix

Development of the Equation for the Receiver Current of an Isolated Rhombic Antenna Due to a Wave in the Vertical Plane Containing the Principal Axis, Using the Side Apexes as the Reference for Phases

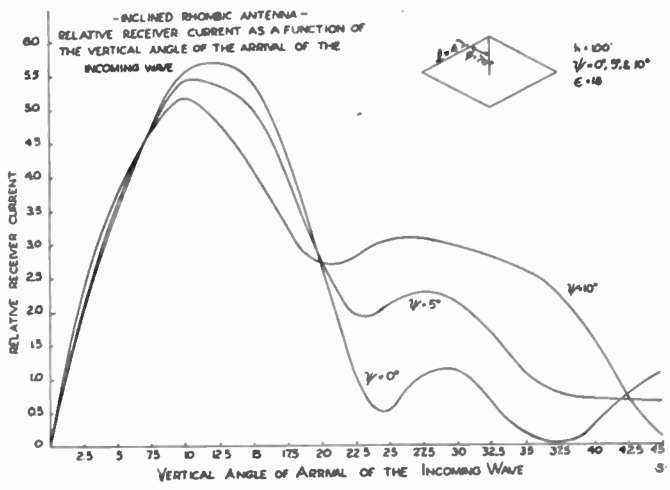

Fig. 7

Let E be the field strength and E_1, E_2, E_3, and E_4 represent the voltages induced in wires 1, 2, 3, and 4, respectively. The side apexes are the reference for phases. Then, for a wave in the vertical plane containing the principal axis

$$E_1 = E_3 = -E\cos\phi e^{-j(2\pi x/\lambda)\sin\phi\cos\Delta} \quad (5)$$

$$E_2 = E_4 = E\cos\phi e^{+j(2\pi x/\lambda)\sin\phi\cos\Delta} \quad (6)$$

$$I_R = 2\left\{\int_0^l \frac{E_1}{2Z_0}e^{-j(2\pi/\lambda)(l+x)}dx + \int_0^l \frac{E_2}{2Z_0}e^{-j(2\pi/\lambda)(l+x)}dx\right\}. \quad (7)$$

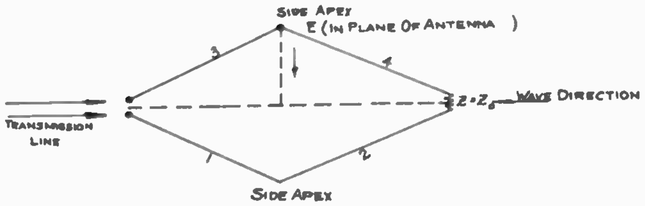

Fig. 8

Substituting (5) and (6) in (7) and integrating,

$$I_R = \frac{jE\lambda}{2\pi Z_0}e^{-j(2\pi l/\lambda)}\frac{\cos\phi}{1-\sin\phi\cos\Delta}\cdot\left[e^{j(2\pi l/\lambda)(1-\sin\phi\cos\Delta)} - 2 + e^{-j(2\pi l/\lambda)(1-\sin\phi\cos\Delta)}\right]. \quad (8)$$

Since

$$e^{ju} - 2 + e^{-ju} = \left[e^{j(u/2)} - e^{-j(u/2)}\right]^2$$

$$= \left[j2 \sin \frac{u}{2}\right]^2 = -4 \sin^2 \frac{u}{2},$$

$$I_R = \frac{-j2E\lambda}{\pi Z_0} e^{-j(2\pi l/\lambda)} \left[\frac{\cos \phi}{1 - \sin \phi \cos \Delta}\right]$$

$$\cdot \sin^2 \frac{\pi l}{\lambda} (1 - \sin \phi \cos \Delta). \tag{9}$$

Equation (9) is the expression for the receiver current due to a wave which lies in the vertical plane containing the major axis of a rhombic antenna located in empty space. It is assumed that the electric vector of the incoming wave lies in the plane of the antenna. The side apexes are taken as the reference for phases.

A Contribution to the Theory of Network Synthesis*

R. A. WHITEMAN†, ASSOCIATE, I.R.E.

Summary—*The design of corrective networks for transient conditions has been based to a great extent upon trial-and-error methods. Such a procedure has practical advantages at times, particularly when intuition may be relied upon for almost the entire solution of the problem. However, a procedure which eliminates trial-and-error methods not only provides a better understanding of the problem but also correlates and clarifies the basic theory involved in the solution of the problem. The results of the analysis presented in this paper extend O. Brune's[1] systematic procedure of synthesizing electrical networks based upon steady-state requirements, to the synthesis of electrical networks satisfying transient requirements. The extension of the method is accomplished with the use of the generalized infinite-integral formula expressed in (16). This equation gives a unique solution for the impedance function in terms of the transient voltage and the transient current. The impedance function obtained by evaluating the infinite integrals of (16) may then be synthesized by using Brune's method.*

I. Introduction

The problem of obtaining the transient current response produced by the sudden application of a voltage to a given network, is a problem in network analysis. If, however, the transient voltage and transient current are known functions of time and it is desired to obtain the requisite network, the problem is classified as a network-synthesis problem. The operational method of solving network-analysis problems was first developed in a heuristic manner by O. Heaviside[2] and systematized in a rigorous manner by T. J. I'A. Bromwich[3] using functions of a complex variable. Since the publication of Bromwich's treatise, complex-function theory has been used advantageously in the analysis and synthesis of electrical networks.[2]

With Bromwich's work in mind, it is very important to appreciate that the complex parameter p, used throughout this paper, is related to the operator d/dt. Heaviside used these symbols as equivalents and derived an infinite-integral theorem of a special form which first appeared in his "Electromagnetic Theory." A more general form of the theorem derived from Laurent's theorem[4] facilitates the synthesis of electrical networks by yielding the desired impedance function.

The following section presents a derivation and interpretation of the generalized formula based upon complex-function theory.

II. The Generalized Infinite-Integral Formula

The presentation given in this part of the paper is based upon Laurent's theorem, which may be stated as follows:

Let $f(\lambda)$ be a function analytic in the ring-shaped region between two concentric circles C_1 and C_2, of radii R_1 and R_2 ($R_2 < R_1$) and center λ_0, and on the circles themselves. Then $f(\lambda)$ can be expanded in a series of positive and negative powers of $(\lambda - \lambda_0)$, convergent at all points of the ring-shaped region.

The algebraic expression of this theorem is given by (1) and (2).
They are

$$f(\lambda) = \sum_{f=-\infty}^{f=+\infty} A(f)(\lambda - \lambda_0)^f \tag{1}$$

where,

$$A(f) = \frac{1}{2\pi j}\oint \frac{f(w)dw}{(w - \lambda_0)^{f+1}}. \tag{2}$$

By considering functions $A(f)$, which do not have singularities between minus infinity and plus infinity, it is possible to extend Laurent's theorem by allowing f to vary continuously between these limits. The summation expressed in (1) becomes an integration with the functions $f(\lambda)$, $f(w)$, and $A(f)$ changing to $\phi(\lambda)$, $\phi(w)$, and $B(f)$, respectively. The new equations become

$$\phi(\lambda) = \int_{-\infty}^{+\infty} B(f)(\lambda - \lambda_0)^f df \tag{3}$$

and

$$B(f) = \frac{1}{2\pi j}\oint \frac{\phi(w)dw}{(w - \lambda_0)^{f+1}}. \tag{4}$$

The complex quantity w locates points on the closed contour C which lies between C_1 and C_2 while λ locates

* Decimal classification: R390. Original manuscript received by the Institute, July 31, 1941; revised manuscript received, January 9, 1942.

† RCA Institutes, Inc., Chicago, Illinois.

[1] O. Brune, "Synthesis of a finite two-terminal network whose driving point impedance is a prescribed function of frequency," *Jour. Math. and Phys.*, vol. 10, pp. 191–236; October, 1931.

[2] O. Heaviside, "Electromagnetic Theory," three volumes.

[3] T. J. I'A. Bromwich, "Normal coordinates in dynamical systems," *Proc. Lond. Math. Soc.*, (2), vol. 15, pp. 401–448; 1916.

[4] Laurent, *Compt. Rend.*, vol. 17, p. 939, 1843.

any point within the same region. The quantities $(w-\lambda_0)$ and $(\lambda-\lambda_0)$ are also complex numbers locating these points within the region bounded by C_1 and C_2 but measured from the point λ_0. In general, these quantities may be given by

$$w - \lambda_0 = M\epsilon^{j2\pi t} \tag{5}$$

where M is a constant greater than R_2 and less than R_1.

Equations (3) and (4) give a mutual relationship between the functions $\phi(\lambda)$ and $B(f)$ and provide a more general pair of transforms than the more familiar Fourier transforms. These equations, however, may be arranged to a more convenient algebraic form by transforming the integrands to functions of t. The functions $\phi(\lambda)$ and $B(f)$ then become $G(t)$ and $F(f)$, respectively. With the aid of the differential of w, which is

$$dw = 2\pi jM\epsilon^{j2\pi t}dt \tag{6}$$

the transformed equations become

$$G(f) = \int_{-\infty}^{+\infty} F(f)M^f\epsilon^{j2\pi ft}df \tag{7}$$

and

$$F(f) = \int_{t_1}^{t_2} \frac{\epsilon^{-j2\pi ft}G(t)dt}{M^f}. \tag{8}$$

The quantity $(w-\lambda_0)^f$ is a periodic function with $1/f$ defined as the period and $F(f)$ as the coefficients of the periodic function. Since $G(t)$ is not periodic in the particular problem under consideration, equation (8) relates the coefficients $F(f)$ of the periodic function to the nonperiodic function $G(t)$. This basic concept is very useful when considering the impedance function as the ratio of the coefficients of the frequency spectra of a transient voltage and current. To transform the variables in (7) and (8) in order to make them applicable to electric-circuit-synthesis problems, let

$$\left.\begin{aligned} 2\pi fj &= p \\ j &= \sqrt{-1} \end{aligned}\right\}. \tag{9}$$

Substituting these transformations in (7) and (8), the equations for $G(t)$ and $F(f)$ become, respectively,

$$g(t) = \frac{1}{2\pi j}\int_{-j\infty}^{+j\infty} \frac{\epsilon^{pt}F(p)dp}{M^{-p/2\pi j}} \tag{10}$$

and

$$F(p) = \int_{t_1}^{t_2} \frac{\epsilon^{-pt}g(t)dt}{M^{p/2\pi j}}. \tag{11}$$

If, in (11), the function $g(t)$ is considered to be a time function specified between the limits t_1 and t_2, then $F(p)$ will represent the coefficients of each element of the periodic function in (10). When $g(t)$ is a transient voltage, then $F(p)$ is the coefficient of the periodic voltages expressed in (10). Let the transformation equations for a voltage be

$$\left.\begin{aligned} g(t) &= e(t) \\ F(p) &= E(p) \end{aligned}\right\} \tag{12}$$

and for a current be

$$\left.\begin{aligned} g(t) &= i(t) \\ F(p) &= I(p) \end{aligned}\right\}. \tag{13}$$

The quantities $E(p)$ and $I(p)$ express the magnitude and phase angle of the corresponding voltage and current coefficients for identical periods of these functions. The ratio of these coefficients express, in accordance with Ohm's law, the impedance function of the requisite network. That is

$$Z(p) = \frac{E(p)}{I(p)}. \tag{14}$$

By substituting (11), (12), and (13) in (14), the impedance function is expressible as

$$Z(p) = \frac{\int_{t_1}^{t_2}\epsilon^{-pt}e(t)dt}{\int_{t_1}^{t_2}\epsilon^{-pt}i(t)dt}. \tag{15}$$

Equation (15) is the solution of the problem of relating the impedance function to the transient voltage $e(t)$ and the transient current $i(t)$. When t_1 approaches minus infinity and t_2 approaches plus infinity, the equation shall be referred to as the *generalized infinite-integral formula*. Upon substituting the new limits, (15) becomes

$$Z(p) = \frac{\int_{-\infty}^{+\infty}\epsilon^{-pt}e(t)dt}{\int_{-\infty}^{+\infty}\epsilon^{-pt}i(t)dt}. \tag{16}$$

It is important to notice that the expression for $Z(p)$ obtained with (16) applies only when a voltage of the form specified by $e(t)$ is applied. If $e(t)$ is the "Heaviside unit-step function," (16) reduces to Heaviside's infinite-integral theorem. The impedance function obtained with (16) in one operation possesses the attenuation and phase-shift characteristics implied in the prescribed transient conditions.

This procedure is more direct than that used when the attenuation and phase-shift characteristics are prescribed explicitly as a function of frequency. In the latter method, it is necessary to solve two problems. The first involves a synthesis of the attenuation network without regard to the phase-shift characteristics, and the second involves a synthesis of the phase-shifting network to obtain the prescribed phase-shift characteristics.

If the impedance function $Z(p)$ represents the behavior of a two-terminal network, the structure may be synthesized by using the method devised by Brune; if however, the function represents the transfer-impedance function of a four-terminal network, the structure may be synthesized by using the method developed by C. M:son Gewertz.[6] The four-terminal

[6] C. M:son Gewertz, "Synthesis of a finite four-terminal network from its prescribed driving-point functions and transfer function," *Jour. Math. and Phys.*, vol. 12, pp. 1–257; January, 1932–1933.

network problem requires, for its solution, supplementary data in the form of prescribed driving-point impedance functions.

An interesting illustrative example of applying (16) is presented in the following part of this paper, where a nonlinear resistance is replaced by a linear passive equivalent network.

III. Equivalent Linear Network of Nonlinear Resistance

A linear passive network shall be considered equivalent to a nonlinear resistance, when both networks behave in an identical manner under prescribed transient conditions. If the nonlinearity of the resistance is expressed by relating the current as a function of the voltage as

$$i = f(e) \tag{17}$$

the impedance function of the linear passive equivalent network is given by substituting $e(t)$ and $i(t)$ in (16). After making this substitution, the function becomes

$$Z(p) = \frac{\int_0^\infty \epsilon^{-pt} e(t) dt}{\int_0^\infty \epsilon^{-pt} f(e) dt} \tag{18}$$

which may be evaluated for a typical nonlinear resistance such as Thyrite.[6]

For a certain sample of this material, the relation between the current and the voltage was expressed by

$$i = Ae^m \tag{19}$$

where i is the current in amperes, e is the voltage in volts, and

$$A = 3.2 \times 10^{-8}$$

and $$m = 3.57.$$

Now, let the transient voltage be

$$e = B\epsilon^{-\alpha t}$$

where $$B = 10$$

and $$\alpha = 10.$$

Then $$i(t) = AB^m \epsilon^{-m\alpha t}$$

and the impedance function after evaluating the integrals becomes

$$Z(p) = \frac{p + m\alpha}{AB^{m-1}p + AB^{m-1}\alpha}. \tag{20}$$

A network obtained by Brune's method, which satisfies $Z(p)$, consists of a resistance R_1 in series with a parallel combination of a resistance R_2 and a capacitance C. The numerical values of these parameters are

$$R_1 = 8.44 \times 10^4 \text{ ohms}$$

$$R_2 = 21.7 \times 10^4 \text{ ohms}$$

$$C = 0.46 \times 10^{-6} \text{ farad.}$$

This passive linear network is equivalent to the nonlinear resistance made of Thyrite.

IV. Conclusion

The synthesis of networks that has been described provides a means of designing electrical circuits which behave in accordance with prescribed transient conditions. This procedure extends and incorporates methods now employed using steady-state conditions as prescribed conditions. The attenuation and phase-shift characteristics implied in the prescribed transient conditions are automatically taken care of in the procedure of the synthesis and thereby introduce an unusual simplification.

[6] T. Brownlee, "The calculation of circuits containing Thyrite," *Gen. Elec. Rev.*, April, 1934.

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(1) H. Nyquist, "Regeneration theory," *Bell Sys. Tech. Jour.*, vol. 11, pp. 126–147; January, 1932.

(2) H. W. Bode, Patent No. 2,123,178.

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(4) N. W. McLachlan, "Complex Variable and Operational Calculus with Technical Applications," Cambridge University Press, New York, N. Y., 1939.

Butler Airphotos

Air View of Downtown Cleveland.

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

Cleveland, sixth city of the United States, is located on the southern shore of Lake Erie. Settled at the close of the eighteenth century, primarily as a trading center, it has developed into one of the country's most important transportation centers, produces huge quantities of iron and steel, and is the home of a long list of diversified industries. Iron ore from the Lake Superior region and limestone and coal from Ohio and Pennsylvania provide the raw materials for Cleveland's major industries—iron and steel.

Our Cleveland Section will be hosts at this convention which is scheduled for June 29, 30, and July 1. Headquarters will be at the Statler Hotel. Under war conditions, no program can be guaranteed so the following schedule must be considered as being our best approximation at this time.

PAPERS PROGRAM

The program of papers to be presented follows. All technical sessions will be in the Euclid Ballroom and there will be no duplicate sessions. Sufficient time is available for presenting all papers substantially in full and the time allotted for each paper is indicated. No papers are available in preprint or separate form. There is no assurance that any of the papers will appear in the PROCEEDINGS although it is expected that many will be published during the next several months. Summaries of the papers are given at the end of this convention announcement and are arranged alphabetically by names of the authors. The order of presentation is indicated by numbers.

PROGRAM

Monday, June 29

10:30 A.M.–1:00 P.M.

Addresses of Welcome by A. F. Van Dyck, President of the Institute; P. L. Hoover, Chairman of the Cleveland Section; and Carl E. Smith, Chairman of the Convention Committee.

1. "Recording Standards," by I. P. Rodman, Columbia Recording Corporation, New York, N. Y. (20 Minutes)
2. "A New Approach to the Problem of Phonograph Reproduction," by G. L. Beers, and C. M. Sinnett, RCA Manufacturing Company, Camden, N. J. (25 Minutes)
3. "Measuring Transcription-Turntable-Speed Variations," by H. E. Roys, RCA Manufacturing Company, Indianapolis, Ind. (20 Minutes)
4. "A New Type of Practical Distortion Meter," by J. E. Hayes, Canadian Broadcasting Corporation, Montreal, Que., Canada. (20 Minutes)

Rose Garden at the Cleveland Art Museum.

5. "Frequency-Modulation Distortion in Loudspeakers," by G. L. Beers and H. Belar, RCA Manufacturing Company, Camden, N. J. (25 Minutes)

2:30 P.M.–5:00 P.M.

6. "Radio-Frequency Oscillator Apparatus and Its Application to Industrial Process-Control Equipment," by T. A. Cohen, Wheelco Instruments Company, Chicago, Ill. (30 Minutes)
7. "The Scanning Microscope," by V. K. Zworykin, J. Hillier, and R. Snyder, RCA Manufacturing Company, Camden, N. J. (45 Minutes)
8. "Spectroscopic Analysis in the Manufacture of Radio Tubes," by S. L. Parsons, Hygrade Sylvania Corporation, Emporium, Pa. (30 Minutes)
9. "Minimizing Aberration of Electron Lenses," by H. Poritsky, General Electric Company, Schenectady, N. Y. (20 Minutes)

Tuesday, June 30

10:00 A.M.–1:00 P.M.

10. "Maintenance of Broadcasting Operations During Wartime," by J. A. Ouimet, Canadian Broadcasting Corporation, Montreal, Que., Canada. (30 Minutes)
11. "High-Power Television Transmitter," by H. B. Fancher, General Electric Company, Schenectady, N. Y. (40 Minutes)
12. "Frequency-Modulation Transmitter-Receiver for Studio Transmitter Relay," W. F. Goetter, General Electric Company, Schenectady, N. Y. (40 Minutes)
13. "Effect of Solar Activity on Radio Communication," by H. W. Wells, Carnegie Institution of Washington, Washington, D. C. (30 Minutes)

2:30 P.M.–4:30 P.M.

14. "Television Video Relay System," by J. E. Keister, General Electric Company, Schenectady, N. Y. (40 Minutes)
15. "Mercury Lighting for Television Studios," by C. A. Breeding, General Electric Company, Schenectady, N. Y. (25 Minutes)
16. "The Focusing-View-Finder Problem in Television Cameras," by G. L. Beers, RCA Manufacturing Company, Camden, N. J. (20 Minutes)
17. "Automatic Frequency and Phase Control of Synchronization in Television Receivers," by K. R. Wendt and G. L. Fredendall, RCA Manufacturing Company, Camden, N. J. (20 Minutes)

Wednesday, July 1

10:00 A.M.–1:00 P.M.

18. "Radio Strain Insulators for High Voltage and Low Capacitance," by A. O. Austin, A. O. Austin, Barberton, Ohio. (30 Minutes)
19. "Improved Insulators for Self-Supporting or Sectionalized Towers," by A. O. Austin, A. O. Austin, Barberton, Ohio. (30 Minutes)
20. "Brief Discussion of the Design of a 900-Foot Uniform-Cross-Section Guyed Radio Tower," by A. C. Waller, Truscon Steel Company, Youngstown, Ohio. (30 Minutes)
21. "Circular Antenna," by M. W. Scheldorf, General Electric Company, Schenectady, N. Y. (30 Minutes)
22. "Stub-Feeder Calculations," by H. A. Brown and W. J. Trijitzinsky, University of Illinois, Urbana, Ill. (20 Minutes)

REGISTRATION

The registration desk will be opened from 4:00 P.M. to 6:00 P.M. on Sunday, June 28. Those who arrive early and can register during this period will assist greatly in relieving the peak load which occurs on the opening morning. During the three days of the meeting, the registration desk will be open from 9:00 A.M. until the last technical session for the day terminates.

TRIP

Wartime conditions have placed severe limitations on inspection tours of the type usually included in Institute convention programs. Only one trip has been arranged for the men and is scheduled for Wednesday afternoon and evening. Restrictions having already

The General Electric Group at Nela Park.

been placed on the use of busses for such purposes, the trip will be made by street car.

The first stop will be at the Picker X-Ray Corporation where the manufacture of X-ray and accessory equipment will be inspected. From there we will go to Nela Park, which is the lighting research laboratories of the General Electric Company. A discussion and demonstration of industrial lighting for war production and home lighting will be given. After dinner, which will be served at Nela Park (tickets $1.50 each), a joint session with the women will be devoted to blackout lighting, light magic, and colored movies.

The Warner-Swasey Observatory will be our final stop and we are invited to inspect the largest Schmidt telescope in the world. This telescope uses a 36-inch reflector in conjunction with a 24-inch lens and provides greater light-gathering power than any telescope of comparable size. The weather permitting, we'll have a look through it.

BANQUET

On Tuesday evening at 7:00 o'clock, the doors of the air-conditioned Euclid Ballroom will be opened for the banquet. Tickets will be on sale at $3.00 each at the registration desk. Dress will be optional.

"Strange Adventures in Discomania" will be presented by George C. A. Hantelman, Manager of the Cleveland Engineering Society.

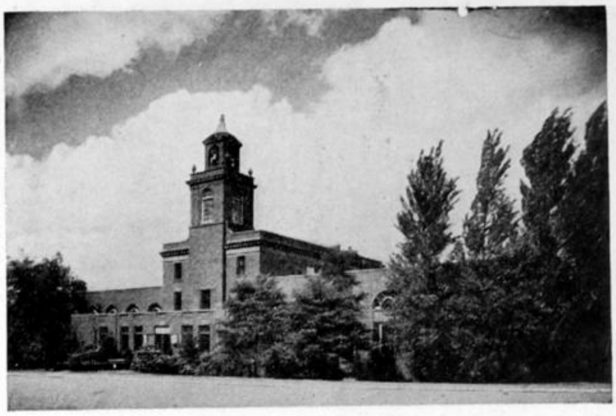
EXHIBITION

There will be no exhibition at this convention.

SECTIONS COMMITTEE

A meeting of the Sections Committee will be held in the Pine Room at 7:00 P.M. on Monday, June 29. It is hoped that most of the Sections will be represented at the meeting which will be devoted to a discussion of Section and Institute affairs. If you plan on attending the Convention and are not sure your Section has arranged for this meeting, get in touch with the Secretary of the Section and offer your services. It is only by getting opinions from our Sections and members that the Institute can make its services most effective.

The Cleveland Museum of Art and Fine Arts Gardens.



General Electric Institute at Nela Park.

WOMEN'S PROGRAM

The women are invited to avail themselves of the advance registration provided from 4:00 P.M. to 6:00 P.M. on Sunday, June 28. Their registration desk will be open at 9:00 o'clock each morning thereafter.

Monday, June 29

Monday morning, June 29, will be devoted to getting acquainted. From 2:00 P.M. to 3:00 P.M. a musical program will be given in the studios of WGAR on the fourteenth floor of the hotel. Tea will be served from 3:00 P.M. to 5:00 P.M. The evening will be free; the program of the men is similarly open and this will be a good time to see some of the town.

Tuesday, June 30

At 10:30 A.M. a tour through the Higbee Company Department store will start. The WHK Studios will then be visited and the morning will close with a look at Cleveland from the Terminal Tower.

At noon, luncheon will be served as a style show is presented. The afternoon may be spent at the Higbee store or on a trip to the Art Museum or any of the other interesting places around University Circle. The banquet is the feature of the evening.

Wednesday, July 1

The morning is reserved for shopping or catching up on sleep. Those who are interested, may go with the men to the Picker X-Ray Corporation plant to see the manufacture of X-ray and related equipment.

At 3:30 P.M. those who do not go with the men, will leave for Nela Park where a demonstration of household lighting will be given. The women will join the men for dinner and the later demonstrations of blackout lighting, light magic, and colored movies. All will then visit the Warner-Swasey Observatory to inspect its outstanding collection of astronomical equipment. It is hoped that the weather will permit some high-powered star gazing by everyone.

Downtown Cleveland Pictured from the Air.

SUMMARIES OF TECHNICAL PAPERS

18. RADIO STRAIN INSULATORS FOR HIGH VOLTAGE AND LOW CAPACITANCE

A. O. Austin
(A. O. Austin, Barberton, Ohio)

The high-voltage requirements to withstand both the radio-frequency and lighting loads will be given. There will then be considered the effects of the electrical requirements on the mechanical reliability. The use of radio-type insulators in the electrical power field to eliminate interference to radio systems will be discussed. Renewed interest in low-frequency transmission is recognized by a discussion of strain insulators for high-powered stations of that type.

19. IMPROVED INSULATORS FOR SELF-SUPPORTING OR SECTIONALIZED TOWERS

A. O. Austin
(A. O. Austin, Barberton, Ohio)

Mechanical requirements for insulators for self-supporting or sectionalized towers will be discussed and include working loads and mechanical hazards. The mechanical properties of several types of insulators will be covered. The effect of the electrical requirements on the mechanical design will then be considered. The advantages of the heated types of insulators and limitations resulting from thermal stresses will be treated.

16. THE FOCUSING-VIEW-FINDER PROBLEM IN TELEVISION CAMERAS

G. L. Beers
(RCA Manufacturing Company, Camden, N. J.)

The technical excellence of a television program may frequently depend on the characteristics of the view finder used in the television camera. Conditions peculiar to television make it desirable that television camera view finders be of the focusing type. The requirements of an ideal view finder of this type are discussed. During the past ten years a number of view-finder arrangements have been investigated in connection with the development of television cameras. Several of these are described and their relative merits indicated.

5. FREQUENCY-MODULATION DISTORTION IN LOUDSPEAKERS

G. L. Beers and H. Belar
(RCA Manufacturing Company, Camden, N. J.)

As the frequency-response range of a sound-reproducing system is extended, the necessity for minimizing all forms of distortion is correspondingly increased. The part which the loudspeaker can contribute to the over-all distortion of a reproducing system has been frequently considered. A type of loudspeaker distortion which has not received general consideration is described. This distortion is a result of the Doppler effect and produces frequency modulation in loudspeakers reproducing complex tones. Equations for this type of distortion are given. Measurements which confirm the calculated distortion in several loudspeakers are shown. An appendix giving the derivation of the equations is included.

2. SOME RECENT DEVELOPMENTS IN RECORD-REPRODUCING SYSTEMS

G. L. Beers and C. M. Sinnett
(RCA Manufacturing Company, Camden, N. J.)

Several factors of importance in obtaining satisfactory reproduction of sound from lateral-cut phonograph records are considered. An experimental record reproducing system employing the principles of frequency modulation is described and data are supplied on the measured and calculated performance characteristics of the system. Curves are included showing the vertical force required for satisfactory tracking with the experimental frequency-modulation pickup as compared with other pickups of conventional design. The paper covers development work which was done several years back.

15. MERCURY LIGHTING FOR TELEVISION STUDIOS

C. A. Breeding
(General Electric Company, Schenectady, N. Y.)

General Electric experiments with water-cooled mercury lamps for television studio lighting began at the New York World's Fair in 1939 and finally culminated in a complete installation in the new modern studios of WRGB at Schenectady in the fall of 1941. The installation in Schenectady is described and experience to date with these lights is outlined.

22. STUB-FEEDER CALCULATIONS

H. A. Brown and W. J. Trijitzinsky
(University of Illinois, Urbana, Ill.)

For reasons of economy, radio-frequency power is usually transmitted over nonresonant lines which are terminated in their characteristic impedances. The terminating impedance is usually a transforming network to match the impedance of the driven load to obtain maximum power transfer.

At high radio frequencies, short sections of open or closed lines are used to terminate the transmission line. These short sections, often termed stub feeders, can be made to transform from the nonresonant-line characteristic impedance to the given antenna input impedance. In this paper, derivations of formulas for the lengths of the matching-line sections are given.

6. RADIO-FREQUENCY OSCILLATOR APPARATUS AND ITS APPLICATION TO INDUSTRIAL PROCESS-CONTROL EQUIPMENT

T. A. Cohen
(Wheelco Instruments Company, Chicago, Ill.)

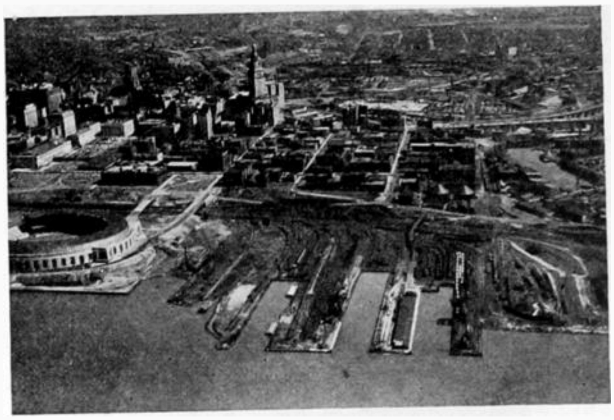
Industrial process-control instrumentation has been an art which has occupied the study of many engineers in the past decade. One of its major problems has been the development of sensitive relay equipment, capable of responding to the command of a sensitive measuring instrument or apparatus without reactively disturbing the primary measurements.

The advent of electronic industrial apparatus has given a great impetus to sensitive-instrument-control mechanisms, especially where radio-frequency oscillators and kindred apparatus are included in the control instrument. The development and application of this type of instrumentation to such problems as temperature control, liquid-level control, and the like are of great interest, due to the fact that communication arts are intimately tied up with manufacturing problems of many types in which such process-control equipment is of essential importance.

11. HIGH-POWER TELEVISION TRANSMITTER

H. B. Fancher
(General Electric Company, Schenectady, N. Y.)

The design factors involved in the development of Station WRGB, located in the Helderberg Mountains, including a 40-kilowatt visual transmitter and a 20-kilowatt aural transmitter are described. The part of the visual transmitter located at the main station consists of a high-frequency receiver, a converter, and a chain of linear class B push-pull amplifiers. The transmitter receives a standard modulated vestigial-sideband signal from its Schenectady studio or from the New York relay station, in each case over a high-frequency radio link. The two final stages each consist of a pair of water-cooled triodes (GL-8009) which were especially designed for television service. The special features involved in building stable circuits for these tubes are considered. The visual antenna consists of a conical doublet fed by an open-wire shielded line which matches the antenna impedance. The output frequency is maintained constant independent of variations in the incoming signal by means of a unique automatic frequency-control system.



A View of Cleveland from the Lake.

The aural transmitter consists of a 50-watt exciter unit containing the oscillator, modulator, and frequency-control unit; a 2-kilowatt amplifier consisting of air-cooled triodes; and a 20-kilowatt amplifier using a pair of tubes similar to those in the visual power amplifier. The antenna consists of a cubic array giving a circular field pattern.

12. FREQUENCY-MODULATION TRANSMITTER-RECEIVER FOR STUDIO-TRANSMITTER RELAY

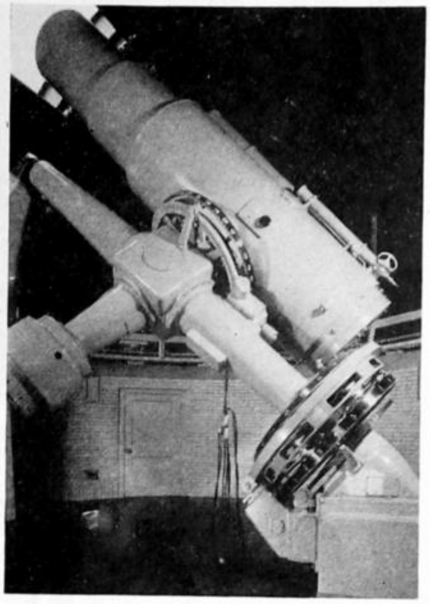
W. F. Goetter
(General Electric Company, Schenectady, N. Y.)

A complete system for high-fidelity relaying between the studio and main transmitter is described. The entire equipment was designed considering simplicity and reliability to be of prime importance.

The 25-watt transmitter incorporates several novel features which account for the excellent performance obtained. New design tubes especially suited for ultra-high-frequency operation are used.

A crystal-controlled, double-conversion, superheterodyne receiver, employing such features as cascade limiting, carrier-off noise suppression, and vertical chassis construction, is also described. Harmonics from the same crystal oscillator are used in performing both conversions, resulting in an extremely stable unit. Both transmitter and receiver may be remotely controlled when proper compliance is made with Federal Communications Commission regulations.

A high-gain studio-transmitter antenna, which meets all Federal Communications Commission requirements, is totally enclosed against the weather to avoid ice-melting problems, etc. A frequency-modulation station monitor, for studio-transmitter



Schmidt-type Telescope at the Warner-Swasey Observatory.

applications, indicates carrier frequency continuously, as well as per cent modulation and carrier level. Aural monitoring is also obtained from the same unit.

4. A NEW TYPE OF PRACTICAL DISTORTION METER

J. E. Hayes
(Canadian Broadcasting Corporation, Montreal, Que., Canada)

This paper gives a description of a distortion meter embodying circuits which differ somewhat from those previously employed for this type of instrument. It consists essentially of a bridged-T audio-frequency bridge circuit, in which the inductance element is replaced by a reactance-tube circuit. Because of the flexibility obtainable in vacuum-tube circuits, it is a relatively simple matter to vary the effective inductance continuously over a fairly wide range, and thus allow the distortion meter to be used at any frequency in the audio range.

Certain precautions must be taken in a circuit of this type in order to avoid difficulties due to nonlinear action of the reactance-tube circuit. Application of negative feedback to the reactance-tube circuit effectively reduces the nonlinearity, increases stability, and at the same time keeps tube noise and hum at a minimum level. An analysis of the reactance part of this circuit, together with formulas for calculating the effective Q and the optimum operating conditions are including.

14. A TELEVISION RELAY SYSTEM

J. E. Keister
(General Electric Company, Schenectady, N. Y.)

A television "repeater" type of station is described. The application is unique in that the station is 130 miles distant and over a mile below line of sight from WNBT, the station originating the programs. The visual signal is received on a 400-foot by 150-foot rhombic antenna. It is then amplified, converted, and further amplified to be retransmitted to the main station, WRGB, at 163.25 megacycles.

A novel system of output-frequency control is described. This system accurately maintains the output frequency regardless of error in the received signal. It also is capable of handling the frequency-modulation type of synchronizing pulse should the occasion arise.

The aural signal is demodulated in the conventional manner and carried to the main station over telephone lines.

10. MAINTENANCE OF BROADCASTING OPERATIONS DURING WARTIME

J. A. Ouimet
(Canadian Broadcasting Corporation, Montreal, Que., Canada)

The technical problem of wartime broadcasting will be discussed in its three main aspects.

1) The maintenance of normal operations under conditions of war economy.
2) The maintenance of normal or restricted operations during actual war action.
3) The maintenance or resumption of essential operations after partial or complete destruction of broadcasting facilities.

Within this broad outline, certain details of plant and operations protection which are of interest to the broadcast engineer will be discussed. Whenever possible this will be illustrated with practical measures taken at various plants of the Canadian Broadcasting Corporation.

8. SPECTROGRAPHIC ANALYSIS IN THE MANUFACTURE OF RADIO TUBES

S. L. Parsons
(Hygrade Sylvania Corporation, Emporium, Pa.)

This paper will describe the use of spectrographic methods in attacking some of the problems encountered in the manufacture of radio tubes. Slides of the laboratory and apparatus will be shown and discussed. The techniques and use of qualitative spectrographic analysis in connection with chemical, metallurgical, ceramic, and fluorescent problems will be described and illustrations of the application of quantitative spectrographic analysis to problems of routine inspection and control will be given.

9. MINIMIZING ABERRATION OF ELECTRON LENSES

H. Poritsky
(General Electric Company, Schenectady, N. Y.)

This paper is concerned with the question of minimizing spherical aberration. Several possible ways of obtaining sharp focusing are investigated. A geometrical analogy between the electron paths and geodesics on a proper surface, referred to in the paper as a characteristic surface, enables the author to utilize various results familiar in differential geometry of surfaces in the investigation of the electronic problem. This analogy to geodesics suggests a study of the class of fields for which the electronic problem corresponds to the two geometric problems of geodesics on surfaces of constant curvature and on surfaces of rotation. It is shown that only for proper fields possessing space charge can the characteristic surfaces be of the above kind. A further case of sharp focusing, which is investigated, is the case where, for all points on the same plane perpendicular to the axis of symmetry, the electron paths all point to the same point on that axis. This case is essentially equivalent to the one investigated by Gray. Again it is shown that no such family of motions can arise for axially symmetric electrostatic fields.

Necessary and sufficient conditions are then investigated for focusing which is sharp to within a certain order. A condition for third-order focusing is established which consists in the vanishing of a certain integral, the integrand of which involves the solution of the first-order equation of electron motion as well as the potential and its derivatives along the axis of symmetry. Similar conditions for higher-order focusing are established and written in many equivalent forms.

1. RECORDING STANDARDS

I. P. Rodman
(Columbia Recording Corporation, New York, N. Y.)

Last June, under the sponsorship of the National Association of Broadcasters, a committee was established to prepare standards for recordings and associated equipment used for broadcasting. The report of this committee has recently been approved.

Mechanical dimensions of records and direction and speed of rotation, are covered. Also included are electrical characteristics, recording level, signal-to-noise ratio, and wow-factor measurements. Definitions of terms used in the field are incorporated in these standards.

3. MEASURING TRANSCRIPTION-TURNTABLE-SPEED VARIATIONS

H. E. Roys
(RCA Manufacturing Company, Indianapolis, Ind.)

One of the important aspects of turntable design is that of constancy of speed, or freedom from "wows." This paper outlines previous methods of "wow" measurement in turntables and describes improvements both in apparatus and method. Factors involved in the measurement of turntable-speed variations are given together with means of minimizing error caused by different rates of variation.

21. CIRCULAR ANTENNA

M. W. Scheldorf
(General Electric Company, Schenectady, N. Y.)

A new horizontally polarized antenna is described. An outstanding feature is the radiation of substantially uniform energy in all directions about the antenna without resorting to a complex structure having phasing networks to secure this pattern. Its low vertical radiation gives a twofold improvement, optimum spacing in an arrangement having more than one bay giving more gain per bay than previously experienced and coupling between bays in such an arrangement having been reduced so that adjustments are simplified.

A given physical structure may be adjusted to resonate over a wide frequency range by a simple means. The structure has a pleasing appearance particularly suited to mounting on simple vertical masts.

20. BRIEF DISCUSSION OF THE DESIGN OF A 912-FOOT UNIFORM-CROSS-SECTION GUYED RADIO-TOWER

A. C. Waller
(Truscon Steel Company, Youngstown, Ohio)

A general presentation is given of the structural factors involved in the design of a 190-degree antenna on the 570-kilocycle frequency of radio station WNAX at Yanktown, South Dakota. (The design was prepared concurrently with that for a 900-foot Franklin antenna for radio station WKY of Oklahoma City, Oklahoma.)

Resistance to wind involves a study of meteorological data for the area in question and determination of probable future frequency and combination of wind intensity, temperature range, and ice deposit. Effect of anticipated wind velocity in terms of actual pressure on the complex truss forms used is empirically fixed as based on experience, known outdoor measurements, and wind-tunnel tests on similar forms. Correct balance of anticipated external loads and steel stresses results in proper factors of safety.

View of Electrical Assembly Department at the Picker X-Ray Corporation.

Effect of temperature on guy catenary lengths, elasticity and coefficients of expansion of porcelain and steel, and effect of connections upon deflection, are among other variables to be taken into account.

Probable relation between wind velocity at various elevations, "scale" factor of wind pressure on cylindrical members, and foundation problems introduced by electrically favorable radio-antenna locations, are also of interest.

13. EFFECTS OF SOLAR ACTIVITY ON THE IONOSPHERE AND RADIO COMMUNICATION

H. W. Wells
(Carnegie Institution of Washington, Washington, D. C.)

Unusual solar activity produces severe disturbances in the ionosphere which in turn directly influence radio communication. Solar flares or sunspot eruptions have been definitely identified as the origin of short-period radio fade-outs. The ultraviolet radiation associated with the solar flare immediately produces intense ionization in the lower part of the ionosphere which results in complete absorption of all normal sky-wave radio transmission. Disturbances of this nature seldom last longer than an hour.

Union Terminal Building housing the Higbee Store and WHK Studios.

However, the radio disturbances which are most severe are coincident with intense magnetic storms. Such magnetic storms are frequently associated with active sunspot areas. It is generally believed that streams of corpuscles are shot out from the active sunspots. These streams travel to the earth in one to four days and produce magnetic storms, auroral displays, and radio disturbances. The severe disturbances can disrupt normal radio communications for several days and various occasions of interruption to land wire circuits have also been reported. Various investigators have shown the effect of magnetic disturbances on radio communication to be more pronounced as the wave path approaches the higher latitudes. Ionospheric recordings, both by the fixed-frequency and the multifrequency techniques, provide fundamental information regarding the development and effect of such disturbances on radio communications.

17. AUTOMATIC FREQUENCY AND PHASE CONTROL OF SYNCHRONIZATION IN TELEVISION RECEIVERS

K. R. Wendt and G. L. Fredendall
(RCA Manufacturing Company, Inc., Camden, N. J.)

One of the problems in the reception of television images is to provide satisfactory synchronization in the presence of noise. During the past several years considerable experience has been gained with respect to this problem under various receiving conditions. The system of synchronization which has given satisfactory results up to the present time has depended for its operation on the reception and separation of individual pulses. In general it can be said that with this system satisfactory synchronization can be obtained from those signals which will in all other respects provide an entirely acceptable picture. However, for limiting conditions of service, particularly during early operation where field strength may be low, an improvement in synchronization will be effective and desirable provided that it does not involve other complications or disadvantages.

This paper describes a synchronizing means at the receiver that employs a new principle in the field of synchronization. The principle is automatic frequency and phase control of the saw-tooth scanning voltages. In such a system, synchronization depends on the average of many regularly recurring synchronizing pulses. Noise has insufficient energy at the scanning frequencies to effect control through the direct-current link from which all but relatively long-time variations are filtered out.

Experimental receivers, in which automatic frequency control of the scanning oscillators has been incorporated, have operated with high immunity to noise. The degree of immunity is of a different order of magnitude from that found in conventional synchronizing systems.

Noise cannot affect horizontal resolution or interlacing. An intrinsic property of the new system is perfect interlacing. The return line in an automatic frequency-controlled system may start before synchronization.

Consideration of this new development indicates that its use would result in several improvements in television service: (1) when severe noise conditions occur, superior performance is realizable within the service area; (2) under such noise conditions the useful service area is extended; (3) the maximum resolution permitted by a television channel is attained; (4) it is expected that the cost of television receivers will not be increased by the use of this circuit.

7. A SCANNING ELECTRON MICROSCOPE

V. K. Zworykin, J. Hillier, and R. L. Snyder
(RCA Manufacturing Co., Camden, N. J.)

In order to examine the surface of bulk material with the high resolving power afforded by the use of an electron beam a new electron microscope of the scanning type has been developed in which an extremely fine and stationary electron probe is produced by a two-stage reducing electrostatic-lens system. The specimen is moved mechanically in such a way that each point of its surface is scanned in a systematic fashion by the electron probe. The secondary electrons which are emitted from the point of the specimen bombarded by the electrons of the probe are accelerated and projected on a fluorescent screen. The intensity of the light emitted by the fluorescent screen varies in accordance with the secondary-emission properties of successive points of the specimen. This modulated light signal is converted into an electrical signal by means of a multiplier phototube and then synthesized in a printed picture by an amplifier and facsimile printer system. The use of the electronic-light-electronic transformation of the image signal improves the signal-to-noise ratio by at least an order of magnitude over that found in conventional methods of collection and voltage amplification. An experimental model has been constructed and has been successful in producing images of etched metal surfaces at magnifications as high as 10,000 diameters and with a resolving power considerably better than 50 millimicrons. The blackening of the image points has been found to be a function of the three-dimensional contour of the specimen as well as of its secondary-emission properties.

Membership

The following indicated admissions and transfers of memberships have been approved by the Admissions Committee. Objections to any of these should reach the Institute office by not later than May 29, 1942.

Transfer to Member

Aston, J. P., 1336 Crawford Bridge Ave., Verdun, Que., Canada
Felch, E. P., 44 Elmwood Ave., Chatham, N. J.
Hoffman, E. J., 409 E. Third St., Emporium, Pa.
Hogencamp, H. C., 492 Prospect St., Maplewood, N. J.
Katzin, M., 3538 A St., S.E., Washington, D. C.
LeBel, C. J., 370 Riverside Dr., New York, N. Y.
McCoy, J. C., Bell Telephone Laboratories, Inc., 180 Varick St., New York, N. Y.
McKechnie, J. S., 124 Lincoln Ave., Little Falls, N. J.
Morris, R. M., 22 Mountain View Rd., Millburn, N. J.
Nichols, W. A., Canadian Broadcasting Corporation, 1012 Keefer Bldg., Montreal, Que., Canada
Olmstead, N. C., Bell Telephone Laboratories, Inc., Whippany, N. J.
Schlesinger, L., c/o Naval Research Laboratory, Anacostia, D. C.
Smith, J. P., 39 Harrison Ave., Erlton, N. J.
Somers, R. M., c/o Thomas A. Edison, Inc., Lakeside Ave., West Orange, N. J.
Young, C. H., Bell Telephone Laboratories, Inc., 463 West St., New York, N. Y.

Admission to Member

Anderson, F. B., 188 Elmwood Rd., Oakhurst, N. J.
Black, H. S., Bell Telephone Laboratories, Inc., 180 Varick St., New York, N. Y.
Bollman, J. H., Bell Telephone Laboratories, Inc., 180 Varick St., New York, N. Y.
Bond, W. L., Bell Telephone Laboratories, Inc., 180 Varick St., New York, N. Y.
Dorsey, J. W., University of Manitoba, Winnipeg, Manit., Canada
Edson, J. O., 10 Rustic Pl., Great Kills, S. I., N. Y.
Kinzer, J. P., Bell Telephone Laboratories, Inc., 180 Varick St., New York, N.Y.
Krist, H. K., 100 Lake Dr., Mountain Lakes, N. J.
Maggio, J. B., 548 Springfield Ave., Summit, N. J.
McCurdy, R. G., Bell Telephone Laboratories, Inc., 463 West St., New York, N. Y.
Page, E. C., 504 Munsey Bldg., Washington, D. C.
Stratton, J. A., Massachusetts Institute of Technology, Cambridge, Mass.
Van Tassel, E. K., 419 S. Chestnut St., Westfield, N. J.

The following admissions or transfers (where indicated as such) of memberships were approved by the Board of Directors on April 1, 1942.

Transfer to Member

Chamber, C. C., University of Pensylvania, Philadelphia, Pa.

Admission to Associate

Ahern, W. R., 43 Woodland St., Worcester, Mass. (Transfer)
Alger, P. B., 2519 S. Colorado St., Philadelphia, Pa.
Amman, E. A., 1560 Lansing Ave., Detroit, Mich.
Anderson, G. P., 513 N. James Ave., Minneapolis, Minn. (Transfer)
Arndt, W. R., N. Tenbroeck St., Scotia, N. Y. (Transfer)
Atherton, J. B., 2234 Kamehameha Ave., Honolulu, T. H. (Transfer)
Bailey, V. G., Amenia, N. Y. (Transfer)
Baker, H. F., R.D. 2, Uniontown, Pa. (Transfer)
Bangs, J. R., RCA Manufacturing Company, Inc., Harrison, N. J.
Barbour, C. W., Jr., 830 Beacon St., Boston, Mass. (Transfer)
Baum, E., 24 Lenox Ave., Mount Vernon, N. Y. (Transfer)
Beck, G., 404 College Ave., Ithaca, N. Y.
Belov, F. I., 210 Madison Ave., New York, N. Y.
Benson, S., 706 N. 28th St., Belleville, Ill.
Bersbach, A. J., 9 Clough St., Lynn, Mass. (Transfer)
Blake, L. V., 1527—28 St., S.E., Washington, D. C.
Bomar, L. C., c/o Southern Natural Gas Company, Wetumpka, Ala.
Bradley, J. A., 1705 W. Clinch Ave., Knoxville, Tenn. (Transfer)
Breeze, J. E., 76 Grosvenor Ave., Ottawa, Ont., Canada (Transfer)
Brelsford, H., 947 Maple Ave., Schenectady, N. Y. (Transfer)
Briggs, D. K., 50 Center St., Chatham, N. J.
Brummer, S. H., 3569 DeKalb Ave., New York, N. Y.
Buss, R. R., 70 Kirby Pl., Palo Alto, Calif. (Transfer)
Carter, S. T., Germantown Manor, Greene and Hortter Sts., Philadelphia, Pa.
Cheng, K. Y., 515 W. 124 St., New York, N. Y.

Clark, J. R., Purdue University, West Lafayette, Ind. (Transfer)
Cochran, R. G., 3345 N. Tacoma Ave., Indianapolis, Ind.
Cosgrove, T., 1555 Odell St., New York, N. Y.
Cristina, H. J., 267 Winsor St., Ludlow, Mass. (Transfer)
Crooks, R. K., Union Switch and Signal Company, Swissvale, Pa.
Crossan, G. W., Jr., 3 Du Maurier Blvd., Toronto, Ont., Canada.
Crow, H. E., 420 N. Oliver St., Wichita, Kan. (Transfer)
Cushing, T. D., 94 Gilchrist Ave., Ottawa, Ont., Canada (Transfer)
Davies, H. E., 5211—37, N.E., Seattle, Wash. (Transfer)
Davison, J. M., 54 William St., St. Thomas, Ont., Canada (Transfer)
Dearborn, F. K., Jr., 32 Union St., Portsmouth, N. H.
Dickey, D. W., 39 E. Water St., Middletown, Pa.
Durkee, C. E., 4245 Elston Ave., Chicago, Ill. (Transfer)
Duszak, H., 613½ N. Dearborn St., Indianapolis, Ind. (Transfer)
Ellis, G. M., 929 N. Stanley Ave., Hollywood, Calif.
Ferguson, K., University of Western Ontario, St. Thomas, Ont., Canada (Transfer)
Ferry, T. R., 616—47 St., Oakland, Calif. (Transfer)
Fitzmorris, S. R., 1102 Yeadon Ave., Yeadon, Pa. (Transfer)
Flyr, J. G., Radio Station WCED, Dubois, Pa.
Fox, R. C., 17 Beaumont Ave., Catonsville, Md. (Transfer)
Friedland, M. S., 411 Federal Annex, Atlanta, Ga. (Transfer)
Funk, L., 525 E. Washington, Belleville, Ill.
Garblik, A. M., 1256 E. 13 St., Brooklyn, N. Y.
Garik, S., 171 W. 81 St., New York, N. Y.
Garrat, P. M., 1802 Hambur St., Schenectady, N. Y. (Transfer)
Giacoletto, L. J., Signal Corps Laboratories, Fort Monmouth, Red Bank, N. J. (Transfer)
Gise, F. G., Jr., 4022 Magee St., Philadelphia, Pa.
Glan, G., 3513 W. Jackson Blvd., Chicago, Ill.
Glyptis, N., 3 Clinton St., Cambridge Mass.
Goldstein, G. D., 7516 Harford Rd., Baltimore, Md.
Green, J. A., 411 W. Stadium, West Lafayette, Ind. (Transfer)
Greenaway, L. R., 272 Main St., East Hartford, Conn. (Transfer)
Griffith, B. W., 1619 Commonwealth Ave., Brighton, Mass. (Transfer)
Griffith, D. J., R.R. 1, Box 172, Moline, Ill. (Transfer)
Haddock, F. T., Jr., 614 Bashford Lane, Alexandria, Va.
Hale, C. R., 2333 Pine St., Abilene, Texas (Transfer)
Hamilton, S. H., 208 Maple Pl., Cranford, N. J.
Hansel, P. G., 11 Elm Pl., Red Bank, N. J. (Transfer)
Hansen, H. M., 3240 Briggs Ave., Alameda, Calif. (Transfer)
Harland, R., Box 721, Caldwell, Idaho (Transfer)
Harris, M. S., 614 E. Union, Seattle, Wash.
Harris, W. A., 121 Payton Ave., Haddonfield, N. J. (Transfer)
Hasbrook, A. F., 1815 W. Hildebrand, San Antonio, Texas
Hatchwell, J., 725 W. Roosevelt Blvd., Philadelphia, Pa. (Transfer)
Hegbar, H., 64 Park Ave., Bloomfield, N. J. (Transfer)
Herman, H., 1729 Webster St., N.W., Washington, D. C. (Transfer)
Hill, B. C., Jr., 422 Construction Bldg., Dallas, Texas
Hoffman, R. Y., Jr., 921 Pine St., Winnetka, Ill.
Hogg, J. E., 1625 Ave. B, Schenectady, N. Y. (Transfer)
Hooper, J. R., 621 Union St., Schenectady, N. Y. (Transfer)
Hopper, A. L., 85 Spring Valley Ave., River Edge, N. J.
Hutchins, W. R., 606 W. 113 St., New York, N. Y. (Transfer)
Jackson, J. J., 454 E. Locust, Philadelphia, Pa. (Transfer)
Jacobs, L. F., St. Horbert College, West De Pere, Wis. (Transfer)
Jacobs, O. B., 20 Hamilton Rd., Morristown, N. J.
Johnson, P. V., 1011 W. Indiana Ave., Elkhart, Ind. (Transfer)
Johnston, D. L., 31 Wilmington Way, Brighton 6, Sussex, England (Transfer)
Jones, B. K., 7607 Eastern Ave., Takoma Park, D. C.
Kappelman, G. E., 2234 E. First St., Brooklyn, N. Y. (Transfer)
Keith, E. G., 208 Union St., Schenectady, N. Y. (Transfer)
Kidd, W. E., 355 Duane St., Glen Ellyn, Ill.
Kirkland, R. E., 304 W. 102 St., New York, N. Y.
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Papas, C. H., 4409—18 St., N.W., Washington, D. C. (Transfer)
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Rugg, H. H., 31 Lambton Rd., Ottawa, Ont., Canada (Transfer)
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Sampathu, S., c/o G. Ramaswami, Narsing, Bhuvan, Brahmin, Wada Rd., Matunga, Bombay, India. (Transfer)
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Segal, B., 5 S. Winfield Ave., Bywood, Upper Darby, Pa. (Transfer)
Selig, B. L., 5340 Crown St., Indianapolis, Ind. (Transfer)
Sensiper, S., 111, Locust St., Garden City, L. I., N. Y. (Transfer)
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Strothers, H. C., 6227 La Mirada Ave., Hollywood, Calif. (Transfer)
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Thomforde, C., 11411 Sandy Blvd., Portland, Ore. (Transfer)
Tiffany, W. D., 2755 Franklin St., San Francisco, Calif. (Transfer)
Toward, T., "Little Thatch," Winterbourne, Monkton, Wiltshire, England
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Wahl, J., 1531 O St., N.W., Washington, D. C. (Transfer)
Walker, R. T., Route 2, Box 68, Beaverton, Ore. (Transfer)
Warriner, B., Bendix Radio Corporation, Baltimore, Md. (Transfer)
Welch, H. E., Stockton Junior College, Stockton, Calif.
Wertz, H. E., Bell Telephone Laboratories, Inc., 463 West St., New York, N.Y.
White, G. E., 71 Fairview Blvd., Hempstead, L. I., N. Y. (Transfer)
Wild, J. H., 1224 Ransom St., Sandusky, Ohio.
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Woo, L. N., Mandarin Restaurant, 102 Market St., Newark, N. J.
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Yost, K., Dayton, Ind. (Transfer)

Books

Electric Circuits, by the Electrical Engineering Staff of Massachusetts Institute of Technology

Published by John Wiley and Sons, 440 Fourth Avenue, New York, N. Y. 764 + xxxiii pages + 14-page index. 420 figures. 6¼×9¼ inches. Price, $7.50.

Electric Circuits differs from most books in that it was written by a group rather than one or two individuals. It was the belief of the electrical engineering faculty of the Massachusetts Institute of Techology that the large staff "with its varied interests in teaching and related research could effect a new synthesis of educational material in the field of electrical engineering . . . with a breadth of view not easily approached by an author working individually. It should be free from duplication, repetition, and unbalances so often present in unintegrated textbooks. It should possess a unity and a breadth arising from the organization of the subject as a whole." These are worthy objectives and undoubtedly the plan possesses merit in producing a standardized and well-balanced treatment.

The introduction provides a broad historical perspective and shows how electrical engineering contributes to modern life. It emphasizes that the systems dealt with consist of tangible elements—generators, circuits, electron tubes—and intangible elements—charges and fields. The solution of problems involving these elements may be obtained by the use of circuit theory or field theory. The method to be used in any given case depends upon the nature of the problem. The choice for the beginning student is very well stated near the end of the introduction. "The chapters which immediately follow are concerned with circuit theory, in which an electrical system may with sufficient precision be thought of in terms of combinations of elements characterized by resistance, inductance, and capacitance, together with voltage and current . . . . Circuit theory is therefore a convenient starting point for a study of electrical engineering, because the tangible elements of so many electrical systems may be expressed by means of these five fundamental circuit components and because, inherently, circuit theory is more readily grasped at first than is the basic field theory from which it is derived." In view of this clear statement it is surprising that the first chapter starts out with an attempted correlation of circuit theory and field theory. Since, in general, the student knows little or nothing of either an explanation of one in terms of the other is meaningless and discouraging.

Since the success of the field theory as applied to problems relating to the transmission of radio waves supports the conclusion that Maxwell's equations provide a complete expression for the quantitative relationship of the electric and magnetic fields associated with an arbitrary distribution of charge and current, it would seem that the circuit relations could be derived directly from the field concept. However, this is possible only in special idealized cases where the field equations can be solved. In view of this limitation, a useful and rigorous development of circuit relations from the field equations appears impracticable. There are border-line cases where it would be not only possible but helpful in giving a more complete understanding of the phenomena. It is hoped that in the next edition the discussion of this aspect of the subject will come later in the book and be considerably expanded.

An extensive list of problems covering the broad field of electrical circuits in both transient and permanent states, is included; many of these are worked out in detail for illustrative purposes. The method of symmetrical components as applied to the analytical solution of unbalanced polyphase circuits is included. The table of electrical and mechanical analogies will prove useful in clarifying some relationships, particularly for the student who is interested in sound-conversion units. A section is devoted to the important subject of nonlinear circuit elements making use of graphical and semigraphical solutions as well as analytic approximations. In solving network problems involving linear algebraic equations, the appendix on determinants will be of assistance. In places the notation differs from established standards, and in others appears unwieldly, although this is largely a question of becoming accustomed to it. Fortunately a list of the symbols used is included.

The book contains much of value both from the standpoint of material and method.

H. M. Turner
Yale University
New Haven, Connecticut

Fundamentals of Vacuum Tubes (Second Edition), by Austin V. Eastman

Published by McGraw-Hill Book Company, 330 West 42 Street, New York, N. Y. 569+xxi pages+13-page index. 464 figures. 6¼×9¼ inches. Price, $4.50.

The second edition of this well-known text has been improved by rearrangement and the addition of new material. The book is divided into two parts covering first a description of almost every type of electronic tube having engineering use, and then giving discussions of the basic circuits to be used in applying the tube. The tube section includes some material on the underlying physics and treats gas-filled and photoelectric tubes as well as high-vacuum thermionic devices. The second part, on applications and circuits, contains material on rectifier and control circuits although much more space is devoted to the uses of tubes in amplifiers, oscillators, modulators, and detectors.

The title may be somewhat misleading to some who may expect a detailed treatise on the fundamental physics underlying vacuum tubes. This is decidedly an engineering text. It will be most useful to practicing circuit engineers and students who want to grasp the basic principles of operation of tubes in order to use them effectively in the most important and commonest applications. It will be less useful to vacuum-tube engineers or physicists whose problem is the design or development of the tubes themselves since the discussions of basic physics are quite sketchy and some statements may create erroneous impressions. There is; for example, the statement on page 14 implying that the constant relating the emission current density to $T^2\epsilon^{-b_0/T}$ is a universal constant which theoretically should be equal to 120.4 but proves experimentally to vary widely. Actually, this wide variation is in general accordance with present theory.

The discussion of amplifiers, oscillators, and modulation is especially good, with clear, physical explanations and convenient notation. The author makes good his promise to provide descriptive discussions preceding extensive mathematical analyses and there are many excellent problems. Oscillator circuits are not only described as to their mechanism of operation but, unlike some radio texts, the problem of design is given careful treatment. Frequency modulation is discussed by use of vector diagrams as well as some mathematical analysis and all important methods of obtaining and utilizing properties of amplitude-limited frequency modulation are treated competently and rather extensively considering the relative newness of this application.

The only phase of vacuum-tube work which the author might be criticised for treating too briefly is the ultra-high frequencies. The problems in applying tubes at frequencies where, for instance, transmission lines take over the jobs from lumped circuits are in many cases unique. They have been discussed extensively in the literature and a good deal of this material needs now to be assembled in a text for the same convenient study and reference which the author makes possible for the ordinary radio frequencies.

Simon Ramo
General Electric Company
Schenectady, New York

Tables of Integrals and Other Mathematical Data, by H. B. Dwight

Published by The Macmillan Company, 60 Fifth Avenue, New York, N. Y. 1934. 220+viii pages+2-page index. $5\frac{3}{4}\times 8\frac{3}{4}$ inches. Price, $1.75.

This book has, over a period of several years, proved to be of very great value in all fields of engineering requiring mathematical analysis. It contains tables of integrals and of infinite-series expansions that are sufficiently complete for almost all ordinary engineering and scientific needs. The various items are adequately cross-referenced.

Integrals and series expansions, also the more common identities, are included for rational and irrational algebraic functions, trigonometric, exponential, logarithmic, hyperbolic, inverse trigonometric, inverse hyperbolic, and elliptic functions. Integrals for Bessel functions, also for zonal harmonics, are included. There is also a table of definite integrals and a brief but very useful outline of the more important methods used in solving differential equations.

Numerical tables included comprise Bessel functions, trigonometric tables, gamma functions, a four-place table of logarithms to the base 10, a five-place table of natural logarithms, a table of values of exponential and hyperbolic functions, and tables of values of elliptic functions and Bessel functions.

Unfortunately there is no table of error-function values. The infinite series expansions, both direct and asymptotic, for the error-function integral are, however, included. The book would be appreciably more useful if it had included additional integrals, both indefinite and definite, of the error-function type.

On the whole, this book has over a period of years proved itself to be, in the reviewer's opinion, by far the best general-purpose table of integral and mathematical data available for engineering and scientific use at the present time.

W. G. Dow
University of Michigan
Ann Arbor, Michigan

Radio Reception in Theory and Practice, by Virendra Kumar Saksena

Published by Industry PublishersLtd., Keshub Bhaban, 22, R. G., Kar Road, Shamhazar, Calcutta, India. 210+v pages +5-page index. 104 figures. $5\times 7\frac{1}{4}$ inches.

This is a very satisfactory introductory book. It is well written and covers most of the elementary subject material needed by beginners. The author received much of his technical education in England.

H. M. Turner
Yale University
New Haven. Connecticut

Contributors

F. A. Everest

F. Alton Everest (A'38) was born on November 21, 1909, at Gaston, Oregon. He received the B.S. degree in electrical engineering from Oregon State College in 1932 and after two years of graduate work received the E.E. degree from Stanford University in 1936. For six months he worked on television development for the Don Lee Broadcasting System after which he became an instructor in electrical engineering at Oregon State college and in 1939, assistant professor. Mr. Everest is a member of Eta Kappa Nu and Sigma Xi and is an associate member of the American Institute of Electrical Engineers.

❖

Dudley E. Foster (A'26–M'37) was born at Newark, New Jersey, on December 12, 1900. He was grauated from Cornell University in 1922 with the E.E. degree. Following his graduation he became associated with the Electrical Alloy Company and Driver-Harris Company. In 1925 he joined the Malone-Lemmon Products Company and the next year became chief engineer of the Case Electric Company. Two years later that company was merged with the United States Radio and Television Company and soon thereafter Mr. Foster was promoted to chief engineer. In 1933 he became chief radio engineer of the General Household Utilities Company and from 1934 to 1941 he was division engineer in the RCA License Laboratory. He was also a lecturer in television engineering in the Graduate School of Stevens Institute of Technology. At the present time Mr. Foster is vice president and technical director of Rogers Majestic (1941), Limited, Toronto, Ont., Canada.

Dudley E. Foster

❖

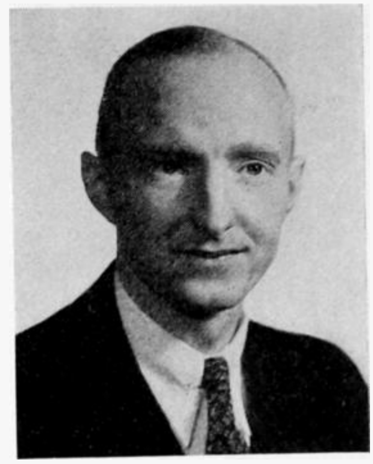
Charles W. Harrison, Jr. (A'36) was born in Virginia in 1913. In 1931 he was enlisted in the U. S. Navy as a candidate for the Naval Academy. During this year he was graduated from the Radio Operators School, and served as a radio operator aboard a vessel of the Pacific Fleet. The years 1932 to 1936 were spent in the U. S. Naval Academy Preparatory School, and in the Academy. In 1939 he received the B.S. degree in engineering, and in 1940 the degree of electrical engineer from the University of Virginia. While in the University he was an operator at WCHV. Mr. Harrison is licensed to practice both electrical and radio engineering in Virginia.

C. W. Harrison, Jr.

The summers of 1939 and 1940, together with the entire year of 1941, were spent in research work relative to antennas and transmission lines at the Navy Department in Washington. For the last two years Mr. Harrison has held scholarships covering tuition at the Cruft Laboratory, Harvard University. He is at present a student officer on duty at the Cruft Laboratory, where he is a candidate for the S.M. and Sc.D. degrees.

❖

William G. Hutton (A'38) was born on March 9, 1904, at Batavia, Iowa. He received his A.B. degree in 1928 at Parsons College and his M.S. degree in physics in 1932 at Iowa State College. Mr. Hutton was with the Western Electric Company in 1929 and 1930 and was professor of physics at William Penn College from 1932 to 1935. He has been a member of the engineering staff of WGAR since 1937.



William G. Hutton

❖

Paul A. Loyet (A'37–M'38) was born at Ottumwa, Iowa, on January 10, 1906. He received the B.E. degree in electrical engineering from the State University of Iowa in 1927. Mr. Loyet became interested in amateur work in 1919 and entered the commercial broadcasting business in 1924 with WOC as an operator. After graduation, he became chief engineer of WOC until February 15, 1930, when the Central Broadcasting Company was formed, incorporating WOC, Davenport, and WHO, Des Moines, Iowa. He has been the chief engineer and technical director from that time to date; also a member of the board of directors of the Central Broadcasting Company.

❖

Garrard Mountjoy (A'30–M'40) received the B.S. degree from Washington

Paul A. Loyet

Garrard Mountjoy

University in 1929. That same year he joined the Sparks-Withington Company as a development engineer, spending some time in Europe for that company designing radio sets for local conditions. He later became chief engineer of the Sparks-Withington Radio Development Department which position he held until 1935. Prior to entering college he spent a year with

R. Morris Pierce

the Westinghouse Company. Since 1935, Mr. Mountjoy has been on the staff of the RCA License Laboratory. He received a "modern pioneer" award from the National Association of Manufacturers in 1940. He is a member of Sigma Xi.

❖

R. Morris Pierce (M'42) was born December 17, 1906, at Chicago, Illinois. He attended Cornell College at Mt. Vernon, Iowa, 1924–1925, and Case School of Applied Science, Cleveland, Ohio, 1930–1932. He was employed in the engineering and tests departments of the Zenith Radio Corporation, 1927–1928; chief engineer, Cleveland Police Radio System, 1929; chief engineer, WWVA, Wheeling, West Virginia, 1930; chief engineer WGAR, Cleveland, Ohio, 1930 to date.

❖

Wilson S. Pritchett (S'41) was born at Vale, Oregon, on December 8, 1910. He received the B.S. degree in 1940 and the M.S. degree in 1942 from Oregon State College where he has been a graduate assistant in electrical engineering. He is a member of Eta Kappa Nu, Sigma Pi Sigma, and an associate of Sigma Xi.

❖

For a biographical sketch of Karl Spangenberg, see the Proceedings for March, 1942.

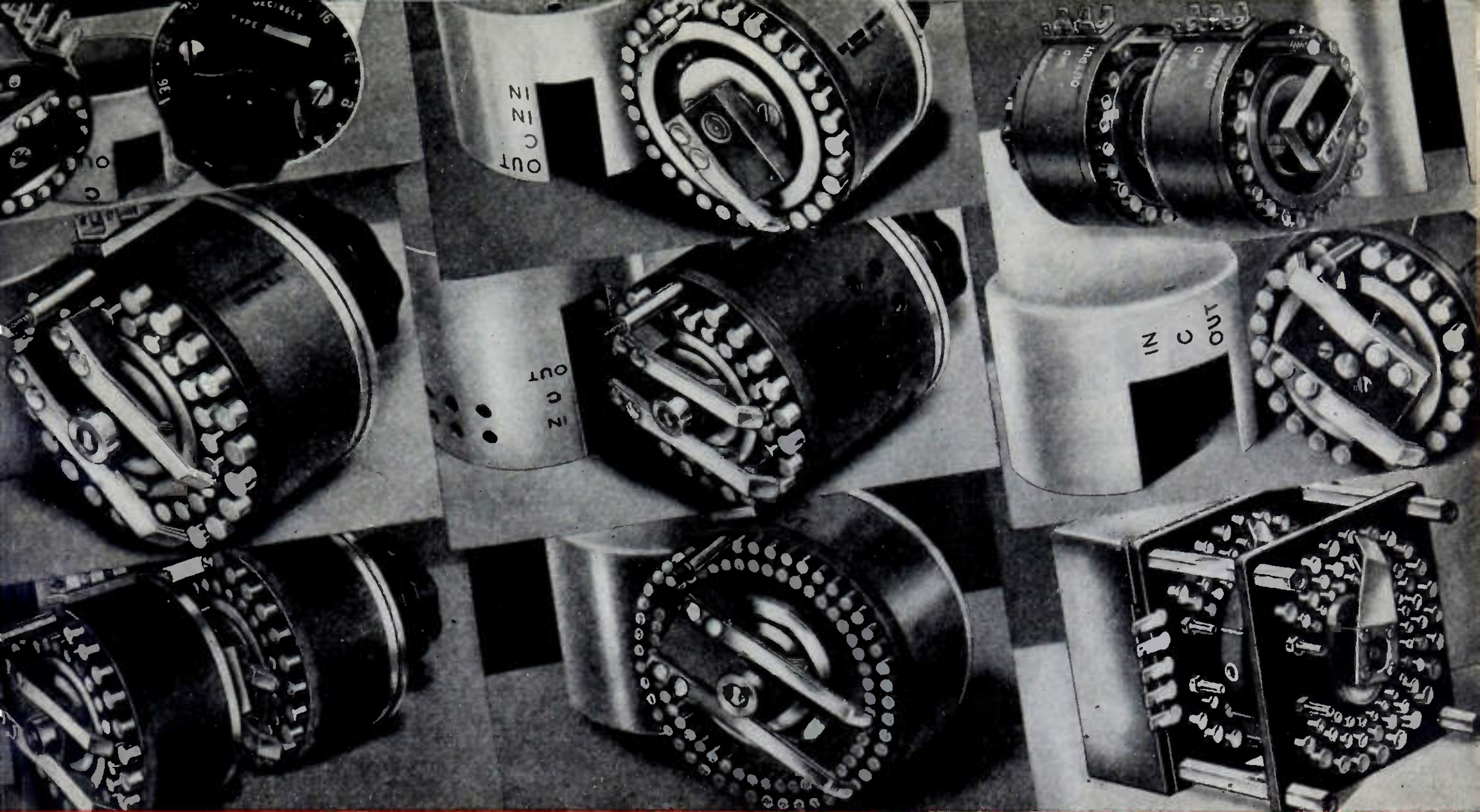
❖

Russell A. Whiteman (A'41) received his B.S. degree in 1933 and E.E. degree in 1934 from Columbia University. He was awarded the Trowbridge Fellowship to continue graduate work at Columbia and was simultaneously employed by American Machine and Metals, Inc., as research engineer on accoustical measurements, in which he continued until 1936. Prior to that, he was engaged in the Bell Telephone Laboratories from 1927 to 1929, on the development of permalloy and perminvar dust cores applied to lumped loading. In

Wilson S. Pritchett

1936 he went with Carrier Corporation as development engineer on acoustical design. In 1937 he joined RCA Institutes specializing in radio receiving and transmitting circuits and was transferred to Chicago as chief instructor. Mr. Whiteman is a member of Epsilon Chi and Sigma Xi.

R. A. Whiteman



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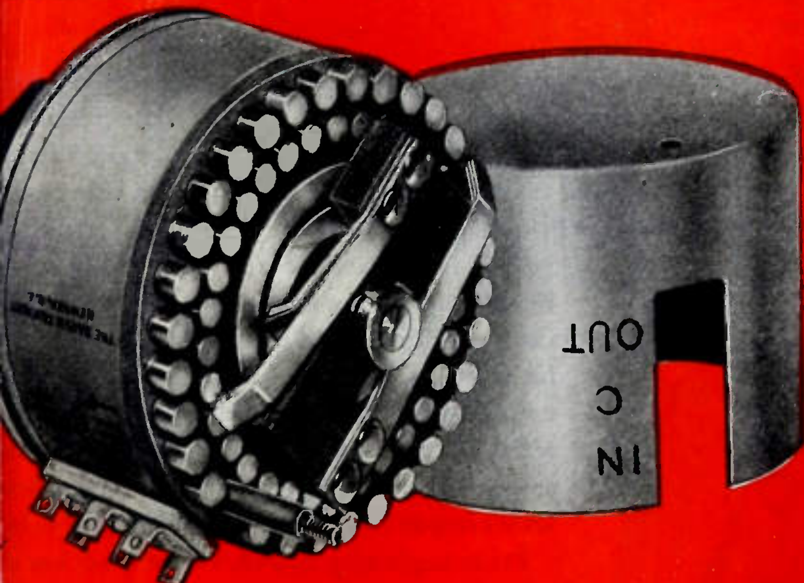

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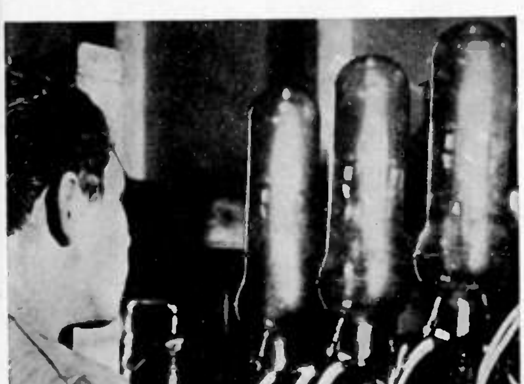

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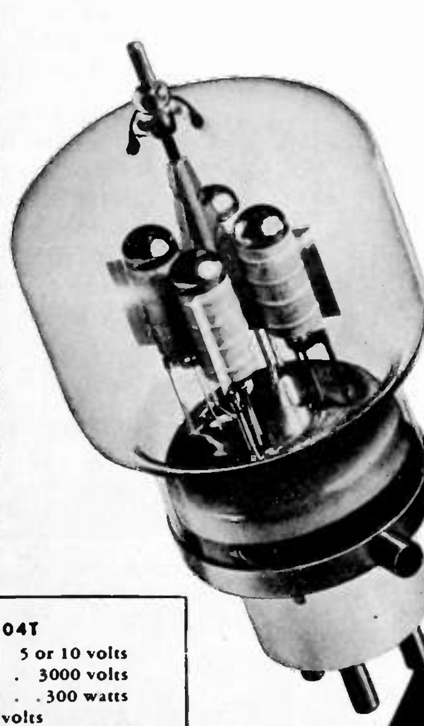

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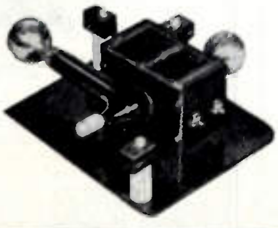

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Current Literature

New books of interest to engineers in radio and allied fields—from the publishers' announcements.

A copy of each book marked with an asterisk (*) has been submitted to the Editors for possible review in a future issue of the Proceedings of the I.R.E.

* ACOUSTIC DESIGN CHARTS. BY FRANK MASSA, in Charge of Acoustic Division, The Brush Development Company. Philadelphia: The Blakiston Company, April, 1942. 228 pages, illustrated, 6×9 inches, cloth. $4.00.

* AUTOMATIC RECORD CHANGERS AND RECORDERS. BY JOHN F. RIDER. New York: John F. Rider Publisher, Inc., 1941. 744 pages, illustrated, 8½×11 inches, cloth. $6.00.

* DICTIONARY OF RADIO AND TELEVISION TERMS. BY RALPH STRANGER. Brooklyn: Chemical Publishing Company, Inc., 1941. 252 pages, illustrated, 5½×7½ inches, cloth. $2.50.

* ELECTROMECHANICAL TRANSDUCERS AND WAVE FILTERS. BY WARREN P. MASON, Member of Technical Staff, Bell Telephone Laboratories, Inc., New York: D. Van Nostrand Company, Inc., 1942. 329+3 index pages, illustrated, 6×9 inches, cloth. $5.00.

* HANDBOOK OF TECHNICAL INSTRUCTION FOR WIRELESS TELEGRAPHISTS (Seventh Edition). BY H. M. DOWSETT and L. E. Q. WALKER. London: Iliffe & Sons, Ltd., 1942. 658+6 index pages, illustrated, 5½×8½ inches, cloth. 25 shillings.

* MATHEMATICS FOR ELECTRICIANS AND RADIOMEN. BY NELSON M. COOKE, Chief Electrician, U. S. Navy. New York: McGraw Hill Book Company, 1942. viii +604 pages, illustrated, 6×9 inches, cloth. $4.00.

* OUTLINE OF AIR TRANSPORT PRACTICE. BY A. E. BLOMQUIST, Transportation Engineer, Airport Engineer, Eastern Air Lines, Inc. New York: Pitman Publishing Corporation, 1941. 412 pages, illustrated, 6×9 inches cloth. $4.50.

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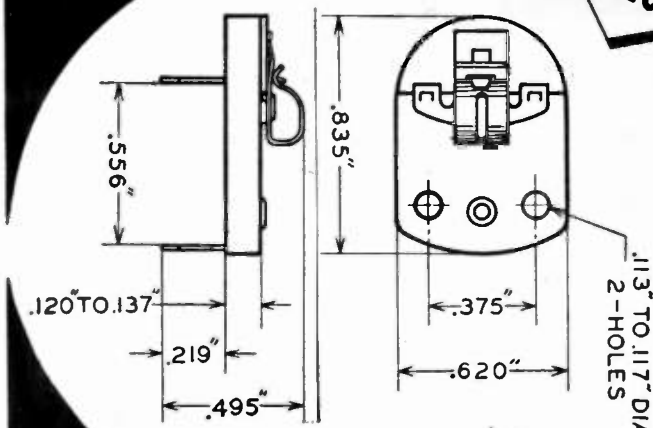

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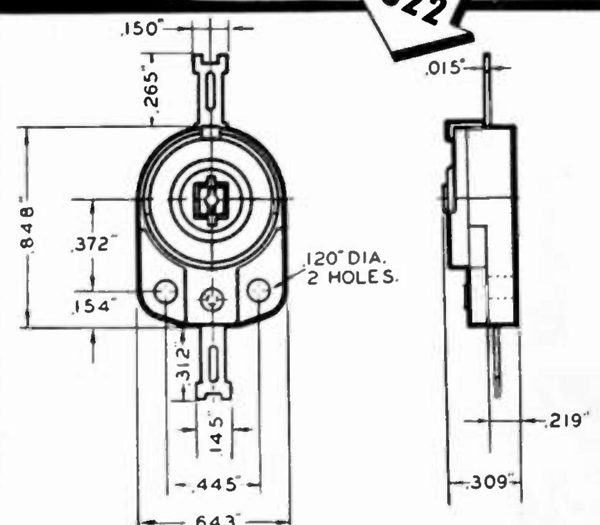

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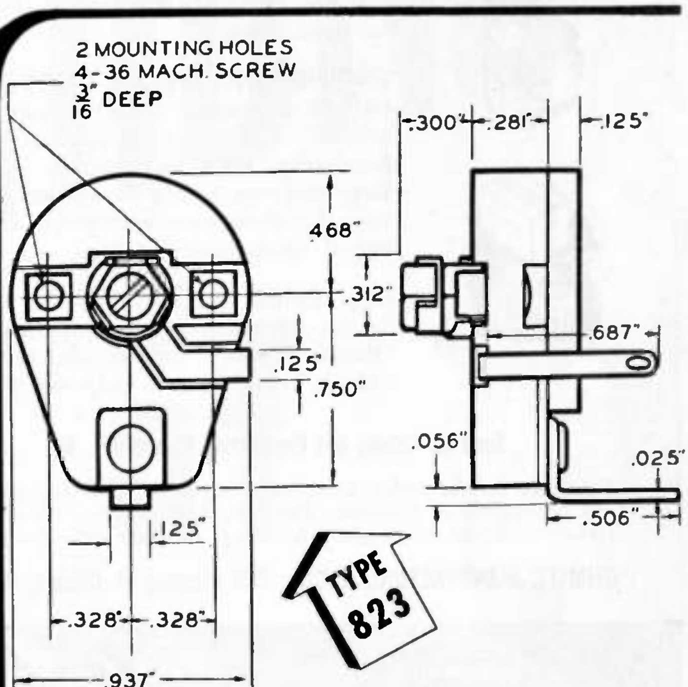

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(cp) < .75 MMF to > 2 MMF

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(Continued from page xiv)

• RADIO HANDBOOK (Eighth Edition). By the Editors of "Radio." Santa Barbara: Editors and Engineers, Ltd., 1941. 624+16 index pages, illustrated, 6½×9½ inches, cloth. $1.75 in Continental U.S.A.; $2.00 elsewhere.

• RADIOTRON DESIGNER'S HANDBOOK (Third Edition). Edited by F. LANGFORD SMITH. Amalgamated Wireless Valve Company, Australia. Harrison: Commercial Engineering Section, RCA Manufacturing Company, Inc., 1941. iv+344+8 index pages, illustrated, 5½×7½ inches, cloth. $1.00.

• STANDARD FOR WET TESTS (No. 29). New York: The American Institute of Electrical Engineers, November, 1941. 8 pages, illustrated, 7¾×10½ inches, paper. 40 cents.

• THE RADIO AMATEUR'S HANDBOOK: Special Defense Edition. BY HEADQUARTERS STAFF of American Radio Relay League, 1942. West Hartford: American Radio Relay League. 280+8 index pages, illustrated, 6½×9½ inches, paper. $1.00.

• BASIC RADIO: The Essentials of Electron Tubes and Their Circuits. BY J. BARTON HOAG, Head of the Department of Science, U. S. Coast Guard Academy. New York: D. Van Nostrand Company, Inc., 1942. 358+21 index pages, illustrated, 5½×8½ inches, cloth. $3.25.

• THE MATHEMATICS OF WIRELESS. BY RALPH STRANGER. Brooklyn: Chemical Publishing Company, Inc., 1942. 210+5 index pages, illustrated, 5½×7½ inches, cloth. $3.00.

• THE SUPERHET MANUAL. EDITED BY F. J. CAMM. Brooklyn: Chemical Publishing Company, Inc., 1942. 131+3 index pages, illustrated, 5½×7½ inches, cloth. $2.50.

• AN INTRODUCTION TO THE OPERATIONAL CALCULUS. BY WALTER J. SEELEY, Professor of Electrical Engineering, Duke University. Scranton: International Textbook Company, 1941. xi+161+3 index pages, illustrated, 5¼×8¼ inches, cloth. $2.00.

• CAREER IN ENGINEERING. BY LOWELL O. STEWART, Head of Department of Civil Engineering, Iowa State College. Ames: Iowa State College Press, October, 1941. 92 pages, illustrated, 6×9 inches, paper. 75 cents.

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