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In This Issue...

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Editorial . . .

EQUIPMENT RELIABILITY

by John E. Remich
Manager, Technical Department

As electronics assumes a more critical position in modern warfare techniques, the demands upon electronic equipment become more and more exacting. In accomplishing new functions, the equipment is required to perform a multitude of complex operations with greater accuracy and reliability than ever before. In the case of aircraft, an equipment failure today not only can nullify the entire purpose of a military mission but also can place in jeopardy the lives of the crew members. It is clear, therefore, that the highest possible degree of reliability is imperative.

Many electronic equipments have already attained a satisfactory degree of reliability—others have not. Electronic equipment designed to provide complex functions is, by necessity, complex to manufacture and maintain. Unreliability may be caused by such things as a defective component or faulty design, or by inability of maintenance personnel to cope with the equipment complexities.

In newly designed equipment, complete reliability cannot be "built in" by the manufacturer. Certain design defects and sources of trouble will be revealed only by use of the equipment in the field under actual operating conditions. Much progress is being made in simulated environmental testing, but only accurate and complete reports on all failures will enable the manufacturer to make the modifications and production changes required to achieve truly satisfactory performance. The field engineer, through his reports and recommendations, holds a key position in the "feedback" network that leads to satisfactory reliability. It is a responsibility not to be taken lightly.

COLLEGIATE USE OF THE PHILCO TRAINING MATERIALS

by Harvey Mertz
Headquarters Technical Staff

(Editor's Note: The Philco Training Materials, including the Lecture Demonstrator and the Laboratory Chassis, have long been used in courses of instruction for training electronics technicians in technical schools and in the armed forces. In this application they have been eminently successful. The same features which contributed to their success in these applications are very useful in collegiate courses in electronics. This article was written to show how the Philco Training Materials will aid Collegiate Training.)

THE COLLEGIATE TRAINING PROBLEM

THE WIDE DIVERSIFICATION of the electronics field is currently placing great demands upon those responsible for the professional training of our electronics engineers. It is becoming increasingly difficult to impart a sound knowledge of the fundamentals of the electronics art in a reasonable length of time. To meet this problem, most educators have been forced to be more selective in the material chosen for the curricula in electrical engineering. The desirability, therefore, of increasing the efficiency of both the teaching and learning processes so as to permit more material to be covered is quite obvious.

Like the electronics technician, the electronics engineer must have a good grasp of fundamental circuit concepts, but for greatly different reasons. The technician works with circuits which have already been designed and built, and his interest is not in design but in the proper operation of such circuits. The problem of the technician is to find the malfunctioning component or components, and to be able to repair or replace such items in the shortest possible time. He should know how to use prepared tables of information and schematic diagrams. The engineer, on the other hand, must concern himself primarily with the problems of design. His job is to design a circuit or circuits which will perform given functions and meet a given set of specifications. However, the details of circuit construction and operation should not obscure the place of the circuit as a "building block" tool of the engineer. Since each circuit of the Philco Training Materials is such a "building block," the engineering student may quickly see the circuit function in its proper perspective when several of the circuits are combined to perform a given job.* The Philco approach to the training problem is to break down complex electronic equipments into a number of basic circuits. Each of these basic circuits is constructed in two equivalent forms: (1) the individual Lecture Demonstration Panel, a number of which are mounted on the Lecture Demonstration Console shown in figure 1, and (2) the individual Circuit Analysis Chassis, a number of which are mounted on the Laboratory Rack shown in figure 3. This construction of com-

*At the present time, many Colleges, Universities, and Technical Institutes are using the Philco Training Materials. Some of these are the United States Military Academy, the United States Naval Academy, the United States Coast Guard Academy, Virginia Polytechnical Institute, and Oklahoma Institute of Technology of Oklahoma A. and M.

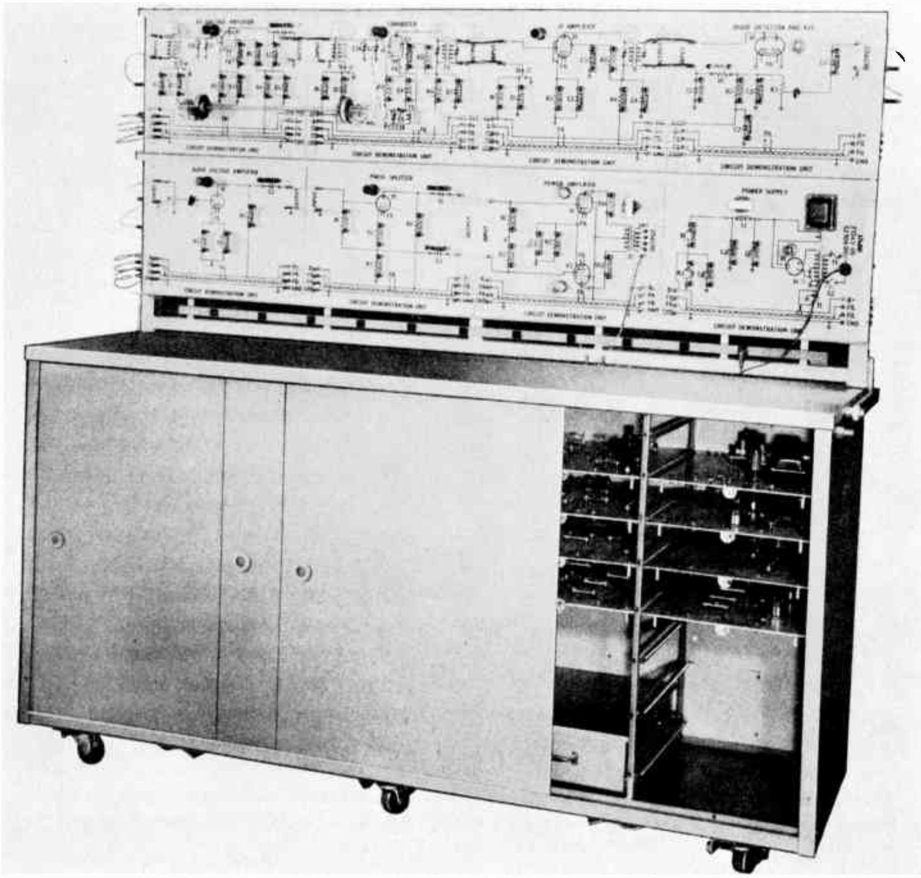

Figure 1. The Philco Lecture Demonstration Console Complete with Circuit Demonstration Panels Which It is Designed to Store and Display

plex circuits from fundamental "building blocks" is a natural way of stimulating imaginative thinking in circuit design during undergraduate studies.

Electronics undergraduate students ordinarily spend considerable time in the laboratory fabricating the circuits with which they experiment. An engineer must, of course, know a good deal about circuit fabrication, but in some cases there is doubtless a point of diminishing returns in terms of knowledge gained versus time expended. Too many key points of circuit operation may become lost in the myriad details of wiring the equipment. Use of the Philco Training Materials will allow the student engineer to become familiar with the physical circuitry without unduly preoccupying him with its construction (which will come in due course). These remarks are especially applicable in view of recent advances in circuit fabrication.

The outstanding advantage of the Philco Training Materials is the great variety of circuits which may be quickly set up and analyzed. The flexibility inherent in the construction of the Training Materials may substantially decrease the time required to construct circuitry for instructional purposes. The effect of variation of circuit design parameters upon the operation of the circuit may be experimentally demonstrated with great ease by the use of the Philco Training Materials. It has been said that

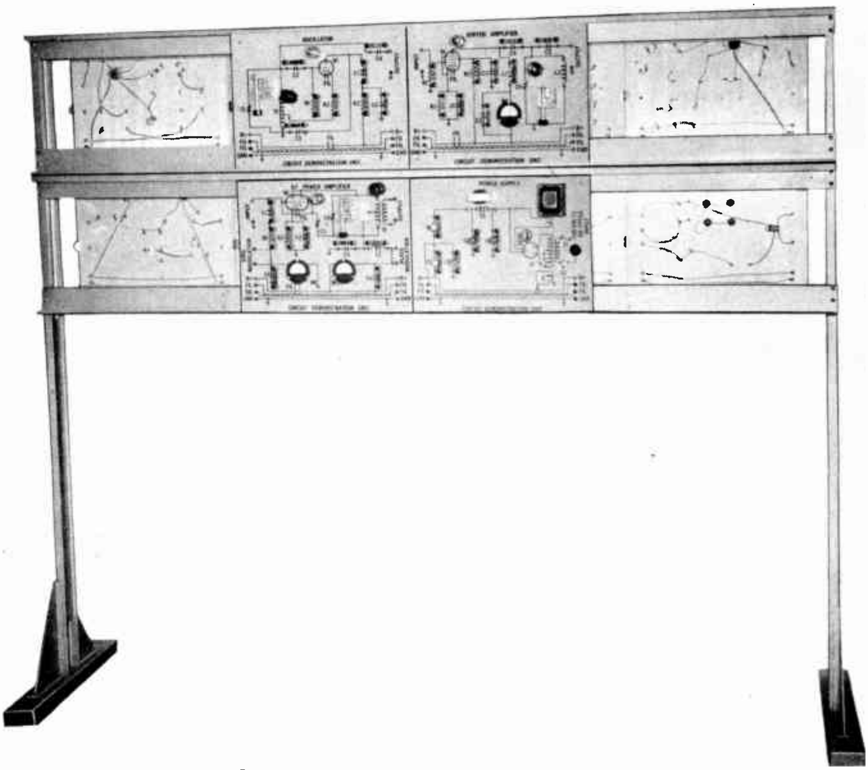

Figure 2. The Philco Lecture Demonstrator Rack with Transmitter Circuit Demonstration Panels Displayed. Notice the Backs of the Circuit Demonstration Panels Which Are Displayed on the Other Side of the Rack.

"repetition is the essence of learning." Regardless of whether or not this is completely true, the duplication of the circuit on the lecture demonstrator and on the laboratory chassis (refer to figure 4) allows the student to apply directly the knowledge gained from lectures employing the lecture demonstrator. This feature makes the transition from schematic diagram to construction of the actual circuit considerably less difficult.

By means of a variation of circuit parameters, circuit application can be readily studied. The completeness of each of the circuit panels emphasizes the fact that comparatively few basic circuits need be used to construct more complex circuits for a variety of tasks.

Circuit design frequently consists of making preliminary calculations concerning a particular circuit and then constructing a "breadboard" of the circuit and actually testing its performance. The measured performance is then compared with the calculated performance, and the discrepancies explained or investigated. Instruction in this design approach is greatly facilitated by the Philco Training Materials without an unreasonable amount of time being spent in soldering wires and components into place. In the completed circuit, any point in the circuit is readily accessible for test measurements as well as for the evaluation or investigation of the circuit operation. The characteristics of real circuit components are introduced, as they must be, at an early stage in the course of study.

APPLICATION OF THE PHILCO TRAINING MATERIALS

The following paragraphs of this ar-

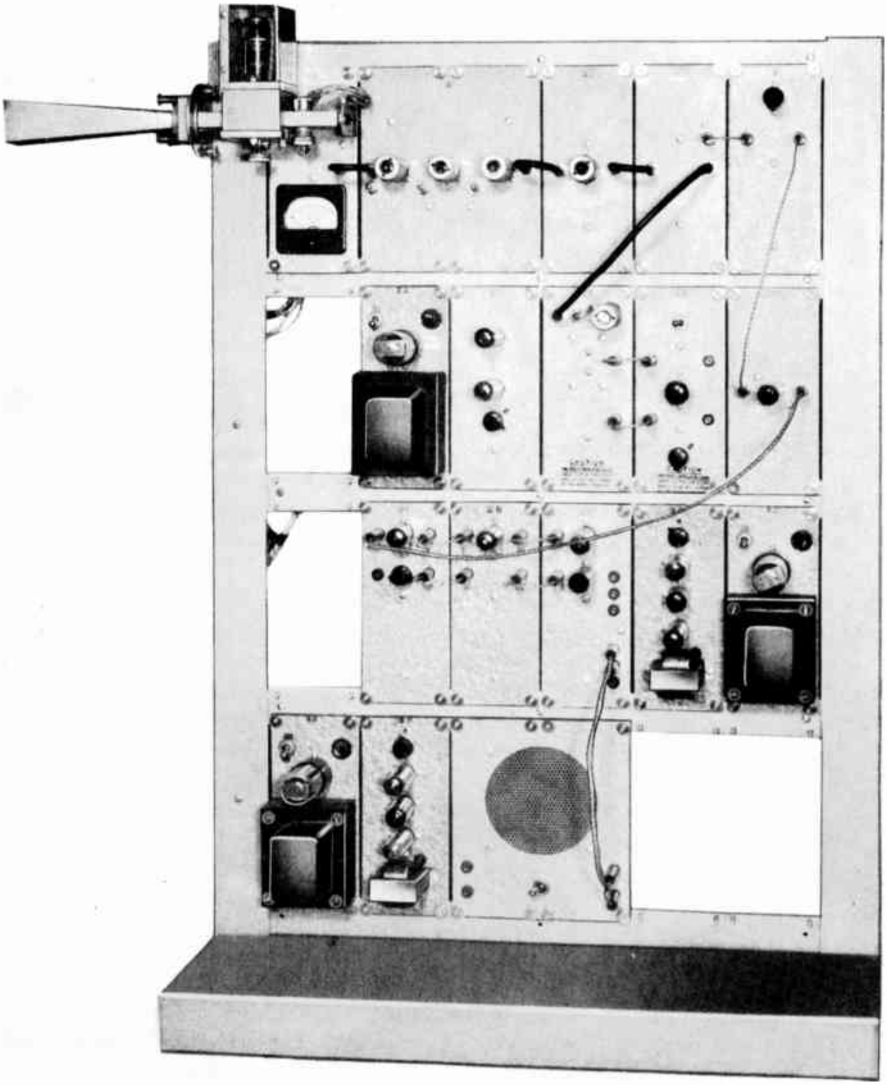

Figure 3. The Philco Laboratory Circuit Analysis Chassis Rack With a Number of Circuit Analysis Chassis Displayed

ticle show how a typical training panel can be used to aid both the student engineer in his studies and the professor in his teaching. As an illustration let us take the basic audio amplifier, a comparatively simple circuit, and consider the approach of the engineer who must design the circuit. For this particular demonstration, Philco Training Materials Audio Amplifier Chassis 27 (shown in figure 4) will be used.

The Design Problem

Assume that the hypothetical training exercise is to design an audio amplifier having a bandpass characteristic flat to within $\pm \frac{1}{2}$ db between the limits of 100 c.p.s. and 5000 c.p.s. The voltage gain must be 50 and the circuit capable of reproducing an input signal up to 6 volts peak-to-peak with no appreciable distortion.

NOTE: The audio amplifier chassis provides further limitations which would be found in designing a circuit as part of a complete equipment. For instance, the choice of

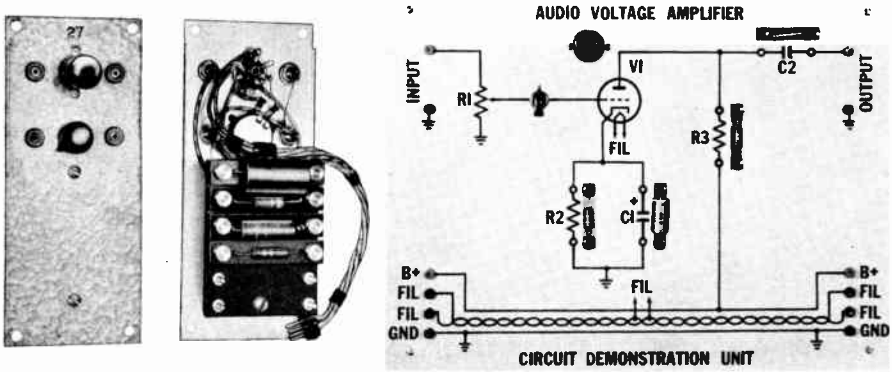

Figure 4. Philco Audio Amplifier Laboratory Circuit Analysis Chassis Number 27 (Front and Rear Views) at Left, and Audio Amplifier Lecture Demonstration Panel at Right

B+ voltage and tube type may be limited by other design considerations.

The Approach

The following material could be used to form the basis of the lecture portion of the hypothetical design problem.

1. A drawing of the schematic diagram of a basic audio amplifier circuit.
2. A discussion of the practical considerations necessary in the choice of components.
3. A review of the derivation of required formulas and their application to the circuit under consideration.
4. Computation of the results of the application of these formulas.
5. Computation of the results for different values of some of the components, and discussion of the reasons for these results.

The Philco Lecture Demonstration Panels can be used to advantage in the lecture for graphically and effectively demonstrating the results of the computations.

The laboratory work for the hypothetical design problem can be given as follows:

1. The student can be assigned the task of designing an audio amplifier which will fit the given specifications, using the formulas which were derived in the lecture class.
2. The design can then be built and tested.
3. A series of tests to demonstrate the results can be run, to determine whether the circuit falls within the desired specifications.
4. Changes can be made in various components and the results of these changes plotted and compared with the original results. Any changes can be discussed in detail in terms of the pertinent theoretical formulas.

The Philco Laboratory Chassis No. 27—Audio Amplifier—can be used to advantage in the laboratory because of the accessibility of test points and the ease with which components may be changed. These features eliminate much of the work required in the building and subsequent modification of the circuit under consideration. Building and modification by conventional means ordinarily require considerable time, which is a complete loss insofar as increasing the understanding of circuit theory is concerned. Also, test points for connecting measuring instruments are often inaccessible in conventionally constructed circuits.

Practical Considerations

1. The tube type. Characteristic curves of tubes should be reviewed and a suitable tube type selected. The lecture must also take into account

the available B+ voltage and the required amplification and input signal level.

2. The plate load resistor. The value of this component depends upon several factors such as current capacity and amplification factor of the tube, and available B+. Several static load lines for various values of resistors are plotted on the characteristic curves; these should show the necessary magnitude of the plate load resistor.
3. Since self bias by means of a bypassed cathode resistor will be used, the approximate value of the cathode resistor can be determined from the bias voltage required and the static value of the plate current by application of Ohm's law. At this point in the design development, a demonstration of the effect of a high-amplitude input signal on distortion can be shown.
4. The coupling capacitor, the grid resistance of the following stage, and the input capacitance of the following stage can be discussed in general, and an idea of the magnitude of these components can be obtained.

Review of Derivation of Applicable Formulas

1. The equivalent circuits for low, medium, and high frequencies are drawn, and derivations for stage gain and phase shift at these frequencies are presented.
2. The formulas are applied to the audio amplifier, keeping practical considerations in mind, and the components are chosen to provide the desired gain and bandwidth characteristics.
3. The gain, bandwidth characteristics, and phase-shift changes with changes in various components should be shown. The calculation and plotting of these changes will be a part of the laboratory work.

Approximation of Plate Load and Cathode Resistor Values

1. The tube must have 3 volts bias to ensure no distortion by grid current flow.
2. With a 6-volt signal applied, the current and voltage data for various values of the plate load resistor are given in Table 1.
3. From these figures it is seen that in order to give the desired gain, the value of the plate load resistor must be 100k.
4. For each plate load resistor listed, the cathode resistor necessary to give the quiescent bias of 3 volts must be as shown in Table 2.

TABLE 2

PLATE LOAD RESISTOR (R_L)	CATHODE RESISTOR (R_k)
100k	Approx. 1.9k
50k	Approx. 1.2k
25k	Approx. 550 ohms

5. Since the plate load resistor must be on the order of 100k, the cathode resistor must therefore be on the order of 2k.

The above calculations assume a B+ voltage of 340 volts, an input signal of no more than 6 volts peak-to-peak, and an adequate bypass capacitor in the cathode circuit. Any changes from these figures will make the calculations inaccurate.

TABLE 1

R_L	QUIESCENT E_p	QUIESCENT I_p	E_p LIMITS	I_p LIMITS
100k	145v	1.6 ma.	45 to 302v	0.3 ma. to 3 ma.
50k	165v	2.5 ma.	70 to 310v	0.6 ma. to 5.2 ma.
25k	238v	5.4 ma.	105 to 320v	0.8 ma. to 9.2 ma.

Discussion of Coupling Capacitor and Grid Resistor Sizes

The maximum frequency is 5000 c.p.s. (period 200 μsec.), and the minimum frequency is 100 c.p.s. (period 10,000 μsec). Therefore, the time constant given by $C_c \cdot R_g$ must be greater than 25,000 μsec. or C_c may charge appreciably during one alternation. The reactance of C_c at the required frequencies must be small compared to the resistance of R_g, while R_g is limited to comparatively large values by the fact that, insofar as an a-c signal is concerned, it is in parallel with the plate load resistance of the preceding circuit.

CONCLUSIONS

In most instances the circuits employed in the Philco Training Materials are comparable with those used in commercial electronic equipment. This fact makes the study of a particular circuit considerably more practical in nature than might otherwise be the case. If it is desired to test certain components under typical operating conditions, the circuit of the Philco Training Materials provides a convenient means of doing so. The rugged construction of the laboratory chassis makes them ideal for repeated use and handling by various types of students. Moreover, parts which must be replaced for any reason can be inserted quickly and easily.

As stated previously, adequate training of electronics engineers is becoming increasingly difficult because of the ever-increasing scope of the electronics field. Therefore, maximum use must be made of available time to ensure that the undergraduates' training is thorough, up-to-date, and practical. Proper application of the Philco Training Materials, which are especially designed for instructional purposes, will go a long way in accomplishing this important goal.

Inquiries concerning the Philco Training Materials should be addressed to Philco TechRep Division, Sales Department, 22nd St. and Lehigh Avenue, Philadelphia 32, Pennsylvania.

"What's Your Answer?"

This issue's problem, which is contributed by John A. Meehan, Jr., Headquarters Technical Staff, concerns the distribution of power in a pulse-amplitude-modulated wave.

It is a well-established fact that any modulated wave consists of a carrier frequency and a group of sideband frequencies. Assume that a radio-frequency oscillator is pulse-amplitude-modulated by a perfectly rectangular pulse whose amplitude is 10,000 volts, whose duration is 1 microsecond, and which recurs at a rate of 1000 pulses per second. If the measured peak power output of the modulated oscillator is 100 kilowatts, what is the average power output at the radio frequency of the oscillator, and how much of the output power is radiated at the sideband frequencies?

D-C ANALOG COMPUTERS

PART 2

by Allen J. MacKinnon
Philco TechRep Field Engineer

This is the second of a series of articles concerning d-c analog computers. The first article of the series appeared in the January-February issue of the BULLETIN, and dealt with the general concepts of d-c analog computers and with the summing circuits employed in these computers.

MULTIPLICATION CIRCUIT

MULTIPLICATION IS ONE of the mathematical operations which must be performed in a computer. The simplest type of electronic multiplication employs a potentiometer as the basic element, as shown in figure 1. When one end of the potentiometer is grounded, the open-circuit voltage appearing between the arm of the potentiometer and ground will be the product of the applied voltage (E_{in}) and R_1/R, where R_1 is the resistance between the arm and ground as shown in figure 1. If the input voltage is 10 volts and the arm is set at the midpoint of the resistance element, the open-circuit output voltage (E_o) is ½ of 10, or 5 volts.

The primary disadvantages of a potentiometer multiplier are that it cannot multiply a quantity by a factor greater than one, and that loading will change the output voltage from the desired value.

The inherent disadvantages of the potentiometer multiplier are largely overcome by the use of a computing amplifier such as that illustrated in figure 2. It was shown in Part 1 of this series that the output voltage (E_o) of such an amplifier is related to the input voltage (E_{in}), the input resistance (R_{in}), and the feedback resistance (R_{fb}) by the equation

$$\frac{-E_o}{E_{in}} = \frac{R_{fb}}{R_{in}},$$

or

$$E_o = -E_{in}\frac{R_{fb}}{R_{in}} \qquad (2\text{-}1)$$

From the latter equation, it can be seen that the input voltage is multiplied by the ratio of R_{fb} to R_{in}, and that the sign of the product is reversed. Multiplication by a factor either greater or less than one is thus possible.

DIVISION CIRCUIT

Division is equivalent to multiplication by the reciprocal of the divisor. In view of this fact, any circuit which can

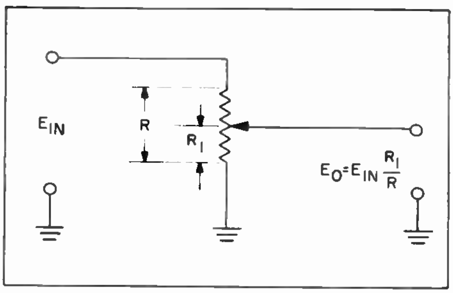

Figure 1. Multiplier Employing Only a Potentiometer

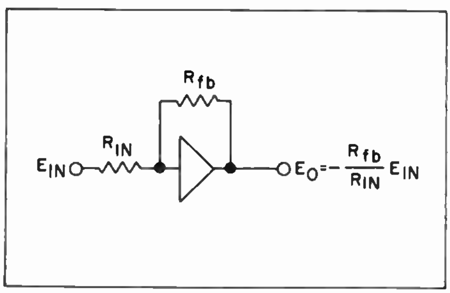

Figure 2. Multiplier Circuit Employing a Computer Amplifier

perform multiplication, such as one of those already described, may also perform division. Since the quotient must be finite, the divisor cannot be less than a certain small quantity in an actual circuit, and obviously cannot be zero.

DIFFERENTIATING CIRCUIT

The derivative of a quantity with respect to time can be thought of as the time rate change of that quantity. For motion along a straight line, for example, the derivative of the distance traversed with respect to time is the velocity or the time rate of change of distance. Similarly, the derivative of a voltage with respect to time is the time rate of change of that voltage. Figure 3 shows a graphical representation of the derivative of a voltage. If a voltage (E_{in}) is changing at a constant rate (has a constant slope) as shown in part A, then the derivative (E_o) of that voltage has a constant value, as shown in part B. A common symbol for the time derivative of a quantity is $\dot{y}$, usually read y-dot.

A simple differentiator circuit consists of a series-connected resistor and capacitor, as shown in part A of figure 4. Notice that the output voltage of this circuit appears across the resistor. With the proper values of R and C to provide a short RC time constant and with a square-wave input voltage (E_{in}), the output voltage (E_o) will be that shown in part B of the figure. When the rate of change of E_{in} is greatest, the value of E_o is greatest, and when the rate of change of E_{in} is zero, the output (E_o) tends toward zero. These facts illustrate

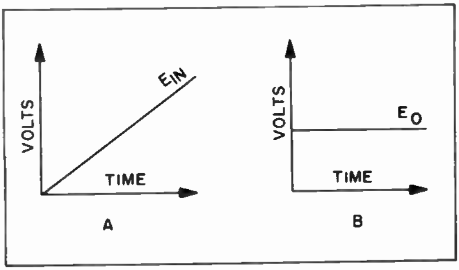

Figure 3. Graphical Representation of the Derivative of a Voltage

that the output voltage is approximately equal to the rate of change (derivative) of the input voltage, or $E_o = \dot{E}_{in}$. The square-wave derivative of a triangular-wave input is shown in part C of figure 4. Here, the rate of change of E_{in} is linear, and E_o is of constant amplitude as illustrated in figure 3. When E_{in} abruptly reverses direction, E_o abruptly reverses polarity.

The primary disadvantage of the simple differentiating circuit is the time required for the output voltage to become equal to the derivative of the input voltage. Shortening the RC time constant to decrease this time decreases the amplitude of the output voltage.

A feedback amplifier differentiator is shown in figure 5. This type of differentiator is superior to the simple differentiator circuit, because its output voltage approximates the derivative of the input voltage in a much shorter time and with greater acccuracy. As in the amplifiers previously discussed, the voltage at the amplifier grid is held at a value close to ground potential, and all

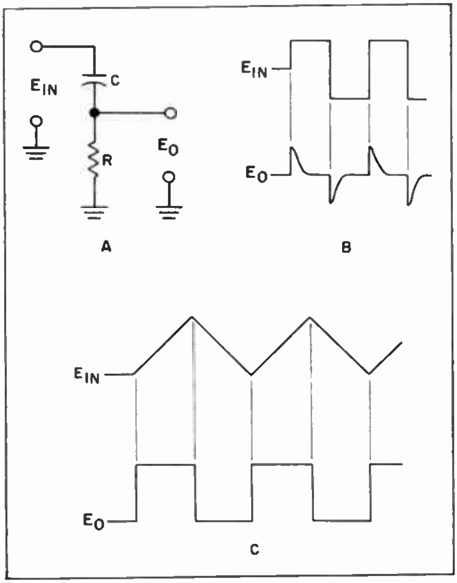

Figure 4. Differentiating Circuit (Part A), Square-Wave Input and Differentiated Output (Part B), and Triangular-Wave Input and Differentiated Output (Part C)

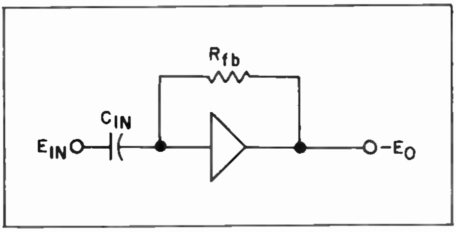

Figure 5. Differentiating Circuit Employing a Computer Amplifier

of the current (I) that flows in the input impedance also flows in the feedback impedance. The output voltage (E_o) is then given by the formula

$$E_o = -I\,R_{fb}. \qquad (2\text{-}2)$$

Hence, it is necessary to find the current (I) in order to find the output voltage.

Since the grid side of C_{in} in figure 5 is always nearly at ground potential, it is apparent that the voltage across C_{in} is approximately equal to E_{in}. When the input voltage increases linearly, as in part C of figure 4, a constant charging current (I) through C_{in} results. Since this same current flows through R_{fb}, the output voltage (E_o) must also be constant. It will be observed by consulting figure 5 that the output voltage has a polarity opposite to that of the input voltage, because of the inversion characteristic of the amplifier. Thus it can be seen that it is now only necessary to determine the amplitude of E_o in equation (2-2).

A coulomb is defined as the quantity of electric charge conveyed by a current of one ampere flowing for a period of one second, or

$$Q = It, \qquad (2\text{-}3)$$

where Q is the charge in coulombs, I is the current in amperes, and t is the time interval in seconds. The charge accumulated on a capacitor is dependent upon the voltage across the capacitor and upon the size of the capacitor, or

$$Q = CE_c, \qquad (2\text{-}4)$$

where C is the capacitance in farads and E_c is the voltage across the capacitor. As was previously mentioned, the voltage across the capacitor is, for all practical purposes, equal to the input voltage, or

$$E_c = E_{in}. \qquad (2\text{-}5)$$

Substituting E_{in} for E_c in equation (2-4), and substituting the result into equation (2-3) solved for I gives

$$I = C\frac{E_{in}}{t}. \qquad (2\text{-}6)$$

Substituting this value of I into equation (2-2) gives

$$E_o = -C\frac{E_{in}}{t}R_{fb}. \qquad (2\text{-}7)$$

$\frac{E_{in}}{t}$ may, in this case, be written as $\dot{E}_{in}$, since here the derivative of E_{in} has a constant value for any time t. Making this substitution in equation (2-7) gives

$$E_o = -R_{fb}C\,\dot{E}_{in}. \qquad (2\text{-}8)$$

If R_{fb} were 1 megohm and C_{in} were 1 μf., the RC time constant would be 1 second; for this condition, if the rate of change of the input voltage were 5 volts per second, the output voltage would be -5 volts.

INTEGRATING CIRCUIT

The process of integration is essentially a process of summing all of the parts of a quantity. The integral of a d-c voltage is a voltage with a constant slope, as shown in figure 6. In this example, the voltage at the end of one second is A volts, at the end of two seconds, B volts, etc., the value of the voltage at any instant representing the sum of the inputs for the specified time. A simple integrator is shown in part A of figure 7. Here a square-wave voltage is applied to the input, and the output voltage (E_o) appears across the capacitor. During the positive portions of the input voltage, the output voltage is the sum of all the positive quantities, which

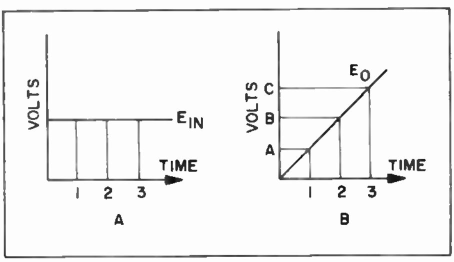

Figure 6. Graphical Representation of Integration of a Voltage

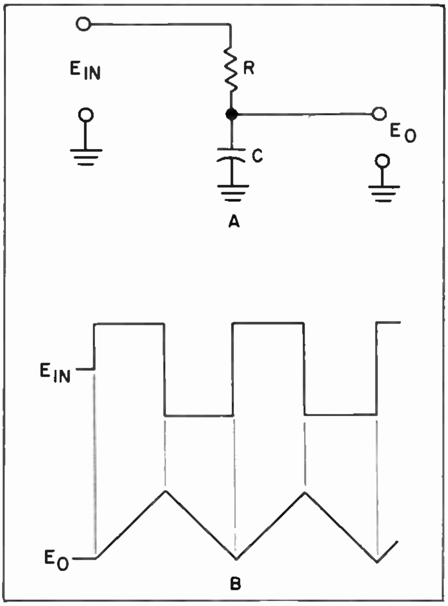

Figure 7. Simple Integrating Circuit (Part A), and Square-Wave Input and Integrated Output (Part B)

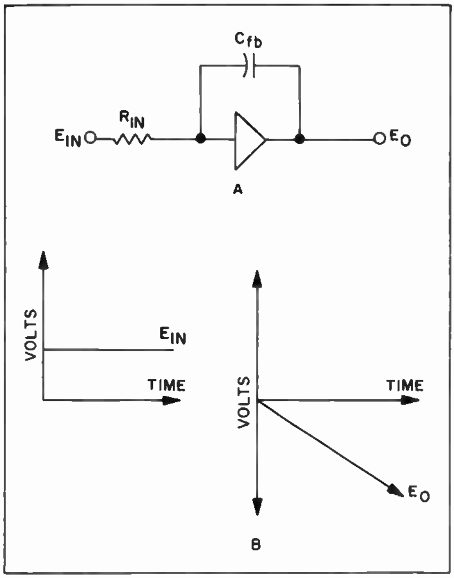

Figure 8. Integrating Circuit Employing a Computing Amplifier (Part A), and Steady Input Voltage and Integrating Output Voltage (Part B)

result in an increasing voltage. During the negative portions of the input voltage, the output voltage is the sum of all of the negative quantities in the input, which results in a decreasing voltage. Comparing the waveforms of figure 7 with those of figure 4 shows that differentiation and integration are inverse processes, i.e., if a square-wave input were applied to a differentiator and the output of the differentiator, in turn, applied to an integrator, the output of the integrator theoretically would be identical to the square-wave input to the differentiator.

An integrator employing a computing amplifier is shown in part A of figure 8.

When the input voltage is constant, the input current is also constant. As this charging current flows through the capacitor, the voltage across the capacitor increases linearly, as has already been indicated. Because of the direction of the capacitor charging current, the ouput voltage (part B of figure 8) is a negative-going voltage which is the integral of the input voltage.

FORWARD SCATTER OF RADIO WAVES

PART 1

A New Means of Communications

(Editor's Note: This article is the first of a two-part discussion on scatter propagation which appears through the courtesy of NBS. This part deals principally with ionospheric forward scatter, and the second part will deal with tropospheric scatter experiments and the theory of forward scatter.)

GENERAL

FORWARD SCATTER, a recently discovered mode of radio propagation, promises to extend greatly the limits of long-distance communication. Through application of scatter propagation, new frequency channels can be opened up to long-range use, new path lengths can be utilized, and reliable service can be provided far beyond the line-of-sight limit in the extreme high-frequency ranges. Both industry and the armed services anticipate savings through elimination of many relay stations.

Radio propagation by forward scatter is believed to result from small inhomogeneities due to turbulence in the atmosphere. These inhomogeneities scatter radio waves in all directions, but predominantly forward. With properly designed equipment, the scattered signal can be received over long distances along the earth's surface even though the direct wave has gone off into outer space.

While this phenomenon has been known to science for several years, extensive research has been necessary before it could be successfully applied to radio communication. For more than 5 years, forward scatter has been the subject of intensive experimental and theoretical investigation at the National Bureau of Standards. Sponsored by the Department of Defense, this work has not only provided insight into the nature and causes of the scattering process but has also laid much of the groundwork for effective application of forward scatter in practical communication circuits.

The NBS program has included both ionospheric and tropospheric forward scatter. Ionospheric scattering takes place in the lower part of the ionosphere —a region of electrified particles 40 to 200 miles above the surface of the earth. Tropospheric scattering occurs in the part of the atmosphere that lies below the ionosphere.

Tropospheric forward scatter appears to be useful for transmissions over distances up to 600 miles, such as in air-to-ground communication between a plane in flight and its control tower, at frequencies from 100 to at least 10,000 mc. Ionospheric forward scatter permits communication in the frequency range from 25 to approximately 60 mc., and over distances extending from approximately 600 to 1200 miles.

Beginning with frequencies where radio waves are no longer reflected by the ionosphere as they are in ordinary long-distance transmissions, ionospheric forward scatter extends into the v-h-f (30 to 300 mc.) region. This part of the radio-frequency spectrum was formerly considered useless for regular long-distance communication. In the regions where transmission is by tropospheric scatter, the available frequencies are increased many fold. Scatter propagation also helps to fill in gaps of path lengths that have not been satisfactorily covered.

NBS research in both areas has progressed simultaneously. The tropospheric program has been centered at Boulder, Colo., with the principal transmitters on Cheyenne Mountain near Colorado Springs and receiving stations in eastern Colorado and Kansas. The ionospheric work has utilized a variety of paths to evaluate variables such as the effects of the auroral zone. One of these paths—between Cedar Rapids, Iowa, and Sterling, Va.—has been used continuously for nearly 5 years. Refer to figure 1.

IONOSPHERIC FORWARD SCATTER

When radio waves were first discovered, they were shown by physicists to be transverse electromagnetic waves identical to those of light except for their frequency of vibration. It was therefore believed that they would no more go beyond the visible horizon than would light rays bend around a corner. However, in thus simplifying the problem, the physicists of that time apparently overlooked the fact that light rays are reflected from mirrors and that they are scattered by atmospheric irregularities such as clouds or fog. When Marconi demonstrated in 1901 that radio waves could be received across the Atlantic Ocean, there was at first no satisfactory explanation for his observations. Search for such an explanation eventually led Heaviside and Kennelly to propose that high up in the atmosphere there are layers of electrified particles, or ions, capable of reflecting radio waves much as a mirror reflects light waves. Thus radio waves can be propagated far beyond the visible horizon by reflection from these ionized layers, known collectively as the ionosphere. Ionospheric reflection has been for years the basis of all radio circuits more than a few hundred miles in length.

Scattering of radio energy by the ionosphere is similar to the scattering of light by small water droplets in a cloud or fog. When a beam of light shines through the mist, moving points of lights are seen in the beam. These are water droplets, ordinarily too small to be seen, scattering light out of the beam toward the observer. Similarly, irregular variations in the structure of the ionosphere are believed to result in the scattering of radio waves.

Ionospheric forward scatter is characterized by its high reliability under prac-

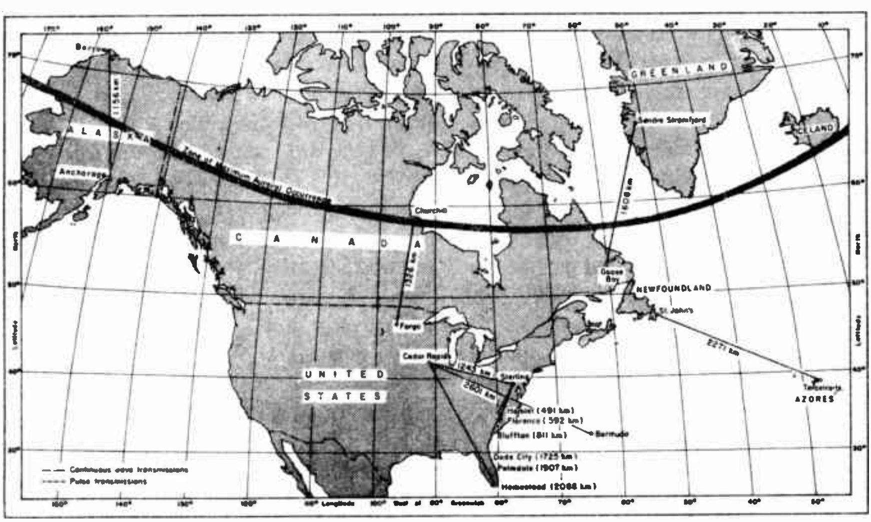

Figure 1. Map Showing the Various Experimental Paths Used in NBS Studies of Ionospheric Forward Scatter

tically all conditions. During auroral disturbances, which cause polar blackouts of ordinary high-frequency ionospheric communication, scattered signals usually become stronger. Also, strengthening is often observed during sudden ionospheric disturbances, which are associated with solar flares and cause short-term blackouts of high-frequency communication throughout the daylight hemisphere.

Auroral storms and polar blackouts constitute one of the most serious problems in arctic communications. Hitherto attempts have been made to obtain reliability during such disturbances by changing the frequency of operation. This, of course, requires installation of multiple transmitters, antennas, and receivers. A single ionospheric scatter circuit, while expensive in itself, can provide more reliable service at a single frequency and at a lesser cost.

It is generally agreed that scattering takes place from that part of the lower ionosphere known as the *E* region. Ionospheric layers exist in this region at various heights, usually between 80 and 110 km. above the earth's surface. Ordinarily, radio waves (i.e., up to about 25 mc.) pass directly through the *E* layers and are reflected from outer layers, usually in the *F* region as shown in figure 2. However, there is some quality about the *E* region which causes VHF radio waves to be scattered in all directions when they strike it. This scattering can occur from a large number of small irregularities in the ionization of the atmosphere, much as light is scattered from the large number of small water droplets in a cloud or fog. A very small portion of the scattered energy will reach a receiver which is pointed directly toward the spot where the incident wave strikes. This principle may be utilized for radio communication if (1) the radio wave is transmitted with sufficient power, (2) the transmitting antenna has high gain and is properly directional, and (3) a high-gain directional receiving antenna is used.

At least three major factors contribute to the ionization of the *E* region, which causes ionospheric scatter. These are direct solar radiation, corpuscular radia-

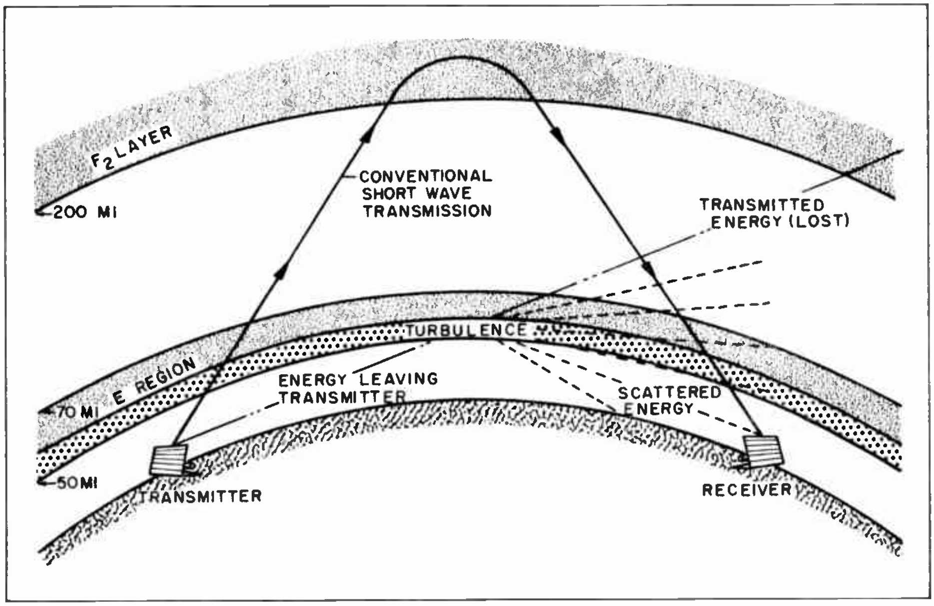

Figure 2. Comparison of Conventional Ionospheric Reflection of Radio Waves and Ionospheric Scatter of Radio Waves

tion, and meteors. The contributions of each of these factors are not known exactly and need more study. Some observations, however, seem to identify behavior noted at different times with specific mechanisms.

In a series of measurements over an arctic path (Anchorage to Barrow, Alaska), correlation with magnetic activity indicated a rise in signal strength at the receiver with increasing magnetic activity at the midpoint (Fairbanks). One explanation is that the contribution of the corpuscular radiation factor at these times is high. Corpuscular radiation, presumably of solar origin, consists of atomic particles of matter which take from 18 to 30 hours to reach the earth from the sun. When they enter the earth's magnetic field, they are caught up by it and spiral toward the pole. The net result is more ionization in the lower region of the ionosphere.

According to one school of thought, ionization by meteors is the dominant mechanism. Literally millions of tiny particles of meteor dust enter the earth's atmosphere every day but usually burn themselves out before striking the surface. Their passage through the atmosphere produces trails of intensely ionized gas which scatter radio waves. The meteor theory says that there are enough meteors to produce a continuous mechanism for propagation by the overlapping trails of ionization, and that this is the dominant contribution to the received signal. There is experimental evidence that this is true at some times of the day, particularly in the early morning hours.

The principal factors that limit the utilization of ionospheric forward scatter are frequency and distance. The experimental approach has been to isolate the different variables and evaluate them individually. However, it has been difficult to express a relationship with frequency.

The useful frequencies are between 25 and 60 mc. The signal strength falls off very rapidly as the frequency is increased. Thus the lower frequencies of the range will generally permit more efficient transmission. However, if too low a frequency is chosen, ionospheric reflection will also take place at certain times, and then the scatter circuit may be affected by multipath propagation of its own transmission or by other transmissions, or it may cause interference in other circuits.

Distance dependence is governed largely by the magnitude of the scattering angle, and the height of the layer of the ionosphere that produces the scatter. The scattering angle is defined as the angle between the incident wave and the scattered wave that reaches the receiver.

As observed in the NBS tests, the strength of the scattered signal falls off very rapidly as the scattering angle is increased. This occurs as the path is shortened. At less than about 600 miles the ionospheric signal becomes too weak for practical use.

At the longer distances, the signal intensity is usable because the angle is small, until the limiting distance is reached. The maximum distance is reached when the portion of the ionosphere from which scattering occurs meets the horizon as "seen" from the transmitter. Increasing antenna height helps slightly toward increasing the distance range.

In 1913, Kennelly suggested the possibility that radio waves might be scattered as well as reflected by the ionosphere. In 1932, Eckersley proposed a theory of scattering to account for certain special effects he observed, but no one appears to have seriously proposed ionospheric scattering as a useful communication mechanism until 1950. In December of that year a group of radio scientists holding a conference at Massachusetts Institute of Technology suggested that scattering in the ionosphere might produce a signal sufficiently strong to be useful, provided a strong transmitter, high-gain antennas, and a sensitive receiver were used.

A cooperative experiment was promptly arranged in which the M. I. T. Lincoln Laboratory sponsored operation of a transmitting site by the Collins Radio Co. at Cedar Rapids, Iowa, while the National Bureau of Standards operated a receiving site at its field station at Sterling, Va. The distance is 773 miles.

After experimental transmissions during the next month had demonstrated that signals were continuously received and that they exhibited interesting and promising behavior during ionospheric disturbances, a program of further experimentation was formulated. Transmissions on 49.8 mc. over the original path have continued without interruption since January 23, 1951. The program has subsequently been enlarged in various ways, but always with the dual objective of (1) investigating the physical principles and theories involved in ionospheric scatter, and (2) providing engineering information for the design, construction and operation of communication systems. The work at NBS has been carried out under the sponsorship of the Department of Defense and the Air Force.*

In addition to the Cedar Rapids-Sterling path, which has been operated continuously, several other paths have been studied for shorter periods of time. Since the method offers the advantage of reliability during ionospheric disturbances such as auroral storms and polar blackouts, three paths were established in regions where the aurora is most active.

One, from Fargo, North Dakota, to Churchill, Canada, terminated practically in the zone of maximum auroral occurrence. The other, from Anchorage to Barrow, crossed the maximum auroral zone practically at the midpoint of the path. An ionospheric station at Fairbanks provided soundings at the approximate midpoint. These paths were almost the same length as the Cedar Rapids-Sterling path so that effects due to differing distances were eliminated. Another auroral-zone path was also tested in 1952 and 1953 from Goose Bay Labrador, to Sondre Stromfjord, Greenland. Much valuable data on the characteristics of these paths were obtained. The investigators found that with careful design of the path components, it is possible to communicate with very high reliability even in these regions where communication by ionospheric reflection is often badly disrupted by natural phenomena.

A path from St. John's, Newfoundland, to Terceira Island in the Azores gave an opportunity to study problems associated with extreme distance—1411 miles in this case. Here the terrain was such that high antenna sites could be chosen overlooking the ocean. Under these circumstances it was verified that a distance of 1400 miles is feasible for reliable operation. An interesting feature of this test is that it combined frequency-standard techniques and propagation measurements in improving the efficiency of the test setup. Portable standard-frequency oscillators developed by P. G. Sulzer of NBS were employed at both the transmitter and receiver to permit using a 40-cycle receiver bandwidth at 36 mc. transmitter frequency. This reduced the necessary transmitter power and antenna gain by improving the signal-to-noise ratio of the circuit.

In addition, short-term experiments for studying other variables utilized other paths, mostly from Cedar Rapids and from Sterling. Tests of reception near the maximum distance were carried out from Cedar Rapids to Dade City, Palmdale, and Homestead, all in southern Florida. In 1952 and 1954 short-distance tests were made from Sterling, Va., to Bluffton, S. C. In these

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*For technical details, see "Radio Transmission at VHF by Scattering and Other Processes in the Lower Ionosphere," by D. K. Bailey, R. Bateman, and R. C. Kirby, Proc. IRE 43, 1181 (1955); and "Some Fading Characteristics of Regular VHF Ionospheric Propagation," by G. R. Sugar, Proc. IRE 43, 1432 (1955).

tests the time of arrival of ionospheric scatter waves was compared with that for more direct tropospheric signals. This gave direct evidence concerning the height of the scattering layers. In 1951, an attempt was made to look for scattering from higher layers such as the F_2, on a path from Cedar Rapids to Bermuda. This path was chosen as being sufficiently long so that the lower layers would be well beyond the horizon. At the frequency used—near 50 mc.—occasional weak signals were observed, and these were attributable to other mechanisms.

As part of the United States participation in the International Geophysical Year, it is planned that NBS will carry out in 1957-1958 an ionospheric scatter experiment in Peru, where the midpoint of the path will be nearly at the geomagnetic equator. This experiment will dedetermine scattering behavior in a geographical region having ionospheric characteristics not previously studied from the scattering standpoint. The possibility of high-level scattering in that region will be investigated.

Large rhombic antennas have been used at both ends of the path for most of the experiments. Where space and suitable terrain were not available for rhombics, Yagi arrays were usually substituted. Such an array is shown in figure 3. In order to obtain a suitable comparison between the different types of antennas, various combinations were installed at several transmitting and receiving sites. In some experiments, transmitter and receiver were switched alternately between two antennas at half-hour intervals. For example, a combination of rhombic to rhombic and Yagi, and of Yagi to rhombic and Yagi, was operated for a time between Anchorage and Barrow. Here the two receiving antennas were used simultaneously and continuously. For continuity in observations on the Cedar Rapids-Sterling path, the rhombic antennas were always used as the reference pair.

Rhombic antennas up to 25 wavelengths on a side have been constructed. For a frequency of 50 mc., approximately that used on all paths except Newfoundland-Azores, this means 500 feet on a side. Construction and successful application of antennas of such size confirmed that conventional design of rhombic antennas could be extended to larger sizes.

In order to study antenna patterns and to help in predicting performance, scaled systems were constructed. If the system is scaled small enough, experimental measurements can be made within a reasonable distance of the antenna. Precise scale models of the antennas, an example of which is shown in figure 4, were thus studied systematically at the Bureau's model antenna ranges at Sterling, Va., and Boulder, Colorado.

In view of the promising characteristics of the new mode of ionospheric propagation in arctic regions, the U. S. Air Force asked the Bureau in 1951 to assume responsibility for setting up the first regular communication circuit using ionospheric scattering in the VHF. Initially a 48-mc. experimental 1-way circuit was tested between Labrador and Greenland. This was followed by a complete prototype system operated at frequencies between 30 and 40 mc., consisting of three 2-way circuits extending from Maine to northern Greenland. Recently this system has been extended under NBS supervision from Goose Bay, Labrador, via Narsarssuak, Greenland, to Reykjavik, Iceland. Detailed engineering design was worked out with the help of the E. C. Page Consulting Radio Engineers, and the installation was contracted to that firm.

Since only a small portion of the transmitted signal is scattered by the ionosphere, it is necessary to use large amounts of gain in a communication system. This calls for powerful transmitters, high-gain antennas, and sensitive receivers. The operating circuits that have been installed use 40-kw.

transmitters, high-gain antennas designed by NBS, and sensitive receivers designed by M.I.T. Lincoln Laboratories.

The complete four-channel multiplex radio teletype systems began operation in 1953. After a brief trial period, the prototype circuits were made available to the Air Force for test traffic. A care-

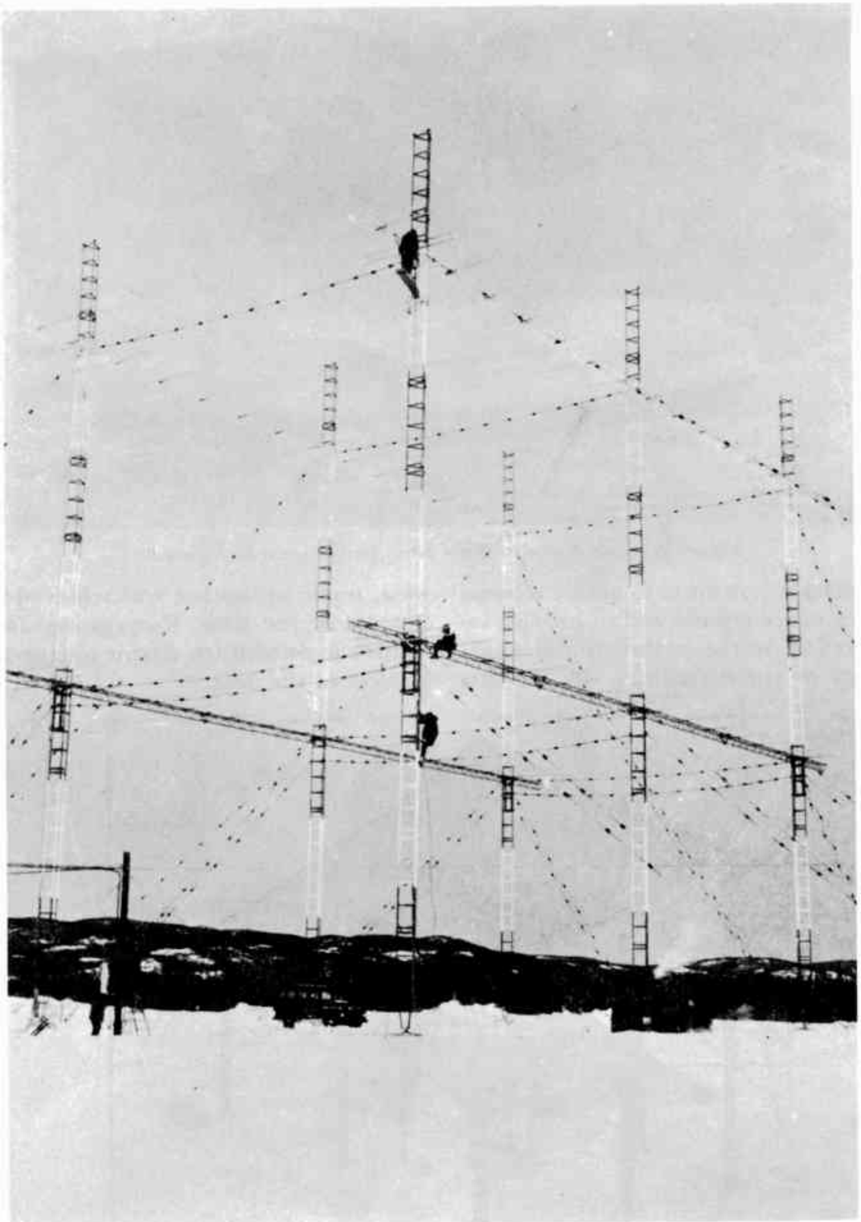

Figure 3. A Yagi Array Used in an Arctic Installation

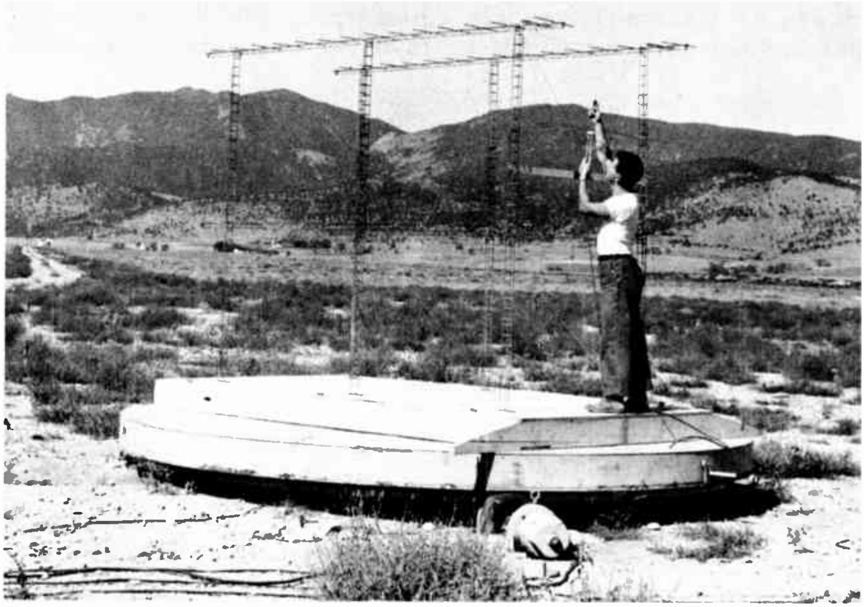

Figure 4. Scale Model of Huge Yagi Array Shown in Figure 3

ful attempt was made to obtain accurate performance reports and to provide information on the problems and weaknesses of these facilities. In the first year, traffic utilization was achieved 91 percent of the time. Propagation difficulties accounted for circuit outage 1.1 percent of the time.

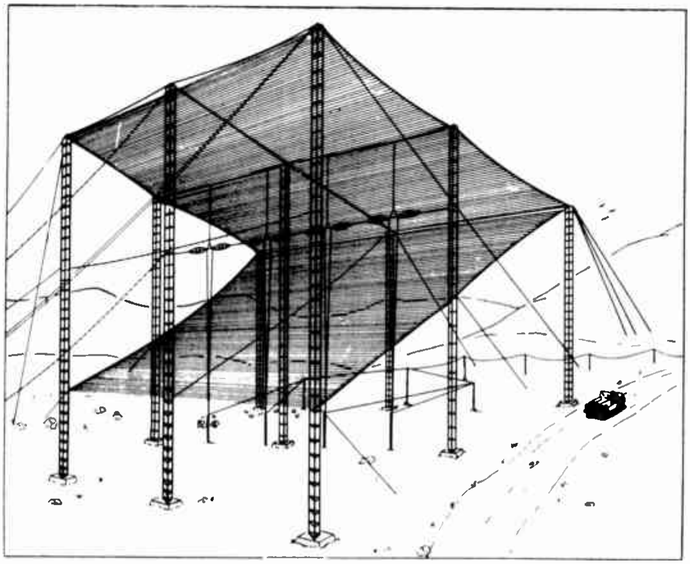

Figure 5. Large Corner Antennas, Designed and Constructed by E. C. Page Consulting Radio Engineers in Cooperation with the National Bureau of Standards, Are Being Studied for Effectiveness in Comparison With Other Types of Antennas

MTI VECTORS

by Robert D. Hunter
Headquarters Technical Staff

The following is a qualitative discussion of the r-f portion of MTI (Moving Target Indicator) operation using "vectors." The vector approach is especially suited for instruction purposes on MTI radar equipments, which are currently in wide use. Although this material is specifically slanted toward the AN/CPS-6B system, it is applicable in general to other MTI systems such as the AN/TPS-1D, AN/FPS-3, and AN/CPS-5D.

GENERAL

Alternating voltages and currents are commonly pictured by the use of vector diagrams such as that shown in figure 1. This rotating vector has an absolute magnitude E, and rotates with an angular velocity of ω radians per second, where ω equals $2\pi f$, and f is the frequency of the alternating voltage or current. The instantaneous voltage (e) at any time (t) is the projection of this vector on the reference axis OA at P′, and is equal to E cos $(\omega t + \phi)$, where ϕ is the initial phase angle.

It can be shown that an alternating voltage can be represented by a pair of contra-rotating vectors.* This method of representation must be used whenever it is desired to superimpose vectors representing signals of different frequencies. Contra-rotating vectors are illustrated in figure 2, where the vector V and its conjugate V′ rotate in opposite directions with an angular velocity ω. The resultant R of these two vectors always lies on the reference axis AA′. The remaining vector diagrams in this discussion omit the conjugate vectors, in order to improve the clarity of the diagrams, particularly of the phase angles involved; however, it should be remembered that conjugate vectors are actually also being considered.

Figure 3 illustrates the addition of ordinary vectors, E_1 and E_2, to obtain the resultant vector, E_3. At this point it is desired to show that the phase angle, ϕ_3, of E_3 depends upon phase angles ϕ_1 and ϕ_2 of both of its components. In each of the vector diagrams used in this article, the vectors are shown at instantaneous positions with their phase angles being measured with respect to the reference axis AA′.

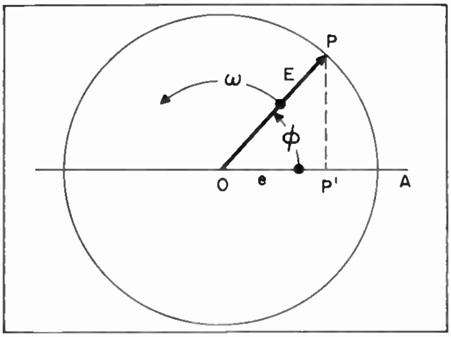

Figure 1. Vector Representation of a Sinusoidal Voltage of Frequency f and Initial Phase Angle ϕ

*Goldman, S., *Frequency Analysis—Modulation and Noise*, p. 160.

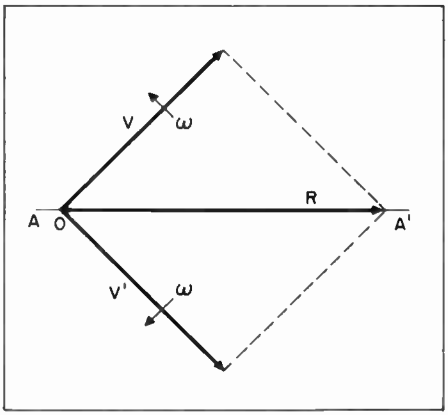

Figure 2. Representation of an Alternating Voltage by Contra-Rotating Vectors

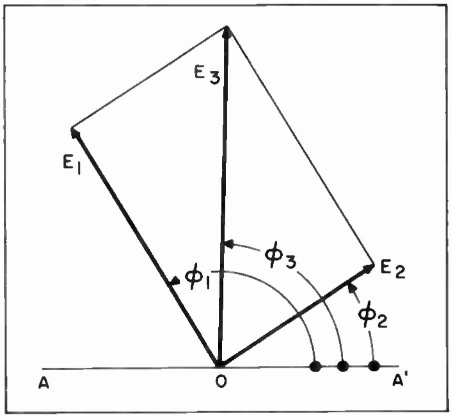

Figure 3. Addition of Conventional Vectors E_1 and E_2 to Obtain Resultant Vector E_3

VECTOR APPLICATION TO MTI PRINCIPLES

A block diagram showing the major r-f components of an MTI system is shown in figure 4. Assume that all of the oscillators and other circuit elements are perfectly stable, and consider the mixing of the stable local oscillator (stalo) signal with the transmitter pulse in the a-f-c mixer. The i-f signal which results, is referred to as an i-f lock pulse. As shown in figure 5, the starting phase of the lock pulse depends upon the starting phases of the transmitter and stalo signals, henceforth abbreviated TX and LO, respectively. The instant chosen to "stop" the rotating vectors in figure 5 is the instant at which locking is completed. Although the i-f lock pulse disappears when the transmitter pulse ends, the signal is preserved in its frequency and phase angle by the coherent oscillator (coho), assuming that the coho is ideally stable and locked.

The timing diagram of figure 6 shows the significance of the numerical subscript used in the applicable vector diagrams starting with figure 7. Referring to figure 6, instant 1 is the instant at which locking is completed, instant 2 is a corresponding point during the reception period. There are similar points in the second pulse period (and all succeeding pulse periods), and these are

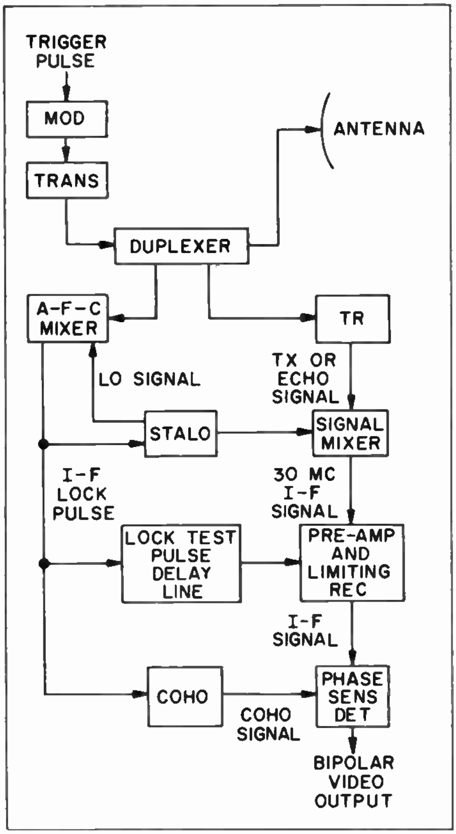

Figure 4. Block Diagram of R-F Components of an MTI System

noted on the applicable vector diagrams. The phase angle between the coho and i-f echo vectors will be designated by θ (theta), while ϕ (phi) will be used to designate other phase angles.

Referring to figure 4, suppose that considerable transmitter signal leaks through the TR tube and mixes with the stalo signal in the signal mixer to produce an i-f echo signal similar to the i-f lock pulse. When this occurs, the resulting signal proceeds through the preamplifier and the limiting receiver to the phase-sensitive detector. At this point the phase angle between the i-f echo signal and the coho signal would be zero except for the difference in the circuit path length, because both signals are made up of the same transmitter and local oscillator frequency compo-

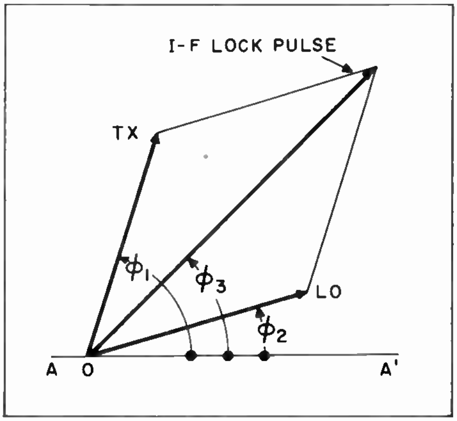

Figure 5. Combination of TX and LO Signals to Provide Lock Pulse

nents. Hence, in this case, θ is the same for all transmitter pulses *regardless of the starting phases of the transmitter and stable local oscillator.*

A situation similar to the one just described will prevail in the case of a fixed target return. Figure 7 shows representative positions for the vectors involved in the "first pulse period" for a fixed target situation. At instant 1, referred to in figure 6, the positions of the transmitter (TX_1), stalo (LO_1), and i-f (IF_1) lock pulse vectors are in the positions shown in figure 7. For discussion purposes suppose that the IF_2 coho vector executes (N + ½) revolutions between instants 1 and 2. (Actually, the number of revolutions could be almost any amount depending upon the target distance.) Also shown in figure 7 are the positions of stalo LO_2 and r-f echo signal TX_2 at instant 2. These are added so

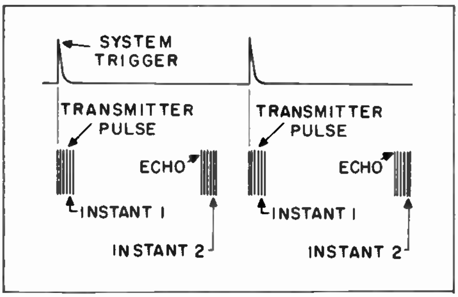

Figure 6. Timing Diagram for Radar Pulses

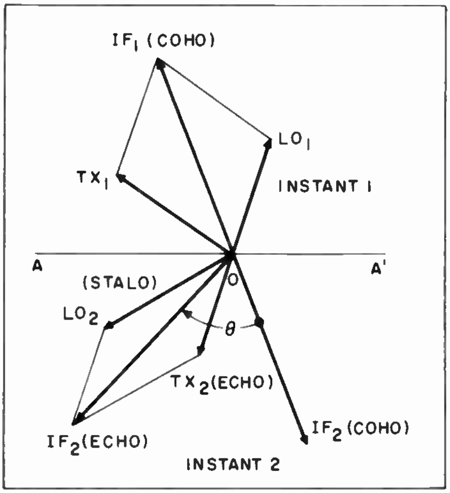

Figure 7. Vectorial Picture of Conditions for a Fixed Target

that i-f echo signal IF_2 may be obtained. In reality, the echo signal TX_2 is much smaller than shown; hence the phase of the i-f echo signal is more dependent upon LO_2. Since the IF_2 echo and coho signals have the same frequency, the phase angle θ between them will remain constant for the duration of the pulse. As long as the elapsed time between the transmission and reception of a pulse remains the same, the angles between LO_1 and LO_2, between TX_1 and TX_2, and between the coho and i-f echo signals will be identical.

For the case of a moving target, assume the conditions for a fixed target during the first pulse period as shown in figure 7. If the target range decreases between the first and second pulse periods, the transmitter, stalo, and coho vectors will not have time to sweep through the same angle as they swept through during the first pulse period. Figure 8 shows the conditions for the second pulse period in this case. For ease of comparison with figure 7, the vectors in figure 8 at instant 1 are shown in the same position that they occupied in figure 7. A small change in range (or elapsed time) would not appreciably change the position of the

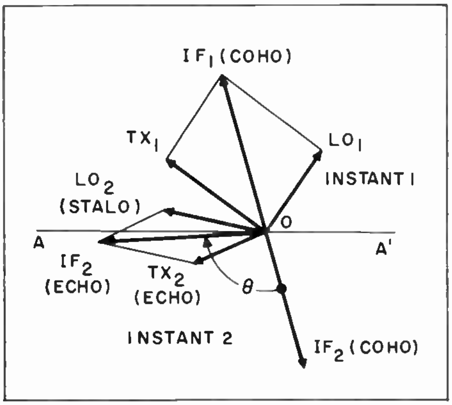

Figure 8. Conditions for a Moving Target, Second Pulse Period

coho vector at instant 2, but would definitely change the positions of the transmitter and stalo vectors at this time. As shown in figure 8, this changes angle θ between the IF_2 coho and the IF_2 echo signal. Since the LO_2 vector rotates with a much higher angular velocity than the coho vector, it is easy to see the reason for the stringent stability requirement for the local oscillator. When the target range increases between pulse periods, angle θ is decreased, as may be concluded by reasoning conversely.

It is the function of the r-f section of the MTI system to produce signals whose pulse-to-pulse characteristics distinguish them as either fixed or moving targets. Since the output of the phase-sensitive detector depends upon the angle between the coho and i-f signals fed to it, this purpose is accomplished. A fixed target observed at the output of the phase detector will ideally have constant amplitude and polarity from one pulse period to the next, while a moving target will generally change in amplitude or polarity (or both) from pulse to pulse. The response of a typical phase-sensitive detector is shown in figure 9. Notice that the output can be either positive or negative (bipolar signal). Figure 7 illustrates that a pulse-to-pulse change in target range of an integral multiple of one-half wavelength at the transmitter frequency would cause the TX and LO vectors to occupy the same positions in which they are shown, because of the full wavelength change in r-f path length. For this condition no apparent change in the target range can be detected; hence, the target speed at which this condition occurs is called the "blind speed."

VECTOR APPLICATION TO MTI SYSTEM TROUBLES

Vectors are useful in analyzing certain imperfections in the system, to determine their effect on MTI performance. The diagram of figure 10 shows the effect of coho mistuning. It is assumed for this discussion that the i. f. is 30 mc., the pulse duration is 1 microsecond, and the detuning is slightly less than 1 megacycle. Using a 1-microsecond test pulse obtained from the delay line test circuit shown in figure 4, or using a fixed target

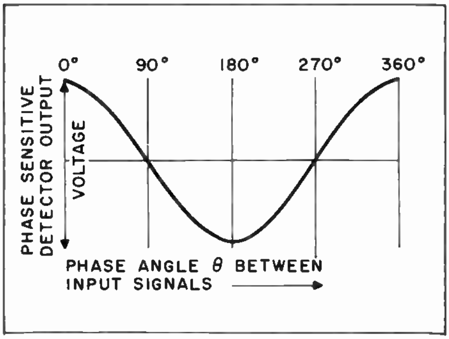

Figure 9. Response Curve for a Typical Phase-Sensitive Detector

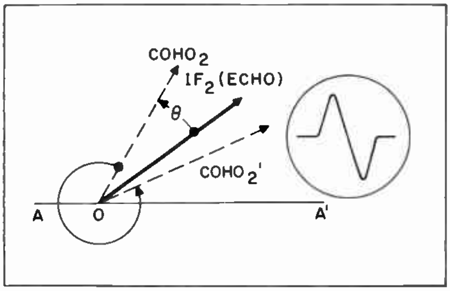

Figure 10. Coho Mistuned by Nearly 1 Megacycle for an I. F. of 30 Mc and a Pulse Duration of 1 μsec (Lock Test or Fixed Target Signal)

return, the vector labeled $coho_2$ would execute nearly one more revolution (if the coho frequency were too high) than would the i-f echo vector IF_2, which is shown in a fixed position for the duration of the pulse. This phase variation during the pulse interval will cause the output of the phase-sensitive detector to vary in the manner shown in the inset of figure 10. Lock test pulses are, in effect, ideal fixed targets which may be used to check the tuning of the coho. They are simply a series of reflections of the i-f lock pulse which is introduced into the MTI receiver. An examination of this signal with a test scope will show the extent to which the coho is mistuned.

Figure 11 illustrates the effect of a pulse-to-pulse drift in coho frequency. Here, the i-f echo vector is shown in the same position for two successive pulse periods, while the coho vector is shown in different positions. The phase angle θ between them is seen to change by an amount $\Delta\theta$ between successive pulses. The term $\Delta\theta$ is proportional to the target range for a linear drift of coho frequency. It is the practical limit of stalo and coho stability which causes fixed targets to lose their "coherency" at longer ranges. (Actually, scanning and other variations are also greater for long range targets.)

The effect of a pulse-to-pulse drift in stalo frequency is shown in figure 12. Here, for the sake of clarity, all of the vectors except the stalo and i-f echo vectors are shown in the same position for two successive pulses. It is apparent that the drift in stalo frequency produces a change $\Delta\theta$ in the angle θ between the coho and i-f echo signals. As before, $\Delta\theta$ is proportional to the range of the target.

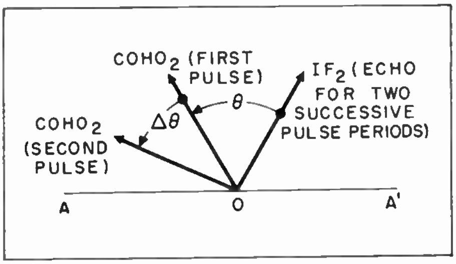

Figure 11. Effect of Pulse-to-Pulse Frequency Drift in Coho Signal Causes θ to Change by an Amount Δθ (Δθ Is Proportional to Range of Fixed Target)

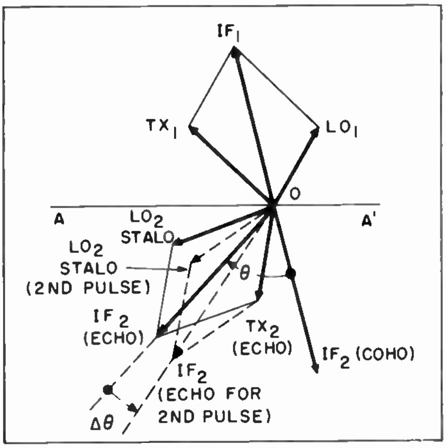

Figure 12. Effect of Pulse-to-Pulse Drift in Stalo Frequency

IMPORTANT MTI CHARACTERISTICS

In mobile MTI systems or those which have off-set antennas that cause the fixed targets to appear to move, the coho signal must be accurately shifted in phase in order to restore coherency to the fixed targets. Figure 13 illustrates this situation. It is assumed that for the first pulse period the phase angle between the coho and i-f echo signal is θ.

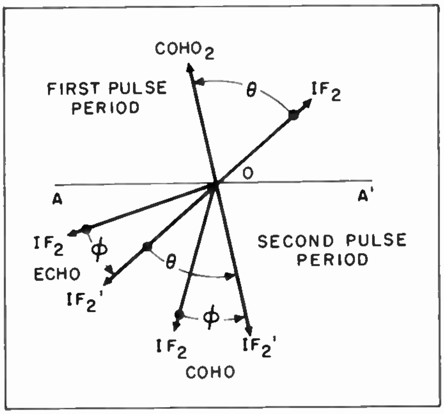

Figure 13. Phase-Shifting of Coho Signal in Order to Maintain Angle θ

For the second pulse period a phase shift of θ is introduced into the i-f echo signal from IF_2 to IF'_2 by the antenna rotation or the motion of the system. It is obvious that the phase angle of the coho signal must be shifted by an angle ϕ in order to restore the angle between the echo and coho signals to θ. For each successive cycle of operation an additional phase shift ϕ will be introduced into the i-f echo signals, and this must be compensated for by a continual shift in the phase of the coho signal at a constant rate of ϕ radians per pulse period.

In an MTI system of the type under consideration, the response of the phase-sensitive detector is very important. Both i-f echo and coho signals are passed through limiting circuits, to make the output of the phase-sensitive detector independent of the amplitude of the input signals. Assuming that only phase-angle changes take place, the output of the phase-sensitive detector should change as much as possible from pulse to pulse for moving targets, and remain as constant as possible for fixed targets. This is best accomplished by a detector having a linear characteristic like the ideal response curve shown in figure 14. If the levels of the coho and i-f echo signals are not matched, however, the output of the phase-sensitive detector will be governed by the "undesirable response" curve shown in figure 14. Consider first that part of the undesirable response curve which lies between 0 and approximately 60 degrees. It can be seen that the rate of change of output for increasing phase angles through this range of values will be small. As a result, both fixed and moving targets will be cancelled over this range. The slope of the curve is greater between 60 and 120 degrees, but it will be more difficult to cancel clutter signals having angles in this region. From 120 to 240 degrees, the conditions are the same as from 0 to 60 degrees, etc.

Since the basic purpose of MTI is to enhance the appearance of a moving target in the presence of clutter, the desirability of the linear response for the phase-sensitive detector can be appreciated. The levels of the coho and i-f echo signals must be carefully matched in order to achieve this linearity.

Until recently the only available oscillators for S-band radars which were of practical value were the magnetron and the reflex klystron. The reflex klystron is used for the local oscillator of most of the S-band and higher-frequency radars. The fact that these oscillators are inherently unstable introduces a major complication into the MTI system of an S-band radar. It has been shown that the stability of the local oscillator is very important for successful MTI operation. The effect of a frequency-stabilizing cir-

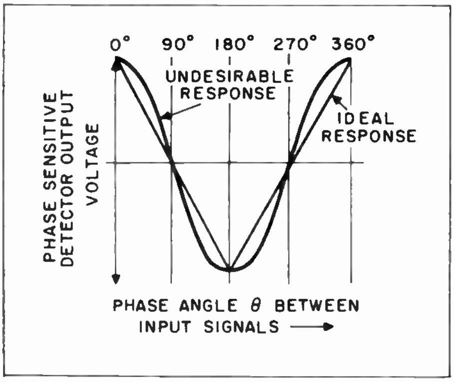

Figure 14. Comparison of Response Curves for a Phase-Sensitive Detector

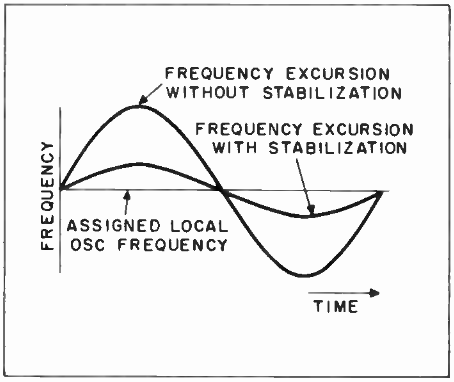

Figure 15. Effect of Frequency Stabilization Upon Local-Oscillator Signal

cuit is shown in figure 15. It is assumed here that the local oscillator is frequency-modulated in a sinusoidal manner as shown, the deviation being symmetrical about the assigned local-oscillator frequency. The extent of the frequency excursion is decreased by the gain of the stabilizing loop, as shown in the figure. Therefore, the rate of change of frequency (slope of curve) with time is also decreased. This in turn decreases the pulse-to-pulse change in θ produced by the drift of local-oscillator frequency, and contributes much to successful MTI operation at S-band frequencies. For the stabilizing circuit, the requirements are stability and high gain. In the circuit used in the AN/CPS-6B, the attainment of these requirements calls for a small frequency separation between the resonant frequencies of the reference cavities.

CONCLUSION

The foregoing remarks have been restricted to the r-f portion of a typical MTI system, and even then have not included such important items as the frequency modulation of the transmitter and the timing stability of the modulator. It is hoped that those who teach MTI operation will find the approach indicated here of some value in explaining the general principles of MTI operation, and in showing the importance of certain MTI system characteristics.

Solution to January-February "What's Your Answer?"

The open-circuit impedance (Z_{oc}) of the circuit shown in the illustration is the impedance seen at terminals 1 and 2 with an open circuit between terminals 3 and 4. Therefore, in this case,

$$Z_{oc} = \frac{Z_1}{2} + Z_2$$

The short-circuit impedance (Z_{sc}) is the impedance seen at terminals 1 and 2 with a short circuit between terminals 3 and 4. Therefore, in this case;

$$Z_{sc} = \frac{Z_1}{2} + \frac{\frac{Z_1}{2} \cdot Z_2}{\frac{Z_1 + Z_2}{2}} ,$$

and

$$Z_{oc} \cdot Z_{sc} = \left(\frac{Z_1}{2} + Z_2\right)\left(\frac{Z_1}{2} + \frac{\frac{Z_1}{2} \cdot Z_2}{\frac{Z_1}{2} + Z_2}\right)$$

By carrying out the indicated multiplication:

$$Z_{oc}Z_{sc} = \frac{Z_1^2}{4} + Z_1Z_2$$

The characteristic impedance of a transmission line is the impedance Z_a which, if connected across the output terminals, would not change the value of impedance seen at the input terminals. Since this must be true for any number of cascaded T sections, then $Z_a = Z_x$.

Connecting a value Z_x of impedance across terminals 3 and 4 gives the following equation:

$$Z_x = \frac{Z_1}{2} + \frac{\left(\frac{Z_1}{2} + Z_x\right) Z_2}{\frac{Z_1}{2} + Z_x + Z_2}$$

Solving this equation gives the following result:

$$Z_x^2 = \frac{Z_1^2}{4} + Z_1Z_2$$

From the calculations above:

$$Z_{oc}Z_{sc} = \frac{Z_1}{4} + Z_1Z_2$$

Therefore:

$$Z_x = \sqrt{Z_{oc}Z_{sc}}$$

SCREEN ROOM MAINTENANCE

Pertinent hints on how to get the most from your "Cell-Type" screen rooms.

(Editor's Note: This article is reprinted from the Instruction Handbook for Cell-Type Wire Mesh Shielded Enclosures through the courtesy of Ace Engineering and Machine Company, 3644 N. Lawrence Street, Philadelphia 40, Pennsylvania.)

GENERAL

Purpose and Basic Principles

THE GENERAL PURPOSE of a shielded enclosure is the elimination of radio interference. It is used to create an interference-free region in which accurate measurements of desired signals can be made, and also to surround a source of radio interference so that the external region will be free of this particular interference.

Ideally, a shielded enclosure should provide one or more layers forming a continuum of metal or wire mesh around the room. The objective is achieved in a practical manner by *tightly* bolting together adjacent panels wrapped with wire mesh and by using special electrical contacting fingers around the access door to obtain continuous peripheral connection to the jamb when the door is closed. It is also necessary not to allow metal to pierce the shielding material unless special precautions are taken, as on the power filter panel, the r-f connector panel and the ventilator panel. Without these special precautions, wire or other metal would act as a radio antenna to pick up a signal on one side of the enclosure and transmit it to the other, thus defeating the objective of the enclosure.

Metallic Objects Penetrating Screen

Whenever a metallic object penetrates the shielding material, the object must be peripherally soldered thereto at the point of entry. It is desirable further that the penetrating object, if not on the power filter panel, be insulated from conducting objects external to the enclosure. Staples used by the manufacturer in construction of the panels are a permissible exception to this rule because of their extremely small size.

MAINTENANCE

Panel Bolts

It is vital to continued high performance of the enclosure that good electrical contact be maintained at all points between adjacent panels. For normal interior use, seams of the enclosure should be re-tightened to approximately 140 inch-pounds at the following periods after erection:

1. Two weeks
2. Six months
3. One year
4. Yearly thereafter

In abnormally corrosive atmospheres it may be necessary occasionally to disassemble the enclosure, clean the contacting surfaces of screening material with steel wool or a fine grade of sandpaper, and re-assemble the enclosure. An attenuation check using the methods of specification MIL-E-4957A(ASG), 17 November, 1954, should be compared with TABLE 1 as a criterion to determine when cleaning is necessary.

Check for Leakage

If, after a period of use, the shielding effectiveness falls short of that required by specification, it is recommended that all seams, contacts, and the screening material itself be given a visual inspection for apparent damage. If after correction the shielding performance is still inadequate, a check should be made for r-f leakage.

TABLE 1
SHIELDING EFFECTIVENESS

FREQUENCY	TYPE OF TEST	ACCEPTABLE ATTENUATION (db)
200 kc.	Loop to loop	70
200 kc.	Rod-to-rod	100
1 mc.	Rod-to-rod	100
19 mc.	Rod-to-rod	100
400 mc.	Dipole-to-dipole	100

This test should be conducted by a radio engineer or experienced technician at that radio frequency for which the shielding is defective. However, experience indicates that a check at 400 mc. provides the most sensitive test for the enclosure, and it is recommended that this frequency be used if equipment is not available for the frequency of concern.

Place a transmitter and its radiating antenna within the shielded enclosure and the sensitive receiving equipment outside the enclosure (equipment specified in MIL-E-4957A(ASG). The antenna of the receiving equipment should be used to explore the exterior surface of the enclosure, and maximum signal will normally indicate the point of difficulty. Occasionally, however, external reflecting objects may cause standing waves of sufficient magnitude to give misleading results, and good engineering judgment will be necessary. (Note that the transmitter and receiver positions are reversed from that required for the measurement of shielding performance by the MIL specification. The severity of standing waves inside the enclosure make the reversal of position necessary when leakage occurs.)

Serrated Access-Panel Fingers

Access-panel fingers and the surfaces they contact should be kept free of corrosion by wiping with steel wool when necessary. In normal atmospheres, this cleaning will be required at intervals of three months, although the actual period will depend upon local conditions at the installation.

Repair to damaged fingers is made by unsoldering the faulty unit and soldering a new piece in its place. A convenient way of doing this is to solder the new unit at various points along its length in order to hold it in position, and then to solder completely along the entire length using a 50/50 tin-lead solder and a noncorrosive rosin flux.

Repairs to Torn Screening

Any place at which screening material is torn will permit severe r-f leakage into or from the room at the higher frequencies. It is recommended that a patch of the same type copper screening material (a temporary repair may be effected using any other metallic screen or sheet metal) be placed over the tear and soldered around the entire periphery of the patch using soldering irons and noncorrosive rosin or activated rosin flux. Torches must not be used in running a seam because of the difficulty of controlling the solder flow and the greater tendency toward undue burning or carbonization of the surrounding material.

Modern screen rooms are the result of years of development. Barring accidents, they are practically trouble-free, and thus require only simple periodic checks for their maintenance.

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