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PROCEEDINGS OF THE IRE®

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CONTENTS

ADVERTISING SECTION

THE COVER—Outlined against the dramatic sweep of the Aurora Borealis, or northern lights, is a 200-megacycle radar antenna used in auroral propagation research. Since the discovery 18 years ago that vhf radio waves are scattered by the aurora, the phenomenon has been investigated by scientists and radio amateurs in several countries, who have observed auroral echoes ranging from 35 mc to 220 mc and at heights estimated at 50 to 70 miles above the earth. This and other vhf and uhf ionospheric effects is the subject of the second of two review papers that open the issue, the first being on the important topic of heat transfer in electronics.

Photo—*Geophysical Institute, University of Alaska*

Scanning the Issue

Review of Industrial Applications of Heat Transfer to Electronics (Kaye, p. 977)—Continuing its series of invited review papers, the PROCEEDINGS presents this month a survey of a subject which has its roots in the mechanical engineering field but which in the last few years has become a prime consideration in the design of a wide variety of electronic equipment and devices. The recent upsurge of heat transfer as a science as well as an art has stemmed to a large extent from the unusual demands placed on the size, operating temperature, etc., of electronic equipment by extremely high-speed high-altitude aircraft and missiles. This paper reviews the various forms of heat removal, *i.e.*, conduction, radiation, natural convection, forced convection and evaporative cooling; presents four well-chosen examples of specific applications of heat transfer to the design of electronic components and equipment, and includes a liberal number of references to other published works on the subject. Novice and initiate alike will find this paper well worth reading.

Review of Ionospheric Effects at VHF and UHF (Little, *et al.*, p. 992)—In the past our interest in the ionosphere has been confined mostly to frequencies below 30 mc where its effect on the propagation of radio waves is very pronounced. Above 30 mc it was assumed that propagation was almost entirely tropospheric. Ionospheric effects above 30 mc, although less evident than at lower frequencies, nevertheless exist. These effects include refraction, reflection, absorption, change of polarization, scatter and diffraction, and, as is now realized, in some cases they can be significant. A good case in point is the recently discovered phenomenon of forward scatter of radio waves by the ionosphere in the 30 to 60 mc range, which makes possible reliable vhf communication at distances of 1000 miles or more. In addition to instances involving ground-to-ground communication, more and more antennas are now being pointed skyward to observe various phenomena in and beyond the ionosphere—making radar soundings of the aurora, meteor trails and the moon, and picking up radiations from the sun and stars (to which we must soon add radio tracking of earth satellites). These recently developed techniques have greatly increased our interest in the ionosphere at vhf and uhf and, at the same time, have provided us tools for learning more about it. This paper presents a comprehensive summary of what is now known of ionospheric effects in the 30–3000 mc range, covering radar echoes from aurora, meteors and the moon, radio noise from aurora, absorption and refraction of radio waves in the ionosphere, and the scintillation of radio stars. The subject of forward scatter is omitted, since this was treated exhaustively in the October, 1955 special issue of the PROCEEDINGS. There remains, however, an excellent review of a relatively new and important frontier of radio science, generously supplemented by references to 182 original papers in this field.

Directional Channel-Separation Filters (Cohn and Coale, p. 1018)—An important new class of frequency-selective networks is described which combines the properties of directional couplers and conventional filters. These "directional filters" consist of four-arm networks in which one arm is always isolated from the input arm and the other two arms perform respectively as band-pass and band-rejection filters with complementary frequency responses. They are very well suited for combining or separating signals of different frequencies in communication systems and for multiplexing several equipments to a single antenna. The many new directional-filter circuits described for the first time in this paper include circuits in the form of strip lines, coaxial lines, waveguides and lumped constants. This is a valuable piece of work which will lead to the design and application, especially at microwave frequencies, of a wide variety of filters embodying the principles revealed here.

A New Technique for the Measurement of Microwave Standing-Wave Ratios (Macpherson and Kerns, p. 1024)—The conventional moving-probe technique of standing-wave measurements is extremely simple to use and understand as long as we can neglect the effect of reflections caused by the slot and by the moving probe. When in the interest of greater precision it becomes necessary to take these reflections into account, the equation which must be used to give a corrected value for the standing-wave ratio becomes very complicated. In the new method presented in this paper, the detector is held stationary and a sliding-load technique is employed to vary the phase of the load, yielding a characteristic equation that is simpler to use than in the moving-probe case and presenting a number of advantages over previous phasing methods. This technique will be of especial interest to those involved in standards or other high precision work.

Novel Circuit for a Stable Variable Frequency Oscillator (Makow, p. 1031)—An important problem in designing variable frequency oscillators is the matter of frequency drift due to aging and to changes in temperature, humidity, CO_2 content and pressure of the atmosphere. Crystal control cannot be applied directly since the oscillator could not then operate over a continuous range of frequencies. In this paper the author has developed a novel circuit in which crystal control is applied indirectly to reduce frequency drift to a considerable degree. The scheme employs a multiloop feedback circuit which sustains three oscillations at three different frequencies in such a way that two of the frequencies, which are variable, always add up to the third, which is precisely and rigidly fixed by a quartz crystal. Thus the two variable frequencies, although they are free to change in equal and ***opposite*** directions because the sum of the frequencies would not be changed thereby, are prevented from changing in the ***same*** direction by the fixed sum frequency. Since the two frequencies would in most cases try to drift in the same direction, this circuit reduces to a substantial degree an important cause of frequency instability in vfo's.

IRE Standards on Electron Devices: TR and ATR Tube Definitions (p. 1037)—Some two score terms relating to TR and ATR tubes are authoritatively defined in this IRE-approved standard.

IRE Standards on Methods of the Conducted Interference Output of Broadcast and Television Receivers in the Range of 300 KC to 25 MC (p. 1040)—This long-titled IRE standard is a supplement to a similar standard issued in 1954 on methods measuring interference outputs in the 300 kc to 10 mc range. The present standard extends that range to 25 mc. This involves a revision in the characteristics of the line impedance network, which is the subject of this supplement to the earlier standard.

Some Limiting Cases of Radar Sea Clutter Noise (Schooley, p. 1043)—A major limiting factor in the ability of an airborne or shipboard radar to pick up a target on or near water is the amount of transmitted energy returned to the radar by the surface of the water. If the sea is rough the return may be increased to the point where it masks the target echo completely. This paper investigates two limiting cases of sea roughness, namely, perfectly smooth and perfectly rough. These two extremes are not met in actual practice but they do suggest the maximum and minimum effect of actual sea-clutter noise and thus contribute to our understanding of this important problem.

C. Frederick Wolcott

DIRECTOR, 1956–1957

C. Frederick Wolcott was born January 11, 1907 at Anderson, Indiana. For a while he attended Northwestern University, and in 1925–1926, Mr. Wolcott came into possession of one of those now-extinct tickets, an unlimited first class radio operator's license, and was Chief Engineer of Broadcast Station WGES in Chicago. He was a member of the Revell-Field Museum Expedition to Alaska, 1927, as radio operator of WNBE. In 1928 he was a consultant to the Coyne Electrical School, giving a series of lectures which were the basis for the formation of their radio division. From 1928 to 1931 he was chief engineer of the L. S. Gordon Co., Chicago. In 1931 he was an instructor at Coyne Electrical School, and he was an engineer for the Transmitter Division of deForest Radio Co. during 1932–1933. He was in charge of radio engineering for Arvin Industries, Inc. from 1933–1938, and during 1938–1939 he held a similar position with Thos. L. Siebenthaler Mfg. Co.

Since 1939, Mr. Wolcott has been successively chief television engineer, chief engineer, assistant to the president, and Technical Director of Gilfillan Bros., Inc., Los Angeles, California.

Mr. Wolcott is a registered professional engineer and a member of the Society of Automotive Engineers, the Academy of Television Arts and Sciences, and the Society of Television Engineers of which he was twice president. He held membership on Panels 3 and 7 of the National Television System Committee, and represented STE on the Radio Technical Planning Board.

Mr. Wolcott is an author and lecturer on television topics with a number of patents in the electronic field, issued and pending.

He is a Senior Member of the IRE. Active in the formation of the Indianapolis Section in 1936–1937, he became its chairman in 1938. He was a member of the Receiver Committee 1946–1949. He is a member of the Nominations and the Policy Advisory Committees.

Mr. Wolcott held the following offices in the Los Angeles Section: Secretary, 1941; Chairman, 1942; Chairman, Nominating Committee, 1946; Chairman, Nominating Committee, 1951; Member, Nominating Committee, 1952; Member, Advisory Group to Program Committee, 1952; Chairman, Section Relations Committee 1954–1956; Member, Executive Committee, 1954–1956. He received the 1954 Section Recognition Award.

In 1948 Mr. Wolcott handled liaison between IRE-WCEMA for the West Coast Convention; he was in charge of registration for the 1950 West Coast Convention; Chairman of the Advisory Group on WESCON, 1953; and 1954 Convention Vice-Chairman. He has been appointed to the WESCON Board for the term 1954–1957, to serve as Chairman of the Board of Directors for 1956.

Poles and Zeros

Growth. As anticipated here last month, the IRE membership roster has just passed the 50,000 mark. As of July 1st, in fact, the Institute comprised 50,738 dues-paying members. The passing of this milestone is worthy of more than casual notice, because it emphasizes the impressive position that electronics and the allied arts have taken among the professions and underscores their importance to society at large.

So far as can be determined from circulation figures published by other societies, the IRE is now the second largest association of engineers and scientists, second only to the American Chemical Society. The latest figures (December, 1955) place the paid membership of our esteemed sister societies as follows: American Chemical Society, 77,121; American Institute of Chemical Engineers, 14,391; American Institute of Electrical Engineers, 47,900; American Institute of Physics, 16,946; American Society of Civil Engineers, 35,210; American Society of Mechanical Engineers, 35,266; Institute of Aeronautical Sciences, 12,435.

Of even more significance than the present level of IRE membership is the growth trend. Since 1952 the net annual increase in membership has averaged 12.5 per cent; last year it was 13.4 per cent, and the rate this year thus far is 15 per cent. If our growth continues to be compounded as it has since 1952, in fact, we can expect to pass the 100,000 mark by 1962. This projection is not so startling as it might appear, since we have in fact more than tripled our membership in the past decade.

Continued growth is not guaranteed by mere extrapolation of statistics, of course. History shows that professional societies have not escaped the heavy hand of economic catastrophe. During the depression, from 1931 to 1935, the IRE membership dropped from 6,700 to 4,500 and did not recover this loss until 1941.

Quite possibly a similar "lost decade" may occur in the future, but there are strong reasons for doubting it. First, we have better mechanisms for preventing economic blackouts than we did in 1929. Second, the character of our profession and its impact on every phase of life have broadened to an extent that could not have been foreseen in the pre-war decade. When one segment of our industry suffers an economic setback, two others appear to take up the slack. All considered, it does not take much foresight to predict that the IRE membership will be 100,000 strong before 1970, even if military support of electronics does not continue at its present level. Applications of electronics to industry, commerce and the home, based on developments as yet dimly seen, will provide the base.

Just about the only uncertainty in the future growth pattern, as we see it today, is whether the IRE organization will continue to provide a natural habitat for so large and diverse a group of technologies as is implied in a membership of 100,000. Here, again, there is little cause for concern because the answer is already with us in the Professional Group system. The PGs have an impressive growth pattern of their own. In 1952 there were 12,482 paid PG memberships spread among fifteen Groups; today there are 47,243 paid memberships in 24 Groups, and the number has increased 10,000 in the past six months! Moreover there is still a good way to go, since each PG member belongs, on the average, to two groups. This implies that half of the IRE members have not yet joined a Group. The expansion possible within the PG system is terrific. After all, if every IRE member joined all the Professional Groups (Heaven forbid!) we'd be printing over a million copies of PG TRANSACTIONS each year. If even a small part of this potential is realized, the Institute will not only continue to grow, but will continue to contain its parts in a harmonious whole.

Reminders. In the prosecution of editorial affairs, the uneasy thought occurs that many of our input and output termini, authors and readers, are not aware of the existence of certain aids in the preparation and utilization of IRE publications. For example, at a recent meeting of the Editorial Board, it turned out that five out of the seven members were unaware that there is on hand a Cumulative Index for 1947–53 which covers not only the PROCEEDINGS, but all the then-published TRANSACTIONS and the CONVENTION RECORD as well. This covers an extraordinarily active period in electronic technology and is a correspondingly useful document. It sells for $1 to members, $3 to nonmembers.

Similarly, in the November 1954 issue of the PROCEEDINGS, *Information for Proceedings Authors* was printed. This was a well-thought-out statement of how to make a technical paper effective, and it should be read by every author at the rough-draft stage. But the November, 1954 issue is, by actual measurement, sixteen inches below the top of the pile of the PROCEEDINGS and Directories since published and the chances of its being pulled out are correspondingly slim. To make things easy, reprints are available, without cost, on request to the Editorial Department. A supply will gladly be sent to any person, editorial committee or other organization having responsibility for procuring and clearing technical papers. Pass the word along to the man in your organization.—D.G.F.

Review of Industrial Applications of Heat Transfer to Electronics*

JOSEPH KAYE†

The following paper is one of a planned series of invited papers, in which men of recognized standing will review recent developments in, and the present status of, various fields in which noteworthy progress has been made.—*The Editor*

Summary—**Applications of heat transfer to electronic components and devices are presented and discussed. The major objectives are to review the state of the art in such applications and to indicate the need for a better grasp of the science of heat transfer in aiding the creative engineering of new electronic devices operating at extreme conditions of temperature, heat flux, air density, etc.**

The recent rapid growth of heat-transfer applications to electronics is analyzed in terms of the changing specifications resulting from the introduction of the supersonic and hypersonic aircraft and missiles.

The general thermal problem of a given piece of electronic equipment is discussed and analyzed in terms of different modes of heat transfer. A simple method of comparing heat removal by means of natural convection, radiation, forced convection, and evaporative cooling is presented in chart form. The design philosophy of heat-transfer applications to industrial electronic equipment is discussed.

Four specific illustrations of industrial applications of heat transfer to electronic components and equipment are presented. The first describes use of heat transfer and fluid mechanics to predict accurately the thermal performance of a newly developed vacuum tube which dissipates 25 kw. The second presents the analysis and some samples of the cold-plate technique used to cool highly compact miniaturized electronic equipment. The third example discusses the recent development of high temperature vacuum tubes, which can operate reliably at ambient temperatures in excess of 250°C. The final illustration discusses evaporatively-cooled magnetic components, in particular, transformers, by means of fluorochemical dielectrics.

The educational aspects of the problem of training and providing needed manpower are also discussed.

Symbols

a = length of radial fin (10).
A = heat-transfer area.
A_x = cross sectional area for fluid flow.
A_0 = base area of tube.
A_f = area of extended surfaces or fins.
b = thickness of radial fin (10).
c_p = specific heat at constant pressure.
d_1, d_2 = dimensions of cooling duct (18).
D_e = equivalent diameter ($4\ r_H$).
f = friction coefficient.
G = mass velocity (w_a/A_x).
g_c = acceleration given to unit mass by unit force.
h = heat-transfer coefficient.
K_c, K_e = entrance and exit loss coefficients for flow in and out of heat exchanger (12).
k = thermal conductivity.
L = axial length of fin, in air-flow direction.
n = number of tubes or ducts in heat exchanger.
Pr = Prandtl number ($c_p\mu/k$).
p = pressure.
Δp = pressure drop through heat exchanger.
q = rate of heat transfer.
r_H = hydraulic radius of duct ($4A_x$/perimeter).
r = aspect ratio of duct (d_1/d_2).
Re = Reynolds number (D_eG/μ).
St = Stanton number (h/c_pG).
T = temperature (°Fahrenheit absolute).
t = temperature (°Fahrenheit).
Δt = temperature difference.
t_a = temperature of air.
t_f = temperature of fin.
t_p = temperature of plate.
w_a = mass rate of flow of air.
ϵ = emissivity.
μ = viscosity.
ρ = density.
ρ_s = standard density.
η_f = fin effectiveness.
σ = ratio of cross sectional area of header to that of duct.

Objectives

THE OBJECTIVES of this review are to examine the state of the art in applications of heat transfer to electronic equipment, to discuss some recent applications, and to indicate the need for a better grasp of the science of heat transfer in the creative engineering of new devices and in the extension of the thermal range of operation of many different types of electronic components.

Introduction

The art of heat transfer has been known and applied by man since antiquity. The science of heat transfer, however, has been rather slow in development. This science has grown and expanded mainly in the last half century until today it is an integrated structure of logic and experiment based on the principles of thermodynamics, fluid mechanics, electromagnetic radiation,

* Original manuscript received by the IRE, April 16, 1956.
† Prof. of Mech. Engrg., Dir., Res. Lab. of Heat Transfer in Electronics, M.I.T., Cambridge, Mass.

etc. The applications of this science to the creative design of electronic equipment have become of vital importance today.

The engineering literature before 1930 contains only a few applications of this science to electrical machinery and electronic devices. From 1930 to 1950, the number of applications of heat transfer in the design and in the safe operation of electronic equipment increased slowly. The number found in the literature since 1950 has increased exponentially. A recent survey of the unclassified literature,[1] covering the period from 1950 to 1954, disclosed several hundred references on applications of heat transfer to electronic equipment and electrical machinery. It is evident from an examination of the papers listed in this incomplete survey that the science of heat transfer has been utilized as an important tool for the design and reliable operation of electronic equipment and electrical machinery.

What has caused this exponential increase in the number of applications of heat transfer to electronic equipment? Several factors are discovered quite easily, but most of them stem from a single major cause. This has been the introduction of supersonic and hypersonic aircraft and missiles, whose high speed requires extremely high rates of reaction and control, thus making them almost completely dependent on electronic systems for reliable operation during a given mission. Furthermore, these high speed devices operate more efficiently with respect to fuel consumption at high altitude. These devices inevitably require a significantly larger fraction of their total weight to consist of electronic equipment than do low speed aircraft. All these factors can be summarized by listing the new conditions facing the designer of airborne electronic equipment as follows:

1) The electronic equipment must be miniaturized but still must not exceed safe levels of temperature for reliable operation.
2) The equipment must operate reliably in environments of higher temperatures caused by the aerodynamic heating at supersonic speeds. Fig. 1, borrowed from a previous study,[2] shows the order of magnitude of temperatures to be expected in airborne devices moving at supersonic speeds up to Mach 10 at an altitude of 50,000 feet.
3) The equipment must operate reliably at the lower air densities found at the higher altitudes.
4) The electronic equipment should not be cooled piecemeal because of the high cost, in terms of weight and volume, required by the distribution of separate coolant systems for different items of equipment. On the contrary, the electronic equipment, power sources, auxiliaries, cabin, etc., should be cooled by a properly designed centralized cooling system of high effectiveness with a minimum penalty in terms of weight and volume.

Other factors accounting for the growth of heat-transfer problems in electronic equipment can undoubtedly be found but are of less importance, on the whole, than the factors listed above. These secondary factors arise from the growing awareness in the electronics industry that the science of heat transfer can be used to predict answers in the design and operation of diverse equipment, and thus eliminate the need for expensive trial-and-error type of engineering solution.

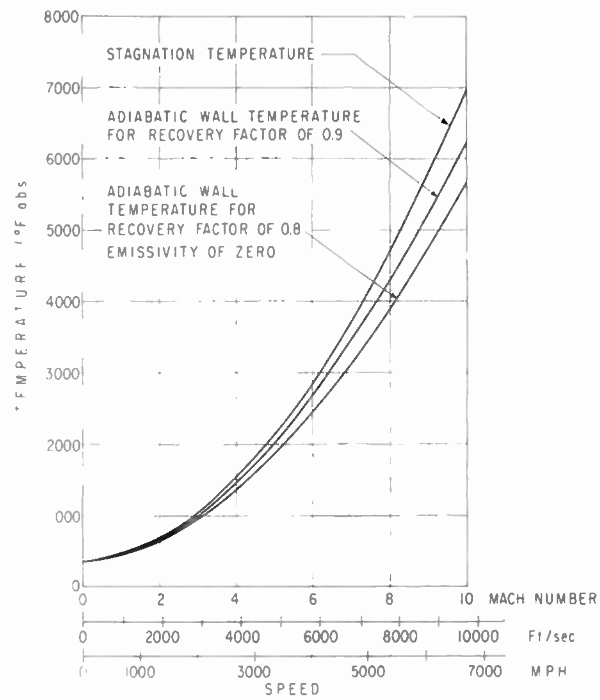

Fig. 1—Stagnation and adiabatic wall temperature for flight at constant speed in air at 50,000 feet altitude.

General Thermal Problem

All electronic devices operate with "losses" which are dissipated as heat. From the thermodynamic viewpoint, all electronic devices undergo irreversible processes which result ultimately in the generation of heat and usually in a significant increase of temperature of the device over the temperature of the surroundings. This temperature difference, from device to environment, is the driving potential for the heat-transfer rate required to dissipate the losses to a thermal sink. Thus Fig. 2 shows schematically an electronic device receiving energy input, which may be electrical, magnetic, mechanical, etc., in nature, and delivering useful energy output in different forms. The losses, which for many items may range from 95 to 50 per cent of the energy input, are rejected to a heat sink.

Several thermal problems arise in considering the device in Fig. 2 in the design stages and also in the utilization stages. In the former, the problem is to design the

[1] J. Kaye and M. A. Rives, Jr., "Bibliography on Thermal Effects in Electronic Equipment and in Small Electrical Machines" Rep No. RLHTE-9, M.I.T.; May 12, 1955.

[2] J. Kaye, "The transient temperature distribution in a wing flying at supersonic speeds," *J. Aero. Sciences*, vol. 17, pp. 787–808; December, 1950.

thermal characteristics within and without the device so that it will operate reliably in accordance with its functional design, or briefly at its electronic rating. Usually only elementary thermal considerations are required if the device operates at one extreme condition, namely, at a low level of heat flux with a large available external surface area, but exceptions are known today even for this extreme condition. If, on the other hand, the device operates at very high levels of heat flux with limited amounts of external surface area, the thermal design often becomes the major design problem of the device. Mouromtseff[3] in his well-known 1942 paper opens with: "Without exaggeration one may state that in designing electronic tubes there are many more mechanical, metallurgical, and heat engineering problems than those of pure electronic character. One may also admit that quite frequently nonelectronic problems are solved by the cut-and-try method rather than by calculation." These statements are still valid today, in that the final economic design or, for military applications, the final reliable design, of many devices is limited by nonelectronic considerations, such as the thermal problem.

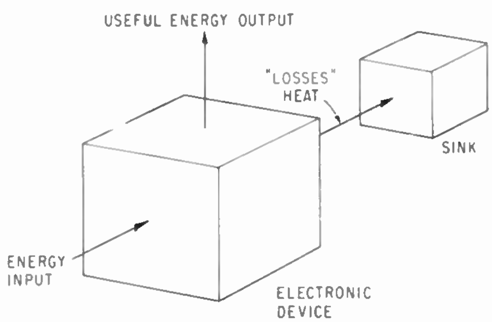

Fig. 2—Functional representation of thermal problem in electronic equipment.

When the electronic device is employed in a given piece of equipment, thermal problems can arise in two distinct ways. First, the device may be utilized in a normal environment which is compatible with the intent of the designer in regard to its cooling for safe and reliable operation. Second, the device may be used in an abnormal environment which may be incompatible with the specifications of the designer, such as excessively high environmental temperatures or very low density of cooling air. If the level of heat flux is high enough, then both methods of utilization may require careful design of the package in order to achieve reliable operation of the components without derating them.

Methods of Heat Transfer

The general thermal problem can be solved by use of one or more different types of heat transfer. Space does not permit a detailed discussion of these various methods of heat transfer so that only a general comparison will be attempted here. The black box or electronic device in Fig. 2 may have its losses removed by heat-transfer methods such as conduction, radiation, natural convection, forced convection, evaporative cooling, or by a combination of these.

A fairly simple comparison can be made on a single chart of all these heat-transfer methods except for conduction. Although conductive cooling is important for many applications, it is perhaps best understood since it is analogous to conduction of electricity, and it can be omitted here in this particular comparison. Consider the methods of cooling the electronic device in Fig. 2 for the simple case where it has the shape of a circular cylinder, one foot in diameter, infinitely long, and exposed to cooler ambient air. The significant parameters are then as follows:

1) Temperature of ambient air.
2) Temperature rise of the cylindrical surface over ambient air.
3) Heat flux.
4) Velocity of ambient air.
5) Density of ambient air.
6) Emissivity of surface of cylinder.

Figs. 3 and 4 have been constructed to illustrate the effects of large variations in the values of these parameters, using log-log scales.

Cooling in an Ambient of 70°F

Natural Convection

Consider first the cooling of this cylinder at sea-level conditions of one atmosphere pressure and an air temperature of 70°F, as shown in Fig. 3. If cooling occurs only by natural convection, *i.e.*, if no fan or blower forces the air to circulate around the cylinder, then the curve marked Natural Convection at Sea Level shows that a temperature rise of about 700°F is needed to reject a heat flux of 2 watts/inch.^{2} This curve is a straight line on the log-log diagram.

Radiation

If only radiative cooling were used, and the maximum emissivity of 1 were assumed, then the curve marked Radiation shows that a surface temperature rise of about 350°F would be required for a heat flux of 2 watts/inch^2. If the cylinder were cooled by the combination of both natural convection and radiation, then the surface temperature rise would be reduced to about 200°F for a heat flux of 2 watts/inch^2. Of course, if the emissivity were reduced from 1 to 0, then the combined effects of cooling by natural convection plus radiation are reduced to that of cooling only by natural convection. Hence for values of emissivity between 0 and 1 the temperature rise will lie between the curve shown for natural convection and that shown for radiation plus natural convection.

[3] I. E. Mouromtseff, "Water and forced-air cooling of vacuum tubes," Proc. IRE, vol. 30, pp. 190–205; April, 1942.

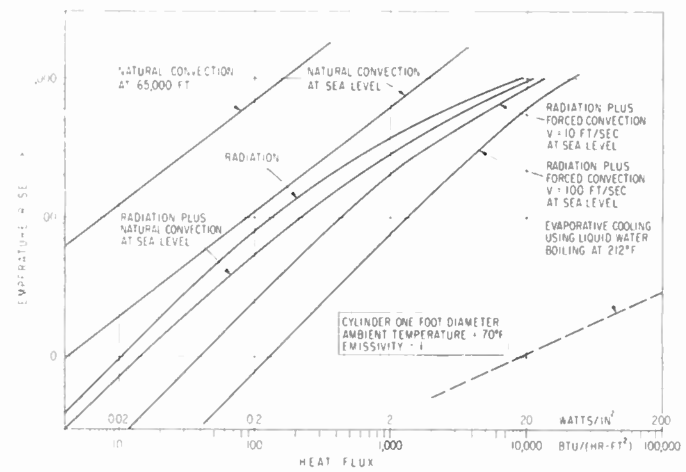

Fig. 3—Temperature rise vs heat flux.

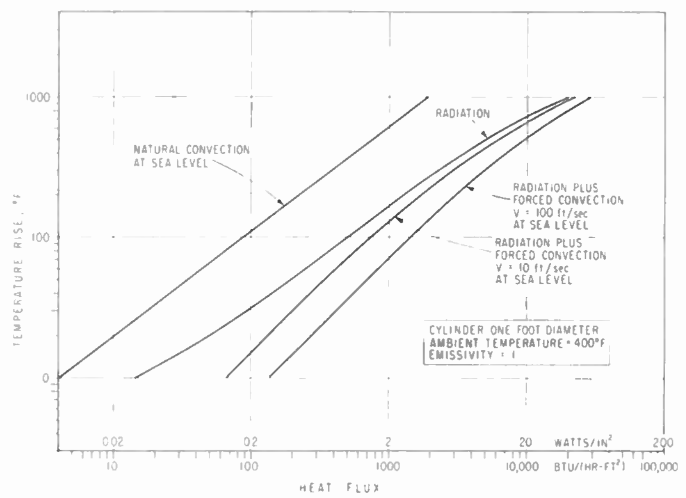

Fig. 4—Temperature rise vs heat flux.

Forced Convection

If the cylinder is cooled by forced convection by blowing air at sea-level conditions normal to its axis, and if radiation is used simultaneously, the results for air velocities of 10 and 100 feet/second are also shown in Fig. 3. For a velocity of 100 feet/second a temperature rise of only about 80°F would be required for a heat flux of 2 watts/inch². But such a reduction in temperature rise is not attained at small cost, since the air-pressure drop and power required to force the air to move at this speed are by no means negligible in cost, whether expressed in terms of dollars, or weight, or volume of auxiliary equipment. Hence the effectiveness of cooling the cylinder by means of forced convection must be determined by several considerations rather than by only the temperature rise of the cylinder surface.

Evaporative Cooling

Consider finally the removal of heat from the cylinder by means of evaporative cooling. Although the dashed curve labelled Evaporative Cooling Using Liquid Water Boiling at 212°F, is also shown in Fig. 3, its interpretation must be considered differently from that for the other curves. This curve refers to a temperature rise equal to cylinder-surface temperature minus 212°F; *i.e.*, the ordinate at a value of 10°F means that the cylinder surface is at 222°F if the water is boiling at 212°F and at 1 atm. Hence a temperature rise of only about 5°F is required for a heat flux of 2 watts/inch². Furthermore, this type of evaporative cooling is an extremely important method of heat transfer which offers the possibility of removing 200 watts/inch², with a reasonable value of the temperature rise, since all the other types or combinations shown will require a temperature rise in excess of 1,000°F for this large heat flux. Since evaporative cooling can be used for extremely large values of heat flux, this method of heat transfer appears attractive for many future applications in the fields of electronic devices and electrical machinery.

Low Density

Fig. 3 shows the effect of cooling by natural convection if the ambient pressure is reduced from atmospheric at sea level to the low density air at 65,000 feet altitude for the same ambient temperature of 70°F. At an air density corresponding to 65,000 it is seen that a temperature rise of greater than 2,000°F is required for a heat flux of 2 watts/inch²; this curve indicates that only natural-convection cooling cannot be used for present-day equipment at high altitudes.

Cooling in an Ambient of 400°F

Fig. 4 shows the calculated results for the same cylinder, as in Fig. 3, except that the ambient temperature is fixed as 400°F. For high speed devices this value represents temperatures attainable today whereas considerably higher temperatures will be reached in the future. The curve for natural-convection cooling shown in Fig. 4 is the same as that shown in Fig. 3; this follows since the temperature rise depends mainly on the coefficient of heat transfer for natural convection, and this in turn depends mainly on the temperature difference, not on the temperature level.

The curve for radiation, for maximum emissivity of 1, shown in Fig. 4, indicates that a smaller temperature rise is required than in Fig. 3, since the rate of heat transfer by radiation varies as the fourth power of the level of the absolute temperature. Hence the cylinder will require a temperature of about 200°F above the ambient of 400°F, or a surface temperature of about 600°F to dissipate 2 watts/inch² by radiation plus natural convection. This value of 600°F is in the range of the new ceramic vacuum tubes designed for operation at high temperatures and cooled by natural convection plus radiation; some additional information on these new tubes will be given later.

For cooling by forced convection, with an air velocity of 100 feet/second, and simultaneously by radiation,

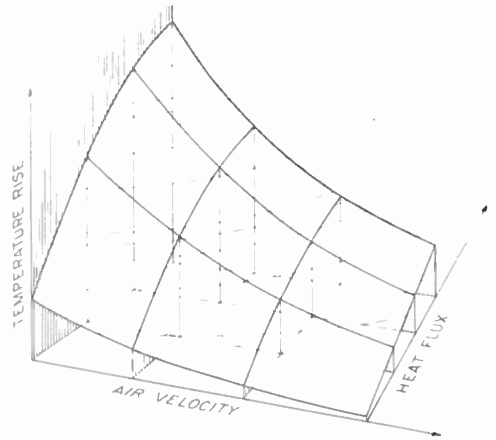

Fig. 5—Surface illustrating relation between temperature rise, heat flux, and air velocity.

the temperature rise of the cylinder is about 70°F above the ambient of 400°F for a dissipation of 2 watts/inch². Hence it is evident that electronic components can be cooled under high temperature ambient conditions by forced convection if one is willing to pay the price of pump weight and power required to provide the necessary pressure drop of the cooling air. Heat could be removed from the cylinder under discussion by use of evaporative cooling, under the conditions of high ambient temperatures, but such a curve is not shown in Fig. 4 for lack of data.

Illustrative Surface

The calculations and curves shown in Figs. 3 and 4 are limited to the specific geometry of a cylinder. It is evident, however, that many different geometrical configurations will be encountered in practical applications. Excellent reference books are available today, such as McAdams,[4] which contain a wealth of heat-transfer data and correlations; these permit the computation of curves similar to those shown in Figs. 3 and 4 for other geometries and other flow arrangements.

In order to illustrate the relation between the important variables of temperature rise, heat flux, and ambient air velocity, a three-dimensional surface has been sketched schematically in Fig. 5 above. Dashed curves refer to lines of constant temperature rise. This surface shows the great reduction in temperature rise obtained by increasing air velocity for a fixed heat flux and the rapid increase in temperature rise obtained by increasing heat flux for a given air velocity. This surface illustrates the importance of air velocity for many different kinds of geometrical configurations. However, this surface does not depict the important simultaneous effects of air velocity on required pump power to deliver the necessary pressure drop of the air.

[4] William H. McAdams, "Heat Transmission," McGraw-Hill Book Co., New York, N. Y., third ed., 1954.

Design Philosophy of Heat-Transfer Applications

The general thermal problem in electronic equipment has been presented together with a discussion of various methods of heat transfer which can be utilized for its solution. Many possible solutions exist for each given problem. An important question which arises often in practical industrial problems is whether an optimum design exists. The criterion for an optimum thermal design can be based on familiar parameters, such as minimum yearly cost for the sum of fixed charges and operating charges, or on minimum weight, or on minimum volume, etc. Relatively few pieces of electronic equipment have been constructed on the basis of an optimum thermal design!

The design philosophy which is employed in a given situation depends on the type of heat-transfer problem. In this connection, three fairly simple categories of problems can be distinguished, as follows:

1) Problems where large quantities of surface area or large amounts of inexpensive coolants are available. Either small or large values of heat flux are possible.
2) Problems involving relatively large values of heat flux where surface area, coolants, weight of auxiliaries, etc., are all considered as "expensive" in an over-all sense.
3) Problems involving new engineering efforts in the direction of extremely large values of heat flux or of extremely high ambient temperatures (such as 500°C), or of significant creative efforts in the art of energy conversion where the thermal problems are, perhaps, the most difficult to solve in terms of available materials.

Considering both categories 1) and 2), the design philosophy of the solution of the thermal aspects of these problems is well known. Starting at the bottom of the scale of values of heat flux, the designer's philosophy is to depend on conduction, natural convection, and radiation to transfer the heat to the walls of the electronic package and thence to the outside air. Since the level of heat flux is assumed to be low, sufficient surface area is usually available to alleviate the thermal problem. Consider next an increase in value of the heat flux by one or two orders of magnitude, caused either by an increase in heat dissipation for the same volume of package or by a decrease in size for the same dissipation. Usually forced convection is used, in the form of one or more fans or blowers, to prevent overheating of local components or formation of hot spots in the package. This solution of blowing air randomly in the package is a reasonable solution up to certain limits of heat flux. This solution might work satisfactorily at sea-level conditions but perhaps not at high altitude; it might work for certain ambients but perhaps not for the highest temperatures encountered in military specifications.

Consider now that the value of the heat flux is increased a second time by one or two orders of magnitude, and that we are dealing with problems in category 1) above. For this case, where the coolant is plentiful and inexpensive, the philosophy of the designer is to execute a thermal design which is still fairly simple and is not critical, since an overprotected thermal circuit is fairly easily attainable. However, for category 2), where the coolant is expensive and where weight must be minimized, the philosophy should be to use utmost care in an attempt to attain an optimum thermal design. The coolant flow path must then be investigated in detail for both effectiveness and cost of pumping power. Many solutions should be obtained and studied in order to achieve an optimum thermal design.

Consider problems in category 3) where extremely large values of heat flux may occur. In order to operate the device safely at its electronic rating, the solution of the thermal problem may become the most important item in the creation of the device. An example will be given later illustrating this view when the design of a new electron tube dissipating 25 kw will be discussed. A similar design philosophy applies to creation of new electron devices designed to operate safely at extremely high ambient temperatures, such as in the development of ceramic vacuum tubes with surface temperatures near or greater than 500°C; in this instance the thermal problem should be solved in detail, and should be near the optimum solution for reliable operation. Finally the introduction of novel energy conversion devices produces challenging thermal problems which require a fresh approach for their solutions. Thus the introduction of the miniature transistor seems to have accomplished a large reduction in the size of electronic packages until one becomes aware of the temperature limitation of this device. Furthermore the thermal problems of a transistor are greatly aggravated when a miniature power transistor is utilized since the thermal paths from the collector and emitter junctions to the heat sink introduce significantly important difficulties in its design and usage.

Illustrative Applications of Heat Transfer to Electronic Devices

In the final portion of this summary, several illustrative examples of recent applications of heat-transfer techniques and applications to electronic equipment will be discussed. Since only a few such cases can be considered here for reasons of space, the coverage of this field is not intended to be complete. However, additional cases of interest will be mentioned and references given. The illustrative examples have been selected to emphasize those problems where the creative engineering development required fairly detailed and intensive applications of the art and science of heat transfer. The following list summarizes the examples considered here.

1) Development of an air-cooled vacuum power tube with a heat dissipation of 25 kw.
2) Development of cold-plate techniques for cooling electronic equipment.
3) Development of vacuum tubes to operate with surface temperatures of about 500°C.
4) Development of gaseous and liquid filled magnetic components using evaporative-cooling techniques.

1) Development of a 25 kw Vacuum Tube

One of the best illustrative examples available in the unclassified literature of the application of heat-transfer techniques to electronic devices is the recent paper by London.[5] The basic problem was to design an adequate air cooling system for a new type of vacuum tube dissipating 25 kw. The emphasis was placed on analytical derivation of the cooling performance by use of basic data from heat transfer and fluid mechanics and the comparison of the predicted results with experimental test results. The optimization of the design was not emphasized since this investigation was to be continued.

Fig. 6 (opposite) sketches the 25 kw electron tube, showing the heavy copper anode as the outer tube wall, the radial fins for extended-surface cooling, and the axial flow of cooling air in approximately trapezoidal passages formed by the fins, the anode wall, and the outer duct wall. Specifications for this tube included a maximum allowable inside wall temperature of 275°C, as shown in Fig. 6 for an air temperature of 49°C, and for a maximum heat dissipation of 25 kw. As an approximation, the basic data for heat-transfer rate and friction factor for fluid flow in the trapezoidal passages were taken as for a rectangular passage of equivalent aspect ratio, shown in Fig. 7.

The analysis for prediction of thermal performance and of fluid-flow characteristics can be summarized briefly as follows:

First Law:

$$q = w_a c_{pa}(t_{a2} - t_{a1}). \tag{1}$$

Definition:

$$a \equiv \eta_0 h(A_0 + A_f)(t_{f,\max} - t_{a2}). \tag{2}$$

Definition:

$$D_e \equiv 4A_x L/(A_0 + A_f). \tag{3}$$

Definition:

$$St \equiv h/c_p G. \tag{4}$$

Definition:

$$G \equiv w/A_x. \tag{5}$$

[5] A. L. London, "Air-coolers for high power vacuum tubes," Trans. IRE, vol. ED-1, pp. 9–26; April, 1954.

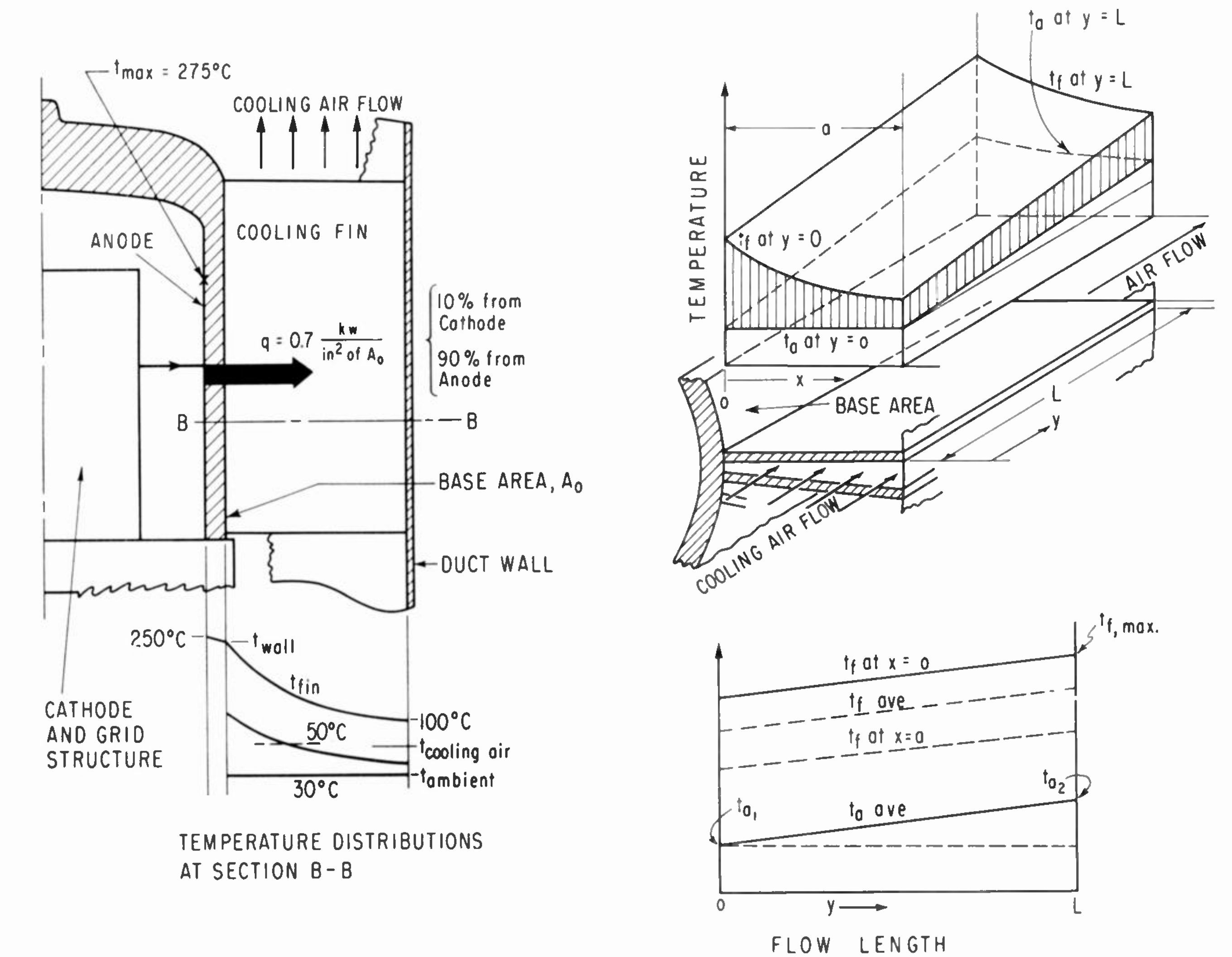

Fig. 6—Cooling of 25 kw tube, London.[5]

Definition:

$$Re \equiv D_e G/\mu. \tag{6}$$

Definition:

$$\text{C.P.} = q/(t_{f,\max} - t_{a1}) \equiv (hA)_{\min}. \tag{7}$$

If (1) to (7) are rearranged and solved for the "cooling power" as used by London,[5] which in essence represents a minimum thermal conductance or a specification which the tube designer desires to be built into this cooling system, the result is

$$\text{C.P.} \equiv (hA)_{\min} = \frac{w_a c_{pa}}{1 + \dfrac{1}{\eta_0}\left(\dfrac{D_e}{4L}\right)\left(\dfrac{1}{St}\right)}. \tag{8}$$

Hence (8) can be used to predict the thermal performance of this air-cooling system if data for fin effectiveness and if heat-transfer correlations for Stanton numbers are available for a wide range of laminar and turbulent flow conditions. The fin effectiveness can be

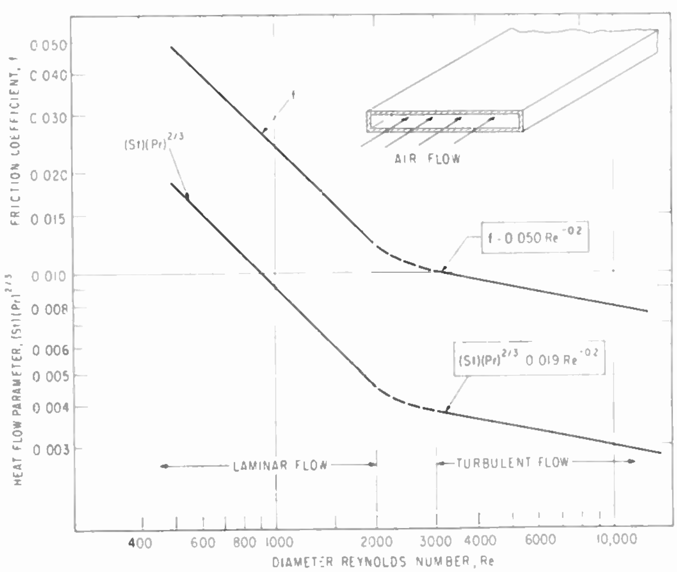

Fig. 7—Data for heat transfer and fluid flow in rectangular duct, London.[5]

obtained for this type of radial fins as follows:

$$\eta_0 = 1 - \frac{A_f}{A}(1 - \eta_f) \tag{9}$$

where

$$\eta_f = \frac{\tanh\,(a\sqrt{2h/kb})}{a\sqrt{2h/kb}}. \tag{10}$$

The heat-transfer correlations for Stanton number are taken from Fig. 7; thus for turbulent flow in this rectangular channel, the correlation is:

$$StPr^{2/3} = 0.019Re^{-0.2} \tag{11}$$

Hence (8) through (11) may be used to predict the variation of minimum thermal conductance required for a given flow rate of cooling air. The predicted results for the plain fin are shown in the lower portion of Fig. 8 by the heavy continuous curves, after adjustment was made for nonuniformity of the heat losses from the anode structure. The test results, carried out at 8.3 kw, agree within 15 per cent of the corresponding predicted results.

The prediction of the pressure-drop requirements for this air-cooling system depends on the following equation, derivable from Newton's second law and using various definitions of irreversible head losses:

$$\Delta p p_1/p_s = (G^2/2g_c\rho_s)\left[(K_c + 1 - \sigma^2) + 2(\rho_1/\rho_2 - 1) + (fL\rho_1/r_H\rho_m) - (1 - \sigma^2 + K_e)\rho_1/\rho_2\right] \tag{12}$$

where the first, second, third, and fourth terms in the bracket represent the effects of entrance-loss, air compressibility, wall friction, and exit loss, respectively. The friction-factor correlations are shown in Fig. 7 for laminar and turbulent flow. The values of the other parameters may be found in Kays, London, and Johnson,[6] as functions of area ratio and Reynolds number. Hence (12) with the necessary additional data permits the prediction of air-pressure drop and pumping-power requirements for this plain-fin air cooling system. The predicted results for pressure drop are shown in Fig. 8 by the heavy continuous curves for the two cases of zero heat loss and 25 kw heat loss from the tube. The pressure drop measured at a loss of 8.2 kw agrees within a few per cent with the predicted results over the entire range of air flows used in these tests.

The results obtained by London for the plain-fin cooler can be summarized as follows. For a specified minimum thermal conductance of 0.120 kw/°C, the plain-fin performance can be predicted accurately for engineering purposes; an air pressure drop of 7.9 inches of water and a pumping power of 4 hp would be required for cooling this tube.

[6] W. M. Kays, A. L. London, and D. W. Johnson, "Gas turbine plant heat exchangers—basic heat transfer and flow friction design data," ASME Res. Monograph; April, 1951.

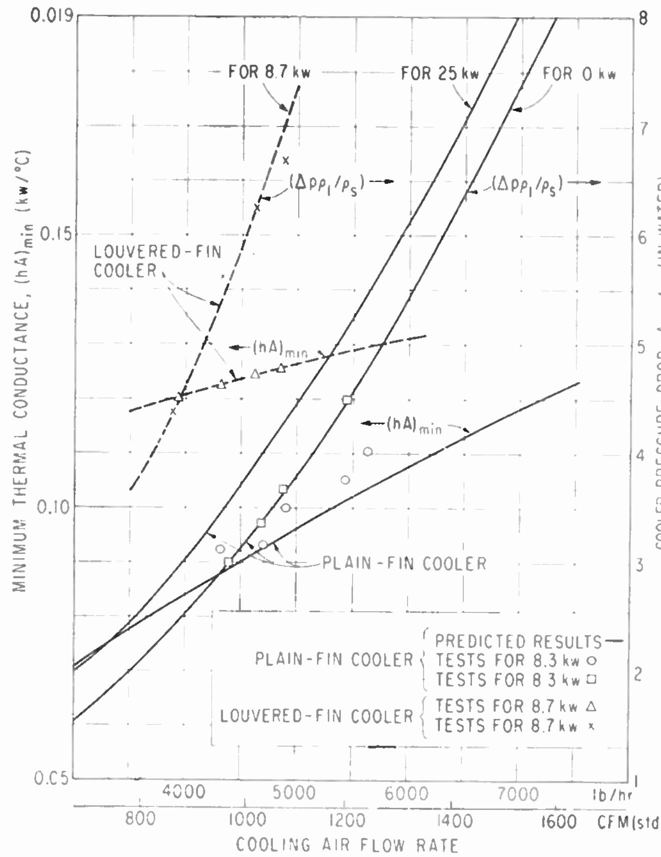

Fig. 8—Comparison of plain-fin cooler with louvered-fin cooler for 25 kw tube, London.[5]

One can sense intuitively that the above solution using plain fins is not an optimum design. London proposed to improve the design by replacing the plain fins with louvered fins in order to achieve two effects. One effect of the louvered fin would be to force more of the cooling air to flow closer to the fin root at the anode wall. A second effect would be to increase the Stanton number by periodically interrupting the boundary-layer flow. However, when the louvered fin is used, the understanding of the fluid-flow and heat-transfer phenomena became hazy so that one can not quantitatively predict the results based on the new flow conditions. Recourse to experiment led to the results shown by the dashed curves in the upper portion of Fig. 8. It is evident, for the same specified value of the minimum thermal conductance of 0.120 kw/°C, given above, that the performance of the louvered fin is considerably better than that for the plain fin. The air-pressure drop is now 5.2 inches of water and the air-pumping power is now 1.5 hp.

In concluding the discussion of this particular illustrative example, the following should be noted.

Where the fluid flow and heat transfer correspond to well-defined situations, accurate engineering predictions of performance of such cooling systems are possible based on the present science of heat transfer.

Optimization of the design of these cooling systems is possible but due regard must be given to the regions where intuitive solutions based on engineering judgment are likely to provide important results even though predictions can not be made easily.

Other examples of applications of heat transfer and fluid flow to vacuum tubes and related devices may be found in papers and reports by Bauermeister,[7] Beurtheret,[8] deBrey and Rinia,[9] Buckland,[10-13] Erdle,[14] Hickey,[15] Jenny,[16] Lemeshka and Nekut,[17] Matthews,[18] Mouromtseff,[19,3,20] Ostlund,[21] Schärli,[22] Young,[23] Ziefvert.[24]

2) Development of Cold-Plate Techniques for Electronic Equipment

During the past decade the cold-plate techniques of cooling miniaturized and subminiaturized electronic equipment have become highly developed and quite versatile. It will be recalled that at one extreme electronic equipment may be cooled by blowing or wafting air in a general random direction over the hot spots in a package; *i.e.*, for no particular ordered arrangement for the flow of cooling fluid. In contrast with this extreme, the cold-plate technique achieves an extremely well-ordered arrangement of both thermal paths and fluid flow paths. Usually the hotter heat generating components are carefully fastened to a chassis or plate by means of well-designed thermal conductors which maintain a finite but small temperature difference between hot spot and chassis, *i.e.*, the "colder" or more properly the "cooled" plate. The heat is then removed from this chassis or plate by forcing a coolant, air or liquid, to pass through well-defined channels or ducts which form part of the chassis for structural purposes. The net result is a highly ordered arrangement, *i.e.*, the cold plate, whose thermal performance and fluid-flow characteristics can be predicted accurately in most cases before construction of the electronic package.

The advantages of cold-plate techniques are numerous when the object is to design a unit which will operate successfully the first time it is constructed. These techniques are very successful for design of electronic items used in cramped quarters such as in missiles and similar gear; they are practical, lend themselves to easy fabrication, and are not expensive compared with other means of cooling expressed either on a dollar or on a weight basis.

An approximate analysis which can be used for rapid prediction of thermal performance and fluid-flow characteristics is summarized below to illustrate the usual case where the cold-plate temperature and cooling-air temperature both increase linearly with length of cooling duct. This simplified temperature distribution diagram is shown on the next page in Fig. 9(a), together with a sketch of the cold-plate dimensions in Fig. 9(b). In the following approximate analysis, the origin and restrictions for the various independent relations are adjacent to the relation in order to identify it. For this simplified case shown in Fig. 9(a), the local temperature difference from the plate to the air is constant throughout the cold plate.

Approximate Analysis for Turbulent Flow

First Law:

$$q = w_a c_{pa}(t_{a2} - t_{a1}). \tag{13}$$

Definition:

$$q \equiv hn2(d_1 + d_2)L\Delta t. \tag{14}$$

Definition:

$$\Delta t \equiv t_p - t_a = t_{p1} - t_{a1} = t_{p2} - t_{a2}. \tag{15}$$

Experimental correlation for turbulent flow ($Re > 5{,}000$):

$$hD_e/k = 0.023(D_eG/\mu)^{0.8}(c_p\mu/k)^{0.4}. \tag{16}$$

Definition:

$$D_e \equiv 4d_1/d_2/2(d_1 + d_2). \tag{17}$$

Definition:

$$r \equiv d_2/d_1. \tag{18}$$

[7] H. A. Bauermeister, "Metal-ceramic air-cooled microwave triodes," AF Translation F-TS-4945-RE (ATI-X28181); 1944.

[8] C. Beurtheret, "Evaporation-cooled power tube" *Electronics*, vol. 25, pp. 106–107; March, 1952.

[9] H. deBrey and H. Rinia, "An improved method for the air cooling of transmitting valves," *Phil. Tech. Rev.*, vol. 9, pp. 171–178; April, 1947.

[10] B. O. Buckland, "Electron—tube heat-transfer data" *Elect. Engrg.* vol. 70, pp. 962–966; November, 1951.

[11] B. O. Buckland, "Basic heat-transfer data in electron tube operation," *Proc. AIEE*, vol. 70, pt. 1, pp. 1079–1085; 1951. Also, *Elect. Engrg.* vol. 70, pp. 962–966; November, 1951.

[12] B. O. Buckland, "Heat transfer from electronic tubes at high altitudes and high ambient temperatures," *Trans. AIEE Comm. and Elec.*, vol. 72, pt. 1, pp. 698–703; November, 1953.

[13] F. F. Buckland, "Application of fundamental heat transfer data to component thermal analysis," Proc. First Conf. Cool. Air. Elec. Equip., Ohio State U., vol. 21, no. 2, Eng. Exp. Stat. Bull. 148, pp. 107–108; 1952.

[14] P. J. Erdle, "Heat transfer from vacuum tubes" *Sylvania Tech.*, vol. 7, pp. 80–82; July, 1954.

[15] J. S. Hickey Jr., "Heat transfer at high power densities," *J. Appl. Phys.*, vol. 24, pp. 1312–1317; October, 1953.

[16] F. Jenny, "The development of a water-cooled transmitting triode of 50 kw anode dissipation," *Brown Boveri Rev.*, vol. 33, pp. 211–214; August, 1946.

[17] B. Lemeshka and A. G. Nekut, "High efficiency coolers for power tubes," *Tele-Tech.* vol. 11, pp. 69–71; November, 1952.

[18] J. G. Matthews, "The reduction of temperature of vacuum tube envelopes," Proc. First Conf. Cool. Air. Elec. Equip., Ohio State U., vol. 21, no. 2, Eng. Exp. Stat. Bul. 148, pp. 150–152; 1952.

[19] I. E. Mouromtseff, "Temperature distribution in vacuum tube coolers with forced air cooling," *J. Appl. Phys.*, vol. 12, no. 6, pp. 491–497; June, 1941.

[20] I. E. Mouromtseff and H. N. Kozanowski, "Grid temperature as a limiting factor in vacuum tube operation," PROC. IRE, vol. 24, pp. 447–454; March, 1936.

[21] E. M. Ostlund, "Air cooling applied to external-anode tubes," *Electronics*, vol. 13, pp. 36–39; June, 1940.

[22] O. Schärli, "New type of radiator for high power transmitting tubes," *Brown Boveri Rev.*, vol. 36, pp. 311–315; 1949.

[23] A. J. Young, "Radiators for transmitting valves," *Marconi Rev.*, vol. 12, pp. 85–91; July–September, 1949.

[24] B. Ziefvert, "Air cooled steel tank mercury arc rectifiers," *Jour. Allmande Svenska Electriska Aktiepalogel*, vol. 22, pp. 127–149; 1949.

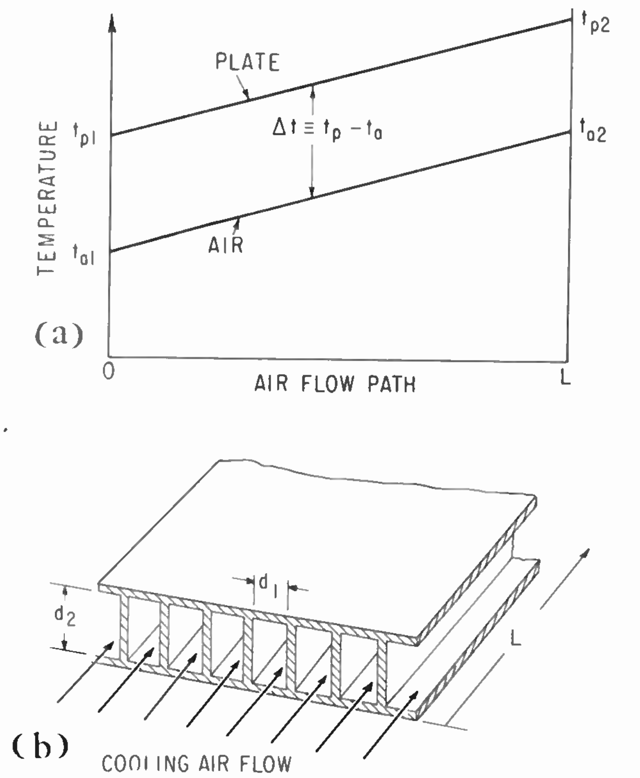

Fig. 9—(a) Approximate temperature distribution for cold plate. (b) Section showing cold-plate dimensions.

Definition:

$$Re \equiv D_e G/\mu. \tag{19}$$

Definition:

$$G \equiv w_a/nd_1d_2. \tag{20}$$

Newton's Second Law for the approximation of incompressible flow:

$$\Delta p = 4f(L/D_e)G^2/2g_c. \tag{21}$$

Experimental correlation for turbulent flow ($Re > 5{,}000$):

$$f = 0.046Re^{-0.2}. \tag{22}$$

Equation of state (air is a perfect gas):

$$p_i - \Delta p/2 = R(T_{a2} + T_{a1})/2. \tag{23}$$

Eqs. (13) through (23) contain a total of 22 different variables. The design or specification of the cold plate usually fixes ten of these 22 variables; for example, many designs fix the following set: q, c_p, t_{a1}, L, t_{p2}, k, μ, Δp, r, p_i. Hence there remain 11 independent equations, *i.e.*, (13) through (23), to solve for the 12 remaining variables.

If the specifications are used to fix ten variables, then an additional variable must be specified in order to obtain a unique solution. Since the above analysis is limited to the *turbulent* flow regime, the additional variable which may be specified can be the Reynolds number, or some other variable. If (16) and (22) are replaced by similar correlations for *laminar* flow, the Reynolds number may again be selected as the additional parameter to obtain a unique solution for design of the cold plate in the laminar flow regime.

The above analysis does not, however, provide a means of demonstrating that an optimum design exists and can be obtained. This analysis provides only a means of obtaining a large number of possible designs for a given set of specifications, and does not yield a means of solution of the best design. The problem of optimization, *i.e.*, of selection of an optimum or best design, from this large number of designs must be based on a criterion which is, of necessity, completely independent of this type of analysis. Different independent criteria exist for optimization of the design of a cold plate. Two of these criteria are as follows:

Economic Criterion: Many designs of diverse equipment are optimized in the process industries, in electrical manufacturing, etc. by using the important economic criterion that the optimum design corresponds to the case where the sum of the yearly fixed charges and yearly operating charges *shall be a minimum.*

Combination of Cold-Plate Weight and Weight of Auxiliaries, etc.: For airborne cold-plate equipment, the sum of the weight of cold plate plus equivalent weight of aircraft resulting from its growth factor caused by air flow and auxiliaries for the cold plate *shall be a minimum.*

Instead of optimization of the designs, it is also found convenient to employ criteria which limit or fix a boundary condition for a series of designs. Three such bounds can be described as follows:

- The weight of the cold-plate equipment *shall be less than A pounds.*
- The total volume of cold-plate equipment *shall be less than B cu feet.*
- The rate of flow of cooling air *shall be less than C lb/min.*

An example of the design of a cold plate is shown in Fig. 10, supplied by the courtesy of the Missile Systems Division of Raytheon Manufacturing Company. At the bottom are shown the entrance and exit headers together with the ducts used for flow of cooling air through this cold plate. In the sketch above are shown the details of the tube clamping device used to obtain excellent conductive paths from the hot tube surface to the cooler cold-plate surface. Although the sketch shows only tubes mounted on one side of the chassis, it is evident that in actual practice tubes can be easily mounted on both sides of this cold plate. The topmost photograph shows the milled ducts of an experimental cold plate; for production purposes other techniques of fabrication would be employed. The remaining photographs show the partially assembled cold plate with headers and the array of subminiature tubes mounted

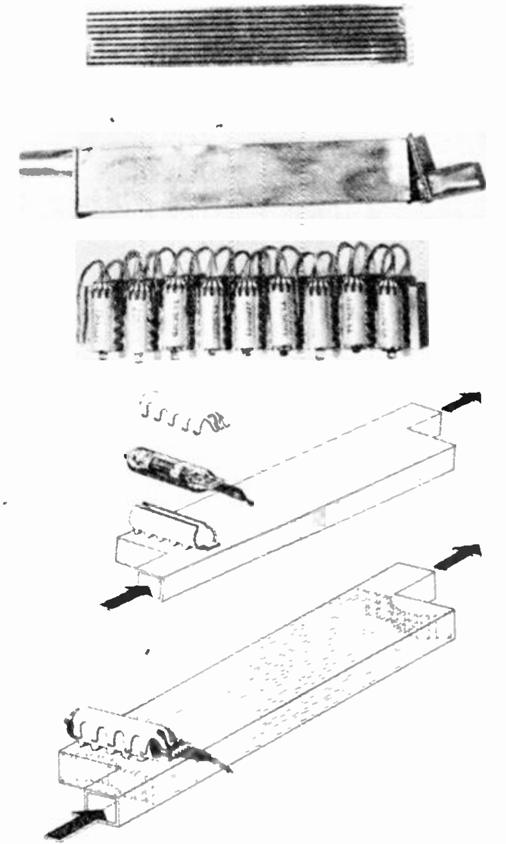

Fig. 10—Details of Raytheon cold plate.

on one side of the cold-plate surface. This type of cold plate can be designed accurately and quickly for either laminar or turbulent air flow to meet a particular set of specifications which are strongly dependent on the end use of the electronic equipment.

Additional information on design, construction tests, and other details of cold-plate heat exchangers used for cooling of electronic equipment can be found in Jerger,[25] Katz,[26] Mark and Stephenson,[27] Mathis,[28,29] and Robinson, Jones, and Sepsy.[30]

3) *High-Temperature Vacuum Tubes*

One of the most interesting developments in recent years, which can be characterized as a result of creative engineering, especially electronically and thermally, is the appearance of vacuum tubes designed specifically to operate with surface temperatures near 500°C. If the ordinary type of vacuum tube, with a filament dissipation of 1 to 3 watts is miniaturized or subminiaturized, the heat flux begins to approach values of the order of 1 watt per square inch of external surface. It was shown in the previous discussion that effective cooling of such a cylindrical tube by combined radiation and natural convection would require surface temperatures in the range of 500 to 800°F. Hence, the problems faced in the development of these new tubes were to find materials with suitable thermal and mechanical properties for these high temperatures and to discover simple methods of construction which could maintain the required high vacuum and still allow the possibility of automation in mass production. Several manufacturers of vacuum tubes have investigated this problem in the last decade but only few of the available results will be discussed here.

One of the first attempts to determine the feasibility of high temperature vacuum tubes was made by Raytheon Manufacturing Company. Over a period from June, 1949 to March 2, 1953 when the final report[31] was issued, the objective was "to develop rugged electronic tubes similar to the 6L6, 6AG7, and 1/2 of the 6AS7, which shall be housed in small envelopes and be operated at high power levels at high altitudes and at high ambient temperatures. The most drastic condition under which these tubes will be expected to operate efficiently will be at an ambient temperature of 250°C (480°F) and at an altitude of 60,000 feet." Other principal objectives were attainment of full ratings at 60,000 feet altitude, capability of operation from −55°C to 250°C ambient temperature, and dependable (free-air) long life performance.

The results of this research at Raytheon led to the successful development of tubes with 3 to 5 the normal previous ratings in about 1/20 the volume of the prototypes. These new tubes were operable for long periods, during life tests, in an ambient of 250°C which was, of course, a temperature far in excess of that of which the prototypes were capable. Many novel materials, methods of sealing, manufacturing, etc. were discovered in this search for a new high temperature tube.

Wheeler and Evans of Sylvania Electric Products Inc., have described[32] the development of reliable miniature tubes for automatic production which will withstand continuous operation in high ambient temperatures. Fig. 11 shows a semiexploded view and a photograph of the SN-1724 Fe stacked tube, supplied through the courtesy of Sylvania Electric. This type of

[25] J. J. Jerger, "The Application of Proven Techniques to the Cooling of Airborne Electronic Equipment," Raytheon Rep., Missile Div. February 3, 1956.

[26] L. Katz, "The design of properly cooled electronic equipment for operation at high altitudes," Proc. First Conf. Cool. Air. Elec. Equip., Ohio State U., vol. 21, no. 2, Eng. Exp. Stat. Bul. 148, pp. 25–43; 1952.

[27] M. Mark and M. Stephenson, "The Air-Cooled Electronic Chassis," ASME Paper No. 55-A-55; 1955.

[28] G. Mathis, "Experiments and design data for air-cooled electronic chassis," Proc. First Conf. Cool. Air. Elec. Equip., Ohio State U., vol. 21, no. 2, Eng. Exp. Stat. Bul. 148, pp. 83–94; 1952.

[29] G. Mathis, "Air-cooled chassis design and application," Proc. Second Conf. Cool. Air. Elec. Equip., Ohio State U., vol. 1, pp. B-4-1–B-4-9; 1953.

[30] W. Robinson, C. D. Jones, and C. F. Sepsy, "Conduction and Radiation Cooling of Electronic Units with Cooling Plates," Ohio State U. Res. Foundation Rep. No. 46; August, 1955.

[31] G. Freedman and M. L. Anderson, Progress Reps. 1 to 22, from June 30, 1949 to Feb. 28, 1953. Raytheon Manufacturing Co. Contract No. AF33(038)-5734 and No. S5(S51-2008); Final Rep. No. 1041-22.

[32] W. R. Wheeler and T. L. Evans, "Reliable tubes for automatic production," *Rad. and Telev. News*, vol. 59, pp. 10 11, 36, 37; September, 1954.

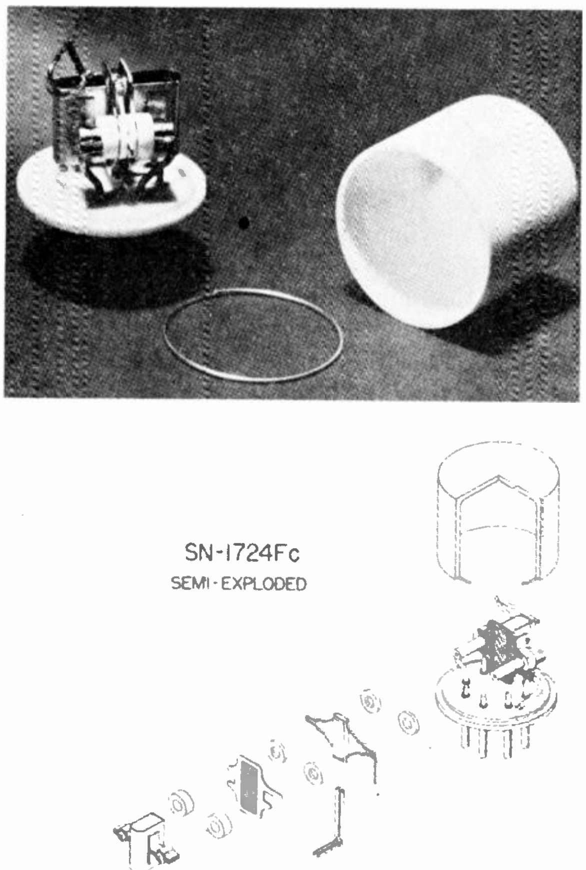

Fig. 11—High temperature ceramic vacuum tube, SN-1724Fc.

stacked tube has been subjected to intensive life tests at room temperature, at 300°C ambient, and at 400°C ambient. The results after tests of 1,000 hours indicate that the upper limit of ambient operation has not been reached. The results of life tests, shock tests, electrical ratings, etc., indicate that the stacked designs are more rugged, highly stable under adverse conditions, more uniform than the conventional prototypes, and that they will improve receiving tube reliability.

A third example of a new high temperature vacuum tube is the micro-miniature triode 6BY4, developed recently by General Electric Company.[33] The bottom portion of Fig. 12 shows the details of the various parts of this new ceramic tube, while a cutaway view and a photograph of the finished tube are shown in the upper portion. An ordinary pencil is used to indicate the size of this ceramic tube. This tube can be operated at very high temperatures. The tube was designed to operate at very high frequencies, for ruggedness under all types of shock conditions, and for automatic production. Fig. 13 compares the size of this new micro-miniature ceramic tube with the sizes of the conventional glass tube with a plastic base and of a miniature glass tube. The size reduction attained in the 6BY4 tube is a significant advance in the art of high temperature vacuum tubes.

[33] GE Tube Dept. Data Sheets on 6BY4; October 11, 1955.

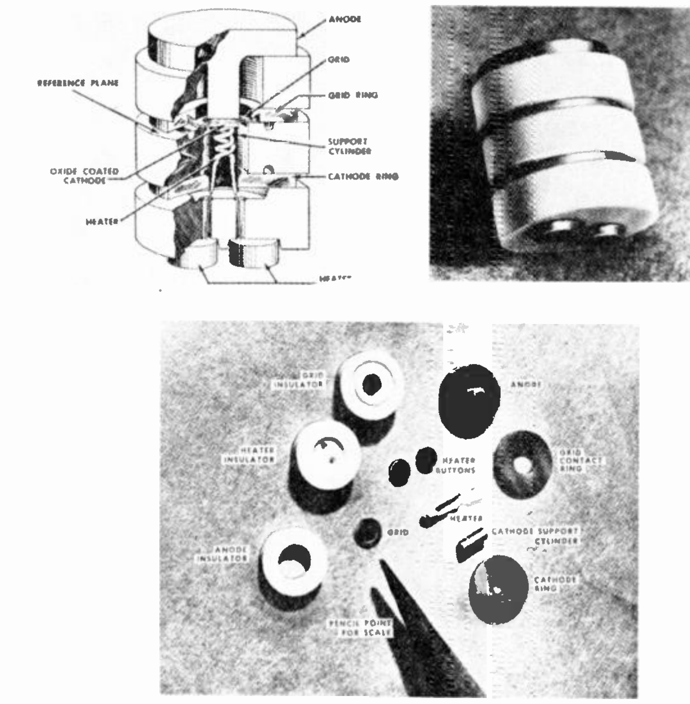

Fig. 12—High temperature ceramic vacuum tube, 6BY4.

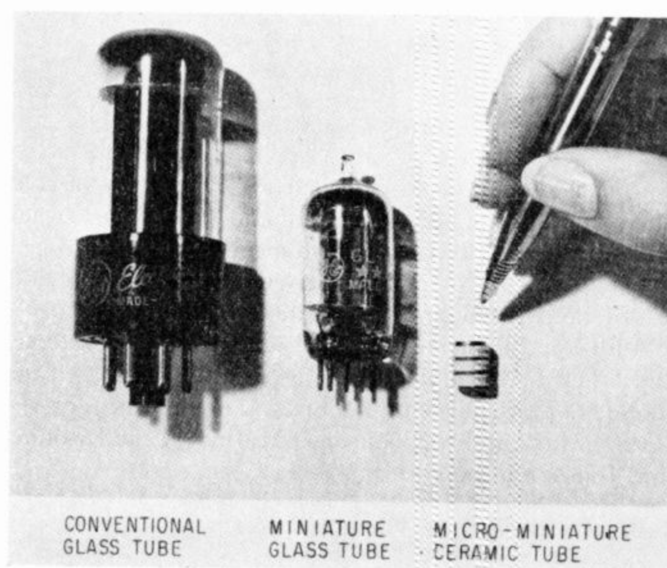

Fig. 13—Size reduction obtained with new ceramic tube, 6BY4.

Other examples, studies, etc. of thermal problems of high temperature components may be found in Frey,[34] Schell,[35] and other reports.[36,37]

[34] H. A. Frey and J. M. Jesatko, "The electrical performance of ceramic dielectrics at elevated temperatures," *Trans. AIEE*, vol. 65, pp. 911–920; 1946.

[35] J. W. Schell, "High-temperature capacitors," Proc. First Conf. Cool. Air. Elec. Equip., Ohio State U., vol. 21, no. 2, Eng. Exp. Stat. Bul. 148, pp. 70–72; 1952.

[36] "High Temperature Miniaturized Capacitors," Sprague Electrical Co. Contract No. AF-33(038)-9328 (Third Progress Report October 1, 1950.

[37] "A Study of High Dielectric Constant Films for High Temperature Operation," Balco Res. Labs. Final Report, Contract No. W36-034-32361.

4) *Evaporative Cooling in Design of Transformers*

The miniaturization of magnetic components by use of evaporative-cooling techniques has been successfully accomplished in the past few years. One of the best examples was given by Kilham, Ursch, and Ahearn,[38] of Raytheon Manufacturing Company. The major objective of this study was to apply fluorochemical dielectric materials to obtain efficient evaporative cooling of transformers in order to effect a significant reduction in weight and size for the same power and to enhance their electrical performance. This final report contains an excellent summary of the properties of these fluorochemicals and of mixtures with other gases, sources of supply, and a useful extensive bibliography.

Fig. 14 shows a cutaway view of a typical transformer using fluorochemicals for evaporative cooling. The sealed transformer contains some liquid phase at the bottom. In steady-state operation of this transformer, a wick of woven-glass insulating material, which is easily wetted by the liquid, is used to transport this liquid by capillary action to the interstices of the transformer coil. This liquid phase rises in temperature and finally evaporates at the hotter surfaces of the coil windings. The vapor phase passes outward and condenses at the cooler inside surface of the container. The condensate is then returned by gravity to the liquid phase at the bottom of the container.

There are many advantages of this type of evaporative cooling in this transformer design. First, the heat-transfer coefficients for a boiling liquid on the hot winding surfaces and for a condensing vapor on the cooler container surface are about one order of magnitude greater than those obtainable with only a gas phase. These high values of heat-transfer coefficient cause a large and significant reduction in required heat-transfer area for a given power dissipation. Second, significant reductions in size and weight of evaporatively-cooled transformers are possible; an example is shown in Fig. 15 of the size reduction obtained with a magnetron filament transformer. Third, this type of cooling can result in transformers capable of safe operation for long periods under conditions of high ambient temperatures. Finally, the fluorochemical dielectrics used in this type of cooling have enhanced the electrical performance of the transformers.

Fig. 14 shows that internal copper plates are also used in this transformer to conduct heat generated in the winding interstices to a cold base or cold plate. The evidence obtained in this study by Kilham, *et al.*[38] indicates that cold-plate techniques are quite applicable to these transformers for use on land or shipboard and that considerable further size reduction of these transformers is easily attainable.

[38] L. F. Kilham, Jr., R. R. Ursch, and J. F. Ahearn, "Final development report on 'Gaseous or Liquid Filling of Magnetic Parts' " Contract No. NObsr-63239; January 1, 1956.

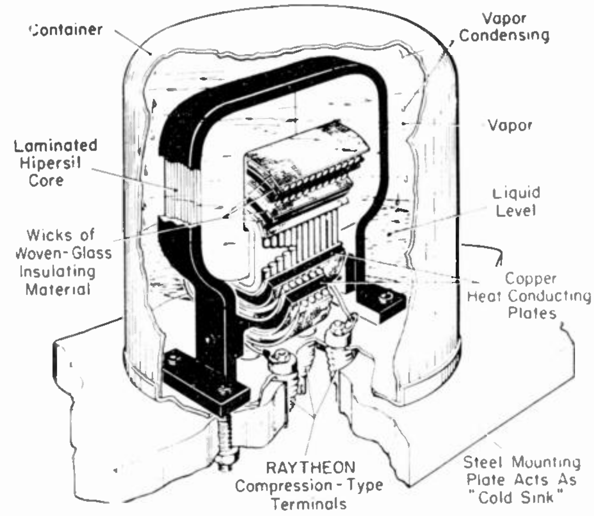

Fig. 14—Cutaway view of fluorochemical transformer using evaporative cooling technique.

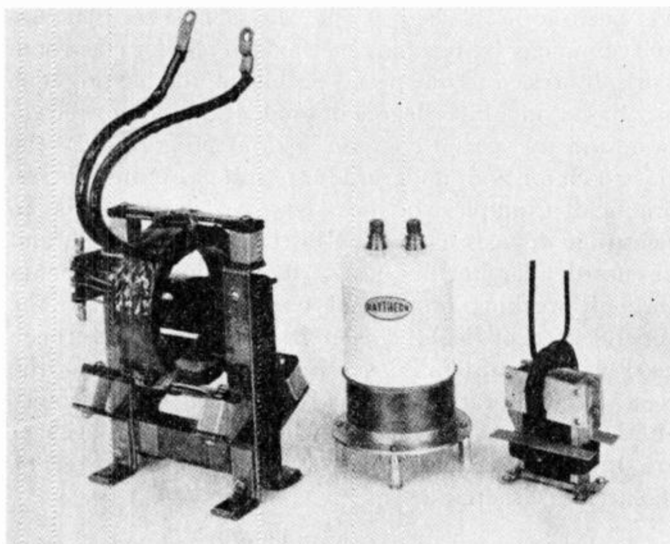

Fig. 15—Size reduction of magnetron filament transformer obtained with evaporative cooling using fluorochemical dielectric.

Other examples of evaporative cooling techniques may be found in Beurtheret,[8] Dawes,[39] Fritz,[40] Greene and Wightman,[41] Kilham and Ursch,[42,43] de Koning,[44]

[39] C. L. Dawes and W. R. Mansfield, "Built-up mica plate for high temperature applications" *Elect. Engrg.* vol. 72, pp. 145–150; January, 1953.

[40] D. Fritz, "Environment-free alternator," *Aero-Dig.*, vol. 64, p. 82; June, 1952.

[41] A. D. Greene and J. C. Wightman, "Engineering study of direct evaporative cooling for electronic equipment" AMC MCREE 48-32; 1948.

[42] L. F. Kilham, Jr. and R. R. Ursch, "Fluorochemical liquids and gases as transformer design parameters," IRE Natl. Convention, New York, N. Y. March, 1955.

[43] L. F. Kilham, Jr. and R. R. Ursch, "Transformer 'miniaturization' using fluorochemical liquids and conduction techniques," Natl. Electr. Components Symposium, Los Angeles, Calif.; May, 1955.

[44] T. de Koning, "Vaporization cooling of large electrical machines," *Elect. Engrg.* vol. 68, pp. 385–392; May, 1949. Also, *Power Gen.*, vol. 53, pp. 67–69; May, 1949. Also, *J. Franklin Inst.*, vol. 248, pp. 49–68; July, 1949.

Kroko,[45] Martin and Hambor,[46] Olyphant and Brice,[47] Robinson and Han.[48]

Educational Motivation for Heat Transfer in Electronics

Industrial demand for trained manpower in the field of heat-transfer applications to electronic equipment has grown very rapidly in the last five years. There has been considerable delay, however, on the part of educational institutions to meet this demand. One reason for this delay has been the slowness with which fictitious barriers between mechanical and electrical engineering are being overcome. In my opinion, the engineer who works in both fields must be ready and able to adapt his thinking and his design procedures to encompass both fields at will, if he is interested in obtaining an optimum over-all design of an electronic device, from thermal and electronic viewpoints.

The education and training of manpower in this area has been accomplished at the Massachusetts Institute of Technology by teaching methods in regular classroom work, by research on special problems, theses, projects, etc. based on a high degree of student and staff motivation, and by special conferences and programs. In the classroom for both undergraduate and graduate courses, practical examples of heat-transfer applications to electronic devices have been introduced, analyzed, and discussed. The techniques of motivation of students through working on special research projects in the general area of heat transfer in electronics have been used with considerable success. For example, in the Research Laboratory of Heat Transfer in Electronics at M.I.T., many of the following topics have been or are under investigation as student theses, projects, research tasks, etc.:

1) Design and construction of very small heat meters.
2) Design and construction of thermoelectric generators.
3) Analyses and tests on small rotating electric machinery to unravel the complicated relations between thermal, magnetic, and electric processes. The preliminary results of this type of research for a dc motor may be found in the literature.[49,50] Further work is continuing on analyses and tests of an induction motor.
4) Temperature control of gyros.[51]
5) Temperature distribution in missile components and systems.
6) Means of protection of electronic equipment from abnormal environmental conditions by means of energy storage utilizing chemicals undergoing solid to liquid phase transformations.[52]
7) Analysis, design, and construction of cooling devices based on the reversible Peltier effect.
8) Design, construction, and testing of devices for measurement of thermal conductivity of plastics and related insulating materials at high temperatures.
9) Analysis, design, and construction of novel energy conversion devices.
10) Analytical problems which are based on step changes in heat generation in bodies with flow of electricity in order to establish design charts for transient temperature response.
11) Analysis and experimentation on thermal characteristics of power transistors and means of cooling them under various environmental conditions.
12) Evaporative cooling techniques applied to various electronic equipment.
13) Basic studies of heat flow and fluid flow in annuli, such as may be encountered in the case of rotating machinery, in the cooling of tubes and components with chimneys or surrounding ducts, etc.

This list of topics indicates that the main task in motivation is to get the student or staff member interested in studying and solving problems requiring a high degree of creative engineering in the fields of heat transfer and electrical engineering.

It is only recently that special programs have been organized and presented to aid and inform engineers

[45] L. J. Kroko, "Vapor cooling for high altitudes," *Elec. Mfg.*, vol. 50, p. 120; October, 1952.

[46] C. G. Martin and V. F. Hambor, "Survey of evaporative and liquid coolants for rotating electrical machines," AIEE Publ. no. S-57, Paper no. CP-7; 1953.

[47] M. Olyphant, Jr. and T. J. Brice, "Dielectric and coolant studies of inert fluorochemical liquids," Proc. 1954 Electronic Components Symposium, Washington, D. C.

[48] W. Robinson and L. S. Han, "Cooling Systems for Electronic Equipment Utilizing Expendable Evaporative Coolants," Ohio State U. Res. Foundation, Report 30B; 1949.

[49] J. Kaye and S. W. Gouse, Jr., "Thermal analysis of a small dc motor—Part I: Dimensional analysis of combined thermal and electrical processes." To be published by AIEE.

[50] J. Kaye, S. W. Gouse, Jr., and E. C. Elgar, "Thermal analysis of a small dc motor—Part II: Experimental study of steady-state temperature distribution in a dc motor with correlations based on dimensional analysis." To be published by A.I.E.E.

[51] G. N. Hatsopoulos and J. Kaye, "A novel method of obtaining an isothermal surface for steady-state and transient conditions." In press.

[52] J. Kaye, R. M. Fand, W. G. Nance, and R. J. Nickerson, "Final Report on Heat-Storage Cooling of Electronic Equipment," final rep. on AF Contract No. AF 33(616)-2358; December, 1955

who are working in the general area of heat-transfer design and of applications to electronic devices. Two of these conferences on "Cooling of Airborne Electronic Equipment" have been held at Ohio State University in 1952 [2] and in 1953 [4]. In both conferences a large number of papers and reports was presented and discussed. A special program was held at M.I.T. during the summer of 1955 [7], on "Industrial Applications of Heat Transfer to Electronics," in which M.I.T. staff members and invited industrial lecturers reviewed, lectured, demonstrated, and discussed many phases of heat transfer which could be applied to electrical engineering in a broad sense. In addition to these particular conferences, numerous others have been held at meetings of various engineering societies, at meetings of different contractors working on airborne electronic equipment, etc. These conferences are bringing the important educational aspects of these thermal problems out into the open, and are making many groups cognizant of the important tasks still to be performed in this area.

Finally, it should be mentioned that efforts are underway by various groups to prepare and publish suitable texts, manuals, design charts, etc., for further education of the engineer and stduent in the general area of heat-transfer applications to electronics. Examples may be found in the publications by Booth and Welsh [1], Welsh [5], and Kays and London [6].

Conclusion

The science of heat transfer is exceedingly useful to the designer of electronic components and devices; it can lead to accurate predictable engineering designs before construction if sufficient details of the fluid-flow and thermal-flow paths are available. The art of heat transfer based on accumulated experience can be used, intuitively, to produce useful solutions for many complex heat-flow problems in electronics where basic thermal and fluid-flow data are not available.

The need for creative engineering efforts, based on thermodynamics, heat transfer, fluid mechanics, and mass transfer, to help solve problems in electronic components and devices, has increased rapidly in the last decade with the introduction of supersonic and hypersonic aircraft and missiles. These high speed devices require developments of new electronic components and devices which will operate reliably at extremes of high ambient temperature, at low air density, and in the presence of shock effects, etc. These challenging problems can be solved only by concerted efforts which recognize no artificial barriers between diverse fields but rather which emphasize the over-all unity of engineering science.

Acknowledgment

It is a pleasure to acknowledge the discussions over the past year of the contents of this report with many colleagues who aided materially in broadening of its scope. Special thanks are due to R. M. Fand, E. F. Kurtz, R. H. Shoulberg, and F. Viles, for detailed criticism of the final draft.

I am greatly appreciative also of the excellent cooperation given my requests for illustrative material by the various divisions of Raytheon Manufacturing Company, by General Electric Company, and by Sylvania Electric Products, Inc. Finally, I wish to thank Professor A. L. London and the IRE for permission to reproduce part of his paper on the 25 kw tube.

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[4] *Cooling of Airborne Electronic Equipment and Rotating Electrical Machines.* Second Conference at Ohio State University (June, 1953).
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Review of Ionospheric Effects at VHF and UHF*

C. G. LITTLE†, W. M. RAYTON†, AND R. B. ROOF†

Summary—**This paper summarizes the present-day knowledge of ionospheric effects at vhf and uhf, with the exception of forward scattering of vhf radio waves by the ionosphere. The seven effects covered in the paper are: Radar Echoes from Aurora, Radar Echoes from Meteors, The Faraday Effect and Radar Echoes from the Moon, Radio Noise of Auroral Origin, Absorption of Radio Waves by the Ionosphere, Refraction of Radio Waves by the Ionosphere, and the Scintillation of the Radio Stars. Each ionospheric effect has in turn been divided into separate subtopics, and the main results are given in these subsections, with particular emphasis upon providing references to the original papers. In this way the reader wishing to know the answer to a specific problem will speedily be able to find a summary of the main published results in the field, and also be able to learn which papers deal with the particular topic of interest.**

I. Introduction

The occurrence in the earth's upper atmosphere of ionized regions of differing ionization density and hence of different radio frequency refractive index, has important effects upon the propagation of radio waves through it. These effects include the reflection, refraction, absorption, change of polarization, scatter, and diffraction of some or all of the incident radio energy.

The deviations of refractive index from unity decrease with increasing frequency; at first it was considered that these deviations would have negligible effect upon radio wave propagation at frequencies above about 30 mc. More recently, new techniques, such as auroral and meteor radar, and the radio astronomical studies of the upper atmosphere, have shown that the effects are sometimes significant even at frequencies considerably above 30 mc. It is the purpose of this review to summarize what is now known about some of these effects in the frequency range 30–3000 mc.

Some of these effects (notably refraction, absorption, change of polarization) can be introduced by ionospheric layers which are uniform in the horizontal plane (*i.e.*, vary only with height). In such cases the effects may be expected to vary relatively slowly with time and with position upon the earth. Certain other of the effects, such as auroral and meteor echoes and ionospheric scatter and diffraction effects, are dependent upon the presence of relatively small-scale irregularities in the ionospheric layers. In this case, the variations with time and position can be expected to change much more rapidly. These small-scale irregularities may also be capable of producing refraction, absorption, and polarization effects; since these rapidly varying effects may be superimposed upon a slowly varying effect due to the regular or standard ionosphere, it is sometimes necessary to differentiate between the normal average values of the phenomena and the short-term irregular deviations from these mean values.

Two important limitations of the report should be emphasized. Firstly, the forward scatter of vhf radio waves by the ionosphere is not described. To have included this topic would have meant lengthening the review considerably; it was also felt that little useful purpose would be served in view of the admirable series of papers recently published in the October, 1955, issue of Proceedings of the IRE. Secondly, this review is limited to ionospheric effects, and no attempt has been made to cover tropospheric effects, even though, under some circumstances, these may be more serious than the ionospheric effects.

In order to retain a satisfactory perspective, it should always be remembered that ionospheric effects at vhf and uhf, though possibly important, are nevertheless considerably smaller than those at lower frequencies.

II. Radar Echoes from Aurora

Introduction

Since the discovery and identification of aurorally-scattered vhf radio waves by Harang and Stoffregen [1], the phenomenon has been investigated in several countries. Very considerable progress has been made in determining the main characteristics of the phenomenon; this information is summarized in the succeeding sections. However, although general agreement exists as to the main characteristics of the echoes themselves, there is considerably less certainty about the parameters of the auroral scattering centers which are responsible for the echoes.

Relation of Radar Reflections and Visual Aurora

Many of the observers of ionospheric radar echoes at vhf have attempted to relate the position of the reflecting surfaces with those of the visual auroral forms. While all workers are agreed that there is close statistical correlation between the occurrence of vhf echoes and aurora or magnetic storms, there is disagreement as to the correlation between the positions of individual echoing surfaces and visual aurora. Thus, although Harang and Landmark [2] state that there is no correlation between the positions of auroral forms and the range of the observed echoes, most workers [3–7] imply or state that such a correlation is observed, *i.e.*, that the echoing surfaces are closely associated in space with visual auroral forms.

* Original manuscript received by the IRE, January 16, 1956; revised manuscript received, May 28, 1956. This review is based on a report prepared by the Geophysical Institute of the University of Alaska for the Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y., under Contract No. AF 30(635)-2887.

† Geophysical Institute, University of Alaska, College, Alaska.

The three fullest investigations appear to be those of Currie, Forsyth, and Vawter [5], Harang and Landmark [2] and Bowles [7]. A comparison of the data presented in these three papers suggests that much of the apparent discrepancy is due to the inclusion in the analysis by the Norwegian workers of aurora other than those occurring at low elevations in the north. As explained in the following part of this section, a marked aspect sensitivity occurs, rendering the radars insensitive to aurora at high elevations or in the south. When the analysis is restricted only to those areas of the sky where echoes are obtained, then as shown by the Saskatoon workers [5] and by Bowles at College [7], fair to good correlation between the ranges and azimuths of visual and radar aurora is obtained. However, as pointed out by Bowles, the intensity of the echo is not proportional to the brightness of the visual forms.

The generally accepted conclusion is, therefore, that the auroral radio wave scatterers are closely associated in space and time with visual auroral forms, but that the brightness of the latter is no indication of the strength of the echoes. Owing to the aspect sensitivity factor, only those auroral forms occurring at low elevations toward the north can be expected to give echoes.

Aspect Sensitivity

Observations at several centers in the northern hemisphere have shown that the auroral ionization giving rise to vhf radar echoes possesses a form of aspect sensitivity. Although most observers have used fixed antennas, the azimuthal distribution of auroral echoes has been investigated in Sweden [8], England [9], Canada [5], the United States [10], and Alaska [7, 11, 12]. In each case the echoes were most frequently observed toward the magnetic north, with a scatter in angle of the order ± 60 degrees. That this type of azimuthal distribution is fundamental to the auroral echoes is shown by Dyce's observations at Point Barrow, north of the auroral zone [13]. Despite the fact that auroras are most frequently seen south of the zenith from this point, he found that echoes were most frequently obtained when the antennas were directed north.

Observations of the variation of auroral echoes with elevation angle have also been made at various stations [2, 6, 10–12] and have shown that echoes are rarely observed at elevations greater than about 15 degrees at 100 mc and somewhat higher (say 20 degrees) at 30 mc.

The interpretation of these results has been discussed by Bowles [14] and Dyce [15], on the basis of a suggestion by Moore [16] and others [17, 18] that the ionization which gives rise to vhf auroral echoes is aligned along the lines of the earth's magnetic field. Such a distribution of ionization would most readily produce an echo when the radar beam is perpendicular to the magnetic field in the aurora. At temperate and higher latitudes these lines of force are steeply inclined relative to the earth's surface. The auroral ionization is therefore most readily detected when the radar is directed towards magnetic north at low angles of elevation. Bowles and Dyce have expressed their auroral radar echoes results in terms of "off-perpendicular" angle, and have convincingly demonstrated the magnetic control of this auroral aspect sensitivity. This conclusion that the radar echoes are due to scattering from ionization aligned along the earth's line of force has also been reached by other workers [10, 19].

The College observations have been further analyzed by Dyce [20], to determine the scattering polar diagram of the auroral ionization in more detail. This work indicates that the beamwidth of the scattered radiation is roughly 6 degrees between half-power points at 50–100 mc, indicating that the equivalent length of the elongated scatterers is of the order of 10 wavelengths.

The theory of auroral radar echoes has been discussed by Booker [20], assuming nonisotropic scatterers, and curves have been presented showing the manner in which the aspect sensitivity factor would vary as a function of range in the magnetic meridian, for a radar at magnetic dip ± 75 degrees. Similar curves showing the aspect sensitivity as a function of azimuth were presented, and, assuming certain equipment parameters, curves were given showing the expected range distributions at various frequencies for various amounts of ionospheric attenuation.

Frequency Dependence

Rather little systematic work has been done on the dependence of auroral echo strength upon observing frequency. The first recorded dual frequency observations were those of Lovell, Clegg, and Ellyett at 46 and 72 mc [3], who found on the one occasion that they observed auroral echoes, that they were observable only at 46 mc. Later observations at 56, 106, and 3000 mc by Forsyth [21] and by Currie, Forsyth, and Vawter [5] showed the presence of auroral echoes on the two lower frequencies only. Successful dual frequency observations at 35 and 74 mc have been described by Harang and Landmark [2], and at 52 and 106 mc by Dyce [15].

The most complete analysis of multi-frequency observations is that given by Forsyth [21]. Two similar radars at 56 and 106 mc were used at Saskatoon, and the relative amplitudes of simultaneous echoes obtained on the two equipments were investigated. The results showed that the 56 mc echoes were observed approximately 4 times as frequently as the echoes at 106 mc, and that 56 mc echoes were almost always observed whenever a 106 mc echo was present. In general, it was found that for a given 56 mc echo strength, there was a maximum 106 mc echo strength, and that the 106 mc echo could have any value less than this maximum value.

A decay of echo strength with frequency is also indicated by the observations taken at College, Alaska, by Dyce [15] using 52 and 106 mc. Although the latter equipment was considerably more powerful (approximately 20 times in peak power and roughly 10 times in antenna gain), a similar number of echoes was obtained on each equipment. The correlation in range and azimuth of echoes appearing simultaneously on the two frequencies was often good, but was occasionally lacking. Such discrepancies may be due in part to the different polar diagrams, though they have also been reported by Harang and Landmark [2].

The maximum frequency at which auroral echoes is detectable is, of course, a function of the equipment parameters. So far as is known, the highest frequency at which auroral effects have been observed is about 220 mc, where radio amateurs [22] have reported signals showing the very high fading rate characteristic of auroral echoes.

Diurnal Variation of Auroral Echo Activity

The diurnal variation of auroral echo activity has been investigated by workers in several countries. For temperate latitudes, Bullough and Kaiser [23] have published results based on $4\frac{1}{2}$ years' operation at Manchester, England, on 72 mc, and Dyce [15] has investigated auroral echo activity over a two-year period at Ithaca, N. Y., on 50 mc. In both cases, the observers report a broad minimum of activity during the midday hours followed by an early evening maximum at about 1900h, a late evening minimum ($\sim$2100h Bullough and Kaiser, $\sim$2400h Dyce) and a second rather smaller maximum at about 0200h. The agreement in these diurnal curves, obtained with markedly different techniques, is striking, especially when they are compared with diurnal variations obtained at higher latitudes. Such investigations have been carried out at Kiruna, Northern Sweden, by Hellgren and Meos [8], by the Canadian workers at Saskatoon [5], and at College, Alaska [7]. In each of these cases, a single main maximum was obtained, peaking at about midnight, in good agreement with visual data for high latitude aurora. It would therefore appear that there is a marked difference between auroral echo activity at magnetic latitudes about +65 degrees and +55 degrees. At all latitudes, however, auroral echoes are very rare during the noon hours.

Seasonal Variation of Auroral Echo Activity

It is well known that the visual aurora show a marked seasonal variation, being observed most frequently during the equinoctial months (for example see [24] which analyzes 1627 auroras observed over a 55-year interval at the Yerkes Observatory in Wisconsin). Similar equinoctial maxima, with minima at the solstices, have also been reported by Dyce [15] for the radar echoes, and suggestions that the equinoctial maxima were observed have been made by Bullough and Kaiser, and as a result of observations at College [7]. In general, however, the published data from many of the sites is not sufficient to enable the magnitude of this seasonal variation to be estimated accurately.

Eleven-Year (Sunspot) Variation in Auroral Echo Activity

Continuous observations from 1950 through 1953 reported by Bullough and Kaiser [23] show a steady decrease in the number of aurora observed per year, in agreement with the decrease in sunspot numbers. Observations by Dyce [15] over a 2-year period commencing April, 1952, also show this decrease of activity with sunspot numbers. It would therefore appear reasonable to suppose that the auroral radar echoes will show a sunspot cycle effect similar to that shown by the visual aurora [24]. If this is the case, then vhf auroral radar echoes will be most commonly observed about two years after the occurrence of sunspot maxima, and will be least frequently detected at the sunspot minima. It may therefore be expected that the incidence of auroral echoes will begin to increase during 1956 and will continue to increase for several years.

Polarization of Auroral Radar Echoes

The effect of antenna polarization upon the strength of vhf auroral echoes has been discussed by several workers [2, 25–29]. The fullest (and most recent) investigation appears to be that by the Canadian workers McNamara and Currie [29] who transmitted horizontally-polarized signals at frequencies of 56 and 106 mc. Horizontally, vertically, and circularly-polarized receiving antennas were used at 56 mc; at 106 mc only horizontally and vertically-polarized antennas were used.

The polarization phenomena were found to be very different on the two frequencies. At 106 mc, all the echoes were found to be horizontally polarized, *i.e.*, of the same polarization as the transmitted signal. The same was true of only about one-third of the 56 mc echoes, the majority of which showed appreciable or even complete depolarization of the radiation. The mean power ratio between horizontally and vertically-polarized receiving 56 mc antennas, for all echoes, was about 5 db; restricting the analysis to those echoes showing depolarization, the mean value was 3 db.

The polarization ratio (horizontal to vertical field strength) was found to be a function of range. Echoes of the same polarization as that radiated were most frequently observed at ranges of the order 500–900 km, *i.e.*, in that region of sky where the perpendicularity requirement (See the third part of this section above) is most likely to be met. These observations have been discussed by McNamara and Currie in terms of two possible depolarizing processes: magneto-ionic effect and multiple scatter within the reflecting region.

Some results essentially similar to the above have been reported by Moore [25], Bowles [14, 26], and Harang and Landmark [2, 27]. The latter workers in particular observed a decrease in the depolarizing effect of the aurora with frequency, in that their 35 mc echoes were frequently completely depolarized (except during discrete echoes), and their 74 mc echoes remained linearly polarized.

The question of a possible difference in average echo strengths on two equipments identical except in the use of opposite polarizations, has apparently not been investigated, and whether vertical or horizontal polarization is the more favorable for the observation of auroral echoes is not known.

Height of Scattering Centers Responsible for Auroral Echoes

Experimental Observations: While the range of vhf auroral echoes can usually be determined with considerable accuracy if required, the angle of elevation of the target, and hence its height above the curved earth cannot readily be measured. This is due to the difficulty of obtaining sufficient angular resolution in the vertical plane, and to uncertainties as to the effect of ground reflection upon the vertical polar diagram at low angles of elevation.

Many authors have included comments as to the height of their echoes, *e.g.*, [3–6, 8, 9, 23, 30]. The fullest investigation, however, appears to be that of Currie, Forsyth, and Vawter [5], who made use of single station photographs to determine the elevation angle of the bottom of the auroral forms. The radar measurements were used to determine the slant range of the auroral scatterers; assuming that these scatterers were closely associated in space with the lower borders of the visual forms, the combination of elevation angle and slant range was used to calculate the height of the echoing centers above the (curved) earth. Their results are in good agreement with measurements of the heights of the lower borders of aurora made by parallactic photography, and indicate that the auroral echoing centers are most frequently observed at a height of 100 km. The apparent scatter in height was considerable (from 70 to 160 km), but about 2/3 of the echoes obtained by these Canadian workers occurred in the range 85–115 km. Similar results were obtained from an analysis of the observed range distributions which show maxima and minima corresponding to the lobes in the antenna polar diagrams. The results of other workers seem to be in agreement with those described above.

Theoretical Considerations: As described in the section dealing with the aspect sensitivity of auroral echoes, it has been shown that echoes are most likely to be observed when the radar path is perpendicular to the earth's magnetic lines of force passing through the aurora. Chapman [18] has shown that the echo points which fulfill this perpendicularity requirement lie on a cubic surface if the earth's magnetic field is treated as a centered dipole field. Radio waves emitted at different elevation angles along the magnetic meridian from any given station will fulfill this perpendicularity condition at different heights and ranges. The values of height and range to these perpendicular points as a function of elevation angle have been given for various magnetic co-latitudes by Chapman [18].

His results show that as the magnetic latitude is increased from 50 degrees, the range in height and in elevation angle over which this perpendicularity requirement can be fulfilled is steadily reduced, and that at latitudes greater than 64 degrees this requirement can only be fulfilled at heights less than 90 km. Hence, if this perpendicularity requirement is strictly imposed, the height of the echoes may be expected to decrease with increasing magnetic latitude.

This prediction, however, must be modified by the fact that deviations from this perpendicularity requirement of several degrees are common. Also, visual aurora are rarely seen at heights less than 80 km. The combination of these effects tends to smooth out this change of height distribution with latitude, though it is probable that the height distributions will show a tendency to move to lower heights as the observing station is moved to higher latitudes.

Range Distribution of Auroral Echoes

Factors Determining the Range Distribution: While the expected range distribution of auroral echoes along the magnetic meridian for any radar is, of course, affected by such factors as the vertical polar diagram of the antenna and mean distribution of auroral activity across the region of the ionosphere illuminated by the radar beam, three other important factors must be considered. These are respectively: the minimum height of occurrence of aurora, the curvature of the earth, and the aspect sensitivity of the auroral ionization.

Visual observations have shown that aurora rarely comes down to heights less than 80 km, a result which is in good agreement with the radar observations. This fact, after allowing for the vertical polar diagram, the aspect sensitivity of the aurora and the curvature of the earth, therefore determines the minimum range of the aurora. The radar observations indicate that most of the auroral echoes are obtained from the region below 120 km. The curvature of the earth therefore imposes a maximum value for the range of auroral echoes. Owing to the difficulty at vhf of obtaining adequate coverage at very low angles of elevation (due to ground reflections) the maximum range is usually somewhat less than this limit.

The actual distribution of echoes between these two limits is largely determined by the aspect sensitivity of the auroral ionization. As indicated in the third part of this section, auroral echoes are most likely to be observed when the radar path is perpendicular to the earth's magnetic lines of force in the scattering region. The radar observations have shown that most auroral echoes originate at a height of about 100 km; other things being equal, one would therefore expect a range distribution which peaked at that range which corresponded to the perpendicularity requirement being fulfilled at a height of about 100 km. However, various other factors, including the vertical polar diagram of the antenna, the decrease of echo strength with range, the possible nonuniform distribution of aurora across the region of ionosphere illuminated by the radar beam, and the increase of volume of ionosphere illuminated with range, tend to modify this range distribution.

Since the perpendicularity requirement is likely to be less severe at low frequencies, the range distributions are likely to be decreased in width as the frequency is increased. Such an effect has been reported by McNamara and Currie [29].

Experimental Observations: Many workers have published information concerning the distribution of their auroral echoes in range. In general, echoes were reported as occurring in the range 300–1100 km. A comparison of the observations of the different groups indicates that the most frequently observed range increases with increasing frequency from $\sim$500 km at about 30 mc [2, 8], to significantly higher values ($\sim$750 km) at frequencies in the range 50–110 mc [12, 15, 23, 29]. The actual values obtained depend to some degree upon the magnetic latitude of the observing station.

The problem of the change of auroral response with range towards the magnetic pole has been investigated theoretically by Booker [20], who presents curves showing the variation of auroral response as a function of range and ionospheric absorption for certain operating parameters. His curves show that the median range at which echoes will be observed (assuming a uniform distribution of aurora with latitude) increases with increasing radio frequency in agreement with the experimental observations described above.

Range Drifts of Auroral Radar Echoes

Many observers have reported the occurrence of echoes whose ranges change with time, for example [3, 5, 7, 17, 23]. The fullest investigations appear to be those of Aspinall and Hawkins [17] and Bullough and Kaiser [23] who have described the changes in range of auroral echoes observed at Manchester. Many echoes were found to remain at approximately constant range, but others showed radial velocities between +2.8 km per second and −1.4 km per second. These observations were taken with a directional antenna beamed at low elevation at an azimuth some 50 degrees west of the magnetic meridian, and revealed an interesting variation of radial velocity with local time. The results showed considerable scatter, but the average radial velocity was found to decrease from about +0.5 km per second at 1800h to about −0.5 km per second at 0600h. These radial velocities were interpreted by Bullough and Kaiser as being the components, along the line of sight, of the motion of scattering centers along a line of magnetic latitude. If this interpretation is correct, then the motion of the auroral ionization is westward in the early evening, decreases to about zero at 2100h, and is eastward during the period 2200h–0600h.

If one accepts the above model of auroral scatterers moving along a line of magnetic latitude then the observed radial velocities would be a function of antenna azimuth. The magnitude of the observed radial velocity would be roughly proportional to the sine of the azimuth angle, measured from the magnetic meridian, and would reverse as the antenna was moved from one side of the meridian to the other.

The change in direction of motion at about 2100h apparently corresponds with the minimum in the observed diurnal distribution of auroral echoes at temperate latitudes. Since this evening minimum does not occur at high latitudes, it may be that no such reversal in direction would be expected at the higher latitudes. Observations at College by Dyce [15] indicate that the radial velocities of the scattering centers were often high (radial velocity up to 6 km per second) but showed no simple relationship with time of day.

It should perhaps be emphasized at this point that these motions are not attributed to the drift motion of the individual free electrons, but to the motion of the point of impact of the source of ionization.

Frequency Spectrum of Aurorally-Scattered Radio Waves

One of the most distinctive characteristics of vhf auroral echoes is their rapidly varying amplitude, which has the effect of broadening the frequency spectrum of a cw signal from zero to several hundreds of cycles per second. This phenomenon has been investigated by Bowles [26] who determined the frequency power spectrum of a 50 mc aurorally propagated cw signal. His results showed the presence of frequency components in the fading signal of roughly equal strength from zero frequency to a cutoff frequency between 100 and 200 cps. Assuming that this fading is due to interference between signals scattered back to the receiver from moving regions of auroral ionization, calculations show that the velocities of these scatterers must be of the order of 500 meters per second or about ten times normal E region drift velocities.

The above work by Bowles refers only to the broadening of the frequency spectrum of a cw signal during auroral propagation. In later work [14, 31], Bowles made use of combined cw and pulse operation to investigate the Doppler shift as well as the Doppler broadening of the reflected signal. A small leakage signal from the cw transmitter was picked up in the receiving antenna, and used as a reference frequency against which the frequency of the echo signal could be compared. This work showed that the echo signal was often displaced at the higher frequencies by as much as several hundred cycles. Simultaneous observations on two frequencies showed that the frequency shift was proportional to the observing frequency, thus confirming the Doppler origin of the phenomenon. Typical values of frequency shift were ±100 cps at 25 mc at College, Alaska, and ±100 cps at 50 mc at Ithaca, N. Y.

One important fact discovered by Bowles was that the magnitude and sign of the Doppler shift was not correlated with the changes in range of the auroral ionization, as would be expected if the target were a moving metal sheet. Thus, on several occasions, the Doppler shift from an approaching echo was negative, or was positive for a receding echo. An equally significant observation was the fact that positive Doppler shifts (increases in mean frequency) were associated with visually homogeneous forms, whereas rayed forms showed negative Doppler shifts.

The interpretation of these Doppler shifts has been discussed by Bowles [14], who suggests that the most plausible explanation is that they are due to the mean motion of the free electrons along the lines of magnetic force. On this interpretation, these motions must be downward in homogeneous forms and upward in rayed forms, with velocities of the order of 20 km per second. The lack of correlation between the observed Doppler shifts and the changes in range can be explained by considering the scattering area as composed of many discrete scattering centers whose position in space is approximately fixed, but whose average position (center of gravity of all the scatterers) moves as the source of ionization (the incident corpuscular stream) sweeps across the sky.

The Amplitude Distribution of Auroral Echoes

The amplitude distribution of auroral echoes has been investigated by Bowles [28] who recorded the rapidly fading auroral echoes photographically. The results were expressed in terms of amplitude density plots (histograms of the occurrence of a given echo strength) and in terms of the time auto-correlation function of the fading.

It was found that about half of the experimental amplitude density plots were of the Rayleigh type. This type of amplitude distribution is obtained when the received signal strength is the sum of many scattered signals uncorrelated in phase. Under such conditions, there is a relatively high probability of obtaining low values of signal strength. A second common type was one rather similar to the Rayleigh distribution, except for the occurrence of too many high values of signal strength. As would be expected, when a steady leakage signal was introduced by ground wave, a normal distribution was obtained.

The time auto-correlation functions for these records were found to divide into three classes. About half of the experimental auto-correlation plots were found to be exponential in type, *i.e.*, of the form

$$\rho(\tau) = e^{-\beta|\tau|}.$$

The remainder were of a similar type given by

$$\rho(\tau) = e^{-\gamma|\tau|}(1 + \gamma|\tau|)$$

with the exception of one of the Gaussian form $\rho(\tau) = e^{-\alpha^2\tau^2}$.

The Latitude Distribution of Auroral Echoes

So far as is known, no systematic series of observations to determine the latitude distribution of auroral echoes has been made, although such measurements are expected to be made in Alaska during the International Geophysical Year.

It has been found (see the second part of this section) that auroral echoes are closely associated in space with the visual auroral forms, and it is therefore possible to use the visual data to indicate how the probability of echoes will vary with latitude. Thus, it is known that auroras are most frequently seen overhead at a magnetic latitude of about 67 degrees. Bearing in mind that auroral echoes are most readily obtained when the radar is directed towards the magnetic pole at low angles of elevation, and that echoes are therefore observed most frequently at ranges of about 700 km, the zone of most frequent detection of aurora would appear to lie at about magnetic latitude 60 degrees. Echoes may be expected to occur with lower frequency at other latitudes because the perpendicularity requirement (see the third part of this section) is not fulfilled at the heights at which the auroral scatterers most frequently occur. These effects are clearly brought out in Fig. 8 of [18].

The Effective Scattering Areas of Aurora

It is possible to calculate the "target area" of an aurora (in terms of the equivalent isotropic scatterer) if the signal/noise ratio of the echo and the other operating parameters of the equipment are known. Calculations of this type for various types of equipment have been made by Dyce [15], who suggests that the equivalent target area at 50 mc is of the order 30 square km during strong aurora. It is also possible to calculate the "effective reflection coefficient" of the aurora, defined as the ratio of the echo strength actually received to that which would be received by specular reflection from a large metal sheet at the distance of the auroral scatterers. Using this definition, the maximum effective reflection coefficient observed during strong aurora is of the order of 10^{-4} at 100 mc, indicating that almost all of the vhf radiation incident upon the aurora is transmitted, only a small fraction being reflected by the auroral ionization.

III. Radar Echoes from Meteors

Introduction

When a meteor enters the earth's atmosphere, it forms a trail of ionization at a height of about 100 km. Under certain circumstances, this trail is capable of reflecting back to the earth's surface a sufficient proportion of the incident radio energy to enable the trail to be detected by radio means. Despite the transient nature of the echoes, this technique has proved to be a very valuable one for studying both the astronomical and the physical aspects of meteors and meteor trails. The purpose of this section is to describe the present state of knowledge of the main radio characteristics of the meteor echoes at frequencies above 30 mc; the section is not concerned with the physical or the astronomical implications of the observations.

Early Observations of Radio Wave Scattering from Meteor Trails

The discovery by Eckersley [32] in 1929 of short-lived bursts of signal within the normal skip zone was followed by work by Schafer and Goodall [33] and Skellett [34], who showed that these transient echoes were due to meteoric ionization. No major investigation of these meteor echoes (strictly, meteoric trail echoes) was made until 1946 and 1947, when Hey and Stewart [35–37] convincingly demonstrated that these transient echoes were meteoric in origin, by measuring their height of occurrence and their velocity. Other investigations of meteor echoes were made at about the same time by Appleton and Naismith [38] and by Prentice, Lovell, and Banwell [39].

A valuable summary of this early work has been given by Lovell [40].

The Aspect Sensitivity of the Meteor Trails

As the meteor rushes into the earth's upper atmosphere, it forms a straight line trail of ionization. As suggested by Pierce [41] such a trail may be expected to be markedly aspect sensitive, in that radar echoes should be strongest when the radio waves are incident upon the trail from a direction perpendicular to the trail (specular reflection). The presence of such an aspect sensitivity has been demonstrated by Hey and Stewart [35, 36] and by Lovell, Banwell, and Clegg [42], and forms the basis of almost all the radio determinations of the direction of travel of the meteors, *e.g.*, Clegg [43].

This perpendicularity requirement applies only to the case of a radar in which the receiver is close to the transmitter; in the case of communication between two separated points, the requirement is that there shall be a length of the trail for which the angle of incidence from the transmitter is equal to the angle of reflection towards the receiver.

The Height of the Meteor Echoes

The height of the meteor ionization responsible for meteor echoes has been investigated experimentally by Hey and Stewart [36], and by Clegg and Davidson [44] in England; by Millman and McKinley [45, 46] in Canada, and by Volmer [47] in France. The methods used by these observers differed very considerably; in general, however, the results obtained were in good agreement. For each equipment, meteors were detected at heights ranging from about 80 km to about 110 km, the most probable height being about 95 km.

Later work by Evans [48] has shown that the height of the echoing point increases with meteor velocity from about 89 km for a velocity of 22 km per second to about 101 km for a velocity of 53 km per second. The theory of the meteor height distributions for shower and sporadic meteors has been discussed by Kaiser [49, 50]; using this theory Evans has been able to determine the variation of scale height over the range 88–101 km, and also the variation of atmospheric density and pressure with height over that range.

Range Distribution of Meteor Echoes

The distribution in range of the meteor echoes obtained by a given equipment is chiefly determined by the vertical polar diagram of the radar and by the distribution in height of the meteor trails. Hey and Stewart [36] have shown with a vertically directed antenna that meteor echoes are rarely observed at heights less than 75 km; this therefore places the minimum range at which meteors will be observed as about 75 km. The maximum height at which meteor echoes are obtained is about 120 km; owing to earth curvature, this height imposes a maximum value of about 1200 km for the range of meteor echoes, even if the radar waves leave the earth's surface at very low angles of elevation. The difficulty of radiating vhf radio waves at low angles of elevation due to ground reflection effects often reduces this maximum range to less than 1000 km. The actual distribution of meteor echoes within these limits is chiefly determined by the vertical polar diagram of the radar, although other equipment parameters such as equipment sensitivity and azimuth of antenna can also have an effect.

Amplitude Characteristics of Meteor Echoes

The Head Echo: The main, specularly-reflected echo from a meteor trail is sometimes preceded by a short-duration echo which is apparently due to the head of the meteor trail, in that this "head echo" moves with the velocity of the meteor [37, 51–53]. Such echoes are relatively rare, particularly at the higher frequencies. The origin of this head echo has been discussed by Browne and Kaiser [54], who show that the intensity of the head echo (relative to the body echo) should be proportional to the radio wavelength and inversely proportional to the difference in range between the two echoes.

The Main Echo: It is not possible to describe the "typical meteor echo" with any quantitative exactness, because many of the characteristics of the typical echo are determined by the operating frequency and other parameters of the radar equipment concerned. Also, the echoes obtained on any one equipment vary very considerably with regard to amplitude, duration, and fading rate.

Most meteor echoes are of short duration at vhf radio frequencies. Thus, Prentice, Lovell, and Banwell [39] indicate that at 72 mc, with 150 kw pulses radiated from a single Yagi, over 80 per cent of the echoes lasted less than 0.5 second. However, several long-duration echoes were observed, particularly during the Perseid meteor shower. Some of these echoes had durations in excess of 20 seconds.

The behavior of the electron trails after formation has been discussed by Greenhow [55]. He divides the echoes he observed on wavelengths of 4.2 and 8.4 m into two groups: short duration (low-electron density) and long duration (trails in which the electron density is greater than the critical density for the wavelength considered).

The short-duration echoes tended to show an exponential decay of amplitude with time, the duration to half amplitude being about 0.1 second for a wavelength of 8.4 m. By comparing echoes obtained simultaneously on the two wavelengths, he was able to show that the duration to half amplitude was proportional to λ^2.

The remaining echoes (about 40 per cent for the Perseid shower) had much longer durations, though not markedly different amplitudes. These echoes showed considerable fading with periods between about 0.01 second and 0.1 second, the fading rate being proportional to the observing frequency. This fading phenomenon is associated with the distortion of the meteor trail by wind gradients and/or turbulence.

Both groups of echoes may show the rapid initial rise in amplitude, followed by oscillations of increasing frequency which are due to diffraction effects during the formation of the trail. These Fresnel zone oscillations can be used to determine the velocity of the meteor [56, 57], but only occur in about 10 per cent of the echoes.

Variation of Echo Power from an Individual Meteor Trail with Parameters of the Observing Radar

The echo power received by a radar equipment by specular reflection from a meteor trail whose electron density is less than the critical density, has been shown by Lovell and Clegg [58] to be given by the expression

$$\epsilon = 3.3 \times 10^{-28} P \left(\frac{\lambda}{R}\right)^3 (\alpha G)^2 \text{ watts,}$$

where

ϵ = peak echo power in watts
P = peak transmitted power in watts
λ = observing wavelength in cm
R = distance to meteor trail in cm
G = antenna gain (assumed equal for transmitting and receiving)
α = number of electrons produced per cm of path.

This expression has been shown to be in quantitative agreement with experimental observations of the short duration echoes. The above expression is limited to trails in which the electron density is less than the critical density.

The theory of radio reflections from meteor trails has now been extended to cover the long duration vhf meteor echoes whose electron line density is greater than about 10^{12}/cm. Probably the three most important papers in this field are those of Greenhow [55], Kaiser [59], and Eshleman [60]. Kaiser [59] shows that the short duration types, which are of relatively low electron density, give rise to an echo amplitude which is proportional to α, and that their duration is independent of α, being determined by the wavelength and the diffusion coefficient. For the long duration echoes (those with electron line densities greater than about 10^{12}/cm) the echo amplitude increases only as the fourth root of α, but the duration is proportional to α. In both cases the duration is proportional to the square of the wavelength and inversely proportional to the diffusion coefficient.

The Variation of Meteor Echo Rate with Radar System Parameters

An important factor in determining the average power received by a radar equipment from meteor echoes is the number of meteor echoes observed per minute. The effect of the operating characteristics of a vhf radar equipment upon the observed meteor echo rate has been discussed in a valuable series of papers by McKinley [61–64], and also by the Stanford group [60, 65, 66]. The meteor echo rates are a function of a very large number of variables, including power transmitted, operating wavelength, elevation angle of radar beam, azimuth of radar beam, geographical location of radar, polarization of antenna, receiver sensitivity, time of day, and time of year, etc. These variables are often inter-related, and it is not possible to summarize their individual significance in a report of this size. The reader is therefore referred to the original papers indicated above, and to a review article on meteor echoes which (we understand) is currently being prepared by the Stanford workers.

Polarization Phenomena in Meteor Echoes

In 1948 Herlofson [67] showed that a polarization phenomenon might exist in the echoes from meteor trails. He showed that a narrow trail with high electron density would act as a resonator, giving stronger echoes for waves polarized with the electric vector normal to the trail than for the case when the electric vector is parallel to the trail. The theory has been extended by Kaiser and Closs [68] and by Eshleman [69].

Experimental observations of this phenomenon have now been made by both the Manchester and the Stanford groups [70–72]. The two sets of observations are in good agreement with each other, and with the theoretical investigations by Kaiser and Closs [68] and Eshlemann [69], who showed that the amplitude ratio between the two polarizations, assuming a Gaussian distribution of electrons across the trail, would reach a maximum resonance value of 2.0.

The Motion of the Meteor Trails

The drift motion of the meteor trails after formation has been described by the Manchester, England, group [55, 73–75]; by the Stanford, Calif., group [76–78] and by workers in Australia [79, 80]. Various techniques have been used, including direct measurements of the changes in range using an A-scope presentation and also Doppler techniques with coherent pulse radar or cw transmission.

This work has shown that the meteor trails move with a drift velocity usually in the range 10–100 meters per second. Measurements of the velocities of individual meteors occurring almost simultaneously in the same area of the sky showed very considerable scatter; however, when these values were averaged over a period of one hour, it was found that marked semidiurnal and diurnal variations of the direction of motion existed. These semidiurnal and diurnal variations were superimposed on a prevailing wind motion, which was found to vary from month to month [75]. A positive wind gradient of about 2.7 meters per second per km increase of height was also found.

Meteor Echoes at High Latitudes

Very few observations of meteor echoes have been made at high latitudes. The only published comment is that of Bowles [12], based on observations made with the 106 mc radar at the Geophysical Institute of the University of Alaska. He reported that the rate and strength of meteor echoes was considerably enhanced during periods of aurora. Contrary to the auroral radar echoes, these meteor echoes showed no preferred azimuth; they also showed fading rates about one order of magnitude less than the fading of the auroral echoes.

More recent (unpublished) observations at the Geophysical Institute failed to indicate any significant increase of meteor echo activity during aurora. It is possible that the reason for this discrepancy lies in the use of very different radar equipments, since the 106 mc equipment used by Bowles had approximately 20 times the peak power and 6 times the antenna gain of the 50 mc equipment used more recently. The 106 mc equipment was therefore capable of detecting smaller meteors.

It is hoped that it will be possible to continue this investigation at the Geophysical Institute.

IV. The Faraday Effect and Radar Echoes from the Moon

Since the first radar moon echoes were obtained by the U. S. Army [81] and by Bay [82] in 1946, many such echoes have been secured. Work has been carried on in Australia [83, 84], England [85], and the United States [86, 87]. Theoretical consideration has also been given to the possibility of obtaining echoes from the sun and planets [88, 89].

The various investigators used different transmitter powers and frequencies. Dewitt and Stodola [86] used 15 kw at 111.5 mc in the army tests in 1946; Kerr, Shain, and Higgins [83] used 50 kw at 17.8 mc, and 70 kw at 21.5 mc, and Murray and Hargreaves [85] used 2 kw on 120 mc. Sulzer, Montgomery, and Gerks [87] used 20 kw at 418 mc in their telegraphic relay.

The radar pulse was generally of the order of a second in length, though Murray and Hargreaves [85] used 30 milliseconds. The receivers were all of low-noise design, with a final bandwidth of 30 to 100 cps. The ellipticity of the moon's orbit and the rotation of the earth produce a Doppler effect which may amount to as much as 100 cps. With the very narrow bandwidth receivers used, this frequency shift requires careful retuning of the receiver. High gain antennas were used in all cases, some having a gain of 3000 to 5000 over an isotropic radiator.

Most workers report two types of fading. One, with a period of seconds, is believed to be due to the moon's libration; the second, with a period of about an hour, has been observed by most workers. Murray and Hargreaves [85] found that this type of fading was most intense a few hours before sunrise and continued throughout the daytime observations. This and other evidence [34, 84] points to ionospheric effects as the cause.

Murray and Hargreaves [85] suggested that this long period fading is due to the Faraday effect, *i.e.*, to the fact that the plane of polarization of the radio waves is rotated as they traverse the ionosphere in the presence of the earth's magnetic field. (This rotation is due to the splitting of the incident wave into its ordinary and extraordinary components, which travel through the medium at slightly different velocities.) To confirm that the long period fading was associated with polarization changes, they used a paraboloid receiving antenna with a primary feed consisting of two dipoles at right angles. The receiver input was switched from one to the other at one-minute intervals. During a 45-minute period of echo reception, the change of polarization of the received wave was such that signals were received first on both dipoles, then only on one, and later on both dipoles again.

The army experiment [81, 86] reported instances where no echoes were received, even though the equipment seemed to be functioning properly. This occasional failure to detect echoes may well have been due to the Faraday rotation of the plane of polarization of the radio wave.

Murray and Hargreaves [85] give the following expression for Ω, the total polarization shift, created during the double passage through the ionosphere:

$$\Omega = \frac{7490H \cos \theta}{f^2} \int N dr \text{ complete rotations,}$$

where θ is the inclination of the geomagnetic vector to the line of sight, f is the transmitted frequency in cps,

H the geomagnetic field in emu, N the electron number density per cc, and dr cm the element of path length along the line of sight between the antenna and the moon where the electron density is N. Substitution shows that in order to rotate the polarization through 90 degrees at a frequency of 120 mc, a change in mean electron density of about 5×10^4 electrons per cc over a path length of 200 km is required. Such changes of electron density in the F region are likely to occur during the transition from nighttime and daytime conditions.

The total number of complete rotations of the plane of polarization undergone by the radio waves in the double passage through the ionosphere is not yet fully known, though it is understood that numerical values of this quantity have recently been obtained at the University of Manchester [90]. These results were obtained by comparing the rotations of the plane of polarization on two closely-spaced frequencies.

The application of moon radar to ionospheric research may be expected to lead to important advances in our understanding of the outer regions of the atmosphere. In particular, this new technique offers a valuable means of determining the electron densities above the level of maximum ionization.

V. Radio Noise of Auroral Origin

General Considerations

Ionized gas in which electron-ion or electron-molecule collisions occur would be expected, on purely classical grounds, to generate electromagnetic waves by the mechanism of charge acceleration. The flux of energy escaping from such a region will be determined by the relative importance of generation and absorption effects taking place within the volume. For a large gas volume, under conditions of thermal equilibrium between the electrons and the gas molecules, the amount of radio-wave energy, E_f, lying in a one-cycle band centered on a wavelength λ meters and proceeding outward through a unit of solid angle is given by the Rayleigh-Jeans approximation to the black-body radiation law, *viz.*: $E_f = 2KT/\lambda^2$ (watts meters^{-2}, (CPS)$^{-1}$, steradians^{-1}). K is Boltzmann's molecular constant.

As an example of the magnitude of such thermal radiation, it may be noted that at a wavelength of 1 m and for a temperature of 6000°K, the value of E_f is 1.66×10^{-19} w. As applied to regions of auroral discharge, available data [91, 92] on the temperatures of the gas molecules (on the basis of the broadening of spectral line widths) indicate values not significantly different from the normal atmospheric values at those heights. Detectable amounts of such thermally-generated radiation in the vhf region thus appear unlikely from regions of visible aurora.

If the region occupied by aurora is regarded as a gas permeated by a relatively high-velocity flux of electrons and protons coming from the sun, not in gas-kinetic equilibrium with the atmosphere being penetrated, the possibility of radiation corresponding to considerably higher equivalent black-body temperatures exists. Thus, electrons approaching the earth at 1500 km per second, corresponding to the rate of passage of aurora producing corpuscles from sun to earth, could exhibit electron temperatures equivalent to many tens of thousands of degrees.

Other mechanisms involving more ordered motions of charges can be conceived that would constitute generating mechanisms for electromagnetic waves at radio wavelengths. In particular, in regions of plasma, *i.e.*, where the free charge present per unit volume as electrons is equal to the free charge density present as positive ions, the possibility of plasma oscillations occurring as a result of transient perturbations or local concentrations of the distribution of charges exists. Jaeger and Westfold [93] have discussed this phenomenon with particular reference to the solar corona, a medium in which the electron density decreases toward the observer. They point out that there are three characteristic features of the spectrum:

1) The intensity distribution will show a cutoff below the so-called plasma frequency, numerically given by $f_0^2 = 81N$, where N is the electron concentration per cubic meter. (There will be a maximum of energy emitted at a frequency just about f_0, with the remainder of the curve conforming to a law depending on the spectrum of the exciting transient.)

2) The disturbance observed on different frequencies will show various time delays due to selective retardation, high frequencies preceding low frequencies.

3) The observed disturbance on any one frequency will decay with a time constant equal to the reciprocal electron collision frequency. (While these conclusions apply to a medium without magnetic field, it appears that the behavior with a field will be broadly similar, though there will be two magneto-ionic frequency components to be considered.) This theory which seems to explain many features of solar corona emission, might apply to the regions of ionized aurora. The most obvious perturbing mechanism would be fluctuations in the density or velocity of the corpuscular stream presumed basically responsible for auroral phenomena. The question of whether or not these transient waves in the medium are actually radiated cannot be given an unambiguous answer on the basis of present knowledge, but the possibility exists. If auroral radar echoes at 100 mc and up are due to deviative refraction by clouds of electrons, then plasma oscillations are possible up to the frequency limit of observable radar echoes; and their frequency spread will depend on the spread in electron concentration existing in the ionized aurora.

Experimental Results

The only positive observations of radio radiation from aurora is comprised of the work of Forsyth, Petrie, and Currie [94]. In 1949, observing on 3000 mc, they observed short duration (1–5 μsec) pulses arriving in bursts of short duration on the visual indicator of a radar set. The auroral origin was confirmed by observations of some 20 auroras, with associated photography of the parts of the display to which the receiving antenna was pointed. Some correlation between frequency of pulses and the intensity and type of aurora were noted. The origin of these bursts was considered by these authors [95] as plasma oscillation of the ionized region associated with the auroral display. From a knowledge of the constants of the equipment used, it was deduced that the power density at the receiving antenna was at least 2×10^{-10} w/m².

Subsequent investigation of the phenomenon by the other workers, notably Harang and Landmark [2] and Chapman and Currie [97] has, however, failed to detect the effect. The last mentioned workers believe this may be an effect tied up with the solar sunspot cycle, because in the 1949 observations the effects were only noted when the sunspot number exceeded about 120. The failure to detect the effects in 1951–1952 may thus be attributed to the generally decreased auroral and sunspot activity since 1949.

At frequencies of 65 mc during 1954 and 30 mc during 1955, no clear-cut indications of radio noise associated with auroral displays have been found on the low-noise figure receiving equipment used by Little [98] at College to monitor the absorption of extraterrestrial radio signals. The mean power level of the galactic noise background, above which any auroral noise would have to have risen about 50 per cent to be detected, was about 2×10^{-14} w over a band 30 kc wide centered at 30 mc. The recording instrument has a full-response time of 0.5 second, which would seriously discriminate against the detection of short-duration noise bursts of the type reported by Forsyth.

On frequencies of approximately 4, 8, and 12 mc, monitored continuously at College, no effects consistently relating background noise and visual aurora have been found although it is not uncommon for both noise and signal to increase on the 12 mc circuits during auroral displays. However, in the high frequency region it is particularly difficult to be certain that the increased background noise is not due to aurorally-propagated signals from distant transmitters or to man-made interference. Until further and more careful work is done, the most likely explanation of the observed increases in antenna signal power during aurora at College appears to be that they are due to the auroral propagation of man-made transmissions.

VI. Absorption of Radio Waves by the Ionosphere

Introduction

When a radio wave encounters a region in the atmosphere containing "free" electrons, it experiences two principal effects. Firstly, there is a change in the velocity and, in general, in the direction of travel. Secondly, the energy of the wave is, under certain circumstances, dissipated in the form of weak local heating of the atmosphere. The first mentioned effect, corresponding to a change in refractive index of the medium, is responsible for the bending and return to earth of radio waves from the upper atmosphere, *i.e.*, ionospheric wave propagation. The second effect, referred to as "absorption," manifests itself on radio circuits as attenuation of the radio signal over and above the attenuation to be expected from the geometrical divergence of a pencil of rays proceeding outward from a point source. This section is concerned with some of the details of the mechanism of wave energy absorption, experimental techniques for its observation and measurement, and some of the results so far obtained by various investigators.

It is shown from the basic theory that very little absorption is to be expected, in general, in the vhf region. However, since little direct experimental work has been carried on at vhf, it will be necessary to discuss the results of high frequency measurements and make suitable extrapolation to the vhf range. The effects are shown to be of usually negligible magnitudes.

Basic Mechanism

When a radio wave passes over any charged particle, that particle is subjected to an alternating driving force —electrical in nature and of magnitude equal to the product of the amount of charge of the particle and the maximum electrical field intensity of the wave. The particle, if completely free to move about, responds by accelerating and decelerating as the wave field alternates and thus engages in an oscillating motion with an amplitude governed directly by the strength of the wave field and inversely by the inertial property (mass) of the particle in question. The passing wave thus forces all the free charged particles in the region being traversed to engage in an oscillatory motion at the wave frequency, and the accelerated charges reradiate and generate a new radio wave in the manner illustrated by Huyghen's principle in optics. Thus the energy extracted from the incident radio wave and converted to the energy of motion of the charged particles is not lost from the wave, but it is recovered in the process of reradiation. In the atmosphere the only charged particles of concern at radio wave frequencies are electrons since the positive ions in general are several thousand times more massive and hence remain relatively inert in the passing wave field.

Radio wave absorption is thus primarily concerned with the presence of electrons in the atmosphere, and with any process interfering with the complete efficiency of the conversion process which converts the wave energy from field energy to kinetic energy and back into field energy. Thus any process which acts to extract energy from the kinetic energy stage decreases the conversion efficiency and wave absorption is observed.

The only process other than reradiation which can extract significant amounts of energy from the moving charges is that involving chance collisions between electrons and neighboring neutral gas molecules (or positive ions). If this happens, the electrons recoil with less than their incident energy, having given much of it to the heavier particle involved. The neutral gas molecules will have been slightly speeded up, *i.e.*, the temperature of the gas will have been increased, and although no net energy has been destroyed or created in the process, a certain amount of the radio wave energy has been converted to a form from which it is not possible for it to reradiate completely at the original wave frequency. The frequency of electron-molecule collisions therefore is an important parameter affecting the degree of absorption.

Quantitative Theory

The quantitative theory of dissipative attenuation has been discussed by many workers since the publication of the Appleton-Hartree magneto-ionic theory [99–102]. This theory shows that, providing the ray path is not perpendicular to the earth's magnetic field, K, the absorption coefficient per unit path length can be written, in rationalized mks units, as

$$K = \frac{e^2}{2mc}\cdot\frac{1}{\eta K_0}\cdot\frac{N\nu}{\nu^2 + (\omega \pm \omega_L)^2} \qquad (1)$$

where K_0, e, m, and c have their usual significance, η is the appropriate refractive index, N the number of electrons per unit volume, ν the frequency of electron-molecule collisions, ω the angular wave frequency, and ω_L the gyromagnetic angular frequency corresponding to the longitudinal component of any static magnetic field existing in the region. The positive and negative signs refer to the ordinary and extraordinary modes of wave propagation in an ionized medium pervaded by a magnetic field, with the plus sign going with the ordinary mode.

Since the gyromagnetic frequency is of the order 1.5 mc, its effect can usually be ignored at frequencies well above 30 mc. At 30 mc, however, $\omega+\omega_L$ and $\omega-\omega_L$ differ by about 10 per cent. These quantities appear squared in the denominator of (1), and hence the attenuation of the extraordinary wave (measured in db) is approximately 20 per cent greater than that of the ordinary wave, provided that ν is small compared with 30 mc. Under extreme Arctic conditions, the attenuation of 30 mc radiation may exceed 15 db for a single passage through the ionosphere at vertical incidence; under these extreme circumstances, a significant polarization effect could be introduced as a result of the differential absorption of the ordinary and extraordinary waves.

With respect to the collisional term, ν, rocket flight determinations of pressure and temperature of the atmosphere reported by Havens, Koll, and LaGow [103] combined with the experimental studies of Crompton, Huxley, and Sutton [104] on the collision processes between slow electrons and air molecules make possible the assignment of the following numerical values for ν at various heights. (In Table I the values are shown to be computable from the relationship $\nu = 9.36\times10^7\ P_{mm}$ of mercury.)

TABLE I

h_{km}	ν/sec
65	10^7
80	10^6
95	10^5

Thus for vhf, $\omega^2 \gg \nu^2$, which further simplifies the absorption expression. Subject then to the assumptions that negligible selective absorption of the magneto-ionic modes occurs and that wave frequencies are greatly in excess of collision frequency, the absorption coefficient becomes more simply expressible as

$$K = \frac{e^2}{2mcK_0}\cdot\frac{N\nu}{\eta\omega^2} = \frac{\nu f_N^2}{2c\eta f^2} \qquad (2)$$

where f_N is the "critical frequency" associated with an electron concentration N, *i.e.*,

$$f_N = \frac{1}{2\pi}\sqrt{\frac{Ne^2}{K_0 m}} \qquad (\approx \sqrt{81N}) \qquad (3)$$

if N is electrons/m^3,

$$K_0 = \frac{1}{36\pi}\times 10^{-9}\text{fds}/m,$$

and f in kc. This expresses the fact that the fractional absorption per unit path length varies directly with the product $N\nu$, and inversely as the squared angular wave frequency, ω^2. Further the absorption is inversely dependent on the refractive index, η, and is thus directly dependent on the time spent by the signal in traversing unit path length in the medium. For wave frequencies approaching f_N, η approaches zero, rays are strongly deviated and absorption rapidly rises—approaching infinity in the limit. For wave frequencies far removed from f_N the value of η approaches the value unity according to the expression

$$\eta = \frac{1}{\left[1 - \left(\frac{f_N}{f}\right)^2\right]^{1/2}}.$$

Thus at frequencies in the vhf range it is unlikely that the deviative type of absorption (*i.e.*, $\eta\to0$) will occur since normal ionospheric soundings rarely reveal f_N values greater than 12 mc in the regular ionospheric layers.

Returning to the $N\nu$ term which, along with ω, governs the nondeviative absorption index, the collisional term ν theoretically involves the gas pressure and temperature of the region. Assuming constant temperature, ν falls exponentially with increasing scale-heights according to the expression $\nu = \nu_0 \exp{(h_0 - h)/H}$, where H is KT/mg and K is Boltzmann's molecular gas constant, m the molecular mass, T the absolute temperature, and g the gravitational acceleration constant.

Experimentally, from the values given in Table I it is seen that ν decreases by about one order of magnitude for height increases in steps of 15 km above 65 km and up to E-region height. This implies a scale-height value of about 7.7 km and is thus virtually the same as exists at the earth's surface. Rocket flight observations give a daytime electron concentration, N, value of about $10^9/\text{m}^3$ which changes only slightly with elevation over the 165 to 95 km range. The product $N\nu$ appears therefore governed in large measure by the exponentially rising ν values with decreasing height. It is most unlikely that this increase in ν will be exactly compensated by the falling N values and thus some sort of a broad maximum in $N\nu$ presumably exists at some region below that of the normal E layer. (Height of 100 km.) The D region of the ionosphere seems thus indicated as the most probable region.

To apply the absorption coefficient K to the evaluation of total absorption over a given path, we note that for unit path length the field intensity relationship is $E = E_0 \exp^{-K}$. Thus $E = E_0 \exp \int -K ds$ where the limits of the path are the integration limits. When N, ν, and η vary in complicated fashion along a path, it is necessary to use numerical integration to solve for the total dissipative attenuation.

Radar echoes and vhf radio propagation supported by auroral curtains of ionization are well established phenomena observed by several groups of workers [2, 5, 105]. Abnormally high electron densities up to values of the order of 10^{14} per m^3 have been suggested in auroral ionizations, and strong absorption of the deviative type may thus be possible, since this corresponds to f_N values in the neighborhood of 100 mc. Meteor trails of ionization are also known to possess high electron concentrations, as evidenced by radar echoes and the work of Lovell [58], Manning [106], and others [107]. Linear trail densities of the order of 10^{12} electrons per cm of path length for visual first-magnitude meteors are reported by Lovell and Clegg, and meteoric radar echoes on 100 mc are relatively common. Strong deviative-type absorption may thus be possible at vhf over the rather restricted geometry and time duration of auroral forms and meteoric ionization trails, if the wave frequency were below 100 mc and chanced to lie close to f_N of the region traversed.

An alternative explanation of auroral echoes, held by Booker [20] suggests that since the measured reflective power of auroral radar targets is only about 10^{-4}, the echoes can be explained in terms of partial reflections from relatively small-scale irregularities, of the order of 1 per cent, in an electron concentration of only about 10^{12} per m^3. If this interpretation is correct, the corresponding value of f_N [*i.e.*, (3)] is only about 9 mc and no possibility exists for strong absorption of the deviative type suggested by the previous paragraph. Assuming an auroral discharge at 95 km height, an f_N value of 10 mc, and a collisional frequency (Table I) of 10^5 per second, a signal at 100 mc applying (2) shows a total nondeviative absorption of only 0.14 db over a path 1 kilometer long.

Observational Methods

To measure absorption directly involves measurement of E, the field intensity with absorption present, and E_0 the unabsorbed field value. While it is in principle possible to calculate the unabsorbed field under certain conditions, it is a matter of considerable practical difficulty. A better approach is afforded if it is possible to actually observe the signal E_0, on the same equipment used to indicate E, *i.e.*, under conditions of low or zero absorption. This is approximated under nighttime (low N) conditions with ionospherically reflected cw signals.

By using pulse techniques, the absorption can be obtained directly by comparing the indications of successive multiple reflections of a pulse. The absorption index is equal to the logarithm of the ratio of intensities of successive pulses, corrected for any known losses occurring at the reflecting surfaces involved. This method is directly applicable to vertical incidence ionospheric sounding of the upper atmosphere. Illustrative of this method is the recent work by Davies and Hagg [108].

Another method being exploited is the use of galactic radio noise signals as an extraterrestrial source of constant power for monitoring the transmissivity of the atmosphere at selected frequencies. The galactic noise power received at a given sidereal time will thus be constant except for absorption effects. Work of this sort has been reported by Mitra and Shain [109] using 18.3 mc. Blum, Denisse, and Steinberg [110] have made observations at 29.5 mc. More recently Little [98, 111] has used the technique at high latitudes at frequencies of 30 and 65 mc.

Qualitative information concerning absorption can be inferred from the lowest frequency of detectable echoes in the 1 to 1.5 mc frequency region on vertical ionospheric sweep-frequency recorders. The change of absorption with frequency is great in this frequency region because of the gyromagnetic resonance phenomenon, *i.e.*, absorption rises as ω approaches ω_L. The minimum frequency for just discernible E-layer echoes, $f_{\min}E$, is commonly used as the parameter for scaling the effect [112].

Finally, observations of the received field strengths of radio signals arriving via oblique paths involving known reflection points in regular ionospheric layers can in principle be used to calculate the total absorption over the path. The practical difficulties are however such as to seriously impair the accuracy of the results

[113], particularly situations in which the partial reflections from a lower layer give rise to apparent attenuation of the rays reflected from a higher one.

Recent Theoretical and Experimental Results

Frequency Dependence: Piggott [102], Appleton and Piggott [114], and Allcock [115] have carried out a thorough series of analyses of observations extending over a number of years at stations in England. They find that the variation of absorption with the inverse square of wave frequency, as predicted by theory, is valid to a remarkable extent. Expressing the absorption in the form $K=A/(\omega \pm \omega_L)^2$ the A factor for southern England (where $f_L=\omega_L/2\pi=1.2$ mc) comes out as 505 db. This value is found subject to considerable enhancement at frequencies near the critical frequencies of either the E or F layers, as was to be expected, corresponding to the deviative-type absorption encountered at the slow (group) velocity propagation. (See theory.) At higher frequencies, notably in the vhf region, this inverse frequency-squared dependency has not apparently been tested experimentally, but it appears qualitatively justified.

Localization: A great deal of work points to the height interval 65 to 95 km as the region where the principal absorption occurs. This result is to be expected from the known theoretical dependency of collision frequency ν on height. The recent experimental evidence for the preponderant importance of the D region in this respect includes the work of Shain and Mitra [116] which follows up earlier observations of Bracewell and Straker [117] on the sudden phase anomaly found in low-frequency waves being propagated through D regions during the onset of a Dellinger Effect (SID). Measurements by Gardner [118] show further, from observation of the changes in the ionosphere's thermal radiation (at radio frequencies), that the principal absorption occurs at or below 80 km during an SID. Additionally, the observation by Beynon and Davies [119] that both E and F layer obliquely-reflected radio signals are found to be almost identically affected by absorption changes observed simultaneously by vertical sounding methods implies that the absorption occurs at relatively low levels. A similar effect is reported by Chatterjee [120]. Further, Appleton and Piggott [114], Lindquist [121], Dieminger [122], Gardner and Pawsey [123], and Heppner [124] have reported that high absorption occurs in frequent association with unusually low-lying vertical-incidence sounding echoes of a sporadic nature from the height interval 65 to 95 km. This high absorption, together with the evidence of diurnally-varying ionization of presumed meteoric origin extending down to 85 km found by Forsyth [125] by back scatter of 56 mc signals, leaves little doubt that nondeviative absorption can be explained principally by the occurrence of excess ionization in the regions below normal E-layer heights.

Evidence that there is nevertheless some absorption ascribable to higher regions was obtained by Mitra and Shain [109], who found a correlation between the total absorption of zenithal galactic noise signals on 18.3 mc and f_0F_2 values above 5 mc. Analyses of a similar character carried out by Blum, Denisse, and Steinberg [110] likewise reveal a partial absorption attributable at 30 mc to the F regions—or beyond, since it is uncertain whether the excess attenuation occurs within the ionosphere or perhaps well away from the earth. A possible explanation of this in terms of a significant nondeviative absorption in the F regions has recently been put forth by R. Gallet [126] of the National Bureau of Standards. In a paper presented at the Western Electronic Show and Convention on August 19, 1955, he developed a theory based on electron-positive ion collisions in the F regions. This theory seems adequate to explain the observed absorption effects showing correlation with F-region ionization.

Causative Mechanisms, Solar Control: Several agencies are involved in the production of the ionization at D-layer heights responsible for absorption. The regular diurnal and seasonal variations of absorption appear to depend on the cosine of the sun's zenith angle through a relationship of the form $\alpha=\alpha_0(\cos\chi)^n$ where α_0 is the maximum or noontime value. Such solar angle dependence implies that solar radiation flux density is the agency concerned, and is in agreement with the usual formulations of the Chapman Law [127] of formation of an ion layer under certain simplifying assumptions. While a considerable amount of attention has been directed to this aspect of the subject, *cf.*, that of E. W. Taylor [128], Appleton and Piggott [114], and Rawer (129], it has become increasingly evident that the exponent in the relationship is a quantity which is quite consistently smaller than the theoretical value of 3/2 derived by Appleton. The exponent varies seasonally, with latitude, and with the solar sunspot cycle but with anomalies which lead to the conclusion that something other than solar radiation is in part responsible for the absorption.

Latitude Effects: At high latitudes, particularly, the solar influence weakens and yields increasingly to the effects of the primary agency closely associated with polar blackouts, magnetic disturbances, polar sporadic E layer, and auroral effects. This prime agency is generally presumed to be charged corpuscular streams emanating from the sun and magnetically guided into the earth's higher latitudes by the interaction between the moving charged particles and the earth's magnetic field. Davies and Hagg [108], for example, have analyzed absorption effects at Prince Rupert, British Columbia (54.3°N, 130.3°W), and find the exponent n expressing the influence of solar radiation reduced to a value of about 0.5 at this latitude. This result contrasts with Piggott's [102] value of 0.75 for the lower latitude observations made at Slough (Southeast England). A similar phenomenon is reported by J. C. W. Scott [130] in an investigation of the latitude variations of E and F layer ionization densities. He showed that the value of n in the relation $f_0E\alpha(\cos\chi)^n$ decreased from about 0.25 outside the auroral zone to about 0.1 inside the zone. This suggests a similarity of causative agency in the D and E regions, and it is interesting to note further that this similarity extends to the seasonal variation of the respective absorption and ionization indices, namely that n tends to be higher in summer than in winter. More detailed results are given by Cox and Davies [131] in an analysis of the diurnal variation of the minimum frequency reflected from the ionosphere on a string of four stations distributed from 50°N to 75°N.

Sunspots: Lange-Hess [132] has investigated the absorption-sunspot relations at several latitudes and has found close correlations in equatorial regions and at high latitudes, but with results at Slough quite irregular under wintertime conditions. In all three regions it was possible, under some conditions, to detect a 27-day periodicity which presumably arose from the familiar sunspot recurrence rate. Scott [130] has commented on similar sunspot vs critical frequency correlations in the North American high latitudes.

Polar Blackouts: Wells [133], Cox and Davies [131], Agy [134], as well as Lange-Hess [132], have reported on the phenomenon known as "polar blackout." This intense absorption effect, whose duration may run to many hours, is encountered with increasing frequency and intensity as the auroral zone is approached. Further, the absorption effects at high latitudes show a strong tendency to be associated with magnetically-disturbed conditions and with the occurrence of clouds of abnormal ionization observed as sporadic E and as "spread" F-region echoes particularly as shown by the work in North America by Heppner, Byrne, and Belon [124], and Cox and Davies [131], and observed by Lange-Hess [132] at European longitudes. V. Agy [135] and Meek [136], in particular, have localized a zone of "auroral absorption" from a study of data of radio wave field strength recordings along the 90th meridian, and show it to be approximately (within 6 degrees) in coincidence with the zone of maximum auroral frequency in North America.

Absorption measurements based on variations in the received intensity of extraterrestrial radio signals were made at 18.3 mc by Mitra and Shain [109], at 34°S 151°E, by Blum, Denisse, and Steinberg [110] using 29.5 mc at 49°N, 2°E, and by C. G. Little [98] using 65 mc and 30 mc at 65°N, 148°W. At the temperate latitudes, both south and north, these observations confirm the regular diurnal and seasonal variations found by other methods and investigators. It is found that while the major component of the absorption arises in the D region, there is a perceptible contribution from the F_2 region. This absorption varies with f_0F_2 almost linearly, according to Blum, while according to Mitra and Shain a curvilinear relation holds, over the f_0F_2 range 4 to 8 mc. During diffuse or spread F-layer echo conditions, more frequent in wintertime, there is a much less clear relationship with the f_0F_2 values. This type of absorption may be in fact scattering due to ionospheric inhomogeneities under disturbed conditions; it may be nondeviative absorption caused by electron-ion collisions in the F region as suggested by Gallet [126]; or it may be that it occurs in some region lying outside the ionosphere. This is at least one inference possible from Shain and Mitra's [116] observations of "excess" absorption occurring approximately 30 hours after Class-3 solar flares. At these temperate latitudes the total absorption of extraterrestrial signals received from the zenith at 18.3 mc is only of the order 1 to 2 db, of which approximately 1.0 db, at most, is apparently taking place in the D region at noontime. Values of the solar angle dependency exponent n are found to be 1.1 in summer, 0.9 during equinoctial periods, and 0.5 in winter at 18.3 mc. At 29.5 mc, the total zenithal absorption rarely exceeds 0.5 db, with a maximum of about 0.2 db assignable to D region effects.

High Latitude Effects: At high latitudes, the work of Little [98] on 65 mc and presently on 30 mc has been especially concerned with sudden and intense excess absorptions, amounting in some cases to 50 per cent or more of the signal. These are uniquely high-latitude effects, when unassociated with solar flares. The incidence of the 65 mc absorptions greater than 10 per cent correlates well with the occurrence of polar blackouts recorded on C-3 Ionospheric Sounding equipment. A noontime maximum incidence in both events is found at College, Alaska. The daily sum of the three-hour geomagnetic K indices for College also correlate, with a coefficient of 0.48, with a daily index of the incidence of absorption. Work currently in progress on 30 mc zenithal absorption at College appears to resolve two classes of absorption events: those with sudden commencement and those with rather gradual onset.

The sudden-commencement type appears closely associated with the occurrence of visible aurora in the zenith and has a diurnal maximum of incidence near local midnight. A typical value for the maximum absorption of this type is 40 per cent. The gradual-commencement type seems to occur with a broad maximum of incidence shortly after local noon, and is apparently more associated with the incidence of polar blackout conditions than with auroral conditions. A typical value for this class of absorption is 20 per cent.

That these anomalous absorption effects are taking place preponderantly below the level of the normal E and F ionospheric layers, and thus presumably in the D region of high collisional frequency, is indicated by the fading records obtained concurrently on sky-wave high frequency signals monitored at College. Relatively small absorption effects noted on zenithal equipment at, say, 30 mc are in general noted with great magnification on the high frequency circuits. This magnification would be expected both from the inverse squared-frequency dependency previously discussed and from the different obliquity factors of the paths involved. Thus, 1 db absorption at 30 mc would be expected to amount to about 100 db at a frequency one-tenth as high, neglecting any factor of obliquity, because the absorption exponent will be 100 times greater at the lower frequency. Zenith absorption at 30 mc amounting to only about 1 db is thus also found associated with no-echo conditions on the C-3 Ionospheric Sounder at College. It has not been quantitatively established, however, how much "extra-ionospheric" absorption of galactic signals may be taking place, but the available evidence indicates this is a very small effect. Additional observations employing extraterrestrial signals are known to have been conducted by Rydbeck (Sweden) on 30 mc, but results have not been reported. Further similar work is planned during International Geophysical Year at a site at Godhavn (west Greenland) by the Danish National Committee of URSI under direction of J. Rybner [137].

Lateral Extension of Regions of Excess Absorption: The spatial extent of regions of excess absorption have been explored by Rawer [129] and by Beynon and Davies [119] by examining the correlation between absorption effects as a function of the distance between observing stations. For stations 230 km apart a correlation coefficient of 0.74 was found. For a 330 km separation, the figure was 0.68 while for 400 km apart results likewise yielded the value 0.40. The summertime figures were in general lower than for winter, being a further manifestation of the special "winter-anomaly" (consisting of groups of days exhibiting absorption in excess of the usual day to day value) commented upon by Appleton and Piggott [114], by Dieminger [122], and by Lange-Hess [132].

Work on localized absorption effects in the high latitudes is being carried out at College, Alaska, using extraterrestrial signals at 30 mc and a rotating 3-element horizontally-polarized Yagi antenna. The antenna beamwidth is 55 degrees to the half-power points, and the wave angle is probably about 15 degrees above the horizontal. Analysis shows the sudden absorption effects are in general nonuniformly distributed over the sky, with the northern quadrants, *i.e.*, toward the auroral maximum zone, showing the most frequent and greatest effects. Sudden effects of particular azimuths amounting to as much as 95 per cent absorption have been observed, particularly during the hours of darkness. It is to be noted that radio wave emission arising from thermal excitation processes in the D region contributes a weak (about 2 per cent of mean) signal which tends to obscure observation of extreme cases of absorption.

These absorption equipments measure the average absorption over the polar diagram of the receiving antenna. On many occasions, absorption in excess of 3 db has been observed using a 3-element Yagi directed vertically overhead, implying that at least half of the 60-degree wide polar diagram was being affected at the same time. The observed values, however, give no information concerning the maximum intensity of absorption, and the possibility remains that there are localized areas of absorption of considerably greater intensity.

Summary

It is shown that at low and middle latitudes the absorption effect is preponderantly under solar control, through radiation ionization of the D region. Sudden, short-duration absorption events are directly correlated with solar flares and SID's at all latitudes. It is however, increasingly clear that at high latitudes intense and localized effects, frequently associated with auroral, magnetic, and ionospheric disturbances, become the most significant phenomenon. While it is the prevailing view that streams of charged particles, probably of solar origin, are the primary agency by which the ionization of the D region is enhanced, the details are not yet clear. In some cases it appears, for example by the work of Hagg and Hanson [138], that vertically downward mass-movement of clouds of ionization from the F to the E regions occurs. In other circumstances the observed apparent high-velocity of movement seems to preclude the possibility of mass transport. These cases are thought of as most probably being due to a sweeping motion of the ionizing agent, *i.e.*, the charged corpuscular stream. Effects like these, however, are probably coupled through electromagnetic forces and do not occur in separately resolvable manner. The particularly strong association of moving, *i.e.*, pulsating or flaming, auroral forms in producing absorption effects has been remarked in the north high-latitudes by Heppner [124] and by Major [139] in the far south.

M. Nicolet [140] has also suggested, through a theoretical investigation of the manner in which absorption depends upon variations in the vertical gradient of scale height (hence temperature and composition at D-region heights), that significant variations in absorption could be produced by changes in the temperature and atmospheric composition. That this may be a factor in seasonal anomalies such as the "wintertime" effect previously noted has been considered particularly by Dieminger [122].

In specific reference to nondeviative vhf absorption at high geographic latitudes, the results to date indicate: that the effects due to regular ionospheric layers are negligible compared with the polar blackout phenomenon, and that during conditions of blackout over a 3-month period in 1955, the absorption of zenithal extraterrestrial signals on 30 mc exceeded 1 db about 15 per cent of the total time, and exceeded 3 db less than 1 per cent of the time.

These observations, expressed as per cent signal absorption for one-way passage at vertical incidence are presented in Table II.

TABLE II
PER CENT ABSORPTION OF 30 MC EXTRATERRESTRIAL SIGNALS AT ZENITH

	0–10%	10–25%	25–50%	50–75%	75–100%	Month
% of serving time	94	5	0.6	—	—	July, 1955
	89	8.7	2.1	—	—	August, 1955
	66	26	6.8	0.7	0.1	September, 1955
	83	13	3.2	0.2	0.1	Average

These figures, when expressed as db power attenuation, can be extended to the frequency range 150–300 mc by the inverse frequency-squared relationship. Thus, the averaged data predicts absorption exceeding 0.04 db on 150 mc about 15 per cent of the time, and 0.12 db less than 1 per cent of the time. At 300 mc the expected figures would be 0.01 db exceeded about 15 per cent of the time, and 0.03 db exceeded about 1 per cent of the time.

These figures are subject to some upward revisions, however, to allow for the effects of oblique paths, seasonal and solar-cycle variations and for antenna averaging effects. For an absorbing layer extending from 70 km to 90 km, the obliquity factor amounts to a little over 6 times for ray paths making grazing incidence with the earth. For an angle of 75 degrees, a factor of about 4 is thus introduced by the obliquity. The seasonal variation in incidence of blackout conditions shows pronounced maxima in the equinoctial months, with incidence in the summer months (May through August) only about 1/5 to 1/10 as great. The effect is noticed in the table of values given above, and it has been reported more extensively, in the Geophysical Institute Quarterly Progress Report No. 4 (December 1, 1954, to February 28, 1955) entitled "Auroral Zone Absorption of Radio Waves Transmitted Via the Ionosphere" for Signal Corps Contract No. DA-36-039 SC 56739.

The solar-cycle (sunspot) effect on percentage of blackout time at high latitudes has not been established. Ionospheric soundings at College, suitable for this purpose, extend only from 1954 to date. These indicate a moderate increase in total blackout time in the early months of 1955 as compared with 1954, but extrapolation from this to a time of sunspot maximum is not reliably possible on this basis. An appreciable increase is to be anticipated, however, in view of the frequently observed association of sunspots and disturbed ionospheric conditions.

At vhf and at high latitudes it remains to be determined to what extent transient ionospheric abnormalities such as aurora, meteoric trails, and ionization produced by meteoric dust may be significant in producing deviative absorption.

VII. Refraction of Radio Waves by the Ionosphere

Radio waves from extraterrestrial sources are subject to refraction as they pass through the earth's atmosphere. This refraction occurs in two atmospheric regions, in the ionosphere and in the troposphere. This report is devoted to ionospheric effects, and therefore the tropospheric refraction is not further discussed.

This section is also restricted to the regular ionospheric refraction only; the irregular random changes in apparent position which are associated with the scintillations of the radio stars are discussed in section VIII.

Refraction at Low Angles of Elevation

The first investigation of regular ionospheric refraction was by Payne-Scott and McCready [141], who compared their 60 mc and 200 mc observations of the apparent elevation of a source of solar noise as the sun rose. The 60 mc radiation suffered much the greater deviation, presumably due to the much greater effect of the ionosphere at the lower frequency. Bailey [142] found that observed deviations, amounting to an increase of elevation of about 20 minutes of arc for a 60 mc source at an elevation of 5 degrees, exceeded the theoretical deviations by a factor of from 2 to 4. These observations were based on the assumption that the refraction at 200 mc was negligible; if this were not the case then the discrepancy between theory and experiment was even larger. The refraction decreases with increasing elevation, and it is probably less than 5 minutes of arc for a 200 mc source at an elevation of 10 degrees.

Using theoretical considerations, Bailey [142] has shown how the calculated and observed refractions can be brought into agreement. This he does by assuming that the F_2 layer above the region of maximum ionization is a parabola of a shape different from that assumed for the region of the F_2 layer below the maximum. The problem has also been investigated theoretically by Bremmer [143].

Refraction at Large Elevation Angles

For zenithal observations, Smith [144], using the idea of wedge refraction, develops for the refraction angle θ the equation

$$\theta = \frac{e^2}{2\epsilon_0 \pi m f^2} \cdot \frac{dn}{dx}$$

where m is the mass of the electron, f the wave frequency, n the number of electrons in a vertical column of unit cross section, and ϵ_0 is the permittivity of free space. Thus, measurement of θ enables a determination of the horizontal gradient of n, which in turn can give information about the upper ionosphere.

Independent calculation of the horizontal gradient has been carried out by Osborne [145] using Ratcliffe's approximate method [146] on ionospheric sounding data. His results agreed with those calculated from Smith's formula in the winter but not in the summer.

Further work by Ratcliffe [147] shows that it is likely that a greater proportion of electrons occurs above the region of maximum ionization in the summer than in the winter. If this assumption is made, then the refraction of the radio stars could be explained in winter in terms of a model ionosphere in which the total electron content (n) was two or three times the electron content below the maximum. This would make it similar to a theoretical Chapman layer. In summer the observations could be explained if it were assumed that the electron content increased to approximately twice the winter value. This increase is in agreement with the work of Ratcliffe.

Pawsey and Bracewell [148] describe two effects of the ionosphere: one, a prism-like bending of the rays toward the earth, which can permit the observation of sources below the horizon; and a lens-like bending which normally causes divergence and a reduction in flux density from a source. Reber [149] ascribes ionospheric anomalies to a turbulent scattering extension of the earth's upper atmosphere whose shape approximates to that of a thick disk some 30 degrees to 40 degrees wide.

Until more information is available no quantitative picture of the distribution of electron density above the F region maximum can be made, and numerical calculations of the expected ionospheric refraction would therefore be of doubtful value. However, once the variation of refraction with source elevation has been determined at a frequency well above the critical frequency, it should be possible to calculate the ionospheric refraction at higher frequencies, since the ionospheric refraction will be proportional to λ^2.

VIII. The Scintillation of the Radio Stars

Discovery of the Phenomenon

The scintillation (or "twinkling") of the radio stars was discovered by Hey, Parsons, and Phillips [150] in 1946. These authors were investigating the intensity distribution of extraterrestrial radio waves over the celestial sphere at a frequency of 64 mc, and found that on certain occasions the radio flux from the constellation of Cygnus was very variable with time. From this result, they deduced that there must be one or more localized sources of radio signals in that direction in space. Their suggestion led to the search for, and the discovery of, the localized sources of radio noise now known colloquially as "radio stars."

It was at first believed that the observed variations in signal strength from the localized sources were inherent in the sources themselves. Later work by Bolton, Slee, and Stanley [151] and by Little and Lovell [152] and by Smith [152] has shown that the variations of a single source, as viewed simultaneously from widely different points on the earth's surface, are not correlated. The phenomenon is therefore analogous to the twinkling of the optical stars in that the scintillations are produced in the earth's atmosphere; in the radio case, however, the disturbing region is the ionosphere, rather than the troposphere.

Diurnal Variation of Scintillation Activity

The diurnal variation of scintillation activity has been investigated at several centers. The published results differ widely for the different sites, apparently due to the differences in source elevation and latitude.

The simplest variation is that reported for zenithal observations in England. In this case, both the Manchester [153] and Cambridge [154] workers found that the scintillation activity was most marked shortly after midnight, and was rarely observed during the midday hours. For sources low on the northern horizon, Little [153] observed a similar diurnal variation in scintillation activity, except that it was considerably reduced in magnitude. At these low northerly elevations, scintillations were observable at all hours of the day.

The Australian observations [151] taken at low angles of elevation revealed a diurnal variation with two maxima, one at about midday and the other at about midnight. The discrepancy between the low elevation English and Australian observations is almost certainly due to the difference in latitude of the region of ionosphere traversed by the radio line-of-sight; in England the ionosphere was penetrated close to the auroral zone [156], in Australia between latitudes 20°S and 50°S [151].

Seasonal Variation of Scintillation Activity

The Australian observations [151], made over a period of about five years, revealed the presence of a seasonal variation of scintillation activity. The scintillations were approximately twice as strong during the summer and winter months as in the equinoctial months.

Observations by Hewish [155] taken at high angles of elevation in England have shown that seasonal variation in England is relatively small. Such variation as does exist is, however, similar to that observed by the Australians, since the scintillation index was at a minimum near the equinoxes.

Variation of Scintillation Amplitude with Source Elevation

The effect of source elevation upon scintillation amplitude has been discussed by several workers [153, 155–158]. In every case an increase in amplitude of the scintillations with decreasing source elevation was found. In England, Little [153] and Hewish [155] found that the scintillation index did not vary with source elevation for zenith angles less than 30 degrees, and had increased to about twice the value for a zenith angle of about 60 degrees. Little's observations showed that the factor had increased to about 6 at a zenith distance of 85 degrees.

This increase in scintillation amplitude with zenith distance has been interpreted in two different ways. Little and Maxwell [156] attribute it to the combined effects of the increased effective thickness of the disturbing region and a change in ionospheric conditions with latitude (the low elevation measurements were all taken at a northerly azimuth in England, *i.e.*, towards the auroral zone). On the other hand, Hewish [155] suggests that the increase is due to the increased distance of the disturbing region from the observer. It is now the opinion of the writer (C. G. Little) that the increase is due to a combination of all three effects. Assuming a uniformly disturbed ionosphere, the expected increase between zenith angles of 0 degrees and 85 degrees would be of the order of three times. The larger increases reported by Little and Maxwell [156], by Hartz [157], and by Seeger [158] for observations from northern temperate latitudes at low angles of elevation above the northern horizon are almost certainly due to observing the sources through the disturbed auroral F region.

For a given equipment sensitivity, the probability of observing scintillations increases with increasing zenith distance. Observations by Hartz [157] over a range in zenith angle of from 13.1 to 76.1 degrees indicate that the probability of detecting scintillations is proportional to the logarithm of the secant of the angle of incidence on the 400 km level of the ionosphere. By extrapolation he was able to show that scintillations should always be observed for zenith angles close to 90 degrees, at a frequency of 50 mc.

The Variation of Scintillation Activity with Observing Frequency

Several observers have analyzed records of radio star scintillations made at two or more frequencies simultaneously. For relatively high angles of elevation, Hewish [155] found that the fluctuation index varied approximately as the square of the wavelength and that the rate of fluctuations remained approximately constant with frequency. Observations at low angles of elevation in Australia [151] gave a very different picture, the scintillation index being roughly proportional to wavelength up to 100 mc and almost constant thereafter. The reason for this discrepancy is not clear; possibly the residual fluctuations at the higher frequencies in the Australian work were tropospheric in origin. On any of the present ionospheric theories of radio star scintillations, the scintillation index would be expected to decrease rapidly with increasing frequency; any tropospheric scintillation would, however, be expected to be much less frequency-sensitive.

The correlation between individual fluctuation records at different operating frequencies has been investigated by Burrows and Little [159], working simultaneously on 81.5 and 118.5 mc. The average cross-correlation coefficient between the two simultaneous scintillation records was about 0.6, with a range of from 0.4 to 0.9. No systematic variation of this cross correlation coefficient with fluctuation rate of source elevation was found; the correlation showed a tendency to decrease with increasing amplitude of fluctuations. These observations are in agreement with published comments by Hewish [155] and Bolton, Slee, and Stanley [151], though apparently in disagreement with some of the earlier Australian work [160].

Angular Deviations in the Apparent Position of Radio Stars

The advent of the interferometer technique [161, 162] and particularly of the phase-switch interferometer [163] has permitted measurements of the position of some of the stronger radio stars to be made with an accuracy of better than 1 minute of arc. As a result, it has been discovered that, associated with the scintillations in amplitude there are rapid, irregular changes in the apparent position of the source. These angular deviations have been investigated at Cambridge by Ryle and Hewish [154, 155] and by the Australian workers, Mills and Thomas [164].

Hewish [155] has published curves indicating that during strong scintillations the apparent position of the radio star varies about its mean position, with a standard deviation of about 5 minutes of arc at a wavelength of 8 m. These deviations have a period of the same order as the amplitude variations, but individual deviations and fluctuations are not closely correlated. The average magnitudes of the two phenomena are, however closely correlated, times of strong scintillation being associated with times when large changes in apparent position are observed [164].

In general these observations refer to the horizontal component of the irregular refraction since they were made with interferometers using horizontal baselines. The relative magnitude of the vertical component is not known, and may be significantly larger. It is clear, however, from the sea cliff interferometers that the vertical component cannot be enormously larger; otherwise the narrower lobe interferometers of this type would have been valueless.

The Dimensions of the Irregularities in the Diffraction Pattern

The early spaced receiver observations of radio star scintillations not only showed that the phenomenon was terrestrial in origin but also gave information about the dimensions of the irregularities in the intensity pattern across the ground. Thus Little and Lovell [152] found that there was complete correlation between antennas spaced 100 meters apart, and that the correlation had fallen to values between 0.5 and 0.95 for a separation of 3.9 km. Smith [152] found that the correlation was zero at distances greater than 20 km.

These results, leading to a figure of about 4 km for the size of the irregularities in the diffraction pattern, have since been confirmed by other workers [151, 153–156, 165–167]. A very important new result has recently been obtained in England by both the Manchester and the Cambridge groups, who have found [168, 169] that the irregularities in the pattern tend to lie in a series of parallel bands rather than approximately circular areas. Spencer [168] has shown that the axial ratio is often high (greater than 20), and that the observed orientation of the bands is in good agreement with a model in which the F region ionospheric irregularities creating the scintillations are aligned along the earth's magnetic field. This anisotropy in the diffraction pattern is a function of source elevation, and restricts the measurements of the motion of the diffraction pattern [155, 165–167] to the component of the velocity perpendicular to the bands.

One interesting fact is the relative constancy of the separation of the bands with time. The scintillation rate (number of fluctuations per unit time) has been observed to vary through approximately a factor of 100 to 1; simultaneous observations of the velocity of motion of the diffraction pattern indicate that this variation is due to changes in velocity rather than to changes in the dimensions of the irregularities [153, 166, 167]. Maxwell has shown that this average dimension, which appears to be an approximately fixed characteristic of the disturbing region, is roughly that of the minimum size of eddy which could exist in the upper ionosphere.

Scintillation Phenomena during Aurora and Magnetic Storms

The effect of aurora upon the scintillation of radio stars has been discussed by Little and Maxwell [170]. The most striking effect was upon the scintillation rate, which was found to be increased, on the average, by about a factor of four during aurora. The effect upon the amplitude of the scintillations was much less marked except during the active phases of the aurora (pulsing and flaming forms) which were found to be associated with strong scintillations.

On one occasion, a marked decrease in the intensity of the extraterrestrial signal, associated with the presence of an aurora, was observed at a frequency of 81.5 mc. The decrease (about 10 per cent of the total antenna power) was tentatively attributed to the reflection of some of the incoming radiation back into space; later observations at College [98] suggest that ionospheric absorption is the more likely explanation for the decrease in signal strength.

The effect of magnetic storms upon the scintillation of the radio stars has been considered by Smith [152], who has shown that the periods of strong scintillation activity were not closely related to magnetic storms. An analysis by Little and Maxwell [170] has shown that although the amplitude is little affected by the magnetic activity, the fluctuation rate is notably affected, and that the number of fluctuations per minute is roughly proportional to the magnetic K index.

Fluctuation Rates

A notable feature of the scintillation of the radio stars is the variability of the rate of fluctuation. The average value of this fluctuation rate for a source in the zenith is of the order of 1 or 2 fluctuations per minute, but it may vary from about 10 fluctuations per minute to about 0.2 fluctuation per minute. Little [153] has shown that, in England, the fluctuation rate increases about threefold as the source moves from transit (near the zenith) to lower culmination (at a zenith distance of 86 degrees).

This fluctuation rate is the combination of various factors, including (in decreasing order of importance) the motion of the ionospheric irregularities, the motion of the radio line of sight across the ionosphere as the earth rotates, and the rate of change of the ionospheric structure. The angular motion of the radio stars is a function of their declination; hence, the rate at which the line of sight sweeps across the ionospheric structure is a function of the star's declination as well as its elevation above the horizon.

It has been found that the fluctuation rates at a given time are the same over a wide band of frequencies [151], and, as indicated in the previous section, that the fluctuation rate is roughly proportional to the magnetic K index [170].

The Correlation of Radio Star Scintillations with Ionospheric Phenomena

A number of investigations have been made to determine the location of the disturbing region by correlating the occurrence of radio star scintillation with the appearance of various abnormal features in the structure of the ionosphere.

In England, both the Cambridge [154] and the Manchester [156] workers reported that their zenithal observations showed good correlation between the occurrence of radio star scintillations and the phenomenon known as "spread F." (Normally the virtual height and the critical frequency of the nighttime F region are well defined; on about one night in three in England these characteristics are found to be spread over as much as several hundred km and one or two mc per second respectively.) A similar correlation has been observed by Mills and Thomas [164] in Australia and by Hartz [157] in Canada. No correlation was found with sporadic E or meteor showers.

The agreement between these observations would appear to be convincing were it not for the fact that Bolton, Slee, and Stanley [151] in Australia have reported a lack of correlation with spread F. A similar result has been reported by Dueno at Cornell University. Both these groups of workers report significant correlation between the occurrence of radio star scintillations and sporadic E.

The reason for this discrepancy is not certain, though in the opinion of the writer (C. G. Little) it is probably due to a difference in the elevation angle at which these investigations were carried out. The English work refers to zenithal observations, and no investigation was made to determine the origin of the low-elevations scintillations. On the other hand, the Australian observations (other than those by Mills and Thomas [164] who found good correlation with spread F) were taken at elevation angles less than 10 degrees. It is believed that the sporadic E correlation observed by Dueno was also based upon low-elevation observations. The increasing predominance of E-region phenomena at low angles of elevation can readily be explained as due to the curvature of the earth's surface, which causes the effective thickness of the lower layers to increase by a larger ratio than for higher layers. For example, radiation leaving the earth's surface at zero elevation is incident upon the E layer at an angle of about 10 degrees, and upon a 400 km height layer at an angle of about 20 degrees; for this reason the effective thickness of the E region increases about sixfold from zenith to horizon for the E region, and only about threefold for the F region.

Theory

The following is a simplified treatment of the theory of the scintillation of radio stars, developed in outline to indicate the manner in which the scintillations are produced and the way in which they may be expected to vary with frequency.

The problem may be considered as analogous to that of the diffraction of electromagnetic radiation by an irregular phase-shifting screen. Such a problem has been considered in various papers [155, 165, 172–175]; the treatment below is, however, aimed to give a simple physical picture of what happens, rather than a rigorous mathematical treatment.

Consider the case of a plane wave incident normally upon a horizontal, irregular diffracting screen which introduces irregularities in phase only. The emergent wavefront (surface of constant phase) will therefore no longer be plane but will be distorted in an irregular manner. The amplitude distribution across a plane immediately below the diffracting screen will be uniform, since the screen has been supposed to affect only the phase of the radiation. At some horizontal plane a considerable distance below the diffracting screen, however, the effect of this "corrugated" emergent wavefront is to produce a nonuniform amplitude distribution. The minimum distance at which the amplitude irregularities are fully developed is determined by the size and intensity of the distortions in the emergent wavefront. In this connection, it is convenient to measure the average dimensions of the irregularity in the emergent wavefront in terms of the phase autocorrelation function, across a plane parallel to, and immediately below the diffracting screen. When this is done, it is found that the width of this autocorrelation function is of the same order as the corrugations in the emergent wavefront, only if the average distortions in the wavefront do not exceed about one radian. For an irregular wavefront in which the typical distortion is N radians (N greater than unity), the autocorrelation function is very approximately $1/N$ times as wide as the individual irregularities.

Booker, Ratcliffe, and Shinn [172] have shown in the one dimensional case, and Fejer [174] in the two dimensional case, that this autocorrelation function is the same across all planes below the emergent plane; hence, the average lateral dimensions of the irregularities in the diffraction pattern across some distant receiving plane are about equal to those of the corrugation in the emergent wavefront, if these have depth up to one radian, and decrease steadily in size as these corrugations increase in depth. If the magnitude of the phase changes which cause these corrugations is much less than about one radian, the amplitude variations across the receiving plane can never be large, however distant this plane may be. It can be shown that the amplitude variations will not be fully developed until the receiving plane is sufficiently distant from the emerging plane for the radius of the first Fresnel zone $(R\lambda)^{1/2}$ to be of the same order as the separation along the emerging plane for which the autocorrelation function has effectively fallen to its minimum value.

Let us now apply this to the case of radio star scintillations at vhf and uhf. In this case we have a plane wave (from the radio star) passing through a phase distorting screen (the ionosphere), to be received at a receiving plane (the earth's surface). The phase distortions are produced by irregularities in the ionosphere, whose refractive index μ is given, for a frequency f well above the critical frequency f_c, by

$$\mu = \left[1 - \left(\frac{f_c}{f}\right)^2\right]^{1/2} \text{ or } \quad \Delta\mu = 1 - \mu \cong \frac{1}{2}\left(\frac{f_c}{f}\right)^2 .$$

A given irregularity will therefore distort the wavefront by a distance proportional to λ^2.

It can be shown that the width of the angular cone of plane waves required to build up the emergent wavefront will be proportional to λ^2; the irregular angular deviations observed at the receiving plane will therefore also be proportional to λ^2. Note that the phase change (as opposed to the path change) introduced by a given irregularity is proportional to λ.

Consider what happens as we vary the frequency of the incident plane wave from the radio star, starting first at the top of the uhf band (3000 mc). The total path change introduced by the presence of the ionosphere will be given by the integral $\int\Delta\mu \cdot ds$ along the path through the ionosphere. A rough estimate for maximum conditions (summer midday at low latitudes) can be obtained by suggesting that the effective thickness of the ionosphere at normal incidence would be equivalent to that of a uniform layer 200 km thick with a critical frequency of 10 mc. This estimate is unlikely to be in error by more than a factor of about two. For such a layer, the total path change introduced by the ionosphere would be of the order 10 wavelengths at 3000 mc.

The radius of the first Fresnel zone, assuming a height of 400 km for the disturbing region, would be about 200 m.

The above theoretical considerations therefore indicate that strong scintillations will not be observed unless the effective thickness of the ionosphere varies by more than 1 per cent over a distance of only 200 m. Such rapid variations cannot be expected in an ionosphere which is some hundreds of kilometers thick.

The theoretical work also shows that the irregular gradients in ionosphere thickness required to produce scintillations when expressed as a per cent change of thickness per km normal to the line of sight, are proportional to $f^{3/2}$. (A factor $f^{1/2}$ arises due to the radius of the Fresnel zone, and a further factor of $f^{1.0}$ due to the magnitude of the phase change introduced by a single irregularity.) Thus, at 30 mc the total path change introduced by the ionosphere will be of the order of 1000 wavelengths (at equatorial noon), and the radius of the first Fresnel zone will be about 2 km. The above theoretical work therefore suggests that strong scintillations will be seen unless the irregular variations in equivalent thickness of the ionosphere are less than 10^{-4} over a distance of 2 km. The fact that such scintillations are not observed for more than about 10 per cent of the observing time is an indication of the uniformity of the thickness of the ionosphere over distances of the order of a few km.

Consider now the effect of observations at lower angles of elevation. This change has the double effect of increasing the effective thickness of the ionosphere and also increasing the distance of the observer from the diffracting screen. For elevations of only a few degrees, the distance to the 400 km height will be increased approximately 5 times, and the effective thickness will be increased about 3 times. The combined effect is to reduce the per cent change of effective thickness per km required to produce scintillations by a factor of about $3\sqrt{5}$, say 7 times. It is therefore not surprising that many observers [153, 155–158] have reported an increase of scintillation activity at low angles, although it must be remembered that this may also, in part, be due to a change in the nature of the diffracting screen with latitude [156].

The actual distribution of blob sizes in the diffracting screen is not known. It is probable that the emerging wavefront includes distortions of many sizes. However, the earth will be too near the diffracting screen for any corrugations whose phase autocorrelation distance is large compared with the Fresnel zone to be effective in producing amplitude scintillations. Also, any distortions, even if sufficiently small to meet this requirement, will not produce strong scintillations unless they produce phase changes of the order of one radian. This is almost certainly the reason why the relatively small blobs of ionization which give rise to auroral echoes apparently do not cause radio star scintillations.

When considering the question of communicating with an object not at infinity (*e.g.*, the moon, or the International Geophysical Year satellite), this question of effective blob size will be an important one. In this case, one has to consider a spherical wavefront, and the radius of the first Fresnel zone (which determines the maximum distance over which irregular phase changes of the order of one radian must be produced to cause strong scintillation) will depend on both the transmitter to diffracting screen distance and the diffracting screen to receiver distance. When the object is at the diffracting screen, the first Fresnel zone has dimensions only of the order of one wavelength, and no scintillations will be observed. As the distance of the object from the diffracting screen is increased, the radius of the first Fresnel zone increases, reaching

$$\frac{\sqrt{R\lambda}}{\sqrt{2}}$$

when the transmitter and object distances from the screen are both equal to R, and $\sqrt{R\lambda}$ when the object distance reaches infinity. Since, for small values of phase change per Fresnel radius, the fluctuation index (rms value of scintillations expressed as a percentage of the mean power from the source) is proportional to the square of the phase change per Fresnel radius, this suggests that that fluctuation index observed by an object an equal distance from the diffracting screen as the transmitter would be 0.5 that which the object would see at infinity, and therefore (by inverting the positions of the transmitter and receiver) 0.5 of that observed from a radio star.

Estimates of the Magnitudes of the Effects to be Observed

In summary, as the frequency is decreased from 3000 mc, weak scintillations will begin to appear at a frequency which will be dependent upon the elevation angle of the source and (since the ionospheric conditions vary with latitude) also upon latitude. The intensity of the source at this frequency may be expected to have a Gaussian distribution about its mean level. Present experience suggests that the standard deviation will be not more than 10 per cent at 300 mc for a source in the zenith, but may occasionally reach values three times this at low angles of elevation.

As the frequency is further decreased, the fluctuation index may be expected to increase until the amplitude distribution tends to a Rayleigh distribution. Through this range of frequency, the scale (lateral dimensions) of the diffraction pattern, and hence the fluctuation rate, will be constant with frequency. For zenithal observations, this Rayleigh distribution will be rarely reached at 100 mc (say, a few hours per year). At low angles of elevation (but greater than 3 degrees, to avoid uncertainty as to effects produced by the troposphere) a Rayleigh distribution may be expected for a similar proportion of time at a frequency of 200 mc. As the frequency is still further reduced below that at which a Rayleigh distribution is observed, the scale of the pattern will begin to decrease proportional to $1/\lambda$ and the fluctuation rate to increase proportional to λ. For all frequencies, irregular angular deviations should be proportional to λ^2.

The frequency at which a Rayleigh distribution is observed most of the time for a source in the zenith is not known, but it is believed to be less than 30 mc, even at Arctic latitudes. At low angles of elevation, the author believes a Rayleigh distribution would be observed most of the time at 30 mc in the Arctic, and less than half the time at 60 mc.

The above estimates are what might be termed "educated guesses" and must not be regarded as being the results of scientific experiments. They are included only to give an indication of the magnitude of the effects that may be expected and include no safety factor which might enable them to be used for design purposes.

List of Agencies Active in VHF—UHF Ionospheric Studies

Auroral Radar

- Geophysical Institute, University of Alaska, College, Alaska.
- School of Electrical Engineering, Cornell University, Ithaca, N. Y.
- Radio Propagation Laboratory, Stanford University, Stanford, Calif.
- Physics Department, University of Saskatchewan, Saskatoon, Saskatchewan, Can.
- Jodrell Bank Experimental Station, University of Manchester, Manchester 13, Eng.
- Norwegian Defence Research Establishment, Division of Telecommunication, Kjeller, Norway.
- Research Laboratory of Electronics, Chalmers University of Technology, Gothenburg, Sweden.

In addition to the above agencies, the National Committees for the International Geophysical Year 1957–1958 of Australia, Canada, France, Great Britain, New Zealand, Sweden, United States, and USSR have indicated their intention to use auroral echoes as a technique for studying aurora during the International Geophysical Year.

Meteor Radar

- Radio Propagation Laboratory, Stanford University, Stanford, Calif.
- Harvard Observatory, Cambridge, Mass.
- Geophysical Institute, University of Alaska, College, Alaska.
- Jodrell Bank Experimental Station, University of Manchester, Manchester 13, Eng.
- Department of Physics, University of Adelaide, Adelaide, Australia.
- Radio Physics Laboratory, Defence Research Board, Ottawa, Can.
- Canterbury University College, Christchurch, New Zealand.

Moon Radar

- Signal Engineering Laboratories, Fort Monmouth, N. J.
- Jodrell Bank Experimental Station, University of Manchester, Manchester 13, Eng.

Radio Noise of Auroral Origin

- Geophysical Institute, University of Alaska, College, Alaska.

Ionospheric Absorption

- Geophysical Institute, University of Alaska, College, Alaska.
- Central Radio Propagation Laboratory, National Bureau of Standards, Boulder, Colo.
- Radio Research Station, D.S.I.R., Ditton Park, Slough, Bucks, Eng.
- Division of Radiophysics, Commonwealth Scientific and Industrial Research Organization, Australia.

Ionospheric Refraction

- Geophysical Institute, University of Alaska, College, Alaska.
- National Bureau of Standards, Boulder, Colo.
- School of Electrical Engineering, Cornell University, Ithaca, N. Y.
- General Electric Co., Syracuse, N. Y.
- Massachusetts Institute of Technology, Lincoln Laboratory, Lexington, Mass.
- Radio Physics Group, Cavendish Laboratories, Cambridge, Eng.

Radio Star Scintillation

- Geophysical Institute, University of Alaska, College, Alaska.
- School of Electrical Engineering, Cornell University, Ithaca, N. Y.
- General Electric Company, Syracuse, N. Y.
- Massachusetts Institute of Technology, Lincoln Laboratory, Lexington, Mass.
- Department of Terrestrial Magnetism, Carnegie Institution of Washington, 5241 Broad Branch Road, N.W., Washington, D. C.
- Radio Physics Laboratory, Defence Research Board, Ottawa, Can.
- Jodrell Bank Experimental Station, University of Manchester, Manchester 13, Eng.
- Radio Physics Group, Cavendish Laboratories, Cambridge, Eng.
- Commonwealth Scientific and Industrial Research Organization, Australia.
- Research Laboratory of Electronics, Chalmers University of Technology, Gothenburg, Sweden.

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Directional Channel-Separation Filters*

S. B. COHN†, SENIOR MEMBER, IRE, AND F. S. COALE†

***Summary*—A class of frequency-selective networks is described that combines the properties of a directional coupler and a conventional filter. These networks, which will be called "directional filters," are particularly suited for combining or separating signals of different frequencies in communication systems, for multiplexing several equipments connected to a single antenna, etc. A signal entering one arm of the network will be transferred with essentially zero loss to a second arm over a desired channel of frequencies, while at other frequencies the signal will emerge with little loss from a third arm. Over the entire frequency range the fourth arm is isolated and the first arm is nonreflecting.**

Many new types of directional filters have been found and are described in this paper. Several of these are particularly easy to design and build in strip or coaxial line, while others lead to compact, economical structures in waveguide. Several lumped-constant circuits are also given. Design information and experimental data are included in this paper.

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INTRODUCTION

ONE of the most common uses of microwave filters is in multiplexing (separating or combining) signals of different frequencies. For example, in microwave communication systems a large number of frequency channels are usually combined by filter networks and radiated in a wide bandwidth from a single antenna. At the receiving point of the system, this total spectrum enters the receiving antenna and is then divided by filters into its constituent channels. Other applications for such filters are in multiple-frequency radar systems and in equipment installations where a single antenna must be shared by several electronic devices. One type of multiplexing filter[1] that has had considerable use in microwave communications consists of two hybrid junctions and two band-rejection

[1] W. D. Lewis and L. C. Tillotson, "A non-reflecting branching filter for microwaves," *Bell Sys. Tech. J.*, vol. 27, pp. 83–96; January, 1948.

filters suitably connected in a waveguide circuit. Another makes use of an arrangement of coaxial resonant cavities coupled together in an intricate manner.[2] These two types of filters are characterized by the following basic properties, which are illustrated in Fig. 1: 1) they each have four arms (*i.e.*, four terminal pairs), one of which is always isolated from the input arm; 2) the input signal power emerges from one arm with the frequency response of a band-pass filter, while the remaining power emerges from another arm with the complementary response of a band-rejection filter; 3) the input arm is nonreflecting when the other arms are connected to their characteristic impedances, and hence the filter is a constant-resistance-type network; 4) as in the case of a directional coupler, these properties apply no matter which of the four arms is used as the input arm. As suggested by these properties, the name "directional filter" will be applied to the two filters described above, and to all other filters of this class.

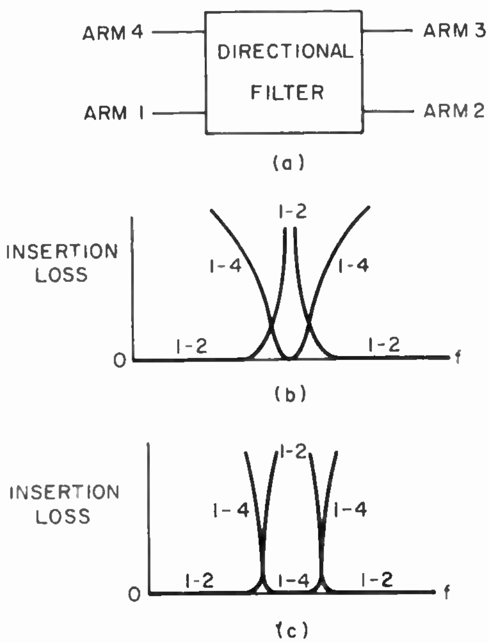

Fig. 1—Insertion-loss response defining performance of directional filter.

Fig. 2 gives an example of the use of directional filters for separating signals of different frequencies. In the case of filters *a*, *b*, and *d* the isolated arm is not used. However, in the case of filter *c* it is used to permit the frequencies f_c and f_d to appear at a common arm. Because the directional filter is a reciprocal device, Fig. 2 may also be used as an example of a frequency-combination network, if all the arrows are reversed. Countless other interconnections of directional filters are feasible to meet specific requirements.

[2] H. J. Carlin, "UHF multiplexer uses selective couplers," *Electronics*, vol. 28, pp. 152–155; November, 1955.

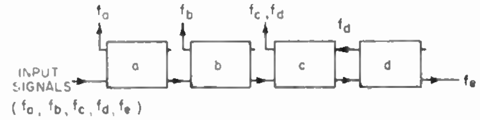

Fig. 2—Example of directional-filter application to channel separation.

Novel Directional-Filter Circuits

Many new directional-filter circuits are described for the first time in this paper. These include circuits composed of strip line or other TEM transmission line, of waveguide, and of lumped constants. The transmission-line and waveguide filters are much more economical to construct than previous directional filters, and are generally more compact. The principles revealed by the examples in this paper can be easily extended to obtain many other circuits having directional-filter properties.

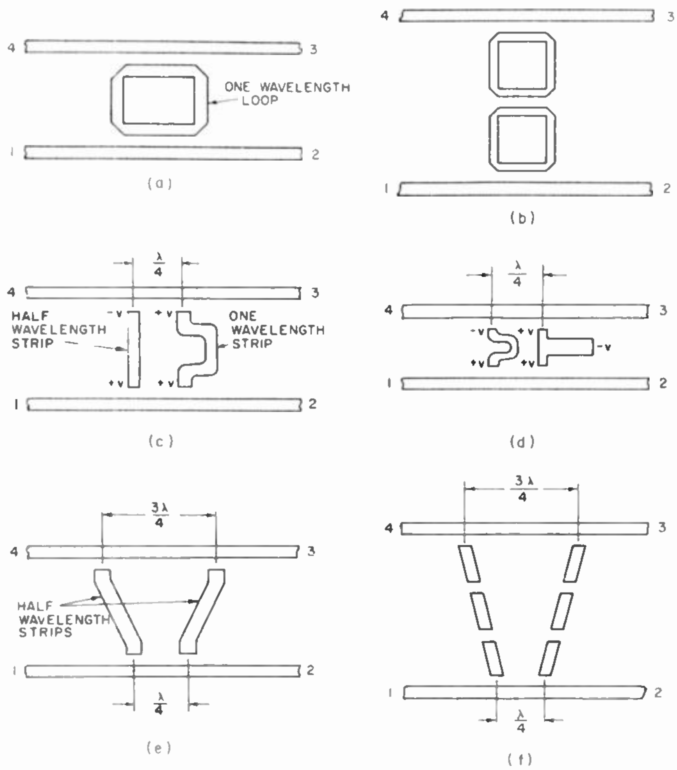

Fig. 3—Strip-transmission-line configurations having directional-filter performance.

A number of strip-line directional filters are shown in Fig. 3. The one in Fig. 3(a) consists of two parallel-strip directional couplers with one strip of each connected in a loop.[3] The parallel-strip directional coupler couples in a reverse direction, and hence a wave entering arm 1 excites a wave traveling clockwise in the loop.

[3] F. S. Coale, "A traveling-wave-ring filter," presented at Natl. Symposium on Microwave Techniques, Philadelphia, Pa., February 2–3, 1956. To be published in Trans. IRE.

At the frequency at which the loop is one wavelength around, the amplitude of this traveling wave builds up to a large value in a manner similar to resonance of a cavity. The second directional coupler removes energy from the loop through arm 4 of the network. It can be shown that all of the input power emerges from arm 4 at this frequency (assuming no dissipation loss in the circuit), while at frequencies sufficiently removed virtually all the power emerges from arm 2. If this *traveling-wave-loop* directional filter is properly adjusted, no power emerges from arm 3 at any frequency, nor is any reflected from arm 1.

The measured frequency response of an experimental traveling-wave-loop filter is shown in Fig. 4 and demonstrates very well the directional filter properties. The arm 4 insertion loss of 1.2 db at midband is due mainly to dissipation in the loop. This can be decreased by reducing the attenuation constant of the loop transmission line, or by increasing the bandwidth. The rejection of arm 2 is very sharp, and could be made deeper, if desired, by more careful adjustment. Similarly, the isolation of arm 3 and the vswr of arm 1 are very good and could also be improved by adjustment. The degree of selectivity observed is what would be expected of a single-resonant-circuit band-pass or band-rejection filter. Greater selectivity may be achieved by the use of two or more loops coupled to each other in cascade, as in Fig. 3(b). In this way the well-defined pass and stop bands illustrated in Fig. 1(c) can be obtained.

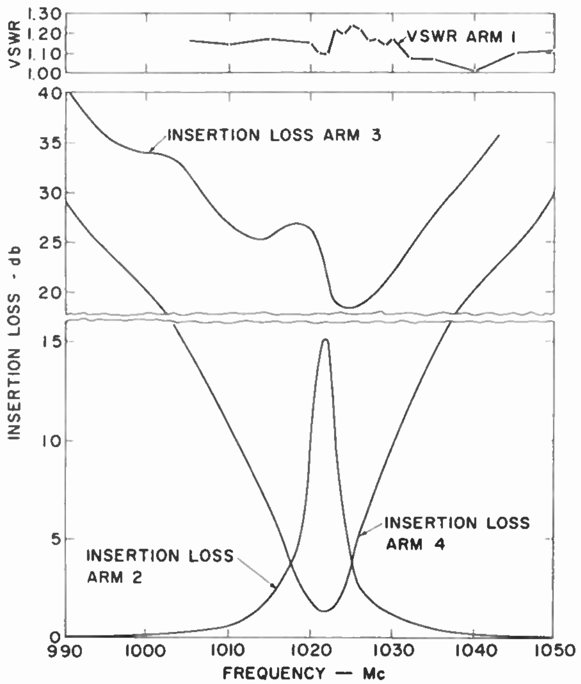

Fig. 4—Measured insertion loss and vswr of traveling-wave-loop filter.

Other strip-line circuits that provide directional-filter performance are shown in Figs. 3(c), (d), and (e). Their principle of operation cannot be stated as simply as in the case of the traveling-wave-loop filter, but a few points significant to their operation are as follows: Each of the circuits of Figs. 3(c), (d), and (e) contains two separate stripline resonators capacitively coupled to a pair of strip lines at points separated by an odd number of quarter wavelengths. In Figs. 3(c) and (d) the resonant strips are made dissimilar, so that the coupling points on one are at opposite polarities and on the other at the same polarity. In Fig. 3(e) the resonant strips are identical, but an effective polarity relationship like that of Figs. 3(c) and (d) is achieved through the use of a separation of the coupling points one-half wavelength greater on one line than on the other. A study of the phase relationships shows that a wave entering arm 1 will couple to arm 4. An analysis proving that all properties of a directional filter are obtained with these circuits is given in the section on Design Equations.

The insertion-loss response measured for an experimental model of the circuit in Fig. 3(e) is shown in Fig. 5. The excellent results verify the performance of the circuit as a directional filter. As in the case of the traveling-wave loop, greater selectivity may be achieved for this filter, or for any other type of directional filter, by utilizing resonators in cascade. This construction is illustrated by the example in Fig. 3(f).

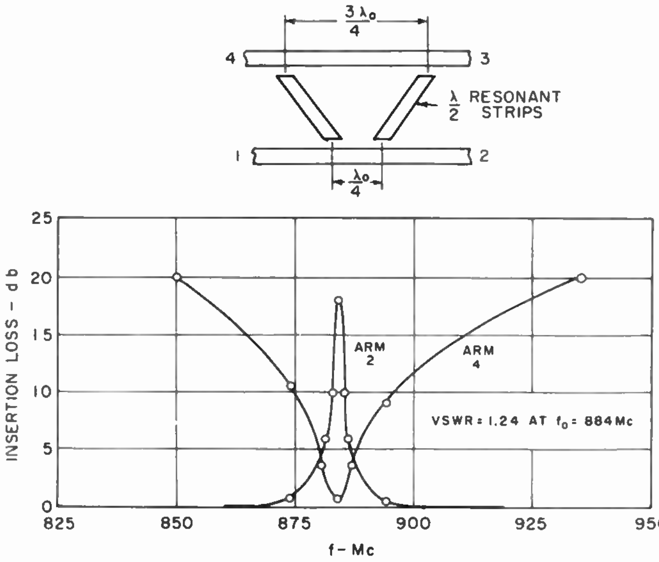

Fig. 5—Response of strip-line half-wavelength-resonator directional filter.

Fig. 6 shows a number of waveguide directional-filter types. The simplest configuration is that of Fig. 6(a), where a single cylindrical cavity is coupled by apertures to two rectangular waveguides.[4] Each coupling aperture

[4] After this paper was presented at the 1956 IRE National Convention, independent work being carried out on the same waveguide configuration was called to the authors' attention. This parallel work is being done by Bell Laboratories personnel, and by C. E. Nelson of Hughes Research Laboratories. Mr. Nelson described his work at the Symposium on Microwave Properties and Applications of Ferrites, Harvard University, April 4, 1956.

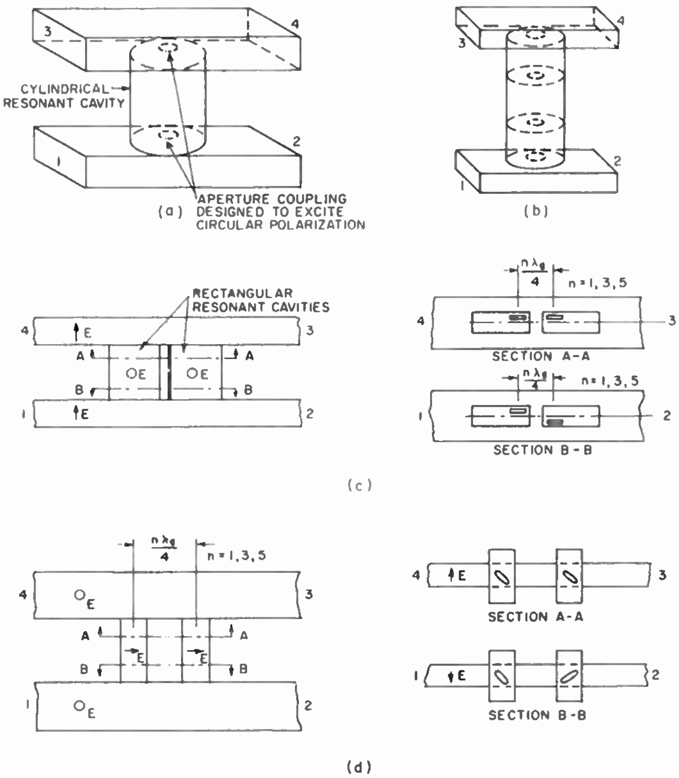

Fig. 6—Waveguide configurations having directional filter characteristics.

is located off-center on the rectangular-waveguide broad wall in such a way that a wave entering arm 1 of the rectangular waveguide will excite a circularly polarized resonant wave in the cylindrical cavity. This will then couple directionally to arm 4 of the second rectangular waveguide. Because of the use of circular polarization, the two rectangular waveguides may be at any angle to each other. In cases where greater selectivity is desired, several cavities may be coupled in cascade, as in Fig. 6(b). Experimental insertion-loss and vswr

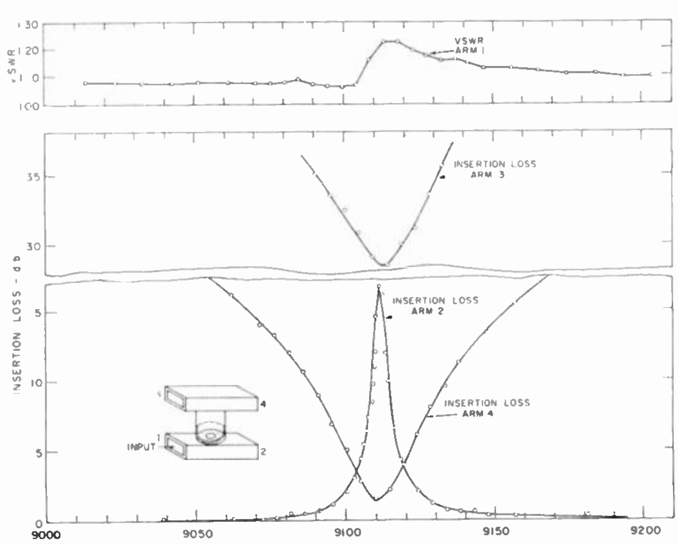

Fig. 7—Measured insertion loss and response of waveguide directional filter.

curves are shown in Fig. 7 for a model of the filter of Fig. 6(a), and again demonstrate directional-filter performance. Design formulas for this filter are given in the next section.

In a very recent paper, another waveguide configuration was presented that utilizes circular polarization and that has directional filter characteristics.[5] However, the method of connection of the rectangular-waveguide arms and the method of exciting circular polarization is completely different from that of Fig. 6(a).

Two other waveguide directional-filter configurations are shown in Figs. 6(c) and (d). They each have two resonant cavities, and are analogous to the transmission-line filters of Figs. 3(c), (d), and (e). The polarity-of-coupling relationship necessary to operation as a directional filter is achieved by proper location and orientation of the coupling slots.

A few lumped-constant versions of directional filters are given in Fig. 8 below. Those of Figs. 8(a) and (b)

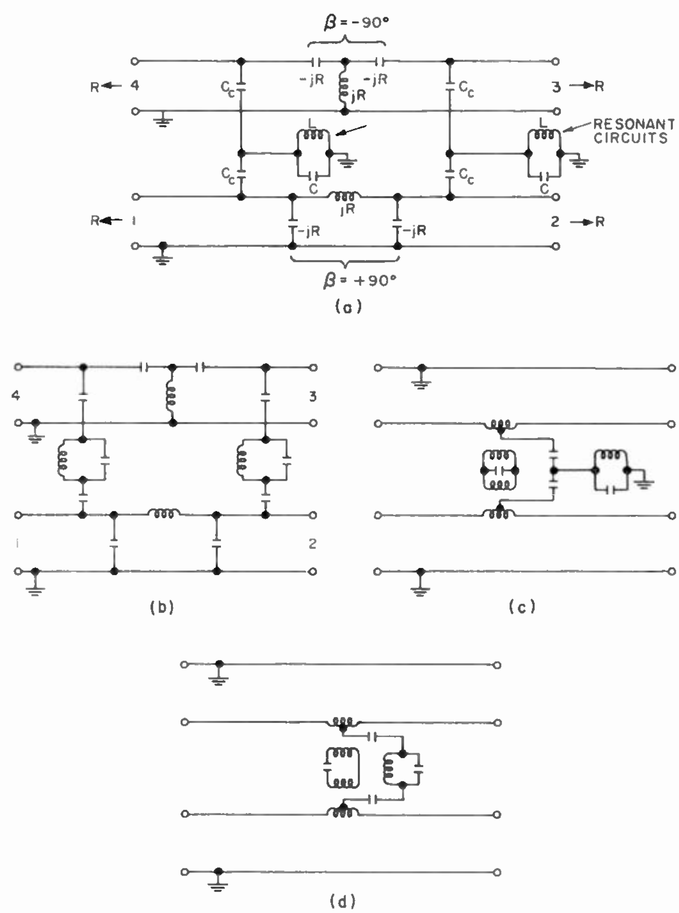

Fig. 8—Lumped-constant directional filter circuits.

are analogous to some of the transmission-line and waveguide filters in that a pair of resonant circuits are separated by lumped-constant equivalents of quarter-wave-

[5] R. W. Klopfenstein and J. Epstein, "The Polarguide—A constant resistance waveguide filter," Proc. IRE, vol. 44, pp. 210–218; February, 1956.

length (90 degree) lines. The necessary polarity relationship is here achieved by designing one phase-shift network to have +90 degrees and the other −90 degrees phase shift. The properties of the circuit of Fig. 8(a) were verified experimentally.

Two other types of lumped-constant directional filters are shown in Figs. 8(c) and (d). In these the directional-filter properties are achieved through the use of two resonant circuits coupled by different means to two lines at the same points. Thus the frequency dependence of the ±90 degree phase-shift networks is avoided, and wide-range tunable filters of these types are possible.

Most of the directional-filter types described above utilize a pair of resonators or resonant circuits. In the types using a single resonant structure, analysis has shown that two orthogonal resonant modes are excited in the structure, and that both are necessary to achieve the performance of a directional filter.

Design Equations

Traveling-Wave-Loop Filter

Design formulas for the traveling-wave-loop filter have been derived by summing the step-by-step build-up components of the wave in the loop when an input wave of amplitude $E_1=1$ is applied at arm 1.[3] The steady-state output-wave amplitude from arm 4 obtained by this method is

$$E_4 = \frac{c_1 c_2 e^{-(1/2)(\alpha+j\beta)l}}{1-(1-c_1^2)^{1/2}(1-c_2^2)^{1/2}e^{-(\alpha+j\beta)l}} \tag{1}$$

and from arm 2 is

$$E_2 = \frac{(1-c_1^2)^{1/2}-(1-c_2^2)^{1/2}e^{-(\alpha+j\beta)l}}{1-(1-c_1^2)^{1/2}(1-c_2^2)^{1/2}e^{-(\alpha+j\beta)l}}, \tag{2}$$

where, with reference to Fig. 3(a), c_1^2 is the power coupling factor of the lower directional coupler and c_2^2 is that of the upper, α is the attenuation in nepers per unit length of the loop transmission line, β is the phase constant of the loop in radians per unit length, and l is the line length around the loop. At the center frequency of the filter $\beta l=2\pi n$, where n is the number of wavelengths around the loop, and hence it is seen that the condition for complete rejection between arms 1 and 2 is

$$\frac{1-c_2^2}{1-c_1^2} = e^{2\alpha l}. \tag{3}$$

Therefore, if the loop has attenuation, the couplings must be made unequal to obtain a perfect null. If the couplings are made equal, the rejection will not be perfect but will generally be substantial, as in Fig. 4.

With $c_1=c_2=c$, the loaded Q of the single-loop filter of Fig. 3(a) is

$$Q_L = \frac{n\pi(1-c^2)^{1/2}e^{-\alpha l/2}}{1-(1-c^2)e^{-\alpha l}}. \tag{4}$$

It should be noted that this expression gives the true loaded Q of the resonator as it would be measured, and takes account of the external loading of the couplers combined with the losses in the transmission line of the loop. That is, $Q_L=f_0/(f_2-f_1)$ where f_2 and f_1 are the 3 db points with respect to the response at the center frequency f_0.

The insertion loss at resonance is given by

$$\text{Insertion Loss} = 20 \log\left[\frac{1-(1-c^2)e^{-\alpha l}}{c^2 e^{-\alpha l/2}}\right]. \tag{5}$$

If a waveguide loop is used in the filter, Q_L as given in (4) must be multiplied by $(\lambda_g/\lambda)^2$ where λ_g is the guide wavelength and λ is the free-space wavelength. Eq. (5) for insertion loss applies without change to the waveguide case.

If several loops are connected in cascade, as in Fig. 3(b), the coupling coefficient between adjacent traveling-wave-loop resonators is given by

$$k = \frac{1}{\pi n}\tan^{-1}\left(\frac{c}{\sqrt{1-c^2}}\right) \approx \frac{c}{\pi n} \tag{6}$$

where c is the amplitude coupling factor of the directional coupler between the loops and n is the number of wavelengths of each loop.

Half-Wavelength-Strip Filter

The filter circuits of Figs. 3(c) to (f) are analyzed most easily by calculating the reflection and transmission of the circuit for a pair of equal waves entering arms 1 and 4 first in phase and then 180 degrees out of phase. For example, in Fig. 3(e), waves arriving in phase at the ends of the left-hand strip resonator will be transmitted past it with virtually no reflection. However, because of the path-length difference, the pair of waves will arrive at the right-hand strip 180 degrees out of phase, and will therefore excite this strip, causing reflected waves to emerge from arms 1 and 4. Similarly, a pair of waves entering arms 1 and 4 out of phase will excite the left-hand strip, while the transmitted waves will be in phase at the right-hand strip, and hence will not be affected by it. When these two cases are superimposed (with circuit dissipation neglected), it is found that the reflected waves out of arm 1 cancel each other, and out of arm 4 add. The transmitted waves out of arm 2 add, and out of arm 3 cancel. Thus arm 1 is nonreflecting and arm 3 is isolated, as is required for directional-filter performance.

It is seen from the above argument that the relative amplitude of the wave emerging from arm 4 is equal to the reflection coefficient caused by one strip resonator with its midpoint at ground potential. Thus the output from arm 4, when a wave of unit amplitude is incident at arm 1, is

$$-\rho = \frac{Y_0 - Y}{Y_0 + Y} = \frac{-jB}{2Y_0+jB} = \frac{-1}{1+j2X/Z_0}$$

where $Z_0 = 1/Y_0$ is the characteristic impedance of the transmission line, $X = -1/B$ is the shunt reactance introduced by a strip resonator with its midpoint grounded, and $Y = Y_0 + jB$. The shunt reactance is

$$X = \frac{-1}{\omega C} + Z_0 \tan \phi$$

where C is the capacitance of the gap between the end of a strip-line resonator and a main strip-line, and ϕ is one-half of the electrical length of the resonator. The insertion loss in *db* between arms 1 and 4 is $20 \log_{10} |1/\rho|$, and therefore

$$\text{Insertion Loss} = 10 \log_{10} \left[1 + \left(\frac{2X}{Z_0} \right)^2 \right]$$

$$= 10 \log_{10} \left\{ 1 + 4 \left(\tan \phi - \frac{1}{\omega C Z_0} \right)^2 \right\}. \quad (7)$$

From this it is seen that the center frequency of the pass band occurs at

$$\tan \phi_0 = \frac{1}{\omega_0 C Z_0} \quad (8)$$

and the 3 db points at

$$\left| \tan \phi - \frac{1}{\omega C Z_0} \right| = \frac{1}{2}. \quad (9)$$

Eq. (7) shows that the mid-band insertion loss between arms 1 and 4 is zero, and therefore between arms 1 and 2 it is infinite. Far from resonance this condition becomes reversed, and hence the directional-filter response shown in Fig. 1(b) is verified. In the case of narrow bandwidth and infinite unloaded Q, the loaded Q of the filter may be expressed as follows in terms of the 3 db bandwidth determined from (9):

$$Q_L = \frac{f_0}{(\Delta f)_{3\text{ db}}} = \frac{\pi}{2} \left(\frac{1}{\omega C Z_0} \right)^2. \quad (10)$$

Circularly Polarized Waveguide-Cavity Filter

The loaded Q of the waveguide directional filter of Fig. 6(a) has been obtained by various methods, including the summation method used for the traveling-wave-loop filter. Eqs. (1), (2), (3), and (5) for transmission response, infinite-rejection condition, and insertion loss hold if dissipation loss in the end walls of the cavity is neglected, and if l is taken to be *twice* the height of the cavity (since one round trip of a wave front propagating axially between the end walls of the cavity corresponds to one traversal of the loop). c_1^2 is the total power coupled into the circularly polarized wave when a wave of unit power travels in the lower rectangular waveguide past the aperture in one direction, and c_2^2 is the corresponding factor for the upper waveguide. In evaluating c_1^2 and c_2^2, the circular waveguide should be assumed to be terminated by a nonreflecting load. When $c_1^2 = c_2^2 = c^2$, (4) holds if, in addition, n is taken to be the cavity height in *half* wavelengths, and if $(\lambda_g'/\lambda)^2$ is included as a factor, where λ_g' is guide wavelength in the circular waveguide. In the case of negligible dissipation loss, Q_L reduces to

$$Q_L = \frac{\pi n (1 - c^2)^{1/2}}{c^2} \left(\frac{\lambda_g'}{\lambda} \right)^2 \quad (11)$$

The power coupling factor c^2 may be calculated from the aperture dimensions by means of Bethe's small-aperture theory. A formula for a broad-band three-slot configuration [Fig. 9(a)] has been derived by one of the authors of this paper.[6] In using that formula it should be noted that it gives the power coupled into only one linearly polarized mode in the circular waveguide. Therefore c^2 is *twice* the value computed from that formula.

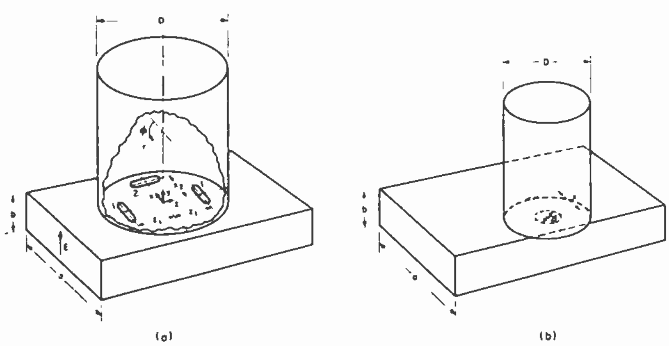

Fig. 9—Aperture arrangements for waveguide filter of 6(a). (a) Broad-band case; (b) narrow-band case.

If a single circular aperture is used as in Fig. 7, the power coupling factor is

$$c^2 = \frac{8\pi^2 \lambda_g d^6 \sin^2 \left(\frac{\pi x}{a} \right)}{27\, a^3 b D^2 \lambda_g'} \quad (12)$$

where a, d, D, and x are defined in Fig. 9(b), b is the cross-section height of the rectangular waveguide, λ_g is guide wavelength of the TE_{10} mode in the rectangular waveguide, and λ_g' is the guide wavelength of the TE_{11} mode in the circular waveguide. The aperture and the circular waveguide cross section are concentric. The spacing x to provide circular polarization is determined from

$$\tan \frac{\pi x}{a} = \frac{2a}{\lambda_g}. \quad (13)$$

The accuracy of (9) may be improved by corrections for wall thickness and for aperture resonance.[7] In the case of moderately large coupling ($c^2 \approx 0.03$) these effects have been found to cancel approximately.

[6] S. B. Cohn, "Impedance measurement by means of a broadband circular-polarization coupler," Proc. IRE, vol. 42, pp. 1554–1558; October, 1954.

[7] S. B. Cohn, "Microwave coupling by large apertures," Proc. IRE, vol. 40, pp. 696–699; June, 1952.

The axial length l of the cylindrical cavity is somewhat less than $\lambda_g'/2$ because of the tuning effect of the apertures. This length is given very closely by

$$\tan\left(\frac{\pi l}{\lambda_{g0}'}\right) = \frac{0.358 D^2 \lambda_{g0}'}{d^3} \tag{14}$$

Lumped-Constant Filter

The center frequency of the filter of Fig. 8(a) is

$$f_0 = \frac{1}{2\pi\sqrt{L(C + 2C_c)}}. \tag{15}$$

The loaded Q for narrow bandwidth and dissipationless elements is

$$Q_L = \frac{\omega_0 L}{R}\left(\frac{C}{C_c}\right)^2. \tag{16}$$

The reactance values in the phase-shift networks are shown in the figure.

Conclusion

The experimental results obtained with several different directional filter circuits verify their anticipated performance. A number of the circuits appear to be particularly practical for use in waveguide, TEM transmission line (*e.g.*, strip line or coaxial line), and lumped constants. Groups of directional filters may be used in many ways to perform frequency separation and combination functions. The matched input and low loss of the filter ensure good operational efficiency of such networks.

Acknowledgment

The authors wish to thank Dr. Otto Heinz for his assistance with the design, analysis, and experimental evaluation of the waveguide directional-filter model. The work reported in this paper was supported by the Signal Corps under Contract No. DA 36-039-sc-64625.

A New Technique for the Measurement of Microwave Standing-Wave Ratios*

A. C. MacPHERSON† AND D. M. KERNS‡

Summary—**A new microwave standing-wave-ratio (or reflection coefficient) measurement technique, apparently suitable for standards and other high precision work, is described. The technique requires that the phase angle of the unknown reflection coefficient be subjected to arbitrary, known variations—as is possible with a sliding load, for example. Accurate measurement of suitable sliding loads furnishes standards, and with the aid of comparison techniques, enables indirect measurement of arbitrary unknowns.**

Generator and detector are connected to two arms of a three-arm waveguide junction. Tuning elements in the junction are (desirably) adjusted for small reflection looking into the third arm, to which the unknown connects. Observation of the detector response vs variation of phase of the unknown yields a curve, similar to a standing-wave pattern, from which the unknown is determinable by procedures that are given.

The technique has the advantages of 1) being amenable to rigorous theoretical analysis, 2) enabling the attainment of heavy coupling to the detector without simultaneous severe distortion of the response pattern, and 3) being, in a sense explained in the text, fundamentally simpler than the conventional slotted-line technique.

Introduction

A NEW MICROWAVE standing-wave-ratio (or reflection coefficient) measurement technique, which appears to be suitable for standards and other high-precision work, is described. An earlier stage in the development of this technique has been reported previously by the authors.[1]

The technique requires that the phase angle of the unknown reflection coefficient be subjected to arbitrary, known variations. (A reflection coefficient whose phase can be subjected to such variations will be termed "phasable.") It is clear that this phasability requirement would be no limitation if an ideal phase-shift device were available.

Thus far the technique has been used primarily with specially constructed "sliding-loads" (terminating structures arranged to slide in the waveguide in which used—see Fig. 5). One other phasing method has been tried and will be mentioned. However, the sliding-load technique in itself offers a solution to most problems of standing-wave-ratio (swr) standardization. A suitable sliding load whose swr has been accurately measured constitutes a standard of swr. A series of such standards could be used, for example, to calibrate a conventional standing-wave machine. With the aid of sliding loads of adjustable swr and microwave-bridge comparison techniques, arbitrary unknowns can be measured indirectly.

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‡ National Bureau of Standards, Boulder, Colo.

[1] A. C. Macpherson and D. M. Kerns, "A New SWR Measurement Technique," URSI-IRE Meeting, Washington, D. C., April 27, 1953.

The experimental work done thus far has been concerned only with the determination of the magnitude of reflection coefficients. There is no reason why angles cannot also be determined, however, and the necessary theory and equations for this will be given.

The basic arrangement (see Fig. 1) consists essentially of a detector and a generator connected to two of the arms of a three-arm waveguide junction. This arrangement (generator and detector included) constitutes an instrument for the measurement of any phasable reflection coefficient presented at the reference plane in the third arm of the junction. The operating procedure consists essentially of observation of detector response as a function of the variation of phase of the unknown reflection coefficient. This yields a curve, similar to a standing-wave pattern, from which the unknown reflection coefficient is determinable by procedures to be given below.

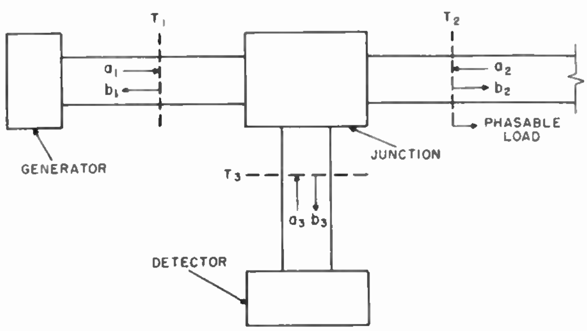

Fig. 1—Basic arrangement.

The junction is provided with tuning elements that are adjusted for zero reflection coefficient as seen looking into the load arm of the junction (with the detector and inactive generator connected). This matching operation is not critical; the technique does not assume that a zero or negligible reflection coefficient is actually obtained. The matching need be performed only once for a given setup operating within a limited frequency range. The matching operation reduces (or may practically eliminate) distortion of the response pattern and enables one to have heavy coupling to the detector without at the same time having a highly distorted response pattern.

It should be emphasized that the waveguide junction used in this technique is not subject to special requirements, such as symmetry or losslessness. The junction may indeed be of a rather general type, in that the waveguide leads may individually be of arbitrary cross section and one or more of them may be two-conductor systems, such as coaxial line. The main assumptions are that the electromagnetic field in a portion of each waveguide lead consists of a single waveguide mode and that the variation of the phase of the unknown is accomplished without affecting its magnitude. These two assumptions are only mildly unrealistic. The further requirement that the system (apart from the rf source itself, which must be well padded) be linear and passive is hardly restrictive. Reciprocity is not used in the main part of the analysis. It may be well to mention explicitly that it is not required that the detector or generator[2] present matched loads to the junction and that, by the same token, the joints in the waveguides connecting detector and generator to the junction are not required to be perfect.

The idea of using a fixed detector and a phasable load is not in itself new.[3] It is felt that the present work represents an advance in three main respects: 1) In significantly increased generality and rigor. Apparently it has not been realized that a much more thorough treatment could be carried out with very little additional complication in the results. 2) In enabling the attainment of heavy coupling to the detector without simultaneous distortion of the response pattern. In the present work the heavy coupling permitted the use of a bolometric substitution method of power measurement as the detector, virtually eliminating the question of detector linearity. Probably the principal advantage of having small distortion is that it permits simplification of the general procedure for the determination of the unknown reflection coefficient from the observed data; it may also enable increased accuracy. 3) In the application of an analysis (developed in a different context by Altar, Marshall, and Hunter[4]) which derives the desired information from the response pattern whether or not the pattern is highly distorted. It is remarkable that the analysis applies more rigorously in the present problem than it does in its original context.

The characteristic equation of the instrument—the analytical expression of detector power as a function of phase of the unknown reflection coefficient—is obtained by a rigorous analysis under the above hypotheses. The characteristic equation is of a simple form and contains effectively just two complex constants. In general both constants must be evaluated or accounted for empirically; the necessary procedures are relatively simple.

The new technique should also be compared with the conventional slotted-line standing-wave measurement technique. In Appendix I the slotted-line instrument is analyzed with rigor comparable to that applied in the text to the new technique, and it is shown that the characteristic equation for the slotted-line instrument (neglecting slot-end effects) contains four complex constants and is of a fairly complicated form. It is considered that the new technique has a fundamental and important advantage in the relative simplicity of its characteristic equation.

Theory

Our first theoretical task is to obtain and discuss the characteristic equation of the system.

[2] Here and subsequently the "generator" is to be understood as the equivalent generator as seen at the junction reference plane.

[3] C. G. Montgomery, "Technique of Microwave Measurements," M.I.T. Rad. Lab. Ser., The McGraw-Hill Book Co., Inc., New York, N.Y., Vol. 11, p. 507; 1947.

[4] Wm. Altar, F. B. Marshall, and L. P. Hunter, "Probe error in standing wave detectors," Proc. IRE, vol. 34, p. 33, January, 1946.

Let reference planes T_1, T_2, and T_3 be chosen in the generator, load, and detector waveguide leads, respectively, as indicated in Fig. 1. For the purpose of analysis, the definitions of the "junction" and the "terminations" are, of course, relative to the choice of reference planes. (It should be noted that any discontinuities introduced by flanged joints in the generator and detector leads will be taken care of implicitly by inclusion either in the junction or in the terminations.) It is convenient to characterize the junction (as defined by the choice of reference planes) by means of the inverse of its scattering matrix. Letting a_1, a_2, a_3 measure the amplitudes of the incident traveling-wave components of the field on the respective terminal surfaces, and letting b_1, b_2, b_3 similarly measure the amplitudes of the emergent traveling-wave components, we then have

$$\begin{aligned} a_1 &= G_{11}b_1 + G_{12}b_2 + G_{13}b_3 \\ a_2 &= G_{21}b_1 + G_{22}b_2 + G_{23}b_3 \\ a_3 &= G_{31}b_1 + G_{32}b_2 + G_{33}b_3 \end{aligned} \tag{1}$$

where the G_{ij} are the elements of the inverse scattering matrix G (G will be called the "gathering matrix" of the junction). The relations imposed at the terminal surfaces by the terminations we express in the form

$$\begin{aligned} a_1 &= S_g b_1 + b_g \\ a_2 &= Se^{-2j\theta}b_2 \\ a_3 &= S_d b_3 \end{aligned} \tag{2}$$

Here the reflection coefficient S_g and the fixed wave-amplitude b_g characterize the equivalent generator at the reference plane T_1, S is the unknown reflection coefficient, -2θ is the phase shift imposed upon S (so that $Se^{-2j\theta}$ is the reflection coefficient presented at reference plane T_2), and S_d is the reflection coefficient of the detector at T_3. (In the case of a sliding-load unknown, θ may be interpreted as the variable electrical distance between T_2 and the reference plane of the unknown.) Of these quantities only S and θ will be explicitly involved in the final results.

Combining (1) and (2) one obtains a set of three equations in three unknowns,

$$\begin{aligned} b_g &= (G_{11} - S_g)b_1 + G_{12}b_2 + G_{13}b_3, \\ 0 &= G_{21}b_1 + (G_{22} - Se^{-2j\theta})b_2 + G_{23}b_3, \\ 0 &= G_{31}b_1 + G_{32}b_2 + (G_{33} - S_d)b_3 \end{aligned} \tag{3}$$

which determines b_1, b_2, and b_3. The desired quantity, the power P delivered to the detector, is proportional to the square of the magnitude of b_3. We introduce the determinant

$$\Delta = \begin{vmatrix} G_{11} - S_g & G_{12} & G_{13} \\ G_{21} & G_{22} & G_{23} \\ G_{31} & G_{32} & G_{33} - S_d \end{vmatrix} \tag{4}$$

(note that this is not exactly the determinant of the above system of equations) and the two minors

$$M_{13} = G_{21}G_{32} - G_{31}G_{22} \tag{5}$$

$$M_{22} = (G_{11} - S_g)(G_{33} - S_d) - G_{31}G_{13}. \tag{6}$$

The power P can be expressed in the form

$$P = C\left| 1 + y + Y(KS, \theta) \right|^{-2} \tag{7}$$

where C is a constant of proportionality,

$$\left.\begin{aligned} K &= G_{31}/M_{13} \\ y &= \frac{-2(M_{22}/\Delta)}{K + (M_{22}/\Delta)} \\ Y(KS, \theta) &= \frac{1 - KS \exp(-2j\theta)}{1 + KS \exp(-2j\theta)} \end{aligned}\right\} \tag{8}$$

the last being a functional notation. The significance of expressing K and y in terms of the quantities (4), (5), and (6) will become apparent. The "constants" C, y, and K are of course frequency-dependent. Eq. (7) is the desired characteristic equation.

The value of the constant C is immaterial as far as the analysis of the response curves is concerned, inasmuch as only relative values of P will be involved. Thus the equation contains effectively just two parameters, y and K, that characterize the instrument. Of course y and K are in general complex so that four real parameters are involved.

We observe that the characteristic constant K and the unknown reflection coefficient S appear in the characteristic equation only in the combination KS. In what follows it will be convenient to deal with the product KS as an intermediate unknown and to postpone the question of determining S itself. The notation $S' = KS$ will be used.

As a function of S' and θ, the characteristic equation is of the same form as the highly simplified equation for the slotted-line standing-wave machine given in Appendix II. (This is the basis of our application of the Altar, Marshall, and Hunter analysis; see below and Appendix II.) Thus it is recognized that in general the graph of the characteristic equation will have the form of a distorted standing-wave pattern, the distortion being introduced by the presence of the quantity y and vanishing for $y = 0$. An experimentally obtained graph in which the distortion is rather small is shown in Fig. 2. It is interesting to note that the quantity y can be regarded as a thoroughgoing generalization of the "probe admittance" y_p appearing in (16); y depends on S_g, S_d, and on all the elements of the matrix G and will in general be nonzero whether or not the system actually contains a probe.

If an ordinary waveguide junction (for which reciprocity holds and in which dissipation is small) is employed, qualitative statements can be made about the magnitudes of K and y:

1) The magnitude of K will be approximately equal to unity. This follows from the fact that under the assumed conditions the gathering matrix will be symmetric

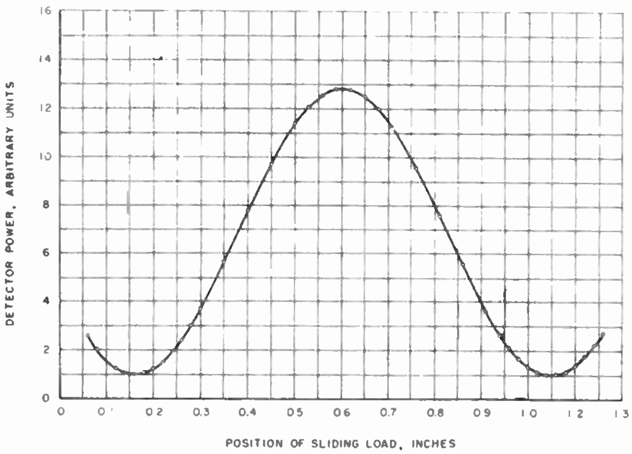

Fig. 2—Detector power vs load position—experimental.

and nearly unitary,[5] K becomes equal to the ratio of the element G_{31} to its cofactor in G, and (arguing by continuity from a well-known property of unitary matrices) the magnitude of this ratio will be approximately equal to unity.[6]

2) The matching operation described previously will make y small (relative to unity). This may be established as follows. Let S_{2i} denote the reflection coefficient that is observed and minimized in the matching operation; it may be shown without difficulty that

$$S_{2i} = M_{22}/\Delta$$

where Δ and M_{22} are defined in (4) and (6). Hence the expression for y may be written

$$y = -\ 2S_{2i}/(K + S_{2i}).$$

Now, although K as well as S_{2i} will be affected by the matching operation, the magnitude of K will remain approximately equal to unity (as shown in the preceding paragraph). Thus it is seen that making S_{2i} small will make y small.

A small or vanishing value for the distortion term y is not incompatible with heavy coupling to the detector. To establish this it is sufficient to consider an ideal case. Assume that a symmetrical lossless H-plane T junction is used in the arrangement of Fig. 4, and assume further that $S_g = S_d = S_{2i} = S = 0$. Then it can be shown with little difficulty that of the power from the generator incident on the junction 25 per cent is reflected, 25 per cent goes to the load, and 50 per cent to the detector.

We turn now to the second main theoretical task, which is to provide techniques for handling the following problem: given an empirically determined graph of P vs θ for a given S', together with the form of the characteristic equation, to determine S' and y. It happens that techniques developed by Altar, Marshall, and Hunter[7] (AMH) in connection with an approximate treatment of probe error in standing-wave machines are rigorously applicable in the present problem.[8] Both graphical and analytical techniques were provided; here we shall give a selection of the analytical results. These results have been checked by employing an independent method of derivation and are presented with minor emendations and modifications in form.

The magnitude of S' may be determined as follows. Let P_{max}, P_{min}, and P_{mid}, respectively, denote detector power readings at maxima, minima, and "midpoints" of the P vs θ curve; P_{mid} being defined by

$$P_{mid} = \frac{2P_{max}\,P_{min}}{P_{max} + P_{min}}.$$

Let θ_{max}, θ_{min}, and θ_{mid} denote angular positions corresponding to P_{max}, P_{min}, and P_{mid}, respectively. In order to avoid ambiguities in sign and in multiples of π, the angles may be chosen in the manner shown in Fig. 3.

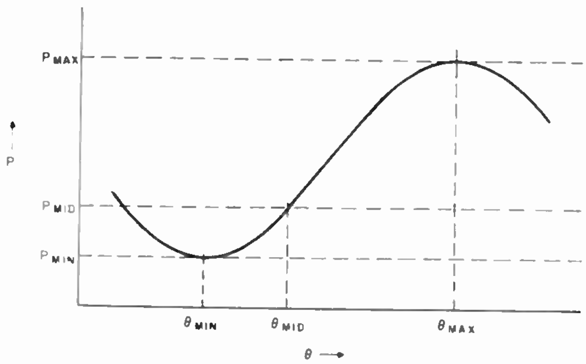

Fig. 3—Illustrating choice of θ_{min}, θ_{max}, and θ_{mid}.

Further, let

$$\eta = \theta_{max} - \theta_{min}$$
$$\epsilon = \theta_{max} + \theta_{min} - 2\theta_{mid}$$

and define the auxiliary quantity $\bar{\eta}$ by

$$\sin \bar{\eta} = \frac{\sin \eta}{\sqrt{1 + \left(\dfrac{\sin \epsilon \sin \eta}{\cos \epsilon - \cos \eta}\right)^2}}$$

with the proviso that $\bar{\eta}$ is to be taken in the first quadrant. Then the magnitude of the intermediate reflection coefficient is

$$|S'| = \frac{1 - \tan(\bar{\eta}/2)}{1 + \tan(\bar{\eta}/2)}. \tag{9}$$

[The voltage standing-wave-ratio (vswr) corresponding to S' is $\rho' = \cot(\bar{\eta}/2)$.]

[5] The argument assumes that the characteristic impedances in the three waveguide leads are chosen equal, as is always possible; the result established is independent of the choice of characteristic impedances.

[6] It is worth noting that a K wholly similar to the above K appears in the rigorous characteristic equation for the slotted-line standing-wave machine (appendix I). In slotted line work it has generally been assumed (in effect) that $K=1$. In the present technique the use of an ordinary junction would permit the assumption of the *magnitude* of K equal to unity with about the same justifiability as in the slotted line case. This assumption is not made in the present work, which is concerned with the development of a measuring technique under minimum assumptions.

[7] Montgomery, *loc. cit.*

[8] The relation between their problem and our problem is explained in Appendix II.

The angle σ' of S' may be determined as follows. The shift of the minima corresponding to the presence of the term y in the characteristic equation may be defined as

$$\delta = \theta_{\min} - \theta_{\min}^{(t)}$$

where $\theta_{\min}^{(t)}$, the "true" minimum position, is the value that $\theta_{\min}$ would have if $Im(y)$ were zero. It may easily be determined that

$$\sigma' = 2(\theta_{\min} - \delta) + \pi \tag{10a}$$

apart from multiples of 2π. The value of δ is given in terms of η and ϵ by[9]

$$\delta = \frac{1}{2}\sin^{-1}\left(\frac{\sin\eta}{\sqrt{1+\cos^2\eta\left(\dfrac{\cos\epsilon-\cos\eta}{\sin\epsilon\sin\eta}\right)^2}}\right) - \frac{1}{2}\eta. \tag{10b}$$

Here the inverse sine is to be taken in the 1st or 2nd quadrant according to whether η is in the 1st or 2nd quadrant.

It will be observed that the above formulas do not involve the characteristic constant y explicitly and are applicable whether or not y is small. However, when y is small, certain approximate formulas involving the real and imaginary parts of y explicitly may be preferable. The value of y is obtainable from an expression given by AMH (p. 38P):

$$y = \frac{2}{(1-|S'|^2)}\left(\left|\frac{S'}{S_0}\right|e^{i\psi} - 1\right). \tag{11}$$

Here S' is as already defined, $|S_0|$ is defined in terms of the directly observed vswr $\rho_0 = \sqrt{P_{\max}/P_{\min}}$ by $|S_0| = (\rho_0-1)/(\rho_0+1)$, and ψ (which is to be taken between $-\pi/2$ and $+\pi/2$) is given by $\tan\psi = \csc\epsilon\cot\eta\cdot(\cos\epsilon - \cos\eta)$. Let us denote the real and the imaginary parts of y by g and b, respectively. It can be shown that, to the first order in g and b, the intermediate vswr is given by[10]

$$\rho' = \rho_0 + \rho_0 g\frac{\rho_0 - 1}{\rho_0 + 1} \tag{12}$$

and the shift of the minima by

$$\delta = \frac{-b}{(\rho_0+1)^2}. \tag{13}$$

Both of these expressions furnish small corrections to be applied to the directly observed quantities.

Thus far we have provided formulas for the determination of S'; in order to be able to obtain the actual unknown reflection coefficient, S, the value of the characteristic constant K must be known. Inasmuch as K and S appear in the characteristic equation only in the combination $S'=KS$, the empirical evaluation of K requires that S' be determined for at least one phasable load whose reflection coefficient ($S=S_s$, say) is known independently (K being then given simply by $K=S'/S_s$). Probably the type of load for which the reflection coefficient can best be determined independently is a "short." A sliding short was used by the authors and the means of evaluating it are described briefly in the following section.

As soon as K is known, the value of any S is of course at once obtainable from the corresponding value of S'.

Experimental

For purposes of illustration, some of the equipment and practices employed in the experimental part of the present work will be described. The arrangement used was designed with the idea of obtaining the highest possible accuracy. Much attention was given to problems of generator stability, detector linearity, detector noise and drift, and mechanical precision. The complete set-up is shown in Fig. 4. The rf source is an X-21 klystron in cw operation. This tube yields about 5 watts output, which easily permits the use of adequate padding, and in addition is very stable in power level and frequency. For ordinary work the X-21 has the disadvantage of being essentially nontunable. (An arrangement that would probably be sufficiently stable and yet adjustable in frequency would be to use a tunable reflex klystron operating in an oil bath, with a unidirectional device for padding.) Means are provided for adjusting the power level and monitoring the frequency. The waveguide junction used is an H-plane tee provided with built-in tuning elements. The detector is a commercial platinum-wire barretter in a commercial mount operated in an oil bath. The barretter is connected into a manually-balanced dc wheatstone bridge.

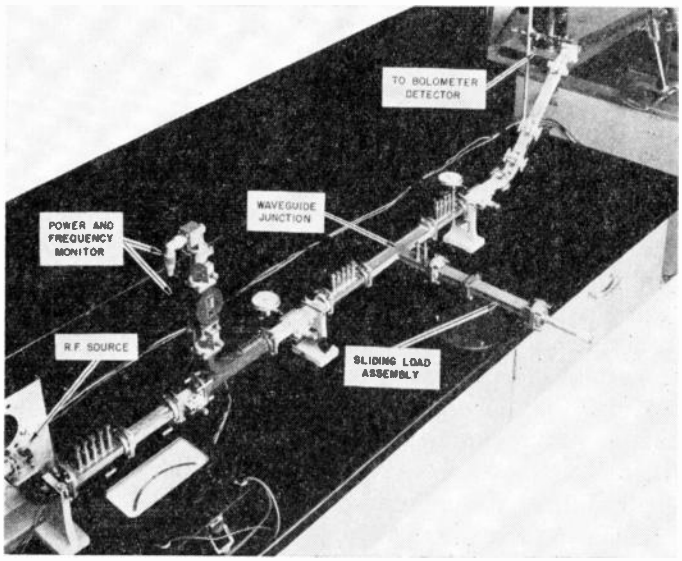

Fig. 4—Laboratory set-up.

[9] In AMH the quantity measuring the shift of the minima is μ, and $\mu=-2\delta$. The factor of 2 is accounted for by the different angular measure used by AMH and the negative sign is presumably due to an opposite definition of the shift. (The approximate formula for μ given by AMH is in error.)

[10] The expressions given here can be derived along the lines indicated by Montgomery, *op. cit.*, pp. 483–488. (The result given for the shift of the minima in this reference should be multiplied by 2π).

Since the bridge is balanced each time a reading is made, the rf impedance of the detector is virtually independent of the rf power level. Power generation and detection are relatively simple since no modulator is needed and the detector requires only a simple dc network and uses no vacuum tubes.

Three more or less typical sliding loads are shown in Fig. 5. The top one in the figure is hand driven and consists of a steel slug (which is a close sliding fit in the waveguide) to which is attached a tapered resistance strip. The load shown in the middle of the figure is similar except that the lossy material is a slug of Catalin and it is driven by a micrometer screw. The load at the bottom, designed by Ivan K. Munson, uses Polyiron as the dissipative material and employs a short transverse wire mounted on the end of an axial dielectric rod that can be rotated. This permits adjustment of the swr. The waveguides in which the loads slide must be of highly uniform cross section; they were made by electro-forming copper on precision mandrels. It should be noted that whatever discontinuity there may be at the flange connection between the junction and the guide in which the load slides is automatically associated with the junction and not with the sliding load.

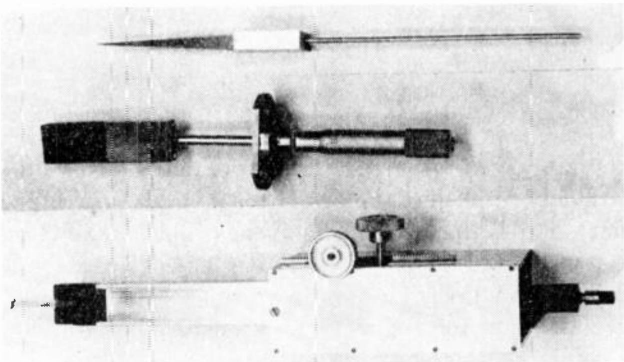

Fig. 5—A selection of sliding loads.

A sleeve-type "line-stretcher" for phase shifting purposes is shown in Figs. 6 and 7. This device permits the direct measurement of portable loads (that are not too bulky or heavy). It is the outcome of a suggestion by H. Lyons and was developed by A. J. Couvillion. An important point to be brought out in connection with this phase shifter is that discontinuities such as those at A and B are quite innocuous (if they are constant). These discontinuities are spatially fixed with respect to the H-plane junction when the sleeve is moved and therefore are automatically taken care of in the measurement procedure (cf previous section). The effective unknown is that seen between B and the joint F and thus the effects of the joint are associated with the unknown in the present instance. Some tests were made using sliding-load assemblies as loads on the "line-stretcher." Preliminary measurements using the "line-stretcher" are encouraging but much more thorough

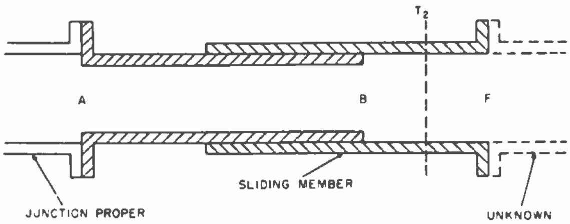

Fig. 6—Sleeve-type "line-stretcher"—schematic; sliding member is waveguide of standard internal dimensions. Slides over fixed member of reduced size attached to junction proper. T_2 is the junction reference plane; it is fixed in position relative to the junction.

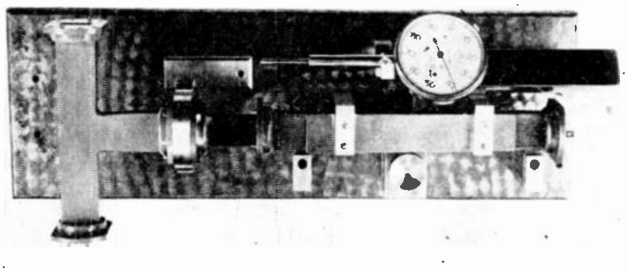

Fig. 7—Sleeve-type "line-strecher"—mechanical.

evaluation is necessary. If mechanically feasible, an rf choke at B would seem desirable in order to reduce the effect of variable sliding contacts.

As mentioned in the previous section a load whose reflection coefficient S_s is known is needed for the evaluation of K. In the experimental work only the magnitude of S_s was considered. A choked sliding short was chosen and two independent methods of evaluating $|S_s|$ were employed. 1) A cavity was formed from a straight section of waveguide terminated at both ends with shorts of the type being evaluated. The Q of this cavity was measured and $|S_s|$ was calculated assuming that the two shorts were identical and that the waveguide was lossless. This is a sensitive method giving a lower limit for $|S_s|$. (Inasmuch as $|S_s|$ cannot exceed 1, the problem can be regarded largely as that of determining lower bounds for $|S_s|$.) 2) Using well-known microwave bridge techniques, the sliding short was compared with a fixed, soldered, silver plate short whose reflection coefficient was calculated from the geometry and the dc conductivity involved. The two methods gave good agreement, and it is considered that $|S_s|$ for the load measured was almost certainly within the range (0.990, 1.000).

In the arrangement described above there are two main sources of random error. Under good conditions, the effects of electrical noise and drift due to both the generator and the detector amount to about one microwatt during the time necessary to make one power-ratio measurement. Since the detector power level for a nearly matched load is about 10 milliwatts, the power resolution is about one part in ten-thousand, and the corresponding minimum detectable vswr is about 1.00005. This figure is not realized because of the additional "noise" associated with the mechanics of the phase shifting process. Some measurements made on a sliding load of nominal vswr 1.01 give an idea of the mechanical stability possible. Six vswr values were computed from readings of maxima and minima with the load at 7 different positions in the waveguide; the mean was 1.0166 and the greatest deviation from the mean was 0.0003. With loads of higher vswr, and with the "line-stretcher" phase shifter, the reproducibility in terms of vswr was poorer in general. Nevertheless, the mechanical problem for a sliding load seems considerably easier than for a slotted line.

Appendix I

Characteristic Equation for Standing-Wave Machine

We give a brief but rigorous derivation of the characteristic equation for a conventional standing-wave machine. The general approach and much of the notation used in the second section will be applied here.

Let reference planes T_1, T_2 be chosen on the generator side and on the load side of the probe assembly, respectively (see Fig. 8). These two reference planes,

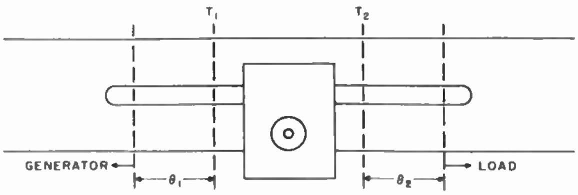

Fig. 8—Notation for standing-wave machine; coaxial line for probe output is indicated in the sketch.

which are to move with the probe assembly, plus a third one T_3 in the rf probe output, define a 3-arm waveguide junction. We denote the gathering matrix of this junction by G and use the notation a_1, a_2, a_3, and b_1, b_2, b_3 for the incident and emergent wave amplitudes, as in the second section. The relations imposed by the terminations are

$$a_1 = S_g e^{-2j\theta_1} b_1 + e^{-j\theta_1} b_g$$
$$a_2 = S e^{-2j\theta_2} b_2$$
$$a_3 = S_d b_3$$

[cf (2)], where θ_1 is the electrical distance between T_1 and the generator reference plane and θ_2 is the same between T_2 and the load reference plane. The "generator" and the "load" dealt with here are, of course, as seen within the slotted portion of the slotted line. When the probe assembly is moved, the line length on one side increases and that on the other decreases in such a way that $\theta_1+\theta_2$ remains constant.

Solving for b_3 in the problem just set up yields a result expressible in the form

$$b_3 = \frac{c \exp(-j\theta_1)}{1 + y + mS_g \exp(-2j\theta_1) + [1 + nS_g \exp(-2j\theta_1)]Y(KS, \theta_2)}$$

where c, y, m, n, and K are constants independent of θ_1 and θ_2, and $Y(KS, \theta_2)$ is the function defined in (8). The characteristic equation, which is an expression for the rf power P delivered to a detector on arm 3, may be written

$$P = C \left| 1 + y + m' \exp(-2j\theta_1) + [1 + n' \exp(-2j\theta_1)]Y(KS, \theta_2) \right|^{-2}, \quad (14)$$

where we have put $m' = mS_g$ and $n' = nS_g$. The four characteristic constants are y, m', n', and K. This K is the same function of the elements of G as the K of the text and thus shares the property $|K| = 1$ for a lossless junction fulfilling the reciprocity condition. It is worth noting that if the generator is matched ($S_g = 0$), then m' and n' vanish and (14) becomes formally identical to (7), as it should.

Appendix II

Problem Treated by AMH

In the treatment of the conventional standing-wave machine given by AMH the two main assumptions involved are 1) that the generator is matched (this was tacit), and 2) that the effects of the probe assembly may be represented by a pure shunt element in a lossless transmission line. These assumptions lead at once to an expression for the voltage V across the equivalent transmission line at the probe position,

$$V = \frac{c}{1 + y_p + Y(S, \theta)} \quad (15)$$

where c is a constant, y_p is the admittance associated with the probe, and $Y(S, \theta)$ is the normalized load-admittance referred to the probe position. If it is assumed that the power delivered to a detector is proportionl to $|V|^2$, (15) leads to the characteristic equation

$$P = C \left| 1 + y_p + Y(S, \theta) \right|^{-2}. \quad (16)$$

Although AMH do not actually write down equations of the form of (15) and (16), their analysis does in fact pertain to such equations. Upon comparing (16) with the characteristic equation of the text (7), the applicability of the analysis given by AMH becomes obvious.

Acknowledgment

The authors are grateful to Robert Sullivan, for help in many measurements and calculations, and to Harold Lyons, for his interest in encouragement in this work.

Novel Circuit for a Stable Variable Frequency Oscillator*

DAVID MAKOW†, MEMBER, IRE

Summary—**In the past, quartz crystals have been used extensively to control single frequencies. This paper describes a circuit where a quartz crystal exercises considerable control over a continuously variable frequency spectrum.**

The circuit is characterized by multiloop feedback, which enables oscillation to be maintained at three frequencies, with the corresponding voltages confined to certain branches of the circuit and coupled to each other through three mixers. The frequencies of oscillation, to be called f_1, f_2, and f_3, are determined by zero phase shift conditions in the feedback loops and by one frequency condition, $f_1+f_2=f_3$. The frequency f_3 is restricted by a quartz crystal resonator, whereas the frequencies f_1 and f_2 are tunable in such a manner, that as f_1 increases, f_2 decreases and vice versa. The LC resonators which control f_1 and f_2 drift in the same direction when the ambient conditions change. Such a drift cannot influence f_1 and f_2, since the latter can change only when the LC resonators shift in the opposite direction. The presence of a quartz crystal in the feedback loop also improves the phase stability of the circuit.

An experimental model has been tested, and shows a thirty to hundredfold reduction of the frequency drift with temperature, excellent short and long term stability and a remarkably small initial frequency drift with switching on.

Introduction

A MAJOR problem in the development of variable frequency LC oscillators is the frequency stability of the generated voltage. Numerous references to circuits exist in the literature, in which the influence of the power supply and amplitude variations upon the frequency of oscillation is minimized.[1–5] Even though such variations were eliminated entirely, the frequency stability of an LC oscillator would never exceed the stability of its resonator, whose center frequency drifts with temperature, humidity, CO_2 content in the air, atmospheric pressure, and with aging. In practice, the influence of the resonator drift on the frequency stability is in most cases many times greater than that of the power supply and amplitude variations. The drift of the resonator can be decreased by controlling some of the factors mentioned above, for example by means of temperature regulations and by sealing. Relatively little, however, has been done to the writer's knowledge to provide circuitry by which the effect of resonator drift on frequency could be reduced or eliminated.

* Original manuscript received by the IRE, November 28, 1955; revised manuscript received, March 14, 1956.

† Radio and Electrical Eng. Div., National Res. Council, Ottawa, Can.

[1] G. G. Gouriet, "High-stability oscillator," *Wireless Engr.*, vol. 27, pp. 105–112; April, 1950.

[2] J. K. Clapp, "An inductance-capacitance oscillator of unusual frequency stability," PROC. IRE, vol. 36, pp. 356–358; March, 1948.

[3] F. B. Llewellyn, "Constant frequency oscillators," PROC. IRE, vol. 19, pp. 2063–2094; December, 1931.

[4] H. A. Thomas, "Theory and Design of Valve Oscillators," Chapman & Hall Ltd., London; 1951.

[5] T. Vackar, "L-C Oscillators and Their Frequency Stability," *Tesla Technical Reports*, pp. 1–9; December, 1949.

This paper describes a novel variable frequency oscillator circuit in which a considerable reduction in frequency variation with resonator drift has been realized. By suitable multiloop circuit arrangement, three oscillations are maintained at three separate frequencies, such that one is the sum of the other two. The sum frequency is fixed and controlled precisely by a quartz crystal, while the other two frequencies are controlled by variable LC resonators which may be set to any pair of frequencies adding up to the desired sum. Then if the center frequency of the LC resonators at a given setting should drift in opposite directions, there would be little or no change in the sum and the crystal would have little or no control over the pair of frequencies of oscillation. But, fortunately, such drifts are almost always in the same direction, because they are usually due to changes in temperature, humidity, atmospheric pressure and so forth, so that the sum frequency tends to change appreciably. Thus in the usual case the crystal is able to exercise a considerable degree of control over the drifts in the frequencies of oscillation in each of the two secondary circuits.

The two variable LC resonators are actually ganged to be tunable in opposite directions, so that the sum of their center frequencies remains fixed. Thus they are readily set to any desired pair of frequencies over a wide range, while the stability of the frequencies of oscillation at any given setting is considerably increased by the crystal in the third circuit.

Theory and Description of Operation

Conditions of Oscillation

Consider the block diagram shown in Fig. 1. *M*1, *M*2, and *M*3 are mixers which generate sum and difference frequency components of the voltages present at their respective signal (s_1, s_2, and s_3) and carrier (c_1, c_2, and c_3) inputs. A_1, A_2, and A_3 are tuned amplifiers which amplify the useful output frequency component of the corresponding mixer. The output of each amplifier is applied to the signal input of the following mixer and to the carrier input of the next following mixer, so that a closed loop system is formed as shown in Fig. 1.

Assume that the branches *A*, *D*, and *G* are opened at the points α, β, and γ, and that the voltages $V_1 \cos(2\pi f_1 t+\phi_1)$, $V_2 \cos(2\pi f_2 t+\phi_2)$, and $V_3 \cos \cdot(2\pi f_3 t+\phi_3)$ respectively, are applied there, where: V_1, V_2, and V_3 are the amplitudes, f_1, f_2, and f_3 are the frequencies, not rationally related, $f_1<f_2<f_3$, and ϕ_1, ϕ_2, and ϕ_3 are the relative phases. These voltages are amplified in the tuned amplifiers A_1, A_2, and A_3 respectively and are then applied to the corresponding mixers. The amplifiers

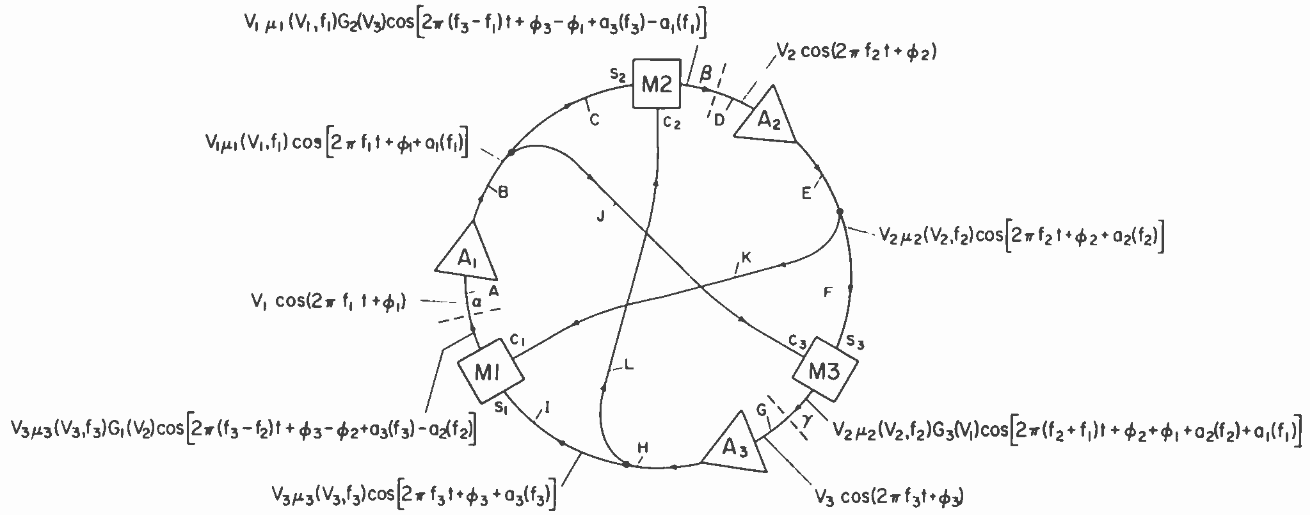

Fig. 1—Block diagram of the novel circuit for a stable variable frequency oscillator. A_1, A_2, and A_3 are tuned amplifiers having the amplification factors $\mu_1(V_1 f_1)$, $\mu_2(V_2, f_2)$, and $\mu_3(V_3, f_3)$ and introducing the phase angles $a_1(f_1)$, $a_2(f_2)$, and $a_3(f_3)$ respectively. $M1$, $M2$, and $M3$ are mixers having the conversion gains $G_1(V_2)$, $G_2(V_3)$, and $G_3(V_1)$.

A_1, A_2, and A_3 have amplification factors $\mu_1(V_1, f_1)$, $\mu_2(V_2, f_2)$, and $\mu_3(V_3, f_3)$ which are, in general, functions of the amplitude and frequency of oscillations, and introduce the phase angles $a_1(f_1)$, $a_2(f_2)$, and $a_3(f_3)$ respectively. The mixers, $M1$, $M2$, and $M3$ have the conversion gains (less than unity) $G_1(V_2)$, $G_2(V_3)$, and $G_3(V_1)$ respectively, which are defined as ratios of the useful output voltage component to the signal voltage. The conversion gain depends, in general, on the carrier voltage, and approaches a constant value G_1, G_2, or G_3 respectively as the carrier voltage becomes large.

The condition for oscillation is that each of the useful voltage components found at the outputs of the mixers $M1$, $M2$, and $M3$ (the difference frequency components of $M1$ and $M2$ and the sum frequency component of $M3$, since $f_1 < f_2 < f_3$) must equal the voltage components assumed at α, β, and γ respectively, in amplitude, frequency, and phase. Then assuming linear conditions, and equating the voltages on each side of the breaks α, β, and γ (see Fig. 1)

$$V_1 \cos(2\pi f_1 t + \phi_1) = V_3\mu_3(V_3, f_3)G_1(V_2) \cos\{2\pi(f_3 - f_2)t + \phi_3 - \phi_2 + a_3(f_3) - a_2(f_2)\} \quad (1)$$

$$V_2 \cos(2\pi f_2 t + \phi_2) = V_1\mu_1(V_1, f_1)G_2(V_3) \cos\{2\pi(f_3 - f_1)t + \phi_3 - \phi_1 + a_3(f_3) - a_1(f_1)\} \quad (2)$$

$$V_3 \cos(2\pi f_3 t + \phi_3) = V_2\mu_2(V_2, f_2)G_3(V_1) \cos\{2\pi(f_2 + f_1)t + \phi_2 + \phi_1 + a_2(f_2) + a_1(f_1)\}\cdot \quad (3)$$

The amplitude conditions are obtained by equating the amplitudes in (1), (2), and (3). These can be readily simplified assuming steady state. During the build-up of the oscillation the amplification factors in general decrease due to overloading, and the conversion gains increase as a result of the rising amplitude. The steady state is reached when the losses in the mixers are balanced by the gains in the amplifiers. The carrier voltages are then relatively large and the conversion gains of the mixers will be constant and equal to G_1, G_2, and G_3. The amplification factors depend little on frequency in the neighborhood of the center frequency of the resonators, and they assume the values μ_1, μ_2, and μ_3. Then considering the amplitudes V_1, V_2, and V_3 to be equal, which is fulfilled when the amplification factors of the amplifiers and conversion gains of the mixers are equal, the amplitude conditions reduce to

$$\mu_3 G_1 = \mu_1 G_2 = \mu_2 G_3 = \mu G = 1. \quad (4)$$

The coupling between the outputs of the amplifiers and the carrier and signal inputs of the following mixer is such that the ratio of the carrier-to-signal voltages is about three or more. The coupling circuit is not considered here, since it is not essential for the discussion.

Equating the phases in (1), (2), and (3), leads to the frequency condition

$$f_1 + f_2 = f_3 \quad (5)$$

and to the phase conditions

$$a_1(f_1) - a_2(f_2) = 0 \quad (6)$$

$$a_3(f_3) + a_2(f_2) = a_3(f_3) + a_1(f_1) = 0. \quad (7)$$

These three conditions determine the values of the frequencies of oscillation f_1, f_2, and f_3, which can be maintained in the system. This is unlike conventional oscillators in which the frequency is determined by one phase condition only. Eq. (6) states that the total phase shift in the loop made up of the branches A, B, C, D, E, and K must be zero. The sign is negative because of the action of the Mixer $M2$. Likewise, (7) states that the total phase shift in the loop made up of the branches G, H, L, D, E, and F or in the loop made up of the branches G, H, I, A, B, and J must be zero.

Frequency Stability

The values of the frequencies of oscillation f_1, f_2, and f_3, which are maintained in the system can be found by solving (5), (6), and (7). The phase angles $a_1(f_1)$, $a_2(f_2)$, and $a_3(f_3)$ are expressed by the fractional detuning of the center frequencies f_{10}, f_{20}, and f_{30} of the series resonant circuits in A_1, A_2, and A_3 from f_1, f_2, and f_3 respectively and their quality factors Q_1, Q_2, and Q_3. Assuming small detuning and $Q_1=Q_2=Q>10$

$$f_1 = \sqrt{\frac{Q}{(Q+Q_3)} f_{10}^2 + \frac{Q_3}{(Q+Q_3)}\left(\frac{f_{30}f_{10}}{f_{10}+f_{20}}\right)^2} \tag{8}$$

$$f_2 = \sqrt{\frac{Q}{(Q+Q_3)} f_{20}^2 + \frac{Q_3}{(Q+Q_3)}\left(\frac{f_{30}f_{20}}{f_{20}+f_{10}}\right)^2} \tag{9}$$

$$f_3 = f_1 + f_2 = \sqrt{\frac{Q}{(Q+Q_3)} f_{10}^2 + \frac{Q_3}{(Q+Q_3)}\left(\frac{f_{30}f_{10}}{f_{10}+f_{20}}\right)^2} + \sqrt{\frac{Q}{(Q+Q_3)} f_{20}^2 + \frac{Q_3}{(Q+Q_3)}\left(\frac{f_{30}f_{10}}{f_{20}+f_{10}}\right)^2}. \tag{10}$$

The drift of the center frequencies f_{10}, f_{20}, and f_{30} of the resonators affects the frequencies of oscillation to an extent which can be calculated by taking the total differentials of (8), (9), and (10) and then assuming that $f_1 \approx f_{10}$, $f_2 \approx f_{20}$, and $f_3 \approx f_{30}$.

$$\frac{df_1}{f_1} = \frac{Q}{(Q+Q_3)}\frac{df_{10}}{f_{10}} + \frac{Q_3}{(Q+Q_3)}\frac{df_{30}}{f_{30}} - \frac{Q_3}{(Q+Q_3)}\frac{S}{(1+S)}\left\{\frac{df_{20}}{f_{20}} - \frac{df_{10}}{f_{10}}\right\} \tag{11}$$

$$\frac{df_2}{f_2} = \frac{Q}{(Q+Q_3)}\frac{df_{20}}{f_{20}} + \frac{Q_3}{(Q+Q_3)}\frac{df_{30}}{f_{30}} - \frac{Q_3}{(Q+Q_3)}\frac{1}{(1+S)}\left\{\frac{df_{10}}{f_{10}} - \frac{df_{20}}{f_{20}}\right\} \tag{12}$$

$$\frac{df_3}{f_3} = \frac{Q}{(Q+Q_3)}\left\{\frac{1}{(1+S)}\frac{df_{10}}{f_{10}} + \frac{S}{(1+S)}\frac{df_{20}}{f_{20}}\right\} + \frac{Q_3}{(Q+Q_3)}\frac{df_{30}}{f_{30}} \tag{13}$$

where

$$S = \frac{f_{20}}{f_{10}}.$$

Consider the case when the resonator in A_3 consists of a quartz crystal in series resonance, then Q_3 is much larger than Q. The first terms in (11) and (12) become very small and the second terms are inherently small due to the high stability of the quartz crystal. The third terms can be made to approach zero by using identically the same construction for the resonators in A_1 and A_2, and by employing inductors and capacitors of the same type, so that they will drift equally with temperature, humidity, atmospheric pressure, CO_2 content in the air, and probably with aging. Under these conditions the stability of f_1 and f_2 will be very high, since the right sides of (11) and (12) are of very small magnitude. Also a partial or total compensation of the three terms may take place because of the negative sign of the third term, and this may further contribute to the stability of f_1 and f_2.

Eqs. (11) and (12) can be interpreted physically with the help of Fig. 2. The functions $a_1 = F_1(f_1)$ and $a_2 = F_2(f_2)$

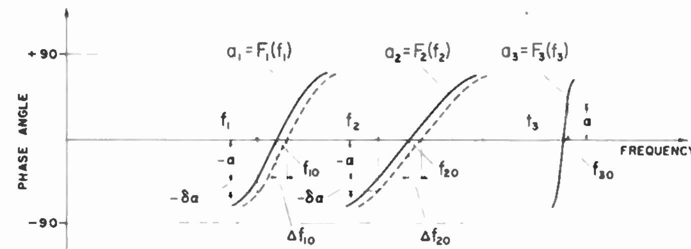

Fig. 2—Phase-frequency characteristics $a_1 = F_1(f_1)$, $a_2 = F_2(f_2)$, and $a_3 = F_3(f_3)$ of the tuned amplifiers A_1, A_2, and A_3 respectively.

are the phase characteristics of the series resonant circuits in the amplifiers A_1 and A_2, and have equal Q values. The function $a_3 = F_3(f_3)$ is the phase characteristic of a high Q quartz crystal in series resonance in the amplifier A_3. The frequencies f_1, f_2, and f_3 are chosen in such a way that $f_1 + f_2 = f_3$ and $-a_1(f_1) = -a_2(f_2) = a_3(f_3) = \alpha$, in accordance with the theory discussed above. It is possible to choose f_{10} and f_{20} so $\alpha = 0$, but in practice there will always be a small phase increment α introduced by the tuned amplifiers. Suppose that the functions $a_1 = F_1(f_1)$ and $a_2 = F_2(f_2)$ drift by the relative amounts $\Delta f_{10}/f_{10} = \Delta f_{20}/f_{20}$ which would take place for example with a change in ambient temperature. The drift of the function $a_3 = F_3(f_3)$ can be neglected since it is comparatively small. As a result of the drift of the former, additional phase increments $-\delta\alpha$ will be introduced to the frequencies f_1 and f_2 which can not follow this drift fully because the condition $f_1 + f_2 = f_3$ would be violated. Then the phase condition $-a_1(f_1) = -a_2(f_2) = a_3(f_3)$ can be fulfilled only when f_3 increases by an amount Δf_3 which corresponds to an increase in the phase angle α by $\delta\alpha$. The shift in f_3 of Δf_3 is associated with an increase in f_1 of Δf_1 and in f_2 of Δf_2, in order to fulfill the frequency condition (5), and the magnitude of Δf_1 and Δf_2 is very small since Q_3 is very high. A minute decrease in the phase increment $-\delta\alpha$ then takes place and in turn causes f_1, f_2, and f_3 to decrease slightly. This action continues in diminishing degrees until the conditions (5), (6), and (7) are fulfilled simultaneously. It is seen that the drift Δf_{10} and Δf_{20} of the resonators in A_1 and A_2 respectively, causes the frequencies f_1 and f_2 to shift by only a small fraction of this value, and not by an equal amount as would be the case in a conventional oscillator.

Although the above discussion was carried out in terms of the drifts of resonators in A_1 and A_2, it is also useful in considering phase stability. Any phase increments introduced in branches of this system, for instance as a result of a change in the interelectrode capacitance of the tubes due to power supply variations, can be represented by an equivalent drift of the center frequency of the corresponding resonators. The improvement in phase stability or frequency stability of f_1 and f_2, assuming equal relative resonator drifts, is of the order of the ratio of Q_3 to Q. This ratio is expressed by the first terms of (11) and (12), and is directly attributed to the presence of a quartz crystal in one of the feedback loops.

The relative shift of the frequency f_3 caused by the drift of the resonant circuits in A_1, A_2, and A_3 is given by (13). It is seen that the magnitude of this shift is approximately that of the crystal resonator drift plus that of the resonator drift in A_1 reduced by the ratios $Q/(Q+Q_3)$ and $1/(1+S)$ and that of the resonator drift in A_2 reduced by the ratios $Q/(Q+Q_3)$ and $S/(1+S)$.

The Variable Frequency Oscillator

The frequencies f_1 and f_2 in this circuit can be varied by tuning the resonators in A_1 and A_2 in such a manner that an increase of the resonant frequency f_{10} of the first one is coupled with a similar decrease of the resonant frequency f_{20} of the second one or vice versa, so that the sum f_1+f_2 remains constant. It is seen that tuning of f_1 and f_2 to any desired pair of values can be accomplished only by shifting the resonators in the opposite direction, whereas changing ambient conditions, aging, and power supply variations tend to shift the resonators in the same direction and thus have little influence upon the frequency.

A stable variable frequency oscillator can be built by employing resonators in A_1 and A_2 which have approximately equal drift characteristics and Q values over the required frequency range. This condition is not too difficult to fulfill in practice. Inequalities of the drift characteristics affect the stability of f_1 and f_2 to an extent which can be evaluated from (11) and (12), or determined graphically as indicated in Fig. 2. Inequalities in the Q's of the resonators in A_1 and A_2 have the same effect on frequency stability as unequal drifts of the resonators. It is observed that according to (11) and (12) the stability of f_1 and f_2 is affected to a lesser degree by the inequality of the resonators at small values or at high values of $S=f_{20}/f_{10}$ respectively.

Thus far the discussion of this system has been confined to a mode of oscillation at nonrationally related values of f_1, f_2, and f_3. When f_1 and f_2 are tuned close enough to a rational value, the phenomenon known as locking is observed[6] which causes f_1 and f_2 to establish a fixed rational relationship to f_3. In this mode of operation the stability of f_1 and f_2 will equal that of f_3. The extent to which the resonators can be shifted without influencing this fixed relationship is defined as the locking range, and is, in general, proportional to the loop gain and inversely proportional to the Q's of the resonant circuits. Within the locking ranges the frequency doesn't change and therefore these ranges represent small holes in the tuning spectrum of the oscillator. The tendency to lock may be observed, in general, at any rationally related values of f_1 and f_2, in the frequency range extending from zero to $f_3=f_1+f_2$. Only the simpler ratios of f_1 and f_2 however, have any real significance since locking is caused by the action of higher modulation components and harmonics, which decrease in strength as their order increases. In applications where continuity of tuning is desirable the tendency to lock can be considerably diminished by measures which reduce the harmonic content of the generated wave form, for instance by additional means of selection of desired frequency components in and out of the mixers and by use of a gain controlled limiter. Of course, the locking properties of this circuit have one decided advantage; they provide a convenient means of checking the calibration of the dial.

[6] R. Adler, "A study of locking phenomena in oscillators," Proc. IRE, vol. 34, pp. 351–357; June, 1946.

Experimental Work

The characteristics of the oscillator, outlined above, have been verified by constructing and testing an experimental model, using the circuit shown in Fig. 3. The mixers $M1$, $M2$, and $M3$ employ balanced ring modulators using matched germanium diodes. The tuned amplifiers A_1 and A_2 consist of double triodes and series resonant circuits, the latter employing precision variable air capacitors C_1 and C_2 tunable over the investigated frequency range, 40 kc to 50 kc, and 60 kc to 50 kc respectively, and ferrite inductors having a Q of approximately one hundred. The plate and grid circuits of the first and second triode sections are shunted to ground through relatively large capacitors. This greatly reduces the influence of interelectrode capacitance changes, caused by power supply variations, on the center frequency of the tuned amplifiers. The amplifier A_3 employs a double triode and a bridge circuit using a quartz crystal resonating at $f_3=100$ kc, as the tuning element. In the bridge circuit, the shunt capacitance of the quartz crystal is compensated to eliminate parallel resonance.

Oscillation is self-starting when grid current bias is employed. The frequency of oscillation can be changed by decreasing C_1 and increasing C_2 or vice versa. This could be realized with only one control, having the capacitors C_1 and C_2 connected back to back and the angle of rotation of these capacitors linearly related with frequency. The purity of the generated waveform depends upon the selectivity of the resonators, the degree of balance of the ring modulators and the amplitude of oscillation.

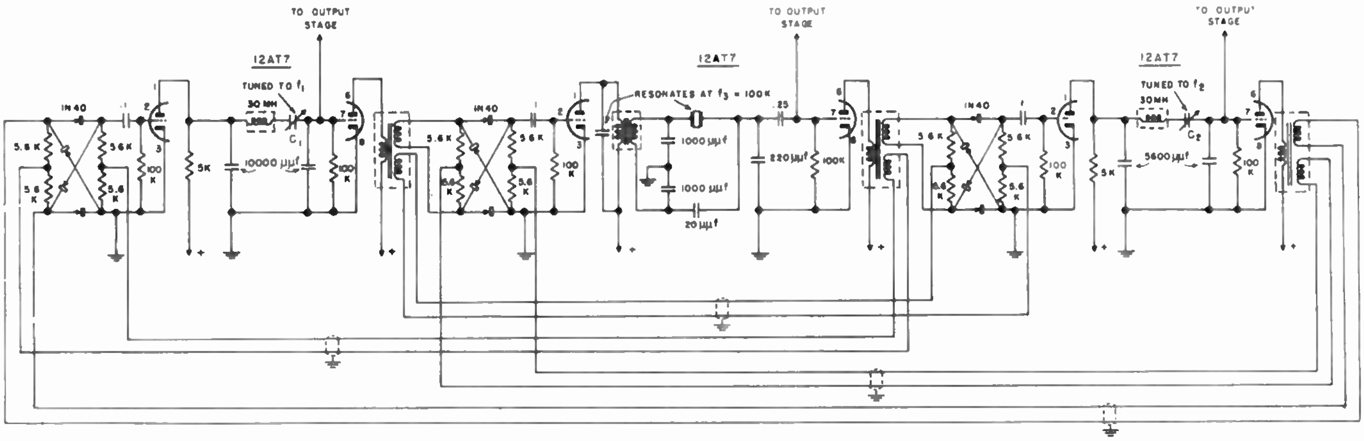

Fig. 3—Detailed circuit of the experimental model of the oscillator shown in block diagram in Fig. 1.

In Fig. 4, graphs (a) and (b) show typical curves of drift with temperature for a pair of nonrationally related frequencies f_1 and f_2 in the range from 40 kc to 50 kc, and from 60 kc to 50 kc respectively. In one set of tests the values for f_1, f_2, and f_3 at 25°C were 45,198.8 cps, 54,802.8 cps, and 100,001.6 cps respectively. The graphs (c) and (d) show the average drift curves for the center frequencies f_{10} and f_{20} of the corresponding resonators in A_1 and A_2. It is seen that an improvement in frequency stability by factors of about thirty and one hundred has been realized for ambient temperature variations of ± 15°C and ± 3°C respectively, about a mean of 25°C. The graph (e) shows the average drift curve for the frequency f_3 and the graph (f) that of the center frequency f_{30} of the quartz crystal resonator in A_3. The drift curves (a) and (b) show a region of zero slope followed by a region of reversed slope. A possible cause of this behavior, not readily apparent from the observed drift curves, which are subject to accuracy limitations of the measurements, may be attributed to the fact that if the resonators in A_1 and A_2 are matched closely, the first and second terms in (11) and (12) may be of comparable magnitude but of opposite sign to the third terms. Since the temperature coefficients and the quality factors of these resonators are functions of temperature, the presence of temperature regions with zero slope or reversed slope is likely. Also it is observed that (11) and (12) were derived on the assumption of equal quality factors Q which is fulfilled only to within a few per cent in the actual circuit. Unequal quality factors, as mentioned before, may have a similar effect as unequal drifts, and may be partially responsible for the reversal of the slope. A detailed investigation of these effects was not possible in the scope of this work.

High frequency stability with power supply variations was measured and is believed to be the result of improved phase stability of the system due to the presence of a quartz crystal in one of the feedback loops. Another contributing factor is formed by the relatively large capacitors which shunt to ground the plates and grids of some of the tubes. The average values observed

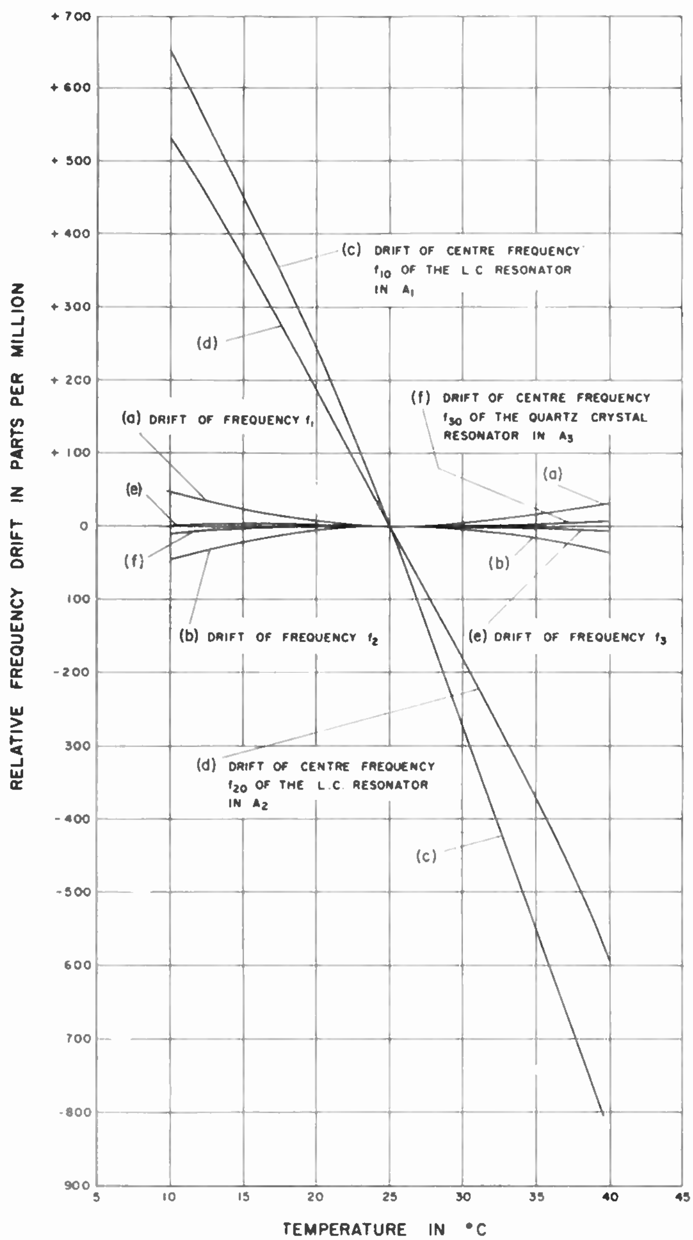

Fig. 4—Relative frequency drift curves with temperature, illustrating the reduced influence of resonator drifts in the tuned amplifiers A_1, A_2, and A_3, on the frequencies of oscillation f_1, f_2, and f_3.

were: ± 10 parts in 10^6 for ± 10 per cent plate voltage variations and ± 3 parts in 10^6 for ± 10 per cent filament voltage variations. In general, operation at lower plate voltage leads to higher frequency stability with filament voltage changes, but to poorer stability with plate voltage changes.

The frequency of a conventional LC oscillator drifts considerably in the period following switching on and this drift is caused mainly by temperature changes in the resonator. Its magnitude after the first 15 minutes of operation is often in the order of 200 in one million and as great as 2,000 parts in one million in the case of high power oscillators.[7] In the tested model, a remarkably small initial frequency drift of 6 parts in one million was measured. In general, smaller initial frequency drifts were observed at lower plate voltages.

The short and long term frequency stabilities of the experimental model were investigated. A regulated power supply was used and the resonators were enclosed in an insulated box, the inside temperature of which varied not more than 3°C over a period of several days. The following values were measured:

Ten minutes stability	3 parts in 10^7
One hour stability	6 parts in 10^7
24 hours stability	3 parts in 10^6
7 days stability	6 parts in 10^6

Frequency ratios f_1/f_3 and f_2/f_3 ($f_3=100$ kc) at which considerable locking was observed were few even though no means were used to reduce this tendency. These ratios are listed below:

f_1/f_3 (f_1 varied from 40 kc to 50 kc);
2/5, 9/22, 7/17, 5/12, 3/7, 7/16, 4/9, 5/11, 6/13, 7/15, 8/17, 9/19, 10/21, 1/2;
f_2/f_3 (f_2 varied from 60 kc to 50 kc);
3/5, 13/22, 10/17, 7/12, 4/7, 9/16, 5/9, 6/11, 7/13, 8/15, 9/17, 10/19, 11/21, 1/2.

The relative magnitude of the locking ranges at these frequency ratios varied from 0.002 per cent to 0.02 per cent, the simpler ratios having the larger values. Immediately outside these locking ranges, the short term frequency stability has been found satisfactory, although it may not be possible there, to maintain a desired frequency within close limits for a long period of time, since a change of operating conditions may cause this frequency to lock.

[7] H. A. Thomas, *op. cit.*, p. 89.

Discussion

The oscillatory system described in this paper departs considerably from prior art circuits. By virtue of the novel multiloop feedback and use of a quartz crystal, a significant reduction of the influence of the resonator drift on the frequency of oscillation is realized and the phase stability is greatly improved. Excellent long and short term stability and a remarkably small initial frequency drift with switching on are properties of this oscillator. The upper limit of the frequency range over which this circuit could be used is determined by the available quartz crystals and with present-day techniques this is 50–100 mc. The possibility of using this circuit at uhf and shf where a high-quality cavity or a molecular resonator replaces the quartz crystal and variable frequency cavities replace the LC tuned circuits is worth considering in the future when microwave techniques are more advanced. In cases where a continuously variable frequency spectrum is required, the locking property of the circuit may prove of value as a means of calibration at rationally related values of f_1, f_2, and f_3. However, in the immediate neighborhood of some of these locking regions, it may not be simple to maintain a desired frequency within close limits for a long period of time. In a suitably designed circuit the waveform of the oscillation will be satisfactory for most purposes.

It is believed that this circuit can be employed to advantage in certain high stability continuously variable frequency generators, certain multichannel communication systems and any place where a large number of very stable single frequencies are required, one at a time.

Acknowledgment

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IRE Standards on Electron Tubes: TR and ATR Tube Definitions, 1956*

(56 IRE 7. S3)

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Definitions Coordinator

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Arc Loss (Switching Tubes): The decrease in radio-frequency power measured in a matched termination when a *Fired Tube*, mounted in a series or shunt junction with the waveguide, is inserted between a matched generator and the termination. In the case of a *Pre-TR or TR Tube*, a matched output termination is also required for the tube.

ATR (Anti-Transmit Receive) Tube: A gas-filled radio-frequency switching tube used to isolate the transmitter during the interval for pulse reception.

Attenuator Tube: A gas-filled radio-frequency switching tube in which a gas discharge, initiated and regulated independently of radio-frequency power, is used to control this power by reflection or absorption.

Band-Pass Tube (TR and Pre-TR Tubes): See *Broad-Band Tube* (*TR* and *Pre-TR Tubes*).

Broad-Band Tube (TR and Pre-TR Tubes): A gas-filled fixed-tuned tube incorporating a band-pass filter of geometry suitable for radio-frequency switching.

Cell Type Tube (TR, ATR, and Pre-TR Tubes): A gas-filled radio-frequency switching tube which operates in an external resonant circuit. A tuning mechanism may be incorporated in either the external resonant circuit or the tube.

Control Electrode Discharge Recovery Time (Attenuator Tubes): The time required for the control electrode discharge to deionize to a level such that a specified fraction of the *Critical High Power Level* is required to ionize the tube.

Critical High Power Level (Attenuator Tubes): The radio-frequency power level at which ionization is produced in the absence of a control electrode discharge.

Direct Coupled Attenuation (TR, Pre-TR, and Attenuator Tubes): The insertion loss measured with the *Resonant Gaps*, or their functional equivalent, short-circuited.

Equivalent Conductance (ATR Tubes): The normalized conductance of the tube in its *Mount* measured at its resonance frequency.

Note: Normalization is with respect to the characteristic impedance of the transmission line at its junction with the tube mount.

Fired Tube (TR, ATR, and Pre-TR Tubes): The condition of the tube during which a radio-frequency glow discharge exists at either the *Resonant Gap*, *Resonant Window* or both.

Flat Leakage Power (TR and Pre-TR Tubes): The peak radio-frequency power transmitted through the tube after the establishment of the steady-state radio-frequency discharge.

Harmonic Leakage Power (TR and Pre-TR Tubes): The total radio-frequency power transmitted through the *Fired Tube* in its *Mount* at frequencies other than the fundamental frequencies generated by the transmitter.

High Level Firing Time (Switching Tubes): The time required to establish a radio-frequency discharge in the tube after the application of radio-frequency power.

High-Level Radio-Frequency Signal (TR, ATR, and Pre-TR Tubes): A radio-frequency signal of sufficient power to cause the tube to become fired.

High Level VSWR (Switching Tubes): The vswr due to a *Fired Tube* in its *Mount* located between a generator and matched termination in the waveguide.

Ignitor Current Temperature Drift (TR, Pre-TR, and Attenuator Tubes): The variation in *Ignitor Electrode* current caused by a change in ambient temperature of the tube.

Ignitor Discharge (Switching Tubes): A dc glow discharge, between the *Ignitor Electrode* and a suitably located electrode, used to facilitate radio-frequency ionization.

Ignitor Electrode (Switching Tubes): An electrode used to initiate and sustain the *Ignitor Discharge*.

Ignitor Firing Time (Switching Tubes): The time interval between the application of a dc voltage to the *Ignitor Electrode* and the establishment of the *Ignitor Discharge*.

Ignitor Interaction (TR, Pre-TR, and Attenuator Tubes): The difference between the *Insertion Loss* measured at a specified ignitor current and that measured at zero ignitor current.

Ignitor Leakage Resistance (Switching Tubes): The insulation resistance, measured in the absence of an *Ignitor Discharge*, between the *Ignitor Electrode* terminal and the adjacent radio-frequency electrode.

Ignitor Oscillation (TR, Pre-TR, and Attenuator Tubes): Relaxation oscillations in the ignitor circuit.

Note: If present these oscillations may limit the tube protection characteristics.

Ignitor Voltage Drop (Switching Tubes): The dc voltage between the cathode and the anode of the *Ignitor Discharge* at a specified ignitor current.

Insertion Loss (TR, Pre-TR, and Attenuator Tubes): The decrease in power measured in a matched termination when the *Unfired Tube* at a specified ignitor current is inserted in the waveguide between a matched generator and the termination.

Keep Alive: Deprecated, see *Ignitor Electrode*.

Leakage Power (TR and Pre-TR Tubes): The radio-frequency power transmitted through a *Fired Tube*.

Loaded Q (Switching Tubes): The *Unloaded Q* of the tube modified by the coupled impedances.

Note: As here used, Q is equal to 2π times the energy stored at the resonance frequency divided by the energy dissipated per cycle in the tube or, for *Cell Type Tubes*, in the tube and its external resonant circuit.

Low-Level Radio-Frequency Signal (TR, ATR and Pre-TR Tubes): A radio-frequency signal with insufficient power to cause the tube to become fired.

Minimum Firing Power (Switching Tubes): The minimum radio-frequency power required to initiate a radio-frequency discharge in the tube at a specified ignitor current.

Mode Purity (ATR Tubes): The extent to which the tube in its *Mount* is free from undesirable mode conversion.

Mount (Switching Tubes): The flange or other means by which the tube or tube and cavity are connected to a waveguide.

Phase Recovery Time (TR and Pre-TR Tubes): The time required for a *Fired Tube* to deionize to such a level that a specified phase shift is produced in the *Low-Level Radio-Frequency* signal transmitted through the tube.

Phase Tuned Tube (TR Tubes): A fixed tuned *Broad Band TR Tube*, wherein the phase angle through and the reflection introduced by the tube are controlled within limits.

Position of the Effective Short (Switching Tubes): The distance between a specified reference plane and the apparent short circuit of the *Fired Tube* in its *Mount*.

Pre-TR Tube: A gas-filled radio-frequency switching tube used to protect the *TR Tube* from excessively high power and the receiver from frequencies other than the fundamental.

Recovery Time (ATR Tubes): The time required for a *Fired Tube* to deionize to such a level that the normalized conductance and susceptance of the tube in its *Mount* are within specified ranges.

Note: Normalization is with respect to the characteristic admittance of the transmission line at its junction with the tube mount.

Recovery Time (TR and Pre-TR Tubes): The time required for a *Fired Tube* to deionize to such a level that the attenuation of a *Low-Level Radio-Frequency Signal* transmitted through the tube is decreased to a specified value.

Resonant Gap (TR Tubes): The small region in a resonant structure interior to the tube, where the electric field is concentrated.

Resonant Window (Switching Tubes): A resonant iris, sealed with a suitable dielectric material, and constituting a portion of the vacuum envelope of the tube.

Spike Leakage Energy (TR and Pre-TR Tubes): The radio-frequency energy per pulse transmitted through the tube before and during the establishment of the steady-state radio-frequency discharge.

TR (Transmit Receive) Tube: A gas-filled radio-frequency switching tube used to protect the receiver in pulsed radio-frequency systems.

Tuning Range (Switching Tubes): The frequency range over which the resonance frequency of the tube may be adjusted by the mechanical means provided on the tube or associated cavity.

Tuning Susceptance (ATR Tubes): The normalized susceptance of the tube in its *Mount* due to the deviation of its resonance frequency from the desired resonance frequency.

Note: Normalization is with respect to the characteristic admittance of the transmission line at its junction with the tube mount.

Unfired Tube (TR, ATR, and Pre-TR Tubes): The condition of the tube during which there is no radio-frequency glow discharge at either the *Resonant Gap* or *Resonant Window*.

Unloaded (Intrinsic) Q (Switching Tubes): The Q of a tube unloaded by either the generator or the termination.

Note: As here used, Q is equal to 2π times the energy stored at the resonance frequency divided by the energy dissipated per cycle in the tube or, for *Cell Type Tubes*, in the tube and its external resonant circuit.

IRE Standards on Methods of Measurement of the Conducted Interference Output of Broadcast and Television Receivers in the Range of 300 KC to 25 MC, 1956*

(56 IRE 27 S1)

This Standard is a supplement to 54 IRE 17 S1, Standards on Receivers: Methods of Measurement of Interference Output of Television Receivers in the Range of 300 to 10,000 kc, 1954.

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1. Introduction

1.1 This supplement is issued to permit the extension of conducted interference measurements defined in Standard 54 IRE 17. S1 from a range of 300 to 10,000 kc to a range of 300 kc to 25 mc and replaces sections 3.2.2 and 3.2.5 of the standard. The extension in frequency coverage involves revising the characteristics of the line impedance network.

2. Methods of Measurement

2.1 General

Except as indicated, the method of measuring the conducted interference output of a receiver in the 300 kc to 25 mc range is identical to that specified by Standard 54 IRE 17. S1. The primary difference involves a change in the line impedance network.

2.2 Line Impedance Network

2.2.1 The line impedance network is schematically illustrated in Fig. 1. The purpose of the one ohm (non-reactive) resistor is to minimize any possible resonance effects of the series circuit of the 5 microhenry inductor and the 1.0 microfarad capacitor. The purpose of the 1000 ohm resistor is to limit the line voltage that may appear on the bottom end of the 0.1 microfarad capacitor.

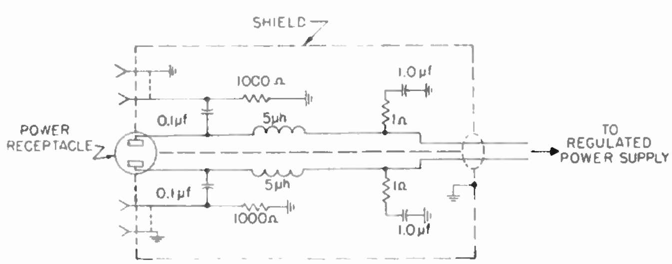


Fig. 1—Power line impedance schematic.

2.2.2 The impedances of the line network measured from each side of the receiver receptacle to chassis must conform within ±5 per cent to the characteristic shown in Fig. 2. (For this requirement the power plug is open-circuited.)

2.2.3 A suitable method of measuring these impedances is shown in Fig. 3. This measurement technique is a substitution method. The reference resistor is chosen so that the voltage drop across this resistor is equal to the voltage across the line-impedance network at each frequency of measurement. The value of the resistor is then taken as the absolute value of the impedance. Since the impedance of the line network is considerably less than that of the 470 ohm resistor, the generator impedance has a negligible effect on the measurements. The accuracy of the voltmeter is unimportant since it is used to hold the voltage constant when the switch is changed. It is important to keep the lead lengths as short as possible.

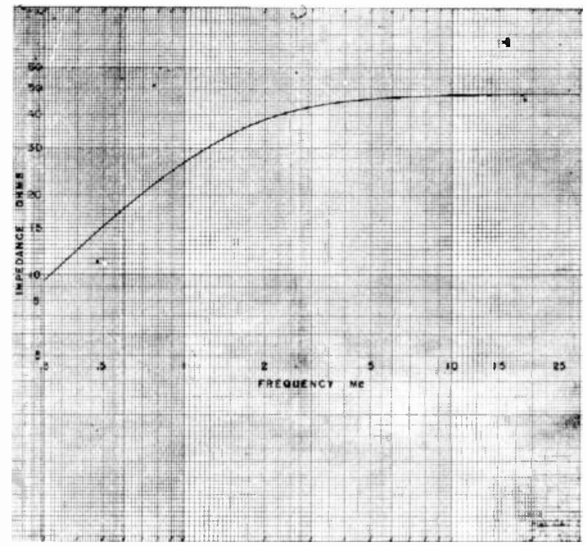


Fig. 2—Impedance characteristics of line measured from either side of the receiver receptable to chassis.

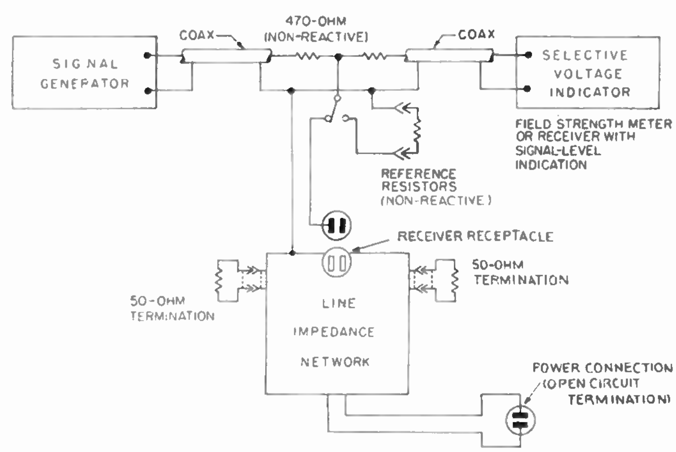


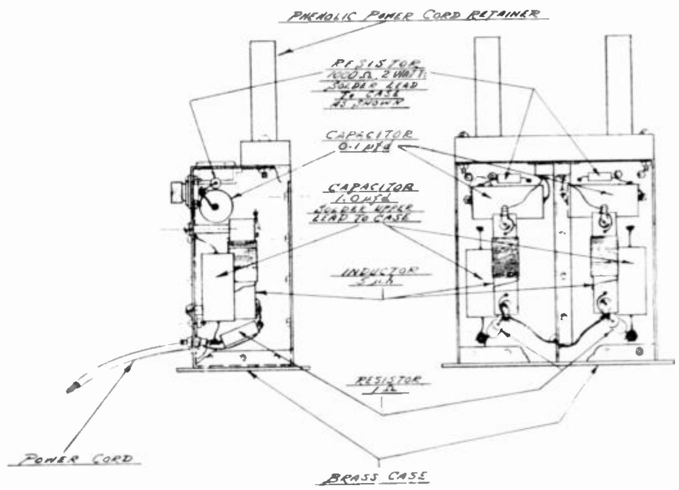
Fig. 3—Method of measurement of impedance characteristic shown in Fig. 2.

2.2.4 To minimize variations which might occur among different line impedance networks and to permit more uniformity in test facilities, detailed construction drawings of a suitable network, of which assembly drawings are shown in Fig. 4(a) and (b), have been prepared.[1] Any network constructed according to these drawings should be tested in order to insure that it meets the requirement of paragraph 2.2.2.

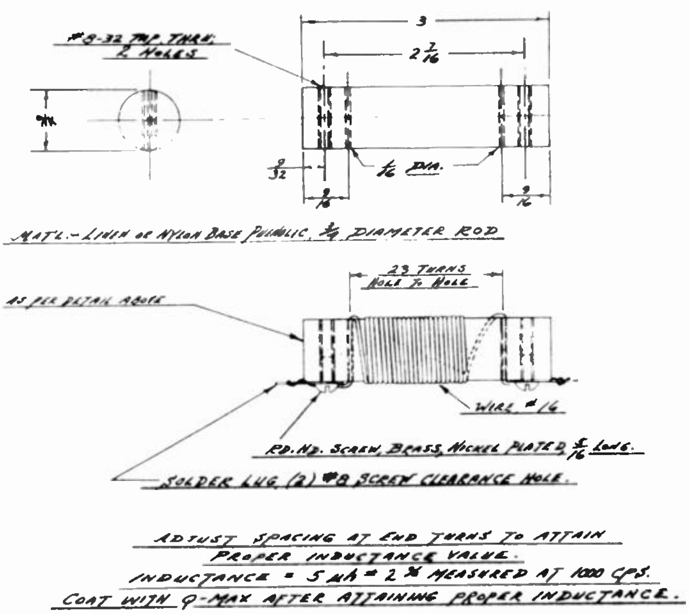
2.3 Methods of Installation and Operation

2.3.1 The line impedance network is interposed between the source of power and the line input to the receiver. The voltage at the receiver receptacle shall be maintained at 117 volts ±2 per cent unless otherwise specified. Voltage regulators with high harmonic content should be avoided. The line from the stabilizing

[1] These drawings may be purchased from the Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y. at a cost of $2.00 per copy.



(a)



(b)

Fig. 4—(a) Line impedance assembly. (b) Inductor 5 μh.

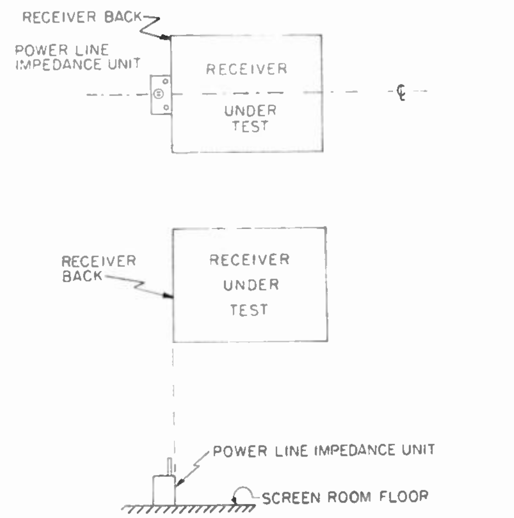


Fig. 5—Placement of receiver with reference to power line impedance unit.

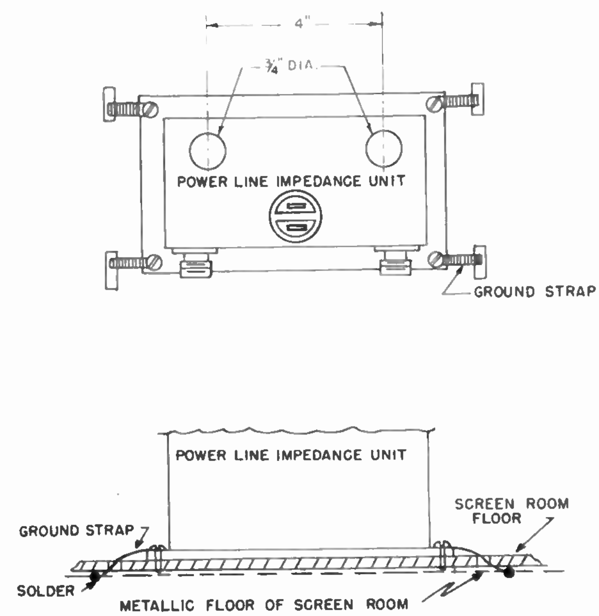


Fig. 6—Suggested method for grounding the line impedance unit to the screen room.

network to the power source should be kept close to the walls or floor of the shielded enclosure. The line impedance network shall be located directly below the back of the cabinet of the receiver under test as shown by the lower view of Fig. 5. The center line of the line impedance unit shall be coincident with the center line of the receiver back, as shown in the upper view of Fig. 5.

2.3.2 The line impedance unit shall be on the floor of the shielded enclosure and shall be connected to the metallic floor by means of four solid copper straps as shown in Fig. 6. The width to length ratio of each strap shall be at least 1 to 5 and the thickness of the strap shall be at least 0.025 inches. In the unit shown in Fig. 2, four holes have been provided for this purpose.

2.3.3 All portions of the receiver under test shall be at least 30 inches from the walls of the shielded enclosure. Floor model receivers shall be placed on a nonmetallic platform 18 inches above the metallic floor of the shielded enclosure and table models placed on a nonmetallic platform 30 inches above the floor.

2.3.4 A 50-ohm resistive load shall be connected to each of the two coaxial connectors of the line impedance network at all times. The voltages developed across these loads represent the conducted interference output of the receiver. A 50-ohm nonreactive resistor or a 50-ohm input impedance field-strength meter or any combination of field-strength meter and external resistor to equal 50 ohms can be used as the resistive load. The field strength meter shall be used with the switch in the position normally designated as *Field Intensity*, *CW*, or *Carrier*.

2.4 Dress of the Receiver Power Line Cord

2.4.1 The cord shall be dressed to this unit through the shortest possible path. The excess cord length shall be taken up by wrapping the cord in a figure eight pattern around the two posts provided on the top of the unit. The receiver power line cord shall be plugged into the receptacle provided in the power line impedance unit. This is shown in Fig. 7.

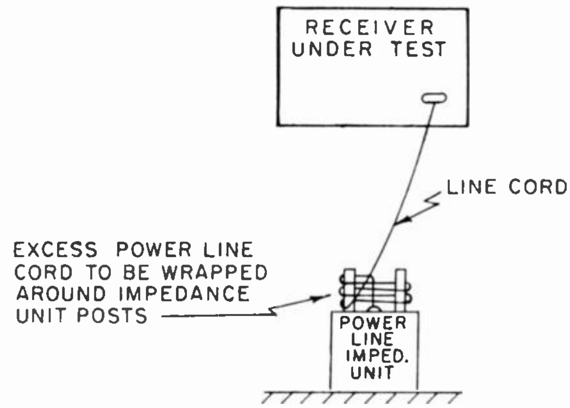


Fig. 7—Method of dressing the receiver power line cord.

Some Limiting Cases of Radar Sea Clutter Noise*

ALLEN H. SCHOOLEY†, FELLOW, IRE

Summary—**Radar return from the rough surface of the wind disturbed sea is a source of noise that affects the over-all signal-to-noise ratio of radars operating over water. This paper presents the results of an analysis showing some limiting values of the effective radar scattering area per unit area of the sea surface (σ^0) vs radar depression angle (Θ) for perfectly smooth and perfectly rough surfaces. Available experimental data are shown to be consistent with the theoretical limits.**

Introduction

THE VARIOUS FORMS of the general radar equation used by radar designers contain a factor which is defined as the effective radar cross section area of the target, usually symbolized by σ. Whether a given target can actually be seen by a radar under specified conditions depends upon the background noise that tends to mask the target echo. For long ranges, receiver noise may be the limiting factor. For an airborne or shipborne radar looking for a target on or near the wind-swept surface of the sea, another noise source may be the limiting factor. This noise is called radar sea clutter and is caused by the reflection and back scattering of transmitted energy from the many randomly oriented facets of the rough sea surface. It is the purpose of this paper to present what are believed to be some limiting values of sea clutter.

If the effective scattering area of the rough sea surface happens to be appreciably greater than the effective cross section area of a wanted target, the wanted target will be obscured. In an area-extensive reflecting surface, such as the wind disturbed sea, the radar cross section area can best be defined as the effective radar cross section per unit area of the mean sea surface,[1] and is symbolized by σ^0.

For all but grazing incidence, σ^0 is primarily a function of three parameters,

$$\sigma^0 = f(\Theta, \lambda, w, \cdots) \tag{1}$$

where

Θ = vertical angle which the axis of the radar beam makes with the horizontal

λ = radar wavelength

w = local wind velocity.

The dots in (1) indicate that other less important factors (not always negligible) are omitted, such as radar polarization angle, wind direction, wind fetch, etc. Sea-clutter research is centered around trying to understand the nature of the functional relationship expressed by (1).

There are two limiting special cases of sea roughness: perfectly smooth and perfectly rough. Even though these conditions are not met in practice they are interesting, nevertheless, because they appear to set the

* Original manuscript received by the IRE, November 21, 1955; revised manuscript received May 31, 1956.
† Naval Research Lab., Washington, D.C. (During 1956 attached to the U. S. Naval Mission to Brazil.)

[1] D. E. Kerr, "Propagation of Short Radio Waves," M.I.T. Rad. Lab. Ser., the McGraw-Hill Book Co., Inc., New York, N.Y., vol. 13, ch. 6; 1951.

limits of the effect of actual sea-clutter noise. Each condition is discussed in the following two sections.

Perfectly Smooth Surface

Fig. 1 shows the approximate σ^0 function derived in Appendix I for a perfectly smooth surface. The ordinate on the left is σ^0 on a logarithmic scale. The ordinate on the right is σ^0 in decibels referred to unity. The abscissa is the radar depression angle in degrees. A depression angle of zero degrees corresponds to a radar antenna looking horizontally in one direction; 90° represents the antenna looking straight down, and 180° represents the antenna looking horizontally in the opposite direction from zero degrees. The peaks of the various curves are given by $4/\Phi^2$ when the radar antenna beam width Φ is expressed in radians. In Fig. 1 both Φ and Θ are in degrees. The shapes of the curves are determined by the radar antenna beam pattern. In the case of Fig. 1, the antenna pattern is assumed to have a Gaussian shape. This assumption is valid for most conventional antennas over a 20 db range. Actual antennas have side lobes that would put "skirts" on the curves of Fig. 1 in the region below 20 db.

It is interesting to note that σ^0 can be considerably greater than unity radar cross section area per unit area of the reflecting surface under the conditions of Fig. 1. This has been borne out in experimental measurements over smooth water.[1]

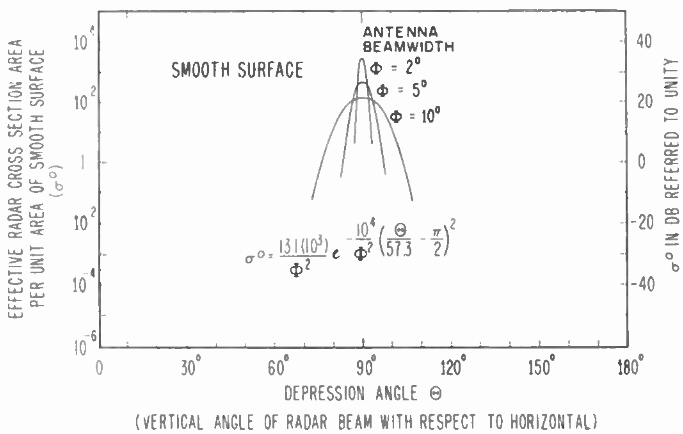


Fig. 1—Approximate σ^0 function for a perfectly smooth surface.

Perfectly Rough Surface

The σ^0 function for a perfectly rough surface depends upon the definition of roughness that is used. Fig. 2 shows two such functions. The approximate $\sigma^0=2 \sin \Theta$ relationship is obtained if one assumes that a perfectly rough or diffuse surface, illuminated by a bundle of parallel rays of fixed cross section and given total power, will reradiate in the direction of the source of the rays a power per unit solid angle independent of the direction of the incident rays. This relationship is derived in Appendix II.

Ament[2] has shown that the approximate $\sigma^0=4 \sin^2 \Theta$ relationship results if an illuminated unit area of lossless diffuse reflecting surface scatters, per unit solid angle, a power proportional to the cosine of the acute angle between the direction of the scattering and the normal to the surface at the unit area. According to Ament this leads to the prediction that when a sphere with an ideally rough surface is illuminated solely by a searchlight held near the axis of the eyes, the outer edge should appear darker than the central region. Unfortunately, there appears to be one known exception to this viewpoint. The full moon has been measured to be of nearly uniform average brightness across its face with a slight amount of unexplained edge brightening.[3] Thus we have interesting contradictions that are not yet completely understood.

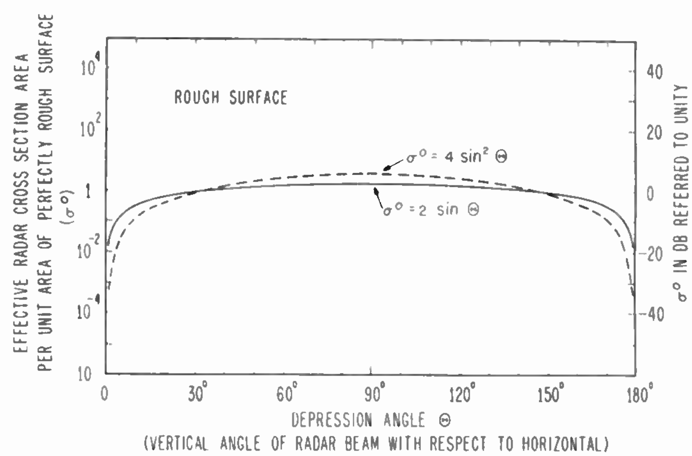


Fig. 2—Approximate σ^0 function for a rough surface.

Fig. 3 shows a composite of Figs. 1 and 2 together with additional data. One curve from Fig. 2 is reproduced in order to simplify Fig. 3. The coordinates of Fig. 3 are the same as the previous figures except that Θ is now measured with respect to the horizontal upwind direction. Thus, zero degrees represents a radar antenna looking upwind and 180° represents the antenna looking with the wind.

Actual Surfaces

One would expect that sea clutter from an actual sea surface would fall below the appropriate smooth surface curve (Fig. 3) in the region around 90° and below the rough surface curve for other values of Θ, because such a surface would unlikely be neither perfectly smooth nor perfectly rough. There is some experimental evidence to substantiate this, and it is shown by the experimental points and empirical equations in Fig. 3.

[2] W. S. Ament, "Forward- and back-scattering from certain rough surfaces," (accepted by the IRE).

[3] A. V. Markov, "Distribution of brightness over the lunar disk at full moon," *Astronomicheskii Zhurnal*, USSR, vol. 25, pp. 172–179; 1948.

The points in the regions between zero to 20° and 160° to 180° are from unpublished data using horizontal polarization and gathered by the Naval Research Laboratory near Bermuda in 1950 by means of a PBM equipped as a flying laboratory. These data indicate that σ^0 is very approximately proportional to the local wind velocity (w) squared and proportional to the reciprocal of the wavelength (λ) over the range of conditions measured. More complete experimental data and discussion of the grazing incidence case are contained in Katzin[4] and MacDonald[5] which were not available at the time this paper was originally prepared. The numerical coefficients of the empirical equations of Fig. 3 show the well established fact that σ^0 is appreciably greater when looking upwind than when looking with the wind. The symbol σ_u^0 is used for the upwind case and σ_w^0 for the with the wind case.

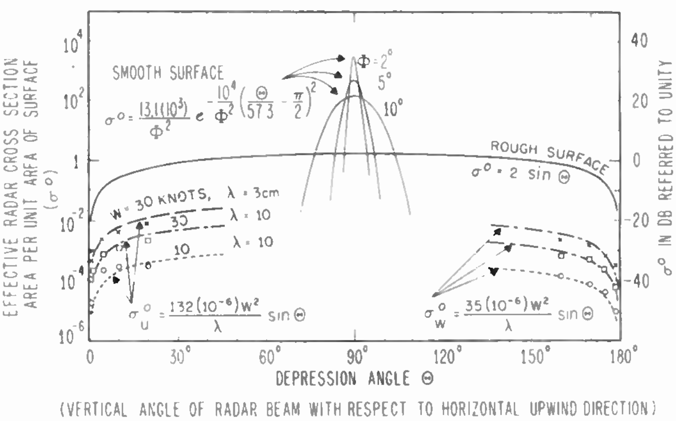


Fig. 3—Theoretical and measured σ^0 functions.

Conclusion

The effective radar cross section area per unit area of a perfectly smooth sea surface is primarily a function of radar antenna beamwidth and radar beam depression angle. The approximate quantitative relationships are given in Fig. 1.

The effective radar cross section area per unit area of a perfectly rough surface is primarily a function of the radar beam depression angle. Two possible approximate quantitative functions relating these variables are shown in Fig. 2.

Available measurements on the actual wind disturbed sea surface are consistent with the theoretical limits that have been derived for the perfectly smooth and perfectly rough cases. Fig. 3 shows the results of some experimental measurements together with the theoretical limits.

[4] Martin Katzin, "Recent Developments in the Theory of Sea Clutter," 1956 IRE Convention Record.

[5] F. C. MacDonald, "Correlation of Radar Sea Clutter on Vertical and Horizontal Polarizations with Wave Height and Slope," 1956 IRE National Convention Record.

Appendix I

Infinite Perfectly Smooth Surface

Fig. 4 shows a radar with an effective beamwidth Φ directed straight down on a smooth perfectly reflecting water surface. It is assumed that the radar beam is conical and illuminates a circular area of the water surface with radius r. At any range R_r the energy is assumed to be uniform within the angle subtended by Φ and zero outside the beam. For this condition the following general free space radar formula applies

$$P_r/P_t = G^2\lambda^2\sigma_s/(4\pi)^3 R_r^4 \tag{2}$$

where

- P_r is received power
- P_t is transmitted power
- G is antenna gain
- λ is wavelength
- σ_s is the effective reflecting area of the smooth surface
- R_r is the radar range or distance to the surface.

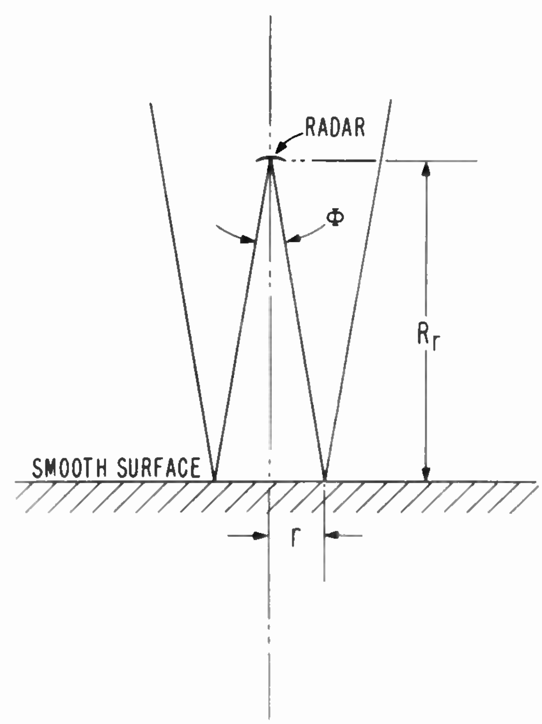


Fig. 4—Radar directed towards a large smooth surface.

Fig. 5 is equivalent to Fig. 4 where the smooth surface is considered to produce an image of the radar antenna. Thus the transmitter is considered to be above the surface and the receiver to be the image below the surface. The general one way free space transmission formula is

$$P_r/P_t = G_t G_r \lambda^2/(4\pi R)^2 \tag{3}$$

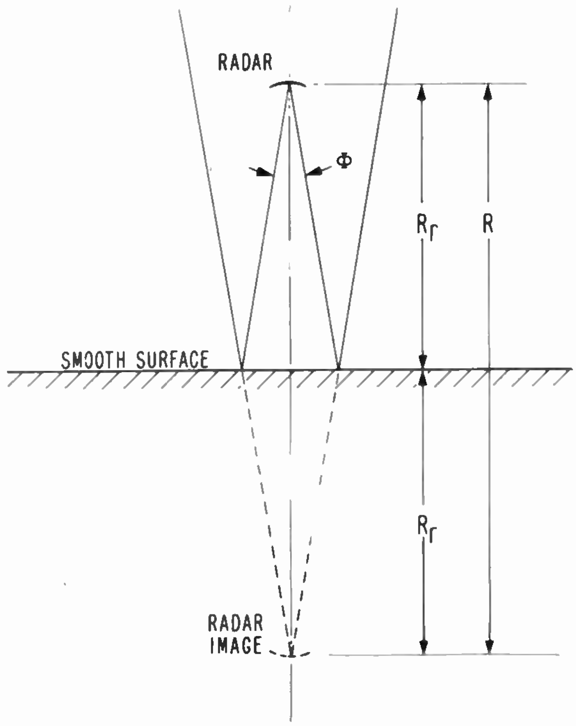


Fig. 5—Radar image beneath a large smooth surface.

where

G_t is transmitter antenna gain
G_r is receiver antenna gain
$G_t = G_r = G$ and $R = 2R_r$
All other symbols are as defined for (2).

Equating (2) and (3), it is found

$$\sigma_s = \pi R_r^2. \tag{4}$$

Now σ_s^0 is defined as the effective reflecting area of the smooth surface per unit actual area A that is illuminated. Thus

$$\sigma_s^0 = \sigma_s / A. \tag{5}$$

From Fig. 4 $r = R_r \tan(\Phi/2) \cong R_r \Phi/2$ where Φ is small and expressed in radians.

Also from Fig. 4

$$A = \pi r^2 = \pi R_r^2 \Phi^2 / 4; \tag{6}$$

combining (4), (5) and (6)

$$\sigma_s^0 = 4/\Phi^2. \tag{7}$$

Eq. (7) gives the effective target area per unit actual illuminated area of the flat sea surface for various effective radar beamwidths measured in radians. The values of σ_s^0, as given by (7) are uniform over the beamwidth angle Φ and zero outside this region due to the assumptions that have been made. In the case of an actual radar the situation can better be approximated by assuming that Φ is the angle subtended by the half power points of the actual radar antenna pattern. It has been found that a pencil beam type of antenna pattern can be approximated for a range of about 20 db down from the maximum by a Gaussian distribution type of function. Using this function in conjunction with (7) the following approximate function can be derived

$$\sigma^0 = \left[13.1(10^3)/\Phi^2\right] e^{-[10^4/\Phi^2][\Theta/57.3 - \pi/2]^2} \tag{8}$$

where

σ^0 is effective radar cross-section area per unit area of smooth surface
Φ is antenna half power beamwidth in degrees
e is the base of natural logarithms
Θ is the radar depression angle in degrees

Equation (8) is shown plotted for three values of Φ in Fig. 1.

Appendix II

Infinite Perfectly Rough Surface

The term perfectly rough surface is somewhat ambiguous because roughness is a matter of definition. For the purpose of this discussion, a perfectly rough or diffuse surface, when illuminated by a bundle of parallel rays of fixed cross section and given total power, will be assumed to reradiate in the direction of the source of the rays a power per unit solid angle independent of the direction of the incident rays. This is the type of reflection that is approximated by the full moon at optical frequencies.[3]

The effective radar scattering cross section area of a given target may be defined in terms of the cross section of a perfectly conducting sphere that returns an equivalent signal to the radar when placed at the target position. Expressed symbolically

$$\sigma_0 = \pi r_0^2 \tag{9}$$

where r_0 is the radius of a perfectly conducting sphere that is large compared to the radar wavelength. Since a perfectly rough or diffuse infinite surface may be assumed to reflect a pencil beam of incident radiation uniformly in 2π steradians, no matter what the angle of incidence may be, it is evident that the effective scattering cross section area of the diffuse surface is

$$\sigma_d = 2\pi r^2 \tag{10}$$

where r is the radius of a unit pencil of radiation incident upon the diffuse surface.

The area of the rough or diffuse surface target may be defined as the area that the unit pencil of radiation makes on a plane through the average value of the rough surface. If the pencil is directed straight down where

$\Theta = 90°$ the area will be πr^2. For other values of Θ the elliptical illuminated area A will be

$$A = \pi r b \tag{11}$$

where r is the minor semiaxis and is always equal to the unit radius of the illuminating pencil and b is the major semiaxis as shown in Fig. 6.

From Fig. 6 it is evident that $b = r/\sin \Theta$ and combining this with (11)

$$A = \pi r^2/\sin \Theta. \tag{12}$$

By definition the effective scattering area per unit actual target area is $\sigma^0 = \sigma/A$ which for the diffuse target is σ_d/A. Thus combining with (10) and (12)

$$\sigma_d{}^0 = 2 \sin \Theta \tag{13}$$

Eq. (13) is plotted as the solid line in Fig. 2.

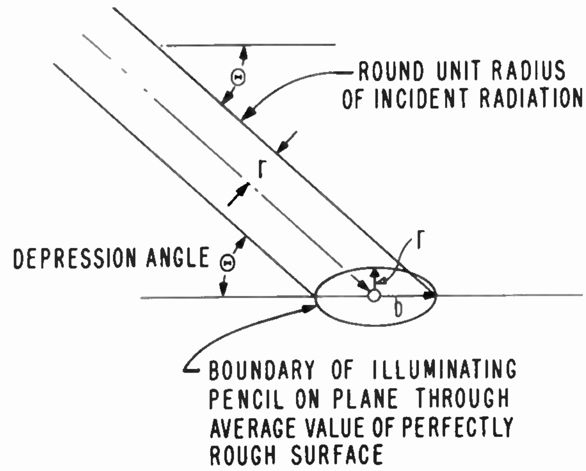


Fig. 6—Geometry of unit pencil of radiation striking a rough surface.

Correction

The following corrections to "Transistor Amplifiers for Use in a Digital Computer," by Q. W. Simkins and J. H. Vogelsong, which appeared on pages 43–55 of the January, 1956, issue of PROCEEDINGS OF THE IRE has been brought to the attention of the authors by Meir Drubin.

Specifically, these corrections apply to the coefficients, b_1 and b_0, on page 53. These should read:

$$b_1 = g_c[n - n^2(1 + r_e g_b)] + n^2 \omega_c c[a_0 - 1 - r_e(g_b + g_c)]$$

$$b_0 = \omega_c\{n a_0 g_b + g_c[n - n^2(1 + r_e g_b)]\}$$

rather than

$$b_1 = g_b + g_c[1 + n r_e g_b - n^2(1 + r_e g_b)]$$
$$+ n^2 \omega_c c[a_0 - 1 - r_e(g_b + g_c)]$$

$$b_0 = \omega_c\{(n a_0 + 1) g_b + g_c[1 - n^2(1 + r_c g_b)]\}.$$

The computed rise times for Region I as shown on page 55 are increased by less than 10 per cent as a result of these changes.

Correspondence

Electron Beam Noisiness and Equivalent Thermal Temperature for High-Field Emission from a Low-Temperature Cathode*

Dyke[1] has demonstrated that substantial current can be drawn from an unheated cathode in the presence of high electric fields. This raises a question as to the utility of field emission at low cathode temperatures in connection with low-noise microwave amplifiers. Since cathode temperature is so important in limiting the minimum theoretical noise figure of such an amplifier when it uses thermal emission,[2-4] some workers have suggested the possibility of obtaining low-noise performance with field emission from low-temperature cathodes. The following is an evaluation of the "noisiness" of field-emitted beams from cold cathodes, and the derivation of an expression for the equivalent thermal-cathode temperature.

Pierce,[5] Bloom, and Peter[2] have shown that an electron beam carrying noise fluctuations which stem from two uncorrelated noise sources is characterized by a noise standing wave with maximum and minimum currents I_{max} and I_{min} such that

$$W|I_{max}||I_{min}| = \text{constant} \tag{1}$$

where W is the characteristic beam impedance. The constant is independent of any lossless region through which the beam might pass. The above expression has been termed the beam "noisiness"[6] and is important in determining minimum noise figure.[2-4] For thermal emission, in which the electrons leave a hot cathode with a Maxwellian velocity distribution, (1) becomes[2,5]

$$W|I_{max}||I_{min}| = 2\alpha k T_c B \tag{2}$$

where T_c is the cathode temperature. Assuming pure shot noise at the potential minimum, α is approximately unity.

Since the constant in (1) is independent of the region through which the beam passes, the constant can be evaluated readily for any process of emission by assuming that the anode is placed very close to the cathode and given a high potential. The emitted electrons then are immediately accelerated by the strong electrostatic fields which prevent the accumulation of space charge in the cathode-anode region. For the above temperature-limited operation, it is valid to consider the entire cathode-anode region as constituting a velocity jump. If one of the independent noise sources is due to current fluctuation at the cathode and the other is due to an uncorrelated velocity fluctuation there, (1) can be written

$$W|I_{max}||I_{min}| = \left[\overline{I^2}\,\frac{\overline{v_0^2}}{\eta^2}\,\frac{2eB}{I_0}\,\overline{\delta v^2}\right]^{1/2} \tag{3}$$

where

$\overline{I^2}$ is the mean-square current fluctuation at the anode,

$\overline{v_0^2}$ is the mean-square velocity of the electrons passing the anode plane, and

$\overline{\delta v^2}$ is the mean-square deviation in velocity, or the variance, of the electrons passing the anode plane.

When electrons emerge from a cathode in completely random fashion, the current fluctuation there is called shot noise and is given by

$$\overline{I^2} = 2eI_0B. \tag{4}$$

Complete randomness characterizes both thermal emission and high-field emission. Under the cathode-anode conditions assumed, (4) also gives the current fluctuation at the anode for both types of emission. Evaluation of $\overline{v_0^2}$ and $\overline{\delta v^2}$ for both cases requires a knowledge of the respective energy distribution functions.

The energy distribution for thermal emission is very nearly Maxwellian and is given by

$$P_t(w) = A_t \exp\left(-\frac{w}{kT_c}\right) \quad \text{for } w > eV_0$$
$$= 0 \qquad \text{for } w < eV_0 \tag{5}$$

where

$P_t(w)dw$ is the number of electrons passing the anode plane per unit time having kinetic energies between w and $w+dw$,

A_t is a constant,

w is the electron kinetic energy (*i.e.*, $mv^2/2$) normal to the cathode and anode planes, and

V_0 is the anode voltage.

Assuming $eV_0/kT_c \gg 1$ we obtain

$$\overline{v_0^2} = \frac{\int_0^\infty v^2 P_t(w)dw}{\int_0^\infty P_t(w)dw} = 2\eta V_0 \tag{6}$$

and

$$\overline{\delta v^2} = \overline{v_0^2} - \left[\frac{\int_0^\infty vP_t(w)dw}{\int_0^\infty P_t(w)dw}\right]^2 = \left(\frac{kT_c}{2m}\right)\left(\frac{kT_c}{eV_0}\right). \tag{7}$$

Using (4), (6), and (7) in (3) gives

$$W|I_{max}||I_{min}| = 2kT_cB \tag{8}$$

which agrees with (2).

Conceptually and mathematically, field emission is more complicated than thermal emission. The wave-mechanical theory as enunciated by Sommerfeld, Bethe, Nordheim, and Fowler[7-9] assumes the electrons within the cathode metal to be incident upon a potential barrier at the metal surface. With no accelerating field the barrier is infinitely thick and the electrons cannot escape by penetration through the barrier. Unless energy is given the electrons by heating the cathode, the electrons cannot escape over the top of the barrier. However, the addition of an accelerating field reduces the barrier thickness and makes possible electron escape from a cold cathode by the so-called "tunnel effect." Field emission occurs when an appreciable number of electrons escape.

A Fermi-Dirac energy distribution for the number of electrons incident upon the barrier per unit time is assumed. The probability of transmission through the barrier, sometimes called the transmission coefficient, is a function of electron energy and barrier thickness, increasing with electron energy and decreasing with barrier thickness. The product of the Fermi-Dirac energy distribution and the transmission coefficient gives the distribution of the normal component of kinetic energy for the emitted electrons. The complete expression for the distribution function is complicated and difficult to use for the calculations of interest here. However, by making simplifying assumptions, Richter[10] has arrived at a tractable, approximate expression for the case of zero absolute temperature. Following Richter, we obtain for the energy distribution function of electrons passing the anode plane per unit time

$$P_f(w) = A_f(w_0 - w)\exp\left(-\frac{w_0 - w}{e\beta}\right) \quad \text{for } w_0 - e\mu < w < w_0$$
$$= 0 \qquad \text{otherwise} \tag{9}$$

* Received by the IRE, April 2, 1956. Prepared under AF Contract AF33(600)-27784.

[1] W. P. Dyke, "Progress in electron emission at high fields," PROC. IRE, vol. 43, pp. 162–167; February, 1955.

[2] S. Bloom and R. W. Peter, "A minimum noise figure for the traveling-wave tube," *RCA Rev.*, vol. 15, pp. 252–265; June, 1954.

[3] J. R. Pierce and W. E. Danielson, "Minimum noise figure of traveling-wave tubes with uniform helices," *J. Appl. Phys.*, vol. 25, pp. 1163–1165; September, 1954.

[4] H. A. Haus and F. N. H. Robinson, "The minimum noise figure of microwave beam amplifiers," PROC. IRE, vol. 43, pp. 981–991; August, 1955.

[5] J. R. Pierce, "A theorem concening noise in electron streams," *J. Appl. Phys.*, vol. 25, pp. 931–933; August, 1954.

[6] S. Bloom, "The effect of initial noise current and velocity correlation on the noise figure of traveling-wave tubes," *RCA Rev.*, vol. 16, pp. 179–196; June, 1955.

[7] A. Sommerfeld and H. Bethe, "Elketronentheorie der metalle," *Handbuch der Physik*, vol. 24/2, pp. 436–443; 1933.

[8] L. Nordheim, "Die theorie der elektronenemission der metalle," *Physik. Z.*, vol. 30, pp. 177–196; April, 1929.

[9] R. H. Fowler and L. Nordheim, "Electron emission in intense electric fields," *Proc. Roy. Soc. A*, vol. 119, pp. 173–181; June, 1928.

[10] G. Richter, "Zur geschwindigkeitsverteilung der feldelecktronen," *Z. für Physik*, vol. 119, pp. 406–414; September, 1942.

where

- A_f is independent of w,
- β is a voltage given numerically by $0.97\times 10^{-10}E/\sqrt{\phi}$,
- E is the accelerating field at the cathode in volts per meter,
- ϕ is the thermionic work function in volts,
- w_0 is an energy given by $(eV_0-e\phi)$ in MKS units, and
- $e\mu$ is the energy range in the Fermi-Dirac distribution for zero absolute temperature.

A plot of (9) for a typical set of parameters is shown by the solid-line curve of Fig. 1. The adjacent dotted-line curve is given to illustrate the effect of temperature. It was computed from an expression, more complete than (9), which takes into account temperature. Note that for the field-emission distribution the electron potentials are below the anode potential by approximately the work function. This is a consequence of the "tunneling." Shown for comparison are plots of the Maxwellian distribution (5) for two different values of cathode temperature.

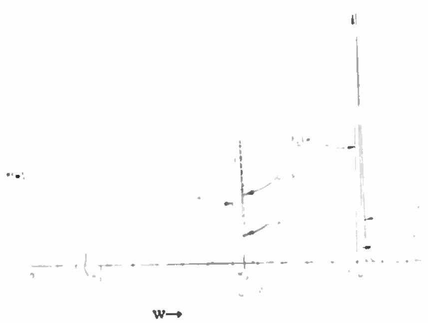


Fig. 1—A comparison of the energy distributions of electrons for high-field emission and for thermionic emission. The high-field emission curves assume a tungsten cathode and a field strength of 3.5×10^9 volts per meter. For tungsten $\phi=4.5$ volts and $\mu=5.7$ volts.

Using (9) and following the procedures indicated in (6) and (7) we obtain for the field emission case

$$\overline{v_0^2}=\frac{2w_0}{m} \qquad (10)$$

and

$$\overline{\delta v^2}=\frac{(e\beta)^2}{w_0 m}. \qquad (11)$$

Here we have assumed w_0 is much greater than the energy spread and that $\exp(-\mu/\beta)\ll 1$. Using (4), (10) and (11) in (3) gives for the beam noisiness

$$W\,|\,I_{\max}\,|\,|\,I_{\min}\,|=\sqrt{2}\,2e\beta B. \qquad (12)$$

By equating (8) and (12) we obtain the equivalent thermal-cathode temperature for field emission from a zero temperature cathode. Thus

$$T_{\text{eq.}}=\sqrt{2}\,\frac{e\beta}{k}=1.6\times 10^{-6}\,\frac{E}{\sqrt{\phi}}. \qquad (13)$$

For the solid-line curve of Fig. 1, $T_{\text{eq.}}$ is 2,640°K.

As (13) indicates, the equivalent temperature can be reduced by decreasing E and by choosing a metal with a higher work function. Since the total current emitted is approximately proportional to[9]

$$\frac{E^2}{\phi}\exp\left(-6.8\times 10^9\,\frac{\phi^{3/2}}{E}\right),$$

a small change in either E or ϕ unfortunately has enormous effect upon the emission. For example, in order to reduce the equivalent temperature to approximately a typical thermal-cathode temperature, say 1,160°K, the E for the above example must be reduced by a factor of 2.27 which results in a current reduction by a factor of 7×10^{10}!

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VHF Diffraction by Mountains of the Alaska Range*

During the winter of 1953–1954 it was learned that television signals of usable quality were being received at Lake Minchumina, Alaska, from two transmitters at Anchorage, 200 miles away. As Mt. McKinley, over 20,000 feet high, lies almost directly on the line of sight between the two locations, it is evident that a diffracted wave is responsible for the result. Signals were received on two frequencies, 57 mc and 200 mc, with roughly equal intensities, the peak transmitted power of the latter signal being 400 watts and the antenna gain about 7.

A preliminary field investigation in the summer of 1954 suggested that this was an interesting opportunity for studying diffraction propagation over very long paths. On either side of the line between Anchorage and Minchumina are two very prominent peaks, Mt. McKinley (20,300 feet) and Mt. Foraker (17,395 feet), which are connected by a rugged ridge 15 miles long, studded with peaks between 10,000 and 14,000 feet high. Mt. McKinley is the highest mountain in North America, and probably the highest in the world above its base plateau. There are no extensive foothill systems. The valleys on either side are extremely flat, over a hundred miles across, and nearly uninhabited.

In order to study temporal variations of signal strength, a receiving station was installed at Lake Minchumina, consisting of a receiver, a graphic recorder, and a direction-finding antenna. The 200 mc (Channel-11) signal was chosen, to simplify antenna construction. Throughout the winter of 1954–1955 occasional severe fades were noted, accompanied by swings in the direction of arrival of up to 12°. On occasion the fading would be frequency selective so that the video signal would fade out while the audio signal remained strong, and vice versa. At all other times the signal varied in a less extreme manner, with periods of from one to several hours. In part, this fading has been correlated with the rise and fall of the tides in the estuary (Knick Arm) at Anchorage; however, it seems certain that the major influences are meteorological. This contrasts with previous experiences with diffracted signals over shorter paths, which show, in general, relatively small variations of signal strength.

The space-variation of signal strength was investigated, during the summer of 1955, by airborne measurements taken from a C-47 airplane equipped with a 200 mc receiver and a steerable, 4-element, Yagi antenna. During these tests, the Anchorage 200 mc station transmitted a "black" picture; that is, a signal unmodulated except for synch pulses. Flights were made along a 100 mile path, centered at Lake Minchumina and perpendicular to the direction of propagation, at altitudes of 5,000, 10,000, 15,000, and 20,000 feet. Navigation was accomplished with the aid of the Minchumina radio range, which has a magnetic bearing of 216°-36°, almost exactly perpendicular to the direction of propagation. Radar fixes were obtained at the beginning and end of each run, and dead reckoning was used to interpolate to positions between these fixes. The most noticeable feature of the results was the very rapid fluctuation of signal strength from point to point over distances of a few hundred feet. Signals of as much as 10 microvolts (antenna output) were obtained on local maxima, and at minima the signal often faded into the noise (about 2 microvolts). At all points in the radio shadow of the mountains the signal displayed this violent variation. During the 20,000 feet altitude run the plane was heading northeast along the radio range, and when it emerged from the radio shadow of the McKinley mountain complex the signal became steady and strong (19 microvolts). At this point the descent was started, and the signal declined steadily and smoothly until it disappeared into the noise at about 12,000 feet. It is evident, then, that the mountains are responsible for the presence of the signal in the vicinity of Minchumina, and that the existence of the pronounced fine structure of the diffraction pattern makes the siting of a receiving antenna extremely important. It was noted during analysis of the flight data that coarse-structure maxima were obtained when the plane was in line with the larger mountains. These maxima, however, were much less important than the fine-structure maxima.

In order to verify the fine structure some additional data were taken on the ground at Minchumina Airport. The transmitter was modulated with a video test pattern, and the audio channel was disabled. In a direction parallel with the direction of propagation the signal voltage (antenna output across 50 ohms from a 10-element Yagi, gain 7.5 db) varied between 2.5 and 10.5 microvolts with periods of from 600 to 700 feet. The maxima were not all of the same intensity, but the pattern repeated itself, approximately, throughout the entire 4,000 foot run. A similar result was experienced

* Received by the IRE May 28, 1956. This work was performed as part of Air Force Contract No. AF19(604)-1089.

when the field intensity was explored in the direction perpendicular to the direction of propagation, except that in this case the minima were spaced from 50 to 100 feet apart.

Such a complicated interference pattern must be quite sensitive to small changes in phase of one or more of the interfering components; such changes might accompany meteorological changes, even though much of the transmission path is at high altitude. It is thought that the time-variations in signal strength and direction of arrival can be explained by changes in meteorological conditions over the paths of various signal components arriving via scattering and/or diffraction by widely-separated mountain peaks.

It was thought desirable to check the experimental signal strengths against those predicted by knife-edge diffraction theory. For the Anchorage-to-Minchumina path an obstacle height (above sea level) of 12,000 feet was assumed, this being approximately the average height of the profile between Mts. McKinley and Foraker. The other distances are as follows: transmitter to knife-edge, 132 miles; receiver to knife-edge, 68 miles; obstacle height above direct line between transmitter and receiver (using 4/3 earth radius): 15,400 feet. Using these data, the transmission loss between transmitter and receiver was calculated to be 150 db, assuming no image antennas. The maximum image gain from a 4-ray treatment would be 12 db. The signal actually received at Lake Minchumina with the receiving antenna at a theoretically optimum height corresponded with a transmission loss of 131 db. The signal was therefore some 7 db stronger than would be expected under the optimum 4-ray situation.

An additional experiment was conducted over a different path, to obtain another check of the validity of the knife-edge approximation. This path extended from College, Alaska (near Fairbanks) to the crossing of the Paxson-Cantwell Road and the Maclaren River, a distance of 135 miles, and crossed the Alaska Range in the vicinity of Mt. Hayes. The height of the knife-edge above sea level was assumed to be 8,000 feet, and the distance from transmitter (at College) to knife-edge was 97 miles. The transmitter was a modified SCR-270 radar set with a frequency of 107 mc. Using a 4/3 earth radius the one-ray transmission loss was computed to be 132 db. Adding 12 db for maximum image gain would give 120 db transmission loss. These figures are in agreement with the measured transmission loss of 126 db. No measurements were made to determine how this signal varied with position or time.

Despite the assumptions made necessary by lack of detailed topographical information, the above results appear to give substantial support to the validity of the knife-edge approximation.

The work reported here required the assistance of many organizations and individuals, to whom thanks are credit are due. The writer wishes to extend his sincere appreciation to Drs. C. T. Elvey, C. G. Little, and R. B. Dyce, and to Joseph Pope and Ernest Stiltner, of the Geophysical Institute of the University of Alaska; to Mr. Kurt Utley of Michigan State University; to the staffs of television stations KTVA in Anchorage and KTVF in Fairbanks; to Communications Equipment and Services Company of Anchorage; and to the personnel of the 5001st Operations Squadron at Ladd Air Base. To Mr. and Mrs. R. H. Collins of Lake Minchumina he owes his warmest thanks, not only for technical assistance but also for hospitality in the best Alaskan tradition.

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Measurement Considerations in High-Frequency Power Gain of Junction Transistors*

Introduction

A recent paper[1] concerning high-frequency power gain of junction transistors included experimental results of measurements of power gain for a number of grown-junction transistors. Since the appearance of the paper, the writer has been criticized for the method used to measure power gain, *viz.*, by employing *no* neutralization and by *not* conjugate matching in the input circuit. For example, one question that has been raised is whether a value of high-frequency gain for an unneutralized transistor has any meaning.

This letter explains in more detail why the measurements were made as they were. This was not done in the original paper because much of the necessary background material more properly belonged in a then-unpublished paper by J. G. Linvill.[2] Also given here is an extension of the theoretical results presented earlier to two types of neutralization commonly used in common-emitter transistor amplifiers.

Transistor Stability

For a particular transistor in a particular configuration, there are ranges of frequencies for which the transistor is *unconditionally stable; i.e.*, if no external feedback is employed, oscillations cannot be obtained with any possible passive terminations.[3] In these frequency ranges the transistor *per se* has an unambiguous maximum available power gain. The value of this gain can be calculated in terms of the four-pole parameters as shown by Linvill, and it can be measured by employing conjugate matching at both input and output with no risk of converting the transistor amplifier to an oscillator. On the other hand, in the other frequency ranges where the transistor is potentially unstable, oscillations may occur for particular terminations, and it is necessary to apply some sort of constraint in defining a "maximum" gain, since in reality the maximum gain is infinity. The constraint may take the form of neutralization,[4] of unilateralization,[5] or of requiring a given degree of "stability," with a suitable analytical definition for the latter.[6]

Since the type of constraint employed necessarily influences the value of "maximum" gain obtained, the writer preferred to avoid this situation if possible and to consider high-frequency power gain under conditions for which the transistor was unconditionally stable.

Investigation of the conditions for unconditional stability[3] for the usual ideal model of the junction transistor including ohmic base resistance r_b' and collector-base capacitance C_c, leads to the following results for *high* frequencies.[7] For the common-emitter configuration, unconditional stability is obtained for radian frequencies ω *greater* than a critical frequency[8]

$$\omega_{\text{crit}} = 0.4(r_e'/r_b')\omega_a, \tag{1}$$

where r_e' is the Shockley *et al* emitter resistance (kT/q_eI_e), and ω_a is the inherent radian alpha-cutoff frequency of the transistor. This equation is valid for either the fused-junction or the grown-junction transistor, since ω_{crit} is always less (by a factor of 2.5) than the frequency for which the distributed nature of the base resistance of a grown-junction transistor becomes significant, *i.e.*, for $\omega < 2.5\omega_{\text{crit}}$, the complex base impedance z_b' of a grown-junction transistor is resistive and is equal to r_b'.[9] On the other hand, for either the common-base or the common-collector configuration, unconditional stability is obtained only at frequencies *greater* than the inherent alpha-cutoff frequency ω_a.

* Received by the IRE, April 23, 1956.

[1] R. L. Pritchard, "High-frequency power gain of junction transistors," Proc. IRE, vol. 43, pp. 1075–1085; September, 1955.

[2] J. G. Linvill, "The relationship of transistor parameters to amplifier performance," presented at IRE-AIEE-Univ. of Penn. Conf. on Transistor Circuits, Philadelphia, Pa., February 17, 1955, submitted for publication, *Bell Sys. Tech. J.*

[3] A stability criterion for determining whether or not a particular four-pole is unconditionally stable has been derived in terms of its four-pole parameters by Linvill (see footnote reference 2), and independently by A. P. Stern, "Considerations on the stability of active elements and applications," 1956 IRE Conv. Rec., part 4. The criterion was quoted by A. P. Stern, C. A. Aldridge, and W. F. Chow, "Internal feedback and neutralization of transistor amplifiers," Proc. IRE, vol. 43, p. 839, (5); July, 1955.

[4] Using the definition quoted by C. C. Cheng, "Neutralization and unilateralization," Trans. IRE, vol. CT-2, p. 138; June, 1955, neutralization refers to "the process of balancing out an undesirable effect."

[5] Using the definition of Cheng, see footnote reference 4, and Stern *et al.*, see footnote reference 3, unilateralization refers to a method of rendering a bilateral network unilateral. Note that unilateralization is a special case of neutralization, but the converse is not necessarily true. Both of these authors also have pointed out the large number of different ways by which neutralization can be effected. See also G. Y. Chu, "Unilateralization of junction-transistor amplifiers at high frequencies," Proc. IRE, vol. 43, pp. 1001–1006; August, 1955.

[6] For example, see J. F. Gibbons, "Transistor amplifier performance," paper presented at IRE-AIEE-Univ. of Penn. Transistor Circuits Conf., Philadelphia, Pa., February 17, 1956. Also, D. D. Holmes and T. O. Stanley, "Stability considerations in transistor IF amplifiers," 1956 IRE Conv. Rec., part 4, and A. P. Stern, footnote reference 3.

[7] High frequencies are defined here as (radian) frequencies ω much greater than the common-emitter current-amplification-factor cutoff frequency $(1-\alpha_0)\omega_a$, where ω_a is the (radian) inherent alpha-cutoff frequency of the transistor; more exactly, as $\omega^2 \gg (1-\alpha_0)^2\omega_a^2$.

[8] Pritchard, *op. cit.*, footnote 13. Also derived independently by Stern, *op. cit.*

[9] Pritchard, *ibid.*, (9a).

The critical frequency of (1) is sufficiently low in general relative to ω_a that the common-emitter configuration provides a fairly wide range of "high" frequencies over which power gain can be measured *without* having to employ neutralization for stability. Accordingly, for the writer's measurements the common-emitter configuration was employed, and a measuring frequency of 5 mc was chosen for convenience as being above the critical frequency in general for the transistors used.

The reason for employing a purely resistive generator impedance rather than using conjugate matching in the input circuit was simply convenience. In general, such a measurement would *not* yield the maximum gain. However, as noted in the original paper, calculations show that for the junction-transistor model at *high* frequencies, greater than ω_{crit}, the maximum gain obtainable with a purely resistive generator is at most a db less than the *true* maximum gain, which *would* be measured with conjugate matching in the input circuit.

Furthermore, an additional calculation for the theoretical model, taking account of low-frequency parameters, has shown that potential instability for this method of measuring power gain with a purely resistive generator impedance occurs only at frequencies *less* than ω_{crit} of (1). In fact, the critical frequency in *this* case is the geometric mean of ω_{crit} and the common-emitter current-amplification-factor cutoff frequency $(1-\alpha_0)\omega_a$. For frequencies greater than *this* critical frequency, the value of gain measured is the available power gain quoted earlier, multiplied by a function of frequency that is equal to two at this critical frequency but then decreases rapidly to one with increasing frequency.

Neutralization or Unilateralization

Although it is *not* necessary to employ neutralization to maintain stability of transistors in the common-emitter configuration for frequencies greater than ω_{crit}, some sort of neutralization often *is* employed to minimize interaction between cascaded amplifier stages. Accordingly, it is of interest to present results of calculations of the maximum available power gain at high frequencies for the transistor model with the y type of neutralization or unilateralization commonly employed,[10] as shown in Fig. 1. In this circuit a feedback admittance y_f is connected between output and input terminals of the transistor through an ideal, lossless, phase-reversing transformer of turns ratio n:1. For neutralization, the feedback admittance generally is a neutralizing capacitor C_N such that

$$\begin{aligned} y_f &= j\omega C_N \\ &= -nIm(y_{12}), \end{aligned} \qquad (2)$$

whereas for unilateralization, by definition,

$$y_f = -ny_{12}, \qquad (3)$$

where y_{12} is the short-circuit feedback admittance of the transistor (numerically negative owing to the current-polarity convention used).

At frequencies $\omega > \omega_{crit}$, the maximum available power gain of the transistor *per se* $(y_f = 0)$ is given in terms of the equation presented earlier (for the constant $-r_b'$ model) as:

$$G_{max} = G_{av} \cdot K_G(c), \qquad \omega > \omega_{crit}, \qquad (4)$$

where

$$G_{av} = \frac{0.2}{\omega^2}\left(\frac{\omega_a}{r_b'C_c}\right), \qquad (5)$$

and $K_G(c)$ is Linvill's gain factor[2] as a function of his criticalness factor c, which for this case is simply $c = (\omega_{crit}/\omega)$. Normally K_G is essentially equal to one, but as c approaches one, indicating impending instability, K_G rapidly approaches the value 2.

If the capacitive-type neutralization is employed (2), the corresponding expression for maximum gain is

$$G_{max} = G_{av}\left[\frac{(\omega/\omega_{crit})^2 + 4}{(\omega/\omega_{crit})^2 + 2}\right] K_G(c'), \qquad (6)$$

where c' is the criticalness factor for the new amplifier and is a function of (ω/ω_{crit}). Note that this expression is independent of the turns ratio n (for the lossless transformer).

For the case of y unilateralization (3), the maximum available power gain *is* a function of the turns ratio n. However, if n is made sufficiently large relative to (r_e'/r_b'), the dependence can be essentially eliminated, in which case the maximum power gain is

$$G_{max} = G_{av}[1 + (2\omega_{crit}/\omega)^2]. \qquad (7)$$

The gain factor K_G for this case is equal to one, since the criticalness factor vanishes for the case of zero feedback.

A comparison between the power gain for these three cases is presented in Fig. 2, which shows the ratio G_{max}/G_{av} as a function of (ω/ω_{crit}). Note that as frequency is decreased below ω_{crit}, the maximum y unilateralized gain increases quite rapidly with respect to G_{av} (which already is increasing at the rate of 6 db per octave of decreasing frequency). Ultimately, however, the power gain is limited by low-frequency parameters which have not been included in these calculations.

It should be emphasized that these results correspond to only *two* of *many* possible methods of neutralizing a transistor amplifier.

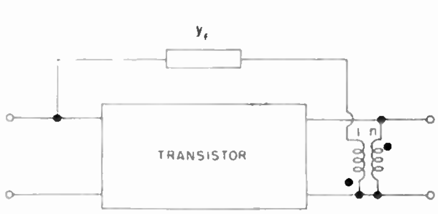

Fig. 1—Transistor with feedback (neutralizing) admittance and phase-reversing transformer.

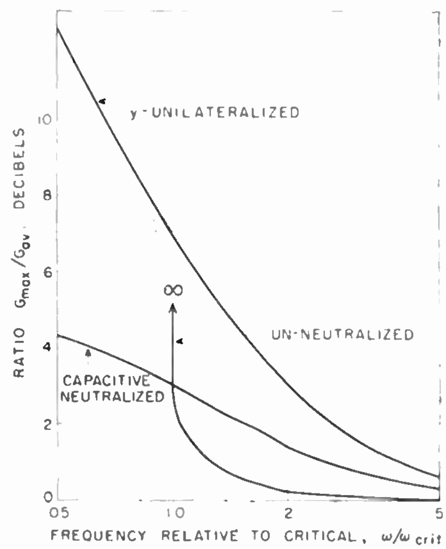

Fig. 2—Ratio of maximum available power gain to figure-of-merit power gain as a function of relative frequency for transistor amplifier; unneutralized, y neutralized, and y unilateralized.

Conclusion

A few additional remarks concerning the critical frequency of (1) may be of some interest. In particular, note that the value of C_c does not appear explicitly. This may be surprising at first inasmuch as C_c is the source of collector-base feedback, which gives rise to instability. However, at high frequencies, where the phase of alpha is significant, collector-base capacitance also contributes the dominant part of collector-emitter (output) conductance, which tends to stabilize the transistor. Hence, if two transistors at the same dc emitter current have identical values of ω_a, and of the *product* $r_b'C_c$, so that their high-frequency gains are identical, the transistor with the lower C_c will be stable down to a lower frequency, not because it has lower capacitance, but because it has higher base resistance!

Since a grown-junction type of transistor generally has a low collector capacitance and high ohmic base resistance relative to the fused-junction transistor, this may explain why the former type of transistor is more likely to be stable at the low end of the high-frequency range, *e.g.*, at 455 kc.

In conclusion, it should be emphasized again that all of the above discussion concerns frequencies that are high relative to the common-emitter current-amplification-factor cutoff frequency $(1-a_0)f_a$. However, a given frequency, *e.g.*, 455 kc, which may be high for one transistor, may be low for another in which $(1-\alpha_0)\omega_a$ and ω_a are both very high. For example, a transistor with f_a of the order of 500 mc, and $\alpha_0 = 0.98$[11] *may* be unconditionally stable at 455 kc simply because this is essentially dc for such a transistor![12]

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[10] See, for example, D. D. Holmes, T. O. Stanley, and L. A. Freedman, "A developmental pocket-size broadcast receiver employing transistors," Proc. IRE, vol. 43, pp. 663–664; June, 1954. Also, see L. J. Giacoletto, "Performance of a radio-frequency alloy junction transistor in different circuits," *Transistors I*, RCA Labs., Princeton, N. J., pp. 431–457; March, 1956.

[11] For example, the diffused-base transistor; C. A. Lee, "A high-frequency diffused base germanium transistor," *Bell Syst. Tech. J.*, vol. 35, p. 23; January, 1956.

[12] At low frequencies a junction transistor is unconditionally stable from dc *up to* a second critical frequency, which generally is somewhat less than $(1-\alpha_0)\omega_a$ and which is inversely proportional to C_c and to collector-base resistance. In this frequency range, C_c gives rise to potential instability in the same manner as does grid-plate capacitance in a conventional triode electron tube (Miller effect).

On the Waveform of a Radio Atmospheric at Short Ranges*

In view of the very great number of past experimental investigations[1] of waveforms of atmospherics, it seems worthwhile to give further consideration to this problem from a theoretical standpoint. Probably the first approach in this direction was carried out by Ollendorf.[2] The lightning flashes were represented by dipole moments which vary with time, so that the fields produced can be obtained with the aid of a Hertz vector. In a previous theoretical analysis, the author[3] considered a vertical electric dipole on a curved imperfectly conducting earth for the case of a current excitation which varied exponentially with time. It is the purpose of this note to present some calculations in graphical form showing the nature of the transient response of idealized lightning discharge at short ranges where the ionospherically reflected wave can be neglected or separately accounted for.

The instantaneous product of the dipole current and vertical height is denoted by $P(t)$ and is represented as follows

$$P(t) = \sum_{n=0}^{\infty} p_n e^{-\Gamma_n t} u(t) \qquad (1)$$

with $u(t) = 0$ for $t < 0$ and $= 1$ for $t > 0$, where p_n and Γ_n are real coefficients. It is known from the work of Bruce and Golde[4] and others[1] that only about two or three terms in the above expansion are adequate to represent the electric moment of a typical discharge. Employing the results of a previous theoretical analysis,[3] the electric field $E(t)$ at a distance D meters on a homogeneous earth of conductivity σ is

$$E(t) = \frac{2 \times 10^{-7}}{D} u(t') \sum_{n=0}^{\infty} p_n B_n(t') \qquad (2)$$

where

$$B_n(t) = \left(\frac{t}{2\alpha^2} + \Gamma_n - \frac{C}{D}\right) e^{-t^2/4\alpha^2} + \frac{C^2}{\Gamma_n D^2} + \left(\frac{C}{D} - \frac{C^2}{\Gamma_n D^2} - \Gamma_n\right) e^{-\Gamma_n t}\left[1 + \pi^{1/2}\alpha\Gamma_n e^{\Gamma_n^2\alpha^2} \cdot \left[\operatorname{erf}\left(\frac{t}{2\alpha} - \alpha\Gamma_n\right) + \operatorname{erf}(\alpha\Gamma_n)\right]\right] \qquad (3)$$

where $t' = t - D/C$, $C = 3 \times 10^8$ meter/sec, $\operatorname{erf}(z) = 2\pi^{-1/2}\int_0^z e^{-x^2} dx$, and

$$\alpha = \left(\frac{D}{21.6\pi\sigma}\right)^{1/2} \times 10^{-9}.$$

In the preceding formula, the effect of earth curvature has not been included. It can be shown[3] that the flat earth approximation is valid for transient fields if t' is greater than about 0.2 microseconds and the range D is not much greater than 50 km. If the earth conductivity is allowed to approach infinity, the preceding equation reduces to

$$B_n(t) = \delta(t) + \frac{C^2}{\Gamma_n D^2}(1 - e^{-\Gamma_n t}) + \left(\frac{C}{D} - \Gamma_n\right) e^{-\Gamma_n t} \qquad (4)$$

where $\delta(t)$ is the unit impulse or "Dirac" function.

As an example, the response $B(t)$ is calculated for a source function, given by

$$P(t) = P_m\left[e^{-\Gamma_1 t} - e^{-\Gamma_2 t}\right] \qquad (5)$$

which is clearly a special case of (1) where only two terms are considered. P_m is a constant which is independent of time and would be proportional to the peak magnitude of the product of the current times the height of stroke column. Choosing $\Gamma_1 = 0.02 \times 10^6$ sec^{-1} and $\Gamma_2 = 0.50 \times 10^6$ the function $P(t)/P_m$, which can be called the stroke dipole moment, is plotted in Fig. 1. It is believed that this pulse shape is representative of the main return stroke of the lightning discharge having a build up time of about 10 μsec and pulse width of about 50 μsec.

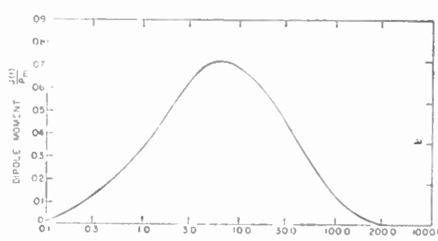

Fig. 1—Idealized stroke dipole moment: $P(t)/P_m = e^{-\Gamma_1 t} - e^{-\Gamma_2 t}$.

The field response $B(t)$ for this particular source is shown plotted in Fig. 2 for various ranges over a perfectly conducting ground. It is interesting to note that at larger ranges the shape of the response curves approach an asymptotic value which is closely proportional to the first time derivative of the source dipole moment. At shorter ranges, the shape of the response curves is very different and approaches the time integral of the dipole moment. The curves illustrated in Fig. 2 would be modified at small times to some extent if the finite conductivity of the ground were considered. As an example, the function $B(t)$ for a range of 100 km is shown plotted in Fig. 3 for ground conductivities of ∞, 10^{-2}, and 10^{-3} mhos per meter. The earth curvature would also have some effect on the response curves at very small times, say less than 0.2 μsec. In the event that one should become interested in this region of the time domain, reference can be made to an earlier paper by the author.[3]

It is hoped that more detailed experimental results on atmospheric waveforms will be forthcoming. For the purpose of comparing the experimental results with theory, it is highly desirable that simultaneous wave forms be made at two or more stations whose distances to the source is in the range from 5 to 200 km. The recent simultaneous recordings taken at the University of Florida[5] seem very promising. Despite the fact that the distances to strokes were not known accurately, the qualitative transformation of the wave shape with distance is in agreement with the theoretical curves presented here.

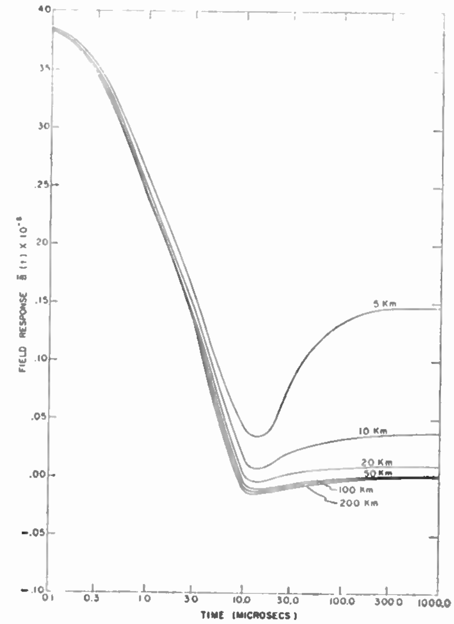

Fig. 2—Field response for idealized stroke dipole moment; $\sigma = \infty$.

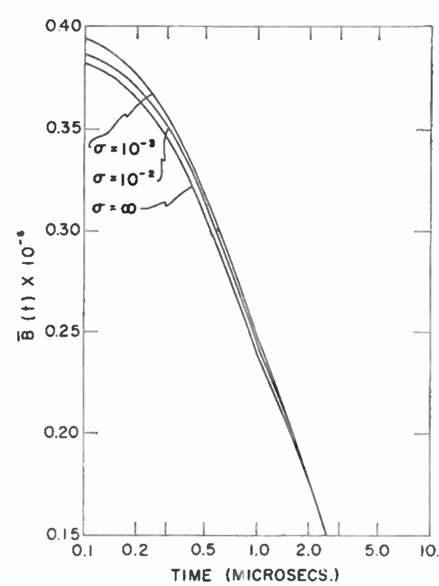

Fig. 3—Field response for idealized stroke dipole moment; $D = 100$ km.

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* Received by the IRE, April 9, 1956.

[1] An excellent summary of past work in this field is given in "Bibliography relating to lightning atmospheric location," Lightning and Transients Res. Inst. Rep. No. 181, Minneapolis, Minn., 1955.

[2] F. Ollendorf, "Radiation field of lightning," *Elekt. Nachr.-Tech.*, vol. 7, pp. 108–119; March, 1930.

[3] J. R. Wait, "Transient fields of a vertical dipole over a homogeneous curved ground," *Can. J. Phys.* vol. 34, pp. 27–35; January, 1946.

[4] C. E. R. Bruce and R. Golde, "The mechanism of the lightning stroke," *J. IEE*, vol. 88, pp. 487–497; December, 1941.

[5] A. W. Sullivan, S. P. Hersperger, R. F. Brown, and J. D. Wells, "Investigation of atmospheric radio noise," Scientific Report No. 9, Dept. of Elect. Eng., University of Florida, October, 1955.

A Balanced, Unregulated, Dual Power Supply*

When using certain types of dc amplifiers requiring two voltages of opposite polarity and equal amplitude or servo potentiometers that require a physically fixed zero voltage reference point or any instrument requiring two voltages of opposite polarity and equal amplitude, it is desirable to have a power supply that will vary both ± voltages simultaneously and by like amount regardless of changes occurring in the ac line.

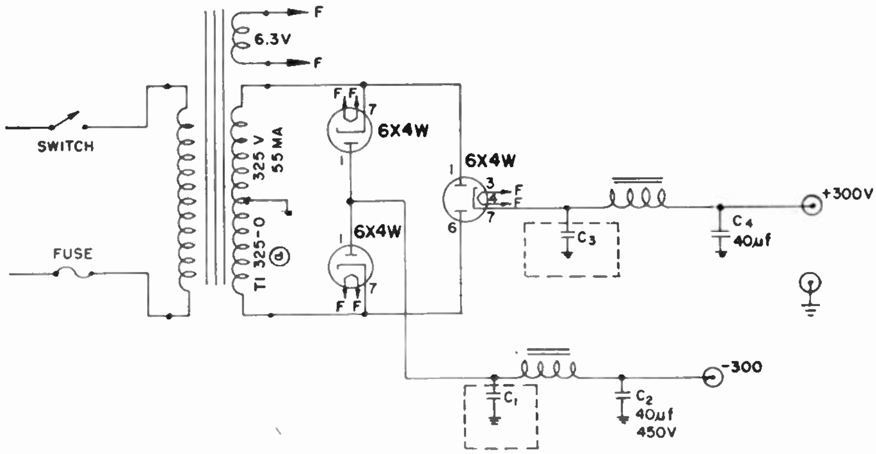

Fig. 1—A balanced dual power supply. (Note: C_1C_3, small value capacitors, used if necessary to adjust the level of $B\pm$ voltages.)

Fig. 1 displays a schematic diagram of a balanced, unregulated, dual power supply that is inexpensive to construct, but one that will hold both ± voltages to an equal amplitude without the use of regulator tubes.

The power supply is of the bridge type except for the grounded center tap of the power transformer secondary which provides two distinct voltages of equal amplitude and opposite polarity. These two voltages are easily adjusted to equal amplitude by adjustment of the value of the capacitors, C_1 and C_3 (Fig. 1).

Due to the type of circuitry used, a transformer with a full wave rectifier rating of 325-0-325 volts at 55 milliamperes of current can be used to provide +300 V at (55 ma) (.707) = 38.9 ma and −300 V at (55 ma) (.707) = 38.9 ma without overrating the transformer.

Hence, the rated current should be specified in rms current within the winding. In practice, however, the current rating of most $B\pm$ supply transformers is given in dc output current to make it easier for the design engineer. This current rating assumes a full wave rectifier with a condenser-input filter. If the type of power supply circuit used is any other type, the current rating of the transformer must be modified accordingly.

In the balanced-bridge type power supply used in this application, the current through the transformer winding is sinusoidal in the absence of filters. When working into a LC filter, the sine wave will become somewhat distorted. In this application, a grounded center tap was used rather than the conventional bridge-type circuit, thus giving two balanced supplies of 300 volts each rather than one supply of 600 volts as would be obtained from a conventional bridge circuit with the transformer secondary center tap ungrounded.

Fig. 2 displays a graphed result of tests made of the balanced, unregulated, dual power supply, used in conjunction with a dc amplifier, offering an essentially balanced load, both at zero volts output and maximum signal output. The only necessary condition was to assure that the two power supply voltages would remain balanced although changes in the ac line voltage occurred. In the test, two dc amplifiers were connected in cascade, and sufficient additional resistance to make up a total load of plus 40 ma and minus 40 ma of current were used. An input signal was fed into the first cascaded amplifier; the output of the second amplifier was then used to buck out the input signal, the difference then being the error voltage of the amplifier's output.

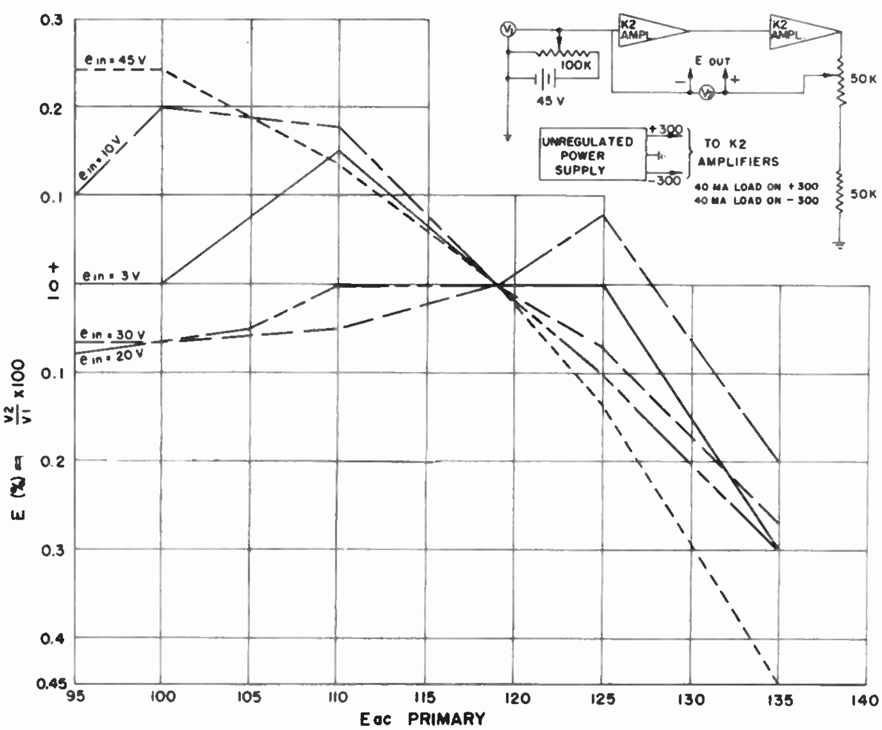

Fig. 2—Varying of ac input voltage into unregulated power supply vs output voltage of Philbrick K^2 amplifier.

A graph (Fig. 2) discloses the percentage of output voltage error of the amplifiers when using a known input voltage to the amplifiers and various transformer input voltages.

The dc amplifiers were zeroed with the ac line voltage set at 118 volts. Voltages ranging from 95 volts to 135 volts ac were fed into the primary of the power transformer. The largest error voltage graphed proved to be less than ½ of 1 per cent, with the ac primary input voltage to the power transformer set at 135 volts and with the dc input voltage to the amplifier set at 45 volts. All other error readings graphed fell within 3/10 of 1 per cent.

The physical size of the dual power supply may be kept very small as compared to regulated power supplies serving the same purpose. Only three tubes and sockets, one power transformer, two chokes, and two filter condensers are required.[1] Identical chokes and filter components are used in both positive and negative legs of the supply to maintain like conditions in both sections.

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* Received by the IRE, March 7, 1956.

[1] Filter condensers C_{1-3} (Fig. 1) are small value capacitors and used only if necessary to adjust $B\pm$ voltage requirements.

Systemic Learning*

W. P. Tanner, Jr.[1] has recently advanced a suggestion which promises to add considerably to our understanding of that basic intelligent process which we call learning. The idea as advanced is embryonic in development, but it can be seen to embody concepts of system design such as those much used by communication engineers, together with dependence on mathematical probability theory and its outgrowth—statistical decision theory. If we accept certain philosophical limitations due to dependence on probability theory as being relatively nonrestrictive, then we may use the first characteristic to suggest the name systemic learning in order to differentiate between Tanner's suggestion and most preconceived notions of learning.

Briefly, Tanner's suggestion begins with the system indicated in Fig. 1. It consists of an input, an output, and a relation between the two. We usually say that learning reflects a change in the *relation.* Tanner continues with an adjunct as indicated in Fig. 2, where the adjunct may be regarded as a (possibly optimum) rule for changing the relation as based on experience (estimates of probabilities) when the system of Fig. 1 operates in an environment.

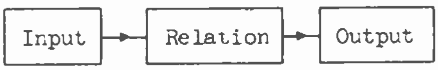

Fig. 1—A general system.

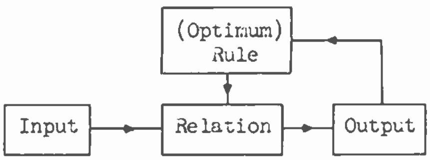

Fig. 2—A self-modifying system.

If now the adjunct is included with the original system to form a new system, we see that the new system is essentially fixed in character, since it is provided, by its design, with the only (fixed) rule which it needs to adapt itself to environmental influences in a manner which may well be optimum. Being fixed, the new system cannot change in over-all design by any experience; hence it need not be regarded as exhibiting learning. The actual "behavior" of the original system we can call systemic learning; and then we say, with Tanner, that there is no such thing as systemic learning, meaning thereby that what we call systemic learning is not unique but depends on our choice of system, and that with proper choice of system there is no systemic learning. If learning as commonly understood should be equivalent to systemic learning as here discussed, then we are free to say that there is no such thing as learning.

It seems to the writer than Tanner was led to the above point of view from his experience with human observers in strictly defined detection experiments modeled after statistical decision theory (hence mathematical probability theory) and in which he found his observers performing in a manner not far from optimum, however defined within the framework of the theory.

Being familiar with this previous work the writer has added his own engineering background and philosophical interests to frame the above discussion for the benefit of others with engineering background. It does not appear proper to go into questions of philosophical and religious implication of the above ideas, but if the writer's feeling is borne out, there will be many of them. They depend in part on the first mentioned philosophical implications of mathematical probability theory.

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* Received by the IRE, May 7, 1956.
[1] In private communication.

On the Use of a Special Word for the Quantity "Angular Velocity"*

In electric-circuit analysis, the use of the quantity "angular velocity," ω, in place of the related quantity "frequency," f, has great mathematical utility and simplicity. However, there is no simple designation for the term when speaking of it, leading to corruptions such as "angular frequency," "radian frequency," or even the erroneous term "frequency" alone. The repeated occurrence of the quantity "angular velocity" in written or spoken technical language makes the two-word designation unwieldy. It is accordingly suggested that a word be coined to designate this characteristic quantity of a circuit or of a sinusoidal time function. The words "pulsatance" and "pulsation" have been so used, but carry little general acceptance, are ambiguous, and may be confused with pulses or pulsation in the ordinary sense, to which they are related only as frequency is related. The lack of acceptance of such terms implies difficulty, so it is proposed that another term be selected. The main purpose of selecting a new term is to obtain one which is not ambiguous.

It is the purpose of this article to invite discussion and suggestions for possible words.

Three possible words are here suggested. The first two end in "cy" to show the close relation to "frequency," and both have stems that imply rotation or angular velocity. They are "rotency" and "angulancy." The third is derived from the reasonably meaningful corruption, "radian frequency": "frecrad." (The simpler spelling is preferred to "frequerad".) A 60-cycle (per second) signal would have a "frecrad" of 377 radians per second, for example (or an angulancy or rotency of 377 radians per second).

Let us further proceed to the generalized complex-quantity angular velocity that is becoming so common in complex-plane plots of poles and zeros of networks. Let us consider the complex S plane where $S=\sigma+j\omega$. In the ω-direction, the units are radians per second; in the σ-direction, the units are nepers per second; in either case the natural time function is a complex exponential $f(t)=K\epsilon^{st}$, regardless of the relative magnitudes of the "real" and "imaginary" components. Here are two distinct units—radians per second and nepers per second—that are intimately related. It is proposed that a common term—the "nerad"—be supplied for the general case. A complex network will thus have for one of its natural "frecrads" a value of so many "nerads." Since the term is coined, the "per-second" is preferably specifically included in its meaning: one "nerad" is equivalent to one *radian per second*, or one *neper per second*, or any combination of the two having unit magnitude. If the behavior is wholly undamped, the "frecrad" should be expressed in radians per second; if completely nonoscillatory, the "frecrad" should be expressed in nepers per second; if the behavior is a damped oscillation, the "frecrad" is complex, and expressed in nerads.

What do you think?

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* Received by the IRE, May 21, 1956.

Frequency Doubling and Mixing in Ferrites*

Ayres, Vartanian, and Melchor[1] have shown recently that the equation of motion for the magnetization vector in a ferrite predicts a frequency doubling, and they have experimentally observed this doubling. They showed that with the usual arrangement, in which the biasing field H_{dc} is applied along the z axis, the rf z magnetization m_z varies at twice the applied frequency; this can be used to induce a double-frequency voltage in a loop with the proper experimental arrangement. It was also pointed out by Ayres, Vartanian, and Melchor that frequency mixing would occur if two frequencies were applied to the ferrite; however, this mixing was not discussed. We have carried out a simple analysis of this effect in terms of applied magnetic fields and demagnetizing factors.[2] The results are valid for sample shapes and sizes for which the concept of demagnetizing factors is valid. Damping is neglected throughout the analysis. It is the purpose here to summarize some of the more interesting results. The equations of motion for the magnetizing vector are

$$\dot{m}_x = \gamma(m_y H_z - M_z h_y), \tag{1}$$

$$\dot{m}_y = \gamma(M_z h_x - m_x H_z), \tag{2}$$

$$\dot{m}_z = \gamma(m_x h_y - m_y h_x). \tag{3}$$

* Received by the IRE, May 14, 1956. This work was carried out under Contract AF19(604)-1084, while the author was an RCA Fellow in Electronics.
[1] W. P. Ayres, P. H. Vartanian, and J. L. Melchor, "Frequency doubling in ferrites," *J. Appl. Phys.*, vol. 27, p. 188; February, 1956.
[2] J. E. Pippin, "Frequency Doubling and Mixing in Ferrites," Sci. Rep. No. 2, AFCRC-TN-56-369, Gordon McKay Lab., Harvard Univ., Cambridge, Mass.; May 5, 1956.

As usual, we let small letters denote time varying quantities and capital letters denote dc quantities. In the first two equations it is assumed that $m_z \ll M_z$ and $h_z \ll H_z$. These equations, which are in terms of internal fields, can be written in terms of applied fields and demagnetizing factors in the usual way. If the applied field consists of two sinusoids with frequencies ω_1 and ω_2, then the first two equations can be solved under the above assumptions of small m_z and h_z to give

$$\boldsymbol{m} = \chi_1 \cdot \boldsymbol{h}_{a_1} + \chi_2 \cdot \boldsymbol{h}_{a_2}, \tag{4}$$

where here the vectors $\boldsymbol{m}$ and $\boldsymbol{h}_a$ refer only to the x and y fields; the subscript a means *applied*. The components χ_{xx}, χ_{xy}, and χ_{yy} of the tensor $\boldsymbol{\chi}$ are just the Polder susceptibility tensor components written in terms of applied fields. (This is the "external susceptibility" tensor.) The subscripts 1 and 2 refer to frequencies ω_1 and ω_2. There will be one such tensor for each applied frequency, the only variation being the value of ω appearing in the components of the tensor. If (4) is substituted in (3), after (3) has been rewritten in terms of applied fields, and the very considerable algebra carried through, it is found that $\dot{m}_z$ contains terms of frequency $2\omega_1$, $2\omega_2$, $\omega_1+\omega_2$, and $\omega_1-\omega_2$. These terms are given in detail elsewhere;[2] they are not written here because of their complexity. However, the special case of a *cylindrically symmetric* ferrite sample will be considered here. Specifically, if

$$h_{x_a} = h_{x_1} \cos \omega_1 t + h_{x_2} \cos (\omega_2 t + \phi)$$

and

$$h_{y_a} = h_{y_1} \cos (\omega_1 t - \alpha) + h_{y_2} \cos (\omega_2 t + \phi - \delta),$$

then for a cylindrically symmetric sample the double frequency terms D in $\dot{m}_z$ are given by

$$\begin{aligned} D = \frac{\gamma}{2} \chi_{xy_1} [h_{x_1}^2 \sin 2\omega_1 t + h_{y_1}^2 \sin (2\omega_1 t - 2\alpha)] \\ + \frac{\gamma}{2} \chi_{xy_2} [h_{x_2}^2 \sin (2\omega_2 t + 2\phi) \\ + h_{y_2}^2 \sin (2\omega_2 t + 2\phi - 2\delta)]. \end{aligned}$$

The sum and difference frequency terms s and d in $\dot{m}_z$ are given by

$$\begin{aligned} \left.\begin{matrix} s \\ d \end{matrix}\right\} = \frac{\gamma}{2} (\chi_{xy_1} \pm \chi_{xy_2}) \{ h_{x_1} h_{x_2} \sin [(\omega_1 \pm \omega_2)t \pm \phi] \\ + h_{y_1} h_{y_2} \sin [(\omega_1 \pm \omega_2)t \pm \phi - \alpha \mp \delta] \} \\ + \frac{\gamma}{2} (\chi_{xx_1} - \chi_{xx_2}) \\ \cdot \{ h_{x_1} h_{y_2} \cos [(\omega_1 \pm \omega_2)t \pm \phi \mp \delta] \\ - h_{x_2} h_{y_1} \cos [(\omega_1 \pm \omega_2)t \pm \phi - \alpha] \}, \end{aligned}$$

where the upper of the double signs applies for s and the lower for d. Upon inspection of these equations the following conclusions are evident.

1) If for a particular frequency ω_1 (or ω_2), the excitation is circularly polarized in either sense ($h_{x_1} = h_{y_1}$; $\alpha = \pm\pi/2$), *there is no doubling*. The amplitude of the double frequency term is zero and the magnetization vector precesses in a circle. Furthermore, the amplitude of the double frequency term is a maximum when the excitation is linearly polarized, for a given amplitude of excitation.
2) If the two excitations are circularly polarized in the same sense ($\alpha = \delta = \pm\pi/2$), then the amplitude of the sum term is zero, and the amplitude of the difference term is, in general, not zero.
3) If the two excitations are circularly polarized in opposite senses ($\alpha = -\delta = \pm\pi/2$), then the amplitude of the difference term is zero, and the amplitude of the sum term is, in general, not zero.
4) If $\dot{m}_z$ is used to induce a voltage in some manner, then statements 1), 2), and 3) show that it is possible to obtain *single sideband* mixing by applying two circularly polarized waves to the ferrite.

The writer wishes to acknowledge the valuable advice of Prof. C. L. Hogan.

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The Optimum Tapered Transmission Line Matching Section*

A recent paper by R. E. Collin[1] treats the problem of the design of an impedance matching transmission line taper of optimum characteristic impedance contour, *i.e.*, optimum in the sense that for a given maximum reflection coefficient tolerance in the pass band the taper has minimum length, or, conversely, for a given length the taper has minimum reflection coefficient tolerance in the pass band. Since the problem treated by Collin is identical with the one I treated in a paper which appeared several months earlier,[2] some comment on the relationship between the two papers seems to be called for.

Since the two papers treat the identical problem, one would think that they should arrive at identical results even though possibly by different routes. That this is, in fact, true is nowhere indicated in the body of Collin's paper, and, instead, it is implied throughout that one must settle for a rather involved compromise design in which the sidelobes of reflection coefficient ultimately decay inversely with frequency. This results from the unnecessary assumption that the characteristic impedance is a continuous function in the closed interval including the end points of the taper. When this assumption is removed, a simple design formula for the exact optimum taper can be obtained as indicated in my paper.[2] Although Collin indicates at the very end of his paper in an appendix that the exact optimum taper can be obtained (which violates his assumption of continuous Z_0) by considering a limiting form of his solution, an erroneous impression has already been given, and unless the reader perseveres to the very end of the paper he will never find out that a much simpler design than that indicated is available and applicable.

The above seems to indicate to me the danger of carrying over too closely the work from one field (in this case, antenna aperture theory) to another field in which the mathematics is analogous. While the mathematics may be exactly analogous, there may be important characteristics of the systems which differ, and, hence, strongly affect the nature of the results. The assumption of a discontinuous change of characteristic impedance in the transmission line taper is completely consistent with the physics of the problem. The corresponding assumption in antenna aperture theory, however, implies the presence of an impulse function in the excitation function, and, hence, implies infinite radiated power—an obviously nonphysical condition. This is exactly the reason that Taylor[3] found it necessary to go to a rather deep analysis in treating the problem of optimum distributions on continuously excited antenna apertures.

While Collin's paper furnishes a very instructive example of mathematical analysis, it could have been considerably simplified by removal at the outset of the unnecessary assumption of continuous Z_0. The results so obtained (as indicated in his appendix) are then equivalent to the results of my earlier paper. The Fourier series representation for the ideal characteristic impedance contour which was ultimately obtained by Collin is quite interesting and converges fairly rapidly (requires twelve terms for three figure accuracy in at least one practical case). There is a lot to be said computationally for the definite integral representation of my paper, however. Not the least of these is that it has been computed and tabulated,[2] thus rendering the design of an optimum transmission line taper particularly direct and convenient.

Finally, I would like to thank Dr. Collin for pointing out the omission in (12) of my paper. The equation will read correctly with the addition indicated by Dr. Collin except at the end point $x = -l/2$, or, more precisely, it may be corrected to read:

$$\begin{aligned} \ln (Z_0) &= \frac{1}{2} \ln (Z_1 Z_2) + \frac{\rho_0}{\cosh (A)} \\ &\quad \cdot \left\{ A^2 \phi(2x/l, A) + U\left(x - \frac{l}{2}\right) - U\left(-x - \frac{l}{2}\right) \right\}, \quad |x| \leq l/2, \\ &= \ln (Z_2), \qquad x > l/2, \\ &= \ln (Z_1), \qquad x < -l/2. \end{aligned} \tag{12}$$

This omission had also been called to my attention in a private communication by G. J. Wheeler of the Raytheon Wayland Laboratory. He also noted that on the last page of the paper one should read $l/\lambda = 0.587$ rather than $\beta l = 0.587$. While I must apologize for my faulty proof reading, I would like to point out that these in no way affect the results of the paper nor do they affect

* Received by the IRE, April 27, 1956.
[1] R. E. Collin, "The optimum tapered transmission line matching section," Proc. IRE, vol. 44, pp. 539–548; April, 1956.
[2] R. W. Klopfenstein, "A transmission line taper of improved design," Proc. IRE, vol. 44, pp. 31–35; January, 1956.
[3] T. T. Taylor, "Design of line-source antennas for narrow beamwidth and low sidelobes," Trans. IRE vol. AP-3, pp. 16–28; January, 1955.

the tabulated values of $\phi(z, A)$ which were given.

I hope that these remarks may have clarified the situation in regard to the design of optimum impedance matching transmission line tapers.

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In connection with R. E. Collin's paper,[1] I should like to make the following comments.[4]

In his paper, Collin refers to my letter to the Editor of PROCEEDINGS[5] in which I showed how a nonlinear Riccati equation for the reflection coefficient ρ could be linearized by neglecting the squared term ρ^2 that is assumed to be small compared to unity. The solution of the resultant linear differential equation was interpreted as being the Fourier transform of $P(x) = (1/2)\cdot(d\ln Z_0(x)/dx)$, where $P(x)$ is the reflection from a length dx of the tapered line at a distance x, and $Z_0(x)$ is the characteristic impedance varying along the line. Collin refers to my letter in connection with tapered transmission lines that have triangular $P(x)$ distributions. He calls these tapers Gaussian. However, in a more extensive research work,[6] indicated in the last part of the letter,[5] I have already introduced the designation Gaussian for tapered lines that have Gaussian $P(x)$ distributions. A set of graphs showing $Z_0(x)$, $P(x)$, and $\rho(l/\lambda)$ (l = length of the tapered line, λ = wavelength) for different Gaussian tapers is shown in reference 6. The latter designation seems to me to be the more consistent one because the Gaussian distribution is an important self-reciprocal Fourier transform that is well known from probability theory. This notation is also analogous to notations used in other engineering fields, for example, in antenna theory.[7] In engineering we usually deal with incomplete Gaussian distributions. The Fourier transforms of these distributions have been studied by Millington[8] by means of the saddlepoint method.

A quick literature survey reveals that the extensive Fourier transform work mentioned above, in addition to being referred to in numerous papers and books, is reviewed in:

1) *Science Abstracts*, Section B, Electrical Engineering, Vol. 54, p. 458, 1951; (Abstract 3439).
2) *Wireless Engineer*, Vol. 28, p. A226, December, 1951; (Abstract 2909).
3) PROCEEDINGS OF THE IRE, Vol. 40, p. 116, January, 1952; (Abstract 2909).
4) *Mathematical Review*, Vol. 13, p. 803, 1952.
5) *Annales des Télécommunications*, Vol. 8, p. A6, 1953; (Abstract 51173).

It appears that these reviews have not been brought to the attention of Dr. Collin.

Lately a brief summary of this work has been given in *Proceedings*.[9]

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[4] Received by the IRE, May 2, 1956. This work was supported in part by the Army (Signal Corps), the Air Force (Office of Sci. Res., Air Res. and Dev. Com.), and the Navy (Office of Naval Res.).

[5] E. F. Bolinder, "Fourier transforms in the theory of inhomogeneous transmission lines," PROC. IRE, vol. 38, p. 1354; November, 1950.

[6] E. F. Bolinder, "Fourier transforms in the theory of inhomogeneous transmission lines," *Trans. Roy. Inst. Tech.*, Stockholm, Sweden, No. 48, pp. 84; 1951.

[7] J. F. Ramsay, "Fourier transforms in aerial theory," *Marconi Rev.*, Part III, vol. 10, pp. 41–58; April/June, 1947.

[8] G. Millington, "The Fourier transform of the incomplete Gaussian function," *Marconi Rev.*, vol. 11, pp. 17–30; January/March, 1948.

[9] E. F. Bolinder, "Fourier transforms and tapered transmission lines," PROC. IRE, vol. 44, p. 557; April, 1956.

Rebuttal[10]

From a theoretical point of view the assumption of a continuous taper is not necessary and, as pointed out by Dr. Klopfenstein, does result in a simpler analysis. The approach to the problem as given in my paper does, however, give additional useful information such as:[1]

1) An estimate of the range of variation in the load impedance and pass band tolerance for which the theory can be expected to give an accurate result.
2) Design information for a taper which departs only slightly from the ideal taper and does not have a discontinuity at each end. A discontinuity in characteristic impedance may be undesirable in certain cases as for example when it leads to a sharp discontinuity in the physical structure of the line with its consequent limitation on the power handling capability of the line.

The tabulated data given in the paper by Klopfenstein is of considerable value.[2] The results of my analysis could of course be tabulated in a similar form. The difference in the two taper designs (continuous taper and taper with a discontinuity at each end) is negligible in practice (the continuous taper is 5–10 per cent longer) considering that either design is only approximate since it neglects the square of the reflection coefficient as compared to unity in the differential equation governing the system.

The terminology introduced by Mr. Bolinder has its own merits but does not appear to be in accordance with the usual terminology used in regard to tapered transmission lines. Southworth mentions the use of the word "Gaussian" to describe a taper with a characteristic impedance varying as an "incomplete" Gaussian function.[11] Other tapered transmission lines have been named after the type of function that describes the characteristic impedance along the taper, *e.g.*, exponential,[12] hyperbolic,[13] parabolic,[14] etc. If the terminology introduced by Mr. Bolinder is used, then, in order to be consistent, one should refer to the exponential taper for which

$$\frac{d}{dx}\ln Z_0(x) = \text{constant},$$

as a constant taper. At the same time one would be somewhat at a loss for a suitable name for the linear taper for which

$$\frac{d}{dx}\ln Z_0(x) = \frac{A}{Ax+B}$$

(A and B suitable constants).

I regret to say that I have not yet read the "extensive Fourier transform work" of Mr. Bolinder, although I have been aware of its existence for some time.[6] In fact I wrote a letter requesting some information regarding this work to the author addressed to the Royal Institute of Technology in Stockholm before submitting the manuscript of my paper for publication but, for some unknown cause, did not receive a reply.

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[10] Received by the IRE, May 14, 1956.

[11] G. C. Southworth, "Principles and Applications of Waveguide Transmission," D. Van Nostrand Co., Inc., New York, N.Y., sec. 9.1; 1950.

[12] H. A. Wheeler, "Transmission lines with exponential taper," PROC. IRE, vol. 27, pp. 65–71; January, 1939.

[13] H. J. Scott, "The hyperbolic transmission line as a matching section," PROC. IRE, vol. 41, pp. 1654–1657; November, 1953.

[14] R. F. H. Yang, "Parabolic transmission line," PROC. IRE, vol. 43, p. 1010; August, 1955.

Marconi's Last Paper, "On the Propagation of Microwaves over Considerable Distances"*

Recently Professor R. M. Fano translated from Italian into English Marconi's last paper (1933), only thirteen sentences long. I have never seen any reference to this paper in any English language periodical, not even in the moving tribute to Marconi by fellow inventor E. H. Armstrong.[1] Fano's complete translation of the text, which follows this letter, will interest readers of PROCEEDINGS, in view of the current revival of interest in this subject.

Considered as a sequel to Marconi's longer paper[2] (also rarely mentioned or read now, alas), this remarkable paper indicates almost certainly that Marconi discovered experimentally the existence of useful microwave radio propagation into the twilight

* Received by the IRE, May 4, 1956.

[1] E. H. Armstrong, "The spirit of discovery: an appreciation of the work of Marconi," *Elect. Engrg.*, vol. 72, pp. 670–676; August, 1953.

[2] G. Marconi, "Radio communication by means of very short electric waves," *Proc. Roy. Inst. G. Brit.*, vol. 27, pp. 509–544; 1933. This paper is expected to be republished in an early issue of the IRE TRANS. ON ANTENNAS AND PROPAGATION.

region beyond the horizon which could not be explained theoretically by diffraction and refraction. The increase in range from 52 to 150 km, from 1.7 to 5 times geometrical optical, between the experiments of 1932 and 1933, with only transmitter power increase from about 10 to 25 watts, and improvements in receiving equipment and reflectors, seems unmistakably to indicate he was experimenting in the twilight region where the field strength attenuation rate is now known to be roughly 0.1 db/km, instead of the value about 10 times higher calculated from diffraction theory with a refraction correction. The skepticism of the experts concerning this last great propagation discovery by Marconi clearly parallels the similar history of his 1901 discovery of transatlantic wireless and his short wave revolution of the early nineteen twenties, except that this time his ill health and death in 1937 caused the skeptical radio world to wait a generation before rediscovering the truth and importance of the phenomenon. Although Hershberger's experiments[3] and those of Trevor and George[4] soon afterwards in the same frequency band apparently agreed with Marconi's observations well beyond the horizon in 1932 and 1933, radio science generally came to rely instead on calculations unchecked by experiment and to delay to the last few years the exploitation of Marconi's vision, heedless of the warning so well phrased in his 1932 paper: "Long experience has, however, taught me not always to believe in the limitations indicated by purely theoretical considerations, or even by calculations. These—as we well know—are often based on insufficient knowledge of all the relevant factors. I believe, in spite of adverse forecasts, in trying new lines of research, however unpromising they may seem at first sight."

While the current controversy over the explanation of twilight region propagation still rages, the importance of checking theory by experiment and vice versa is a lesson retaught by this bit of radio history. Considering the enormous amount of effort which has gone into research and production of microwave systems since 1933, certainly we should ask ourselves how these last two papers on microwaves by so successful a discoverer as Marconi could go practically unread, even granting that one was published in a foreign language and the other in an inaccessible journal. Renewed scholarly attention to this phase of Marconi's career seems unquestionably to be called for, with belated homage to the last neglected discovery of a great radio pioneer, which seems to me to illustrate, in Debye's memorable phrase[5] that: ". . . our science is essentially an art which could not live without the occasional flash of genius in the mind of some sensitive man, who, alive to the smallest of indications, knows the truth before he has the proof." The practical consequences of this last propagation discovery of Marconi's could well rival in importance his discoveries of transatlantic wireless and the short wave revolution.

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[3] W. D. Hershberger, "Seventy-five centimeter radio communication tests," PROC. IRE, vol. 22, pp. 870–877; July, 1934.

[4] B. Trevor and R. W. George, "Notes on propagation at a wavelength of seventy-three centimeters," PROC. IRE, vol. 23, pp. 461–469; May, 1935.

[5] "The Collected Papers of Peter J. W. Debye," Interscience Publishers, Inc., New York, N. Y., p. xxi, 1954.

*On the Propagation of Microwaves over Considerable Distances**

(Professor R. M. Fano of the Department of Electrical Engineering and Research Laboratory of Electronics, M.I.T., has made the following translation into English of the Italian original "Sulla Propagazione Di Micro-Onde A Notevole Distanza," as it appeared in the volume, "Scritti di Guglielmo Marconi," "Collected Papers of Guglielmo Marconi," published by the Royal Academy of Italy, Rome, 1941, pp. 447–449.)

Electromagnetic waves of wavelengths smaller than one meter are commonly known as microwaves; they are also called quasi-optical waves because it was generally believed that with them radio-telegraphic communication would have been possible only when the transmitting equipment and the receiving equipment were within line-of-sight: their practical utility would then have been limited by such restriction.

In the course of experiments carried out in the months of July and August of last year, I was able to discover that the useful range of these waves was not at all limited to the optical distance—depending, in the main, on the height of the equipment—but, that these waves could be received and detected beyond the horizon up to a distance approximately twice the optical one and also between stations shielded from each other by hills.[1]

Between the second and the sixth of this month I was able to carry out further radio-telegraphic and radio-telephonic transmission tests by means of microwaves of approximately 60 centimeter wavelength (500 megacycles) from a transmitter located at Santa Margherita Ligure and a receiver installed on the yacht Elettra which was sailing along the coast of the Tirreno.

The transmitting dipole which was radiating approximately 25 watts was located on the hotel Miramare at Santa Margherita at a height of 38 meters above sea level and was placed near the focus of a parabolic reflector having an aperture of 2 meters.

The receiving dipole was in a similar reflector located on the yacht Elettra at a height of 5 meters above sea level.

In spite of the fact that the optical distance was only 30 kilometers, the radio-telephonic and radio-telegraphic signals sent by the transmitting station were received on the yacht with clarity, great strength, and regularity at a distance of 150 kilometers, that is, at five times the optical distance; during the last year's tests, on the other hand, although the height above sea level of the transmitter at Santa Margherita was greater (50 meters), the maximum distance at which Morse signals were detected feebly was 52 kilometers.

It was not possible during these tests to make continuous observations beyond the above mentioned distance of 150 kilometers because the navigation requirements specified by the configuration of the coast did not permit maintaining the reflector of the Elettra pointing at all times toward the transmitting station. The Morse signals were nevertheless detected very feebly and with slight fading, but often legible, up to the mooring at St. Stefano Harbor, at a distance of 258 kilometers from Santa Margherita—that is, at almost nine times the optical distance—although, in this case, on the straight path joining the two stations there was land broken by high hills for almost 17 kilometers: the promontory of Piombino for 11.482 kilometers and the Punta Troja for 5.566 kilometers.

The greater range obtained in these experiments seemed due to the improved efficiency of the transmitting and receiving equipment and of the reflectors employed.

In these experiments, as in those of last year, I was effectively assisted by Mr. G. A. Matthieu who has personally taken care of the construction and of the initial tests of the new equipment and also by technicians of the Marconi Company.

The theoretical explanation of the results obtained when the wavelength employed is taken into account presents, in my view, serious difficulties even when using the calculations involving diffraction and refraction indicated by Pession in his paper "Considerations on the Propagation of Ultra-Short Waves and of Mircowaves."[2]

The speculations that may arise from such results concern the entire theory of radio transmission over distances greater than the optical one.

After further, more complete, and more extended experiments, I am planning to publish a detailed paper on the methods employed and on the results obtained and I express the hope that in addition to theoretical speculations that might be of scientific interest, the present results will lead to new and substantial progress in the field of radio communications.

* Roy. Acad. of Italy, "Proceedings of the section on physical, mathematical and natural sciences," vol. IV, 1933; paper no. 16, presented at the special meeting of August 14, 1933.

[1] Guglielmo Marconi, paper presented on December 2 1932 at the Roy. Inst. of Gr. Brit., London.

[2] G. Pession, "Alta frequenza," vol. I, no. 4; December, 1932.

The Statistics of Combiner Diversity*

Several papers[1–3] have recently appeared pointing out the advantages of combiner diversity over switch diversity both from a theoretical and practical point of view. Mack,[3] using numerical techniques, presented curves of the statistical distribution of the combined signal for dual and triple

* Received by the IRE, May 23, 1956.

[1] L. R. Kahn, "Ratio squarer," PROC. IRE, vol. 42, p. 1704, November, 1954.

[2] D. G. Brennan, "On the maximum signal-to-noise realizable from several noisy signals," PROC. IRE, vol. 43, p. 1530; October, 1955.

[3] C. L. Mack, "Diversity reception in uhf long-range communications," PROC. IRE, vol. 43, pp. 1281–1289; October, 1955.

diversity. It is the purpose of this letter to point out that an analytic evaluation of the combined statistical distribution is available in terms of a well tabulated function and to present the statistics of the combined signal up to ten-fold diversity. This author has already participated in a systems engineering study utilizing quadruple diversity and feels that even higher order diversity might be used for future longer-range tropospheric scatter circuits.

We begin with Brennan's[2] generalized formula that the optimum combination of many signals provides

$$P = \sum_{i=1}^{n} p_i \tag{1}$$

where P=combined output S/N power ratio, p_i=output S/N power ratio of receiver i, and n=order of diversity. It is usually assumed that the individual S/N ratios are Rayleigh distributed whose statistical distribution is of the form

$$\frac{R}{\psi_0} e^{-R^2/2\psi_0} dR. \tag{2}$$

Actually, it is the voltage S/N ratio which is Rayleigh distributed while the power S/N ratio is exponentially distributed. Since p is proportional to R^2, its statistical distribution is of the form

$$\frac{1}{p_0} e^{-p/p_0} dp. \tag{3}$$

If each of the individual p_i have the statistical distribution given by (3), then $P = \sum_{i=1}^{n} p_i$ has the distribution (assuming that the p_i are independent)

$$\frac{1}{(n-1)!} \left(\frac{P}{p_0}\right)^{n-1} e^{-P/p_0} \frac{dP}{p_0}. \tag{4}$$

This is well-known and is discussed in almost any mathematical statistics book under the topic Chi-Square distribution. To the radar engineer, the same distribution is known as the output of a square law detector and integrator. It should be quite clear that p_0 is simply related to the median of the distribution of (3). Since it is convenient to present our data relative to the median of (3), it is only necessary to evaluate the cumulative probability distribution of $y = p/p_0$. The probability that y will exceed any particular value, y_0, is given by

$$\frac{1}{(n-1)!} \int_{y_0}^{\infty} y^{n-1} e^{-y} dy = \left[1 + y_0 + \frac{1}{2!} y_0^2 + \frac{1}{3!} y_0^3 + \cdots + \frac{1}{(n-1)!} y_0^{n-1}\right] e^{-y}. \tag{5}$$

It is also well-tabulated as the incomplete gamma function.[4] Eq. (5) for n=1, 2, 3, 4, 6, 8, 10 is shown in Fig. 1. When compared with Mack's estimates for dual and triple combiner diversity, differences of about one or two db between our estimate and his appears. This presumably is due to the difference between our exact integration in closed form and his numerical integration.

In designing a scatter circuit (either tropospheric or ionospheric) one must be able to override the expected fading range to some predetermined degree of reliability (*e.g.*, 99 or 99.9 per cent). In another paper,[5] curves are presented for system gain for various orders of diversity and different degrees of reliability. Unfortunately, the curves there assumed that only independent Rayleigh fading occurred at all receivers. As is quite well known this is not true. There are, to a fair degree of approximation, two types of fading. One is referred to as slow fading which describes the variability of hourly medians throughout the whole year and which is statistically described by a db Gaussian distribution. An estimate of this variability on a tropospheric scatter link can be obtained from Fig. 3 of a Lincoln Laboratory paper.[6] The second type of fading is referred to as fast fading and its statistical distribution is described in this letter for various orders of diversity. Diversity action has two effects. It reduces the median scatter loss by the increase of the median signal and it reduces the range of the fast fading only. Therefore, to estimate the required power for a scatter link requiring 99 per cent reliability and utilizing quadruple diversity, we may proceed as follows. We decrease the yearly median scatter loss estimate obtained from the Mellen paper[6] by 7.2 db (as shown in Fig. 1). The total fading range

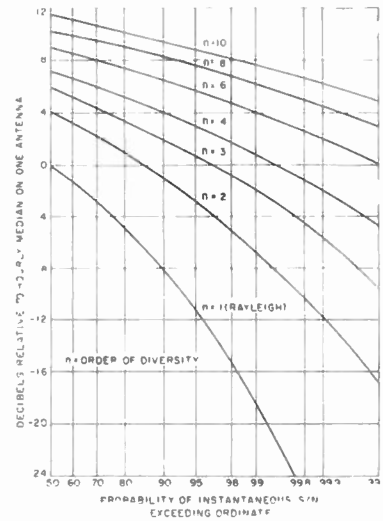

Fig. 1.

to be expected can be taken as the root sum square of the fast fades (6.8 db in this example) and the slow fades as shown by Mellen (this is a function of distance). The procedure just described would be exact if both the fast fades and the slow fades were db Gaussian. For diversity systems even the fast fades seem to become approximately db Gaussian which would be a straight line on the accompanying figure.

It should be pointed out, however, that for an fm system, the fast fade statistics are really a cross between the combiner statistics just discussed and the statistics appropriate to switch diversity which is already well understood by systems engineers. The reason for this lies in the fm threshold effect in that combiner diversity does not operate effectively on signals that are below the fm threshold. Therefore, the statistics of combiner diversity of an fm system are a function of median signal level. A complete analysis of the statistics appropriate to an fm system is a very interesting mathematical statistics problem; however, from a practical point of view, such a complex analysis is not justified at the present state of the art.

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[4] "Tables of the Incomplete Gamma Function" edited by K. Pearson, Cambridge Univ. Press, Cambridge, Eng.; 1946.

[5] F. J. Altman and W. Sichak, "A simplified diversity communication system for beyond-the-horizon links," IRE TRANS., vol. CS-4, pp. 50–56; March, 1956.

[6] G. L. Mellen, W. E. Morrow, Jr., A. J. Poté, W. H. Radford, and J. B. Wiesner, "UHF long-range communication systems," PROC. IRE, vol. 43, pp. 1269–1281; October, 1955.

A Note Concerning the Dirac Delta Function*

In most discussions of the impulse or Dirac Delta function, the authors give several different definitions of the function which they claim to be equivalent.[1]

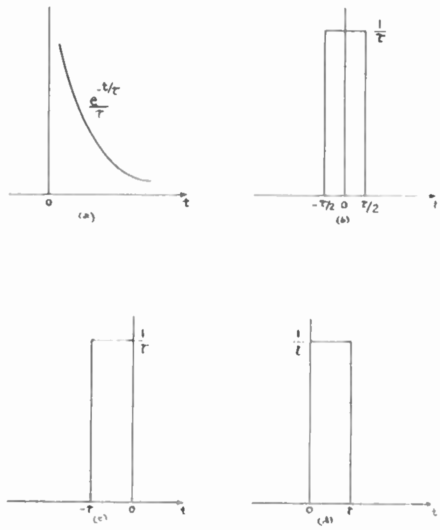

Fig. 1.

Usually the forms given are as shown in Fig. 1 and are defined as:

$$\text{(a)}\quad \delta(t) = \lim_{\tau\to 0} \frac{e^{-t/\tau}}{\tau} \qquad t > 0$$
$$= 0 \qquad t \leq 0$$

$$\text{(b)}\quad \delta(t) = \lim_{\tau\to 0} \frac{U(t+\tau/2) - U(t-\tau/2)}{\tau}$$

$$\text{(c)}\quad \delta(t) = \lim_{\tau\to 0} \frac{U(t+\tau) - U(t)}{\tau}$$

$$\text{(d)}\quad \delta(t) = \lim_{\tau\to 0} \frac{U(t) - U(t-\tau)}{\tau}.$$

It can be shown however that the Laplace transforms of the above functions are

* Received by the IRE, May 21, 1956.

[1] See: S. Goldman, "Transformation Calculus and Electrical Transients," Prentice-Hall, Inc., New York, N. Y.; 1953.

different and hence that the definitions are not equivalent, if it is remembered that in dealing with the δ function the limiting process should be carried out after all other manipulations.[2]

Consider the definition given in (a) above. The Laplace transform is written:

$$L[\delta(t)] = \lim_{\tau\to 0} \int_0^\infty \frac{e^{-t/\tau}}{\tau} e^{-st}dt$$

$$= \lim_{\tau\to 0} \lim_{\substack{T\to\infty \\ a\to 0}} \frac{1}{\tau} \int_a^T e^{-(s+1/\tau)t}dt$$

$$= \lim_{\tau\to 0} \lim_{\substack{T\to\infty \\ a\to 0}} \frac{-1}{\tau\left(s+\dfrac{1}{\tau}\right)} e^{-(s+1/\tau)t}\Big|_a^T$$

$$= \lim_{\tau\to 0} \frac{1}{\tau s + 1}(1-0) = 1$$

which is the value normally given.

The other three cases are specific examples of the general case in which:

$$\delta(t) = \lim_{\tau\to 0} \frac{U[t+(1-\rho)\tau]-U(t-\rho\tau)}{\tau},$$

where $0 \leqq \rho \leqq 1$, the equalities holding for cases (c) and (d) and $\rho = 1/2$ for case (b). The Laplace transform then becomes:

$$L[\delta(t)] = \lim_{\tau\to 0} \frac{1}{\tau} \int_0^\infty \{U[t+(1-\rho)\tau] - U(t-\rho\tau)\} e^{-st}dt$$

$$= \lim_{\tau\to 0} \frac{1}{\tau}\left\{ \int_0^\infty U[t+(1-\rho)\tau]e^{-st}dt - \int_0^\infty U(t-\rho\tau)e^{-st}dt \right\}$$

$$= \lim_{\tau\to 0} \frac{1}{\tau}\left\{ \int_0^\infty e^{-st}dt - \int_{\rho\tau}^\infty e^{-st}dt \right\}$$

since

$$U[t+(1-\rho)\tau] = 1 \text{ for } 0 \leq t$$

and

$$U(t-\rho\tau) = 0 \quad \text{for} \quad t < \rho\tau.$$

Thence,

$$L[\delta(t)] = \lim_{\tau\to 0} \frac{1}{\tau} \int_0^{\rho\tau} e^{-st}dt = \rho$$

which can take on any value from 0 to 1 inclusive.

The symmetrical definition of case (b) has a Laplace transform of $\frac{1}{2}$, a fact which has been used by Spiegal[3] to evaluate certain integrals.

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[2] H. Weyl, "The Theory of Groups and Quantum Mechanics," E. P. Dutton and Co., Inc., New York, N. Y., 1931.
T. Lewis, "Some applications of the Dirac Delta function;" *Phil. Mag. Ser. 7*, vol. 24, p. 329; September, 1937.
[3] M. R. Spiegal, "Applications of the Dirac Delta function to the evaluation of certain integrals," *J. Appl. Phys.*, vol. 25, p. 1302 [see (20)]; October, 1954.

A Sensitive Method for the Measurement of Amplitude Linearity*

Virtually all electronic devices and components exhibit variations in their characteristics which vary with frequency and/or amplitude. The square wave generator is a useful tool for the measurement of frequency response but no comparable expedient is in general use for the measurement of amplitude characteristics. The method proposed below permits the rapid determination of the amplitude response of a system or device with considerable sensitivity and extremely simple apparatus. It consists of the application of a linear sawtooth waveform to the input terminals of the device under test and a differentiator through which the output is passed. The differentiated waveform is then observed on an oscilloscope as a function of the input, output, or other variable. If the device is linear over the dynamic range encompassed by the sawtooth, the derivative, with respect to time, will be a constant proportional to the slope of the input and the gain of the device. Any departure from linearity will result in a corresponding departure of the derivative from the horizontal which can be readily detected and measured. This method has the virtue of practically separating the linear and nonlinear components since the base line can be translated to the horizontal axis by adjustment of the oscilloscope centering control. The more conventional method of harmonic analysis shares this feature of separation (by means of filters) but does not provide an instantaneous display as a function of amplitude and is inherently not as sensitive, as will be shown.

The equipment required is shown in block form in Fig. 1. The sawtooth generator

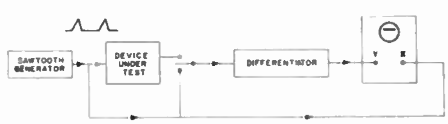

Fig. 1—Block diagram of system to measure linearity.

output consists of repetitive waveforms for convenient display. The amplitude must cover the dynamic range of interest and the linearity should be as high as possible; certainly high compared with the device to be tested. The low duty cycle which can be achieved offers an additional feature in permitting measurements to be made using peak amplitudes which might cause excessive dissipation under steady state conditions.

The differentiator may be of the electronic feedback type but for most applications a simple RC differentiator having a short time constant is satisfactory. The oscilloscope should have a dc amplifier and the sweep can be supplied by the sawtooth generator. Loading of the sawtooth generator by the device under test can be checked by examining the derivative of the input, with the load connected, and this comparison can be used as a basis for gain measurement.

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Since the concept of "degree of nonlinearity" is not as familiar as that of harmonic generation, a quantitative comparison is given below between the results obtained with a generalized nonlinear device, using the proposed method, and the method of harmonic analysis. Consider a nonlinear device whose characteristics can be expressed by the power series

$$y = Ax + Bx^2 + Cx^3 + Dx^4 + \cdots \quad (1)$$

where

$$y = \text{output}$$
$$x = \text{input.}$$

If the input signal is a ramp function then

$$x = kt \quad (2)$$

and

$$y = Akt + Bk^2t^2 + Ck^3t^3 + Dk^4t^4 + \cdots \quad (3)$$

and the derivative of the output with respect to t is

$$\frac{dy}{dt} = Ak + 2Bk^2t + 3Ck^3t^2 + 4Dk^4t^3 + \cdots. \quad (4)$$

A measure of the linearity can be made by comparing the magnitude of the first term with the remaining terms. This is done by taking the ratio of the departure from the initial value, at time T, to the initial value. (See Fig. 2.)

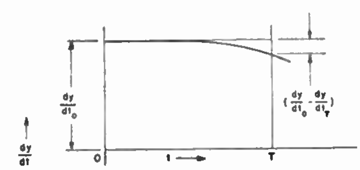

Fig. 2—Curve showing derivative as $f(t)$ with ramp input.

Thus

$$\frac{\dfrac{dy}{dt_0} - \dfrac{dy}{dt_T}}{\dfrac{dy}{dt_0}} = -\left(\frac{2Bk^2T + 3Ck^3T^2 + 4Dk^4T^3}{Ak} + \cdots\right) \quad (5)$$

where

$$\frac{dy}{dt_0} = \frac{dy}{dt} \text{ at } t = 0, \quad \frac{dy}{dt_T} = \frac{dy}{dt} \text{ at } t = T.$$

Eq. (5) shows that the total departure at $t = T$ is a weighted sum of the distortion components whose coefficients are B, C, D, etc. Theoretically, the magnitude of each coefficient can be computed by deriving the equation of the curve of Fig. 2. Practically, this becomes increasingly difficult as the order of the curvature increases and it is proposed that a "figure of demerit," which weighs the various distortion coefficients, be used as a measure of nonlinearity. The relation between this "figure of demerit" and intermodulation distortion should be apparent.

If a sinusoidal waveform, having a peak amplitude of kT, is introduced into the same device the input is given by

$$x = Kt \sin \omega t \tag{6}$$

and the output is

$$y = Akt \sin \omega t + Bk^2T^2 \sin^2 \omega + Ck^3T^3 \sin^3 \omega t + Dk^4T^4 \sin^4 \omega t + \cdots. \tag{7}$$

Expanding each term gives

$$y = AkT \sin \omega t + \frac{Bk^2T^2}{2}(1 - \cos 2\omega t) + \left(\frac{Ck^3T^3}{4} 3 \sin \omega t - \sin 3\omega t\right) + \frac{Dk^4T^4}{8}(3 - 4\cos 2\omega t + \cos 4\omega t) + \cdots. \tag{8}$$

A direct comparison with (5) can now be made on a term by term basis. Considering the second harmonic [in the second term of (8)], the ratio of the coefficient to that of the fundamental is

$$\frac{\text{Second harmonic}}{\text{fundamental}} = \frac{BkT}{2A}. \tag{9}$$

From (5), the departure from the initial value at $t=T$ divided by the initial value is

$$\frac{\dfrac{dy}{dt_0} - \dfrac{dy}{dt_T}}{\dfrac{dy}{dt_0}} = \frac{2BkT}{A} \tag{10}$$

for second order term only.

As an example, if only a second harmonic is present whose amplitude is 1 per cent of the fundamental, $Bkt/2A = 0.01$, then the departure of the sawtooth derivative from the horizontal at $t=T$ will be 4 per cent of the initial value, $2BkT/A = 0.04$. From actual operating data, it has been observed that a departure of 0.1 per cent is readily detected making it completely feasible to detect a second harmonic whose amplitude is only 0.025 per cent of the fundamental.

A comparison of the high order terms shows an even greater sensitivity advantage for this method. For the third and fourth terms the ratio of linearity departure to harmonic amplitude is 12 and 32, respectively.

Some approximations have been made in obtaining these figures, which include neglecting the contributions to the fundamental by the third term, and to the second harmonic by the fourth term, etc. (8). If it is desired to correlate the differentiated sawtooth signal with the distortion coefficients this can be done by a simple comparison between (1) and (4). Thus the ratios B/A, C/A, D/A, etc., are multiplied by 2, 3, 4, etc. A device having a 2 per cent second order nonlinearity ($B/A = 0.02$) would give a departure from the horizontal of 0.04 at $t=T$.

The discussion thus far has been limited to a nonlinear device which exhibits no frequency distortion. In practical applications of this method it is desirable to use a repetitive waveform and it is necessary to know the bandwidth requirements for the particular waveform selected. This bandwidth will be a function of the degree of precision which is sought.

A simple Fourier analysis of the repetitive sawtooth is not applicable because much of the harmonic content is associated with the sharp trailing edge which does not contribute to the measurement. Therefore, the analysis is made using a ramp function in conjunction with several elementary low- and high-pass filters. Consider first an elementary low-pass RC network whose transfer function is $1/(Ts+1)$ where $T=RC$ and the transform of the ramp function is k/s^2. Thus the transform of the output is

$$\mathfrak{L}e_{out} = \frac{k}{s^2(Ts + 1)} \tag{11}$$

and

$$e_{out} = k(t - T(1 - e^{-t/T})). \tag{12}$$

When $t \geqq 3.4\ T$, the departure from linearity of the sawtooth is less than 1 per cent. If the initial portion of the sweep is of interest, it is necessary to use a sweep of sufficient duration to permit the error to drop to the required limit during the early portion of the sweep.

For the corresponding high-pass network, the transfer function is given by $TS/(1+Ts)$ and the transform of the output is

$$\mathfrak{L}e_{out} = \frac{kTs}{s^2(1 + Ts)}, \tag{13}$$

and

$$e_{out} = kT(1 - e^{-t/T}). \tag{14}$$

Eq. (14) is similar to the familiar expression for the current in an RL circuit with a step function input. If $t = 0.02\ T$ the departure from linearity is less than 1 per cent.

This method, where applicable, provides directly the same information which is obtained indirectly by intermodulation or harmonic distortion measurements. In addition, the information is displayed as a function of amplitude and with extremely modest equipment. The bandwidth requirements will prove a drawback for some measurements.

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When is a Backward Wave Not a Backward Wave?*

The following remarks have reference to a number of studies of the large-signal (nonlinear) performance of traveling-wave amplifiers.[1–4] There seems to be some difference of opinion regarding the treatment of the transmission-line theory for the helix. The procedure of Nordsieck and Rowe is essentially to integrate numerically the second-order differential equation for the helix with the proper boundary conditions. Clearly the result is the complete and unique solution of the problem. Poulter and Tien prefer to start the numerical work from the general solution of the differential equation written in semiclosed form. While also formally correct, their procedure is rather involved and leads to some intuitive difficulties. Tien's claim that Rowe omits part of the solution is a direct consequence of this confusion.

The Poulter-Tien general solution referred to above contains an integral which can be considered as the convolution integral of the spatial impulse response of the helix with the current induced in the helix by the beam. The *impulse response* has two components, one of which can be interpreted as a forward wave, the other as a backward wave. However, to consider the corresponding *component integrals* as independent forward and backward waves, respectively, has very serious consequences.

Take a small-signal growing-wave solution as an illustration. Referred to the system *helix-beam* as a whole, this wave constitutes a single component of the complementary function of the differential equation, but referred to the *helix alone* this wave, according to the Poulter-Tien convolution terminology, breaks up into one "forward" and one "backward" wave with the *identical (forward) propagation constant*. Actually these two "waves" have no separate physical existence, and it is hard to see that their mathematical separation serves any useful purpose. Tien is stretching the concept "backward wave" to the point where it becomes very nearly useless. He is making one particular method of derivation rather than the actual space-time variation the criterion for "backward" and "forward" waves.

There is no contradiction in the circumstance that the differential equations for Rowe's $A(y)$ and Tien's $a_1(y)$ and $a_2(y)$ are different. The differential equation for $A(y)$ must necessarily be of the second order; its numerical solution must include all the components as required by the boundary conditions. Tien's variables, on the other hand, do not represent the complete solution but only his "forward wave"; they consequently satisfy first-order differential equations.

The Nordsieck-Rowe procedure is the straightforward approach to the numerical solution of a nonlinear problem. Since superposition of forward and backward waves does not hold for the system as a whole, it seems artificial and pointless to break up the phenomena on the linear part of the system into forward and backward waves, as Poulter and Tien do.

While the amplitude function $A(y)$ represents the complete solution, it should be noted that Rowe's (45) for the power along the helix

$$P \sim \left[\frac{\overline{V}^2(z, t)}{Z_0}\right]_{avg} = 2CI_0V_0A^2(y) \tag{1}$$

involves an approximation that was unfortunately not discussed in his paper.[2] We shall calculate here the magnitude of this approximation.

* Received by the IRE, May 28, 1956.

[1] A. Nordsieck, "Theory of the large-signal behavior of traveling-wave amplifiers," PROC. IRE, vol. 41, pp. 630–637; May, 1953.

[2] J. E. Rowe, "A large-signal analysis of the traveling-wave amplifier; theory and general results," TRANS. IRE, vol. ED-3, pp. 39–57; January, 1956.

[3] H. C. Poulter, "Large signal theory of the traveling-wave tube," Tech. Rep. No. 73, Electr. Res. Lab., Stanford Univ., Stanford, California; January, 1954

[4] P. K. Tien, "A large signal theory of traveling-wave amplifiers," *Bell Syst. Tech. J.*, vol. 35, pp. 349–374; March, 1956.

The total power on the helix in the forward direction is exactly

$$P = \tfrac{1}{2}\,\mathrm{Re}\,[V^*I]. \tag{2}$$

One of the two first-order transmission-line equations may be written as

$$\frac{C\omega}{u_0}\frac{\partial V(y,\phi)}{\partial y} + \frac{Z_0}{v_0}\frac{\partial I}{\partial \phi}\frac{\partial \phi}{\partial t} = 0 \tag{3}$$

where

$$y = \frac{C\omega}{u_0} z \tag{4}$$

$$\phi(y, \phi_0) = y/C - \theta(y) - \omega t \tag{5}$$

and

$$V(y, \phi) = \mathrm{Re}\left[\frac{Z_0 I_0}{C} A(y) e^{-j\phi}\right]. \tag{6}$$

When (5) and (6) are used in the transmission-line equation, (3), the current on the helix is given by

$$I = \frac{I_0}{1 + Cb}\left[\left(\frac{1}{C} - \frac{d\theta(y)}{dy}\right) A(y) e^{-j\phi} + j\frac{dA(Y)}{dy} e^{-j\phi}\right]. \tag{7}$$

Substituting (6) and (7) into (2) and simplifying gives for the power along the helix

$$P = 2CI_0V_0A^2(y)\,\frac{\left(1 - C\dfrac{d\theta(y)}{dy}\right)}{(1 + Cb)}. \tag{8}$$

The efficiency is then given by

$$\eta = \frac{P}{I_0V_0} = 2CA^2(y)\,\frac{\left(1 - C\dfrac{d\theta(y)}{dy}\right)}{(1 + Cb)}. \tag{9}$$

It should be noted that since $\theta(y)$ is a negative function the numerator as well as the denominator of the correction factor is always greater than unity. It was found that for all cases of interest the correction factor varies between zero and four per cent, justifying the use of the approximate relation (1).

A further insight into the equivalence of the differential-equation approach and the convolution-integral approach may be seen in Fig. 1 where Poulter's solution is compared with Rowe's. The case selected for comparison is one for which $QC = 0$ and large $C (C = 0.2)$ so that the effect of the so-called "backward wave" should be large.

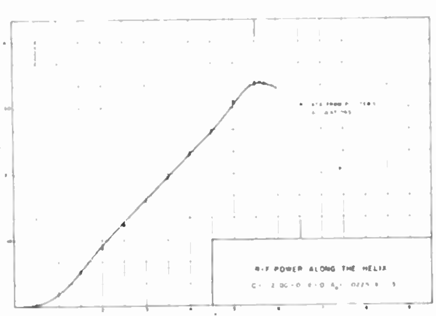

Fig. 1.

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The Noise Factor of Traveling-Wave Tubes*

Pierce, Watkins, Bloom, Peter, and others, have presented various approximate figures for the minimum value of the noise-factor of traveling-wave tubes with an infinite confining magnetic field. The purpose of this letter is to direct attention to the fact that no obvious theoretical lower limit for the noise factor is found when a simple mathematical model with different means for shaping the beam is chosen.

We shall in this discussion assume a positive space charge with infinite mass and with a density exactly neutralizing the dc negative charge of the beam. It appears that any way of shaping the beam that is based on balancing out the dc lateral space-charge forces rather than on brute-force suppression of all lateral motion will give substantially the same result. Other examples of such methods are the E-beam flow and the Brillouin flow. The beam is treated as a continuous fluid transmitting shot noise and velocity fluctuations from a cathode.

Inside the beam, where the dc velocity and space-charge density are uniform and constant, the Maxwell-Lorentz equations have two classes of small-signal solutions.

1) "Space-charge Waves," or "Plasma Waves." In a reference frame moving with the beam velocity these waves are irrotational; the total current density is everywhere zero. The divergence of the rf electric field is *not* in general zero. The shot noise generated by thermionic emission is primarily of this nature.

2) Solenoidal Waves (div $E = 0$).

If a step-function beam edge is postulated and Hahn's method for satisfying the boundary conditions is used the following results are obtained. A space-charge wave with components of its rf velocity and electric field perpendicular to the boundary surface produces a ripple along the edge of the beam. The equivalent surface charge of this ripple is equal and opposite to the normal component of the displacement vector. As far as the velocity and electric-field components parallel to the beam edge are concerned, the boundary conditions require the addition of a solenoidal surface wave inside the beam edge. When the resulting additional ripple is accounted for, it is found that the field external to the beam is zero. The space-charge waves as defined above simply do not interact with external fields.[1] For an ideal model of a traveling-wave tube using the kind of beam discussed here, noise waves of the first class produce no noise in the external circuit.

The study of the matching of the external fields by solenoidal beam waves leads to a theory of traveling-wave amplification very closely analogous to the conventional theory. In this theory, the ripple of the beam boundary plays the same part as the space-charge density fluctuations do in the confining-field type of tubes.

Since the beam in this 0-db noise-factor model of a traveling-wave tube is in a state very far from thermal equilibrium, there is no thermodynamic paradox in the behavior just described.

The important question is of course what obstacles may impede the approximate realization of this noise factor. The three idealizations on which the analysis is based are: homogeneous fluid; ideal flow ($\vec{u}_0 =$ const, $\rho_0 =$ const); step-function beam edge.

The first assumption probably does not produce an appreciable error in the analysis. Most likely the second point is the crucial one in a Brillouin-flow or E-beam tube. It has been established experimentally that the space-charge distribution in a cutoff dc magnetron is very far from the Brillouin solution; by analogy it may be concluded that a beam will also tend to assume a different distribution. However, the means available for control of the space-charge distribution in an electron gun is far superior, and once the desired flow pattern has been established, it can be expected not to change materially during the relatively short time of transit through the tube. For the same reason the third point will probably not cause any additional difficulty; a simple collimation scheme will most likely give a sufficiently sharp beam edge. However, everything else being ideal, a small but finite limit for the noise factor is certainly set by the blurring of the beam edge by the thermal velocities of the electrons. In the Maser, which has been hailed as the 0-db noise-factor microwave amplifier, the beam entropy is also growing and a small amount of noise generated; even if the *mean* life of the active energy state of the molecules is considerably longer than the transit time through the cavity, a small number of spontaneous transitions will necessarily occur. As long as this additional noise is small compared to the thermal noise in the input circuit, it will have a negligible effect on the noise factor.

It is conceivable that also solenoidal noise waves are generated in the electron gun, but their phase velocity is so high that the gun itself or a short subsequent drift tube can be designed as a waveguide below cutoff and serve as an effective attenuator of these waves.

On the basis of some theoretical arguments and rough estimates of the consequences of the oversimplification involved, the prospects of improving the noise factor of a traveling-wave tube consequently appear rather promising.

Once the strong coupling is removed between external "circuit waves" and the noise "space-charge waves" there seems to be no reason why the conventional traveling-wave tube should not be able to hold its own in the competition with the so-called transverse-field tube as far as noise factor is concerned.

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* Received by the IRE, April 20, 1956.

[1] In their original study of plasma oscillations, Tonks and Langmuir pointed out the difficulty of observing the oscillations because of the absence of external fields.

Geophysical Prospection of Underground Water in the Desert by Means of Electromagnetic Interference Fringes*

In his above paper[1] Prof. El-Said writes: "The author wishes also to acknowledge the early work by Dr. H. Löwy under the supervision of Prof. H. M. Mahmoud, Professor of Electrical Engineering, Cairo University, for the determination of the degree of transparency of desert rocks to electromagnetic waves."

To this I note that my early work concerning the electric transparency of desert rocks begins (without "supervision") in the year 1911, with my paper in the *Annalen der Physik*, vol. 36, p. 125. It is the first systematic treatment of the problem, as can be seen from the references of R. L. Smith Rose.[2]

Concerning the method by means of which he succeeded to measure groundwater depths of about 800m, Prof. El-Said quotes Appleton and Barnett, *Nature*, vol. 115, p. 333; 1925. Missing here is the reference to another paper of mine, published in 1912 in the *Physikalische Zeitschrift*, vol. 13, p. 397, concerning the ground-water interference experiments which I made in German potash mines. Reference is missing also to my recent paper in the *Bulletin de l'Institut d'Egypte*, vol. 35, p. 103, 1953, in which I have developed the theory used by Prof. El-Said for calculating the groundwater depth.

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* Received by the IRE, May 8, 1956.

[1] Proc. IRE, vol. 44, pp. 24–30; January 1956.

[2] *Jour. IEE*, vol. 75, p. 221; 1934.

Increasing the Accuracy of CRO Measurements*

We have developed several convenient ways of increasing the accuracy of measurements made on a cathode ray oscillograph. Because of deflection nonlinearities, curvature of the face plates, the time lag between measurement and calibration, and other factors, amplitude measurements made on the face of a cathode ray oscillograph are, as a rough rule of thumb, considered accurate to ±5 per cent. With the techniques to be described accuracies of ±0.1 per cent are possible.

An essential feature of the technique is that the oscillograph is used as a null detector. The unknown and a standard voltage are displayed simultaneously and the standard is adjusted until it equals the unknown. The standard is usually a dc voltage obtained from a battery and checked occasionally against a standard cell. A precision attenuator is used with the standard. The simultaneous display is obtained by means of a relay, usually operated directly from the 60 cycle line, which alternately connects the standard and unknown to the input of the oscillograph. Electronic switching means could also be used. One observes the unknown waveform and a straight line due to the standard. It is simple to adjust the standard until it coincides with unknown waveform at the point of measurement. With instruments that have dc amplifiers which have a wide dynamic range, such as the DuMont type 304, the gain may be increased so that the zero is far off scale and high sensitivity and repeatability in setting the null are obtained. With that instrument the gain can be adjusted so that the pattern has a height six times the tube diameter and the centering controls can be adjusted so that any portion of the waveform can be seen. On other instruments the same effect can be obtained by introducing a bucking voltage in series with both the unknown and the standard.

Details of the application of the principle vary with the particular situation and there is considerable room for exercise of ingenuity. Several other examples will be given:

1) In measuring 60 cycle hysteresis loops[1] it is desirable to measure the peak drive current. Distortion in the waveform which may be caused by the nonlinear core can make meter measurements inaccurate. In this case we switch both the X and Y axes of the oscillograph. During one period the loop is displayed. During the next 1/120 second the X axis is switched to the standard and the Y axis to 3 kc relaxation oscillator. One sees the right half of a hysteresis loop and a vertical line. The line is adjusted until it touches the tip of the loop.
2) In the application above it is necessary to measure the coercive force of the material. Here the hysteresis loop is reflected about the Y axis by passing the X axis signal through a full wave rectifier. Now on alternate 1/120 second periods one displays the right half of the hysteresis loop and a straight line, which is obtained by shorting the Y axis and passing the X-axis signal through a precision 10-turn window reading potentiometer. The potentiometer is adjusted until the line touches the side of the loop. The potentiometer than reads the ratio of the coercive force to the peak magnetic field.
3) To measure the gain of an amplifier to a repetitive signal the output and input could be displayed simultaneously. In many cases it is most convenient to obtain the input from a standard attenuator which is connected to a signal generator. The attenuator is adjusted until its attenuation is equal to the gain of the amplifier. With both traces present distortion is quite evident.
4) To measure the percentage droop in pulse transmission the synchronization of the oscillograph could be switched so that the leading edge of the input lined up with the trailing edge of the output when input and output are simultaneously displayed. The input is sent through a precision attenuator which is adjusted until the leading and trailing edges are equal in amplitude.

The possibilities of the technique should be apparent. Its advantages are:

1) The deflections for both measurement and calibration are identical, eliminating errors due to lack of linearity in the system.
2) Measurement and calibration are made simultaneously. The possibility of error due to drift in gain or unintentional change in settings is eliminated.
3) The accuracy can be made dependent on a dc source and a calibrated attenuator, two components which can readily be made accurate and stable.

William J. Bartik and Fred Bernstein contributed to these techniques. We have observed a pulse amplitude calibration method similar to the one described here in use at the Digital Computer Laboratory, Massachusetts Institute of Technology.

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* Received by the IRE, May 31, 1956.

[1] F. Bernstein, T. H. Bonn, and R. D. Torrey, "A sensitive single-turn hysteresis-loop tester," presented at the Special Magnetic Amplifier Conf., Syracuse, N.Y.; April 5, 1956.

Optimum Slicing Level in a Noisy Binary Channel*

Suppose that a binary signal, consisting of the familar "on-off" type in radio telegraphy, is transmitted over a noisy channel. At the receiver a very small amplitude slice of the envelope is selected and amplified, as is shown in Fig. 1.

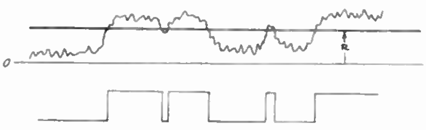

Fig. 1.

If the sine wave frequency is located at the center of a flat noise spectrum of width β, the expected number N_R of crossings with positive slope at the level R by the envelope is given by Rice[1] as

$$N_R = \left(\frac{\pi}{6b_0}\right)^{1/2} \beta R I_0\left(\frac{RQ}{b_0}\right) \cdot \exp\left(-\frac{R^2+Q^2}{2b_0}\right) \quad (1)$$

* Received by the IRE, April 11, 1956; revised manuscript received May 28, 1956.

[1] S. O. Rice, "Statistical properties of a sine wave plus random noise," *Bell Syst. Tech. J.*, vol. 27, pp. 125–126; January, 1948.

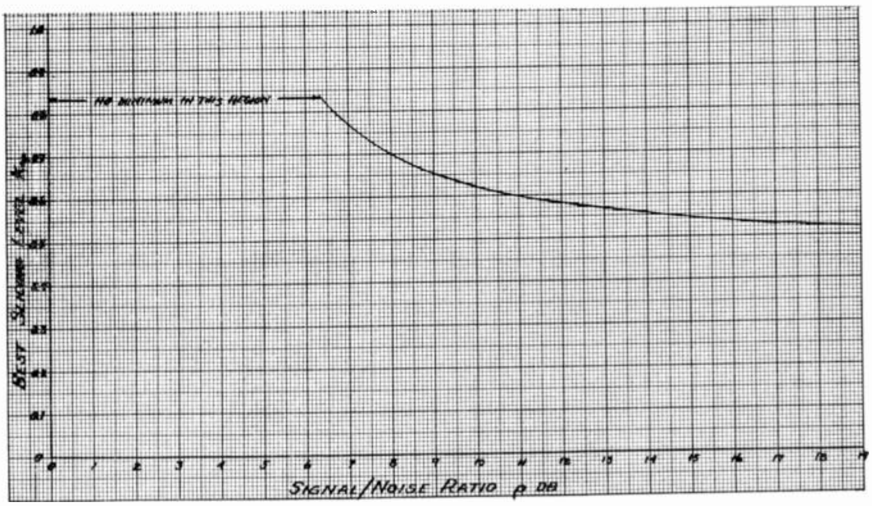

Fig. 2—Best slicing level for various signal/noise ratios.

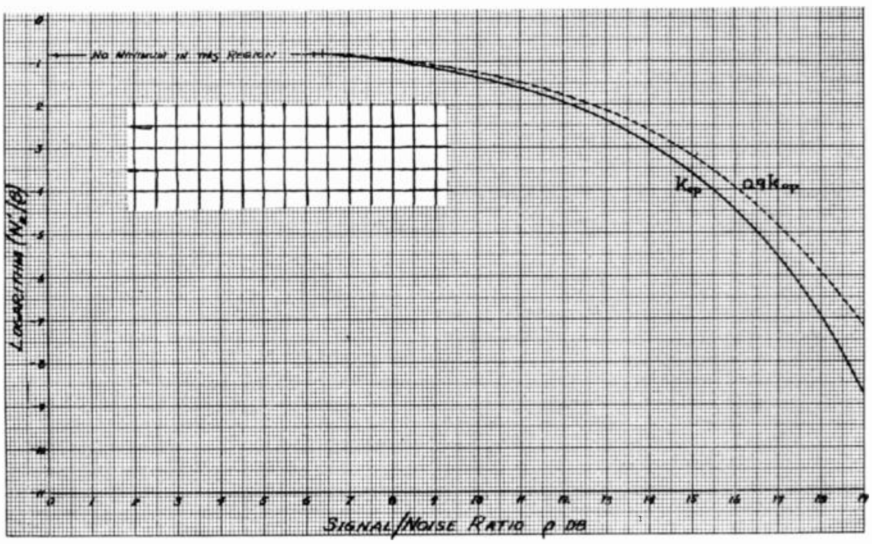

Fig. 3—Common logarithm of number of expected positive crossings per second per cycle bandwidth.

where b_0 is the mean noise power and Q is the peak amplitude of the sine wave. In many cases of interest the total duration of "signal" about equals that of "no signal." The crossings that are considered herein are only those caused by noise and are not due to the changes in the signal amplitude. Assuming this to be true the expected number of positive crossings is

$$N_{R}' = \left(\frac{\pi}{6b_0}\right)^{1/2} \cdot \frac{\beta}{2} \cdot R \exp\left(-\frac{R^2}{2b_0}\right) \cdot \left\{1 + I_0\left(\frac{RQ}{b_0}\right) \exp\left(-\frac{Q^2}{2b_0}\right)\right\}. \qquad (2a)$$

We set the slicing level R at some fraction $k(0<k<1)$ of the peak signal amplitude Q so that $R=kQ$ and let ρ equal the power signal/noise ratio $Q^2/2b_0$. Eq. (2a) becomes

$$N_{R'} = \left(\frac{\pi}{3}\right)^{1/2} \cdot \frac{\beta}{2} \cdot k\sqrt{\rho} \exp(-k^2\rho) \cdot \{1 + I_0(2k\rho) \exp(-\rho)\}. \qquad (2b)$$

The derivative of $N_{R'}$ with respect to k, when set equal to zero, yields the necessary condition which stationary points must obey. This condition is

$$2k \exp(-\rho) I_1(2k\rho) = \left(2k^2 - \frac{1}{\rho}\right) \cdot \{1 + I_0(2k\rho) \exp(-\rho)\}. \qquad (3)$$

The best slicing level k_{op} occurs when the expected number of crossings is a minimum. For ρ less than about 6.4 db the curve (2b) has no minimum over the range of k of interest. For values of ρ up to 19 db, this optimum level has been computed and is plotted in Fig. 2. As ρ increases indefinitely, (3) shows that k_{op} approaches 1/2 as one would expect. Fig. 3 is a graph of the logarithm of the minimum number of expected crossings, corresponding to the values of k_{op}, against the signal/noise ratio. For large ρ the curve varies like the function -0.11ρ (ρ a ratio). For a nonoptimum slicing level ($k=0.9\ k_{op}$) the logarithm of the expected crossing rate is shown as a dashed curve. A value of $k=1.1\ k_{op}$ yields closely the same curve.

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Power Transfer in Double-Tuned Coupling Networks*

In tuned amplifiers using amplifying elements with finite power gain (such as transistors) the interstage coupling network must, in addition to the required selectivity, provide optimum power transfer from driving to driven stage.

The design of single-tuned and transitionally coupled double-tuned networks for tuned transistor amplifiers has been described elsewhere.[1] In particular, it is known that good power transfer and narrow bandwidth are conflicting requirements which can be satisfied only by the use of inductances having very high *unloaded* Q's. The purpose of this note is to present considerations applicable to the general double-tuned case (with inductive coupling).

In the circuit of Fig. 1 transistor T_1 can be represented by a current source I_1 and its output admittance (g_0 and C_0) and the second transistor T_2 by its input admittance (g_i and C_i), resulting in the equivalent circuit of Fig. 2. g_1 and g_2 are the parallel loss conductances of the transformer windings. Lumping the capacitances and conductances together, we obtain the circuit of Fig. 3.

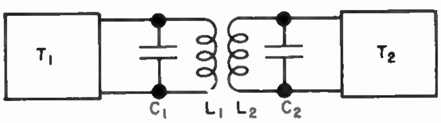

Fig. 1—Transistor amplifier stages coupled by double-tuned network.

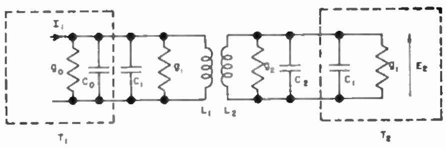

Fig. 2—Equivalent of Fig. 1.

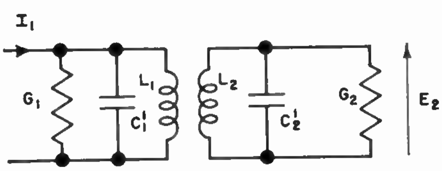

Fig. 3—Simplified equivalent of Fig. 1.

If Q_0 is the *unloaded* Q of the transformer windings (primary and secondary unloaded Q's are assumed equal) the *loaded* Q's are:

$$Q_1 = Q_0 g_1/G_1 = Q_0/x \qquad (1)$$

$$Q_2 = Q_0 g_2/G_2 = Q_0/y. \qquad (2)$$

The absolute value of the transfer impedance z_{21} is given by[2,3]

$$z_{21} = \frac{s}{(1+s^2)\sqrt{G_1 G_2}} \times \frac{1}{\sqrt{1 - \alpha v^2/(1+s^2) + v^4/(1+s^2)^2}} \qquad (3)$$

* Received by the IRE, March 20, 1956.

[1] R. R. Webster, "How to design IF transistor transformers," *Electronics*, vol. 28, pp. 156–160; August, 1955.

[2] C. B. Aiken, "Two mesh-tuned coupled circuit filters," Proc. IRE, vol. 25, pp. 230–272; February, 1937.

[3] G. E. Valley and H. Wallman, "Vacuum Tube Amplifiers," McGraw-Hill Book Co., Inc., New York, N.Y.; 1948.

with

$$\alpha = \frac{2(s^2 - b/2)}{1 + s^2}. \quad (4)$$

The symbols in (3) and (4) have the following significance: $s = k\sqrt{Q_1Q_2}$, k being the coupling coefficient; $b = Q_1/Q_2 + Q_2/Q_1$; $v = \sqrt{Q_1Q_2}\ (\omega/\omega_0 - \omega_0/\omega) \cong Q_1Q_2\ (2\Delta\omega/\omega_0)$, where $\Delta\omega = (\omega - \omega_0)$.

α can be called the *shape factor* of the coupled-tuned circuit response: $\alpha = 0$ corresponds to the case of transitional coupling (maximum flat response), whereas $\alpha < 0$ and $\alpha > 0$ are characteristic of the undercoupled and overcoupled cases respectively.

Using the above approximation for v, the fractional bandwidth[4] B_n is determined by

$$Q_1^2Q_2^2B_n^4/(1 + s^2)^2 - \alpha Q_1Q_2B_n^2/(1 + s^2) - n = 0. \quad (5)$$

At the center frequency, the power consumed by the load g_i is

$$P_d = E_2^2g_i = |I_1|^2[s^2/(1 + s^2)^2](g_i/G_1G_2) \quad (6)$$

and dividing P_d by the available power $(|I_1|^2/4g_0)$, we find the power transfer efficiency from T_1 to T_2

$$\eta = 4[s^2/(1 + s^2)^2](g_0g_i/G_1G_2). \quad (7)$$

Using the variables x and y, (5) and (7) can be written:

$$F(x, y) = Q_0^4B_n^4 - [\alpha/(2 - \alpha)]Q_0^2B_n^2(x + y)^2 - [n/(2 - \alpha)^2](x + y)^4 = 0 \quad (8)$$

$$\eta = 4(2 - \alpha)(x^2 + y^2 + \alpha xy)(x - 1)(y - 1)/(x + y)^4. \quad (9)$$

The maximum value of η can be found by setting

$$\partial\eta/\partial x + \lambda\partial F/\partial x = 0 \quad (10)$$

$$\partial\eta/\partial y + \lambda\partial F/\partial y = 0. \quad (11)$$

This results in the simple condition for η_{max}

$$x = y. \quad (12)$$

Substituting (12) into (8) and solving for x(or y), we find

$$x = y = (Q_0B_n\sqrt{2 - \alpha}/2\sqrt[4]{n}) \times \sqrt{\sqrt{1 + \alpha^2/4n} - \alpha/2\sqrt{n}}. \quad (13)$$

With (13) the circuit elements and coupling coefficient required for maximum power transfer can be calculated

$$L_1 = (x - 1)/\omega_0Q_0g_0 \quad (14)$$

$$L_2 = (x - 1)/\omega_0Q_0g_i \quad (15)$$

$$k = (x/Q_0)\sqrt{(2 + \alpha)/2 - \alpha}. \quad (16)$$

In the case of *transitional coupling* $\alpha = 0$ and, therefore, (13) becomes

$$x = Q_0B_n/\sqrt{2}\sqrt[4]{n}. \quad (17)$$

Substituting (12) and (13) into (9) we find the maximum power transfer efficiency

$$\eta_{max} = (1 - \alpha^2/4)(1 - 1/x)^2 = (1 - \alpha^2/4) \times \left[1 - \frac{2\sqrt[4]{n}}{Q_0B_n\sqrt{2 - \alpha}\sqrt{\sqrt{1 + \alpha^2/4n} - \alpha/2\sqrt{n}}}\right]^2 \quad (18)$$

For *transitional coupling*:

$$\eta_{max} = (1 - \sqrt{2}\sqrt[4]{n}/Q_0B_n)^2. \quad (19)$$

In solving practical design problems n, B_n, and Q_0 are usually known. The shape of the response curve is often of minor importance and the question may arise: What value of α should be chosen for maximum power transfer efficiency?

The optimum α can be determined from (18) by setting $d\eta_{max}/d\alpha$ equal to zero. This procedure leads to computational difficulties and, in general, one can only state that maximum power transfer efficiency occurs for $\alpha < 0$, *i.e.*, in the case of undercoupled circuits. Transitionally coupled circuits lead to less than maximum power transfer efficiency. The loss in the case of transitional coupling is very small for large values of B_nQ_0 but significant if B_nQ_0 is small. The situation is illustrated by Fig. 4. Undercoupled circuits should be used for small values of B_nQ_0.

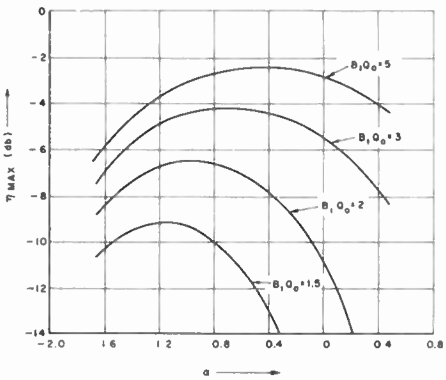

Fig. 4—η_{max} as a function of α with B_1Q_0 as parameter.

Acknowledgment

The subject of this note has been discussed with K. Fong. His valuable advice is gratefully acknowledged.

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[4] The bandwidth B_n' is defined as the interval between frequencies at which the power response is $1/(n+1)$ times the response at the center of the band. The fractional bandwidth B_n is $2\pi B_n'/\omega_0$.

Time Signals for the Determination of Longitude*

During the forthcoming International Geophysical Year, it is proposed to determine the relative positions of certain points on the earth's surface with increased precision.[1] At these points astronomical observations are to be made at times that are closely related to the readings of clocks at Greenwich.

This raises afresh the problem of defining simultaneity at distant points. Radio time signals are propagated from point-to-point at velocities that are less than that in free space, and that are not known with complete certainty.[2]

To judge by preliminary papers issued by the Bureau de l'Heure, the velocities that are considered normal are: 252,000 km per sec for frequencies of the order of 16 kc; 274,000 km per sec for frequencies of the order of 10 mc, by the short path, and 286,000 km per sec for frequencies of the order of 10 mc, by the long path. For single observations, the spread from these values may be considerable.

For a path such as that from Europe to New Zealand, which are approximately antipodal, the distinction between long and short path is by no means clear, and it may be difficult to tell whether a short-wave time signal has been 70 msec on the way at the *short path* velocity, or 67 msec at the *long path* velocity. Since OSAGI recommends that times should be established to within 1 msec for longitude determinations, and to "very high precision" for propagation studies,[3] the uncertainty of velocity is serious.

OSAGI suggests that the velocity of propagation may be investigated from signals initiated by a quartz clock at each end of a transmission path, so that each terminal station both transmits and receives a signal. The relative short-term instabilities of two crystal clocks enter into this experiment, and the measurement can be successful only if cooperation between distant points is uniformly good.

The experiment can be freed from all reliance on crystal clocks, and the radio measurements can be made entirely at one end of the transmission path if a respondor is installed at the remote end. In the simplest case, the respondor consists of a receiver which energizes a transmitter within microseconds of the receipt of a signal. The time between the transmission from the master end of the path and the receipt of the return signal is the time of a double traverse of the path. To a first approximation the midpoint of this interval is the time of arrival of the signal at the remote end of the path.

So simple a respondor is impractical except for short distances, but it may be replaced by existing communication receiving and transmitting stations, as shown in Fig. 1.

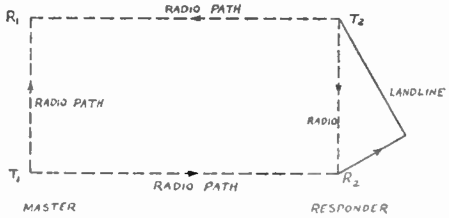

Fig. 1.

* Received by the IRE, April 3, 1956.

[1] 2nd Assembly of OSAGI, Rome, 1954, VIII, p. 1.

[2] See, for example, Mme. A. Stoyko, "Sur la variation de la vitesse de propagation des oudes radio electroniques," *Compt. Rend.* (Paris), vol. 232, pp. 1916–1920; May, 1951.

[3] 2nd Assembly of OSAGI, Rome, 1954, VIII, 2.4.1 and 2.4.4.

T_1 transmits short (1 msec or less) pulses which are received by R_2, some thousands of kilometers away, and are passed on to T_2 over an existing land line. T_2 retransmits at the same frequency as T_1 and is received at R_1, where the total travel time is measured. For the present purpose each transmitter-receiver pair should be as close as possible, but since existing stations are to be used, the existing distances (up to 100 or so km) must be accepted. The connection from R_2 to T_2 will normally be by a devious land line, and the travel time in it must be determined. There may also be instrumental delays in R_2 and T_2. If T_1 and T_2 operate at the same frequency the transmission from T_2 will be received by R_2 which will re-energize T_2, so leading to a train of pulses with a period of the total time of travel from R_2 to T_2 and back to R_2.

This train, received at R_1, will show how much of the total travel time is occupied between R_2 and T_2, and so will lead to the time taken for the two main radio paths $T_1 \cdots R_2$ and $T_2 \cdots R_1$. It will be expedient to interrupt the train of pulses after say 0.1 second. The times for the local radio transmissions, $T_2 \cdots R_2$ and $T_1 \cdots R_1$, will be so short that they may be computed. The times to be expected are:

For the main radio path (double journey) for intercontinental distances.	10–140 msec.
For the land line	up to 1 msec.
For the local radio paths	up to ½ msec.

High-power very low frequency transmitters are so rare that few responding circuits using them can be arranged, but the transmission time for these frequencies can be found by using short waves for the return path. If these measurements are made immediately after those described above, or, better still, interposed among a series of high frequency determinations, the difference between the total travel times will be the difference between the single transmission time at high frequency and by very low frequency. Since the shape of the pulse received at R_1 can be matched against the shape of the pulse transmitted from T_1, there should be little trouble from the slow rise-time of the very low frequency signal.

Measurements, of course, will need to be made to determine the variation of apparent velocity during the day, and during the year.

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Application of Equipartition Theory to Electric Circuits*

In the derivation of a theoretical formula for Johnson noise Nyquist[1] said: "To each degree of freedom there corresponds an energy equal to kT on the average, on the basis of the equipartition law, where k is the Boltzmann constant." This formulation could cause difficulty because the usual statement on equipartition is that with each degree of freedom there is associated *kinetic* energy of mean value *one half* of kT, and in lumped-circuit applications we do in fact find an energy of $\frac{1}{2}kT$ per degree of freedom.

Since equipartition is founded in the science of mechanics, we start with the definition that the number of degrees of freedom of a mechanical system is equal to the number of coordinates which must be specified in order to define the *position* of the system; but the full specification of a mechanical system requires in general a knowledge of *two coordinates* for each *degree of freedom, e.g.*, position and momentum as used in the Hamiltonian form of the equations of motion. The equipartition theorem then says that if there is associated with a *coordinate* an energy proportional to the square of that coordinate, the mean-square value of that energy will be $\frac{1}{2}kT$. If one of the coordinates is momentum and the mass is constant, the kinetic energy being proportional to the square of momentum must have the mean value $\frac{1}{2} kT$, as assumed above. The magnitude of any potential energy which may exist depends on the law of the field of force to which it is due, but in the special case of a harmonic oscillator the mean potential energy is equal to the mean kinetic energy. The total energy associated with the harmonic oscillator is thus kT, but any experiment in which one observes a single coordinate gives a direct measure of only $\frac{1}{2} kT$.

For example, in deriving from the Nyquist formula for noise in a narrow bandwidth the expected total (all-frequency) thermal energy in both a CR circuit and an LCR circuit, Moullin[2] arrived at a value of $\frac{1}{2} kT$. By applying the well-known technique of contour integration it can be shown[3] that this value of $\frac{1}{2} kT$ will be found for any two-terminal lumped network, however complicated its internal structure. Yet we can correctly assume the "available noise power" in a channel of bandwidth B to be kTB. This apparent paradox is eliminated if in Nyquist's formulation one replaces "degree of freedom" by "mode of oscillation," since a harmonic oscillator is known to have mean potential energy equal to its mean kinetic energy and hence to be entitled to a total of kT.

It remains to show how we reconcile $\frac{1}{2} kT$ total for a two-terminal lumped circuit with kT for a transmission line or distributed circuit. (A circuit having a truly flat response over bandwidth B is not physically realizable with lumped elements.) One normally equates magnetic and electrostatic energies to kinetic and potential energies respectively, and each mode of oscillation of a transmission line will have associated with it a total energy kT, made up of $\frac{1}{2} kT$ of electric-field energy and $\frac{1}{2} kT$ of magnetic-field energy. Only the former half should be relevant to the measurement and exchange of energy through the mechanism of the potential difference between a pair of fixed points on the line. In a line which is open-circuited at both ends there is a total standing-wave energy $(kTl/v)d\nu$ which is equivalent to pairs of traveling waves with velocity $\pm v$ in the forward and backward directions. Assuming for the moment that the total energy can simply be halved to find the energy associated with waves traveling in one direction, this is equivalent to a power flow of $\frac{1}{2} kTd\nu$ in each direction. Since this is independent of line length, the latter can now be allowed to tend to infinity. On subsequently cutting the infinite line at an arbitrary point in the neighborhood of the observer, a single matching resistance R can be connected to one of the remaining parts of the line, whereupon conditions in the line will be the same as when the line was of infinite length. The power exchange between this one resistor and the line should then, according to the above argument, be $\frac{1}{2} kTd\nu$, which is only half that required by the Nyquist formula.

The fallacy in the above arguments seems to be in the assumptions about the energy of a standing wave. If the voltage standing wave is to be described as the resultant of two traveling waves of amplitude A each, the formula may be written.

$$\begin{aligned} V &= A \cos(\omega t + \beta x) + A \cos(\omega t - \beta x) \\ &= 2A \cos \omega t \cos \beta x \end{aligned} \tag{1}$$

where x is distance from the end of the line and β is the phase constant of the line. Now the total mean energy stored in the line is proportional to $\overline{\overline{V^2}}$ where the double bar indicates that an average must be taken over the length of the line, as well as a mean in time. From (1) it follows that

$$\begin{aligned} V^2 &= 4A^2 \cos^2 \omega t \cos^2 \beta x \\ \therefore \overline{\overline{V^2}} &= A^2 \end{aligned} \tag{2}$$

and therefore A^2 is proportional to $\frac{1}{2} kT$. But at the input end of the line where $x=0$,

$$\begin{aligned} V\big|_{x=0} &= 2A \cos \omega t \\ \overline{V^2}\big|_{x=0} &= 2A^2. \end{aligned} \tag{3}$$

Thus the time average of the squared-voltage seen at the end of the line is twice as great as the over-all average-squared-voltage which corresponds to the equipartition energy, and viewed from its terminals the line behaves as though it had capacitive energy kT per degree of freedom instead of the equipartition value $\frac{1}{2} kT$ per degree of freedom.

Finally, the writer wishes to emphasize that although he criticizes some details of the original proof, he has always strongly supported the importance and validity of Nyquist's theoretical derivation of the thermal-noise theorem.[4]

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* Received by the IRE, February 10, 1955; revised manuscript received, November 14, 1955.

[1] H. Nyquist, "Thermal agitation of electric charge in conductors," *Phys. Rev.*, vol. 32, p. 110; July, 1928.

[2] E. B. Moullin, "Spontaneous fluctuations of voltage," Oxford Univ. Press, New York, N.Y.; 1938.

[3] D. A. Bell, "Johnson Noise and equipartition," *Proc. Phys. Soc. B*, Vol. 66, p. 714; August, 1953.

[4] D. A. Bell, "On the general validity of Nyquist's theorem," *Phil. Mag.*, Ser. 7, vol. 27, p. 645; June, 1939.

Russian Condenser Terminology*

The Russian word for condenser, конденсатор is simply a transliteration of the German *Kondensator;* Russian has no word equivalent to the English capacitor, although емкость, capacitance is sometimes used in this respect. The adjective эталонный is derived from the French *étalon*, standard.

adjustable c		регулируемый к.
air c		воздушный к.
air-dielectric c		к. с воздушным диэлектриком
antenna c		к. антенны
automobile c		автомобильный к.
bakelite c		к. с диэлектриком из бакелита
balancing c		уравновешивающий к.
band-spread c		к. растянутого диапазона
blocking c		блокировочный к.
buffer c		буферный к.
bypass c		пропускающий к.
ceramic c		керамический к.
charged c		заряженный к.
charging c		зарядный к.
compressed-air c		к. со сжатым воздухом в качестве диэлектрик
compression-type c		к. емкость которого регулируется изменением сжатия пластин
continuously adjustable c		плавно-переменный к.
coupling c		к. связи
cylindrical c		цилиндрический к.
dual c		двойной к.
electric c		электрический к.
electrolytic c		электролитический к.
dry		сухой
semidry		полусухой
wet		жидкостный
experimental		экспериментальный к.
filter		к. фильтра
fixed c		постоянный к.
fixed-capacity c		к. постоянной емкости
flat-plate c	1	плоский к.
	2	к. с плоскими пластинами
gang c		конденсаторный агрегат (блок)
two-gang c		агрегат двух к—ров
three-gang c		агрегат трех к—ров
general-purpose c		универсальный к.
glass c		стеклянный к.
glass-plate c		к. со стеклянными пластинами
grid c		сеточный к.
high-voltage c		к. высокого напряжения
insulating c		изолирующий к.
inverse-square-law c		обратно-квадратический к.
leaky c		к. с утечкой
Leyden jar		лейденская банка
low-capacity c		к. небольшой емкости
low-loss c	1	малопотерьный к.
	2	к.с малыми потерями
metallic-plate c		к. с металлическими пластинами
metallized-paper c		к. из металлизованной бумаги
mica c		слюдяной к.
microfarad c		микрофарадный к.
micromicrofarad c		пикофарадный к.
midget c		крохотный к.
mid-line c		среднелинейный к.
miniature c		миниатюрный к.
molded c		формованный к.
multiple c		многократный к.
neutralizing c		нейтрализующий к.
noninductive c		безындукционный к.
oil c		масляный к.
oil-impregnated-paper-dielectric c		к. с диэлектриком из пропитанной маслом бумаги
paper c		бумажный к.
paper-dielectric c		к. с бумажным диэлектриком
parallel c		параллельный к.
perfect c		совершенный к.
pigtail c		к. с гибкими электродами
plate c		пластинчатый к.
plate (anode) c		анодный к.
plug-in c		сменный к.
polarized c		поляризованный к.
porcelain c		фарфоровый к.
power c		силовой к.
primary c		первичный к.
protective c		предохранительный к.
punctured c		к. с пробой изоляции
receiving c		приемный к.
root-mean-square c		среднеквадратичный к.
series-gap c		к. представляющий собой последовательное соединение
semiadjustable c		секционно-переменный к.
semivariable c		полупеременный к.
shielded c		экранированный к.
short-circuited c		короткозамкнутый к.
shortening c		укорачивающий к.
short-wave c		к. коротких волн
single-section c		к. одной секции
smoothing c		сглаживающий к.
spherical c		сферический к.
square-law c		квадратический к.
standard c		эталонный к.
steatite c		стеатитовый к.
stopping c		преграждающий к.
straight-line-capacity c		прямоемкостный к.
straight-line-frequency c		прямочастотный к.
straight-line-wavelength c		прямоволновый к.
subdivided (sectionalized) c		секционированный к.
synchronized c		синхронизированный к.
taper-plate c		к. с клинообразными пластинами
telephone c		телефонный к.
trimmer		равнитель
trimming c		подстроечный к.
transmitting c		передающий к.
tubular c		трубчатый к.
tuning c		к. настройки
twin c		сдвоенный к.
unshielded c		неэкранированный к.
vacuum c		вакуумный к.
vane c		к. с крыльчатыми пластинами
variable c	1	переменный к.
	2	к. переменной емкости
vernier c	1	верньерный к.
	2	к. точной настройки

George F. Schultz
Indiana Univ.
Bloomington, Ind.

* Received by the IRE, April 13, 1956.

Contributors

Franklin S. Coale (A'53) received the B.S. degree in engineering physics from Lehigh University in 1952, subsequently taking post graduate work in applied mathematics at New York University. In 1953 he joined the Sperry Gyroscope Co. as an associate engineer in microwave components and antennas engineering. In September, 1955, Mr. Coale joined Stanford Research Institute as a research engineer in the Microwave Group of the Antenna Systems Laboratory. He is engaged in research and development on filters, multiplexers, ferrite devices, and antennas.

F. S. Coale

He is a member of the American Physical Society and the Research Society of America.

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Seymour B. Cohn (S'41–A'44–M'46–SM'51) was born in Stamford, Conn. in 1920. He received the B.E. in electrical engineering from Yale University in 1942; the M.S. degree in communication engineering in 1946, and the Ph.D. degree in engineering sciences and applied physics in 1948, both from Harvard University. From 1942 to 1945, he was employed as a special research associate by the Radio Research Laboratory of Harvard University, also representing that Laboratory as a technical observer with the U. S. Army Air Force in the Mediterranean Theater of Operations. Dr. Cohn worked at Sperry Gyroscope Co. from 1948 to 1953, where he held the position of research engineer in the Microwave Instruments and Components Department.

S. B. Cohn

Since February, 1953, he has been with the Stanford Research Institute, as head of the Microwave Group of the Antenna Systems Laboratory.

He is a member of Tau Beta Pi and Sigma Xi.

❖

Joseph Kaye was born in Malden, Mass., in 1912. He received the B.S. degree in Physical Chemistry from Massachusetts Institute of Technology in 1934, and was awarded the first Stratton Prize the same year. Remaining at M.I.T., he was Fellow of the Institute in 1936, received the Ph.D. degree in Physical Chemistry in 1937, and was an A. D. Little Post-Doctorate Fellow from 1937 to 1939.

During the summer of 1939, Dr. Kaye was a turbine design engineer for General Electric in Lynn, Mass. He returned to M.I.T. later that year and since that time has worked there in various capacities. For one year, he was a research engineer in the Mechanical Engineering Department. From 1940 to 1944 he was an instructor, and from 1944 to 1955 was an assistant and then an associate professor. Since 1955 he has been a full professor in Mechanical Engineering, and Director of the Research Laboratory of Heat Transfer in Electronics.

Joseph Kaye

Dr. Kaye has been an engineering consultant to several firms, among them United Aircraft Corp., Raytheon Manufacturing Co., RCA, Federal Telecommunications Laboratory, Inc., American Locomotive Co., and to the U. S. Naval Underwater Ordnance Station.

He is a registered professional engineer in Massachusetts and a member of the American Chemical Society, American Physical Society, American Society of Mechanical Engineers, American Society for Engineering Education, Institute of Aeronautical Sciences, Sigma Xi, and Pi Tau Sigma, and is a Fellow of the American Academy of Arts and Sciences.

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David M. Kerns is a native of Minneapolis, Minn., and received the degree of B.E.E. from the University of Minnesota in 1935. He received the Ph.D. in Physics from Catholic University in Washington, D. C., in 1951.

After receiving his degree from the University of Minnesota he was employed there until 1940 as a teaching assistant and as a research assistant. This was followed by approximately two years in the Aerial Camera Laboratory at Wright Field, Dayton, Ohio, where he was in charge of the Electrical Unit. In 1942 Dr. Kerns began his service in the Signal Corps, U. S. Army; his principal tour of duty was at the wartime M.I.T. Radiation Laboratory, Cambridge, Mass., as project liaison officer associated with microwave fire control radar and microwave test equipment.

D. M. Kerns

Dr. Kerns joined the Microwave Standards Section of the National Bureau of Standards Central Propagation Laboratory in 1946 and in 1949 became assistant chief of the Section. He has worked primarily with the electromagnetic theory of waveguides and waveguide junctions. Dr. Kerns is now assistant chief of the Radio Standards Division and chief of the Microwave Physics Section in the Boulder Laboratories of the National Bureau of Standards.

Dr. Kerns is a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

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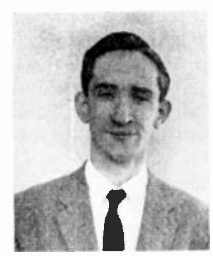
C. Gordon Little was born in Liu Yang' Hunan, China on November 4, 1924. Before graduating from the University of Manchester, England in 1948 with the B.Sc. degree with honors in Physics, he worked in industrial research laboratories for three years on the design of high voltage rectifiers and electrometer tubes.



C. G. Little

After carrying out graduate studies at the Jodrell Bank Radio Astronomy Research Center of the University of Manchester, Dr. Little was awarded the Ph.D. degree in 1952. In January, 1954 he took up an appointment as visiting professor of Geophysics and senior research scientist at the Geophysical Institute of the University of Alaska. Since July, 1954, Dr. Little has been assistant director of the Geophysical Institute, in charge of research programs in ionospheric physics and arctic radio wave propagation.

Dr. Little is a Fellow of the Royal Astronomical Society, London, and an Associate Member of the Arctic Institute of North America.

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Alan C. Macpherson was born in Washington, D. C., on December 24, 1920. He received the B.S. degree in Physics from the University of Maryland in 1943 and the M.A. in Physics from George Washington University in 1950. Serving in the Signal Corps from 1943 to 1946, his duties included work on radar, proximity fuse production, and special purpose electronic tubes.

A. C. Macpherson

In 1947 he joined the National Bureau of Standards where he worked on precision measurements of power and impedance at microwave frequencies. He is now with Naval Research Laboratory, Washington.

He is a member of Sigma Pi Sigma.

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David M. Makow (A'52) was born on December 13, 1923, at Lodz, Poland. He received the Dipl.-Ing. degree in E.E. in 1950 from the Swiss Federal Institute of Technology in Zurich, and was employed in the development of pulse-time multi-channel communication systems with the Brown Voveri Company in Baden, Switzerland.

D. M. Makow

In 1951 he joined the National Research Council of Canada in Ottawa where he participated in radio astronomy work. For the past three years he has been engaged in research and development of certain multiloop feedback systems, and more recently of variable frequency oscillators.

Mr. Makow is a member of the Association of Professional Engineers of Ontario.

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Willis M. Rayton was born on May 17, 1909, in Rochester, N. Y. He received the B.S. degree from Hamilton College in 1931, and the M.A. and Ph.D. from the University of Rochester in 1936 and 1939. He was with the Physics department at M.I.T. from 1936 to 1939, and since then has been teaching Physics at Dartmouth College. He was appointed Professor in 1947.

Willis Rayton

During World War II he did research work in underwater sound. Since 1953 he has been associated with ionospheric radio wave propagation research at the Thayer School of Engineering at Dartmouth. In 1955 he spent one year at the Geophysical Institute at College, Alaska conducting radio propagation studies particularly in connection with the anomolous absorption effects especially noticeable at high latitudes.

Dr. Rayton is a member of the American Physical Society, the American Association for the Advancement of Science, and Sigma Xi.

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Raymond B. Roof was born on December 22, 1900 in Climax, Mich. He received the B.S. degree in Physics from the University of Michigan in 1925. After teaching on the secondary school level for several years he obtained the M.S. degree in Physics from the University of Michigan in 1940.

R. B. Roof

Concurrently with teaching, he was chief engineer of radio station WELL from 1930 to 1942. From 1942 to 1949 he was electrical superintendent of Eaton Manufacturing Company in Battle Creek, Mich.

He then returned to teaching until 1954 when he joined the staff of the Geophysical Institute and the University of Alaska.

Mr. Roof is a member of Phi Beta Kappa.

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Allen H. Schooley (A'35–SM'47–F'54) was born on December 16, 1909, in Terril, Iowa. He received the B.S. degree in electrical engineering from Iowa State College in 1931, and the M.S. degree from Purdue University in 1932. Prior to World War II he was a research and development engineer with RCA Radiatron, Harrison, N. J., where he designed and built the first miniature radio tubes. Since 1940 Mr. Schooley has been at the U. S. Naval Research Laboratory, where he is the superintendent of the Electronics Division. He has been selected by the chief of Naval Research to assist the Brazilian Navy establish a naval research laboratory in Rio de Janeiro. For this assignment he will be on leave-of-absence from the Naval Research Laboratory during 1956.

A. H. Schooley

In 1946 Mr. Schooley received the Navy Distinguished Civilian Service Award. He is a member of several scholastic and professional societies and is the author of numerous technical articles.

IRE News and Radio Notes

Four Sections Enter The IRE

On May 10, the IRE Board of Directors approved the establishment of the following new Sections: Regina, Southern Alberta, Alamogordo-Holloman, and Tucson (date of establishment retroactive to April 1, 1956).

With the establishment of the Southern Alberta Section, the name of the former Alberta Section was changed to the Northern Alberta Section.

On May 9, the establishment of the following Subsections were officially approved; San Fernando Valley, Los Angeles Section; and Memphis, Huntsville Section.

On June 5, the IRE Executive Committee approved the establishment of the Western North Carolina Subsection of the North Carolina-Virginia Section.

Nat'l Elec. Conference Sets 1957–1963 Meeting Dates

In order to avoid conflicting dates for technical meetings, the National Electronics Conference has released a seven-year schedule of dates for future conferences. They are:

1957—October 7–9
1958—October 13–15
1959—October 12–14
1960—October 10–12
1961—October 9–11
1962—October 8–10
1963—October 7–9

The 1956 National Electronics Conference will be held in Chicago October 1–3.

Semiconductor Symposium is Slated for October 1–4

The Semiconductor Group of the Electrochemical Society will sponsor a symposium on semiconductors at the Statler Hotel, Cleveland, Ohio, October 1–4, 1956. Papers on semiconductor materials, elemental alloys and compounds, surface controlled phenomena, and chemical process technology will be presented. Recent news-type papers of ten-minutes' length will also be given. Mimeographed abstracts of all papers will be available at the registration desk.

For those who are planning to submit papers to the symposium, seventy-five word abstracts should be mailed before September 1 to M. F. Lamorte, Program Chairman, Westinghouse Electric Corporation, Semiconductor Department, Youngwood, Pennsylvania.

MIT and PGIT Sponsor Symposium on Information Theory

The Professional Group on Information Theory and the Research Laboratory of Electronics of the Massachusetts Institute of Technology are organizing an Information Theory Symposium, to be co-sponsored by the Office of Naval Research, the Air Research and Development Command, the Signal Corps Engineering Laboratories and the U.R.S.I., on Sept. 10–12, 1956, at M.I.T., Cambridge, Mass.

Registration material will be available in July and will automatically be mailed to members of the PGIT; others may obtain it on request from the chairman of the organizing committee, Professor Peter Elias at the Research Laboratory of Electronics, Massachusetts Institute of Technology.

There will be morning and afternoon sessions on each of the three days on such topics as coding theory, information sources (including language and pictures), and filtering and detection problems. There will also be an evening meeting on Monday, September 10, to discuss computing-machine simulation of learning and problem-solving behavior, and a banquet on the following day.

All papers to be presented will be printed in an issue of the PGIT Transactions, which will be available before the meeting, permitting informed discussion, which is expected to occupy about half of the available 45-minute period assigned to each paper.

The organizing committee for the symposium includes T. P. Cheatham of Melpar, and J. B. Wiesner, R. M. Fano, Y. W. Lee and W. A. Rosenblith of M.I.T.

Sir George Nelson (*center*), president of the Institution of Electrical Engineers, London, England, and chairman and managing director of English Electric Co., Ltd. is shown with IRE President A. V. Loughren (*left*) and M. D. Hooven (*right*), president of the American Institute of Electrical Engineers at a recent AIEE luncheon honoring him. Also present at the luncheon were W. K. Brasher, IEE secretary, G. W. Bailey, IRE executive secretary, and officers of other American engineering societies. The IEE sponsors a convention on ferrites Oct. 29–Nov. 2.

Calendar of Coming Events

National Telemetering Conference, Biltmore Hotel, Los Angeles, Calif., Aug. 20–21
IRE-West Coast Electronic Manufacturers' Association, WESCON, Pan Pacific Auditorium and Ambassador Hotel, Los Angeles, Calif., Aug. 21–24
Annual Summer Seminar, Emporium Section, American Legion Hall, Emporium, Pa., Aug. 24–26
Symposium on Information Theory, Cambridge, Mass., Sept. 10–12
Second RETMA Conference on Reliable Electrical Connections, U. of Pa., Philadelphia, Pa., Sept. 11–12
PGBTS Sixth Annual Fall Symposium, Pittsburgh, Pa., Sept. 14–15
Conference on Communications, Roosevelt Hotel, Cedar Rapids, Iowa, Sept. 14–15
Transistor Reliability Symposium, New York City, Sept. 17–18
Instrument-Automation Conference & Ex., Coliseum, N. Y. C., Sept. 17–21
Symposium on Radio-Wave Propagation, Paris, France, Sept. 17–22
PGNS Third Annual Meeting, Mellon Institute Auditorium, Pittsburgh, Pa., Sept. 20–22
Industrial Electronics Symposium, Manger Hotel, Cleveland, Ohio, Sept. 24–25
National Electronics Conference, Hotel Sherman, Chicago, Ill., Oct. 1–3
Canadian IRE Convention & Exposition, Automotive Bldg., Exhibition Park, Toronto, Can., Oct. 1–3
Second Annual Symposium on Aeronautical Communications, Hotel Utica, Utica, N. Y., Oct. 8–9
URSI Fall Meeting, Univ. of Calif., Berkeley, Calif., Oct. 11–12
IRE-RETMA Radio Fall Meeting, Hotel Syracuse, Syracuse, N. Y., Oct. 15–17
Conference on Magnetism & Magnetic Materials, Hotel Statler, Boston, Mass., Oct. 16–18
PGED Annual Technical Meeting, Shoreham Hotel, Washington, D. C., Oct. 25–26
Aircraft Electrical Society, Annual Display of Aircraft Elec. Eequipment, Pan-Pacific Audit., Los Angeles, Calif., Oct. 25–26
East Coast Conference on Aeronautical & Navigational Electronics, Fifth Regiment Armory, Baltimore, Md., Oct. 29–30
Convention on Ferrites, Institute of Electrical Engineers, London, England, Oct. 29–Nov. 2
Conference on Electrical Techniques in Medicine and Biology, Governor Clinton Hotel, N. Y., Nov. 7–9
Kansas City IRE Technical Conference, Town House Hotel, Kansas City, Kan., Nov. 8–9
Symposium on Applications of Optical Principles to Microwaves, Washington, D. C., Nov. 14–16
PGVC Eighth National Meeting, Fort Shelby Hotel, Detroit, Nov. 29–30
Second Instrumentation Conference & Exhibit, Biltmore Hotel, Atlanta, Ga., Dec. 5–7
IRE-AIEE-ACM Eastern Joint Computer Conference, Hotel New Yorker, New York City, Dec. 10–12

PIB Awards Certificate of Achievement to L. A. de Rosa

L. A. deRosa, Laboratory Director at Federal Telecommunication Laboratories, Nutley, N.J., was awarded a Certificate of Achievement by the Polytechnic Institute of Brooklyn recently for outstanding work in science and engineering.

L. A. deRosa

Mr. deRosa, who began his engineering career a quarter century ago, has been with the Laboratories, a division of International Telephone and Telegraph Corporation, since 1942.

The award was made at the Centennial ceremonies of Polytechnic Institute on Alumni Day, May 19, in Brooklyn.

Mr. deRosa, a native of Tenafly, N.J., was graduated from Polytechnic Institute in 1932 with a degree in electrical engineering. He did graduate work at Polytechnic Institute until 1934, and later at Columbia University.

During his professional career he has made contributions to the fields of electronics and acoustics. In acoustics he is known for his studies of physiopsychological phenomena and his "two-canal" theory of hearing. In electronics he holds forty patents in radar, direction finding, navigational aids, communication systems, antennas, electronic warfare devices and computers.

A Fellow of the Institute of Radio Engineers, Mr. deRosa is also chairman of the Professional Group on Information Theory. He is a member of the American Physical Society and the Acoustical Society of America.

East Coast Conference on Aero. & Nav. Electronics is Planned for Oct. 29–30

Baltimore's Fifth Regiment Armory will be the scene of the Third Annual East Coast Conference on Aeronautical and Navigational Electronics on October 29–30, 1956. About 1,500 engineers and scientists are expected to attend the technical sessions, which will feature the conference theme of "Electronics in the Jet Air Age." This conference is sponsored jointly by the Baltimore Section of the IRE and the IRE Professional Group on Aeronautical and Navigational Electronics. Over 50,000 square feet of floor space will be available on the street-level exhibition floor area of the Armory. Conference headquarters will be the Sheraton Belvedere Hotel, where the annual banquet will be held on the evening of October 29.

The conference steering committee is composed of representatives of the two sponsoring organizations and is headed by Conference Chairman Joseph General, who is deputy chief of the Plans and Programs Office at the USAF's Air Research and Development Command headquarters in Baltimore. C. F. Miller, of the Johns Hopkins University faculty, is Vice-Chairman.

Other steering committee members are T. T. Eaton, *Finances;* J. A. Houston, *Local Arrangements;* A. A Nims, *Exhibits;* Harald Schutz, *Technical Program;* W. D. Crawford, *Publicity;* Norman Caplan, *Baltimore Chapter, PGANE;* and C. D. Pierson, Jr., and C. E. McClellan, advisory members from the 1955 committee.

Second IRE Instrumentation Conference Invites Papers

The Professional Group on Instrumentation and the Atlanta Section of the IRE will sponsor the Second IRE Instrumentation Conference at the Biltmore Hotel in Atlanta on December 5-7, 1956. B. J. Dasher, Director of the School of Electrical Engineering, Georgia Institute of Technology, will serve as conference chairman. Sessions will be devoted to the following subjects: industrial nuclear instrumentation and instrumentation for radiological safety, industrial application of instrumentation, guided-missile range instrumentation and wind-tunnel instrumentation, solid-state devices and other components, laboratory instrumentation and test equipment, and recorders.

Prospective authors are invited to submit abstracts of 200 words or less, not later than September 15, to the program chairman, M. D. Prince, Engineering Experiment Station, Georgia Institute of Technology, Atlanta, Georgia.

Nat'l Elec. Conference Proceedings Now Available

The theme of the twelfth annual National Electronics Conference, "Fifty Years of Progress Through Electronics," will mark the golden anniversary of the electronics industry. Approximately a hundred technical papers will be presented and 240 exhibits will be on display at the conference scheduled for Hotel Sherman, Chicago, Ill., Oct. 1–3. Attendees will also hear luncheon speakers on such topics as space satellites, economic trends, and Russian vs U. S. technical education.

The registration fee will be $3.00 for all technical sessions. Advance orders are now being accepted for copies of the proceedings of the 1956 conference at $5.00 per copy. Back issues of the proceedings of the 1946, 1949, 1950, 1951, 1952, 1953, 1954 and 1955 conferences can also be ordered at $5.00 per copy. All orders should be sent to the National Electronics Conference, 84 E. Randolph St., Chicago 1, Ill.

Audio Society Joins Hi-Fi Show Sept. 26–29 at New York

The 1956 convention of the Audio Engineering Society will be held September 26–29, in participation with the New York High Fidelity Show, which is sponsored by the Institute of High Fidelity Manufacturers, Inc. Both events will take place in the New York Trade Show Building, Eighth Avenue between 35th and 36th Streets, New York. W. O. Stanton will be general chairman of the 1956 convention.

Topics to be covered at the AES convention will include: magnetic recording and reproduction, disc recording and reproduction, transistor application problems, audio systems and components, home music system design, loudspeakers, and audio standards and measurement methods. The annual banquet and presentation of awards will be held on the evening of September 27.

Baker is New Head of RETMA

The Radio-Electronics-Television Manufacturers Association recently elected W. R. G. Baker (A'19–F'28) president at a meeting in Chicago. He has been a member of the RETMA Board of Directors and Director of its Engineering Department since 1934. He is vice-president of General Electric Company.

W. R. G. Baker

Dr. Baker is Treasurer, Chairman of the Professional Group Committee and one of the directors of the IRE. He was awarded the 1952 IRE Medal of Honor.

Dr. Baker was chairman of two National Television Systems Committees which developed standards for black-and-white television in 1941 and color television in 1953.

November 2 Deadline for 1957 IRE Convention Papers

Original papers only shall be submitted, not published or presented prior to the 1957 IRE National Convention.

Prospective authors are requested to submit *all* of the following information: (1) 100-word abstract *in triplicate*, title of paper, name and address; (2) 500-word summary *in triplicate*, title of paper, name and address; (3) indicate the technical field in which your paper falls: Aeronautical & Navigational Electronics, Antennas & Propagation, Audio, Automatic Control, Broadcast & Television Receivers, Broadcast Transmission Systems, Circuit Theory, Communications Systems, Component Parts, Electron Devices, Electronic Computers, Engineering Management, Industrial Electronics Information Theory, Instrumentation, Medical Electronics, Microwave Theory & Techniques, Military Electronics, Nuclear Science, Production Techniques, Reliability & Quality Control, Telemetry & Remote Control, Ultrasonics Engineering, Vehicular Communications.

Address all material to: Ben Warriner, Chairman, 1957 Technical Program Committee, The Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y.

Activities of IRE Sections and Professional Groups

Mayor Thomas D'Alesandro, Jr. of Baltimore has proclaimed the week of Oct. 29–Nov. 3 as "Airborne Electronics Week," in recognition of the annual East Coast Conference on Aeronautical and Navigational Conference to be held in that city Oct. 29–30. Shown *(left to right)* are members of the conference steering committee: W. D. Crawford, *Publicity;* A. A. Nims, *Exhibits;* C. D. Pierson, Jr., *1955 Conference Chairman;* T. T. Eaton, *Finance;* Mayor D'Alesandro; F. T. McHugh, Aviation Director, Baltimore Association of Commerce; Joseph General, *1956 Conference Chairman;* J. A. Houston, *Arrangements;* and Harald Schutz, *Technical Program.*

A. C. Beck *(right)*, New York Section chairman, presents R. R. Batcher with a certificate of appreciation for his active work on IRE committees and as chairman of the PG on Production Techniques.

The New York Section recently held a dinner in New York City to honor its 1956 Fellow members. With A. V. Loughren *(extreme left)*, 1956 IRE President, shown are *(left to right)*: W. J. Barkley, G. W. Bailey, N. G. Anton, H. J. Carlin, J. Z. Millar and C. E. Scholz. G. H. Brown of RCA gave a dinner address on "Whither the Scientist."

At the recent New York Section dinner certificates of service were presented to past chairmen of the Section. Shown *(left to right)* are: H. M. Lewis, L. Espenschied, G. B. Hoadley, J. T. Cimorelli, J. E. Shepherd, R. D. Chipp, A. C. Beck, J. H. Mulligan, H. T. Budenbom and S. Shamis. A. B. Giordano was not in the picture, and J. W. McRae and H. F. Dart were unable to be present at the dinner. Mr. Beck gave the address of welcome and I. S. Coggeshall was toastmaster at the dinner.

Professional Group News

IRE Establishes 3 Chapters

The following Professional Group Chapters were approved by the IRE Executive Committee at its meeting June 5: PG on Circuit Theory, Quebec Subsection of the Montreal Section; PG on Military Electronics, Long Island Section; and PG on Production Techniques, Joint New York-Northern New Jersey-Long Island Section. The name of the New York Chapter of the PG on Component Parts was converted to the Metropolitan New York Area Chapter. This latter chapter also belongs to the Joint New York-Northern New Jersey-Long Island Section.

PG on Nuclear Science Holds Third Annual Meeting

The Professional Group on Nuclear Science will hold its Third Annual Meeting at Pittsburgh, Pennsylvania, September 20–22. Technical sessions on nuclear instrumentation, reactor control, industrial applications, analog simulation and digital computation will be held at the Mellon Institute Auditorium September 20 and 21. Hotel headquarters for the meeting will be at Webster Hall, across the street from the auditorium. A tour of the Shippingport nuclear power station will be made on September 22, and a ladies' program is also planned.

PGANE Announces Winners of 1956 Pioneer Awards

At the May National Aeronautical and Navigational Conference at Dayton, Ohio, Brig. Gen. A. W. Marriner (USAF, retired) and Brig. Gen. W. G. Smith (USAF, retired) were named joint recipients of the 1956 Pioneer Award presented by the Professional Group on Aeronautical and Navigational Electronics. They were cited for their part in the promotion, development, procurement, organization and operation of electronic facilities for Air Force air communications, navigation and traffic control; and in the inauguration, training and administration of Air Force organizations set up for the operation of these facilities. The citation referred specifically to the role these two officers played in the original planning, establishment and early growth of the Airways and Air Communications Service, formerly called the Army Airways Communications Service.

Posthumous citations of Pioneer Awards, included for the first time in the awards program, also went to John Stone-Stone, F. A. Kolster, J. W. Grieg, W. H. Murphy, Harry Diamond, W. G. Eaton, and Thorp Hiscock.

PGED Meets October 25–26

The Second Annual Technical Meeting of the IRE Professional Group on Electron Devices will be held at the Shoreham Hotel, Washington, D. C., October 25–26, 1956.

R. L. Pritchard, in charge of the meeting's technical program, said that there will be sessions on microwave tubes, semiconductor devices, and display and storage tubes. It is hoped that there will also be papers on gas tubes and ferromagnetic and ferroelectric devices.

Members of the general committee in charge of the 1956 meeting include: *General Chairman*, T. M. Liimatainen, Diamond Ordnance Fuze Laboratories; *Vice-Chairman*, H. L. Owens, Texas Instruments, Inc.; *Technical Program*, R. L. Pritchard, General Electric Co.; *Secretary*, A. K. Wing, Federal Telecommunications Laboratories; *Publications*, E. L. Steele, General Electric Co.; *Local Arrangements*, J. H. Wright, National Bureau of Standards; *Publicity*, Prall Culviner, Sylvania Electric Products, Inc.

PGVC Initiates Papers Award

The Professional Group on Vehicular Communications has announced the initiation of an annual Papers Award. The award, consisting of a suitable certificate and two hundred dollars in cash, will be made annually to an IRE member for a technical paper in the field of interest of the Group. To be eligible, the paper must be published in the IRE Proceedings, the IRE Convention Record, the PGVC Transactions or any other duly authorized publication of the IRE within the period beginning with July 1 of a calendar year and ending with June 30 in the next succeeding calendar year. Transactions authors should submit papers to A. A. Macdonald, Chairman, PGVC National Papers Committee, Motorola Inc. 4545 W. Augusta Blvd., Chicago 51, Illinois.

Obituaries

George H. Clark, a pioneer in wireless telegraphy and for twenty-seven years associated with the Radio Corporation of America, died recently.

Mr. Clark retired in 1946 after having served since 1931 as custodian of historical archives for R.C.A.

He joined R.C.A. in 1919, when the newly formed corporation acquired the Marconi Wireless Telegraph Company of America. He had been affiliated with Marconi following pioneer wireless service with the Navy during World War I.

From 1919 to 1931 Mr. Clark was manager of the R.C.A. exhibit division. He organized and directed the operations of an exposition that toured the nation.

As custodian of historical archives, he collected and catalogued files of early radio companies, photographs, blueprints and similar matter concerning radio pioneers. In 1952, this "R.C.A.-Clark Collection of Radioana" was presented by R.C.A. to his alma mater, the Massachusetts Institute of Technology.

Mr. Clark was the author of biographies of Roy A. Weagant, early radio engineer, and John Stone-Stone, radio inventor and engineer.

Born on Feb. 15, 1881, at Alberton, Prince Edward Island, Mr. Clark came to the United States with his parents at the age of fourteen. He received his high school education at Everett, Mass., and worked summers as a telegraph operator for the Boston and Maine Railroad.

After graduation from M.I.T. with a Bachelor of Science degree in electrical engineering in 1903, he went to work for the Stone Telegraph & Telephone Co., Boston.

From 1908 until after World War I, he served as a civilian assistant to the Navy as a subinspector of wireless telegraph stations and as an expert radio aide. In 1912 he had received Government License No. 2 for commercial wireless operators.

Mr. Clark, a former IRE Fellow, was a founder and past president of the Veteran Wireless Operators Association.

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Arthur V. Hollenberg (M'50–SM'50), a research physicist with the Bell Telephone Laboratories in Murray Hill, died recently.

Dr. Hollenberg had been with Bell Laboratories since 1946. During his first six years he specialized in research on microwave electron tubes, particularly traveling wave and double-stream amplifiers. Since 1952 he had been in charge of a laboratory engaged in the construction of complex experimental apparatus for physical research at Columbia University, where he carried on research and development work on microwave magnetrons for radar applications.

Dr. Hollenberg received a Bachelor of Arts degree in 1931 from Willamette University, Salem, Ore., and Master of Science and Doctor of Philosophy degrees from New York University, in 1933 and 1938 respectively.

Dr. Hollenberg was an instructor in physics at Queens College in New York from 1938 to 1942. For the next three years he was a member of the scientific staff of the division of war of Radio Engineers and a member of Sigma Xi.

He had been granted several patents for his inventions in the field of electron tubes and was the author of several technical articles on this subject. Dr. Hollenberg was a fellow of the American Physical Society.

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James R. Nelson (A'27–M'29–SM'43), an authority during the 1930's on the application of vacuum tubes to home, automobile and portable radio sets, died recently.

For twenty-six years, Mr. Nelson had been with the Raytheon Manufacturing Company, Waltham, Mass.

In 1946 he received a Certificate for Exceptional Service to Naval Ordnance Development for his work on proximity fuse tubes. He also served on many industry standardization committees on vacuum tubes, including the War Production Board industry committee on miniature tubes.

Mr. Nelson was the author of more than fifty technical articles. For the last four years he had directed work on transistor applications.

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Hillel I. Reiskind (SM'52), manager of the engineering division of the Radio Corp. of America plant in Indianapolis, Ind., died recently.

In 1945, he became chief engineer of the RCA Victor Record Division at Camden, N. J. and in 1948 came to Indianapolis as manager of the engineering division.

In 1936–37, Mr. Reiskind engineered all the sound recording installations for 20th Century Fox and Warner Bros. in New York and in 1938 went to Hollywood as technical supervisor for RCA film recording operations.

In 1939–40, he was development engineer for the sound track of Walt Disney's "Fantasia."

Before joining RCA, Mr. Reiskind had worked for Paramount Pictures and Eastern Service Studios, Inc.

He was a member of the Society of Motion Picture and Television Engineers, and the technical committee of Radio-Electric-Television Manufacturers' Association.

Technical Committee Notes

Chairman Henry Jasik presided at a meeting of the **Antennas and Waveguides** Committee on May 9 at IRE Headquarters. The entire meeting was devoted to the discussion of the Proposed Standards on Antennas and Waveguides: Waveguide and Waveguide Component Measurement.

The **Electron Tubes** Committee met on May 11 at IRE Headquarters with Chairman P. A. Redhead presiding. The chairman reported that the Proposed Standards on Electron Tubes: TR and ATR Tube Definitions had been approved by the Standards Committee on May 10. T. J. Henry reported the completion of the proposed Standard on Electron Tubes: Methods of Test. The following items have not been considered by the subcommittee: microphonics; noise, other than shot; environmental tests; cross modulation; hum; hot capacitances with current flowing; snibbets (an interesting disease of deflection output tubes). It was agreed that IRE should not attempt to formulate methods of test for reliability or environmental effects. The Methods of Test, when completed, will be published in two parts: "Part A—General Test Method" and "Part B—Tests for Special Tubes."

Chairman K. R. McConnell presided at a meeting of the **Facsimile** Committee held at the Times Building on May 18. The major portion of the meeting was devoted to a review of the Proposed Standard on Facsimile: Methods of Test. It was suggested that a bibliography for each item in the test standards be included as it became necessary. The committee believes that an overall up-to-date bibliography on facsimile would be within its scope, since the scope of the committee does include the dissemination of information.

The **Information Theory and Modulation Systems** Committee met at IRE Headquarters on May 25 with Chairman J. G. Kreer, Jr. presiding. There was a discussion on the Proposed Standard on Computer Definitions on which the committee was asked to comment. In a number of areas, there were conflicts between identical terms as defined by the Committee on Computers and the Information Theory and Modulation Systems Committee. The comments made by the committee members will be passed on to the Standards Committee at their next meeting.

The committee spent the remainder of the meeting reviewing the Proposed Standard on Information Theory and Modulation Systems: Definitions of Terms.

Chairman H. R. Goldberg presided at a meeting of the **Radio Transmitters** Committee held at IRE Headquarters on May 11. A major part of the meeting was given to discussion of the Proposed Standard on Definitions of Terms.

A letter from I. Kaar of the Joint Technical Advisory Committee (JTAC) was read by the chairman. In it an invitation to submit papers on single sideband for inclusion in a special issue of the PROCEEDINGS OF IRE was extended. Although it appears improbable that the Standard Methods of Test for Single Sideband Transmitters can be processed in time for such a special issue, it was suggested that a paper might be prepared by the chairman of Subcommittee 15.5 on the background of the standardization problem.

The **Standards** Committee met at IRE Headquarters on May 10 with Chairman M. W. Baldwin, Jr. presiding. On a motion by J. G. Kreer which was seconded by R. Serrell, the scope of the Medical Electronics Committee was unanimously approved, subject to the amendment of the Bylaws by the Board of Directors in August.

The Proposed Standard on Methods of Testing Transistors was discussed and amended, and review will continue at the next meeting.

The Proposed Standards on Electron Tubes: Definitions of Terms Related to Non-Transit Time Tubes was discussed, amended and unanimously approved on motion by P. A. Redhead and seconded by C. H Page.

Mr. Redhead, chairman of the Electron Tubes Committee, reported that the committee did not want this standard published until the committee completes all electron tube definitions.

The Proposed Standard on Electron Tubes: Cathode Ray Tube Definitions was discussed and amended, and review will continue at the next meeting.

Books

Reliability Factors for Ground Electronic Equipment, ed. by Keith Henney

Published (1956) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 288 pages+4 page index+vii pages. Illus. 8½×11. $7.50.

This book is a qualitative treatment of electronic equipment design as related to reliability. Though the material was prepared primarily for ground-based equipment, most of the information is readily applicable to other classes of equipment.

The material presented is largely a digest drawn from technical journals and reports of government contract activities. While there is very little new material, to one familiar with the literature in the field of reliability, the value of a good digest is self-evident. The copious references and bibliographies are particularly valuable.

The broad scope of the factors influencing reliability is reflected in the organization of the book which starts with reliability concepts, causes of unreliability, systems aspects, and the mathematical approach to reliability. The material is then expanded through discussions of various aspects of equipment design. Specifically, these aspects of design are separated into electrical and electronic factors, mechanical and environmental factors, human engineering, component parts, interference, automatic production, and design for ease of maintenance. A further chapter is included on the importance of equipment publications, *i.e.*, instruction books. It is unfortunate, but understandable, that in a rapidly progressing field—as reliability is today—many aspects of a book on the subject are obsolete before the book can be published. The latest reference in the book is 1954. Substantial progress in the quantitative treatment of reliability has come about since that time. The objective of the book, nevertheless, is very worthwhile and the material could serve as an introduction to the neophyte in the field of reliability. While the objectives are excellent, the execution (from the standpoint of editorial accuracy, consistency, and technical accuracy even in well-established fields) falls far short of the excellent reputation of this publisher. In many respects the errors could be dangerous if accepted uncritically by an inexperienced designer.

Here follow examples of the types of errors referred to in the review:

Page 1–7:

> "It is now recognized that tube failures are most likely to occur during their early history, say, the first 50 or 100 hours. Like human beings, whose mortality is greatest during their childhood, if tubes get over this critical period, they are likely to have a normal life expectancy. To avoid these "quickie" troubles it is becoming customary to "burn in" tubes or to age them before they are accepted or are permitted to go into the equipment. After this period, failures

seem to follow an exponential pattern; in a given time a given percentage of those still in service must be replaced."

This is not so much an error as reflection of the date of the material. This statement was believed to be true rather generally up through 1953. Growing masses of data since that time, however, have quite firmly established that in receiving type electron tubes this "infant mortality" is no longer apparent and that the probability distribution of time to failure in receiving tubes is most frequently Gaussian. The unnecessary "burning in" of tubes by equipment manufacturers and the attendant economic waste is a primary concern of military procurement agencies at this time.

Page 2-5:

"Under no type of contract can the project engineer or the service he represents compel manufacturers to build reliability into the ultimate equipment. Reliability is a function of time, and by the time the lack of reliability shows up, the production run is over, and nothing can be done about it. No guarantees of reliability can be anything more than a statement in the specification, for the simple reason that nothing can be done to penalize the manufacturer if his equipment fails to come up to this paragraph in the specification."

The above paragraph is followed later (pages 2-6 and 2-7) by the following paragraph:

"It is certain that the future specifications for equipment will contain requirements for reliability. At the moment, performance, cost, size, and weight are cited, but in time reliability will surely be one of the prime objectives in development. It is the belief of many in the components industry that if reliability were established as a basic requirement equipment engineers would approach this problem differently. They would exercise more care in the assembly of components; they would study the limitations of components more carefully and use them in ways which would not tax them unduly; they would devise systems and circuits, perhaps unorthodox, which would permit the use of components with wider tolerance and variations; they would use more rather than fewer, larger rather than smaller components to achieve reliability. The burden would then be more equitably divided, and the incidence of unreliability considerably reduced."

The complete inconsistency between these two statements does not reflect a careful digest of the references from which these two statements were drawn.

Chapter 4—"Mathematical Approach to Reliability."

The author fails to distinguish between the "population" and the "sample," defining probability relations and concepts only in terms of the sample. This is dangerous, for it infers to the uninitiated that one failure in two observations has precisely the same meaning as ten failures in twenty observations.

Page 6-4:

"*Air Cooling.* Hermetically-sealed assemblies have been cooled by a number of methods. Nitrogen filling has been used by the National Bureau of Standards. Several groups are considering the use of hydrogen in sealed packages. . . . "

The use of hydrogen under pressure for cooling of any electronic equipment employing electron tubes is doomed to failure. It is well known that hydrogen penetrates the glass envelope of electron tubes to a serious extent over long periods of time. Even helium is unsatisfactory from this standpoint.

Page 6-9:

The legends for figures 6-5, 6-6 and 6-7 are incomplete.

Page 8-5:

"It may astonish the equipment designer to know that it was not generally agreed upon until the last few years that the basic mechanism of cathode emission is the migration of free electrons through the cathode coating and into space. This theory is not yet completely developed, nor is the mechanism fully explained. Not until a complete and accepted emission mechanism is available can a tube manufacturer hope to control electron tube emission between lots in production over long periods of time and between tubes in a single production run."

Anyone familiar with electron tube physics would recognize the above as an inexcusable misstatement of fact.

Page 8-23:

Figure 8-20 shows ionic grid current increasing to a maximum and then decreasing as plate current in the electron tube is increased. That ionic grid current is directly proportional to plate current has been an established fact of physical electronics for many decades.

Page 8-24:

"Contact potential is rather difficult to define because it means different things to different people. For the moment, assume that it arises from the same mechanism as having two dissimilar materials in contact at a high temperature. The contact potential at the grid surface varies throughout the life of the tube. The rate of variation is accelerated by higher heater voltages. The cause of this variation may be thought of as the result of the grid base material being slowly covered by free barium and strontium that have continuously evaporated from the cathode coating and condensed on the grid. The most commonly used measure of contact potential in the tube industry is the value of grid bias that must be used to reduce the grid current to 0.1 microamperes. From figure 8-20, it is obvious that this arbitrary contact potential is not independent of contributions from gas currents, leakage currents or emission currents existing in the tube under test."

The discussion of the term "contact potential" as used by tube manufacturers is inexcusably confused with "contact difference of potential" as used by physicists.

Page 8-27:

"Closed-spaced, high-performance tubes are more subject to wider initial tolerances, greater changes with life, more difficulties with leakage, gas, grid emission, and most of the other problems in uncontrolled detrimental properties that beset the tube industry. They are also more susceptible to microphonics, impact shock damage, and vibration fatigue. Used wisely in the relatively low-impedance, high-frequency applications for which they are designed, these tubes are an important asset to the equipment designer's bag of tricks. A close-spaced, high figure-of-merit tube should only be used where no other alternative exists. The two types (6AK5 and 6J6) that head the ARINC list of tube failures are close-spaced, high-performance types."

The ARINC General Report referred to in establishing the above statement clearly states that over-all failures of the 6AK5 are large in number because the type has the highest socket population of any military tube type. The ARINC report goes on to say that the actual 6AK5 failure rate per socket is one of the lowest of the military receiving tube types.

The above examples are given as illustrations. The list should not be considered comprehensive.

C. R. Knight
Aeronautical Radio, Inc.
Washington 6, D. C.

Recent Books

Industrial Research Laboratories of the United States. Published by the National Academy of Sciences, National Research Council, Publications Office, 2101 Constitution Ave., Washington 25, D.C. $10.00.

Keen, A. W., *Electronics, The Science of Electrons in Action.* The Philosophical Library, Inc., 15 E. 40 St., N.Y. 16, N.Y. $7.50.

Nord, Melvin, *Legal Problems in Engineering.* John Wiley & Sons, Inc., 440 Fourth Ave., N.Y. 16, N.Y. $7.50.

Operations Research for Management, Vol. II, ed. by J. F. McCloskey and J. M. Coppinger. The Johns Hopkins Press, Homewood, Baltimore 18, Md. $8.00.

1956 Who's Who in Electronics, ed. by R. A. Harris, Radio and Electronic Jobber News, Inc., 2775 S. Moreland Blvd., Cleveland 20, Ohio. $7.50.

Quality Control and Applied Statistics Abstracts, Vol. I, Issue I, ed., by R. S. Titchen, A. J. Rosenthal, Bruce Bollerman and Frank Nistico. Interscience Publishers, Inc., 250 Fifth Ave., N. Y. 1, N. Y. $60.00 per year.

Lord Rayleigh, *The Theory of Sound, Vols. I and II.* First American edition of the book which was first published in 1877. Dover Publications, 920 Broadway, N.Y., N.Y. $1.95 per volume.

Stockman, Harry, *Time-Saving Network Calculations, 2nd ed.* SER Co., 543 Lexington St., Waltham, Mass. $1.75.

Symposium on Monte Carlo Methods, ed. by H. A. Meyer. John Wiley & Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $7.50.

U.R.S.I. Proceedings of the XIth General Assembly, Vol. X, Part Two. General Secretariat of U.R.S.I., 42, rue Minimes, Brussels, Belgium. $2.00.

National Telemetering Conference

The 1956 National Telemetering Conference will be held at the Biltmore Hotel, Los Angeles, California, August 20–21, 1956. It is sponsored by the IRE, AIEE, ISA, and IAS.

Registration arrangements should be made with E. W. Robischon, 7660 Beverly Blvd., Los Angeles 36, Calif. The registration fee for members of the sponsoring organizations is $3; for non-members, $4. Admission to the banquet and cocktail party is $8, and luncheon reservations can also be obtained for $3.50.

Monday Morning

Session I

Systems I

A Wide Bandwidth Telemetry System, D. E. Henry, Sandia Corporation.

Advanced Design Telemetry for Vehicles Subjected to Severe Environmental Conditions, D. W. Blancher, Bendix Pacific.

Radio Frequency Link Design for Telemetering, Hans Scharla-Nielsen, Radiation, Inc.

Beacon Telemetering System, Stanley Berinsky, Stavid Engineering.

A Comparison of Transmission Quality in Multiplexed Telemetering Systems, H. S. McGaughan, Naval Ordnance Laboratory.

Session II

Data Processing

A New Development in the Processing of PDM Telemetry Data, W. E. Leever, Douglas Aircraft Company.

An Automatic Date Reduction System for Pulse Width Telemetry, F. T. Chambers III, Applied Science Corporation of Princeton.

A Method of Separating Vibration Data into Its Frequency and Amplitude Components, J. M. Hawk, White Sands Proving Ground.

A High-Speed, High-Accuracy Automatic Digital Data Reduction System for FM and PWM Telemetered Information, William Kroll, Lockheed Aircraft Corporation.

Monday Afternoon

Session III

Components & Equipment I

Temperature Measurements on High Speed Missiles, G. E. Reis, Sandia Corporation.

Measurement of Millivolt Data by Pulse Width Multiplexing, A. S. Westneat, Applied Science Corporation of Princeton.

Commutation and Amplification of Low Level Signals, W. J. Ross, Bendix Aviation.

A Method for Commutating Low Level Signals, Owen Ott, Electro-Mechanical Research, Inc.

Frequency Stability Investigations on FM/FM Telemetering Equipment, J. W. Prast and K. F. Hartmann, Bell Aircraft Corporation.

Session IV

Panel on "Crowding of the Telemetering Frequency Bands."

Chairman: G. S. Shaw, Radiation, Inc.

Monday Evening

Banquet

Welcome: H. L. Hull, Chairman 1956 NTC, Farnsworth Electronics Co.

Address: "Telemetering and Pointer Readings," by C. R. Pion, Vice-President for Research, Avco Manufacturing Corp.

Tuesday Morning

Session V

Systems II

An Integrated Sub-Miniature Digital Airborne Ground Data Transmission System, B. M. Gordon, Epsco Incorporated.

Pulse Coded Messages Over Radar Beacon Equipments, Kurt Merl and Richard Rabin, Ford Instrument Company.

A Comparison of Encoding Techniques for Telemetry and Data Handling, R. A. Runyan, Electro-Mechanical Research, Inc.

A Wide Band Radio Link Telemetering System, T. D. Warzecha, Convair.

Session VI

Recorders

A High-Speed, General-Purpose, Mobile Magnetic-Tape Data Recorder for Use with a Digital Computer, E. R. Pelta, The Rand Corporation.

Direct Writing Continuous Recording Above 100 Cycles, H. I. Chambers and J. C. Riedel, Consolidated Electrodynamics Corp.

A Miniature Digital Recorder, C. P. Hedges, Aerophysics Development Corp.

A Small, Fast Digital Data Printer, E. A. Hilton and Harold Elliott, Hewlett-Packard Company.

Tuesday Noon

Luncheon

Address: "Ballistic Missile Telemetry," E. R. Toporeck, Director of the System Test Staff, The Ramo-Woolridge Corp.

Tuesday Afternoon

Session VII

Components & Equipment II

Receiver Design Considerations for Future Telemetry Requirements, R. E. Grimm, Nems-Clarke, Inc.

A Continuously Tunable Discriminator for FM-FM Telemetering Systems, G. E. Tisdale, Electro-Mechanical Research, Inc.

A Transistorized Telemetering System as a "Watch Dog" on Guided Missile Performance, F. M. Riddle, Jet Propulsion Laboratory.

Transistors Applied to an Operational FM-FM Telemetry System, C. B. McCampbell, R. H. Gablehouse, P. S. Scheele, R. P. Matthews, Sandia Corporation.

Session VIII

Panel on "Telemetry Technique in Flight Testing of Aircraft."

Chairman: J. J. Dover, Air Force Flight Test Center, Edwards, California.

Western Electronic Show and Convention

Ambassador Hotel and Pan Pacific Auditorium, Los Angeles, California
August 21–24, 1956

Approximately 30,000 engineers, scientists and business representatives of the electronics industry will be attending the Western Electronic Show and Convention in Los Angeles. 706 exhibits will be shown.

Lee DuBridge, Calif. Inst. of Technology, will address the convention on "Adventures in Science."

WESCON, held in alternate years at Los Angeles and San Francisco, is sponsored jointly by the West Coast Electronic Manufacturers' Association, and the Los Angeles and San Francisco IRE Sections. The general chairman of Wescon this year is C. F. Wolcott, and Willard Fenn heads the technical program committee in charge of presenting over two hundred technical papers.

Five field trips, an all-day cruise on the United States aircraft carrier *USS Bremmerton*, a tour of medical electronics facilities at UCLA, a trip to Pacific Semiconductors to observe the manufacture of miniature diodes and transistors, a tour of the radio division of Hoffman Electronics Corporation, and an inspection of the color television studios of the Columbia Broadcasting Company, are also scheduled.

The cocktail party will be staged between 5:30 and 7:30 P.M., August 21, at the convention center of the Ambassador Hotel. W. B. Knight, D. D. Dressen and J. F. O'Halloran of the cocktail party committee suggest that those planning to attend write the WESCON Business Office, 344 N. La Brea Ave., Los Angeles 36, Calif., for reservations. Tickets are $5.50 per person.

Women's activities will feature trips to Disneyland and Forest Lawn, a luncheon and aquacade at Hotel Ambassador.

TUESDAY, AUGUST 21

9:30 A.M.–11:30 A.M.

Session 1
Earth Satellite

Chairman: Capt. D. P. Tucker, USN, Office of Naval Research.

The Professional Group on Military Electronics, Capt. C. L. Engleman, USN.

The U. S. Earth Satellite Program Project VANGUARD, M. W. Rosen, Naval Research Laboratory, and D. J. Markcarian, Glenn L. Martin Company.

Session 2
Microwave Theory and Techniques

Chairman: Charles Chandler, Gilfillan Brothers, Inc.

Applications of a Waveguide Slot Coupler with Unity Coupling, R. W. Clapp, Hughes Aircraft Company.

The Effect of Mode Filters on the Circular Electric Wave Transmission Properties of a Long Multi-Mode Circular Waveguide, W. D. Warters, Bell Telephone Laboratories.

Equivalent Circuits for Thin Waveguide-Coaxial Junctions, C. A. Levis, Ohio State University.

Exponential Transmission Lines as Transformers and Cavities, R. N. Ghose, Radio Corporation of America.

Session 3
Engineering Management

Chairman: E. F. Carter, Director, Stanford Research Institute.

Management of Large Research and Development Organizations, N. I. Hall, Vice-President, Hughes Aircraft Co.

People, Things and the Engineer, J. F. Gordon, Chief Development Engineer, Helipot Division, Beckman Instruments, Inc.

Future of Engineers in the Management of Industrial Enterprise, A. C. Fontaine, Vice-President, Bendix Aviation Co., Inc.

Situations that Effect the Productivity of Engineers, M. C. Batsel, Chief Technical Administrator, Defense Electronic Products, RCA.

Session 4
Medical Electronics

Chairman: Alexander Kolin, University of California, Los Angeles.

Exposure Hazards from Cosmic Radiation in Flight in Extra-Atmospheric Regions, Hermann J. Schaefer, Naval School of Aviation Medicine.

Progress in the Field of Xerography, Speaker from General Electric Co.

Methods of Measuring Blood Pressure and Blood Flow, J. P. Meehan, University of Southern California.

Session 5
Tube Techniques

Chairman: M. C. Long, Hughes Aircraft.

Development of a Line of Microminiature Ceramic Tubes, R. E. Moe, G. E. Co.

Noise Performance of a Microminiature Triode, J. W. Rush, G. E. Co.

Flat Display Device with Matrix Display Selection, B. A. Findeisen, G. E. Co.

A Flash Coding Tube, R. W. Sears, BTL.

Top, *left to right*—B. S. Angwin, Convention Vice-Chairman; C. F. Wolcott, 1956 WESCON General Chairman; E. P. Gertsch, Show Vice-Chairman. Below, *left to right*—D. B. Harris, W. E. Noller, N. E. Porter and N. H. Moore, members of the 1956 WESCON Board of Directors, working on last-minute plans for the convention.

Ceramic Seals for Microwave Tubes, Curtis Ward, Varian Associates.

New Techniques Used in the Development of a 40 Watt Ceramic UHF Power Tetrode, M. B. Shrader, RCA.

Session 6

Not scheduled

2:00 P.M.–4:30 P.M.

Session 7
Propagation

Chairman: R. B. Muchmore, The Ramo-Wooldridge Corporation.

Reflection of a Transient Electromagnetic Wave from a Conducting and Anisotropic Medium, J. R. Wait, National Bureau of Standards.

The Prediction of Tropospheric Propagation from Meteorological Data, L. J. Anderson and J. B. Smyth, Smyth Research Associates.

Some Observations of Antenna-Beam Distortion in Trans-Horizon Propagation, A. T. Waterman, Jr., N. H. Bryant, and R. E. Miller, Stanford University.

A Study at 1046 Megacycles of the Reflection Coefficient of Irregular Terrain at Low Angles of Incidence, R. E. McGavin and L. J. Maloney, National Bureau of Standards.

Radar Terrain Return at Near-Vertical Incidence, R. K. Moore and C. S. Williams, Jr., University of New Mexico.

Session 8
Microwave Theory and Techniques—Active Elements

Chairman: Kiyo Tomiyasu, General Electric Microwave Laboratory.

Recent Developments on Transmit-Receive Switch Tubes, L. W. Roberts, Bomac Laboratories.

Characteristics of Crystal Video Receivers Employing B-F Pre-Amplification, W. E. Ayer, Applied Electronics Laboratory, Stanford University

Coupled-Cavity Stabilization of Klystrons, Maurice St. Clair, Varian Associates.

Burnout Experiments on S-Band and X-Band Crystals, Ben Hecht, Bomac Laboratories, Inc.

Session 9

Circuit Theory—Frequency Domain Techniques

Chairman: Louis Weinberg, Hughes Aircraft Company.

Realization of Driving Point Impedance Functions without Mutual Inductance, R. H. Pantell, Stanford University.

RLC Transfer Function Synthesis, E. C. Ho, Logistic Research Inc. and University of California, Los Angeles.

Synthesis of Grounded Two-Element-Kind Symmetrical Networks, P. M. Lewis, Massachusetts Institute of Technology.

Some Simplifications for Analysis of Linear Circuits, G. L. Matthei, Ramo-Wooldridge Corporation.

Network Design by First Order Predistortion Technique, C. A. Desoer, Bell Telephone Laboratories.

Session 10

Instruments

Chairman: C. P. Spaulding, G. M. Giannini and Co.

A Practical Application of Phase Measuring Techniques to Precision Angle and Distance Measuring Equipments, W. T. Thompson Cubic Corp.

A Multi-Pressure Measuring and Recording System for Wind Tunnels, M. Bain, Jet Propulsion Lab.

Electromagnetic Flowmeters—Operating Characteristics and Applications, Eugene Mittelmann, Consulting Engineer.

Three-D Magnetic Flux Meter, M. Muller and R. Feldt, Federal Telephone and Radio Co.

Session 11

Television Receivers

Chairman: William Milwitt, Industry Service Laboratory, Radio Corporation of America.

Stagger-Tuned Transistor Video Amplifiers, V. H. Grinich, Stanford Research Institute.

Analytical Approaches to Local Oscillator Stabilization, W. Y. Pan and D. J. Carlson, Radio Corporation of America.

Retrace Driven Deflection Circuit, W. B. Guggi, Stanford Research Institute.

Graphic Derivation of the Chromaticity Diagram, E. L. Michaels, Hughes Aircraft Co.

Session 12

Military Electronics

Chairman: J. F. Byrne, Motorola Research Laboratory.

Electronic Circuit Standardization, J. H. Muncy and G. J. Rogers, National Bureau of Standards.

Application of Solar Furnaces to High Temperature Research, Lt. Col. R. H. May, Armed Forces Special Weapons Project.

Nuclear Weapons Effects on Communication Systems, Jack Eggart, Signal Corps Engineering Laboratories.

Silicon Solar-Cell, M. B. Prince, National Semiconductors Products.

WEDNESDAY, AUGUST 22

9:30 A.M.—Noon

Session 13

Antennas I

Chairman: V. R. Rumsey, University of Illinois.

Radiation and Diffraction Problems Involving Wedges and Cones, L. B. Felsen, Polytechnic Institute of Brooklyn.

External Fields Produced by a Slot on a Cone, G. Held and G. Hasserdjian, University of Washington.

A Slot Analog of the Loop-Dipole D-F Antenna, F. D. Clapp and H. Masuda, University of California, Berkeley.

The Effects of Thin Conductive Coatings on Low-Frequency Aircraft Antenna Performance, C. W. Steele, Stanford Research Institute.

The TACAN Antenna, A. Casabona, Radio Navigation Laboratory, Federal Telecommunication Laboratories.

Session 14

Control Theory and Methods

Chairman: W. R. Evans, North American Aviation.

Consideration Regarding Design of Stabilizing Elements for Control Systems, T. H. Chin, University of Pittsburgh.

Solution of Statistical Problems by Differential Operator Approach, R. L. Cosgriff, Ohio State University.

Describing Function Technique Applied to the Analysis of a Non-Linear Control System Used in Space Stabilizing a Missile In Roll, Leonard Atran, Westinghouse Electric Corporation.

Statistical Theory and Analysis of Multiply-Instrumented Control Systems, R. M. Stewart, Jet Propulsion Laboratory.

A Time-Domain Synthesis of Optimum Extrapolators, C. W. Steeg, Dynamic Analysis and Control Laboratory, MIT.

Session 15

Transistor Circuits—Linear, Low and High Frequency Applications

Chairman: W. W. Wells, North American Aviation Co.

Design of Negative Feedback Transistor Amplifiers for Hi-Fi Equipment, (with demonstration of equipment). H. R. Lowry, General Electric Co.

Silicon Transistor Video Amplifier, R. E. Leslie, Sperry Gyroscope Co.,

Stability and Power Gain of Tuned Transistor Amplifiers, A. P. Stern, Electronics Lab., General Electric Co.

High Frequency Equivalent Circuits for Junction Triodes, R. M. Scarlett, Stanford University.

Some Circuit Applications of Silicon Tetrodes, R. R. Webster and R. F. Stewart, Texas Instruments, Inc.

Session 16

Cybernetics

Chairman: Don Taylor, University of California.

Spatial Perception, Robert Tscheirgi, UCLA.

The Use of Digital Computers in the Study of Neural Networks Models, Stanley Frankel, Conoco Corporation.

Engineering Studies of Sensory Organs, Joseph Hirsch, Aerophysics Corporation.

Cybernetics Past, Present and Future, Leonard Gardner, Roketdyne.

Session 17

Microwave Tubes I

Chairman: L. M. Field, Hughes Aircraft Co.

Design of Laminar Flow Electrostatically Focused Electron Beams, W. M. Mueller, University of California, Berkeley, and Hughes Aircraft Co.

Growing Waves in Magnetically Focused Electron Beams, R. W. Gould, California Institute of Technology.

Gain of a Low Level Signal in the Presence of a Large Signal, H. L. McDowell, BTL.

Noise and Spurious Oscillations in Backward Wave Oscillators, W. Rorden and C. Conner, Varian Associates.

Gridded Pulse TWT Serrodynes for Application to Continuously Coherent MTI Radars, R. M. Whitehorn and H. Mandoli, Varian Associates.

A Backward-Wave Oscillator for the 4 MM Wavelength Region, C. F. Hempstead, Bell Telephone Laboratories.

Session 18

Reliability—Organization, Systems and Equipment

Chairman: Bernard Hecht, Consultant.

Organizing for Reliability, A. M. Okun and J. Cohen, Bell Aircraft Corporation.

Price of Reliability in Airborne Electronic Equipment, A. H. Wulfsberg, Collins Radio Co.

Missile Unreliability Cost Evaluation, A. L. Stanly and J. Tampico, Associated Missile Products.

Calculations of the Risk of Component Applications in Electronic Systems, J. A. Connor, RCA.

Reliability as a Management-Engineering Responsibility, H. S. Hansen, Servo Mechanisms, Inc.

2:00 P.M.–4:30 P.M.

Session 19

Program to be arranged

Session 20

Control Mechanisms and Automation

Chairman: E. M. Grabbe, Ramo-Wooldridge Corporation.

Mechanization and Application of Vertical Reference System, G. H. Singer, Kearfott Company.

The Eye as a Control Mechanism, R. B. Lockard, U. S. Naval Ordnance Test Station.

Automatic Determination of Missile Air Foil Characteristics in Mass Production, E. L. Watkins, Convair.

A New Digital Path Control System for Machine Tools, Jack Rosenberg, Electronic Control Systems.

Session 21

Circuit Theory—Switching Theory Topology and Time Domain Synthesis

Chairman: L. A. Pipes, University of California, Los Angeles.

Switching Theory, D. A. Huffman, MIT.

Some Aspects of the Network Analysis of Sequence Transducers, J. M. Simon, Sperry-Rand Corporation.

Some Topological Considerations in Network Theory, F. Reza, Syracuse University.

Network Synthesis for Prescribed Impulse Response Using A Real Part Approximation, R. A. Pucel, Raytheon Manufacturing Company.

Time Domain synthesis by Delay Line Analogy, M. Strieby, Ramo-Wooldridge Corporation.

Session 22
Cybernetics Symposium

Chairman: Leonard Gardner, Rocketdyne.

The Philosopher's Approach to Cybernetics, Abraham Kaplan, UCLA.

Studies of Neural Physiology in the Learning Processes, J. A. Gingerelli, UCLA.

Panel discussion with the following panel members: Robert Tschirgi, Stanley Frankel, Joseph Hirsch, Leonard Gardner, Abraham Kaplan, and J. A. Gingerelli.

Session 23
Microwave Tubes II

Chairman: H. R. Johnson, Hughes Aircraft Co.

Characteristics of Modern External Tuning Cavity Reflex Klystrons, T. Moreno, Varian Associates.

A Rugged 8 MM Reflex Klystron, W. G. Abraham and F. L. Salisbury, Varian Associates.

An X-Band Folded Line Type Backward-Wave Oscillator, R. H. Winkler, Cascade Research Corporation.

A Demountable Tube for Gas Discharge Microwave Detection Studies, H. Farber, Polytechnic Institute of Brooklyn,

Means for Electronically Controlling the Current from a Magnetron Cathode, J. S. Needle, Northwestern University.

Design and Calculation Procedures for Low-Noise TWTs, G. Wade, R. W. DeGrasse and L. D. Buchmiller, Stanford University.

A Rugged 915 MCCW Magnetron for Electronic Cooking, P. C. Gardiner, G. E. Co.

Session 24
Component Reliability and Test Methods

Chairman: G. H. DeWitz, Hoffman Laboratories.

The Unreliable Universal Component, M. A. Acheson, Sylvania Electric Products, Inc.

Improved Guided Missile Tube Reliability, Alfred Blattel, Raytheon Mfg. Co.

Tube Failure Rate Variations, M. B. Feyerhern, RCA,

Cine-Radiography, A New Testing Concept of Component Reliability, W. H. Grumet, Rototest Labs.

Thursday, August 23

9:30 a.m.—Noon

Session 25
Panel on Education

Moderator: George Tenny, Vice-President of McGraw-Hill.

Members: Frederick Terman, Dean of Engineering of Stanford University, T. L. Martin, Jr., Chairman of the Electrical Engineering Department of the University of Arizona, Dan Nobel, President of Motorola, Burgess Dempster, President of Electronic Engineering Company.

Session 26
Computers: Analog Techniques

Chairman: M. E. Mohr, Ramo-Wooldridge Corporation.

Magnetic Amplifier Analog Computation Techniques, H. W. Patton, Airpax Products Company.

Wide-Dynamic Range Analog Integrator, George Myers, Rome Air Development Center, Air Research and Development Command.

A New Computer for Complex Functions, M. L. Morgan, Electro-Measurements, Inc.

Automatic Computation of Fourier Expansion Coefficients by Analog Means, R. W. Hubbard and W. E. Johnson, National Bureau of Standards.

Session 27
Transistor Circuits—Switching and Computer Applications

Chairman: M. S. Kesselman, Hughes Aircraft Company.

Critical Survey of Fundamentals of Junction Transistor Pulse and Switching Circuits, D. O. Pederson, University of California, Berkeley.

Maximum Efficiency Switching Circuits, R. H. Baker, Lincoln Laboratory, M.I.T.

Transistor Circuitry for Analog to Digital Conversion, F. H. Blecher, Bell Telephone Laboratories.

Session 28
Instrumentation Techniques

Chairman: F. S. Atchison, U. S. Naval Ordnance Lab.

Distributed-Parameter Variable Delay Lines Using Skewed Turns for Delay Equalization, F. D. Lewis and R. M. Frazier, General Radio Co.

Application of Scale Factors in Data Recording Systems, T. L. Greenwood, Redstone Arsenal.

Adjustable Pulse Width High Voltage Power Supplies by Victor Wouk, Beta Electric Corp.

Transistorized Logarithmic Time-and-Amplitude Quantizer, Euyen Gott, Johns Hopkins University.

Session 29
Antennas II

Chairman: J. T. Bolljahn, Stanford Research Institute.

Relation of Scattering Problems to Radiation from Simple Shapes, K. M. Siegel, Willow Run Laboratories, University of Michigan.

Launching Efficiency of Wires and Slots for a Dielectric Rod Waveguide, R. H. DuHamel and J. W. Duncan, Electrical Engineering Research Laboratory, University of Illinois.

Analysis of a Shunt Excited Airframe Antenna, T. G. Dalby, Boeing Airplane Company.

A New Method for Optimum Yagi Design, H. W. Ehrenspech and Poehler, Air Force Cambridge Research Center.

Session 30
High Temperature Component Parts

Chairman: Floyd Paul, Castell Corp.

Miniature High Altitude and High Temperature Connectors, C. H. Stuart and R. F. Dorrell, American Phenolic Corporation.

Impregnation of Toroids for High Temperature Service, E. O. Deimel, G. E. Co.

Typical Expected Performance Characteristics of Extreme Temperature Range Tantalum Capacitors, J. W. Maxwell, P. R. Mallory.

Some Basic Physical Properties of Silicon and How They Relate to Rectifier Design and Application, G. P. Finn, Sarkes-Tarzian.

2:00 p.m.–4:30 p.m.

Session 31
Reliability of Electronic Tubes

Chairman: To be announced.

White Noise Vibration Test for Electronic Tubes, J. D. Robbins, Sylvania Electric.

Environment Effects of Vacuum Tube Life, H. D. Pleak, Sylvania Electric.

Increased Reliability of Guided Missile Tubes Through Comprehensive Quality Control, H. H. Hoyle and N. J. Davis, Raytheon Mfg. Co.

New Filamentary Tubes of High Reliability, Ross Wood, Raytheon Mfg. Co.

Session 32
Computers: Solid State Switching Devices

Chairman: L. L. Kilpatrick, Autonetics, North American.

Magnetic Cores as Generalized Nonlinear Components, W. H. Bridge, Airtronic Research, Inc.

Pulse Response Properties of Magnetic Materials for Switching Applications, J. D. Childress, General Ceramics Corp.

High-Current Switching Applications of Low Power Transistors, Chaang Huang and W. F. Palmer, Sylvania Electric Products, Inc.

High-Gain Transistor for Very High Switching Speeds, J. B. Angell, Philco Corp.

Session 33
Circuit Theory—Transistor and Active Circuits

Chairman: R. D. Middlebrook, California Institute of Technology.

Power Transistor Specifications for Circuit Applications, H. T. Moore, Minneapolis-Honeywell Regulator Company.

Transistor Bootstrap and Miller Sweep Circuit, J. S. Sherwin, University of California, Berkeley.

An All Transistorized Pulse Generator, E. J. Fuller, Sperry Gyroscope Company.

Negative Impedance Converter Design and Its Use in Active Circuits, A. I. Larky, Stanford University.

A D-C Negative Immittance Converter, M. A. Karp, Applied Physics Laboratory, Johns Hopkins University.

Session 34
Broadcast Transmission Systems

Chairman: John Knight, KRCA—National Broadcasting Co.

The Aplication of Modern Techniques to the Determination of Service Areas of Television Stations in Smooth and in Mountainous Terrain, A. E. Cullum, Jr., Earl Cullum, Jr. and Associates.

Achievement of Practical Tape Speed for Recording Video Signals, C. P. Ginsburg, Ampex Corporation.

A Transistorized Television Camera Chain, W. Ussler, R. J. Decredico, G. W. Trebing, and J. W. Smiley, RCA.

Design Considerations for a High Quality Transistorized Program Amplifier for Remote Broadcast Use, K. J. Birch, Gates Radio Company.

Session 35
Microwave Theory and Techniques—Ferrites and Test Equipment

Chairman: Eric Strumwasser, Hughes Aircraft Company.

Coupling Through an Aperture Containing Ferrites, D. C. Stinson, University of California.

Microwave Frequency Doubling in Ferrites from 9 to 8 KMC, P. H. Vartanian, J. L. Melchor, W. P. Ayres, Electronic Defense Laboratory of Sylvania Elec. Prod., Inc.

A Ferrite Image Rejection Filter for Use in S-Band Microwave Receivers, J. H. Burgess, Electronic Defense Laboratory of Sylvania Electric Products, Inc.

Calorimeters for Centimeter and Millimeter Waves, Sam Hopfer, Polytechnic Research and Development Company.

Session 36
Environmental Effects on Component Parts

Chairman: A. W. Rogers, Signal Corps Engineering Labs.

Environmental Testing of Precision Potentiometers, Shultz-Green, Helipot Corp.

Capabilities of Hot-Molded Composition Resistors, A. C. Pfister, Allen-Bradley Co.

Hermetic Sealing of Precision Potentiometers, I. W. Braun, Circuit Instrument Incorporated.

The Effects of Nuclear Radiation on Electronic Components, C. C. Robinson, W.A.D.C.

Friday, August 24
9:30 A.M.—Noon

Session 37
Semiconductors I

Chairman: J. W. Peterson, Pacific Semiconductors.

Semiconductor Materials and Properties, P. E. Stello, D. M. Van Winkel, and J. D. Turner, Hughes Aircraft.

Developments in Transistor Physics, L. B. Valdes, Shockley Semiconductor Labs.

High Current Amplification—Forward, and Reverse—At High Collector Current in P-N-P Transistors, A. P. Kordalewski General Electric Company.

Very High Power Transistors with Evaporated Aluminum Electrodes, Gene Strull and H. W. Henkels, Westinghouse Electric Corporation.

Silicon Alloy Power Transistors, R. E. Anderson, Texas Instruments.

Session 38
Computers: Advanced Design Techniques

Chairman: J. L. Barnes, Systems Laboratories Corp.

The Effect of a Transverse Field on Switching Rates of Magnetic Memory Cores, T. D. Rossing and S. M. Rubens, Remington Rand Univac.

High Speed Coincident-Flux Magnetic Storage Principles, E. W. Bauer and L. P. Hunter, International Business Machines Corp.

Investigation of Problems Using Analog-Digital Computer Combinations, H. E. Salzer, Convair.

The Synthesis and Analysis of Digital Systems by Boolean Matrices, J. O. Campeau, Hughes Aircraft Company.

Session 39
Radar Systems

Chairman: F. J. Nichols, Sprague.

Taxi Radar—Eyes of the Tower, The Radar Approach Control (RAPCON) Center, D. R. Kirschner, Rome Air Development Center.

Extending Radar Range Timing Techniques, A. I. Mintzer, R.C.A.

TACAN Bearing and Distance Measurement Techniques, J. B. Majerus and K. W. Porter, Collins Radio Company.

Session 40
Vehicular Communications I

Chairman: Newton Monk, Bell Telephone Labs.

A Communication System for the New York Throughway, D. S. DeWire, New York Telephone Company.

Transistorized Communications Receivers, Seymour Schwarts, Massachusetts Institute of Technology.

VHF Radio Coordinated Traffic Light Control System, E. W. Hammel, General Electric Company.

VHF-UHF Communications System Interference Reduction through Use of Selective Filters, M. W. Caquelin, Collins Radio Company.

Session 41
Antennas III

Chairman: M. D. Adcock, Hughes Aircraft Company.

Fundamental Problems in the Antenna Field, S. Silver, University of California, Berkeley.

Scanning Lens Design for Minimum Mean-Square Phase Error, E. K. Procter and M. Rees, General Electric Company.

A Duplexer for Sweep-Frequency Pulse Transmitters, R. Silberstein, National Bureau of Standards.

The Triport Conical Scan Antenna, E. Wantuch and L. A. Kaiser, Raytheon Manufacturing Company.

Session 42
New Trends in Components

Chairman: R. H. Baker, R.C.A.

The Impact of New Electronics on Mercury Battery Designs, J. L. Dalfonso, P. R. Mallory and Company.

Criteria For Selection of Magnetron Beam Switching Tube as a Circuit Component, S. Kuchinsky, Haydu Brothers.

Application of Large Capacitors for Energy Storage, D. F. Warner, General Electric Company.

Film Type Precision Resistors, B. Solow and C. Wellard, International Resistance Co.

2:00 P.M.–4:30 P.M.

Session 43
Semiconductors II

Chairman: D. M. Van Winkle, Hughes Aircraft Co.

Point Contact Diode Theory, Melvin Cutler, Hughes Aircraft Co.

Component Improvement, Germanium and Silicon Low Power Rectifiers, G. N. Hall, General Electric Company.

A Miniature Silicon Diode for Both Power and Circuit Applications, A. L. Rossoff, Radio Receptor Company, Inc.

A New UHF Transistor Structure, W. W. Gartner, Evans Signal Laboratory.

Transistors with Vacuum Tube Properties, M. N. Ross and H. E. Hollmann, National Aircraft Corporation.

Session 44
Computers: Data Gathering and Presentation Systems

Chairman: G. W. Brown, International Telemeter Corporation.

Electrofax Printer, A Continuous Direct Dry Process Enlarger for Digital Computing Systems, H. G. Reuter, Jr., Radio Corporation of America.

Ultratype Camera, A High-Speed Electron Optical Printer for Digital Computing Systems, A. M. Spielberg, S. R. Parker, and K. G. Kaufmann, Radio Corporation of America.

A High Speed, Precision, Large Quantity Data Processing System: IDIOT II, M. L. Klein, Rocketdyne Division, North American Aviation.

A Centralized Data Processing System for the Air Force Flight Test Center, O. F. Vogel, Electronic Engineering Company of California.

Session 45
Airborne Electronics

Chairman: C. A. Ripinski, Electronic Industries.

Thermal Design and Evaluation of An Airborne Electronic System, R. E. Burton, Collins Radio Company.

Flight Control Systems for Jet Transports, H. Miller and R. H. Wagner, Sperry Gyroscope Company.

Electronic Problems Encountered in High-Speed Commercial Jet Aircraft, Dick Hedges, Douglas Aircraft Company.

Bearings and Their Properties as a Design Consideration in Electronic Systems, H. F. Stern, Industrial Tectonics, Inc.,

Session 46
Vehicular Communications II

Chairman: Maurice Kennedy, County of Los Angeles.

Communication with Moving Trains in Tunnels, Newton Monk, Bell Laboratories.

New Developments in Two-Way Communications, Angus MacDonald, Motorola.

Railroad Communications, D. L. Kesselhuth, Bendix.

Session 47

Information Theory

Chairman: R. H. DeLano, Systems Laboratories Corp.

Synthesis of the Linear Time-Varying Predictor for Stationary Signals, Cheng Ling, Minneapolis-Honeywell Regulator Company.

The Relationship of Sequential Filter Theory to Information Theory and Its Application to the Detection of Signals in Noise by Bernoulli Trials, Herman Blasbalg, Johns Hopkins University.

Theory of Weighted Smoothing, L. A. Ule, Gilfillan Bros., Inc.

The Response of a Phase-Locked Loop to a Sinusoid Plus Noise, S. G. Margolis, Jet Propulsion Laboratory.

Autocorrelator for Radioactive Sample Noise Generator, G. W. Anderson and J. E. Murrin, J. B. Rea Co.

Session 48

Production Consideration of Electronic Equipment

Chairman: Marvin Whitney, Hoffman Electronics Corporation.

Problems of Semi-Automatic Assembly of Electronic Test Equipment, C. S. Selby, Hewlett-Packard Company.

Eyelet Failure in Etched Wiring, W. J. Hodges, Hughes Aircraft Company.

Mechanical Design Considerations in the ERMA System, R. W. Melville, Stanford Research Institute.

The Selection of Coatings for Printed Circuits, R. A. Martel and L. J. Martin, Hughes Aircraft Company.

Sixth Annual Fall Symposium

PITTSBURGH, PENNSYLVANIA

SEPTEMBER 14–15, 1956

SPONSORED BY THE PROFESSIONAL GROUP ON BROADCAST TRANSMISSION SYSTEMS

MELLON INSTITUTE AUDITORIUM

FRIDAY, SEPTEMBER 14

8:30 a.m.–9:30 a.m.

Registration

9:30 a.m.–12:15 p.m.

Moderator: Lewis Winner, Publications Director, Bryan Davis Publishing Co.

A Television Aural Transmitter Stand-by System, Benjamin Wolfe, WAAM, Baltimore, Maryland.

Non-Rigid Transmission Lines in Broadcasting, H. R. Kaiser, WWSW, Pittsburgh, Pa.

Remote Control of High Power and Directional Antenna Systems, G. W. Bartlett, NARTB, Washington, D.C.

Automatic Gain Control in TV Automation, Diehl, Hoffman and Shepard, General Electric Company.

Conelrad Developments for the Broadcaster, Ralph Renton, Federal Communications Commission.

A Method of Preventing Burn-in On I.O. Tubes, J. T. Wilner, WBAL-TV, WISN, Hearst Corporation.

2:00 p.m.–5:30 p.m.

Moderator: To be announced.

A Station Approach to TV Field Strength Measurements, R. M. Crotinger, WHIO-TV, Dayton, Ohio.

Measurement of Service Area for Television Broadcasting, R. S. Kirby, Boulder Laboratories, National Bureau of Standards.

Modern Techniques for the Determination of Service Areas of Television Broadcast Stations, A. E. Cullum, Jr. and T. A. Wright, Consulting Engineers, Dallas, Texas.

Color TV Equipment Measurements—Demonstration, J. R. Popkin-Clurman, Telechrome, Inc.

Color Kinescope Recording on Embossed Film, C. H. Evans and H. B. Smith, Eastman Kodak Company.

Video Tape Recording Development and Techniques, speaker to be announced.

Webster Hall Hotel

6:30 p.m.

Cocktails

8:00 p.m.

Banquet

Toastmaster: G. H. Brown, RCA Princeton Laboratories.

SATURDAY, SEPTEMBER 15

9:30 a.m.–12:15 p.m.

Moderator: R. D. Chipp, Allen B. DuMont Laboratories.

A Panel on Development of New Television Studio Facilities for Modern Operating Requirements.

Members: W. J. Purcell, WRGB, J. B. Epperson, WEWS, C. F. Daugherty, WSB-TV, D. M. Weise, WTTW-TV.

2:00 p.m.–5:30 p.m.

Moderator: To be announced.

Monitoring Mobile Unit for Television Service, Speaker from the Federal Communications Commission.

Sawtooth Testing of Audio Amplifiers, R. C. Hitchcock, Syntron Company.

Improvements in Television Picture Quality (with demonstration), Arthur Anderson, Westinghouse Research Laboratories.

Fifth Annual Industrial Electronics Symposium

SPONSORED BY THE PG'S ON INDUSTRIAL ELECTRONICS AND PRODUCTION TECHNIQUES, AND THE A.I.E.E.

HOTEL MANGER, CLEVELAND, OHIO

SEPTEMBER 24–25, 1956

Monday, September 24

8:00 a.m.

Registration

9:30 a.m.

SESSION I

ELECTRONIC EQUIPMENT DESIGN

Development of Interconnecting Wiring, D. J. Keller, Sperry Gyroscope Co., Great Neck, N.Y.

Human Engineering—An Aid to Improving Electronic Equipment, Maurice Rapaport, Stanford Research Institute, Menlo Park, Calif.

A Modern Concept of Electronic Packaging, R. P. Noble, Sandia Corp., Albuquerque, N.M.

Packaging of Transistorized Assemblies, A. A. Lawson and R. J. Simms, Melpar Inc., Falls Church, Va

12:30 p.m.

Luncheon

Speaker: General B. W. Chidlaw, Vice-President, Thompson Products, Cleveland, Ohio, on the subject "Electronics in Air Defense and Air Offense."

2:00 p.m.

SESSION II

AUTOMATIC PRODUCTION

Automation for Electronics, A. R. Gray, Glenn L. Martin Co., Baltimore, Md.

Manufacture of Wire-Spring Relays for Commercial Switching Systems, J. W. Rice, Western Electronic Co. Inc., Chicago, Ill.

Status of Standardization in Electronics Production and Machine Tool Controls, E. H. Bosman, International Business Machines, Endicott, N.Y.

Production Testing in the Automatic Factory, H. S. Dordick, Radio Corporation of America, Camden, N.J.

Tuesday, September 25
9:30 a.m.

SESSION III

TESTING, GAUGING, AND PROCESS CONTROL

The Industrial Electronics Concept, Eugene Mittelmann, Consulting Engineer, Chicago, Ill.

Variable-Speed Electronic Drives for Process Control, A. J. Humphrey, Reliance Electric and Engineering Co., Cleveland, Ohio.

Automatic Process Control with Radiation Gauges, W. H. Falkner, Jr. and G. F. Ziffer, Tracerlab, Inc., Boston, Mass.

Measurement and Control in a Large Steam-Turbine Generator Department, R. G. Goldman, Large Steam Turbine-Generator Department, General Electric Co., Schenectady, N.Y.

12:30 p.m.
Luncheon

IRE—PGIE Papers Award Presentation

2:00 p.m.

SESSION IV

DATA REDUCTION AND ANALYSIS

Data Reduction with a Large Scale Digital Computer, H. A. J. Grosch, General Electric Co., Cincinnati, Ohio.

Analog-Computer Study of an Electrical Constant Cutting-Speed Machine Tool Control Drive System, W. J. Bradburn, Louis Allis Co., Milwaukee, Wis.

Analog vs. Digital Techniques for Engineering Design Problems, D. B. Breedon, Westinghouse Corp., Pittsburgh, Pa.

A Tank Farm Data Reduction System, D. J. Gimpel, Panellit Inc., Chicago, Ill.

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5	Microwaves & Instrumentation **Sessions: 1, 26, 34, 47, 48, 54**	Instrumentation Microwave Theory & Techniques	2.75	6.60	8.25
6	Manufacturing Electronics **Sessions: 6, 8, 17, 27, 35, 44, 45, 52**	Component Parts Engineering Management Industrial Electronics Production Techniques Reliability & Quality Control	3.25	7.80	9.75
7	Audio & Broadcast **Sessions: 12, 13, 20, 21, 25, 55**	Audio Broadcast Transmission Systems	2.25	5.40	6.75
8	Aeronautical, Communications & Military Electronics **Sessions: 3, 4, 11, 15, 19, 31, 36**	Aeronautical & Navigational Electronics Communications Systems Military Electronics Vehicular Communications	2.75	6.60	8.25
9	Ultrasonics, Medical & Nuclear Electronics **Sessions: 2, 9, 18, 51**	Medical Electronics Nuclear Science Ultrasonics Engineering	1.50	3.60	4.50
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The following issues of "Transactions" have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

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Antennas and Propagation

Vol. AP-4, No. 2, April, 1956

News and Views

Contributions—Variational Principles for Electromagnetic Resonators and Waveguides—A. D. Berk

Variational expressions are presented for the propagation constants of a waveguide and for the resonant frequencies of a cavity directly in terms of the field vectors for situations not covered in the literature. These situations occur when the electromagnetic problem is not susceptible of a scalar formulation and are typified by the presence of inhomogeneous or anisotropic matter. The variational expressions are applied to several illustrative examples and the results are compared to known exact solutions. The variational expressions are also applied to the derivation of certain perturbation formulas, some of which were previously obtained by different methods.

Nonlinearity of Microwave Ferrite Media—N. G. Sakiotis, H. N. Chait, and M. L. Kales

Existing theories of propagation in magnetized ferrite media predict propagation constants which are independent of the rf field strength only if a number of restrictive conditions are satisfied. In general, however, the propagation constants are functions of the rf field strength and nonlinear propagation is to be expected. One of the conditions for linearity is that the rf magnetic field intensity be small compared to the static magnetizing field intensity. This condition can be violated when the peak power level of the wave incident on the ferrite medium is sufficiently high and indeed, it has been observed by the authors that ferrite loaded waveguides can become nonlinear at peak power levels as low as 1 kw.

Results are presented of a study of the behavior of ferrite loaded waveguides at 9,400 mc over input peak power levels from 0.1 kw to 100 kw which includes power levels commonly encountered in radar applications. The dependence upon input power level of the ferrite losses, phase shift, and rotation of the plane of polarization is described for a number of ferrite materials. The effect upon the degree of nonlinearity of the intensity of the static magnetizing field is discussed as well as the dependence upon the dimensions of the ferrite.

Diffraction of Electromagnetic Waves Caused by Apertures in Absorbing Plane Screens—H. E. J. Neugebauer

The perturbation field caused by holes in a plane, infinitely thin screen of arbitrary material on which electromagnetic radiation is incident, can be split up in two fields, one of which is symmetrical in the tangential components of the electric, the other in those of the magnetic vector. Both fields satisfy equations within the holes which are generalizations of Bethe's conditions for the perfectly conducting screen and which can be used to derive approximate solutions of Kirchhoff type. The solution for a screen which absorbs practically all of the incident radiation is essentially different from Kirchhoff's result for scalar radiation incident on a so-called perfectly absorbing screen.

Side-Lobe Suppression by Pattern Multiplication—Raymond Justice

It is shown that the properties of the radiation patterns of broadside and end-fire line radiators can be used to minimize the side-lobe level of the radiation patterns of uniformly excited rectangular arrays.

On the Conductance of Slots—J. R. Wait

The external radiation conductance G of narrow half-wave axial and transverse slots on a circular cylinder, a ribbon, a knife edge, and a right-angle wedge are computed. The method is based on finding the total power radiated assuming a sinusoidal variation of voltage along the slot. The theoretical values of G for the ribbon of width $2d$ are applied to the case of a half-wave slot cut in the narrow face of a rectangular waveguide which has a flush mounted plate or flange of width $2d$.

A Method of Analyzing Coupled Antennas of Unequal Sizes—C. A. Levis and C. T. Tai

The impedance parameters of coupled antennas are calculated by a variational method. The unknown coefficients pertaining to the trial functions for the current distributions on various elements are shown to satisfy a set of linear equations involving the excitation conditions, but the final formulas for the impedance parameters involve only the trial functions and the geometry. The method is directly applicable to the analysis of Yagi-Uda antennas, and the technique can be extended to coupled slots inside a waveguide or coupled radiating slots excited by waveguides.

Radiation Characteristics of the Spherical Luneberg Lens—E. H. Braun

Antenna Pattern Distortion by Dielectric Sheets—J. H. Richmond

The radiation pattern of an antenna covered with a radome can be determined from the following data: the antenna Fresnel-Zonefields in the absence of the radome, and the fields inside the radome produced by a plane-wave incident from an external source.

In certain cases the pattern calculations may be simplified without excessive loss of accuracy by using optical theory to describe approximately the plane-wave transmission through the radome, and neglecting scattering by the antenna in determining the fields inside the radome from the external plane-wave source.

Multiple Scattering by Randomly Distributed Obstacles—Methods of Solution—C. M. Chu and S. W. Churchill

Methods of solution of problems of multiple scattering of electromagnetic radiation by randomly distributed obstacles are evaluated with respect to generality, accuracy, and suitability for numerical calculations. Exact solutions are found to be limited to infinite regions and are difficult to apply for anisotropic scattering. Approximate solutions of the transport equation are limited to simple regions and are very complicated for anisotropic scattering. Approximate representation of the scattering process with the classical diffusion equation permits solution for almost all conditions, but the solution is inaccurate near the boundaries of the region of scattering. Approximate representation of the intensity by a number of discrete components permits solution for many geometries. Several two-component representations have been developed which yield more accurate results than diffusion theory for anisotropic scattering. A six-component model appears to be considerably more accurate for most geometries but requires machine computations.

Diffraction of Plane Electromagnetic Waves by a Rectangular Aperture—Michio Suzuki

The Amplitude Concept of an Electromagnetic Wave and Its Application to Junction Problems in Waveguides—J. A. Ortusi

Phenomenological Vector Model of Microwave Reflection from the Ocean—C. I. Beard, I. Katz, and L. M. Spetner

A model of one-way transmission of microwave electromagnetic signals over the ocean surface is developed from experiment. The received signal is described as a vector sum of a constant direct signal, a coherent reflected signal, whose amplitude and phase are fixed by geometry and sea state, and a fluctuating reflected component of random amplitude and phase. By interpreting experimental data in the light of this phenomenological model it has been possible to relate, quantitatively, the coherent and incoherent reflected signal and total signal to geometry and sea state. The results give support to the theoretical expression previously derived by Ament and others relating the coherent reflected signal to "apparent ocean roughness." In addition, the general shape of the curve relating the incoherent scattering to "apparent ocean roughness" has been established and its asymptotic value found.

Report on Comparative 100 MC Measurements for Three Transmitting Antenna Heights—A. P. Barsis and R. E. McGavin

This report evaluates measurements taken during August, 1952, at a frequency of 100 mc as transmitted from three transmitting sites at elevations ranging from 6,220 feet to 14,110 feet above mean sea level. Six receiving sites were used ranging in distance from approximately 50 miles to 620 miles from the transmitter sites. Results are presented in terms of hourly medians of recorded field intensity and their distributions, as well as the over-all median values and deviations derived from these distributions.

The Effect of Superrefractive Layers on 50–5,000 MC Nonoptical Fields—E. E. Gossard and L. J. Anderson

A body of radio data taken over an 80 nautical mile overwater link on four frequencies is analyzed, and the relationships to the vertical structure of the atmosphere determined statistically. The results are compared with the attenuation coefficients yielded by the theory for the first mode of a bilinear model as given by Furry.

It is found that observation agrees reasonably well with theory at about 100 mc, but departs considerably for lower or higher frequencies. It is found that for a given height of the top of the superrefractive layer, an optimum frequency generally exists for which attenuation is a minimum.

Communications—Effect of the Ground Screen on the Field Radiated from a Monopole—J. R. Wait

Admittance of Thin Antennas—Giorgio Barzilai

Contributors

IRE Transactions on Antennas and Propagation—Index to Volume AP-3—1955

Automatic Control

PGAC-1, May, 1956

Automatic Control and the IRE

Synthesis of a Nonlinear Control System—I. Flügge-Lotz and C. F. Taylor

The investigated nonlinear control system consists of a linear member with an ensemble of possible discrete combinations of proportional and derivative feedback around the linear member. The particular combination of proportional and derivative feedback employed at any instant is determined by a feedback switching circuit which is in turn operated by sensed binary information obtained from the output, output derivative, error and error derivative [namely, the signs (sgn) of these variables]. Techniques that are common to the digital computer field are used to implement this switching circuit.

For a linear member of second order the feedback circuit is comprised of four discrete values of proportional feedback and four discrete values of derivative feedback.

Simulation techniques have been used to study and evaluate the performance of the nonlinear control system and to compare it with a linear system for a wide variety of inputs. A sample of these experimental results is presented. The experiments show that the nonlinear system performance is much better than that of a linear system of comparable power handling capability.

Preliminary studies of a nonlinear control system of third order show that the basic idea can be extended to higher order systems.

Nonlinear Compensation of an Aircraft Instrument Servomechanism by Analog Simulation—D. Lebell

Combination of certain analog computer techniques with the direct synthesis philosophy applied to a particularly appropriate servo configuration results in a method for direct specification of compensator design. The method centers about a calibration technique which is simple and easily applicable to nonlinear control systems of considerable generality. Mathematical representation of the physical system is maintained at the realistic level typical of analog simulation.

A detailed description of the method is presented by means of its application to a practical example—nonlinear compensation of an instrument servomechanism. Equations governing the behavior of the unalterable components were written and substantiated experimentally. These equations uniquely determine the form of the ideal compensation for the designated system performance and system configuration. The actual unalterable elements are then employed as an analog simulator to "compute" the specified compensator characteristics. A compensator possessing these characteristics is assembled and over-all system performance obtained. An electronic analog computer is employed as a computer simulator for this work.

The majority of servo compensators designed in the past has been of the linear type. It is well known that this partiality is not due to superiority of the linear over the nonlinear kind but rather that linear designs have been much simpler to perform and nonlinear components hard to come by. The advent of analog computers, the extensive progress made in components development, plus the increasing need for "getting the most" out of available power transducers combine to encourage removal of the highly restrictive linearity condition.

Nonlinear design techniques available to the engineer have frequently suffered from mathematical complexity or computational drudgery on the one hand, or unrealistic simplification on the other. Recently, however, considerable progress has been made in the region between these extremes. This paper attempts to further such progress and its dissemination.

A Steady State Approach to the Theory of Saturable Servo Systems—J. C. Lozier

It is shown that simple saturable servo systems can have two modes of response to a given input signal—one a linear mode as predicted by linear feedback theory, and the other a saturated mode predictable from a large signal analysis presented here. This dual mode of response theory provides a reasonable explanation for premature saturation, hysteresis in the input-output characteristics, jump resonance and similar anomalous effects that so often degrade the steady state performance of saturable servo systems. Furthermore, systems which exhibit a large degree of anomalous steady state behavior can be expected to exhibit a correspondingly poor transient response under large signal conditions such as obtained when the system is coming out of saturation.

Demonstration of the existence of this dual mode of response and prediction of the range of signal amplitude and frequencies for which it exists are both built around the "saturated" transfer characteristics of the control loop. This analysis is quantitatively useful only in certain simple cases where the necessary saturated transfer characteristics can readily be found. Its chief value lies in the insight it provides to the cause of anomalous behavior. In this respect is it particularly useful to the designer because it relates this behavior to the loop gain and phase characteristics of the saturated transfer characteristics.

The analysis is applied here to two simple examples, and the results are verified by an analog computer study on simulations of these systems.

A Numerical Method for Determining a System Impulse Response from the Transient Response to Arbitrary Inputs—N. J. Zabusky

The transfer functions of large physical systems must be known if they are to be controlled efficiently. Environmental transient test procedures are described and their advantages presented. Because impulse and step functions cannot be obtained from practical test equipment, the problem of deducing the impulse response from arbitrary input and response data arises.

A successful iteration procedure for determining $h(\tau)$, the impulse response, by using the convolution equation, $r(t)=\int_0^\infty \theta(t-\tau)h(\tau)d\tau$, is described. An actual test transient is analyzed. An analog computer technique which mechanizes this iteration procedure is presented. The appendix contains a tape multiplication procedure for performing numerical convolutions.

Synthesis of Feedback Control Systems with a Minimum Lead for a Specified Performance—G. S. Axelby

Compensation is usually needed in a high performance feedback control system to realize a desired frequency response. Direct synthesis procedures to determine the necessary compensation are described. In particular, consideration is given to " minimum lead" systems which have an optimum gain.

Performance of Drive Members in Feedback Control Systems—F. M. Bailey

Limitations of system performance due to a mechanical structure are discussed, and practical methods of improving the performance of fire control systems with this limitation are illustrated.

Automation as the Engineer Sees It—W. R. G. Baker

Problems of automation are discussed with particular emphasis on their importance to society and to the engineer.

IRE Activities in the Field of Automatic Control—J. E. Ward

Professional Groups on Automatic Control—R. B. Wilcox

Information for Authors—G. S. Axelby

Because the first PGAC Transactions papers will be reproduced from photographs of original manuscripts, it will be necessary to maintain a standard form to provide uniformity, neatness and legibility.

This manuscript is written and typed in the desired form. Detailed instructions for authors are included.

Committees and Chapters of the IRE Professional Group on Automatic Control

Electronic Computers

Vol. EC-5, No. 2, June, 1956

PGEC Papers Awards for 1955

A One-Microsecond Adder Using One-Megacycle Circuitry—A. Weinberger and J. L. Smith

An analysis of the functional representation of the carry digits in the addition process shows that the one-megacycle circuitry of SEAC and DYSEAC can be organized logically to permit the formation of many successive carries simultaneously. The Boolean expression for any carry digit C_k can be expanded so as to be an explicit function of only the input digits of orders k to $k-p+1$ and of the carry digit C_{k-p}. Certain factorizations can then be made to simplify these expressions so that all of them fall within the limitations on the gating complexity imposed by the circuitry.

A parallel adder utilizing this principle is developed which is capable of adding two 53-bit numbers in one microsecond, with relatively few additional components over those required in a parallel adder of more conventional design.

A Small Coincident-Current Magnetic Memory—W. J. Bartik and T. H. Bonn

This paper describes a small coincident-current memory used for buffer storage. Such a memory as part of the self-checking card-to-magnetic-tape converter, an auxiliary of the Univac system, is now in production. Typical advantages of a small coincident-current memory in computer input-output equipment, as well as some of the problems encountered in its application, are described.

This memory affords the card-to-tape converter a great degree of flexibility, making it possible to read cards sidewise and to check and edit information with a minimum of hardware and complexity.

Memory cells consist of metallic-tape cores wound with multi-turn coils. The low currents required permit operation of the memory directly from the card-sensing brushes on writing and from a diode function-table on reading.

The functional aspects of the memory and its associated electrical circuitry are described. Information concerning the physical nature of the memory, specifications of the cores, and some of the tests performed in their inspection is also presented.

Reflected Number Systems—Ivan Flores

Many papers have been wirtten about the reflected binary system and it is well known in the computer field for analog-to-digital conversion. The method used in creating this system may be extended to systems of bases other than two. It is the purpose of this paper to carry this extension to its logical conclusion. The author describes how reflected systems of different bases may be composed. The equations for translating between the conventional and reflected systems are then derived. It is also demonstrated how the reflected binary system is a special case of reflected number systems and how the general case simplifies for the reflected binary case.

Analog Multipliers and Squarers Using a Multigrid Modulator—R. L. Sydnor, T. R. O'Meara, and J. Strathman

This article describes the use of a multigrid vacuum tube as an AM multigrid modulator multiplier. The accuracy of the multiplier is dependent only upon the linear properties of the vacuum tube used and not upon careful adjustment of the operating potentials. It is unusual that such a simple device should give a range of 78 db with only a ± 2 per cent full scale error. The advantages and also restrictions of this device along with a complete range of dynamic performance are included in this article.

Transistors in Current-Analog Computing—B. P. Kerfoot

Correspondence

Contributors

PGEC News

Reviews of Current Literature

Nuclear Science

Vol. NS-3, No. 3, June, 1956

The Electron Linear Accelerator as a Pulsed Radiation Source—M. G. Kelliher, J. C. Nygard, and A. J. Gale

One of the inherent properties of the electron linear accelerator is that it produces radiation in pulses whose duration is less than the pulse duration of the radio frequency source from which it is fed. In practice the pulse length lies in the range from one to five microseconds. However, by independently pulsing the electron source or injector, intense pulses of radiation of duration 0.1 μsec or less can be easily achieved and the lower limit is set only by stray capacitance effects. The radiation produced can be electrons, X-rays, or neutrons (by gamma-*n* reaction).

One of the more important practical applications in nuclear research is as a pulsed neutron source in time-of-flight neutron spectroscopy where the short duration and high flux, when used with appropriate analyzing circuitry, provide a spectrometer of high resolving power in the intermediate neutron range. Short bursts of radiation also find application in radio-chemistry where physico-chemical changes often occur in a very short interval after the absorption of radiation.

Recent Neutron Detector Studies at Argonne National Laboratory—G. F. Erickson, S. G. Kaufmann, and L. E. Pahis

A neutron sensitive photomultiplier tube has been built and tested in line with a program to investigate unconventional mechanisms of neutron detection. A fission counter following established principles has been designed with emphasis on minimum demands on skill and labor during construction and assembly, and minimum over-all cost.

Determination of Neutron Intensity and Gamma Spectrum of Neutron Sources—L. B. Gnagey

The Literature of Instrumentation for Radiological Studies—Marjorie Comstock

Instrumentation for radiological studies is interpreted as relating to device or tools for measurement purposes or for implementing research studies in which radioactive material is used. In radiological studies, instrumentation plays a vital part due to the hazards involved in handling radioactive material, and due to the necessity for accurate measurements.

The varied fields of research include use of the nuclear reactor, hot laboratory installations, accelerators, cloud chambers, and many other pieces of apparatus. All of these require accessory instruments in order to use them successfully. The list of instruments is lengthy and is mentioned merely to illustrate the spread of the subjects covering the literature of instrumentation for radiological studies. Therefore, it is necessary to have the subject as clearly defined as possible before attempting to search the literature. A good introduction to the literature of instrumentation is *Guide to Instrumentation Literature*. This contains an extensive bibliography. The various abstract journals, books, report literature, and other approaches to the literature are discussed.

Precision Pulse Generator—H. L. Miller

A pulse generator was developed for the range 10 to 100 volts to calibrate pulse height analyzers. Amplitude calibration depends on a standard cell. The generator can synchronize with power line frequency, an external oscillator, or furnish its own repetition rate. An advance pulse for triggering an oscilloscope is provided.

Factors Affecting the Application of Halogen-Quenched G-M Tubes—W. G. Egan

Pulses obtained from halogen-quenched G-M tubes operated under minimum circuit loading conditions were investigated. Oscillograms are shown and pulse characteristics discussed. The data show that the plateau obtained is determined, in part, both by the pulse shapes and by the electrical characteristics of the associated equipment. Spurious peaks in the halogen-quenched pulses occurring at normal operating voltages may be registered as counts, at certain voltage discrimination levels and circuit resolving times. This is of practical importance in radiation measurements.

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The Index to the Abstracts and References published in the PROC. IRE from February, 1955 through January, 1956 is published by the PROC. IRE, June, 1956, Part II. It is also published by *Wireless Engineer* and included in the March, 1956 issue of that journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

ACOUSTICS AND AUDIO FREQUENCIES

534.121 1928

Forced Vibrations of a Rigid Circular Plate on a Semi-Infinite Elastic Space and on an Elastic Stratum—G. N. Bycroft. (*Phil. Trans. A*, vol. 248, pp. 327–368; January 5, 1956.) An analytical paper.

534.2-14-8 1929

Measurement of Attenuation of [ultrasonic] Sound in the Isothermal Surface-Layer of Water—A. N. Barkhatov. (*Akust. Zh.*, vol. 1, pp. 315–320; October/December, 1955.) Results obtained experimentally using a tank of water with a vertical temperature gradient are in reasonable agreement with the theory given by Brekhovskikh and Ivanov (3132 of 1955).

534.21-8 1930

Acoustic and Other Physical Properties of Shallow-Water Sediments off San Diego—E. L. Hamilton, G. Shumway, H. W. Menard, and C. J. Shipek. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 1–15; January, 1956.) The investigation reported covers in situ measurements of sound velocity on various sediments at 100 kc and laboratory measurements of velocity and attenuation at 23–41 kc on samples. In some cases the velocity in the sediment was less than the velocity in the bottom water; this result is discussed in a separate paper [*ibid.*, pp. 16–19 (Hamilton)].

534.213 1931

Propagation of Sound Pulses in a Dispersive Medium—J. M. Proud, P. Tamarkin, and E. T. Kornhauser. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 80–85; January, 1956.) Experiments are described in which rectangular pulses comprising many sound-wave cycles were propagated in a long rectangular channel containing water. The velocity of the main signal is different from that of the initial and final transients. Comparison of the observed beats between the initial transient and the main signal with results predicted from analysis shows good agreement, hence the technique can be used to predict the effect on more complex pulses after the system has been calibrated with rectangular pulses.

534.22:551.510.53 1932

Introductory Theory for Upper Atmosphere Wind and Sonic Velocity Determination by Sound Propagation—Groves. (See 2046.)

534.231 1933

On the Dependence of Directivity Patterns on the Distance from the Emitter—J. Pachner. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 86–90; January, 1956.) A method is presented for the numerical computation of the directivity pattern at any distance from the source based on the expansion of a diverging wave into a series of spherical wave functions. In a separate paper (*ibid.*, pp. 90–92) the method of analysis is extended to cover the investigation of standing-wave as well as traveling-wave field components.

534.231:519.24 1934

Propagation of Correlation Functions in Continuous Media—R. H. Lyon. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 76–79; January, 1956.) The properties of noise fields produced by the random superposition of elementary sources are derived, using the method of analysis developed by Rice in connection with shot noise (2169 of 1945 and back references). A calculation is made of the correlation properties of the response of a continuous linear system exposed to a noise field, from a knowledge of the system impulse response and the correlation function of the source; the latter may be determined experimentally or by calculation.

534.232 1935

Sound Radiation from the Acoustic Boundary Layer—U. Ingard and D. Pridmore-Brown. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 128–129; January, 1956.) "The sound radiated from a finite rectangular plate in an infinite wall oscillating in shear motion is calculated. The order of magnitude of the intensity of this sound field is compared with that which is produced when the plane oscillates with the same amplitude in a direction normal to its surface."

534.232:546.431.824-31 1936

On the Resonant Vibrations of Thick Barium Titanate Disks—E. A. G. Shaw. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 38–50; January, 1956.) Optical interference technique has been used to study the surface motion of $BaTiO_3$ disks with radius/semithickness ratios ranging from 1.14 to 6.63. Vibration patterns, resonance frequency, and electro-mechanical coupling coefficients are given for 12 modes. There is no single mode that can be uniquely identified as the fundamental dilatational thickness resonance, but there is some evidence suggesting an optimum value of the ratio for transducer design.

534.232:546.431.824-31 1937

Radial Vibrations in Short, Hollow Cylinders of Barium Titanate—C. V. Stephenson. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 51–56; January, 1956.) A formula is developed, based on electrostriction equations, for determining the coupling coefficient of annular elements vibrating in the radial mode from the resonance and antiresonance frequencies. Many harmonies are found to be forbidden in this mode. Experimental evidence supports the theory presented.

534.24-14 1938

Underwater Sound Reflection from a Corrugated Surface—E. O. LaCasce, Jr., and P. Tamarkin. (*J. Appl. Phys.*, vol. 27, pp. 138–148; February, 1956.) Experimental results are compared with results predicted by three different published theories. Good agreement is obtained in respect of direction of reflection and cutoff frequencies. The closeness of the agreement in respect of intensity depends on the surface slope; the theories appear to be valid only for small slope.

534.52 1939

Problem of Interaction of Sound Waves—A. G. Gorelik and V. A. Zverev. (*Akust. Zh.*, vol. 1, pp. 339–342; October/December, 1955.) Report of an experimental investigation of the interaction of two sound waves traversing a liquid in mutually perpendicular directions.

534.6-8 1940

Experimental Determination of Ultrasonic Wave Pressure on Obstacles—B. Ozdogan. (*J. Phys. Radium*, vol. 16,pp. 902–907; December, 1955.) Experiments using waves of frequency 1.5 mc in water show that the coefficients of absorption for paraffin-wax and stearin surfaces increased with the angle of incidence. Curves show the variations of reflection and absorption coefficients with angle of incidence and with thickness of the absorbing layer.

534.61-8:535.314 1941

Investigation of Stationary Ultrasonic Waves by Light Refraction—A. P. Loeber and E. A. Hiedemann. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 27–35; January, 1956.) Continuation of work reported previously [2288 of 1954 (Kolb and Loeber)].

534.64 1942

Method of measuring Acoustic Impedance based on Measurement of the Geometrical

Difference of Sound Pressures—V. N. Fedorovich. (*Akust. Zh.*, vol. 1, pp. 360–367; October /December, 1955.)

534.75 1943
Sensitivity to Changes in the Interruption Rate of White Noise—G. H. Mowbray, J. W. Gebhard, and C. L. Byham. (*J. Acoust. Soc. Amer.* vol. 28, pp. 106–110; January, 1956.)

534.78 1944
Automatic Extraction of Formant Frequencies from Continuous Speech—J. L. Flanagan. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 110–118; January, 1956.) Two electronic devices are described for deriving direct voltages corresponding to the first three formant frequencies. The performance of the devices is evaluated in a separate paper (*ibid.*, pp. 118–125.)

534.845 1945
Absorption Characteristics of Upholstered Theater Chairs and Carpet as measured in Two Auditoriums—R. N. Lane. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 101–105; January, 1956.) Measurements on a 496-seat and on a 738-seat auditorium are reported. The value found for the absorption per seat over the af band is much lower than reported in various previous publications [*e.g.*, 2198 of 1953 (Parkin *et al.*)]

534.85/.86:534.76 1946
Two-Channel Stereophonic Sound Systems—F. H. Brittain and D. M. Leakey. (*Wireless World*, vol. 62, pp. 206–210; May, 1956.) The basic requirements for sound location are discussed and the effects of the directional characteristics of the microphones and loudspeakers and of the position of the listener on the sound image are described. Experimental results show that differences of sound intensity from the two loudspeakers afford better guidance in positioning the sound image than do the time differences.

534.86:621.396.66 1947
Monitoring Sound Broadcast Programmes—Somerville (See 2221.)

621.395.623.7 1948
Loudspeakers with Spherical Radiation—H. Schiesser. (*Rev. Son.*, pp. 4–10; January, 1956.) A discussion is presented of the departure from linearity of the frequency response of a cone loudspeaker as a function of the polar coordinates of the auditor's position. Improvements are possible by use of a spherical loudspeaker system. Practical approximations comprising a number of loudspeakers mounted in the faces of a polyhedron are described, with an indication of suitable feed amplifier circuits. For practical purposes a hemispherical arrangement may be satisfactory. Systems for domestic receivers giving stereophonic reproduction are discussed.

534.839 1949
Notes on Applied Science No. 10. Noise Measurement Techniques [Book Review]—Publishers: H. M. Stationery Office, London, 1955, 40 pp. (*Brit. J. Appl. Phys.*, vol. 7, p. 41; January, 1956.) "... a guide to the choice of methods and techniques for determining the physical characteristics of noise, based on the experience of the National Physical Laboratory."

ANTENNAS AND TRANSMISSION LINES

621.315.212:621.3.013.78:621.317.3 1950
Measurement of Coupling Impedance and its Application to the Study of Cable Screens—J. Bourseau and H. Sandjivy. (*Câbles and Transm.*, vol. 10, pp. 11–30; January, 1956.) Various definitions of coupling (transfer) impedance are discussed, together with known methods for measuring it. A new direct method of measurement is described, using a bridge. An account is given of an application of the method to an investigation of screens for reducing crosstalk between coaxial pairs. Helical as well as cylindrical screens are discussed.

621.315.212.011.21 1951
Influence of Standard Splicing on the Uniformity of Impedance of a 2.6/9.4 Coaxial-Pair Cable—R. Roch and J. Bouzitat. (*Câbles and Transm.*, vol. 10, p. 3–10; January, 1956.) Formulas are derived for the components and the modulus of the impedance deviation at the splices, which are assumed to comply with French Post Office standards. The frequency range considered is 2.45–4.1 mc. A numerical example is included.

621.372.029.6+621.385.029.6 1952
Report of Advances in Microwave Theory and Techniques—1954—D. D. King. (Trans. IRE, vol. MTT-3, pp. 4–7; April, 1955.) A review of guided-wave transmission and the circuit aspects of microwave generators and amplifiers, comprising a classified bibliography of 167 items.

621.372.43 1953
The Optimum Tapered Transmission-Line Matching Section—R. E. Collin. (Proc. IRE, vol. 44, pp. 539–548; April, 1956.) The matching section is analyzed as for a high-pass filter. The results obtained previously for the n-section $\lambda/4$ transformer (1250 of 1955) are adapted by allowing n to become infinite; theory developed in relation to antenna design [984 of 1956 (Taylor)] is used. The optimum taper determined yields a matching section 13.9 per cent shorter than the exponential taper and 27 per cent shorter than the Gaussian taper for the same cutoff frequency and pass band tolerance.

621.372.8 1954
A New Annular Waveguide Rotary Joint—K. Tomiyasu. (Proc. IRE, vol. 44, pp. 548–553; April, 1956.) A description is given of a joint designed to permit multiple stacking on a common axis. The joint will carry high power and permit operation with low swr and low-insertion loss throughout the full rotation. Theory and performance characteristics are included.

621.372.8:512.3 1955
Paired Systems of Infinite Linear Algebraic Equations, linked with Infinite Periodic Structures—Ya. N. Fel'd. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, pp. 215–218; January 11, 1956. In Russian.) The method developed is applied to determine the currents in a rectangular waveguide containing an infinite array of transverse rods.

621.372.8:621.318.134 1956
Propagation in a Ferrite-Filled Waveguide—L. G. Chambers. (*Quart. J. Mech. Appl. Math.*, vol. 8, Part 4, pp. 435–447; December, 1955.) "Perturbation methods are used for the solution of the fields in a waveguide filled with ferrite material which is subjected to a static magnetic field in the direction of its axis. It is shown that quasi TE and quasi TM modes exist and the first terms in the expansion are calculated for the case of a waveguide of rectangular cross section."

621.372.8:621.318.134 1957
Some Applications and Characteristics of Ferrite at Wavelengths of 0.87 cm and 1.9 cm—C. Stewart. (Trans. IRE, vol. MTT-3, pp. 27–31; April, 1955.) The use of ferrites to produce Faraday rotation in waveguides is discussed and experimental results are presented, with details of the construction of a unidirectional waveguide for λ 0.87 cm. The improvement of the Dicke-type radiometer by use of devices based on this principle is described.

621.396.674.3 1958
Radiation Resistance of Dipoles in an Interface Between two Dielectrics—J. R. Wait. (*Canad. J. Phys.*, vol. 34, pp. 24–26; January, 1956.) Exact expressions are derived for electric and magnetic dipoles located in a plane interface.

621.396.674.3:621.396.11 1959
Transient Fields of a Vertical Dipole over a Homogeneous Curved Ground—J. R. Wait. (*Canad. J. Phys.*, vol. 34, pp. 27–35; January, 1956.) Analysis is given first for the flat-earth case; the earth's curvature is then taken into account by means of appropriate modifications. When the antenna current is a linear function of time, the radiation field on a flat perfectly conducting earth is of step-function form; departures from this form are caused by the finite conductivity and dielectric constant of the ground, the induction and static fields of the antenna, and the earth's curvature, the last-named factor becoming effective at distances >50 km.

621.396.677.3 1960
The Optimum Current and Field Distribution for Broadside and End-Fire Radiators with Continuous Illumination—A. Heilmann. (*Nachrichtentech. Z.*, vol. 9, pp. 1–9; January, 1956.) A method is presented for calculating the distribution of illumination to obtain maximum side-lobe attenuation for given beam width. The distribution function is built up from a limited number of terms of a Fourier series. The required number of terms and the beam width increase as the requirements for side-lobe attenuation become greater. With three terms, an attenuation >70 db can be attained with a broadside array and >80 db with an endfire array.

621.396.677.3.029.62 1961
LONG, Long Yagis—J. A. Kmosko and H. G. Johnson. (*QST*, vol. 40, pp. 19–24; January, 1956.) An experimental investigation of the characteristics of Yagi arrays for 2 m λ is reported; results for arrays comprising up to 68 elements are presented graphically. Constructional details are also given.

621.396.677.7 1962
The Phase Centre of Aperture Radiators—K. Baur. (*Arch. Elekt. Übertragung*, vol. 9, pp. 541–546; December, 1955.) Calculations are based on Kirchhoff's aperture field method. Simplifications are introduced enabling the problem to be dealt with by means of tabulated functions. An approximate formula useful in practice is given. Results are compared with values obtained by measurements on horn radiators.

621.396.677.7 1963
Two New Modifications for Microwave Aerials—G. von Trentini. (*Rev. Telegr. Electronica, Buenos Aires*, vol. 44, pp. 715–718; December, 1955.) Constructional details and performance figures are presented for simple designs of a) pyramidal horns with good directivity and gain, using trolitul masks for the counterphase Fresnel zones in the aperture, and b) wide-band antennas for circular polarization, using dielectric rods in circular waveguides.

621.396.677.8 1964
Passive Reflectors for Radio Beams (Experimental Investigation)—G. Andrieux. (*Onde Élect.*, vol. 36, pp. 57–72; January, 1956.) Experiments based on theoretical work by Jakes (1243 of 1953) were carried out on a wavelength of 1.25 cm. Auxiliary tests proved that the results were applicable to an installation for 8.75 cm λ. The performance of a given radio link is not appreciably affected by the use of a reflector system; risk of increased coupling between neighboring antennas does exist, but can be overcome.

621. 396.677.859 1965
Design of Randomes—L. Thourel and S. Herscovici. (*Ann. Radiolect.*, vol. 10, pp. 163–

173; April, 1955.) Formulas developed previously [651 of 1950 (Cady *et al.*)] are extended to deal with multiple sandwich constructions. Simple methods of calculation are presented, together with experimental results.

AUTOMATIC COMPUTERS

681.142 1966
A Mechanical Binary-Decimal Converter—M. Setterwall. (*J. Sci. Instrum.*, vol. 33 pp. 18–19; January, 1956.)

681.142 1967
An Electronic Generator for Functions of Two Independent Variables—V. Wentzel. (*Ericsson Tech.*, vol. 11, pp. 183–225; 1955.) A unit built at the Chalmers University of Technology is described. The function is recorded on a photographic plate in the form of variable-width columns; these are scanned by a cr tube. The output is either in the form of width-modulated pulses or in the form of a voltage proportional to the function.

681.142:519.272:534.6 1968
Measurement of Correlation Coefficient—S. G. Gershman and E. L. Feinberg. (*Akust. Zh.*, vol. 1, pp. 326–338; October/December, 1955.) The determination of the correlation coefficient of af noise is based on the measurement of the coincidences of sign of rectangular pulses triggered by the incoming signals. The instrument is described in detail and the theory of operation is given. See also 2542 of 1955 (Goff).

681.142:621.3:620.16 1969
Shock Spectrum Computer for Frequencies up to 2000 c/s—C. T. Morrow and D. E. Riesen. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 93–101; January, 1956.) An arrangement for investigating the effects of mechanical shock on electronic equipment installed, *e.g.*, in guided missiles comprises an analog computer which operates on an accelerometer signal, either direct or recorded, and solves the differential equation for the part involved, the result being displayed on a cro screen. The computer can be tuned up to 2 kc thus covering the range within which the fundamental resonance of any delicate structure is likely to occur.

681.142:621.314.7 1970
Transistor Digital Computers—(*Wireless World*, vol. 62, pp. 210–212; May, 1956.) A brief account is given of circuit techniques involving the use of transistors and ferrite two-state storage devices described at the *IEE* convention on digital computers held in London in April, 1956.

681.142:621.318.5:537.312.62 1971
The Cryotron—a Superconductive Computer Component—Buck. (See 1980.)

681.142:621.387 1972
The Gas-Filled Diode as a Digital Storage Element—B. R. Taylor and R. Bird. (*Electronic Engng.*, vol. 28, pp. 151–155; April, 1956.)

681.142 1973
Proceedings of the Eastern Joint Computer Conference [Book Review]—Publishers: The American Institute of Electrical Engineers, New York, 92 pp. (*Brit. J. Appl. Phys.*, vol. 7, p. 41; January, 1956.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.2 1974
Parallel-Connected Nonlinear Impedances—S. Mayr. (*Elektrotech. u. Maschinenb.*, vol. 73, pp. 31–38; January 15, 1956.) Analysis based on a matrix method of representing complex vectors (847 of 1956) and locus diagrams are used to determine the equivalent single impedance for an arbitrary system of parallel impedances; values of voltage resulting from given current loads and of current resulting from given applied voltages are hence derived. The methods are illustrated by a numerical example taking two nonlinear impedances with ohmic resistance components. The complete working diagram of the parallel system is obtained by combining the locus diagram with the I/V characteristic.

621.314.2+621.318.43 1975
Subminiature Transformers and Transductors—E. F. Dunkin and D. L. Johnston. (*Electronic Engng.*, vol. 28, pp. 144–150; April, 1956.) Limitations associated with subminiature design are discussed in relation to audio and control-frequency transformers and transductors. Below a certain size, a toroidal-shell construction has given results comparing favorably with laminated assemblies; the signal-power level of the transformers is sufficient for junction-transistor circuits. When this construction is applied to transductors, the excitation and control fields are orthogonally related. A paper covering much of the same ground appears in TRANS. IRE PGCP-3, pp. 30–44; April, 1955.)

621.314.213 1976
Transformer "Miniaturization" using Fluorochemical Liquids and Conduction Techniques—L. F. Kilham, Jr., and R. R. Ursch. (PROC. IRE, vol. 44, pp. 515–520; April, 1956.)

621.316.825 1977
The Stability of Thermistors—A. Beck. (*J. Sci. Instrum.*, vol. 33, pp. 16–18; January, 1956.) Experimental results indicate that though the constants of two thermistors tested undergo changes over a period of months, the changes are slow enough for temperature measurements to be made to an accuracy within 0.02°C. over a range of 10°C. in experiments lasting up to 24 hours.

621.318.4.002.2 1978
Design of Modern Winding Machines—(*TSF et TV*, vol. 32, pp. 10, 15; January, 1956.) A short review in which U. S. and European methods are contrasted.

621.318.5 1979
Transfer Function of Relays with Inactive Zone and Hysteresis—R. Setton. (*C.R. Acad. Sci., Paris*, vol. 242, pp. 1138–1140; February 27, 1956.) The equation of transfer for the relay is calculated by the Laplace-transform method.

621.318.5:312.62:681.142 1980
The Cryotron—a Superconductive Computer Component—D. A. Buck. (PROC. IRE, vol. 44, pp. 482–493; April, 1956.) "The study of nonlinearities in nature suitable for computer use has led to the cryotron, a device based on the destruction of superconductivity by a magnetic field. The cryotron, in its simplest form, consists of a straight piece of wire about one inch long with a single layer control winding wound over it. Current in the control winding creates a magnetic field which causes the central wire to change from its superconducting state to its normal state. The device has current gain, that is, a small current can control a larger current; it has power gain so that cryotrons can be interconnected in logical networks as active elements. The device is also small, light, easily fabricated, and dissipates very little power."

621.318.57:621.374.3:621.387 1981
Batching and Counting using Gas-Filled Decade Tubes—W. Grimmond and W. H. P. Leslie. (*Electronic Engng.*, vol. 28, pp. 138–143; April, 1956.) A range of units is described from which a variety of frequency meters, batching counters, etc. can be quickly assembled; circuit diagrams are given.

621.319.4 1982
The Impedance of Shorted-Edge Wound Capacitors—H. Heywang. (*Arch. Elekt. Übertragung*, vol. 10, pp. 29–44; January, 1956.) Results on the frequency variation of the impedance of this type of capacitor, obtained by Leiterer (449 of 1944), are discussed, using simple mathematics. The analysis is given first neglecting losses and is then extended to take account of losses in the dielectric and in the thin metal films. Eddy currents in the projecting contact layer and in thicker metal foils are also taken into account. The method is used to determine the transfer impedance of lead-through capacitors.

621.319.45 1983
The Forming of the Negative Electrode of Electrolytic Capacitors—T. Bohlin and Å. Lagercrantz. (*Ericsson Tech.*, vol. 11, pp. 263–278; 1955.) An examination is made of the conditions under which undesired forming occurs; formulas are presented for the decrease of the capacitance on repeated discharging; these enable load behavior to be predicted and optimum design to be attained. The theory is supported by experimental results.

621.372:621.314.7 1984
Principles of Transistor Circuits—J. P. Vasseur. (*Ann. Radioélect.*, vol. 10, pp. 99–162; April, 1955.) A comprehensive review covering amplifiers, oscillators and flip-flops. Over 70 references.

621.372.012 1985
Tolerance Limits in Matching—W. Alexander. (*Electronic Engng.* vol. 28, pp. 162–164; April, 1956.) Curves are given showing the level of power transfer from source to load for various degrees of mismatch.

621.372.029.6+621.385.029.6 1986
Report of Advances in Microwave Theory and Techniques—1954—King. (See 1952.)

621.372.5 1987
Impedance Synthesis without Minimization—A. Fialkow and I. Gerst. (*J. Math. Phys.*, vol. 34, pp. 160–168; October, 1955.) The synthesis procedure described for realizing $Z(p)$, a rational positive real function of the complex-frequency variable p, by means of a network containing no mutual inductances, is performed in three steps in which the computionally difficult minimization process is avoided.

621.372.5 1988
The Geometrical Representation of Combined Linear Quadripoles—J. de Buhr. (*Arch. Elekt. Übertragung*, vol. 9, pp. 561–570; December, 1955.) Concepts of non-Euclidean geometry are used. The representation of reactance quadripoles by a rigid system of two straight transformation lines and the determination of impedance transformations from geometrical reflections afford useful methods for treating composite quadripoles.

621.372.5 1989
The Geometrical Quadripole Representation of the Double Transformer—J. de Buhr. (*Arch. Elekt. Übertragung*, vol. 10, pp. 45–49; January, 1956.) The representation is effected by adapting the technique described previously (1643 of 1956).

621.372.5.011.1 1990
The Operating-Parameter Cascade Matrix of Quadripoles—F. L. Bauer. (*Arch. Elekt. Übertragung*, vol. 9, pp. 559–560; December, 1955.)

621.372.54:621.372.8:537.226 1991
Waveguide Filters—M. H. N. Potok. (*Wireless Engr.*, vol. 33, pp. 79–82; April, 1956.) Filters having high Q and low insertion loss are produced by arranging inside a wave-

guide a number of suitably spaced dielectric sections constituting $\lambda/4$ transformers.

621.372.543.2 1992

Band-Pass Characteristics of Low Asymmetry—B. Easter. (*Electronic Engng.*, vol. 28, pp. 156–158; April, 1956.) An empirical design procedure is formulated, leading to filter networks which compare favorably with conventional designs.

621.372.543.3:621.397.62:535.623 1993

New Rejector Circuit—W. T. Cocking. (*Wireless Engr.*, vol. 33, pp. 77–79; April, 1956.) A circuit used in some color television receivers [1882 of 1956 (Fairhurst)] is described. It uses a conventional parallel-LC trap connected into circuit via a center-tapped bifilar-wound coil with a resistance across one half; it is characterized by negative inductance and capacitance.

621.372.56.029.6:621.372.8:621.318.134 1994

A Double-Slab Ferrite Field Displacement Isolator at 11 kMc/s—S. Weisbaum and H. Boyet. (Proc. IRE, vol. 44, pp. 554–555; April, 1956.) Brief description, with performance figures, of an isolator comprising a waveguide with two slabs of ferrite arranged with transverse symmetry and subjected to equal but oppositely directed magnetic fields; the forward loss is <1 db over most of the range from 10.7 to 11.7 kmc, and varies by <0.1 db over any 20 mc channel, while the reverse loss is 64–70 db. See also 972 of 1955 (Lax *et al.*).

621.372.6 1995

Impedance Transformation of Linear 2n-Terminal Networks—H. Kleinwächter. (*Arch. Elekt. Übertragung*, vol. 10, pp. 26–28; January, 1956.) If $(n-2)$ outlets of a $2n$-terminal network are terminated by variable reactances, the resulting quadripole represents a variable transformer. As the value of these reactances varies between $-j\infty$ and $+j\infty$, the input impedance assumes all values of the transforming range. This range does not include the entire right-hand half of the complex impedance plane, but only circular areas thereof. An equation is derived for determining these circles. The method is demonstrated in relation to a waveguide hybrid-T.

621.373 1996

Fluctuations in Self-Oscillating Systems of Thomson Type—S. M. Rytov. (*Zh. Eksp. Teor. Fiz.*, vol. 29, pp. 304–328; September, 1955.) Amplitude and phase fluctuations in weakly nonlinear self-oscillating systems are considered, using a symbolic differential equation describing the fluctuations of the random functions and the methods of the correlation theory. A system with one degree of freedom is considered first and the general theory is applied to the case of a tube generator operating in the "soft" state. The theory is then applied to passive systems with one and two degrees of freedom. The effect of stabilizing the frequency of an oscillator by means of a high-Q circuit is discussed. The method of taking into account thermal fluctuations is shown.

621.373.4.029.6:621.396.822 1997

Fluctuation of Oscillations of Klystron Generator—I. L. Bershtein. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, pp. 453–456; January 21, 1956. In Russian.) A theoretical investigation of noise in a reflex klystron is presented. In a typical case the natural bandwidth of the oscillations is 0.1 cps; the results obtained by Shimoda (2943 of 1953) are believed to be in error.

621.373.42 1998

Frequency of the Three-Phase R-C Coupled Oscillator: Part 1—Non-reactive Anode Load Resistance—H. Rakshit and M. C. Mallik. (*Indian J. Phys.*, vol. 29, pp. 534–547; November, 1955.) Report of an investigation of the effect of different types of cathode impedance on the oscillation frequency.

621.373.43:621.314.7 1999

Application of Junction Transistors to the Generation of Linear Sawtooth Waveforms—(*Mullard Tech. Commun.*, vol. 2, pp. 134–139; December, 1955.)

621.373.44 2000

Two Trigger Circuits Useful as Sources of Rectangular Pulses—G. G. E. Low (*Electronic Engng.*, vol. 28, pp. 158–159; April, 1956.) A modified Eccles-Jordan circuit having low output impedance and a development of Schmitt's cathode-coupled trigger circuit are presented.

621.375:[621.314.7+621.385 2001

Comparison of Junction Transistor and Amplifier Valve—G. Ledig. (*Arch. Elekt. Übertragung*, vol. 10, pp. 1–9; January, 1956.) The quadripole equations are developed for the commonly used circuits; important parameters and relations are tabulated. For linear operation the tube can be regarded from the network point of view as a simplified limiting case of the class containing both tubes and transistors

621.375.221.2 2002

Analysis of a Regenerative Amplifier with Distributed Amplification—B. S. Golosman. (Proc. IRE, vol. 44, pp. 533–534; April, 1956.) Analysis indicates that the application of distributed amplification in regenerative circuits, such as the monostable multivibrator, is limited by its inherent time delay.

621.375.23 2003

Component Tolerance Effects in Feedback I. F. Amplifiers—H. S. Jewitt. (*Electronic Engrg.*, vol. 28, pp. 165–167; April, 1956.) Analysis relating to T and Π feedback networks is given. Resistor variation produces no asymmetry of the response curve; in this respect the feedback IF amplifier (1009 of 1954) is superior to the stagger-tuned amplifier.

621.375.232.3.024 2004

A Direct-Current Amplifier Stage with Asymmetrically Earthed Input—E. G. Schlosser and S. Götze. (*Frequenz.*, vol. 10, pp. 19–24; January, 1956.) A push-pull cathode-follower circuit is used, with two tubes having dissimilar characteristics and a special resistance coupling, the usual phase-reversing stage being omitted. Analysis is presented on the basis of a parabolic approximation to the characteristics. Design procedures for insuring the requisite quiescent-current and linearity conditions are indicated. A numerical determination is made of the operating point for a circuit using one Type-EL90 and one Type EL-32 tube.

621.376.22:621.318.134 2005

A Note on Sidebands produced by Ferrite Modulators—P. A. Rizzi and D. J. Rich. (Proc. IRE, vol. 44, p. 556; April, 1956.)

621.376.23:621.385.029.6 2006

Microwave Detector—Mendel. (See 2259.)

621.376.332.029.6:621.372.413 2007

A Simple Microwave Discriminator—C. Colani. (*Frequenz.*, vol. 10, pp. 25–26; January, 1956.) A cylindrical resonator with slightly disturbed symmetry has two closely spaced resonance frequencies for the H_{11} mode, corresponding to two waves with slightly different propagation velocities. These can be detected separately, by means of rectifiers. A discriminator characteristic is given by the difference of the rectified currents as a function of frequency. The dimensions of the resonator for a given frequency can be reduced by capacitive loading. Discriminators of this type are suitable for measurement and control purposes rather than for demodulating fm.

621.39.03 2008

Miniaturization and Quality Improvement of Circuit Parts—T. Nijo. (*Rep. Elect. Commun. Lab., Japan*, vol. 3, pp. 22–28; August, 1955.) Coils, transformers, resistors, capacitors, quartz crystal units, and filters developed by the Nippon Telegraph and Telephone Public Corporation are described and illustrated.

621.397.6.001.4 2009

Circuit Technique for Generation of Electrical Test Patterns in Television—Pilz. (See 2231.)

GENERAL PHYSICS

53.05 2010

A Method of analysing Periodicity—F. Mosetti. (*Ann. Geofis.*, vol. 8, pp. 331–349; July, 1955.) A method of analyzing an empirical function such as that corresponding to the record of an oscillation process is based on asymmetrical rather than symmetrical linear combinations of ordinates; it permits detection of variations of phase as well as amplitude.

530.145:535.14 2011

Theory of Radiation—J. C. Gunn. (*Rep. Progr., Phys.*, vol. 18, pp. 127–183; 1955.) A survey of the application of quantum field theory to the interaction between charged particles and the electromagnetic field; within this domain it is now possible to calculate any process with a precision only limited by the labor of the calculations. About 60 references.

534.01 2012

Intensity of Harmonic and Combination Components in the Nonlinear Distortions of Complex Oscillations—V. M. Vol'f. (*Akust. Zh.*, vol 1, pp. 321–325; October–December, 1955.) The effect of linear, quadratic, and cubic response characteristics on triangular, sawtooth, and rectangular signals is considered; results are tabulated.

535.34:537.56:546.17 2013

Ultraviolet Absorption of Atomic Nitrogen in its Ionization Continuum—A. W. Ehler and G. L. Weissler. (*J. Opt. Soc. Amer.*, vol. 45, pp. 1035–1043; December, 1955.) The absorption of radiation of 400–800 Å λ by the plasma of a discharge in an ionization gauge was measured by passing the transmitted radiation into a spectrograph. Various considerations indicate that the absorption is due to atomic nitrogen, the absorption cross section of which is hence deduced.

537.1 2014

The Physical Interpretation of the Self-Acceleration of Electrons—K. Wildermuth (*Z. Naturf.*, vol. 10a, pp. 450–459; June, 1955.) The forces governing the motion of electrons are more easily understood if the point electron is considered as the limiting case of the finite-size electron. For self-acceleration to occur, it is not essential that the em field energy be infinite but it is important that the mechanical mass of the electron be negative.

537.122 2015

Measurement of the Specific Charge of Conduction Electrons—V. M. Yuzhakov. (*Zh. Eksp. Teor. Fiz.*, vol. 29, pp. 388–390; September, 1955.) The theory of a method of determining epm by means of a special dynamo is given. The device comprises a pair of concentric coils which together rotate with angular velocity o, about an axis parallel to the applied magnetic field. The rectangular inner coil can also rotate about its own axis, which is perpendicular to the magnetic field. The two resulting emfs in the inner coil are due to a) induction and b) Coriolis force acting on the conduction electrons. If these emfs are equal then epm$=2co/H$, where c is the velocity of light and H the magnetic-field strength. The estimated experimental errors are not greater than 1 per cent.

537.2 2016

Potential due to a Uniformly Charged Disk—É. Durand. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 887–889; February 13, 1956.) Formulas are derived applicable to any point.

537.311 2017

The Effect of Free Electrons on Lattice Conduction—J. M. Ziman. (*Phil. Mag.*, vol. 1, pp. 191–198; February, 1956.) "The scattering of phonons by electrons is calculated, assuming the usual electron-phonon interaction, for a parabolic band whose degeneracy temperature is comparable with the temperature of the lattice. The contribution to the thermal resistance is given by an exact formula, subject only to justifiable assumptions concerning phonon-phonon interactions. With rising temperature the apparent mean free path of the phonons at first decreases as $1/T$ (or, if there are very few electrons, as $\exp(a/T)$), but reaches a minimum and then increases as T^2. Energy and momentum conservation then allow only the tail of the electron distribution to contribute to the scattering. The model is thought to apply to certain observations on p-type germanium, irradiated sapphire and conducting diamond."

537.5 2018

Ionization by Relativistic Particles—B. T. Price. (*Rep. Progr. Phys.*, vol. 18, pp. 52–82; 1955.) A summary is given of theories of the relativistic increase of energy loss by ionization and of the density effect. The experimental evidence, as obtained with various types of counter, is discussed and compared with theoretical predictions. Over 100 references.

537.5 2019

The Ionization and Dissociation of Complex Molecules by Electron Impact—J. D. Craggs and C. A. McDowell. (*Rep. Progr. Phys.*, vol. 18, pp. 374–422; 1955.) A review covering theoretical and experimental aspects of collision processes in polyatomic molecular gases. Over 100 references.

537.525:537.534 2020

Ion Oscillations in a Cathode Potential Minimum—K. G. Emeléus and N. R. Daly. (*Proc. Phys. Soc.*, vol. 69, pp. 114–115; January 1, 1956.) A brief theoretical note.

537.525:538.569.029.6:538.6 2021

The Rotation of Plasmoids in a Magnetic Field—H. Pupke and H. G. Thom. (*Naturwissenschaften*, vol. 43, p. 32; January, 1956.) The motion of a luminous low-pressure gas discharge in Ne in the presence of a magnetic field and a superposed rf field of frequency about 75 mc was found to be a function of the anode voltage of the output tube of the 1-kw rf generator used.

537.528 2022

Formative Time Lags in the Electric Breakdown of Liquid Hydrocarbons—R. W. Crowe. (*J. Appl. Phys.*, vol. 27, pp. 156–160; February, 1956.)

537.533 2023

The Work Function and Patch Field of an Irregular Metal Surface—M. J. Morant and H. House. (*Proc. Phys. Soc.*, vol. 69, pp. 14–20; January 1, 1956.) Calculations show that the lowering of the work function due to irregularities of the surface is entirely compensated by a patch field.

537.56:538.569.029.4/.6 2024

The Breakdown of Gases subject to Crossed Electric Fields—W. A. Prowse and P. E. Lane. (*Proc. Phys. Soc.*, vol. 69, pp. 33–46; January 1, 1956.) The effect of an auxiliary electric field, acting at right angles to a 10-kmc field, on the breakdown stress of various gases is investigated experimentally. When the auxiliary field is unidirectional or of relatively low frequency (0.86 mc) its application raises the breakdown stress, but when its frequency reaches 9.7 mc the two fields appear to act independently. A partial explanation is advanced. The gases studied include air, oxygen, nitrogen hydrogen, and neon.

538.114 2025

Spin-Deviation Theory of Ferromagnetism: Part 2—The Non-ideal Spin Deviation Gas—J. Van Kranendonk. (*Physica*, vol. 21, pp. 925–945; December, 1955.) Part 1: 1368 of 1956.

538.114 2026

Application of the Bethe-Weiss Method to Ferrimagnetism—J. S. Smart. (*Phys. Rev.*, vol. 101, pp. 585–591; January 15, 1956.)

538.3 2027

Classical Electrodynamics as a Distribution Theory—J. G. Taylor. (*Proc. Camb. Phil. Soc.*, vol. 52, Part 1, pp. 119–134; January, 1956.)

538.3:535.13 2028

A Derivation of Generalized Macroscopic Electrodynamic Equations: Part 1—Non-relativistic—B. Podolsky and H. Derman. (*J. Math. Phys.*, vol. 34, pp. 198–207; October, 1955.) Maxwell's equations and the constitutive relations are derived from the classical microscopic theory, without making use of extensive assumptions, so that the effects of the higher-order electric and magnetic moments are retained in the equations.

538.56:537.56 2029

On the Theory of Stationary Waves in Plasmas—N. G. Van Kampen. (*Physica*, vol. 21, pp. 949–963; December, 1955.) A mathematical treatment is presented appropriate to the case of a continuous nonvanishing distribution of particle velocities. A complete set of stationary-plane-wave solutions can be constructed. There is no dispersion equation because for a given wave vector a continuous range of values of frequency is possible.

538.566:537.56 2030

Growing Electromagnetic Waves—J. H. Piddington. (*Phys. Rev.*, vol. 101, pp. 9–14; January 1, 1956.) The growth of em waves is considered in terms of Bailey's electromagneto-ionic theory (see, *e.g.*, 105 of 1952). Of the 12 different wave modes predicted by this theory, four are unreal; the remainder comprise two pairs of hydromagnetic waves which become pure em waves at a sufficiently high frequency, one pair of modified sound waves, and one pair of modified electron-sound waves. The growth of the em waves may result from the trapping of ions between potential troughs of the space-charge wave and the subsequent surrender of energy by the ions. Ion drifts may introduce important effects not indicated by the wave equations.

538.566:538.221:538.6 2031

Theory of Wave Propagation in a Gyromagnetic Medium—P. S. Epstein. (*Rev. Mod. Phys.*, vol. 28, pp. 3–17; January, 1956.) The theory presented is relevant to propagation in ferrites, but the physical nature of ferrites is not discussed.

538.569.4 2032

On the Absorption of 3.18-cm Microwaves in some Substituted Phenols in the Liquid State—D. K. Ghosh. (*Indian J. Phys.*, vol. 29, pp. 581–586; December, 1955.)

538.569.4:538.221 2033

Possible Source of Line Width in Ferromagnetic Resonance—A. M. Clogston, H. Suhl, L. R. Walker, and P. W. Anderson. (*Phys. Rev.*, vol. 101, pp. 903–905; January 15, 1956.) Brief theoretical discussion indicating the effect of the finite size of a sample on the dispersion of relaxation times.

539.1:[537.311.31+537.311.33 2034

The Displacement of Atoms in Solids by Radiation—G. H. Kinchin and R. S. Pease. (*Rep. Progr. Phys.*, vol. 18, pp. 1–51; 1955.) A survey covering theoretical and experimental aspects of the irradiation of solids by particles or γ rays; effects produced in metals and semiconductors are described. Nearly 200 references.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 2035

Halo of Radio Emission and the Origin of Cosmic Rays—G. R. Burbidge. (*Phys. Rev.*, vol. 101, pp. 906–907; January 15, 1956.) Observations of rf radiation from nebulas [*e.g.*, 104 of 1955 (Baldwin)] indicate that a large fraction of the radiation comes from roughly spherical regions centered on the galactic center and having radii of about 15 kiloparsecs. The particle density and energy conditions in these regions are considered in relation to the acceleration of cosmic rays.

523.16:523.45 2036

A Search for Radiation from Jupiter at 38 Mc/s and at 81.5 Mc/s—F. G. Smith. (*Observatory*, vol. 75, pp. 252–254; December, 1955.) Following the observation of radiation from Jupiter at 22 mc [2933 of 1955 (Burke and Franklin)], records obtained at Cambridge, England, of radiation received at 38 mc and 81.5 mc were searched for evidence of radiation from Jupiter at these frequencies; results were negative. Inferences are drawn regarding the source of radiation on Jupiter.

523.5:621.396.96 2037

Characteristics of Radio Echoes from Meteor Trails: Part 2—The Distribution of Meteor Magnitudes and Masses—I. C. Browne, K. Bullough, S. Evans, and T. R. Kaiser. (*Proc. Phys. Soc.*, vol. 69, pp. 83–97; January 1, 1956.) The distributions are deduced from observations of sporadic meteors and various showers. Part 1: 2782 of 1948 (Lovell and Clegg).

523.5:621.396.96 2038

Characteristics of Radio Echoes from Meteor Trails: Part 4—Polarization Effects—E. R. Billam and I. C. Browne. (*Proc. Phys. Soc.*, vol. 69, pp. 98–113; January 1, 1956.) The theoretical estimates of plasma resonance effects by Kaiser and Closs (2208 of 1952) were tested at a frequency of 55.3 mc. The polarization effects observed are in good agreement with the predictions for both short- and long-duration echoes, but some unexpected results were obtained for meteor trails with line densities in the transition region of 10^{12} electrons/cm. The echo amplitude A was found to be proportional to $T^{0.3}$ instead of the predicted $T^{3/16}$, where T is the duration of the echo. The discrepancy may be the result of diffusion by turbulence in the atmosphere. Part 3: 2493 of 1952 (Greenhow).

523.5:621.396.96 2039

Meteor Echo Durations and Visual Magnitudes—P. M. Millman and D. W. R. McKinley. (*Canad. J. Phys.*, vol. 34, pp. 50–61; January, 1956.) The statistical relation between the radio-echo duration and the visual magnitudes was analyzed using observations obtained during the years 1948–1950 on about 3300 meteors. Over a range of magnitudes the relation is linear or nearly linear. It is inferred that a meteor of absolute magnitude +5 produces 2×10^{12} electrons per cm of path length.

523.5:621.396.96 2040

Radar-Echo Duration and Height of a Perseid Meteor—D. W. R. McKinley. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 76–82; February, 1956.) Triangulation data for a meteor echo lasting 549 seconds, observed on 12th August 1948, are discussed.

523.7 2041

Daily Maps of the Sun—P. A. Wayman. (*Nature, Lond.*, vol. 177, pp. 518–519; March 17, 1956.) A brief note announcing the inauguration of a service in which a number of observatories are cooperating to produce comprehensive records of solar phenomena.

523.75:550.385 2042

Solar Corona and Geomagnetism—M. Notuki, Y. Nakagomi and M. Fukatsu. (*Rep. Ionosphere Res. Japan*, vol. 9, pp. 215–221; December, 1955.) Both coronal activity and geomagnetic disturbance vary directly with general solar activity except at the period of sunspot minimum, when correlation between the two is negative. Geomagnetic disturbance is greatest when the earth is in the radial line of an active coronal center. An active center appears to have a shielding effect on the outflow of corpuscles from the undisturbed solar areas.

523.78:523.72.029.6 2043

A Model for the Solar Enhanced Region at Centimeter Range derived from Partial Eclipse Observations—T. Hatanaka, K. Akabane, F. Moriyama, H. Tanaka, and T. Kakinuma. (*Rep. Ionosphere Res. Japan*, vol. 9, pp. 195–204; December, 1955.) Records taken at three Japanese observatories on frequencies between 3 and 4 kmc during the partial eclipse of the sun on June 20, 1955 show a marked decrease in the observed flux during the period when a large sunspot group was eclipsed; the location, size and brightness distribution of the enhanced-radiation region are derived. A model for the quiet sun, having a bright region near the limb on the equator is also suggested.

550.385 2044

S_q-Field in the Polar Region on Absolutely Quiet Days—T. Nagata and H. Mizuno. (*J. Geomag. Geoelect.*, vol. 7, pp. 69–74; September, 1955.) Analysis of data for the Second Polar Year indicates that the S_q field, which is recognized as applying to the region between latitudes 60 N and 60 S, also represents the daily variation in the geomagnetic field over the rest of the earth for absolutely quiet days.

550.385.523.78 2045

Report of Observations of Geomagnetic Variations at Aso and Naze (Amani-Oshima). during the Solar Eclipse of June 20th, 1955—M. Ota, H. Maeda, H. Yasuhara, and S. Hashizume. (*J. Geomag. Geoelect.*, vol. 7, pp. 86–90; September, 1955.)

551.510.53:534.22 2046

Introductory Theory for Upper Atmosphere Wind and Sonic Velocity Determination by Sound Propagation—G. V. Groves. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 24–38; February, 1956.) Theory applicable to a rocket-grenade experiment is developed.

551.510.534 2047

The Atmospheric Ozone as Indicator of Air Streams in the Stratosphere: Part 1—The Photochemical Bases—H. U. Dütsch. (*Arch. Met. A, Wien*, vol. 9, pp. 87–119; December 1, 1955.)

551.510.535 2048

A Method of determining the Relative Amounts of D- and E-Region Absorptions of Medium and Short Radio Waves—A. P. Mitra. (*Indian J. Phys.*, vol. 29, pp. 518–521; November, 1955.) The method is based on the concept of "relaxation time" [3289 of 1953 (Appleton)], the value of which for the D layer is appreciably different from that for the E layer.

551.510.535 2049

Region E and the S_q Current System—W. J. G. Beynon and G. M. Brown. (*Nature, Lond.*, vol. 177, pp. 583–584; March 24, 1956.) The complete equilibrium equation for the electron density in a solar-controlled ionospheric region includes a term which depends on vertical drift and is usually neglected; the E-layer critical frequency f_E is then related to the zenith angle χ by the equation $(f_E)^n = K \cos \chi$, where K is a constant and, for an isothermal region, n has the value 4 for recombination and 2 for attachment. Values of n deduced from actual observations lie between 2 and 4; these findings are discussed on the assumption that the variation of n results from neglect of the vertical-drift term. The theory is supported by observations of singularities in the noon curves of n/latitude for several longitudes; these singularities, corresponding to "normal" values of f_E, coincide approximately with the position of the foci of the S_q current system.

551.510.535 2050

A New Method of analysing Ionospheric Movement Records—G. L. Rogers. (*Nature, Lond.*, vol. 177, pp. 613–614; March 31, 1956.) Technique based on optical diffraction systems is outlined. The ionosphere is provisionally assumed to be a plane reflector, and the image of a ground transmitter in it is regarded as a source of coherent radiation giving rise to a diffraction pattern on the ground corresponding to an inhomogeneity moving in or below the ionosphere. From the geometry of the system a focal length is deduced, and when the record of the diffraction pattern is converted into a variable-density photographic record a corresponding visual focal length is obtained, from which the velocity of the moving inhomogeneity can be determined. The method has been tried at the New Zealand Dominion Physical Laboratory, using antennas arranged at the corners of a triangle as suggested by Mitra (96 of 1950). Holograms showing some results are reproduced; from these it is clear that moving objects far below the iosnosphere can affect the records significantly.

551.510.535 2051

Movements of Irregularities in the E Region: Part 2—T. Obayashi. (*J. Radio Res. Labs., Japan*, vol. 2, pp. 413–417; October, 1955.) Continuing the work referred to in 418 of 1956, results for the period July, 1954–March, 1955 show that an east wind with an average velocity of 80 m/second predominates in the evening, continuing throughout the night in winter; towards sunrise winds in the reverse direction are frequent. Irregularities are estimated to be 100–200 km in extent.

551.510.535 2052

Sequential E_s and Lunar Effects on the Equatorial E_s—S. Matsushita. (*J. Geomag. Geoelect.*, vol. 7, pp. 91–95; September, 1955.) Among the various types of E_s which have been observed, one shows apparent vertical movement on the ionogram, and has been termed "sequential E_s" by investigators at the National Bureau of Standards. A study is is made of this phenomenon using records from a number of stations; the subtype investigated is that involving an E_s region which first appears at a height of about 200 km in winter and 180 km in summer and then drops to normal E level, where it persists for some hours. The latitude and time distributions of the phenomenon are briefly discussed.

551.510.535 2053

On Anomalous Variations of Critical Frequencies and Virtual Heights of the F_1 and E regions of the Ionosphere—T. Sato. (*Rep. Ionosphere Res., Japan*, vol. 9, pp. 205–214; December, 1955.) Latitudinal and seasonal variations of critical frequencies and virtual heights of the F_1 and E layers, for years of high and low sunspot activity, are investigated. Anomalous variations are found, analogous to those in the F_2 layer (3254 of 1955) similar explanations are suggested.

551.510.535 2054

The Diurnal and Annual Variations of F_1 Ionization. An Interpretation—O. Burkard. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 83–90; February, 1956. In German.) Values of x and $\log C$ in the formula $(f_0F_1)^2 = C^2 (\cos \chi)^x$, calculated from CRPL records for 1934–1954, are tabulated. The variation of these values is discussed and a model is derived for F_1-layer conditions in which the temperature gradient is smaller in summer than in winter.

551.510.535 2055

A Study of the Total Electron Content of the F-Region of the Ionosphere over Ahmedabad (23°N, 72° 38′E), India—R. M. Sheriff. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 91–97; February, 1956.) The total number of electrons in a column of unit cross section in the F_1 and F_2 regions up to the height h_p of maximum electron density has been calculated for three magnetically quiet and three disturbed days in each month from February, 1953 to January, 1954. The method of analysis is that suggested by Ratcliffe (1292 of 1952) assuming parabolic electron density distribution. The relation between the semithickness y_m and h_p for the F_2 layer shows that thick layers are associated with higher values of h_p and that the parabolic-distribution law does not hold for very thick layers.

551.510.535 2056

Determination of the F-Region Collisional Frequency (over Calcutta)—M. Ghosh. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 116–118; February, 1956.) The collision frequency ν reaches its maximum ($\sim 5\times 10^2$/second) at about noon, and its minimum ($\sim 10^3$/second) at about midnight. The rate of rise of ν increases with $\cos \chi$.

551.510.535 2057

A Possible Explanation of the Drop in F-Region Critical Densities accompanying Major Ionospheric Storms—M. J. Seaton. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 122–124; February, 1956.) An explanation in terms of increased recombination rates is suggested, based on evidence that the abundance of O_2 at great heights is governed by vertical transport. ["Rocket Exploration of the Upper Atmosphere," pp. 361–365; 1954. (Nicolet)]

551.510.535:523.16 2058

The Spectrum of Radio-Star Scintillations and the Nature of Irregularities in the Ionosphere—J. P. Wild and J. A. Roberts. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 55–75; February, 1956.) Simultaneous observations of the intense source in Cygnus were made with a) a frequency-sweep spectroscope, b) a frequency-sweep interferometer, and c) a triangular spaced-antenna receiving system. Most fluctuations are due to focusing by single lens-like irregularities. The fluctuation amplitude shows two maxima: one near midnight (winter); the other near midday (summer). For daytime conditions at least, the elongation of the pattern at the ground indicates marked anisotropy in the ionosphere irregularities.

551.510.535:523.746 2059

Correlation between Noon f_0F_2 and Sunspot Number—J. M. Roy. (*J. Instn Telecommun. Engrs., India*, vol. 2, pp. 45–47; December, 1955.) Analysis of data from four Indian stations shows a linear relation between the running means of the sunspot number R and the noon value of f_0F_2, up to a limiting value of *R*.

551.510.535:523.78 2060

Tilts in the Ionosphere during the Solar Eclipse of 30 June 1954—E. N. Bramley. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 98–104; February, 1956.) "Directional measurements at nearly vertical incidence on the *F* layer during the eclipse showed the existence of a tilt whose direction agreed with that expected from the geometry of the eclipse. The magnitude of the tilt was also of the same order as that calculated from observed changes in the height of reflection. Oblique-incidence bearing measurements on signals reflected from the normal *E* layer failed to reveal any eclipse effect, and theoretical calculation showed that no detectably large effect would have been expected in this case."

551.510.535:523.78 2061

Preliminary Results of the Ionospheric Solar Eclipse of 25 December 1954—M. E. Szendrei and M. W. McElhinny. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 108–114; February, 1956.) An analysis of critical-frequency data obtained in Grahamstown during this annular eclipse. The effects may be explained to a first approximation by assuming that the radiation responsible for the ionosphere structure is distributed uniformly over the sun's disk. Values of recombination coefficient for the E_1, F_1 and F_2 layers are in good agreement with values obtained recently by independent methods.

551.510.535:523.78 2062

Observations of the Lower Ionosphere during the Solar Eclipse on 30th June 1954—K. Sprenger and E. A. Lauter. (*Gerl. Beitr. Geophys.*, vol. 64, pp. 284–312; 1955.) Propagation tests were made at Kühlungsborn on frequencies of 185 kc—1.223 mc during the eclipse. The expected increase of reflection coefficient in the lower ionosphere was considerably delayed after first contact, and normal attenuation values were restored before the end of the eclipse. The magnitude of the effect (about 40 db) and the delay of the maximum (about 4 minutes) agreed with predictions from normal diurnal and seasonal variations based on a single-layer model of the D layer, but the limited duration of the effect was consistent with a multilayer model. Observations at complex variations of very-long-wave atmospherics are also reported.

551.510.535:550.38 2063

The Influence of the Geomagnetic Field on Turbulence in the Ionosphere—J. W. Dungey. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 39–42; February, 1956.) EM damping of turbulent motion of neutral particles is negligible for the smaller eddies. In the higher regions electrons are constrained to move nearly parallel to the magnetic field; this results in variations of electron density, even when the density of neutral molecules does not vary. These variations may be appreciable above 100 km, *e.g.*, ~5 per cent at 110 km.

551.510.535:550.385 2064

On the Disturbance Daily Variations and the Lunar Daily Variations in the F_2 Region of the Ionosphere on the Magnetic Equator—H. Maeda. (*J. Geomag. Geoelect.*, vol. 7, pp. 75–85; September, 1955.) Continuation of work reported previously [*e.g.* 425 of 1956 (Maeda *et al.*)] on the effects of vertical electron drift due to the electrical fields associated with the currents responsible for the diurnal geomagnetic variations.

551.510.535:621.396.11 2065

D-E Layer Electron Model reduced from Considerations of M.F. and H.F. Wave Absorption—Kobayashi. (See 2194.)

551.510.535:621.396.11 2066

The Z Propagation Hole in the Ionosphere—Ellis. (See 2195.)

551.510.535:621.396.11 2067

The Interpretation of Measurements of Radio-Wave Interaction—Huxley. (See 2197.)

551.543 2068

A Comparison of the Annual Mean Solar and Lunar Atmospheric Tides in Barometric Pressure, as regards their Worldwide Distribution of Amplitude and Phase—S. Chapman and K. C. Westfold. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 1–23; February, 1956.)

551.594.221:551.508.94 2069

The Influence of Individual Variations in the Field Changes due to Lightning Discharges upon the Design and Performance of Lightning Flash Counters—E. T. Pierce. (*Arch. Met. A, Wien*, vol. 9, pp. 78–86; December 1, 1955. In English.) Flash counters responding only to the es component of the field change are likely to be least subject to error; this implies a preferential selection of very low frequencies. Examples are given indicating the extent of the inaccuracies encountered. From statistics of rates of flashing and duration of discharges it is concluded that counters should have an insensitive period of about 1 second. The design of counters is discussed and a simple circuit is presented.

551.594.6 2070

Low Audio-Frequency Electromagnetic Signals of Natural Origin—R. E. Holzer and O. E. Deal. (*Nature, Lond.*, vol. 177, pp. 536–537; March 17, 1956.) Signals in the frequency range 25–130 cps have been recorded in California over long periods. The diurnal variation of signal amplitude closely resembles the diurnal variation of atmospheric potential gradients measured at sea; it is inferred that these low-frequency signals are atmospherics and that their mean amplitude is roughly proportional to the number of storms in progress over the whole world.

523.746+550.385 2071

Sunspot and Geomagnetic-Storm Data derived from Greenwich Observations 1874–1954 [Book Review]—Publishers: H. M. Stationery Office, London, 1955, 106 pp. (*Nature, Lond.*, vol. 177, p. 499; March 17, 1956.) Includes explanatory notes and literature references and some diagrams.

LOCATION AND AIDS TO NAVIGATION

527.6:531.383 2072

Inertial Air Navigation Systems—(*Tele-Tech and Electronic Ind.*, vol. 15, pp. 61, 122; January, 1956.) Systems are discussed in which the velocity and position of an aircraft at any time are determined by integrating the outputs of three accelerometers; a two-axis gyro device is used. No information is required from outside sources, and no detectable signal is radiated.

621.396.933+621.396.969 2073

Communication and Navigational Aids for the Bristol Britannia—N. G. Anslow. (*Brit. Commun. Electronics*, vol. 3, pp. 6–9; January, 1956.) A brief description of the radio and radar installation in a modern long-range airliner is given.

621.396.96 2074

Radar Operation and Data Collection desired during Tornadoes and Other Severe Weather Conditions—(*Bull Amer. Met. Soc.*, vol. 36, pp. 289–291; June, 1955.) Procedures recommended by the Committee on Radar Meteorology of the American Meteorological Society are indicated.

621.396.96:621.396.662 2075

Radar A.F.C. System uses Mechanical Tuning—J. L. Confalone and W. R. Rambo. (*Electronics*, vol. 29, pp. 138–141; April, 1956.) For searching, the radar receiver local oscillator is tuned to within a few mc by a motor-driven system controlled by signals from the IF amplifier; the system then switches to afc action, effected by means of a discriminator controlling a two-phase motor. A bistable trigger circuit provides automatic switching between the two states. The complete circuit is shown.

621.396.963 2076

Clutter on Radar Displays—J. Croney. (*Wireless Engr.*, vol. 33, pp. 83–96; April, 1956.) "An analysis is developed of the action of an idealized logarithmic receiver followed by a differentiating circuit (high-pass filter), upon inherent receiver noise, sea-clutter, and rain-clutter echoes. The analysis is extended to estimate the extent to which the performance of the practical logarithmic receiver may depart from that of the idealized receiver. Results are given of experiments with logarithmic receivers on both *S*- and *X*-band radars. The loss which occurs when a differentiating circuit follows a logarithmic receiver is stated, the cause examined and a method of minimizing the loss suggested. The important design parameters of a logarithmic receiver for clutter reduction are dealt with." See also 2953 of 1954.

621.396.963:621.385.832 2077

Storage-Tube Device simulates Radar Net—S. Shenfeld and M. Finkle. (*Electronics*, vol. 29, pp. 181–183; April, 1956.) Signals in the form in which they would be received from two or more geographically separated stations are applied on a time-sharing basis to a storage-type cathode-ray tube provided with means for off-centering the beam in the *X*- and *Y*-directions. On reading out, the signals are separated and directed to individual indicator tubes corresponding to the individual stations.

MATERIALS AND SUBSIDIARY TECHNIQUES

533.5+544.4 2078

The Analysis of Gases at Low Pressures—M. A. Cayless. (*Brit. J. Appl. Phys.*, vol. 7, pp. 13–16; January, 1956.) Apparatus for handling gas in measured quantities of 10^{-6} cm^3 and over at pressures down to 10^{-6} mm Hg, and for effecting analysis at pressures between 10^{-3} and 1 mm Hg is described.

535.37:546.41.33.185-85 2079

Modified Calcium Pyrophosphate Phosphors—D. E. Kinney. (*J. Electrochem. Soc.*, vol. 102, pp. 676–681; December, 1955.) The partial substitution of Na for Ca increases the luminescence efficiency.

535.37:546.472.21 2080

Some Optical Properties of New Zinc-Sulphide Phosphors activated with Rare-Earth Elements—Z. A. Trapeznikova and V. V. Shchaenko. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, pp. 230–232; January 11, 1956. In Russian.)

535.376 2081

The Enhancement Effect of Electric Fields on some X-Ray-Excited Phosphors—G. Destriau, J. Mattler, M. Destriau, and H. E. Gumlich. (*J. Electrochem. Soc.*, vol. 102, pp. 682–684; December, 1955.) Experimental study of the influence of the strength and frequency of the applied field, the X-ray beam intensity and the temperature on the "permanent" enhancement effect. See also 2959 of 1954 (Destriau).

535.376:546.472.21 2082

Frequency Dependence of Electroluminescent Brightness—W. Lehmann: C. H. Haake. (*Phys. Rev.*, vol. 101, pp. 489–491; January 1, 1956.) ZnS phosphors activated with Cu and containing small quantities of Fe, Co or Ni are discussed; two different theories are advanced regarding the frequency variation of the electroluminescence.

537.226/.228.1:546.431.824-31 2083
Electromechanical Properties of Barium Titanate Ceramics—G. Mesnard and L. Eyraud. (*J. Phys. Radium*, vol. 16, pp. 926-938; December, 1955.) The elastic, electrostrictive, and dielectric properties of $BaTiO_3$ ceramics have been investigated by a resonance method, using disk specimens with their faces silvered to form capacitors; the equivalent circuit is derived, also the Q factor and the coefficient of electromechanical coupling for static fields up to 20 v/cm and for the temperature range from −150° to +150° C.

537.226/.227:546.41/.431].824-31 2084
Structural Behaviour in the System (Ba, Ca, Sr) TiO_3 and its Relation to Certain Dielectric Characteristics—M. McQuarrie. (*J. Amer. Ceram. Soc.*, vol. 38, pp. 444-449; December, 1955.) For compositions near the solubility limits in the ternary system (Ba, Ca, Sr) TiO_3 firing temperature has a marked effect on dielectric properties. No evidence of ferroelectric properties was discovered in $CaTiO_3$-$SrTiO_3$ systems.

537.226/.227:546.431.824-31 2085
Twinning in Barium Titanate Crystals—E. A. D. White. (*Acta Cryst.*, vol. 8, Part 12, p. 845; December 10, 1955.)

537.226/.227:546.431.824-31 2086
Dependence of the Coercive Force and Permittivity of Ceramic Barium Titanate on Mechanical Strains—N. A. Roi. (*Akust. Zh.*, vol. 1, pp. 352-355; October/December, 1955.) Experimental results indicate that the coercivity can be increased by means of mechanical tension. The form of the permittivity/temperature curves cannot be explained on the basis of a simple thermodynamic theory neglecting the effects of changes in the domain structure.

537.226/.227:[546.48+546.33].882.5 2087
Solid-Solution Effects, Structural Transitions and Ferroelectricity in Sodium-Cadmium Niobates—B. Lewis and E. A. D. White. (*Acta Cryst.*, vol. 8, Part 12, p. 849; December 10, 1955.) Results of a detailed experimental investigation of $NaNbO_3$-$Cd_2Nb_2O_7$ ceramics indicate that at any temperature the local Cd concentration within each crystallite determines whether the structure is ferroelectric or antiferroelectric. Macroscopically, the ratio of the two modifications depends on the overall Cd concentration and on the temperature.

537.227/.228 2088
Electrostriction—H. F. Kay. (*Rep. Progr. Phys.*, vol. 18, pp. 230-250; 1955.) A survey paper. Ferroelectric materials are discussed with particular emphasis on the ceramic-oxide group, in which striction coefficients can be obtained which are high compared with those of true piezoelectric materials having irreversible polarity. There is still considerable doubt as to the exact mechanisms involved. About 50 references.

537.228.1:547.476.3 2089
Low-Temperature Infrared Absorption Spectrum of Crystallized Rochelle Salt (4400-7100 cm^{-1})—M. P. Bernard. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1012-1013; February 20, 1956.

537.311.1/.31 2090
Relaxation Fluctuations in Condensed Systems—P. S. Zyryanov. (*Zh. Eksp. Teor. Fiz.*, vol. 29, pp. 334-338; September, 1955.) The classification of fluctuations into small-scale (relaxation-type) and large-scale (vibration-type) is considered. The role of relaxation fluctuations in the electrical conductivity of metals is discussed.

537.311.33 2091
New Semiconducting Compounds—E. Mooser and W. B. Pearson. (*Phys. Rev.*, vol. 101, pp. 492-493; January 1, 1956.) Experiments have shown that a large number of compounds of the metalloids Se and Te, which are designated particularly, exhibit typical semiconductor properties.

537.311.33 2092
Conductivity and Hall Effect in Semiconductors—G. Della Pergola. (*Ricerca, Sci.*, vol. 25, pp. 3269-3314; December, 1955.) Theoretical and experimental methods of investigating semiconductors are reviewed. Ionization energies of Ge and Si with various impurities are tabulated. 56 references.

537.311.33 2093
Electronic Properties of Aromatic Hydrocarbons—D. C. Northrop and O. Simpson. (*Proc. Roy. Soc. A*, vol. 234, pp. 124-149; January 24, 1956.) Report of an experimental investigation of electrical conductivity and fluorescence transfer in solid solutions of these substances.

537.311.33 2094
Theory of High-Conductivity Semiconductors—L. L. Korenblit and T. Ya Shraifel'd. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1019-1025; June and pp. 1182-1189; July, 1955.) The electrical properties of some semiconductors over certain ranges of temperature and impurity concentration are quasi-metallic. A detailed mathematical analysis is presented for equilibrium conditions of the current carriers. Formulas are derived for the temperature dependence of conductivity, thermo-emf and Hall effect for both degenerate and nondegenerate semiconductors.

537.311.33 2095
Formation of Mixed Crystals of $A^{III}B^{V}$ Compounds—O. G. Folberth. (*Z. Naturf.*, vol. 10a, pp. 502-503; June, 1955.) Investigations on InAs-InP and GaAs-GaP compounds are briefly reported. Crystals with any composition in these systems can be formed, the width of the energy gap ranging from 0.33 to 1.25 ev and from 1.45 to 2.25 ev respectively.

537.311.33 2096
Isomorphism of Compounds of Type $A^{III}B^{V}$—N. A. Goryunova and N. N. Fedorova. (*Zh. Tech. Fiz.*, vol. 25, pp. 1339-1341; July, 1955.) A property important for many technical applications, *viz.*, high mobility of current carriers, has been observed in inorganic semiconductors with a covalent type of bond. In order to be able to vary the properties of such substances, use can be made of the phenomenon of isomorphism, *i.e.*, the ability of the substances to form substitution solid solutions. A report is presented on a radiographic investigation into this property for a wide range of concentrations of the arsenides and antimonides of Ga and In.

537.311.33 2097
Bipolar Diffusion of Charge Carriers in Semiconductors in the case of Spherical Symmetry in the Presence of an External Field (Linear Approximation)—M. F. Deigen. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1175-1181; July, 1955.)

537.311.33 2098
The Dependence of the Lifetime of Excess Current Carriers on the Concentration of Equilibrium Charge Carriers—R. Paramonova and A. Rzhanov. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1342-1344; July, 1955.) In connection with an experimental verification of a formula for the lifetime of holes [420 of 1953 (Shockley and Read)], specimens were prepared from a single crystal of Ge. It is not possible to check directly whether the concentration of recombination centers remains constant along the crystal when the concentration of impurities is varied. Indirect arguments support this view, as does also the experimentally obtained linear variation of the lifetime with the inverse value of the concentration of equilibrium charge carriers.

537.311.33:535.33/.34 2099
Study of Absorption and Reflection Spectra in the Visible and Near-Ultraviolet Regions, for some Semiconductors in the form of Thin Plates, at the Temperature of Liquid Nitrogen—S. Nikitine and R. Reiss. (*C.R. Acad. Sci., Paris*, vol. 242, pp. 1003-1005; February 20, 1956.) Spectral lines attributable to excitons were observed with Cu and Tl halides.

537.311.33:538.63 2100
Theory of the Magnetic Blocking Layer in Semiconductors—O. Madelung, L. Tewordt, and H. Welker. (*Z. Naturf.*, vol. 10a, pp. 476-488; June, 1955.) Previous work [3590 of 1954 (Weisshaar and Welker)] is extended, with particular attention to the distribution of the electron-hole pairs under the influence of the crossed fields, and to the current/voltage characteristic and its relation to surface recombination and specimen dimensions. Photoeffects, frequency variation, and the growth and decay of the blocking layer in response to the field variations are also discussed.

537.311.33:[546.23+546.28+546.289 2101
The Measurement of the Energy Gap of Semiconductors from their Diffuse Reflection Spectra—P. D. Fochs. (*Proc. Phys. Soc.*, vol. 69, pp. 70-75; January 1, 1956.) The values deduced by this method for the energy gaps at room temperature are: amorphous Se 1.86 ev, metallic Se 1.74 ev, Si 1.20 ev and Ge 0.69 ev.

537.311.33:546.23 2102
Electron-Optical Investigations of Selenium—W. Theis. (*Z. Naturf.*, vol. 10a, pp. 503-504, 464b; June, 1955.) Crystal structures observed by electron-diffraction technique in Se films deposited on preheated base plates are illustrated and discussed.

537.311.33:[546.289+546.682.86 2103
Impurity Scattering in Semiconductors—R. Mansfield. (*Proc. Phys. Soc.*, vol. 69, pp. 76-82; January 1, 1956.) Theory is developed and the combination of impurity and lattice scattering is considered for the general case of any degree of degeneracy of charge carriers. Theoretical and experimental results are compared for Ge and InSb.

537.311.33:546.289 2104
The Effect of Heat Treatment on the Concentration and Mobility of Charge Carriers in Germanium—V. V. Ostroborodova and S. G. Kalashnikov. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1163-1167; July, 1955.) A report is presented on experiments in which Ge specimens were heated to a temperature of 500°C. or higher, and then quenched in oil at room temperature. Measurements indicate that the mobility of majority and minority carriers decreases with the approach to the transformation temperature; this may be due to an increase in the volume heterogeneities in the crystal. The concentration of thermal acceptors varies linearly with the reciprocal of temperature in passing from the p- into the n-region.

537.311.33:546.289 2105
The Recombination of Non-equilibrium Charge Carriers at Thermal Acceptors in Germanium—V. V. Ostroborodova and S. G. Kalashnikov. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1168-1174; July, 1955.) The effect of heat treatment on the rate of the volume recombi-

nation of nonequilibrium electrons and holes in Ge was investigated experimentally; the lifetime varies in inverse proportion to the concentration of thermal acceptors. The effective recombination cross section of the thermal acceptors was 2.5×10^{-17} cm^2. An estimate of the upper limit of the recombination cross section for donors shows that its order cannot exceed 10^{-19} cm^2.

537.311.33:546.289 **2106**

Action of Etchants on Germanium Single Crystals—G. Della Pergola and D. Sette. (*Alta Frequenza*, vol. 24, pp. 499–518; December, 1955.) Four different etchants were applied to the (100) and (111) faces of the crystals for various periods; microscope examinations and resistivity measurements were made after each treatment. The action of the etchants is discussed in terms of energy released, and is illustrated by photomicrograms.

537.311.33:546.289 **2107**

Magnetic Susceptibility of Low-Resistivity *n*-Type Germanium—F. T. Hedgcock. (*Canad. J. Phys.*, vol. 34, pp. 43–49; January, 1956.) Measurements on polycrystalline specimens over a range of temperatures below room temperature are reported. From the experimental results, the contributions of the free and bound charge carriers to the susceptibility are determined; these appear to vary inversely with temperature. The value deduced for the effective mass of the carriers is 0.16 times that of the free electron mass.

537.311.33:546.289 **2108**

Review of Germanium Surface Phenomena—R. H. Kingston. (*J. Appl. Phys.*, vol. 27, pp. 101–114; February, 1956.) "In general the surface may be treated as an assemblage of allowed electron states occurring in the normally forbidden energy range. A review of the measurements of the electrical properties suggests that there are two distinct types of state. The *fast* state has a hole or electron capture time not greater than a microsecond and is chiefly involved in the recombination process. The *slow* state has capture times from a millisecond to several minutes and determines the density and type of carrier at the surface. *Fast* states are believed to occur at the interface between the germanium and the oxide layer, and their density of about 10^{11} cm^{-2} is determined by the initial surface treatment. *Slow* states are associated with the structure of the oxide layer and the gaseous ambient, and have a density greater than 10^{13} cm^{-2}. Since these states determine the conductivity type at the surface, they contribute to surface *leakage* in diodes and transistors and, because of their long equilibrium times, to low-frequency noise. The adsorption of gases such as water vapor, not only controls the density and energy of the *slow* states but also leads to possible electrolytic conduction along the surface, in addition to the normal electron flow in the bulk semiconductor."

537.311.33:546.289 **2109**

Use of Infrared Absorption to determine Carrier Distribution in Germanium and Surface Recombination Velocity—N. J. Harrick (*Phys. Rev.*, vol. 101, pp. 491–492; January 1, 1956.) Experiments are briefly described which indicate that surface recombination velocity can be directly evaluated from measurements of infrared absorption. In a particular case, values of 250 cm and 1900 cm were obtained for surfaces which had been etched and ground respectively.

537.311.33:546.289 **2110**

Infrared Absorption in *n*-Type Germanium—H. Y. Fan, W. Spitzer, and R. J. Collins. (*Phys. Rev.*, vol. 101, pp. 566–572; January 15, 1956.) Measurements were made at wavelengths from 5 to 38 μ, at temperatures from 78° to 450°K. At the higher temperatures the absorption is proportional to carrier concentration, lattice scattering beging the dominant effect. At 78° the absorption per unit carrier concentration comprises a constant term together with a term proportional to the impurity concentration. The absorption increases with wavelength more rapidly at the lower temperatures. The results are in quantitative agreement with theory if the effective carrier mass is assumed to be about 0.1 *m*.

537.311.33:546.289 **2111**

Deformation Potential Theory for *n*-Type Ge—W. P. Dumke. (*Phys. Rev.*, vol. 101, pp. 531–536; January 15, 1956.) A calculation is made of the mobility of electrons in Ge using the deformation potential theory of Bardeen and Shockley (3032 of 1950) and taking into account the effect of shear-wave scattering. Values obtained range between 4550 and 6700 cm per V/cm at 300°K, with a temperature-variation index of $-3/2$. Discrepancies between these results and experimental values are discussed.

537.311.33:546.289 **2112**

Hall Effect in Oriented Single Crystals of *n*-Type Germanium—W. M. Bullis and W. E. Krag. (*Phys. Rev.*, vol. 101, pp. 580–584; January 15, 1956.) Experimental evidence confirms the variations of Hall coefficient with direction and magnitude of the magnetic field and with direction of current which are predicted by the eight-ellipsoid energy-surface model. The observations can be explained by assuming an energy-independent scattering time. This type of measurement could be used to determine symmetry properties of energy surfaces near a band edge.

537.311.33:546.289:535.215:538.6 **2113**

Saturation of the Photomagnetoelectric Effect in Germanium as a Function of the Magnetic Field—A. A. Pires de Carvalho. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 745–747; February 6, 1956.) Measurements on small plates of Ge subjected to illumination and to magnetic induction flux densities *B* up to about 2.7 Wb/m^2 indicate that the short-circuit current becomes saturated at values of *B* consistent with the formula $V=CB/[1+(\mu B)^2]$, where μ is the charge-carrier mobility, *C* is a constant depending on the wavelength and intensity of the illumination, and *V* is the photomagnetoelectric voltage.

537.311.33:546.289:538.63 **2114**

Magnetic Blocking Layers in Germanium: Part 2—E. Weisshaar. (*Z. Naturf.*, vol. 10a, pp. 488–495; June, 1955.) Experimentally determined *I/V* characteristics of symmetrical and asymmetrical blocking layers developed in magnetic fields [3590 of 1954 (Weisshaar and Welker)] and their deviations from the theoretically predicted curves are discussed. The absence of a saturation effect is attributed to the impurity concentration of the Ge specimen used. Measurements of the growth and decay of the layers and of the frequency variation of the current at constant voltage are in good agreement with values calculated from theory for small values of the applied field.

537.311.33:546.46-31:535.215 **2115**

Photo-induced Hall Effect in MgO—E. Yamaka and K. Sawamoto. (*Phys. Rev.*, vol. 101, pp. 565–566; January 15, 1956.) Short report of an investigation of the sign of the charge carriers produced by light in the various absorption bands. A hole mobility of about 2 cm per V/cm was found.

537.311.33:[546.472.21+546.682.86]:537.228.1 **2116**

Mobility in Zinc Blende and Indium Antimonide—W. A. Harrison. *Phys. Rev.*, vol. 101, p. 903; January 15, 1956.) The relations between the piezoelectric properties and the charge-carrier mobility of these materials are discussed.

537.311.33:546.561-31 **2117**

Surface Conductivity of Copper Oxide—V. E. Lashkarev and V. I. Lyashenko. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, pp. 243–245; January 11, 1956. In Russian.) Results of experimental determinations of conductivity and carrier mobility in specimens placed in vaccum and in ethylene alcohol vapor at various pressures are discussed.

537.311.33:546.682.86 **2118**

Optical Properties of Indium Antimonide in the Region from 20 to 200 Microns—H. Yoshinaga and R. A. Oetjen. (*Phys. Rev.*, vol. 101, pp. 526–531; January 15, 1956.) Reflectivity and transmission curves were obtained for large *n*-type single crystals. The curves indicate a strong lattice vibration at 54.6 μ; absorption and transmission in this spectral region are not greatly affected by the presence of free electrons.

537.311.33:546.682.86 **2119**

Infrared Absorption of Indium Antimonide—E. Blount, J. Callaway, M. Cohen, W. Dumke, and J. Phillips. (*Phys. Rev.*, vol. 101, pp. 563–564; January 15, 1956.) Observations are interpreted in terms of transition between valence and conduction bands.

537.311.33:546.682.86:538.6 **2120**

Thermomagnetoelectric Effects in Indium Antimonide—P. Aigrain, C. Rigaux, and J. M. Thuillier. (*C.R. Acad. Sci., Paris*, vol. 242, pp. 1145–1148; February 27, 1956.) Observations of Seebeck and Nernst effects were in good agreement with theoretical predictions; these techniques are useful for deterermining the ratio of the effective masses and the mobilities of the carriers.

537.311.33:[546.817.231+546.817.241+546.817.221 **2121**

Photon-Radiative Recombination in PbSe, PbTe and PbS—I. M. Mackintosh. (*Proc. Phys. Soc.*, vol. 69, pp. 115–118; January 1, 1956.) Analysis indicates that in the absence of traps the maximum attainable lifetimes corresponding to direct electron-hole recombination in PbSe, PbTe and PbS are 0.6, 0.8 and 40 μs, to within factors of 2, 3, and 5 respectively.

537.311.33:621.314.63 **2122**

The Static Reverse-Voltage/Current Characteristic of the Barrier Layer formed at the Boundary between an *n*-Type Semiconductor and a *p*-Type Semiconductor—E. I. Rashba and K. B. Tolpygo. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1335–1338; July, 1955.) In existing theories of rectification at *p-n* junctions, recombination and generation of carriers in the space-charge region are not taken into account, and therefore the universal formula (1) for the current through the semiconductor is not sufficient for determining the properties of *p-n* junctions from the observed *V/I* characteristic. These effects are now considered and an examination is made of the extent to which Shockley's theories (379 of 1950) regarding the variation of quasi-Fermi levels are justifiable.

537.311.33:621.396.822 **2123**

High-Frequency Shot Noise in *P-N* Junctions—A. Uhlir, Jr., (PROC. IRE, vol. 44, pp. 557–558; April, 1956.) A quantitative relation is derived between the frequency variation of the junction conductance and the frequency variation of the shot-noise current. The result obtained is valid for nonplanar as well as planar junctions, and in cases where drift is important.

537.311.4 2124

New Low-Contact-Resistance Electrode—S. S. Flaschen and L. G. Van Uitert. (*J. Appl. Phys.*, vol. 27, p. 190; February, 1956.) The contact resistance of various electrodes was investigated by measuring the resistance of Ni-ferrite specimens to which the contacts were applied. A contact made by rubbing an indium pencil moistened with mercury on to the surface gave both a minimum and a constant resistance over the voltage range 0.01–140 v. Liquid gallium may be substituted for the mercury in some cases.

537.311.4:537.226 2125

Mechanism of Forming of Anode [-interface] Layers in Formed Dielectrics—Ya. N. Pershits. (*Zh. Eksp. Teor. Fiz.*, vol. 29, pp. 362–368; September, 1955.) Continuation of earlier experiments (448 of 1956) is reported. The current/time characteristics of the electrode/dielectric contacts were found to be similar for glass, rock salt, skin, and other dielectrics; the curves cannot be explained by assuming motion of cations only; motion of anions must be assumed. No evidence was found of electron conduction.

537.323:546.3–1–86–24 2126

The Thermoelectric Properties of Alloys of the Antimony-Tellurium System—F. I. Vasenin. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1190–1197; July, 1955.) An experimental investigation was carried out, the main results of which are as follows: a) in the region of the secondary system Te-Sb_2S_3 a slow decline of the thermo-emf curve is observed; this decline becomes a sharp fall when the stoichiometric composition is approached; b) using powder-metallurgy technique it is possible to obtain homogeneous Sb-Te alloys with increased thermo-emf and electrical conductivity; c) heating of the alloys to 400°C. increases thermo-emf, while the corresponding decrease in electrical conductivity is relatively small; d) addition of Pb decreases the thermo-emf of Sb and Sb-Te alloys and increases their conductivity.

538.22:546.711 2127

Antiferromagnetic Structure of α-Manganese and a Magnetic Structure Study of β-Manganese—J. S. Kasper and B. W. Roberts. (*Phys. Rev.*, vol. 101, pp. 537–544; January 15, 1956.)

538.22:621.318.134 2128

Comparative Study of the Crystalline Structures of Lanthanum, Praseodymium and Samarium Ferrites—G. Guiot-Guillain. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 793–795; February 6, 1956.)

538.221 2129

Law of Approach to Saturation for a Single Crystal of Fe-Si along the Three Principal Crystallographic Directions—H. Danan. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 748–750; February 6, 1956.) Measurements are reported which indicate that magnetic hardness is greatest for magnetization along the direction of greatest anisotropy energy.

538.221 2130

Demagnetization of Magnetite and of Sesquioxide of α Iron by the Action of Alternating Magnetic Fields—F. Rimbert. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 890–893; February 13, 1956.)

538.221 2131

Influence of Method of Demagnetization of Specimen on Temperature Dependence of Magnetization of Nickel in Weak Fields—A. I. Drokin and V. L. Il'yushenko. (*Zh. Eksp. Teor. Fiz.*, vol. 29, pp. 339–344; September. 1955.) The effect of demagnetization by a) heating to a temperature above the Curie point and b) applying an alternating magnetic field, on the subsequent magnetization in a field of strength 0.39 oersted at temperatures between −183°C. and +360°C. is investigated experimentally. The magnetization curves obtained are not identical, probably because in a) the alignment of the domains is random, in b) the domains are antiparallel. The number and magnitude of jumps in the curves is also different.

538.221:538.569.4.029.6 2132

Resonance in α Fe_2O_3—Y. Kojima. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, vol. 7, pp. 591–594; December, 1955.) Microwave magnetic resonance in natural single crystals of α Fe_2O_3 was observed at several frequencies ranging from 16.5 kmc to 48.3 kmc. This anisotropic energy was studied as a function of crystal orientation. The *g* factor was also obtained from the resonance data.

538.221:538.63 2133

The Problem of Galvanomagnetic Effects in Ferro-magnetic Materials—K. M. Koch. (*Z. Naturf.*, vol. 10a, pp. 496–498; June, 1955.) The tensor nature of the resistance of a ferromagnetic conductor in a magnetic field is discussed. Experiments are reported on strips of Fe, Ni and alloys arranged rotatably in the field of an electromagnet; pseudo Hall effects were observed confirming the theory presented.

538.221:621.317.411.029.6 2134

Measurement of the Magnetic Permeability of Metals by means of Cavity Resonators, and the Permeability of Iron in the Region of Ferromagnetic Resonance—K. H. Reich. (*Frequenz*, vol. 9, pp. 299–305; September, and pp. 414–422; December, 1955, and vol. 10, pp. 11–19; January, 1956.) Various theoretical explanations of the fall of initial permeability with increasing frequency are examined; experiments with ferrites have made it clear that ferromagnetic resonance is an important factor. A method of measurement at cm wavelengths is described in which part of the boundary wall of a cavity resonator is replaced by the ferromagnetic test specimen and the resulting variation of the damping is observed. The effects of pseudo-damping due to the associated circuit arrangements are corrected for. Measurements are reported on iron specimens of various degrees of purity, at room temperature and at 200°C. using direct field strengths up to 11,000 A/cm and a frequency of 24.5 kmc; the reversible permeability is independent of purity, temperature, and crystal orientation relative to the magnetic field. Variation of the permeability with surface roughness and low permeability of unordered specimens are attributed to scatter of the resonance frequencies. For related work, see 829 of 1956 (Standley and Reich). 126 references.

538.221:621.318.122 2135

Improved Permanent Magnet Materials—(*Electronic Engng.*, vol. 28, p. 171; April, 1956.) New types of alcomax with desired anisotropic properties are obtained by making suitable modifications in the composition and applying heat treatment in a magnetic field.

538.221:621. 318.134 2136

Initial Permeabilities of Sintered Ferrites—G. W. Rathenau and J. F. Fast. (*Physica*, vol. 21, pp. 964–970; December, 1955.) Measurements on Ni-Zn ferrites at low frequencies are reported. The results support the conclusion previously reached by Wijn and Went (1633 of 1952) that after demagnetization in an alternating field of decreasing amplitude the magnetization of these materials is almost exclusively caused by simultaneous rotation of the spins.

538.221:621. 318.134 2137

Complex Susceptibility of Magnesium Ferrites—B. Chiron and P. M. Prache. (*Câbles et Transm.*, vol. 10, pp. 73–74; January, 1956.) Experimental results are presented in a curve using frequency as parameter and choosing the coordinates so as to bring out significant properties of the material. Loops in this curve are attributed to causes other than gyromagnetic resonance.

538.221:621.318.134:538.6 2138

Frequency Doubling in Ferrites—W. P. Ayres, P. H. Vartanian, and J. L. Melchor. (*J. Appl. Phys.*, vol. 27, pp. 188–189; February, 1956.) " A high intensity microwave magnetic field and an orthogonal dc magnetic field were applied to a ferrite body. Double-frequency magnetic fields were generated in the same direction as the dc magnetizing field. A primary frequency of 3175 mc was used with peak powers up to 200 watts. The peak output power at 6350 mc was found to increase linearly with the square of the peak input power over a range of 42 db. Reasonable agreement was found between theory and experiment."

538.222:538.569.4 2139

Paramagnetic Resonance II—K. D. Bowers and J. Owen. (*Rep. Progr. Phys.*, vol. 18, pp. 304–373; 1955.) A report complementary to that of Bleaney and Stevens (452 of 1954). Paramagnetic-resonance data on salts containing transition-group ions are collected and are discussed in the light of theory presented in simple form. Various physical properties of crystals comprising such ions are elucidated. Over 100 references.

538.569.4:546.87 2140

Cyclotron Resonance in Bismuth—M. Tinkham. (*Phys. Rev.*, vol. 101, p. 902; January 15, 1956.) Results obtained by previous workers are discussed in relation to the energy-band structure for Bi first proposed by Jones (*Proc. Roy. Soc. A*, vol. 147, pp. 396–417; November 15, 1934.)

538.63:546.87 2141

Galvanomagnetic Effects in Bismuth—B. Abeles and S. Meiboom. (*Phys. Rev.*, vol. 101, pp. 544–550; January 15, 1956.) "Conductivity, Hall effect, an magnetoresistance in single crystals of pure and tin-doped bismuth have been measured as functions of temperature between 80 and 300°K and as functions of magnetic field up to 2000 oersted. A simple many-valley model for the band structure of bismuth is proposed, and explicit expressions for the galvanomagnetic effects are derived. Numerical values are obtained for the number of conduction electrons and holes, their mobilities, and the overlap of valence and conduction bands."

538.65 2142

Magnetostriction and Magnetomechanical Effects—E. W. Lee. (*Rep. Progr. Phys.*, vol. 18, pp. 184–229; 1955.) A review with about 100 references.

538.652:546.74 2143

Theory of Magnetostriction of Single Crystals of Ni—N. S. Akulov. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, pp. 31–34; January 1, 1956. In Russian.)

546.841.05 2144

Preparation of High-Purity Thorium by the Iodide Processes—N. D. Veigel, E. M. Sherwood, and I. E. Campbell. (*J. Electrochem. Soc.*, vol. 102, pp. 687–689; December, 1955.)

548.5:546.482.21 2145

New Method of Preparation of Single Crystals of Cadmium Sulphide—E. Grillot. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 779–781; February 6, 1956.) By sublimation in a tubular furnace heated to above 1000°, tablets of CdS have been obtained in few hours having thicknesses of several mm and surface areas of several cm_3, constituted by a mosaic of crystals joined at their lateral faces.

669 2146
The Effect of Growth Conditions upon the Solidification of a Binary Alloy—W. A. Tiller and J. W. Rutter. (*Canad. J. Phys.*, vol. 34, pp. 96–121; January, 1956.) A theoretical and experimental analysis is reported of the influence of a) the concentration of solute in the melt, b) the rate of solidification, and c) the temperature gradient in the melt at the solid/liquid interface.

669.046.5:[546.289+546.74 2147
Floating Zone Melting of Solids by Electron Bombardment—M. Davis, A. Calverley, and R. F. Lever. (*J. Appl. Phys.*, vol. 27, pp. 195–196; February, 1956.) A rod of the solid material, held upright in chucks in a continuously evacuated enclosure, and maintained at a positive potential of a few kv, is encircled at the level chosen for melting by a heated tungsten filament cathode; a focusing shield is provided to control the length of the zone bombarded. The method is useful for dealing with Si and Ni as well as more refractory materials.

MATHEMATICS

512.37 2148
On the Calculation of the Roots of Equations—E. Frank. (*J. Math. Phys.* vol. 34, pp. 187–197; October, 1955.) The method described, which is an extension of Newton's method of successive approximations, is particularly useful for accurate computations with an ordinary calculating machine.

517 2149
Transformations of the Fourier and Laplace Types by means of Solutions of Second-Order Differential Equations—B. Levin. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, pp. 187–190; January 11, 1956. In Russian.)

517 2150
Determination of the Family of Orthogonal Polynomials whose Derivatives are Orthogonal—R. Campbell. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1110–1111; February 27, 1956.)

517.534.01 2151
The Initial-Value Problem for the Wave Equation in the Distributions of Schwartz—J. O'Keefe. (*Quart. J. Mech. Appl. Math.*, vol. 8, Part 4, pp. 422–434; December, 1955.) "The classic formulas for the solution of initial value problems for the wave equation are found in the case of distributions by a Fourier transform method."

517.942.922 2152
Tables of $\int_0^x J_0(t)dt$ for Large [values of] x—P. W. Schmidt. (*J. Math. Phys.* vol. 34, pp. 169–172; October, 1955.) The Bessel function is tabulated in steps of 0.2 for the range $10<x<40$.

517.942.932 2153
On Periodic Solutions of Duffing's Equation with Damping—W. S. Loud. (*J. Math. Phys.*, vol. 34, pp. 173–178; October, 1955.) Special cases are considered of the equation $\ddot{x}+f(x)\dot{x}+g(x)=p(t)$, where $f(x)$ is an even function, $g(x)$ is an odd function, and $p(t)$ is periodic and odd-harmonic.

518.2 2154
Formulas for Inverse Osculatory Interpolation—H. E. Salzer. (*J. Res. Nat. Bur. Stand.*, vol. 56, pp. 51–54; January, 1956.) The formulas presented provide an improved means for inverse interpolation in tables where the first derivative is either tabulated alongside the function, as in Bessel functions of the first and second kind, or where it is easily obtained, as in the elementary trigonometrical functions and their integrals.

MEASUREMENTS AND TEST GEAR

531.789.1:538.22 2155
Torsion Balance for a Single Microscopic Magnetic Particle—S. P. Yu and A. H. Morrish. (*Rev. Sci. Instrum.*, vol. 27, pp. 9–11; January, 1956.) The balance described may be used for measurements on individual magnetic particles of diameters down to about 1 μ (mass ~10^{-13} g).

538.569.4.029.608 2156
Millimeter and Submillimeter Wave Spectroscopy—C. A. Burrus and W. Gordy. (*Phys. Rev.*, vol. 101, pp. 599–602; January 15, 1956.) Apparatus and results are described. The energy source is a Si-crystal frequency multiplier driven by a cm-γ klystron. An evacuated bolometer is used as detector at wavelengths down to 2 mm. An Si-crystal detector is also used.

621.317.3:621.315.212:621.3.013.78 2157
Measurement of Coupling Impedance and its Application to the Study of Cable Screens—Bourseau and Sandjivy. (See 1950.)

621.317.33:621.372.413.:537.311.33 2158
A Microwave Resonant Cavity Method for measuring the Resistivity of Semiconducting Materials—J. G. Linhart, I. M. Templeton, and R. Dunsmuir. (*Brit. J. Appl. Phys.*, vol. 7, pp. 36–38; January, 1956.) "A small spherical or cubic specimen is placed in a cavity resonating in the H_{011} mode, and the reduction in Q due to eddy-current loss in the specimen is measured. The theory given enables the resistivity of the material to be calculated. It is shown that resistivities in the range 0.005 to 10Ω. cm can be determined."

621.317.335.3 2159
Dielectric Constant of Water from 0° to 100°C—C. G. Malmberg and A. A. Maryott. (*J. Res. Nat. Bur. Stand.*, vol. 56, pp. 1–8; January, 1956.) Measurements made using a low-frequency bridge with a Wagner earth are reported. The accuracy was within 0.1 per cent. The value found at 25°C. was 78.30, about 0.3 per cent less than that usually accepted. A formula is derived for the temperature variation of the dielectric constant. Sources of error in the experimental method are considered in detail.

621.317.335.3.029.6 2160
A Method for determining Complex Dielectric Constants of Very Small Amounts of Material (0.1 cm³) at Decimetre Wavelengths—H. Lueg and H. K. Ruppersberg. (*Arch. Elekt. Übertragung*, vol. 9, pp. 553–540; December, 1955.) A coaxial-line method is described in which the line stands vertical and the inner conductor is adjustable axially. When the inner conductor is pulled down a few tenths of a millimeter, a cavity is formed into which the test material can be inserted. In this position the field strength is a maximum, so that the specimen/air interface is nearly free from induction and field currents. With liquid specimens, two drops suffice; with solid specimens, thin disks are used.

621.317.411.029.6:538.221 2161
Measurement of the Magnetic Permeability of Metals by means of Cavity Resonators, and the Permeability of Iron in the Region of Ferromagnetic Resonance—(See 2134.)

621.317.42 2162
Methods of measuring Strong Magnetic Fields—J. L. Symonds. (*Rep. Progr. Phys.*, vol. 18, pp. 83–126; 1955.) A survey covering methods based on magnetic resonance, magnetoresistance, the Hall effect, peaking strips, the generator principle, the fluxmeter, and forces on current-carrying conductors. The technique of determining the magnetic median plane is discussed, and several less used methods of measurement are mentioned. About 150 references.

621.317.42:621.383 2163
The Photoelectric Fluxmeter—S. P. Kapitsa. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1307–1315; July, 1955.) A detailed description and analysis of operation of the "photoelectric hysteresisgraph" [*Elect. Engng., N. Y.*, vol. 56, pp. 805–809; July, 1937. (Edgar)]. The frequency characteristics are considered. Conditions for distortion correction are established and the advantages of using negative feedback are indicated. The theoretical results are illustrated by oscillograms of transient processes.

621.317.44 2164
Improved Method for observing Hysteresis Cycles with the Iliovici Permeameter—R. Delors and L. Garge. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 751–753; February 6, 1956.)

621.317.7:621.372.8 2165
Bibliography on Directional Couplers—R. F. Schwartz. (TRANS. IRE, vol. MTT-3, pp. 42–43; April, 1955.) Addendum to 208 of 1955.

621.317.7:621.396.81 2166
New Apparatus for determining the Statistical Distribution of Random Electrical Processes—J. Grosskopf, K. H. Kappelhoff, and G. Kopte. (*Nachrichtentech. Z.*, vol. 9, pp. 34–39; January, 1956.) Automatic recording apparatus for studying radio signal-strength variations is described. The time-distribution curve is displayed every half-hour; accuracy of measurements is adequate for fading frequencies up to 10 kc. The evaluation of the 18-hour daily records is performed by one operator in about half-an-hour, using templates.

621.317.7:621.396.822.1:621.396.31 2167
Equipment for Measurement of Interchannel Crosstalk and Noise on Broad-Band Multichannel Telephone Systems—R. W. White and J. S. Whyte. (*P.O. Elect. Engrs.' J.*, vol. 48, Part 3, pp. 127–132; October, 1955.) A band of noise covering the frequency range of a multichannel system is passed through a narrow-band-stop filter, giving the equivalent of a fully loaded system with one quiet channel. Intermodulation measurements may be made in the quiet channel. The method may be used to investigate coaxial-cable systems as well as microwave radio links.

621.317.71:620.193 2168
An Electronic Self-Balancing Zero-Resistance Ammeter—D. R. Maker and H. T. Francis. (*J. Electrochem. Soc.*, vol. 102, pp. 669–670; December, 1955.) Description of an instrument for continuous observation of the short-circuit current flowing in a galvanic corrosion cell. It comprises a 60-cps chopper circuit, amplifier, phase-detector, relay, and balancing circuit. The sensitivity is 50 μv with current range 20 ma.

621.317.729.029.6 2169
Measurement of Time-Quadrature Components of Microwave Signals—J. H. Richmond. (TRANS. IRE, vol. MTT-3, pp. 13–15; April, 1955.) Equipment for measuring the real and imaginary components of a signal is described; it uses two barretters as detectors and is less elaborate than would be required for separate measurement of amplitude and phase.

621.317.729.029.6 2170
Probes for Microwave Near-Field Measurements—J. H. Richmond and T. E. Tice. (TRANS. IRE, vol. MTT-3, pp. 32–34; April, 1955.) Details are given of the construction of waveguide probes which permit the calculation of far-field signal strengths in good agreement with measured values.

621.317.733 2171
Compensating Recording Bridge for Measurement of Capacitance and Loss Factor—H. Poleck. (*Elektrotech. Z., Edn A*, vol. 76, pp. 822–826; December 1, 1955.) A modified Schering bridge is described.

621.317.733 2172
A Simple Auxiliary Bridge for Accurate Measurements of Loss Factor with the Schering Bridge—A. Keller. (*Elektrotech. Z., Edn A*, vol. 76, pp. 826–827; December, 1955.)

621.317.755 2173
Signal-Triggered Sweep Magnifies Pulse Widths—R. L. Kuehn. (*Electronics*, vol. 29, pp. 146–147; April, 1956.) A cro circuit is described which incorporates an automatic change-over from recurrent-sweep to triggered-sweep operation; the display corresponding to a signal pulse can be expanded as desired by increasing the recurrent-sweep frequency.

621.317.755.029.3 2174
Audio-Frequency Spectrometer—W. Kaule and A. Johne. (*Nachrtech.*, vol. 6, pp. 35–39; January, 1956.) The operation of a new cro instrument using only six filters is described. Over the frequency range 0–1 kc the analyzing bandwidth is 50 cps; over the range 0–20.5 kc it is 500 cps.

621.317.76:621.396.61 2175
The Frequency-Band Recorder—H. Fleischer and H. Widdra. (*Nachrichtentech. Z.*, vol. 9, pp. 21–28; January, 1956.) An automatic instrument for measuring and recording the principal features of radio signals in accordance with the international monitoring system is described. The features covered are frequency, field strength, bandwidth, duration, type of modulation, and transmitter frequency deviation. The instrument can be used in conjunction with different receivers to suit the frequency and sensitivity requirements; the construction permits not more than three frequency sweeps of the receiver to be recorded per minute. The receiver feeds an amplifier which modulates a variable-delay thyratron driving the recording pen. Typical records are reproduced.

621.317.761 2176
Phase-Bridge Frequency Meters—C. Févrot. (*Rev. Gén. Élect.*, vol. 65, pp. 34–38; January, 1956.) In this type of instrument, frequency variations are converted into phase variations which in turn are converted into variations of a direct voltage by means of the phase bridge. The design of circuits to give an indication independent of signal amplitude and harmonic content is discussed, and the merits of some known arrangements for converting from frequency to phase variations are compared.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.2-8:539.32 2177
Apparatus for the Measurement of Physical Constants by the Elastic-Vibrations Method—A. V. J. Martin. (*J. Brit. IRE* vol. 16, pp. 167–183; March, 1956.) Ultrasonic techniques for determining elastic constants are surveyed. Equipment is described suitable for testing gaseous, liquid, and solid specimens. The frequency range is 0.5–5.0 mc, the temperature range 0°–100°C., and the pressure range up to 10,000 kg/cm². 40 references.

534.2-8:61 2178
Ultrasonic Irradiation of the Central Nervous System at High Sound Levels—W. J. Fry and F. Dunn. (*J. Acoust. Soc. Amer.*, vol. 28, pp. 129–131; January, 1956.)

534.2-8:61 2179
High-Intensity Ultrasonic Oscillations for Treatment of Malignant Tumours in Animals and in Man—A. K. Burov. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 106, pp. 239–241; January 11, 1956. In Russian.) Experimental results indicate that some tumors on the surface of the skin may successfully be treated using 1.5-mc ultrasonic beams of intensities up to 350 w/cm² (continuous) or 600 w/cm² (pulsed). Quartzplate transducers with surface areas up to 50 cm² were used. No constructional details are given. For experimental results see also *ibid.*, vol. 106, pp. 445–448; January 21, 1956 (Burov and Andreevskaya).

535.42:548:621.397.6 2180
Apparatus for the Electronic Presentation of Optical Diffraction Patterns—A. W. Hanson and A. Menarry. (*J. Sci. Instrum.*, vol. 33, pp. 24–27; January, 1956.) "An image of the diffraction pattern is focused on the photomosaic of a television camera tube capable of accumulating, for a period of about ten seconds, a corresponding charge pattern. This pattern is transferred to a storage device where it remains for a relatively long time during which it can be continuously presented on a television receiver."

537.534.9:621.314.632:546.28 2181
Solid-State Detector for Low-Energy Ions—O. Heinz. E. M. Gyorgy and R. S. Ohl. (*Rev. Sci. Instrum.*, vol. 27, pp. 43–47; January, 1956.) Experimental results of bombardment of silicon by He^+ ions through a silver film deposited on the surface show that the rectifying properties are modified so as to increase the turnover voltage. Ion energies up to 30 kev and dosages up to 1,800 $\mu c/cm^2$ were used.

621.3:620.16:681.142 2182
Shock Spectrum Computer for Frequencies up to 2000 c/s—Morrow and Riesen. (See 1969.)

621.3:681.8:78 2183
Electronic Music—H. Le Caine. (PROC. IRE, vol. 44, pp. 457–478; April, 1956.) A comprehensive discussion dealing with fundamental aspects and particular instruments and techniques. Various means of achieving interpretative effects are indicated. Coded-performance devices are described, including those used by the Musique Concrète group in Paris and the Cologne studio for electronic music.

621.317.39:534.6:621.385.83 2184
Electron-Acoustic Image Converter—P. K. Oshchepkov, L. D. Rozenberg, and Yu B. Semennikov. (*Akust. Zh.*, vol. 1, pp. 348–351; October/December, 1955.) The tube described uses a $BaTiO_3$ screen at which an ultrasonic field is converted into an electric-charge distribution which is scanned by an electron beam. The response at frequencies of a few mc is linear over the intensity range 3×10^{-9}–3×10^{-3} w/cm². The sensitivity of the instrument is 2×10^{-3} v/bar.

621.384.612 2185
Scattering of Protons by Residual Gas in a Synchrotron: Part 1—Elastic Scattering—J. Seiden. (*J. Phys. Radium*, vol. 16, pp. 917–925; December, 1955.) The amplitude of oscillations set up by elastic collisions of the accelerated protons with gas molecules is directly proportional to the square root of the gas pressure and inversely proportional to the square roots of the energy of injection and of the energy gain per revolution. Direct losses of protons by single collisions vary in the same way; the calculated values for certain existing synchrotrons range between 2 per cent and 14 per cent.

621.384.612 2186
Space Charge and Ionization Phenomena in Constant-Gradient Protron Synchrotrons—P. B. Moon. (*Proc. Phys. Soc.*, vol. 69, pp. 153–156; February, 1956.) The calculation of space-charge effects for constant-gradient machines is simpler than for the alternating-gradient ones discussed by Barden (2485 of 1954).

621.385.833 2187
Electron Optics of Cylindrical Systems having a Plane of Symmetry: Part 2—Aberrations—M. Laudet. (*J. Phys. Radium*, vol. 16, pp. 908–916; December, 1955.) Aberration formulas are derived following the method proposed by Durand (1756 of 1955). Part 1: 2391 of 1955.

621.389:535.833 2188
Role of Saturation in the Limitation of [star] Magnitudes attained by Classical Photography and by Use of the Electron Telescope—P. Vernier. (*Compt. Rend. Acad. Sci., Paris*, vol. 242, pp. 1006–1008; February 20, 1956.)

621.397+[621.37/.38:655 2189
Electronic Methods of Pictorial Reproduction—(*J. Brit. IRE*, vol. 16, March, 1956.) The text is given of the following papers, presented at a symposium held in January, 1956:—

Facsimile Transmission of Weather Charts and Other Material by Landline and Radio—J. A. B. Davidson (pp. 115–124).

Facsimile Communication—H. F. Woodman and P. H. J. Taylor, (pp. 129–144).

Electronic Engraving—S. W. Levine and A. B. Welch (pp. 145–152).

Tone Reproduction with Electronically Cut Stencils—R. Lant (pp. 153–157).

Discussion on the above papers is given on pp. 158–161.

621.398 2190
Telemetering—Electronic Data Transmission—A. A. McKenzie and H. A. Manoogian. (*Electronics*, vol. 29, pp. 153–180; April, 1956.) A survey under the headings: systems development; input transducers; signaling methods; commutating devices; output indicators. An extensive bibliography is included.

PROPAGATION OF WAVES

621.396.11 2191
Propagation of Electromagnetic Waves over the Spherical Earth across Boundaries separating Different Earth Media—K. Furutsu. (*J. Radio Res. Labs. Japan*, vol. 2, pp. 345–398; October, 1955.) Formulas for propagation over one or two boundaries agree with those for a homogeneous earth in special cases. A rigorous proof is given of a general formula for propagation over any number of boundaries, and methods of summation are discussed. For the case of a flat earth, see 541 of 1956 and 1848 of 1956.

621.396.11 2192
On the Multiple Scattering of Waves by an Irregular Medium—H. Hojo. (*J. Radio Res. Labs, Japan*, vol. 2, pp. 419–427; October, 1955.) Expressions are derived, assuming irregularities in one direction only, for the power distribution among the incident wave and singly, doubly, and triply scattered waves in the forward and backward directions. Near the critical frequency in an ionized medium, intense and diffuse reflections may occur.

621.396.11 2193
Forward Scatter of Radio Waves—(*Tech. News Bull. Nat. Bur. Stand.*, vol. 40, pp. 8–12; January, and pp. 24–29; February, 1956.) A short survey of NBS investigations of scatter propagation via the ionosphere and via the troposphere. For detailed papers see PROC. IRE, October, 1955.

621.396.11:551.510.535 2194
D-E Layer Electron Model reduced from Considerations of M.F. and H.F. Wave Absorption—T. Kobayashi. (*J. Radio Res. Labs, Japan*, vol. 2, pp. 399–412; October, 1955.) The model developed is used as a basis for calculating the field strength of waves transmitted over a 1200-km path by a single *E*-layer reflection. The results are shown graphically;

they differ markedly from corresponding curves derived from National Bureau of Standards charts published in 1948.

621.396.11:551.510.535 2195

The *Z* Propagation Hole in the Ionosphere—G. R. Ellis. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 43–54; February, 1956.) The transformation of the ordinary wave to an extraordinary wave under certain conditions gives rise to a "propagation hole" at the ordinary reflection level. Two methods of estimating the angular size of the "hole" are based on: a) measurement of *Z*-echo power; b) analysis of the distribution of angle of arrival of *Z* echoes. Measurements at Hobart at a frequency of 4.65 mc show the "hole" to be approximately circular, with an angular width to half-power points of $<0.84°$.

621.396.11:550.551.535 2196

Interaction of Ordinary and Extraordinary Waves in the Ionosphere and the Effect of Multiplication of Reflected Waves—N. G. Denisov. (*Zh. Eksp. Teor. Fiz.*, vol. 29, pp. 380–381; September, 1955.) The propagation of em waves in the ionosphere is considered for the case of a small angle between the direction of propagation and the magnetic field. Strong interaction between the ordinary and extraordinary waves takes place in the region where the refractive indices are nearly equal and, as a result, the ordinary wave becomes an extraordinary wave; this leads to multiple reflections. Application of the mathematical methods originally developed for inelastic collisions between atoms [*Helv. Phys. Acta*, vol. 5, pp. 369–422; 1932. (Stueckelberg)] leads to simple expressions for the reflection and transmission coefficients in terms of the ordinary and extradorinary refractive indices.

621.396.11:551.510.535 2197

The Interpretation of Measurements of Radio-Wave Interaction—L. G. H. Huxley. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 118–120; February, 1956.) The original formula relating the rate *R* at which electrons with energy *Q* lose energy in collisions with molecules is rewritten $R = B_n (Q - Q_o)$, where *B* is a function of *Q* and Q_o that is effectively constant when $(Q/Q_o - 1) \ll 1$, and *n* is the molecular concentration. Earlier results interpreted according to this formula give a range of *n* from $1.2 \times \times 10^{14}$ to 3×10^{14} per cm³ for heights 83–90 km. These heights are consistent with values determined from the phase of cross modulation.

621.396.11:551.510.535 2198

On the Observation of Ionospheric Self-Interaction—F. H. Hibberd. (*J. Atmos. Terr. Phys.*, vol. 8, pp. 120–122; February, 1956.) "Essential precautions are described that must be taken in experimental studies of ionospheric self-interaction to avoid misleading results caused by interference between various rays and by the bandwidth of the receiver."

621.396.11:621.396.65 2199

Physical Problems of Transmission over Radio Links—J. Fagot. (*Onde Élect.*, vol. 36, pp. 7–22; January, 1956.) Basic problems discussed include the development of an adequate signal/noise ratio, the effect of different propagation paths, and distortions due to frequency variation of propagation velocity over the transmitted band.

621.396.11:621.396.65 2200

The Radio Link under Conditions of Atmospheric Superrefraction—L. Sacco. (*Alta Frequenza*, vol. 24, pp. 436–469; December, 1955.) Superrefraction is commonly experienced in Italy at altitudes up to 500 m. An investigation is made of the geometrical features (reflection points, path differences and convergence coefficients) of the radio link under these conditions, and the value of the received field strength is calculated. Focusing and multiple-reflection effects are discussed.

621.396.11:621.396.674.3 2201

Transient Fields of a Vertical Dipole over a Homogeneous Curved Ground—Wait. (*See* 1959.)

621.396.11.029.55:523.746 2202

Oblique Incidence and Sunspots—R. Gea Sacasa. (*Rev. Telecomunicación, Madrid*, vol. 9, pp. 17–31; December, 1955. In Spanish and English.) Twenty-four charts of muf predictions by various national laboratories, based on the index of solar activity, are compared with predictions by Gea's method. For periods of high solar activity, in summer, the former predictions are too low and are in better agreement with the observations if displaced by six months. With the accuracy of prediction at present attained there is no experimental confirmation of the need to vary the predictions over the eleven-year cycle.

621.396.11.029.63 2203

The Influence of Atmospheric Turbulence on Ultra-short-Wave Radio Links across the Mediterranean—F. du Castel. (*Onde Élect.*, vol. 36, pp. 32–42; January, 1956.) Receiving stations working on wavelengths of 10, 22, and 70 cm were sited at various heights above and below the transmitter horizon. The signal-strength patterns caused by turbulence are described and probable values of the parameters involved are deduced. A mean " scale of turbulence" of 17 m is indicated, as against measured values in the range 20–130 m obtained in the USA by Gordon (1136 of 1955).

621.396.81:621.317.7 2204

New Apparatus for determining the Statistical Distribution of Random Electrical Processes—(See 2166.)

621.396.812.3.029.63 2205

Diversity Tests and Study of Focusing Effects on Long Line-of-Sight Radio Links—Rivet. (See 2210.)

621.396.812.3.029.64 2206

An Interpretation of the Statistics of Fading over Optical-Range Microwave Radio Paths—G. Kraus. (*Arch. Elekt. Übertragung*, vol. 10, pp. 19–25; January, 1956.) Fading due to tropospheric ducts is investigated. It is assumed that the signal at the receiver comprises a principal ray together with a number of subsidiary rays with random phases. The degree of stratification is characterized by the fraction *q* contributed by the subsidiary rays; the higher the value of *q* the greater the fluctuations of signal power about the mean value associated with the main ray. When $q = 1$, the depth of fading is at least 20 db for 1 per cent of the time. The results are in fairly good agreement with published observations of fading.

621.396.812.3.029.64 2207

Interference Fading caused by Frontal Discontinuities—P. Misme. (*Onde Élect.*, vol. 36, pp. 43–47; January, 1956.) Microwave fading is produced due to varying convergence of rays in the horizontal plane by sharply defined fronts characterized by stationary or rising temperature and falling humidity. Field-strength records showing the effect of a normal cold front and the passage of a storm are presented.

RECEPTION

621.376.232.2:621.396.62:621.396.822 2208

Effect of Electrical Fluctuations on a Detector (Wave-Envelope Method)—V. I. Tikhonov. (*Bull. Acad. Sci. U.R.S.S., Tech. Sci.* pp. 3–13; October, 1955. In Russian.) The detection of a signal which is modulated in amplitude and phase by voltage fluctuations in the IF amplifier is considered theoretically for a diode detector with a) a linear, b) a quadratic and, c) an exponential characteristic. Results are also given of an experimental investigation of the modification of the detected voltage probability distribution by the time constant of the detector RC circuit.

621.396.62:621.396.3:621.376.3 2209

Automatic Frequency Control in Frequency-Shift Radiotelegraphy—A. Niutta: G. Bronzi. (*Alta Frequenza*, vol. 24, pp. 519–522; December, 1955.) Discussion on 253 of January and author's reply.

621.396.812.3.029.63 2210

Diversity Tests and Study of Focusing Effects on Long Line-of-Sight Radio Links—P. Rivet. (*Onde Élect.*, vol. 36, pp. 23–31; January, 1956.) Signal-strength records ob-obtained with transmissions on $\lambda = 10$, 20 and 30 cm. over a distance of 230 km, are discussed.

Both space and frequency diversity were employed. Fading phenomena affecting large areas and local interferance fading effects were observed. The fine structure of the records cannot be satisfactorily explained without further theoretical and experimental investigation of focusing effects due to variations of atmospheric refraction.

621.396.82:621.397.62 2211

Parasitic Radiations from Television Receivers—Egidi and Maggiore. (See 2237.)

621.396.823:621.316.13 2212

Measurement of H. F. Interference by 380-kV Installations in Sweden—R. Bartenstein, A. Bergman, and L. Menstell. (*Elektrotech. Z., Edn A*, vol. 76, pp. 857–859; December 11, 1955.) Field-strength measurements in the frequency range 150 kc to about 15 mc and symmetrical line-interference-voltage measurements between 40 and 200 kc were made on two hv overhead lines of respective lengths 1.7 km and 207 km. The ratios of the interference voltages and of the field strengths are fairly well represented by the two formulas given. Results are presented graphically.

STATIONS AND COMMUNICATION SYSTEMS

621.376.2 2213

Tables of Bennett Functions for the Two-Frequency Modulation Product Problem for the Half-Wave Square-Law Rectifier—R. L. Sternberg, J. S. Shipman, and H. Kaufman (*Quart. J. Mech. Appl. Math.*, vol. 8, Part 4, pp. 457–467; December, 1955.) A companion set to the tables presented previously by Sternberg *et al.* (1779 of 1955).

621.376.55:621.397.5 2214

Multiple-Pulse-Time Modulation—H. J. Griese. (*Arch. Elekt. Ubertragung*, vol. 9, pp. 571–572; December, 1955.) Signal/noise ratio can be improved in a system in which a multiple pulse is phase modulated, because the width of the transmitted frequency band can be reduced. The phase shifts corresponding to the signal and simultaneous noise are integrated in an oscillating circuit whose output is rectified and smoothed to give a ptm single pulse which is demodulated in the usual way. Individual communication channels can be selected by frequency separation instead of or combined with time separation. The system is suitable for television sound.

621.39(54) 2215

Research in Electrical Communications in India Since 1865—S. P. Chakravarti. (*J. Instn Telecommun. Engrs, India*, vol. 2, pp. 7–18; December, 1955.) A short review with 157 references.

621.39.001.11 2216

Prolongation [extrapolation] of Signals with Limited Spectrum—J. A. Ville. (*Câbles et*

Transm., vol. 10, pp. 44–52; January, 1956.) Analysis based on operational methods indicates that extrapolation to instants corresponding to $t>0$ cannot be effected in practice, though theoretically possible.

621.39.001.11:519 2217

Theory of the Reliability of Operation of Systems comprising a Large Number of Elements—Sh. L. Bebiashvili. (*Bull. Acad. Sci. U.R.S.S., Tech. Sci.*, pp. 29–39; October, 1955. In Russian.) The reliability of a system such as a radio-relay system is calculated as a function of the probable lifetime of its elements (*e.g.*, tubes) and the method of operating the groups and elements of the system.

621.395.44+621.396.41 2218

Carrier-Frequency Installations for Cables, Overhead Lines and Radio Paths—H. Hannemann and H. Piechatzek. (*Nachrichtentech. Z.*, vol. 9, pp. 10–18; January, 1956.) Survey of methods used by a German company to standardize the equipment for carrier systems with different numbers of channels.

621.396.41:621.396.822.1 2219

Nonlinear Cross-Talk in Multicarrier Multichannel Systems—P. Güttinger. (*Arch. Elekt. Ubertragung*, vol. 9, pp. 573–577; December, 1955.) Analysis is presented for a system in which the individual carriers are phase modulated and all lie within one octave. By suitable design and operation of the mixer, interference products can be kept low.

621.396.61:621.317.76 2220

The Frequency-Band Recorder—Fleischer and Widdra. (See 2175.)

621.396.66:534.861 2221

Monitoring Sound Broadcast Programmes—T. Somerville. (*Wireless World*, vol. 62, pp. 228–231; May, 1956.) A brief review is given of the evolution of methods used by the BBC in assessing the quality of the af signal leaving the broadcasting studios. The selection of the loudspeaker and problems of studio and listening-room acoustics are briefly discussed.

621.396.8 2222

Theoretical and Experimental Study of Signal Interference in Radiocommunication Systems—J. Villepelet. (*Ann. Télécommun.*, vol. 10, pp. 264–276; December, 1955; vol. 11, pp. 8–24; January, 1956.) The protective effect of receiver selectivity in amplitude-keyed communication systems is analyzed in detail. The required relation between the dynamic selectivity curve for the system and the static selectivity curves of the receiver filters is examined. If an interfering pulse signal is filtered through a circuit tuned to its carrier at an early stage in the receiver, the width of the dynamic selectivity curve is reduced appreciably. Experimental confirmation of the results is reported. The properties of the dynamic selectivity curve are determined for AM, pbm and fm systems. It is possible to realize a system with a frequency occupancy as low as that of a system in which only static interference is present. To achieve this, the spectral curve of the transmitter must slope off at least as fast as the asymptotic slope of the receiver selectivity curve, which should be as great as possible.

SUBSIDIARY APPARATUS

621.316.7:621.387 2223

A Magnetic Thyratron Grid Control Circuit—J. H. Burnett. (PROC. IRE, vol. 44, pp. 529–532; April, 1956.) A circuit described previously (1983 of 1951) is modified by including an ac source and a rectifier in the secondary circuit of the grid transformer, changing the mode of operation to that of a reset magnetic amplifier. A 180° range of firing-voltage phase is available; the firing voltage builds up rapidly.

621.35 2224

Chemical Generation and Storage of Electricity—A. M. Adams. (*J. IEE*, vol. 2, pp. 7–13; January, 1956.) A review of recent work on fuel cells, whose operation is based on oxidation of carbon and hydrogen, and which give a low-voltage dc output; the particular importance of this type of cell is that it may give higher efficiency in the use of primary fuels.

621.352.32 2225

Some Basic Scientific Problems relating to [Leclanché-type] Manganese Dioxide Electrochemical Cells—J. Brenet. (*Rev. Gén. Élect.*, vol. 65, pp. 61–64; January, 1956.)

TELEVISION AND PHOTOTELEGRAPHY

621.397+[621.37/.38:655 2226

Electronic Methods of Pictorial Reproduction—(See 2189.)

621.397.24:621.372.55 2227

Phase-Correcting All-Pass Sections. Design of a Delay Equalizer for Television Transmission—H. Martin. (*Câbles et Transm.*, vol. 10, pp. 31–43; January, 1956.) Continuation of earlier work (1196 of 1954). Nomograms facilitating design calculations are presented. Calculations are made for equalizing a Paris-Soissons coaxial-pair circuit for television transmission.

621.397.26:621.396.65.029.63 2228

I.T.A. [Independent Television Authority] Midlands Relay—(*Wireless World*, vol. 62, pp. 223–226; May, 1956.) The Lichfield television station is connected by two radio links, using frequencies of 1.712 and 1.784 kmc and one return link on 2.216 kmc, to the Telephone House in Birmingham, 12 miles distant. The three channels share a single antenna at each terminal, horizontal polarization being used for the one direction of transmission and vertical for the other. Frequency modulation is used; the bandwidth occupied by the signal at peak deviation is about 14 mc. The equipment is briefly described.

621.397.5:621.376.55 2229

Multiple-Pulse-Time Modulation—Griese. (See 2214.)

621.397.5(46) 2230

National Plan for Television [in Spain]—J. Sánchez-Cordovès, I. Miró and E. Gavilán. (*Rev. Española Electrónica*, vol. 2, pp. 18–20; December, 1955.) The first stage of the plan envisages one transmitter at Madrid and another at Barcelona, with power initially 1.5 kw for image and 500 w for sound rising later to 15 kw for image and 5 kw for sound. Band-I frequencies of 55.25 mc for image and 60.75 mc for sound are to be used, with 625-line standard. Second- and third-stage plans are outlined.

621.397.6.001.4 2231

Circuit Technique for Generation of Electrical Test Patterns in Television—F. Pilz. (*Arch. Elekt. Übertragung*, vol. 9, pp. 547–558; December, 1955.) Discussion of methods for producing test patterns without using optical or electron-optical arrangements, *i.e.*, using only combinations of pulse signals. A unit developed at the Nuremberg Rundfunk-Technisches Institut is described.

621.397.61:535.623 2232

Correction Circuits for Color TV Transmitters—K. E. Mullenger and R. H. McMann, Jr. (*Electronics*, vol. 29, pp. 130–133; April, 1956.) Gamma and matrix correction circuits designed primarily for reproduction of color films but useful also for live color transmissions are described.

621.397.611.2:621.383.27 2233

Flying-Spot Scanning and the Transmission of Episcopic Pictures—H. Stier, P. Lindner, and E. Kosche. (*NachrTech.*, vol. 5, pp. 537–541; December, 1955.) Description of an electron-optical epidiascope used at the East-Berlin Television Center.

621.397.62 2234

Trends in 1955–56 TV Receivers [in the U.S.A.]—R. F. Scott. (*Radio-Electronics*, vol. 27, pp. 36–39; January, 1956.)

621.397.62 2235

Linear-Phase-Characteristic Television Receivers—A. van Weel. (*Onde Élect.*, vol. 36, pp. 48–56; January, 1956.) Phase distortion is practically eliminated by provision of a linear-phase IF amplifier, using simple circuitry. Selectivity is normal and disadvantages of phase compensations at video frequency are avoided.

621.397.62:621.372.632 2236

Simplified Band-III Convertor—O. E. Dzierzynski. (*Wireless World*, vol. 62, pp. 221–223; May, 1956.) Constructional details are given of a fixed-tuned antenna filter used in conjunction with the converter described earlier (1564 of May).

621.397.62:621.396.82 2237

Parasitic Radiations from Television Receivers—C. Egidi and F. Maggiore. (*Alta Frequenza*, vol. 24, pp. 470–498; December, 1955.) The sources of the parasitic radiations are reviewed; the local oscillator is recognized as being the most troublesome. A field-measuring set developed at Turin is described and relevant formulas for the field close to a radiator are presented. The observed radiation patterns are classified according to whether they are produced by chassis radiation only or not. Methods of reducing the parasitic radiations by suitable design are discussed.

621.397.71:621.325.5 2238

Xenon Arc Discharge Lamps—H. W. Cumming. (*Elect. Times*, vol. 129, pp. 83–85; January 19, 1956.) Details are given of several types; their illumination characteristics are highly suitable for television studios, intensity being controllable without color change.

621.397.7:628.972 2239

Television Studio Lighting—C. L. Faudell. (*Proc. IRE, Aust.*, vol. 17, p. 7–12; January, 1956.) Basic requirements are discussed and types of lamps and accessories available and the techniques of their use are indicated.

621.397.8:621.372.553 2240

Delay Equalization of a Television System—D. Bünemann. (*Arch. Elekt. Übertragung*, vol. 10, pp. 10–18; January, 1956.) The problem is discussed on the basis of a given frequency variation of the envelope delay. A graphical method is used to design an all-pass circuit by combining elementary all-pass units in such a way that the sum of the delays of the transmission system and the all-pass units has a minimum deviation from a constant mean value.

TUBES AND THERMIONICS

621.314.63.012 2241

A Chart for the Evaluation of Crystal Rectifier Constants—I. M. Templeton. (*Electronic Engng.*, vol. 28, p. 172; April, 1956.)

621.314.7 2242

Alloyed-Junction Transistor Development—J. J. Ebers. (*Bell Lab. Rec.*, vol. 34, pp. 8–12;

January, 1956.) A description is given of the method of producing an *n-p-n* or *p-n-p*-transistor by heating a pellet of the alloying material in contact with a wafer of the base material. The properties of this type of transistor are particularly suitable for switching applications.

621.314.7 2243
Factors affecting Reliability of Alloy Junction Transistors—A. J. Wahl and J. J. Kleimack. (PROC. IRE, vol. 44, pp. 494–502; April, 1956.) Report of an investigation of the effects on Ge *p-n-p*-transistors alloyed with In and *n-p-n*-transistors alloyed with As-doped Pb of exposure to oxygen, water vapor and other gases. Changes of the breakdown voltage, reverse current, and current gain are produced by oxygen and water vapor in opposite senses; no observable changes were produced by pure hydrogen, nitrogen or helium. The changes produced are reversible; the true characteristic can be restored by baking in vacuum at a suitable temperature. Very high stability of characteristics can be achieved if water vapor and oxygen are completely excluded in this way.

621.314.7:621.385 2244
Transistors versus Vacuum Tubes—D. G. Fink. (PROC. IRE, vol. 44, pp. 479–482; April, 1956.) The relative merits of these two types of device are compared with reference to particular applications; possible future developments are briefly indicated.

621.314.7+621.385]:621.375 2245
Comparison of Junction Transistor and Amplifier Valve—Ledig. (See 2001.)

621.383.001.4 2246
Simplified Determination of the Spectral Sensitivity of Photocathodes—C. Dufour. (*Ann. Radioélect.*, vol. 10, pp. 174–181; April, 1955.) Methods using monochromatic filters, particularly Fabry-Pérot and multidielectric types, are preferred. Laboratory equipment is described and results obtained with a Ag-Cs photocathode are reported.

621.383.27 2247
Transit-Time Spread in Electron Multipliers—T. D. S. Hamilton and G. T. Wright. (*J. Sci. Instrum.*, vol. 33, p. 36; January, 1956.) Brief report of experiments on commercial multistage photomultipliers, demonstrating how the modulation of hf signals detected by the multiplier decreases as the transit-time spread increases with the reduction of over-all operating voltage.

621.383.4:546.482.21 2248
Industrial CdS Photoresistors—P. Tarbes. (*Bull. Soc. Franç. Élect.*, vol. 6, pp. 73–82; January, 1956.) Properties and applications of commercially available CdS photoconductive cells are described.

621.383.4:546.817.221:621.396.822 2249
On the Specific Noise of Lead Sulfide Photodetectors—B. Wolfe. (*Rev. Sci. Instrum.*, vol. 27, pp. 60–61; January, 1956.) A brief note giving experimental results for several commercial PbS cells.

621.383.5:537.311.33 2250
GaAs Photo-element—R. Gremelmaier. (*Z. Naturf.*, vol. 10a, pp. 501–502; June, 1955.) According to Rittner's findings (1682 of 1955) GaAs, which has an energy gap of 1.38 ev, should be a highly efficient converter for solar energy. Some measurements on a *p-n*-junction made of polycrystalline material are briefly reported; higher efficiency is to be expected if carefully prepared single-crystal material is used.

621.385:621.318.57 2251
Keep-Alive Instabilities in a TR Switch—T. J. Bridges, P. O. Hawkins, and D. Walsh. (PROC. IRE, vol. 44, pp. 535–538; April, 1956.) The risk of the keep-alive discharge being accidentally extinguished as a result of glow-to-arc transitions is lessened by providing two keep-alive electrodes. See also 2252 below.

621.385:621.318.57.032.2 2252
Electrode Deterioration in "Keep-Alive" Discharges in Transmit-Receive Switches—D. Walsh, A. W. Bright, and T. J. Bridges. (*Brit, J. Appl. Phys.*, vol. 7, pp. 31–35; January, 1956.) An investigation of the behavior of electrode materials in a discharge in a medium containing water vapor is described. Most common stable metals and one semiconducting ceramic material were tried. Normal- and abnormal-glow conditions in tr switches are considered separately; the former lead to long life, while the latter give freedom from oscillations.

621.385.004 2253
Sampling of Vacuum Valves for Acceptance Inspection—M. D. Indjoudjian and J. Oswald. (*Câbles et Transm.*, vol. 10, pp. 65–72; January, 1956.) Sampling methods adopted by the French Post Office are described; these have led to saving in testing time without loss of efficiency. The mathematical basis of the technique is given.

621.385.004 2254
Utilization of Electronic Valves on [French] Long-Distance Underground Lines—P. Bassole and J. Eldin. (*Câbles et Transm.*, vol. 10, pp. 53–64; January, 1956.) A study of the methods used for accounting, inspection and analyzing replacements.

621.385.029.6:621.372.029.6 2255
Report of Advances in Microwave Theory and Techniques—1954—King. (See 1952.)

621.385.029.6 2256
A Graphical Method for investigating Travelling-Wave Valves—V. M. Lopukhin and Yu. D. Samorodov. (*Zh. Tekh. Fiz.*, vol. 25, pp. 1265–1275; July, 1955.) A method is proposed, by means of which it is possible to investigate dispersion equations of any algebraic degree. The method is applied to a study of traveling-wave and two-beam amplifiers for a wide range of values of the parameters. Analysis of the interaction between the field and the space charge yields solutions including one corresponding to a wave traveling in the negative direction. Curves of the phase velocity of the waves are plotted for various operating conditions.

621.385.029.6 2257
Using Traveling-Wave Tubes—R. E. White. (*Electronics*, vol. 29, pp. 144–145; April, 1956.) The suitability of traveling-wave valves as amplitude modulators, limiters, and mixers is indicated by reference to the tube characteristics.

621.385.029.6:538.566:537.56 2258
Growing Electric Space-Charge Waves and Haeff's Electron-Wave Tube—J. H. Piddington. (*Phys. Rev.*, vol. 101, pp. 14–16; January 1, 1956.) Analysis is presented indicating that the explanation of wave growth presented, *e.g.*, by Haeff (1825 of 1949) is not the true one; an alternative explanation based on trapping of electrons between potential troughs of a space-charge wave (2030 above) is considered more plausible.

621.385.029.6:621.376.23 2259
Microwave Detector—J. T. Mendel. (PROC. IRE, vol. 44, pp. 503–508; April, 1956.) The detector described is based on the "stop-band" velocity-discriminator properties of periodic magnetic focusing systems used in association with traveling-wave tubes [2546 of 1954 (Mendel *et al.*)]. The focusing system surrounds a drift tube following the wave-slowing circuit, and the beam electrons are sorted according to their rf velocity modulation, thus the collector current is a function of the microwave power. A numerical calculation for an arrangement comprising 20 magnetic lenses indicates that a power level of 0.36 mw corresponds to a velocity change of 0.76 per cent. Other possible applications of the device include a high-level mixer.

621.385.032.216.2:621.37 2260
Heater-Cathode Leakage—(*Tele-Tech and Electronic Ind.*, vol. 15, pp. 56–57, 110; January, 1956.) Various mechanisms by which current leakage takes place between an oxide cathode and its heater are described; observation and test methods are indicated. The effects on circuit performance are discussed with particular reference to a class-A amplifier, a multivibrator and a fm demodulator.

621.385.3 2261
Some New Types of High-Voltage Low-Current Vacuum Triodes—R. Feinberg and K. C. Burn. (*Electronic Engng.*, vol. 28, pp. 160–162; April, 1956.) Characteristics are given of tubes suitable for use as variable high-voltage resistors and in direct-voltage control and stabilization.

621.385.832 2262
Deposition and Removal of Electric Charges on Insulators by Secondary Emission: Part 1—M. Barbier. (*Ann. Radioélect.*, vol. 10, pp. 182–214; April, 1955.) A detailed theoretical study is presented of the processes involved in recording signals in the form of charges on insulator plates in cathode-ray tubes; the field distributions and secondary-electron paths are analyzed for some simple configurations. Possible accumulations of space charge are taken into account, and estimates are made of the amount of charge that can be deposited, the limits of resolution to be expected, the accumulation factor, and the beam intensity required for writing or reading.

621.385.832 2263
Permanent-Writing Cathode-Ray Recorder—L. N. Heynick, R. J. Wohl and D. H. Andrews. (*Electronics*, vol. 29, pp. 148–149; April, 1956.) A special cr tube of rectangular cross section has a linear array of styli projecting inwards and outwards through the end face. The beam is swept across the inner ends of the styli by a signal, and a suitable cooperating potential is applied to an external electrode which serves also to keep a moving sheet of recording paper pressed against the outer ends of the styli. Various applications of the device are mentioned.

621.385.832(083.7) 2264
IRE Standards on Electron Devices: Definitions of Terms Related to Storage Tubes, 1956—(PROC. IRE, vol. 44, pp. 521–522; April, 1956.) Standard 56 IRE 7. S1.

621.387 2265
Thyratron R. M. S. Current Ratings—(*Mullard Tech. Commun.*, vol. 2, pp. 152–156; December, 1955.) The increaseed ratings possible with delayed firing angles are shown.

MISCELLANEOUS

621.3.002.2 2266
New Developments in Automatic Production for Electronics—(*Elect. Mfg.*, vol. 55, pp. 126–131; April, 1955.) An illustrated report on

progress, including descriptions of a machine for inserting 24 different components into a printed circuit board, an automatically programmed punching machine, and an adaptation of the Tinkertoy technique.

621.37/.38].004.5/.6 2267

Increasing the Reliability of Electronic Equipment by the Use of Redundant Circuits—C. J. Creveling. (Proc. IRE, vol. 44, pp. 509–515; April, 1956.) Conditions are considered for equipment comprising a large number of elements, such as that of bombers. Equations are derived relating reliability to the number of circuit elements in the redundant and non-redundant cases; examples indicate the degree of improvement attainable. The improvement effected in this way is increased by arranging for regular maintenance; attention is drawn to the fact that the redundancy obscures the faulty condition.

621.37/.38].004.5/.6 2268

A Systems Approach to Electronic Reliability—W. F. Luebbert. (Proc. IRE, vol. 44, pp. 523–528; April, 1956.) A rationale for the development of reliable systems is presented in the form of a check-list.

621.39 2269

Radiotechnical Literature [in the U.S.S.R.] in 1956—V. Shipov, A. Smirnov, and P. Popov. (*Radio, Moscow*, No. 1, pp. 16–17; January, 1956.) Brief survey of publishing program of three leading publishing houses.

621.39(083.7) 2270

Symbols used in Electrical Communication Engineering—H. Meinke. (*Nachrichtentech. Z.*, vol. 9, pp. 39–46; January, 1956.) German draft standard DIN 1344, October, 1955, is presented and discussed.