MAY · 1953

Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

PREDICTING ELECTRONIC FAILURES

National Bureau of Standards

Among numerous methods under study for improving the reliability of electronic equipment is the experimental technique pictured above. A maintenance man simply plugs a portable failure-prediction test unit into the slightly-modified equipment to be checked and turns a multi-point selector switch; a red light flashes on to identify deteriorating stages before they affect over-all operation.

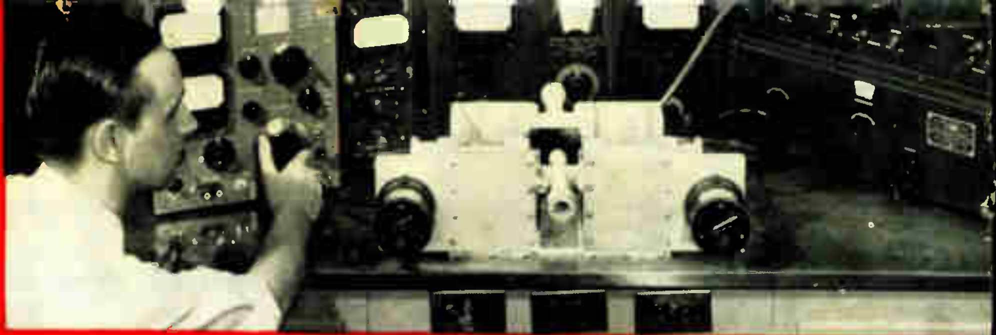
Volume 41 Number 5

IN THIS ISSUE

TABLE OF CONTENTS, INDICATED BY BLACK-AND-WHITE MARGIN, FOLLOWS PAGE 72A

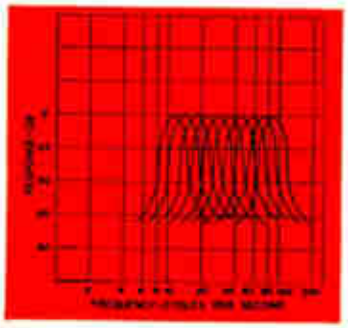
The IRE Standards on Modulation Systems: Definitions of Terms, appear in this issue.

The Institute of Radio Engineers

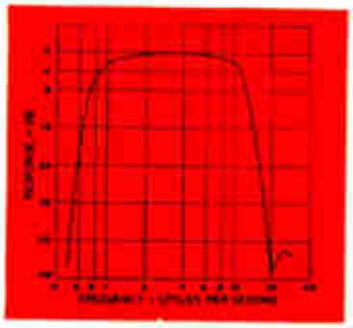


for SPECIALIZED FILTERS

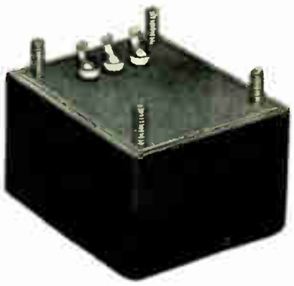
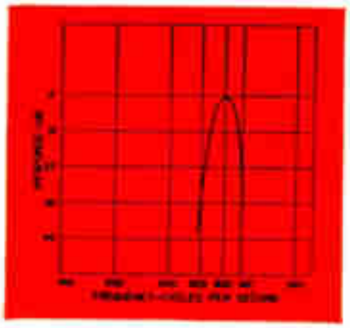
Decades of experience in the design and production of specialized filters have resulted in UTC being a first source for difficult filters. Fifteen years ago UTC was already the largest user of permalloy dust toroids in the world (exclusive of the telephone system). Present designs include a wide variety of core materials, structures, and winding methods to provide maximum performance in electrical requirements and stability. Illustrated below are a few of the thousands of special filter designs in present production.

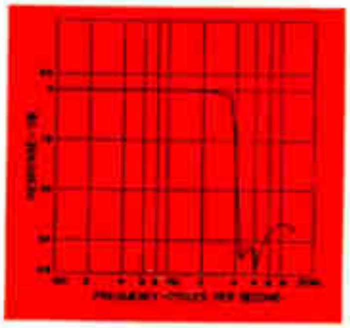
These low frequency band pass filters are held to 1 DB tolerance at the 3 DB crossover . . . 600 ohm . . . 4 filters per 7½" rack panel.



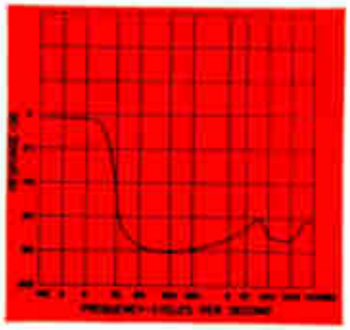

This ultra low frequency filter has a band pass range of one cycle to 10 cycles . . . 50,000 ohms . . . 700 cubic inches.

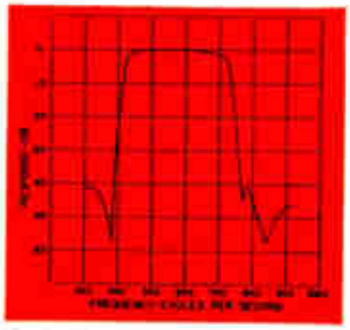
This 600 ohm miniaturized 1 KC band pass filter is housed in a case only 1" x 1¾" x 2½".




This 600 ohm miniaturized low pass filter is housed in a case only 1" x 1¾" x 2½".

This power line filter provides correct output voltages from sources of 50 to 400 cycles . . . noise attenuation is from 14 KC to 400 MC . . . 29 cubic inches.




This band pass filter is designed for sharp cut-off at both ends of the range . . . 10,000 ohms . . . case dimensions 1⅝" x 2½" x 3¼".

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PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

Archie W. Straiton

DIRECTOR, 1953–1954

Archie W. Straiton, Director of IRE Region 6, was born in Arlington, Texas, August 27, 1907. He received the B.S. degree in electrical engineering from the University of Texas in 1929. After a year with Bell Telephone Laboratories, he returned to the University to receive the M.A. and Ph.D. degrees in 1931 and 1939, respectively.

For eleven years Dr. Straiton taught as an assistant professor, associate professor, and professor at the Texas College of Arts and Industries, Kingsville, Texas, where he was in charge of electrical engineering courses. From 1941 to 1943 he was Director of Engineering, and Director of the Engineering Science and Management War Training Program there.

Since 1943 Dr. Straiton has been an associate professor and professor in the electrical engineering department of the University of Texas, and, since 1947, has been Director of the Electrical Engineering Research Laboratory, which has been engaged in radio wave propagation studies for various agencies of the Department of Defense.

Dr. Straiton became a Member of the Institute in 1947 and a Senior Member in 1949; since 1949 he has served on the IRE Wave Propagation Committee. For his work in radio wave propagation he received a Fellow Award at the 1953 National Convention.

Dr. Straiton is a past chairman of Commission II on Tropospheric Radio Wave Propagation of the International Radio Scientific Union, and attended the IX General Assembly of the Union in Zurich, Switzerland, in 1950 as representative of the National Science Foundation. He is a member of Sigma Xi, having served as the University of Texas Secretary and currently, President. He is also a member of the American Institute of Electrical Engineers, the American Society for Engineering Education, Tau Beta Pi, Eta Kappa Nu, and Theta Xi.

Engineering Dependence on Science

JOHN A. HUTCHESON

Engineering progress is built on the firm foundation of the knowledge derived from scientific research. Recognition of this fundamental truth, as well as its implications, are stressed in the following guest editorial by a Fellow of the Institute, who is Vice President and Director of Research of Westinghouse Research Laboratories.—*The Editor.*

The radio engineering profession is deeply indebted to pure science. The radio business owes its very existence to the work of men like Ampere, Oersted, Faraday, Maxwell, Hertz, Thomson, Fleming, and many, many others. The development of the radio industry from the early days of "wireless" to the tremendous business that it is today has been made possible by the work of countless engineers applying the principles first stated by workers in pure science.

As the development of the industry progressed, the need for additional knowledge of a fundamental nature became increasingly apparent. For example, present-day vacuum tubes operate according to the same fundamental principles that governed the operation of the very first tube. However, the vastly improved performance achieved by modern tubes has been made possible through the application of new knowledge, much of which has been obtained since the invention of the first tubes. This new knowledge which made possible vacuum-tube performance was made available through research that was inspired by the realization of the need for new information.

In the past, much of the research upon which our industry is based was the work of scientists in foreign countries. Currently, however, the research programs which are furnishing the new information upon which the continued growth of our industry depends are being carried out in the laboratories of colleges and universities, of industry, and in government operated laboratories. The success of these programs and the programs of the future depends upon a continued supply of trained scientists, a supply which at this time is all too small. Therefore, in order to assure the continued growth of the radio industry, it behooves each of us to do all we can to assure proper utilization of the trained manpower now available, and to encourage and promote the training of scientists and engineers to the end that an adequate supply is available for the future.

The Discovery of Science*

P. W. BRIDGMAN†

Summary—**The scientist has set himself the problem of finding out as much as he can about the world around him and understanding this as well as he can. In accepting this problem, value judgments are involved which, in turn, involve an emotional component. Among the emotional components are: acceptance for its own sake of the discipline of the fact; subject to this discipline, a sense of complete freedom; the challenge of difficulty; the sense of adventure in penetrating into the unknown. In the passion of the scientist for discovering facts irrespective of whether they are pleasant or unpleasant lies the source of the undoubted incompatibility between scientists and non-scientists. One of the tasks of society is to find how to live with this incompatibility.**

THE HUMAN RACE is at present in the process of finding out something new about itself. The process of discovery has been extending over the last few hundred years and is not yet complete. The discovery has two aspects. In the first place, we have discovered that the human animal is capable of engaging in scientific activity, in making scientific discoveries, and making scientific theories to help understand these discoveries. That is, the human animal is capable of "doing" science. In the second place, we have discovered that this scientific activity has emotional involvements and, in particular, is capable of giving a special satisfaction to those who engage in it.

This is not the first time that the human race has discovered that it has unsuspected characteristics, characteristics produced only incidentally in the course of evolution and by no means necessary for survival. A very early discovery must have been that words can be used for pleasure only, apart from their use in communicating factual information, and the art of poetry has been with us ever since. Perhaps a somewhat later discovery was that the human animal responds emotionally to music. We can imagine that music, at least in its instrumental modes, was discovered after poetry, because the fashioning of musical instruments demands a fairly highly developed craftsmanship. The reason that science has appeared on the human scene so much later than poetry or music is that much more elaborate preparation is necessary for it. Science is not possible without a background of an enormous amount of factual information or without the development of technical tools of research.

The human race has by now pretty well found what values to attach to poetry and music and has found how to live with its poets and musicians. I have no doubt, however, that among the early cavemen there were some who groused at the late hours of a too popular bard and perhaps even subjected him to mayhem. The human race, on the other hand, has not yet found how to adapt itself to all the implications in its newly-discovered capacity for science, although adaptation is urgent because the impact of science on the life of everyone is incomparably more formidable than was ever the impact of poetry or music.

There are pleasant and unpleasant aspects of this impact. The pleasant aspects may be typified by the multifarious new amenities of daily life, such as the telephone or the automobile or modern medicine, all of which are an outgrowth of previous scientific activity and would not have been possible without it. The unpleasant aspects may perhaps be typified by the atomic bomb. The impact of the bomb has been so tremendous that at present the unpleasant results of science seem to be uppermost in the minds of many people, and there appears to be an increasing mood of animosity toward science and scientists. Furthermore, the reaction against science is overshooting its mark and is resulting in a general mood of anti-intellectualism and intellectual defeatism. Many people despair of the ability of the human intelligence to grapple with its problems and are grasping at any straw that offers. It is therefore of the greatest importance that the true nature and implications of science be understood. There are implications not only for the more material circumstances of our existence, but also for those more intangible aspects which may be conveniently lumped under considerations of value. It is with these latter that we shall concern ourselves.

What are the drives that make the scientist go, and what are the values implied in these drives? We must also ask what are the values which society must prize if science is to flourish, for the attitude of society does determine to a large extent whether science will flourish or whether it will be driven underground and eventually extirpated. It is to be said in the first place that the question of values is not very consciously present to the scientist himself as he engages in his scientific activities, nor is he conscious of his drives. These have to be inferred by a disinterested spectator from the sidelines, or even by the scientist himself as he thinks over his day in his armchair at night. It seems to me that what the scientist does is perhaps best characterized by saying that he has set himself a problem, and that his values are more or less incidentally involved in its solution—it is the problem that bulks important for the scientist. The problem of the scientist is, as I see it, to find out as much as he can of the world about him, and having found what the facts are, to reduce them to the best sort of understanding that he can.

This characterization of the scientist is obviously not inclusive. It would be difficult to make an inclusive characterization, but, with the exercise of a little good will, I think it will prove adequate for our purposes. There is no compulsion on any individual or even on society as a whole to accept the problem of the scientist, and in fact a large part of the human race feels no compulsion and recognizes no intrinsic interest in the solution of the problem. That the scientist does accept this problem involves what may be called, I suppose, his system of values.

Given then the problem of the scientist—to find out as much as he can and to understand as much as he can about the world around him—it is evident that the fact occupies a position of fundamental importance. The facts must be correctly reported in all pertinent detail, and no theory can endure which is at variance with any single fact. The discipline of the fact, to which the scientist must submit himself, is a discipline of complete rigor, incomparably more rigorous than the discipline of any medieval monk. Acceptance of this discipline demands the suppression of some favorite human frailties, particularly those dealing with enhancement of the ego. We have to mistrust the reliability of our senses, and be willing to submit any report to constant cross-checking and verification. We must similarly submit the results of our reasoning processes to continual questioning and checking. All this is to say nothing of the crasser vices of intellectual conduct, such as wishful thinking or the seeking of personal advantage.

There is, it seems to me, an emotional component in the acceptance by the scientist of the discipline of the fact, just as there was doubtless an emotional component in the acceptance by the medieval monk of his discipline. The emotional factor is not to be dismissed with the observation that acceptance of the discipline of the fact is a necessary precondition to the solution of his problem by the scientist.

Many people, perhaps most, do not regard the discipline of the fact as something to be accepted, but rather as something to be avoided as far as possible. Such people would be willing to live in a world of their own construction, and I can see no reason why they should not if they can find how to get away with it. Acceptance of the fact is, however, so deeply ingrained in the scientist that he finds something unclean in the thought of living in a world of his own construction.

The scientist, having made himself the complete slave of the fact, has at the same stroke won complete freedom in every other respect, for no prohibition or inhibition can be recognized which militates in any way against the recognition of any fact. In designing his experiment or formulating his theory no holds are barred for the scientist—his limitations are set by his own capacities and such limitations are not felt as limitations of freedom. The consciousness of freedom is, I think, always present to the scientist and is itself one of his values.

To the scientist there is no argument about this matter—he feels an inner freedom of which no external social compulsion can deprive him, although he always recognizes that social compulsion can drive him underground. This sense of freedom is, of course, not peculiar to the scientist, but all of us have it to a greater or less degree. I hope I have not misinterpreted the anthropologist in my opinion that the concern of the individual with his own freedom appears comparatively late in social evolution. The scientist's especial feeling of the necessity of

* Decimal classification: 507.2. Copyright 1952 *Harvard Bulletin*, Inc., and reprinted from the *Harvard Alumni Bulletin* of November 8, 1952.

† Physics Department, Harvard University, Cambridge 38, Mass.

freedom and his exultation in it would appear to be consistent with his late appearance on the evolutionary scene. Incidentally, it seems to me that the ideology of communism, in ignoring and suppressing the sense of freedom of the individual, is attempting to live in a world constructed contrary to nature, a course which it seems to me must eventually react unfavorably on the effectiveness of at least its scientific activities.

The problem which the scientist has set himself is not an easy one, and in the challenge offered by the overcoming of difficulty is to be found one of the values of science. Of course every discipline has its own difficulties and the challenge of difficulty can always be found. We can all appreciate the reaction of the mountain climber who has to climb the mountain simply because the mountain is there. In science, however, the challenge of difficulty is particularly insistent and in fact the limits which most hamper the scientist are limits set by difficulties which he has not yet found how to surmount. The scope of his problem is enormous and is taking him further afield than he had ever suspected when he embarked. He is finding, for example, that an essential part of his problem is to understand the nature of his intellectual tools, an understanding which the race has not yet achieved, in spite of long concern with it. The highest capacities of the mind are demanded.

Added to the challenge of difficulty is the challenge of the unknown, which has always appealed to men in all conditions. The scientist gets no less a thrill out of penetrating into a previously unknown domain of knowledge than does the explorer in finding landscapes previously unseen by man. This element of adventure is always present to some degree in every activity of the scientist, because he is almost entirely occupied with finding facts or inventing theories which are new—the scientist does not go over ground already known. Just as the average human being can take vicarious pleasure in the discoveries of the explorer, so he can take a vicarious pleasure in the discoveries of the scientist—once discovered, they are a part of the heritage of the race which cannot be taken from it.

All these drives and values thus far discussed play their part in making the scientist go, but I think the most powerful is still to be mentioned. The scientist would not concern himself with a problem to which the fact is so basic if he did not feel a certain congeniality in the presence of the fact. The scientist likes facts, particularly new ones, and he wants to find what they are. Or, differently expressed, one of the most important drives of the scientist is curiosity. He wants to know just for the sake of knowing, and he also wants to understand just for the sake of understanding. These two drives can hardly be separated, for usually knowledge is a prerequisite to understanding. Although the driving curiosity of the scientist is often recognized as one of his most important characteristics, I think that nevertheless the full implications are not usually appreciated.

Many people have claimed that there is a basic incompatibility between scientist and non-scientist, and there have been many attempts to diagnose and formulate the differences. Thus, it is said that science is materialistic and incompatible with the things of the spirit or with religion, or that science is hard, unemotional, literal, and incapable of poetic feeling. I think, that most of these formulations are beside the mark and can be refuted. Much scientific work is highly imaginative, and the scientist often stands in wonder before the beauties he finds or constructs. At the same time I think that in rejecting the unsound formulations we may minimize what seems to me the incontrovertible, very real difference between scientist and non-scientist. This has got to be squarely faced and analyzed as best we can and not shrugged aside by arguments which to me smack too much of appeasement. I think there can be no question of the existence of an incompatibility or of the reality of the dislike which many people feel toward the scientist. It seems to me that the incompatibility and dislike are tied in pretty closely with the insatiable and universal curiosity of the scientist of which we were speaking a moment ago. The scientist *wants* to know and to discover, even if what he finds is disagreeable.

I think the great majority of people, on the other hand, would rather not know the disagreeable things, a frame of mind expressed in the popular saying "what you don't know can't hurt you." With a frame of mind like this it is only natural to feel resentment against anyone who forces one to think of disagreeable matters and therefore presumably to do something about them. This is understandable enough and is perhaps unavoidable, but I think the resentment overshoots its mark and the scientist is often charged with creating the disagreeable situation which he uncovers.

The atomic bomb is a dramatic example; it is very common for the scientist to be charged with the whole responsibility for the situation created by the bomb, and I have no doubt that most people wish that the scientist never had discovered that the bomb is possible. Many even think that the scientist should not have allowed nature to be so constructed that the bomb is possible. To which, I imagine, the scientist would retort that it is not the knowledge that atom bombs can be produced that is making the trouble, but rather it is that men are willing to produce them. If one does not want atom bombs, the way to deal with the situation is not to make them. I believe the scientist would add that it is not he who is making them but society—at least it is society that is footing the bill, including two billion dollars for the first one.

Even in a case as extreme as this, I think the scientist would maintain that knowledge in and of itself is wholly good, and that there should be and are methods of dealing with misuses of knowledge by the ruffian or the bully other than by suppressing the knowledge. It looks to the scientist as though a large part of society in its present temper is willing to deal with the situation by suppressing the scientist because it is easier to suppress the scientist than to suppress the ruffian and the bully.

To the scientist this fear of knowledge displayed by a part of society appears wholly contemptible, and acceptance of the proposal advocated by many that a moratorium should be declared on science to be wholly unthinkable. Here is what seems to me the fundamental incompatibility that has got to be squarely faced, an incompatibility perhaps more deep seated than the incompatibilities between the common man and the poet or the musician with which the race has long since found how to deal. Scientists are human beings, and the potentiality for science is not the least precious of the great heritages of the human race. What do you propose to do with your heritage?

A Review of VHF Ionospheric Propagation*

M. G. MORGAN†, SENIOR MEMBER, IRE

This tutorial paper was prepared by the Subcommittee on Ionospheric Propagation of the IRE Wave Propagation Committee. Members of the Subcommittee are: M. G. Morgan, Chairman, R. Bateman, K. L. Bowles, R. A. Helliwell, C. W. McLeish (Canada), J. H. Meek (Canada).—*The Editor*

Summary—**Although the very high frequencies (vhf, 30–300 mc) are allocated for utilization almost entirely upon the premise that propagation will be tropospheric, there are very definite ionospheric effects with which one must reckon. These are: (1) regular F_2 ionization, (2) sporadic E ionization, (3) scattering from regular ionization, (4) auroral ionization, (5) meteoric ionization. The first of these is predictable with reasonable accuracy and provides phenomenally good communication over great distances and with extremely low power. The second is far less predictable, but it also can provide good communication with extremely low power. The third is mainly a newly discovered phenomenon; although it offers intriguing possibilities for regular communication, it presents at the same time large problems in achieving that end. The fourth and fifth are so far only of nuisance value to the communicator. However, they are phenomena of great interest to the geophysicist, as indeed are the other phenomena as well.**

In this paper, the nature of vhf propagation arising from these sources of ionization is discussed. The maximum distance between path terminals for single-hop propagation is, of course, about 2,400 km for reflection from E region effects and about 4,000 km for F region effects. Multiple hops are rarely significant except from the regular F_2 ionization at frequencies only slightly above 30 mc.

A considerable amount of information is included on the nature of the ionization as deduced from the propagation effects. The work of individual investigators is extensively referenced.

Introduction

THIS REPORT concerns itself with those phenomena which enable a vhf signal to be propagated, via the ionosphere, between two points on the surface of the earth. The following means are known whereby such propagation can take place under proper conditions:

I. Regular F_2 ionization
II. Sporadic E ionization
III. Scattering from the regular ionization
IV. Auroral ionization
V. Meteoric ionization

In the following sections, these five phenomena are discussed in the order given above. References and notes are numbered consecutively.

I. Regular F_2 Ionization

Frequencies up to about 50 mc when transmitted from the earth's surface in a nearly tangent direction, may be returned to earth by the regular F_2 ionization under not infrequent conditions. Diurnal peak of ionization, low latitude points of reflection, and years of maximum sunspot activity are most favorable for such transmission.[1] Signal strengths tend to be very high but fading due to interference of high and low rays is often great.

During the latter part of the winter of the peak sunspot year 1937, television signals on 40–45 mc originating in London and Berlin were received on Long Island, N.Y. and in Indiana.[2] They were also observed in 1938 and 1939.[3] At the next sunspot maximum, there was much transatlantic activity among radio amateurs on these frequencies. One United States-United Kingdom contact was made on 50 mc in 1946, and in 1947 there were literally hundreds. Daily observations showed that the North Atlantic maximum usable frequency (muf) exceeded 45 mc for several hours almost daily in October and November, 1946, and to a lesser extent in February and March, 1947.[4] The muf for this path exceeded 50 mc almost every day from Oct. 25, 1947 to Dec. 1, 1947 and exceeded 52 mc on some November days. In

* Decimal classification: R112. Original manuscript received by the Institute, April 7, 1952; revised manuscript received February 12, 1953. At the Eighth General Assembly of the International Scientific Radio Union (URSI) held in Stockholm in 1948, a committee on long-distance, vhf, ionospheric propagation was appointed. At the Ninth Assembly, held in Zurich in 1950, the USA National Committee of URSI accepted an assignment from the previously appointed committee to prepare a report on the subject, and the work was carried out as described in the editor's note above. The document was approved on April 1, 1952, by Wave Propagation Committee for recommendation to the Board of Editors for publication in PROCEEDINGS and for transmission to URSI. USA National Committee of URSI approved the paper on April 23, 1952 for transmission to the Tenth General Assembly of URSI held in Sydney, Australia, August, 1952.

The paper consists largely of information available in the United States and Canada. Information from other sources, some recent, appears in the paper: M. G. Morgan, "Progress in ionospheric research during 1952," Dartmouth College Ionospheric Research *Tech. Rpt. No. 5* (contract Nonr 438-02); Dec. 1, 1952. Submitted Feb. 17, 1953 for publication in PGAP TRANSACTIONS OF IRE.

After completion, the existence of an earlier paper with similar title, based upon work done in the U.S.A., came to the author's attention. This paper was prepared by Mrs. M. L. Phillips (now at Mass. Inst. of Tech.) under an assignment accepted by the U.S.A. Delegation of the International Radio Consultive Committee (CCIR), a technical advisory organization serving the International Telecommunications Union (ITU) in its frequency allocation work. The report as adopted by the U.S.A. Delegation of CCIR has not been published. It appears in a very much shortened form, and combined with reports from other sources on the subject, as Report No. 7 in the *Documents of the Sixth Plenary Assembly of CCIR* held in Geneva in 1951.

† Director of Research, Thayer School of Engineering, Dartmouth College, Hanover, N. H.

[1] J. Maire, "Reception of transatlantic signals of frequencies near 30 mc," *Ann. Radioélect.*, (in French), vol. 6, pp. 197–204; July, 1951.

[2] H. O. Peterson and D. R. Goddard, "Field strength observations of transatlantic signals, 40–45 mc," PROC. I.R.E., vol. 25, p. 1291; October, 1937.

[3] D. R. Goddard, "Observations on sky-wave transmission on frequencies above 40 mc," PROC. I.R.E., vol. 27, p. 12; January, 1939; "Transatlantic reception of London television signals," PROC. I.R.E., vol. 27, p. 692; November, 1939.

[4] Due to E. P. Tilton, American Radio Relay League; corroborated by O. P. Ferrell, "Unscheduled Report RASO-2," Mar. 15, 1949, *CQ* Magazine, N.Y. The prediction of these results was described by K. A. Norton, "Sunspots and vhf radio transmission," *QST*, vol. 31, p. 13; December, 1947. For a discussion of co-channel interference of vhf stations due to regular F_2 propagation, see E. W. Allen, Jr., "Vhf and uhf signal ranges as limited by noise and co-channel interference," PROC. I.R.E., vol. 35, p. 128; February, 1947.

the autumn of 1948 and spring of 1949, the muf over the North Atlantic path exceeded 45 mc fairly regularly. Transcontinental communication on 50 mc occurred in the United States in the autumns of 1947, 1948, 1949.

There have been occasions upon which vhf signals were received over long paths and lower frequency vhf signals were not propagated. For example, in the early months of 1947 and 1948 television signals from the United Kingdom were received in the Capetown, S. A., area. In several instances, the 45 mc video was received when the 41.5 mc audio could not be heard. During the early months of 1947, United States FM stations operating in the 45 mc band were heard in England with daily regularity. On numerous occasions, there was a blank band of transmission over the path of 4–10 mc width directly below 45 mc. Whereas the services occupying this region below 45 mc were, in general, using less power than the FM broadcast stations, this seems to be an inadequate explanation and is not applicable to the United Kingdom-South Africa case cited. The effect is difficult to explain on an absorption basis because hf signals continue to be propagated over the path at such times.[5]

Sudden ionosphere disturbance (SID) is now a well-known phenomenon in which it is believed that a burst of ultra violet radiation associated with a solar flare produces increased ionization in the lower ionosphere. This results in virtually complete absorption of hf signals. However, vhf signals are not so seriously affected.[1] (See also Part III.)

Round-the-world echoes frequently constitute a serious limitation to the operation of long-distance hf circuits. This effect is eliminated by working close to the muf.[1]

In middle latitudes, F_2 muf's are, in general, depressed during the immediately following periods of geomagnetic disturbance[6] although exceptions have been noted.[7] It has been found that an increase in F_2 muf's occurs in equatorial regions at these times.[8] This behavior has been related to the theory of magnetic storms, and it is shown that the equatorial enhancement occurs simultaneously with the maximum depression in temperate regions.[9] A large number of radio observations at 50 mc have confirmed this behavior in the western hemisphere. No longitude effect has been established from these data. They indicate that equatorial enhancement may continue for 12 to 18 hours after the cessation of a magnetic storm.[10] Recently, an apparently successful attempt has been made to plot F_2 critical frequencies for the North American continent during disturbances. Even with a paucity of data, surprisingly well-defined contours are obtained. The concentration of established ionosphere stations in northeastern United States and eastern Canada reveals the justification of drawing smooth contours through relatively thin data. Positive and negative deviations from the monthly mean averages as great as 60 per cent are noted. The maps reveal that two disturbances had rather well-defined centers of 2,000—3,000-km diameter which moved across the continent. Two other storms had no centers but exhibited a marked dependence upon geomagnetic latitude. During these latter disturbances, which occurred in winter, the critical frequency was enhanced in regions below and reduced in regions above approximately 50°N geomagnetic latitude. Some additional evidence indicates that the reversal takes place at a lower latitude in the summer.[11]

It must be pointed out that radio-propagation manifestations of geomagnetic disturbances are by no means as simple as the reader might infer from the above remarks which deal only with those features particularly affecting vhf propagation.[9,12]

II. Sporadic-E (E_s) Ionization

Frequencies up to about 150 mcs, when transmitted from the earth's surface in a nearly tangent direction, may be returned to earth by sporadic-E ionization. The occurrence of this ionization has long been studied but its nature and origin remain largely enigmatic.[13] The nature of sporadic-E ionization is ill defined. Even the word "sporadic" is controversial as to whether it applies to time, place, or both.

A number of physical descriptions of sporadic-E ionization have been suggested to account for the observed facts. Some forms of sporadic-E have been explained on the basis of a thin layer[14] but one of the most

[5] Due to E. P. Tilton, American Radio Relay League. The Capetown observations were made by H. and C. Rieder (ZS1P and ZS1T) and the North Atlantic observations by D. Heightman (G6DH).

[6] E. V. Appleton and L. J. Ingram, "Magnetic storms and upper atmospheric ionization," *Nature*, vol. 136, p. 548; Oct. 5, 1935. S. S. Kirby, T. R. Gilliland, E. B. Judson, and N. Smith, "The ionosphere, sunspots, and magnetic storms," *Phys. Rev.*, vol. 48, p. 849; Nov. 15, 1935.

[7] T. R. Gilliland, S. S. Kirby, N. Smith, and S. E. Reymer, "Characteristics of the ionosphere at Washington, D. C., January to May, 1937," Proc. I.R.E., vol. 25, p. 1177; September, 1937.

The last good transcontinental communication in the United States on 50 mc by F_2 transmission occurred on Nov. 20, 1949. There was a pronounced ionospheric disturbance on the previous night. The transcontinental "opening" was about equal to the best experienced in 1947 under undisturbed conditions. The disturbance ended around 2200 EST on Nov. 19, and F_2 long-distance communication on 50 mc existed from 0800 to 1400 EST on Nov. 20. The transcontinental communication commenced within a few minutes after the end of several hours of communication with South America. (Due to E. P. Tilton, American Radio Relay League.)

[8] L. V. Berkner and S. L. Seaton, "Systematic ionospheric changes associated with geomagnetic activity," *Terr. Mag.*, vol. 45, p. 419; December, 1940.

[9] E. V. Appleton and W. R. Piggott, "World morphology of ionospheric storms," *Nature*, vol. 165, p. 130; Jan. 28, 1950.

[10] O. P. Ferrell, "Enhanced transequatorial propagation following geomagnetic storms," *Nature*, vol. 167, p. 811; May 19, 1951. E. P. Tilton "Long-distance communication over north-south paths on 50 mc," International Council of Scientific Unions, Mixed Commission on the Ionosphere, *Proc. of the Second Meeting*, Brussels; September, 1950.

[11] R. S. Lawrence, "Continental maps for four ionosphere disturbances," *Trans. IRE*, PGAP-3, pp. 214–216; August, 1952; abstracted in Proc. I.R.E., vol. 40, p. 747, (paper no. 63); June, 1952.

[12] D. F. Martyn, "A theory of magnetic storms and auroras," 9th General Assembly of URSI, Zurich; September, 1950. Published by Cornell University as *Tech. Rept. No. 3*; Dec. 15, 1950 (contract DA36-039-sc-218). See also, *Nature*, vol. 167, p. 92; Jan. 20, 1951.

[13] For a discussion, see for example: E. K. Smith, Jr. "The sporadic E region of the ionosphere and its effect upon television," *Tech. Rept. No. 7*, Cornell Univ. (contract W36-039-sc-44518), Oct. 1, 1951.

[14] B. H. Briggs, "The determination of the collisional frequency of electrons in the ionosphere from observations of the reflection coefficient of the abnormal E layer," *Jour. Atmos. Phys.*, vol. 1, p. 345; 1951.

plausible pictures at present postulates large clouds with a "blobby" substructure. Propagation is described by the summation of the contributions of closely spaced scattering sources. As the frequency is increased for a given blob size, there is a gradual changeover from back to forward scatter. This explains the fact that transparency increases with frequency. Computations indicate that weak echoes may be accounted for by blobs differing from their surroundings by a maximum of 10 per cent in electron density. The corresponding value for strong echoes is 30 per cent. One point arising from the theory is that the ratio of muf to critical-frequency is the same as for a thin layer—i.e., the secant of the angle of incidence.[15]

Sporadic-E echoes are observed at vertical incidence up to about 25 mc corresponding to a maximum of 135 mc for oblique paths. This is borne out by the observations of long-distance television reception in the United States.[16] Of 441 low-band reports (54–88 mc), 28 are under 800 km and 413 are over 800 km. On the other hand, of 15 high-band reports (174–216 mc), 14 are less than 800 km.[17] Evidence is presented to support the thesis that low-band propagation up to 800 km is generally tropospheric, and beyond is always ionospheric. The predominance of reports is for channel 4 (66–72 mc) because of the greater number of transmitters on this channel. The equivalent, vertical critical-frequency for 1600 km (the most frequently reported distance) is 13.5 mc. Thus an observer near a channel 4 station, surrounded by other channel-4 stations at sufficient distances to prevent interference from tropospheric propagation, may experience interference from the other stations on some occasions. The fact that vertical incidence sporadic-E echoes are not obtained at any individual ionosphere station in excess of this frequency for more than 2 per cent of any month of the year, should not be taken as conclusive evidence that television interference from this cause will never exceed 2 per cent of any month inasmuch as the simultaneity of sporadic-E ionization around the ring of applicable reflection points surrounding the television observer cannot be assumed.[16,18]

Vertical-incidence measurements made in the United Kingdom confirm the foregoing paragraph. It is concluded that co-channel interference of services due to sporadic E, is very significant in the range 30–55 mc, of much less importance from 55–88 mc, and of negligible significance above 100 mc.[19]

In Japan, measurements on 65 mc over a 1,000-km path, made simultaneously with vertical-incidence measurements at the mid-point of the path, show propagation to occur more often than predicted by application of the secant law to the vertical incidence E_s data.[20]

Considerable interest has been shown recently in the determination of the size and motion of the sporadic-E ionization. At least four distinct methods have been used: (1) Analysis of a large number of long-distance oblique path transmissions obtained continuously,[21] (2) observations of backscatter,[22] (3) triangulation from three or more points on the ground,[23] (4) comparison of sweep-frequency, vertical incidence records made at independent observatories spaced about 500 km apart.[19,24] The clouds are found to be of the order of 100 km in size and to move with velocities of a few hundred km/hr or about 100 mc.

Reliable vertical incidence data for studies of geographical and temporal dependence of sporadic E are difficult to obtain because of large differences in the performance characteristics of the various ionosphere stations, particularly with regard to radiated power as a function of frequency. However, the following general characteristics have been discerned: In equatorial and middle latitudes, E_s is most prevalent during summer days,[19] while in polar regions E_s is principally a nighttime winter phenomenon. In the northern hemisphere, the changeover takes place at 55–60°N geomagnetic latitude where both are observed giving semi-diurnal and semi-annual maxima. In the northern hemisphere, the occurrence of E_s increases from middle latitudes northward to a maximum near the normal position of the auroral zone.[25] Less E_s is observed at more northerly points.[26] In general, a corresponding behavior has been found for the southern hemisphere except that in comparing the hemispheres an annual component is also discerned which has its maximum at the

[15] H. G. Booker, E.E. Dept. Colloquium, Cornell Univ., February, 1951, unpublished. See also footnote 31.

[16] E. K. Smith, Jr., *loc. cit.* footnote 13.

[17] It appears certain, so far, that no high-band ionospherically supported television reception has been positively identified in the United States.

[18] E. W. Allen, Jr., "Vhf and uhf signal ranges as limited by noise and co-channel interference," Proc. I.R.E., vol. 35, p. 128; February, 1947.

[19] T. W. Bennington, "Propagation of vhf via sporadic E," *Wireless World*, vol. 58, pp. 5–9; January, 1952.

[20] T. Kono, M. Mayumida, and Y. Abe, "Long-distance propagation of vhf," *Proc. Semi-Annual Meeting of Central Radio Wave Observatory* (Japan); April, 1951. (Document no. 36, 10th General Assembly of URSI, Sydney, Australia; August, 1952.)

[21] N. C. Gerson, "Continental sporadic E activity," *Trans. Amer. Geophys. Union*, vol. 32, p. 26; February, 1951; "Abnormal E region ionization," *Canadian Journal of Physics*, vol. 29, p. 251; May, 1951. P. Revirieux and P. Lejay, *Comp. Rend.*, vol. 227, p. 79; 1948. Also E. K. Smith, Jr., *loc. cit.*, footnote 13.

[22] Using a rotating, directional antenna, range and azimuth of sporadic-E clouds have been presented in PPI display using backscatter on 14 mc; O. G. Villard, Jr. and A. M. Peterson, "Instantaneous prediction of radio transmission paths," *QST*, vol. 36, p. 11; March, 1952; also, O. G. Villard, Jr., A. M. Peterson, and L. A. Manning, "A method for studying sporadic E clouds at a distance," Proc. I.R.E. (Correspondence), vol. 40, pp. 992–994; August, 1952; also, J. T. de Bettencourt, "Instantaneous prediction of ionospheric transmission circuits by the communication zone indicator (COZI)," Trans. I.R.E., PGAP-3, pp. 202–209; August, 1952. Backscatter from sporadic ionization at high northern latitudes has been observed on sweep-frequency ionospheric recorders; see: J. H. Meek, footnote 25 following, and "Oblique reflexion of radio waves by way of a triangular path," *Nature*, vol. 169, p. 327; Feb. 23, 1952.

[23] K. Weeks, "The measurement of wind velocities in the ionosphere," *Proc. Conf. on Ionospheric Physics*, Penn. State College; July, 1950.

[24] S. Matsushita, *Jour. Geomag. and Geoelec.* (Japan), vol. 1, pp. 35–40; October, 1949.

[25] J. H. Meek, "Sporadic ionization at high latitudes," *Proc. Conf. on Ionospheric Research*, Penn. State College; June, 1949.

[26] J. H. Meek, "Characteristics of E region sporadic ionization in Canada," (Canadian Paper) 8th General Assembly of IUGG, Oslo, Norway; Aug. 19–28, 1948.

same latitudes as the semi-annual components and occurs in local summer.[27] The types of E_s observed at vertical incidence may be summarized as follows:

At low latitudes (18°S geomagnetic)[28]

Ess—Sequential type; extends out of the foE decreasing in virtual height as it increases to higher frequencies, later becoming spotty and disappearing. The start of the sequence is most prevalent just after sunrise and in the late afternoon before sunset, sometimes lasting until midnight. It is mainly a summer phenomenon.

Esc—These traces occur with no change in virtual height during their lifetime. They are characterized by a high reflection coefficient and consequently show several multiple reflections. No correlation is found with time of day, season, or sunspot cycle. This type is also seen in polar regions and has been identified with magnetic storms[29] in polar regions although apparently not at low latitudes.

At high latitudes (60°N geomagnetic)[25]

S-type—A solid trace extending up to a maximum frequency less than 6 mc. Presumably a layer of small horizontal extent, or only partially reflecting, inasmuch as the *F*-region trace is not appreciably affected. This type accounts for 55–85 per cent of E_s observations at these latitudes.

L-type—Usually appears as an extension of the S-type to frequencies greater than 6 mc. It takes the form of a sudden increase in frequency for some seconds or minutes and then subsides to near its former frequency (usually between 2 and 4 mc). The peak occurrence is in summer rather than in winter. This type accounts for 5–20 per cent of E_s observations.

M-type—This is a solid trace but with one or more multiple reflections present. Although it often extends to frequencies greater than 6 mc, its top reflection frequency is more stable than the L-type. It blankets much of the *F* trace above it, indicating that it has a larger horizontal extent. Its presence, however, is not usually concurrent at several ionospheric stations. It may be identified with the E_{sc} described at low latitudes. Accounts for 3–10 per cent of E_s observations.

P-type—These are patchy traces extending continuously for only a few tenths of a mc. They occur at 50–55°N geomagnetic latitudes more frequently than further north. Accounts for 4 per cent of E_s observations.

R-type—This type shows retardation near the top-reflection frequencies usually with both ordinary and extraordinary ray traces present. Its peak occurrence is just before sunrise and just after sunset. Its critical frequencies increase and decrease rapidly with a rate of the order of 1 mc per minute. This type is most prevalent in, and north of, the auroral zone. When seen south of the auroral zone, it is less variable in frequency.[30] Accounts for 8–13 per cent of E_s observations.

E_{2s} consists of traces appearing at random heights above the regular E_s trace. It usually first appears at about twice *E* level, then descends to *E* level and rises again. This may be explained by a horizontally moving cloud at *E*-layer height. If the explanation is correct, E_{2s} may not be a particularly appropriate designation.

At middle latitudes

Both high and low latitude types of E_s are observed on occasion.

III. Scattering from Regular Ionization

The discovery of a new type of weak but regular vhf propagation by means of the ionosphere has been reported.[31] Initial experiments on a frequency of 50 mc using a power of about 20 kw and extreme antenna directivity at both transmitter and receiver, reveal the uninterrupted presence of observable signal over a test path of 1,245 km, irrespective of season, time of day, or geomagnetic disturbance, though showing dependence in intensity on these factors.

Preliminary considerations suggest that the mechanism responsible for this type of propagation may be scattering due to ever-present turbulence in terms of parameters describing inhomogeneities in the *E* region. Measured signal intensities during periods of hf radio fadeouts associated with solar flares never show any reduction in signal. On the contrary there is usually an enhancement of signal intensity accompanied by a simultaneous weakening of the background noise. This result suggests strongly that the signals are returned from a part of the *E* region near or just below the absorption region for hf radio waves.

This form of propagation appears to be observable over ranges out to about 2,000 km but is likely to be masked by other kinds of propagation at distances less than 1,000 km.

As possible corroborating material, it is pointed out that scatter-type echoes at constant radar range, have been obtained with high power (500 kw) on 33 mc.

[27] M. L. Phillips, "Seasonal variations of sporadic *E* ionization," Joint IRE-URSI Meeting, Washington, D.C.; Oct. 7–9, 1948.

[28] R. W. E. McNicol and G. de V. Gipps, "Characteristics of the E_s region at Brisbane," *Jour. Geophys. Res.*, vol. 56, p. 17; March, 1951.

[29] H. W. Wells, "Polar radio disturbances during magnetic bays," *Terr. Mag.*, vol. 52, p. 315; September, 1947.

[30] An E_s-type designated N1 has been observed at Kiruna, Sweden (65° N. geomagnetic). This type appears to be correlated with polar blackouts observed at Kiruna and first described by H. W. Wells, "Polar radio disturbance during magnetic bays," Special Report, *Dept. of Terr. Mag.*, Carnegie Institution of Washington; December, 1942. (See reference 17). The polar blackout has not been correlated with solar flares as have the more commonly known radio fadeouts. The N1-type of E_s is most nearly like the R-type described herein. The times of appearance and the occurrence of fast temporal changes are the same. However, the N1-type quite often shows a gradual increase in virtual height with frequency whereas the R-type always shows the features of a definite critical frequency retardation. See Rune Lindquist, "A survey of recent ionospheric measurements at the Ionospheric and Radio Wave Propagation Observatory at Kiruna," *Arkiv För Geofysik* (Stockholm), vol. 1; 1951; and "Polar blackouts recorded at the Kiruna Observatory," *Trans. of Chalmers Univ. of Tech.* (Gothenburg), no. 103; 1951, and *Acta. Polyt.* (Stockholm), no. 85, 25 pp.; 1951.

[31] D. K. Bailey, R. Bateman, L. V. Berkner, H. G. Booker, G. F. Montgomery, E. M. Purcell, W. W. Salisbury, J. R. Wiesner, "A new kind of radio propagation at very high frequencies observable over long distances," *Phys. Rev.;* Apr. 15, 1952.

These are always at 82–85 km range, often several km thick. They grow slowly to a maximum intensity of not more than twice noise and then fade away. The average duration is about half an hour and they may occur several times during a night (or day).[32]

Radio amateur communication in the range 50–54 mc in equatorial regions has been observed in the post-sunset period in equinoctial months. Distances between path terminals range from 2,000–7,500 km. The signals are often garbled and weak and appear to result from *F*-region scattering.[33] It has been noted that *F*-region scatter is observed at these times on the Huancayo vertical-incidence records.[34]

IV. Auroral Ionization

Frequencies up to 150 mc are known to be returned from regions associated with auroral displays under sufficient ionizing activity. Communication over distances up to 2,000 km between path terminals can be obtained in this way when directional antennas are pointed toward these displays.[35]

It seems well established that the aurora is essentially an *E* region phenomenon of intense ionization, existing in bands of a nominal radius of about 23° about the two geomagnetic poles.[36] These so-called "auroral zones" manifest: (1) light radiation normally called "the aurora,"[37] (2) radio wave hf absorption, (3) large variations in the earth's magnetic field, and (4) at least some correlation with *E* sporadic.[38] Visible displays are occasionally seen at much lower latitudes.[39] Low latitude displays often occur above the *E* region. The most probable cause of aurora is the impact of solar particles.[40] Low-frequency measurements have been interpreted as indicating that the auroral ionization spreads to lower latitudes, but not to higher latitudes, at times of increased auroral activity.[41] Microwave emissions from the aurora have been reported and possible causes discussed.[42]

Audio modulation is imparted to signals reflected by the aurora especially when the distance between path terminals is not great. It has been suggested that this is due to rapid changes in the number density of the impinging particles[43] and to Doppler effect.[44] The power spectrum of the fading is found to have frequency components of roughly equal strength from zero to a cut-off frequency of 100—200 cps at a carrier frequency of 50 mc. Amplitude modulation, though badly garbled, is sometimes intelligible on a 50-mc carrier. Received signals usually retain in large degree the transmitter polarization whether horizontal or vertical. In many cases the probability distribution of signal amplitude is close to the Rayleigh distribution. "Openings" probably occur at all hours but have been observed at times varying from shortly after noon to approximately dawn, and over periods of a few minutes to many hours.[35] It has been shown that auroral propagation manifests itself as horizontal bars on television screens when receivers are tuned to distant "low-band" stations (82–88 mc).[45]

Pulse echoes have been received (at the transmitter site) on frequencies of 33,[46] 40,[47] 46,[48] 50,[49] 72,[50] and 100 mc.[48,51] Results, particularly at the higher frequencies, indicate that the echoes arise from within, or close to the lower border of those types of aurora which show visible structure. The echoes occur most frequently at slant ranges of 500–1,000 km, and it has been suggested that the minimum range is due to enhanced reflection

[32] Due to D. W. R. McKinley, National Research Council, Ottawa, Can. See D. W. R. McKinley and P. M. Millman, "Long duration echoes from aurora, meteors, and ionospheric back-scatter," *Can. Jour. Phys.*, vol. 31, pp. 171–181; February, 1953.

[33] O. P. Ferrell, "Vhf propagation in the equatorial region," presented at joint IRE-URSI meeting, National Bureau of Standards, Washington, D. C. April 16–18, 1951.

[34] Due to H. W. Wells, Carnegie Institution of Washington.

[35] R. K. Moore, "A vhf propagation phenomenon associated with aurora," *Jour. Geophys. Res.*, vol. 56, p. 97; March, 1951. K. L. Bowles, "The fading rate of ionospheric reflections from the aurora borealis at 50 mc," *Jour. Geophys. Res.*, vol. 57, pp. 191–196; June, 1952.

[36] E. H. Vestine, "The geographic incidence of aurora and magnetic disturbance, northern hemisphere," *Terr. Mag.*, vol. 49, p. 77; June 1944. F. W. G. White and M. Geddes, "The antarctic zone of maximum auroral frequency," *Terr. Mag.*, vol. 44, p. 367; 1939.

[37] There has been some evidence of two, closely-spaced zones of maximum visible activity. J. H. Meek, "Distribution of aurora in central Canada," Defence Research Board, Ottawa, Can., Radio Physics Laboratory, *Report No. 6;* Aug. 16, 1950.

[38] R. W. Knecht, "Relationships between aurorae and sporadic *E* echoes," Trans. I.R.E., PGAP-3, p. 213; August, 1952 (abstract); also abstracted in Proc. I.R.E., vol. 40, p. 747 (paper no. 64); June, 1952.

[39] C. W. Gartlein, "Southern extent of aurora borealis in North America," *Jour. Geophys. Res.*, vol. 56, p. 85; March, 1951.

[40] D. F. Martyn, "A theory of magnetic storms and auroras," 9th General Assembly of URSI, Zurich; September, 1950, published by Cornell Univ., as *Tech. Rept. No. 3;* Dec. 15, 1950 (contract DA36-039-sc-218); *Nature*, vol. 167, p. 92; Jan. 20, 1951. See also C. W. Gartlein, "Protons and the aurora," *Phys. Rev.*, vol. 81, p. 463; Feb. 1, 1951; *Trans. Amer. Geophys. Union*, vol. 32, p. 120; February, 1951; *Nature*, vol. 167, p. 277; Feb. 17, 1951. D. H. Menzel, "Physics of the aurora borealis," (Harvard Univ.) presented to Amer. Astro. Soc., to be published.

[41] J. H. Meek, "Reception of 2 mc loran in central Canada," Defence Research Board, Ottawa, Can., Radio Physics Laboratory *Report No. 9;* July, 1951.

[42] P. A. Forsyth, W. Petrie, and B. W. Currie, "Auroral radiation in the 3000 mc region," *Nature*, vol. 146, p. 453; Sept. 10, 1949. A. E. Covington, "Microwave sky noise," *Jour. Geophys. Res.*, vol. 55, p. 33; March, 1950. P. A. Forsyth, W. Petrie, and B. W. Currie, "On the origin of ten centimeter radiation from the polar aurora," *Canadian Jour. Res.*, vol. A28, p. 324; May, 1950.

Noise attributed to an ionospheric origin has also been observed on 50 mc. In this case, it has not been correlated with the aurora or any other phenomenon. See H. V. Cottony, "Radio noise of ionospheric origin," *Science*, vol. 111, no. 2872, p. 41; January, 1950.

[43] N. C. Gerson, "Radio observations of the aurora on 19 November 19, 1949," *Nature*, vol. 167, p. 804; May 19, 1951.

[44] R. K. Moore, "Theory of radio scattering from the aurora," Trans. I.R.E., PGAP-3, pp. 217–219; August, 1952. Abstracted in Proc. I.R.E., vol. 40, p. 747 (paper no. 65); June, 1952.

[45] R. Thayer, "Auroral effects on television," Proc. I.R.E., vol. 41, p. 161 (Correspondence); January, 1953.

[46] Due to D. W. R. McKinley, National Research Council, Ottawa, Can., and P. M. Millman, Dominion Observatory, Ottawa, Canada. Also see footnote 32.

[47] L. Harang and W. Stoffregan, "Scattered reflections of radio waves from a height of more than 1000 km," *Nature*, vol. 142, p. 832; 1938.

[48] A. C. B. Lovell, J. A. Clegg, and C. D. Ellyett, *Nature*, "Radio echoes from the aurora borealis," vol. 160, p. 372; Sept. 13, 1947.

[49] Due to K. L. Bowles (and others), Cornell Univ.

[50] A. Aspinall and G. S. Hawkins, "Radio echo reflections from the aurora borealis," *Jour. Brit. Astro. Assn.*, vol. 60, p. 130; April 1950.

[51] P. A. Forsyth, W. Petrie, F. Vawter, and B. W. Currie, "Radar reflexions from aurora," *Nature*, vol. 165, p. 561; Apr. 8, 1950.

when propagation is normal to the earth's magnetic field along which the ionizing particles are thought to travel.[44] At 100 mc, however, no clear evidence has been found that the echoes are enhanced for propagation normal to the earth's magnetic field. The echoes are much more variable at the higher frequencies and for any particular display can be obtained at successively higher frequencies as the intensity of the display increases. The form of the echoes and their distribution in the horizontal plane is consistent with the assumption that they arise from a large number of small reflecting centers distributed along the lower border of the display.[52]

Very recently, measurements of frequency shift of an unmodulated, 50 mc, cw signal reflected from the aurora have been reported. Shifts ranging from zero up to about 200 cps have been observed. Certain obvious conclusions about the motion of the reflecting region can be reached if this is interpreted as a simple Doppler effect.[53,54]

V. Meteoric Ionization

Pulse echoes have been received (at the transmitter site) over a frequency range of 3–200 mc from meteor trails.[55] The lower frequencies give more and longer-duration echoes and it has been shown that the duration depends upon the square of the wavelength.[56] Durations are typically from 1–100 seconds in the vicinity of 30 mc. It has been concluded from theoretical considerations that laminar flow of a meteoric ionization path may increase duration as much as four times.[57]

Strong vhf propagation via meteorically ionized regions of reflection is usually very ephemeral. For a discussion of such echoes at 50 mc, reference should be made to footnote 31 of Part III.[58]

During the spectacular Giacobini-Zinner shower of 1946, meteoric activity was so great that sustained vhf communication over long paths became possible. In the United States, many voice amplitude-modulated contacts were made on 50 mc over paths up to 2,000 km long. Reception was a continuous succession of short bursts of varying intensity, coming so close together that the signal was intelligible. The phenomenon lasted for more than three hours during which time there was only slight change in the character of the signals heard.[59] From measurements made during this shower, it is concluded that only a few watts per km^2 are required to produce an ionospheric layer.[60]

Meteoric ionization has long been under consideration as a cause of sporadic E ionization. The present evidence is that meteoric activity accounts for a separate and distinct phenomenon,[61] although the material in footnote 31 may possibly begin to strengthen again earlier tenets. In the experiments described, there is a pronounced evening minimum of signal intensity and a tendency to show a fairly steady rise during the night. A maximum is reached in the forenoon, followed by a second and greater maximum at noon. It is suggested that the diurnal variation of the solar ultraviolet radiation produces the midday maximum but that the secondary maximum in the forenoon may be accounted for by meteoric activity. Meteoric activity is greatest at about 0600 local time and a minimum at about 1,800. Superposition of this effect upon the regular solar illumination effect accounts reasonably well for the observed diurnal behavior.

Statistical analysis of eleven thousand meteor velocities, observed over a period of fifteen months by vhf pulse-echo technique, has led to the conclusion that all, or nearly all, meteors down to the eighth visual magnitude are members of the solar system.[62] Measurements of meteor decelerations have led to the conclusion that either the formula for the scattering of radio waves from a meteor trail needs revision or that the density of air at 100 km height is lower by a factor of three or more than that in the NACA tables.[63] Measurements of the polarization of meteor echoes indicate that ionization densities may be greater than are usually assumed and that existing theories[64] are not compatible with this result.[65] Radio pulse measurements have shown the effective mean-trail length to be 25–30 km.[66,67]

[52] This paragraph, except as otherwise referenced, is due to B. W. Currie, Univ. of Saskatchewan, and P. A. Forsyth, Radio Physics Laboratory, Defence Research Board, Ottawa, Can.; also corroborated by K. L. Bowles (and others), Cornell Univ.

[53] R. B. Dyce, School of Electrical Engineering, Cornell Univ. From Progress Report No. 1, "Studies on propagation in the ionosphere; Sept. 30, 1952."

[54] Much useful information on the aurora is also contained in: L. Harang, "The Aurorae," The International Astrophysics Series, John Wiley and Sons, Inc., New York, N. Y.; 1951.

[55] For example, see the following and their bibliographies: D. W. R. McKinley and P. M. Millman, "Determination of the elements of meteor paths from radar observations," *Canadian Jour. Res.*, vol. 27A, p. 53; May, 1949; J. A. Clegg, and G. S. Hawkins, "A radio echo apparatus for the delineation of meteor radiants," *Phil. Mag.*, vol. 42, no. 328, p. 504; 1951.

[56] V. C. Pineo and T. N. Gautier, "The wave-frequency dependence of the duration of radar-type echoes from meteor trails," *Science*, vol. 114, no. 2966, p. 460; Nov. 2, 1951.

[57] T. N. Gautier, "The effect of laminar flow on the duration of meteor echoes," presented at joint IRE-URSI meeting, Cornell Univ.; Oct. 8–10, 1951.

[58] See also E. W. Allen, Jr., "Vhf and uhf signal ranges as limited by noise and co-channel interference," Proc. I.R.E., vol. 35, p. 128; February, 1947.

[59] E. P. Tilton, "The world above 50 mc," *QST*, vol. 30, p. 43; December, 1946.

[60] J. A. Pierce, "Ionization by meteoric bombardment," *Phys. Rev.*, vol. 71, pp. 88–92; Jan. 15, 1947. In a report of Oct. 30, 1946 (Cruft Laboratory, Harvard Univ., Contract N5-ori-76, T.O. 1) the author discusses the significance of his conclusions in terms of maintaining a strong ionospheric layer by artificial means.

[61] V. C. Pineo, "A comparison of meteor activity with occurrence of sporadic E reflections," *Science*, vol. 112, no. 2898, p. 50; July 14, 1950.

[62] D. W. R. McKinley, "Meteor velocities determined by radio observations," *Astrophys. Jour.*, vol. 113, p. 225; March, 1951.

[63] D. W. R. McKinley, "Deceleration and ionizing efficiency of radar meteors," *Jour. Appl. Phys.*, vol. 22, p. 202; February, 1951.

[64] J. Feinstein, "The interpretation of radar echoes from meteor trails," *Jour. Geophys. Res.*, vol. 56, p. 37; March, 1951.

[65] L. A. Manning and M. E. Van Valkenberg, "The polarization characteristics of meteoric echoes," *loc. cit.* footnote 57.

[66] L. A. Manning, O. G. Villard, Jr., and A. M. Peterson, "The length of ionized meteor trails," presented at joint IRE-URSI meeting, National Bureau of Standards, Washington, D. C.; Apr. 21–24, 1952.

[67] Many interesting comments on meteoric ionization are also contained in the "Proceedings of the Colloquium on Mesospheric Physics," Air Force Cambridge Research Center, Cambridge, Mass.; July, 1951.

Tropospheric Propagation: A Selected Guide to the Literature*

The following paper was prepared by the IRE Committee on Wave Propagation and its Subcommittee on the Theory and Application of Tropospheric Propagation. It is published with the approval of the above committee.—*The Editor*

Summary—**A subcommittee of the IRE Committee on Wave Propagation selects references designed to direct the non specialist through the mazes of an extensive literature to information on certain specific aspects of the propagation of short radio waves through the troposphere.**

I. Foreword

During 1949 the IRE Wave Propagation Committee undertook to prepare this guide in the hope that this new activity might prove useful to readers of the Proceedings. Tropospheric propagation, or propagation of radio waves on frequencies higher than those normally reflected by the ionosphere (about 30 mc and higher), has an important bearing on the utilization of the vast spectral range so stimulated by the war and post-war development of uses for radio waves. The literature of the subject is extensive, but widely scattered through the technical journals of the last twenty years. On the theoretical side, the mathematical complexities of the subject have tended to repel all but the most mathematically inclined radio engineers. During the last few years, fortunately, several books on the subject have appeared, but it still appeared to the committee that the nonspecialist in the subject might be helped by a selective guide designed to assist him through the maze of literature to some useful answers to certain propagation problems. The guide is intended to be a highly selective list of references for the radio engineer who is a nonspecialist in tropospheric propagation, but who desires either specific quantitative estimates for certain propagation questions, or background discussion to give him a grasp of the meaning of the numerical answers once they are located and calculated. Following a brief account of books and general discussions of tropospheric propagation which are recommended, the selected references comprising the guide are arranged under the topical headings into which a subject is often broken down. It must be emphasized that limitations of space rather than considered critical judgment governed the rather arbitrary selection of papers included. In particular no slight to the pioneers is intended by omission of papers prior to 1940. It is hoped that the papers and books listed will adequately mention these.

II. Caution

The specialists on the committee in the field believe that the nonspecialist reader should arm himself with a healthy note of skepticism while delving through the literature. Nonspecialists almost invariably believe that it should be possible to answer the questions they ask of the propagation "expert" in a positive manner. The experts believe that most of the questions asked can be answered only in a general way. The engineering profession is warned by the committee. therefore, that only a skeleton of working knowledge is available, and that many things in the literature must frequently be taken with a large grain of salt.

III. Books

During the years 1947 through 1951 four books on tropospheric propagation have appeared, which taken together greatly simplify the search for specific information, for the specialist and nonspecialist alike. The first step in approaching the literature for help in solution of almost any tropospheric propagation problem may frequently be consultation of the table of contents or indexes of these books in order to discover the sections bearing most directly on the problem at hand. Important papers which have appeared since the writing of these books are usually mentioned in the pages devoted to tropospheric propagation in the annual progress reviews published each April in the Proceedings. If, in addition, the reader consults the propagation abstracts reprinted each month in the Proceedings from the *Wireless Engineer*, then practically no important paper bearing on a particular tropospheric propagation problem can escape detection. For convenient reference in later paragraphs these four books will be designated by the letters *A*, *B*, *C*, *D*.

A. "Propagation of Short Radio Waves," D. E. Kerr, editor, volume 13 of the Radiation Laboratory Series, McGraw-Hill Book Co. Inc., New York, N. Y.; 1951. The authors of this volume aimed to summarize the state of knowledge in the field at the end of the second world war. The emphasis on MIT Radiation Laboratory contributions is only the natural consequence of the authors' more intimate acquaintance with the propagation work of their own laboratory.

B. "Terrestrial Radio Waves; Theory of Propagation," H. Bremmer, Elsevier Publishing Co., Amsterdam (also New York, London); 1949. This volume will doubtless remain for a long time the starting point for study of the theoretical aspects of tropospheric propagation. incorporating as it does the classic papers of Van

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der Pol and Bremmer in the 1930's on propagation over a curved earth.

C. "Radio Wave Propagation," Technical Report of the Committee on Propagation of the N.D.R.C., C. R. Burrows (chairman), S. S. Attwood (editor), Academic Press Inc.; New York, N. Y., 1949. This volume makes readily available a wide variety of papers, theoretical experimental, and expository in nature, which were declassified at the end of the second world war. At the end of this volume (pages 531–548) is a general bibliography of reports on Tropospheric Propagation. These listings include most of the unpublished wartime reports up to the summer of 1945. A later bibliography which includes references up to the middle of 1938 has been published as CRPL-2-3 by the Central Radio Propagation Laboratory, National Bureau of Standards.

D. "Meteorological Factors in Radio Wave Propagation," Physical Society, London; 1947. This volume conveniently collects together papers summarizing many British wartime contributions to the subject.

A book antedating those above by about 20 years is "Propagation of Radio Waves," P. O. Pedersen, G. E. C. Gad, Copenhagen; 1927. It is still consulted by those who have access to a copy.

IV. General Discussions

While it is the prime purpose of this guide to direct the nonspecialist directly to a place in the literature where quantitative answers may be found to certain specific questions, it would be unwise to start making computations directly without acquiring some general background to illuminate the calculations being made.

During the war, the NDRC Committee on Propagation produced at the request of the armed services a semi-popular account of the vagaries of tropospheric propagation for the information of those concerned with the operation of radar sets.[1] Another longer general account of the subject produced then gives a somewhat more technical treatment, but with a minimum of mathematics.[2] A shorter birdseye view may be found in the first paper of reference (*D*).[3] Readers interested in a general account of super-refraction from the viewpoint of radar application may prefer to read Booker's paper.[4]

V. Diffraction

Many papers on the propagation of radio waves over a homogeneous spherical earth under uniform atmospheric conditions have been published in the last thirty years, mostly based on pioneering work of Watson.[5]

Watson's great contribution was the transformation of a very slowly convergent series of spherical harmonics to a residue series, which expressed the field at distances beyond the optical horizon as a sum of independent modes, each exponentially attenuated but at different rates. For sufficiently large distances only the first mode makes an appreciable contribution to the field.

It is not necessary to list all the papers based on Watson's method since a very complete account of the present state of diffraction theory is now available in volumes (*B*) and (*C*).

Bremmer's book (*B*) includes a historical survey (pages 6 to 8), with references to most of the earlier papers, and a theoretical development of various formulas for the field strengths under different operating conditions. A complete set of numerical formulas adapted to computation under most conditions likely to occur in practice appears on pages 104 to 110. These may be used without reference to the theory in the preceding pages, and include numerical values of all the relevant parameters. Cases considered include both vertical and horizontal polarization; dielectric earth and conducting earth (sea water); transmitter and receiver both on the earth's surface and one or both elevated; distances within the optical horizon and beyond. Examples of computations are shown, and some illustrative curves of the field variation with distance and with height (pages 113 to 120).

Reference (*C*) gives a somewhat more technical account of diffraction theory and includes reports on experimental work. Volume III of this work covers much the same ground as Bremmer's book, and includes formulas for field computations, with curves of various factors which may be used in determining numerical values. In particular, chapter 5, on the calculation of radio gain, should be consulted.

Some may prefer the compact methods of calculation suggested in chapter 2 reference *A*, for the cases covered. For example, the bilinear profile of the horizontally stratified atmosphere treated in section 2.17 constitutes the most exhaustive presentation yet published of the problem of the nonstandard troposphere.

VI. Refraction

A. Rays

It should be stressed that those computational formulas discussed in section V are based on the assumption that the earth is a homogeneous sphere of constant dielectric constant and conductivity, while the atmosphere is also homogeneous with constant refractive index. As the frequencies in use were extended to higher and higher values it became apparent that more attention would have to be paid to variations in the refractive index of the atmosphere. The known fact that the refractive index decreases linearly with height in a well-mixed atmosphere, and thus causes a downward bending of rays towards the earth, was first compensated for by assuming an effective radius of the earth, somewhat larger than its actual radius. This had the effect of straightening the rays, relative to the modified earth,

[1] (*B*) *op. cit.*, "Variations in radar coverage," pp. 127–133.
[2] *Ibid.*, "Tropospheric propagation and radio meteorology," pp. 134–165.
[3] Sir E. Appleton (*D*) *op. cit.*, "The influence of tropospheric conditions on ultra-short-wave propagation," pp. 1–17.
[4] H. G. Booker, "Elements of radio meteorology: how weather and climate causes unorthodox radar vision beyond the geometrical horizon," *Jour. IEE*, vol. 93, IIIA, pp. 69–78, no. 1; 1946.
[5] G. N. Watson, "The Diffraction of Electric Waves by the Earth," *Proc. Roy. Soc. A* (London), vol. 95, pp. 83–99; 1918. The Transmission of electrics waves by the earth, *ibid.*, pp. 546–563.

and the above formulas could be used with this modification to include the refraction effect. *Standard refraction* is then defined as one in which the decrease of refractive index with height corresponds to an effective earth's radius equal to 4/3 of the actual radius.

For wavelengths in the centimeter range this theory was soon found to be inadequate. Meteorological observations have shown that at sufficiently large heights the refractive index usually decreases linearly with height. However, in the immediate neighborhood of the earth's surface there may be large deviations from a linear relationship. In the modified index of refraction, defined as $M=[n-1+(h/a)]\ 10^6$, n may decrease linearly but much more rapidly than is the case for a standard atmosphere. In these circumstances rays which start out at very small angles to the horizon are bent back towards the earth, from which they are reflected, and the process may be repeated. The heights within which the rays are trapped define the confines of a *duct*.

A discussion of the elementary theory of nonstandard propagation will be found in volume I of reference (*D*) (pages 11 to 17). For very short wavelengths the field at any point in the duct may be determined by a ray-tracing method, based on geometrical optics. This method is also discussed in Bremmer's book on pages 131 to 138, and in some of the papers in the Physical Society publication (*C*), notably those by Hartree, Michel and Nicholson (page 127) and by Stickland (page 253). A critical study of ray versus diffraction techniques, earth flattening approximation, and limitation of ray methods is contained in chapter 2 of (*A*).

Bremmer has also written of ray treatment of propagation in a duct.[6] He assumes a quadratic variation of refractive index with height, and discusses multihop transmission over the earth's surface, together with such topics as the time of arrival of consecutive rays originating from a point source. The existence of a shadow zone above an atmospheric duct is indicated in another paper[7] which uses ray-tracing methods to investigate the shadow zone, and shows that measurements of the M variations and of the associated fields agree qualitatively with theoretical values.

B. Modes and Ducts

When the wavelength is not short enough to use the ray theory in a duct, some progress can be made with a mode theory similar in principle to the usual mode theory of propagation in a metallic waveguide. Above the geometrical horizon, too many modes are involved to make the theory useful, but below the horizon it often happens that the lowest mode by itself gives a tolerable description of the field. A drawback in the practical use of the mode theory is the lack of propagation curves covering sufficiently flexible profiles of refractive index.

Booker and Walkinshaw give a general description of the mode theory,[8] followed by propagation curves appropriate to a dependence of modified refractive index μ upon height h of the form

$$\mu^2 = 1 - 2k_\infty[h - (d^{1-m}h^m/m)].$$

Section 8 gives the curves for $m=1/2$ (the square-root profile), and section 9 those for $m=1/5$ (the fifth-root profile). The drawback about these profiles is that although they give an appropriate lapse-rate of refractive index well above the duct, the curves do not tend to an asymptote at great heights. The curves for the fifth-root profile are presented in a manner that is different from, and better than, those for the square-root profile. The method of presentation of these fifth-root profile curves, even when applied to the standard atmosphere without ducts, seems to be the most convenient yet devised for the quick drawing of a field strength curve beyond the horizon.

Reference 9 contains curves of horizontal attenuation for the lowest mode at 200, 3,000, and 10,000 mc for a profile of modified refractive index consisting of two straight lines forming an angular minimum at the top of the duct.[9] Height-gain curves are not given. A drawback of this profile is that the minimum of modified refractive index is usually quite blunt in practice.

Near the surface of the sea there is a semi-permanent surface duct that is particularly well developed in those part of the oceans that are swept by trade winds. The influence that this duct can have on propagation of centimeter wavelengths near the ocean surface has been described.[10] In this article Figs. 10 and 11 summarize the measurements made at 9 and 3 cm, respectively. The comparison of those results with the duct theory has also been discussed.[11] For a discussion of the departure of the observation from the duct theory at long range (beyond about 80 km) at 9-cm wavelength, see Booker and Gordon.[31]

For anyone who proposes to calculate additional propagation curves reference should be made to other works.[12–15]

C. Elevated Layers

Partial and total return of energy from elevated refracting layers in the troposphere at heights of some

[6] H. Bremmer, "On the theory of spherically symmetric inhomogenous wave guides in connection with tropospheric radio propagation and underwater acoustic propagation," *Philips Res. Rep.*, vol. 3, pp. 102–120; April, 1948.

[7] W. L. Price, "Radio shadow effects produced in the atmosphere by inversions," *Proc. Phys. Soc.* (London), vol. 61, pp. 48–59; July, 1948.

[8] H. G. Booker and W. Walkinshaw, "The mode-theory of tropospheric refraction and its relation to wave-guides and diffraction," (*D*) *op. cit.*, pp. 80–127.

[9] (*C*) *op. cit.*, "Attenuation diagrams for surface ducts," pp. 178–180.

[10] M. Katzin, R. W. Bauchman, and W. Binnian, "3 and 9 cm propagation in low ocean ducts," Proc. I.R.E., vol. 35, p. 891; September, 1947.

[11] C. L. Pekeris, "Wave theoretical interpretation of propagation of 10 cm and 3 cm waves in low-level ocean ducts," Proc. I.R.E., vol. 35, pp. 453–462, July, 1947.

[12] C. L. Pekeris, "Asymptotic solutions for the normal modes in the theory of microwave propagation," *Jour. App. Phys.*, vol. 17, pp. 1108–1124; December, 1946.

[13] C. L. Pekeris, "Perturbation theory for an exponential M curve in nonstandard atmospheric conditions," (*C*) *op cit.*, pp. 185–190.

[14] G. G. Macfarlane, "The application of a variational method to the calculation of radio wave propagation curves for an arbitrary refractive index profile in the atmosphere," *Proc. Phys. Soc.*, vol. 61, p. 48; July, 1948.

[15] D. E. Kerr, "Theory of propagation in a horizontally stratified atmosphere," (*A*) *op. cit.*, chap. 2, pp. 27–144.

thousands of feet is likely to be an important phenomenon, especially at meter wavelengths. The phenomenon is likely to occur especially in connection with what meteorologists call a "subsidence" inversion. The phenomenon occurs in quite a persistent form in certain areas of the world, two of which are the coast of California and the coast of New South Wales.

Reference 16 is a description of the distribution of temperature, humidity, and refractive index experienced in practice off the Southern California shore.[16] Radio results in the region have been described also.[17] Typical signal records for low and high reflecting layers are shown in Fig. 5, together with a condensed log of field-strength (Fig. 3) over a 90-mile link over sea-water along the California coast, showing the dependence of field on layer height. Frequencies used were 52 mc, 100 mc, and 547 mc. Theoretical reflection coefficients from a layer having a plausible distribution of refractive index with height ("Epstein" layer) are shown in Fig. 2.

A description is given by Price of experiments showing reflection from an elevated layer off the coast of New South Wales,[7] and the ray theory of refraction in such a layer is developed to show that regions of shadow could develop which might interfere with detection of aircraft by coastal radars. For a general discussion of ray tracing see Section VI (A) of this guide.

Some statistical results concerning the occurrence in anticyclones of the type of elevated layer that is of importance in radio communication are to be found in reference[18].

A description[19] of the semi-permanent elevated duct covering practically the whole of that part of the world which is subject to trade winds is based on upper-atmosphere observations of the research ship "Meteor" in the Atlantic Ocean.

D. Angle of Arrival

Variations in tropospheric refraction give rise to changes in the angle of arrival of radio waves amounting rarely to more than a small fraction of a degree. Measurements of the angle of arrival may be checked with expectations based on observed index of refraction profiles. On the practical side, stimulus for these measurements arises from two typical types of questions. (1) What is the largest microwave antenna which can be used on a relay path without causing fading due to the fluctuations of the angle of arrival comparable with the antenna beam width? (2) What is the ultimate angular accuracy which can be hoped for in the fire-control radar, as set by the unavoidable fluctuations in the angle of arrival through the atmosphere?

Horizontal angle of arrival measurements are relatively simple since it may reasonably be assumed that if more than one wave component is present, like interference will occur over the antenna's horizontal dimension. For the vertical-angle measurements, however, more than a single wave component is commonly received, causing a complex-wave front in the vertical direction.

The studies of the angle of arrival have been concerned with measuring the changes of angle as a function of meteorological conditions, the detection of multiple-path transmission, the separation of wave components for line-of-sight paths, and the measurement of the angle of arrival of the refracted wave front for nonoptical paths. As at lower ionospheric frequencies, two experimental methods have been used: (1) sharp-beamed antennas scanned mechanically in angle, or (2) simultaneous-phase measurements on spaced antennas. The first method is limited in the ability to follow rapid variations; the second gives results in a form difficult to interpret directly. The following papers[20–22] illustrate the available data and methods.

The measurement of horizontal and vertical angles of arrival were made on 13- and 24-mile paths using a wavelength of 3.2 cm. The method employed was that of rocking a 20-foot parabolic antenna with a beam width of 0.36° in the plane of measurement through an angle of $\pm 3/4°$ on either side of the "normal" angle of arrival. The position of the peaks on a continuous-recording chart was used to obtain the angle of arrival.

The 13-mile path was over rolling land and the effective reflection coefficient was found to be 0.18. The longer path was partly over water and partly over land, and the effective reflection coefficient was found to be 0.5. Very little horizontal change in angle of arrival was reported except for an occasional horizontal shift of angle up to 0.1°. Vertical variations were more commonly noted. On the 24-mile path, the direct ray was observed to arrive as much as 0.35° above the angle associated with standard atmospheric conditions.

Additional measurements of the angle of arrival on the 13-mile path mentioned in the preceding paper were made on a 3.2-cm and 1.25-cm wavelength. The 1.25-cm measurements were made with a 20-foot metal lens antenna with a beam width of 0.12°. A collector antenna was 48.6 feet to the rear of the lens. The beam was scanned by moving the lens up or down or sideways to change the focusing on a collector antenna 48.6 feet to the rear of the lens.

The horizontal-plane angular deviations were usually absent, and never more than 0.03°. The vertical-plane angular deviation varied from 0.04° to 0.11° relative to the line of sight on both wavelengths considered. A general correlation with measured index gradients was

16 (C) *op. cit.*, "Meteorology of the San Diego transmission experiments," pp. 202–206.
17 J. B. Smyth and L. G. Trolese, "Propagation of radio waves in the troposphere," Proc. I.R.E., vol. 35, p. 1199; November, 1947.
18 S. Pettersen, P. A. Sheppard, C. H. B. Priestley and K. R. Johannsen, "An investigation of subsidence in the free atmosphere," *Quart. Jour. R. Met. Soc.*, vol. 73, p. 43; 1947.
19 (C) *op. cit.*, "Analysis of ducts in the trade winds," p. 223–225.
20 W. M. Sharpless, "Measurement of angle of arrival of microwaves," Proc. I.R.E., vol. 34, pp. 837–845; November, 1946.
21 A. B. Crawford and W. M. Sharpless, "Further observations of the angle of arrival of microwaves," Proc. I.R.E., vol. 34, pp. 845–848; November, 1946.
22 A. W. Straiton and J. R. Gerhardt, "Results of horizontal microwave angle-of-arrival measurements by the phase difference method," Proc. I.R.E., vol. 36, pp. 916–922; July, 1948.

found. The angles measured on the two wavelengths agreed.

Multiple-path transmission was noted on two clear, calm nights. As many as four separate paths were noted, with the highest angle 0.75°. It was proposed that these additional paths were due to boundary layer reflections.

The horizontal angle of arrival of 3.2-cm radio waves on a 7-mile path along a coast line was observed. The principle employed was that of using the phase difference between the fields of two antennas separated horizontally by 10 feet. Meteorological measurements of the horizontal gradient of the index of refraction were made at two points along the path and at three levels.

For 90 per cent of the measurements the angle of arrival was shifted shoreward from the line of sight, with angles of 0.015° or more existing 40 per cent of the time. This horizontal shift was associated with the index gradient which existed from the more moist and cooler air over the water to the dryer and warmer air over the land. A general correlation was observed between the radio and meteorological measurements.

VII. Lobe Structure

The main feature of the field-strength distribution in connection with propagation through the troposphere above the horizon is the lobe structure formed by the interference pattern between the direct and surface-reflected waves, known as the Lloyd's mirror phenomenon in optics. During the war, contours of constant field strength in a vertical plane through this interference pattern became known variously as the "vertical polar diagrams" or "vertical coverage diagrams." For VHF radar equipments in particular, the field-strength contour corresponding to the minimum detectable signal is a much better indicator of set coverage than the simple "line of sight" criterion of the preceding decade. For a plane earth the lobe structure is especially simple, since the minima separate the lobes at angles of elevation which are integral multiples of $\lambda/2h$, where λ is the wavelength and h the antenna height. Over a curved earth the vertical angular spacing is no longer uniform, and the precise calculation of lobe boundaries becomes rather tedious in practice because of computational difficulties involved in finding the location of the reflection point on a curved earth. The first part of chapter 5 (pages 377 to 404), and chapter 6 (pages 433 to 453) of reference (*C*), as well as sections 2.1, 2.2, and 2.13 of reference (*A*), give considerable detail concerning the calculations, beginning with the elementary calculation of the plane-earth approximation and proceeding to a variety of sophisticated methods which have been devised to circumvent the complications of the curved-earth calculations. Very few of the thousands of experimentally-determined radar coverage diagrams have been published, but the following reference[23] gives some indication of the extent to which the details of the smooth-earth calculation are verified in practice. Section X of this guide should be consulted for data which indicates in a general way the effect of surface land or sea roughness in reducing the reflection coefficient, and thus moderating the depth of the nulls of the interference pattern.

[23] J. A. Ramsay, "The vertical distribution of radar field strength over the sea under various conditions of atmospheric refraction," (*D*) *op. cit.*, pp. 238–242.

VIII. Time and Space Variations of Field Strength

The enormous and sudden expansion of television and FM broadcasting on VHF since the war and the proposed extension to UHF have precipitated intense practical interest in certain propagation questions peculiar to a broadcast service, especially for the purposes of frequency-band allocations and station spacing. The effects of particular importance to broadcasting are: (1) the effects of rough terrain and of urban locations on the field strengths received in home locations in a broadcasting station's service area, compared with calculations based on a smooth earth; (2) the tropospheric interference fields produced well beyond the horizon by a high-power broadcasting station capable of interfering with another station on the same channel. With respect to the rough terrain effect on VHF, the average departure of the median service area field from the conventionally calculated smooth-earth field, and the statistical distribution about this median may be found for a number of VHF broadcasting stations in eastern United States in a report of the Ad Hoc Committee.[24]

Estimates of the shadow loss exceeded at a specified percentage of locations behind a hill of known height between the transmitter and receiving locations may be found on pp. 29–30 of Proceedings.[25] Some information of comparative measurements in different types of terrain at UHF as well as at VHF may be found in the *RCA Review*[26] and Proceedings.[27]

A summary of all long-term recording up to mid-1949 of VHF time variable field strength, and average curves deduced from these as to field strength exceeded 1, 10, 50 and 90 per cent of the time at various distances may be found in an FCC report.[28] Statistical summaries of later VHF recordings and of a few UHF recording may be found in two other FCC reports.[29,30]

[24] K. A. Norton, M. Schulkin, and R. S. Kirby, "Ground wave propagation over irregular terrain at frequencies above 50 mc," reference (*C*) to the report of the Ad Hoc Committee of the FCC; June 6, 1949.

[25] K. Bullington, "Radio propagation variations at VHF and UHF," Proc. I.R.E., vol. 38, pp. 27–32; January, 1950.

[26] G. H. Brown, J. Epstein and D. W. Peterson, "Comparative propagation measurements; television transmitters at 67.25, 288, 510, and 910 megacycles," *RCA Rev.*, pp. 177–201; June, 1948.

[27] A. J. Aiken and L. Y. Lacy, "A test of 450 megacycle urban area transmission to a mobile receiver," Proc. I.R.E., vol. 38, pp. 1317–1319; November, 1950.

[28] E. W. Allen, W. C. Boese, and H. Fine, "Summary of tropospheric propagation measurements and the development of empirical VHF charts (revised)," FCC Report TID 2.4.6; May 26, 1949.

[29] G. V. Waldo, "Summary of recent VHF tropospheric propagation measurements over southern and mid-western paths," FCC report TRR 2.4.8; June, 1950.

[30] H. Fine and F. V. Higgins, "Long-distance tropospheric propagation in the ultra high frequency band," FCC TRR report 2.4.10; October 13, 1950.

One paper has appeared[31] which attempts to interpret the unexpectedly high-tropospheric field strengths measured well beyond the horizon as being caused by scattering from blobs in the atmosphere related to turbulent fluctuations in the index of refraction. Formula (2.23) gives the power scattered by the atmosphere (*i*) per unit solid angle, (*ii*) per unit incident power density, (*iii*) per unit volume, as a function of the scale of turbulence, the mean square deviation of dielectric constant from mean, the angle through which scattering takes place, an angle depending on polarization, and the wavelength. Section III gives various approximations to the general formula. Difficulties arise in applying the formulas because the volume integration is sometimes difficult (*i*). Calculation is done in Section V for beam communication using identical beam-widths for transmission and reception, assuming no variation with height of the scale of turbulence and the mean square deviation of the dielectric constant from mean. Agreement with observation appears possible in this case. Knowledge of the scale of turbulence in the atmosphere and of the mean square deviation of dielectric constant from mean is poor, especially so far as its variation with height is concerned (*ii*). A review of available information on this subject is given in Section I.

IX. Attenuation by Atmospheric Gases and Precipitation

The theory, methods of measurement, and experimental values of the absorption of microwaves by atmospheric oxygen and water vapor are compactly presented by Van Vleck and Purcell in chapter 8 of (*A*), pages 646 to 671, along with references to post-war measurements as well. Attenuation by precipitation (clouds, rain, and snow) is reviewed both on the theoretical and experimental side by Goldstein in chapter 8 of (*A*), pages 671 to 692. A somewhat more extensive account on the theoretical side and a summary of experimental values up to the end of the war is contained in the article by Goldstein in (*C*), pages 269 to 284.

X. Ground and Sea Roughness

Roughness of the ground is of importance at sufficiently short wavelengths both in communication and in radar. It affects the reflection coefficient of the ground, and in extreme cases may even lead to indirect transmission paths. In the radar application surface-roughness leads to echoes that may obliterate wanted targets.

Reference (*C*) contains material on the subject of which the following specific items are typical. (Alternatively, chapter 5 of reference (*A*) discusses much of the same material, and includes an attempt to interpret the meaning of the measurements.)

Chapter 4, page 263, includes some information on the effect of grass-cover on reflection from ground at a wavelength of 9 cm. Magnitudes of reflection coefficients are quoted for both horizontal and vertical polarization for angles of elevation up to 56.5°. The height of the vegetation runs from 0 to 45 cm.

Chapter 4, page 267, gives information about the magnitude of the reflection coefficient of 9-cm waves (both vertically and horizontally polarized) from ground that has been artificially ridged. Angles of elevation from 12 to 46.5° are considered, with the plane of propagation along, across, and at 45° to the run of the ridges. The ridge wavelength varied from 60 to 120 cm and the ridge height from 5 to 15 cm.

Chapter 7, page 306, gives information about radar clutter from sea and land surfaces on 9 and 3 cm for vertical and horizontal polarization. On page 311 figures are quoted for the ratio of scattering cross section of the sea per unit are on 9 to that on 3 cm. Fig. 2 relates to the time constant of sea-clutter at 9 and 3 cm.; Figs. 7 and 8 give similar information for ground clutter at 9 cm, wind speeds of 25 mph and 10 mph being involved in these two diagrams.

Reference (*A*) contains numerous references to these topics of which the following are representative. On pages 411 to 418 the theory of surface roughness is examined qualitatively in terms of half-period zones drawn about the reflection point. Pages 419 to 436 give a summary of experiments and their interpretation which lead to values of the reflection coefficient for various land and sea surfaces. An extensive treatment of both theory and experiment on sea echo and land clutter is given by Goldstein on pages 481–587.

In an article by Davies and Macfarlane[32] equations (7) and (12) are approximate formulas for radar reflection from the sea surface applicable at angles of elevation small and large compared with θ_0 defined a quarter of the way down page 721. Results are expressed in terms of a scattering coefficient $f(\theta)$ of the sea that depends on angle of elevation θ. The dependence of $f(\theta)$ on wave height for $\lambda = 3$ cm and 1.25 cm is given in Fig. 6, on θ for $\lambda = 3$ cm in Fig. 8 and on wind speed for $\lambda = 10$ cm in Fig. 10.

Practically all of the information on surface roughness mentioned above concerns its effects on propagation over optical paths, whether both radar or one-way applications. Recently a suggestion has been outlined by Bullington that surface irregularities may be important for understanding of the unexpectedly high-field strengths from high-powered vhf transmitters well beyond the horizon.[33]

XI. Meteorological Echoes

Chapter 7 (pages 588 to 640) in reference (*A*) gives a comprehensive account of the subject, beginning with a theoretical outline of the origin of meteorological echoes from assemblages of water drops, and concluding with a select but comprehensive illustrative photograph

[31] H. G. Booker and W. E. Gordon, "A theory of radio scattering in the troposphere," Proc. I.R.E., vol. 38, pp. 401–412; April, 1950.

[32] H. Davies and G. G. Macfarlane, "Radar echoes from the sea surface at centimeter wavelengths," *Proc. Phys. Soc.*, vol. LVIII, p. 717; 1946.

[33] K. Bullington, "Propagation of UHF and SHF waves beyond the horizon," Proc. I.R.E., vol. 38, p. 1221, October, 1950.

of rain echoes on PPI tubes of microwave radars. For a period after the war, echoes received from an optically clear atmosphere within a few thousand yards of a microwave radar were provisionally ascribed to reflection from abrupt discontinuities in the refractive index of the troposphere. Crawford has published evidence that many, if not all, such "angel" echoes must be due to "insects."[34] Thus angels have turned out sometimes to be insects!

XII. Velocity of Propagation

A paper by Smith-Rose comprises a review of the knowledge about speed of transmission of radio waves under the practical conditions of certain applications in which such knowledge is important.[35] It is shown first that, for radio waves in a vacuum, their speed of transmission is equal to the velocity of light, to within the limits of experimental error. When waves of frequencies in the neighborhood of 100 kc are propagated at a height of a fraction of a wavelength above the earth's surface, their speed is reduced by an amount dependent upon the electrical conductivity of the earth. For overland transmission, the speed is about 299,250 km. For higher frequencies propagated at a height of several wavelengths, the speed of the waves is determined by the refractive index of the air, rather than by the properties of the ground. Since the refractive index decreases with the height of transmission, so does the speed of the waves increase toward the velocity of light. For example, centimeter waves propagated at heights of a few hundred feet have been observed to travel at a speed of about 299,690 km. When the waves are transmitted between ground and aircraft flying at a height of 30,000 feet (9,800 meters) this speed is increased to about 299,750 km.

Essen[36] has recently proposed a new "most probable" value for the velocity of electromagnetic waves in vacuo, based on a weighted average of all the precision measurements of this fundamental constant, both at optical and radio frequencies.

The pioneering work of Aslakson in the use of radio navigation equipment for the measurement of the velocity of radio waves in the atmosphere is presented in refs. 37–39. The 1949 paper[37] is a summary of the use of Shoran for the measurement of geodetic distances, and of improved values for the velocity of propagation deduced therefrom. The 1950 paper[38] concerns itself with the effect of meteorological conditions in the troposphere on the velocity, and the 1951 paper[39] gives the most recent results obtained with improved radio equipment. The provisional recommended value for the velocity in vacuo is 299,793.1 km/sec, which is 17.1 km /sec higher than the value widely recommended before the war for this very fundamental physical constant. It would appear that post-war radio science may contribute a new order of accuracy to the experimental measurement of this fundamental constant of classical, relativity, and quantum physics. Current views of physicists about the "best" value for the velocity of light, as well as for other related fundamental atomic constants may be found in the *Physics Review*.[40,41]

Acknowledgment

The foregoing is the work of the following contributing members of the Subcommittee on the Theory and Application of Tropospheric Propagation of the IRE Wave Propagation Committee: Henry G. Booker (Chairman, 1949), Thomas J. Carroll (Chairman, 1950–51), A. W. Straiton, R. P. Wakeman, D. E. Kerr, C. R. Burrows, and Miss M. C. Gray.

[34] A. B. Crawford, "Radar reflections in the lower atmosphere," Proc. I.R.E., vol. 37, p. 404; 1949.

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[36] L. Essen, "Proposed new value for the velocity of light," *Nature*, vol. 167, pp. 258–259; February 17, 1951.

[37] C. I. Aslakson, "Can the velocity of propagation of radio waves be measured by Shoran?" *Trans. Amer. Geophys. Union*, vol. 30, pp. 475–487; August, 1949.

[38] C. I. Aslakson and O. O. Fickeissen, "The effect of meteorological conditions on the measurement of long distances by electronics," *Trans. Amer. Geophys. Union*, vol. 31, pp. 816–826; December, 1950.

[39] C. I. Aslakson, "A new measurement of the velocity of radio waves," *Nature*, vol. 168, pp. 505–506; September 22, 1951.

[40] J. A. Bearden and H. M. Watts, "A re-evaluation of the Fundamental Atomic Constants," *Phys. Rev.*, vol. 81, pp. 73–81; Jan. 1, 1951.

[41] J. W. M. DuMond and E. R. Cohen, "Least-squares adjusted values of the atomic constants as of December, 1950," *Phys. Rev.*, vol. 82, pp. 555–556; May 15, 1951.

An Experimental Study of Wave Propagation at 850 MC*

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Summary—**The prediction of area coverage of a broadcast station is of prime importance. To do this requires the evaluation of the effects of such factors as wave refraction, earth reflection, diffraction, and attenuation. The following paper describes an experiment conducted at 850 mc which was undertaken to solve this problem. We have concluded that useful predictions of wave propagation can be made with free space theoretical field strength reduced by theoretical knife edge diffraction shadow loss and by suitable empirical experience factors.**

Introduction

IN SELECTING the site and characteristics for a broadcast transmitting antenna, the foremost consideration is that of obtaining optimum coverage in the area to be served. To effect a reasonable approach to the problem, the engineer must be provided with a means of predicting the coverage to be expected from a given antenna location when terrain characteristics are known. This prediction requires a knowledge of radio propagation together with the ability to evaluate such well-known factors as wave refraction, earth reflection, diffraction, and attenuation.

The goal of our study is to formulate procedures to enable the prediction of field strength throughout a broadcast service area. The achievement of such a goal will depend upon the accumulation of experimental data against which theoretical formulations can be checked.

We shall study propagation characteristics near the upper edge of the uhf television band. We shall further narrow the problem by limiting distances to 30 or 40 miles. This restriction usually will permit the variation of field strength with time[1,2] to be ignored.

* Decimal classification: R122. Original manuscript received by the Institute, January 20, 1953; revised manuscript received January 27, 1953.

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[1] G. S. Wickizer, "Field strength of KC2XAK, 534.75 mc recorded at Riverhead, N. Y." (33 mile path). PROC. I.R.E., vol. 41, p. 140; January 1953.

[2] G. S. Wickizer and A. M. Bratten, "Propagation studies on 45.1, 474, and 2,800 megacycles." PROC. I.R.E., vol. 35, p. 670; July, 1947.

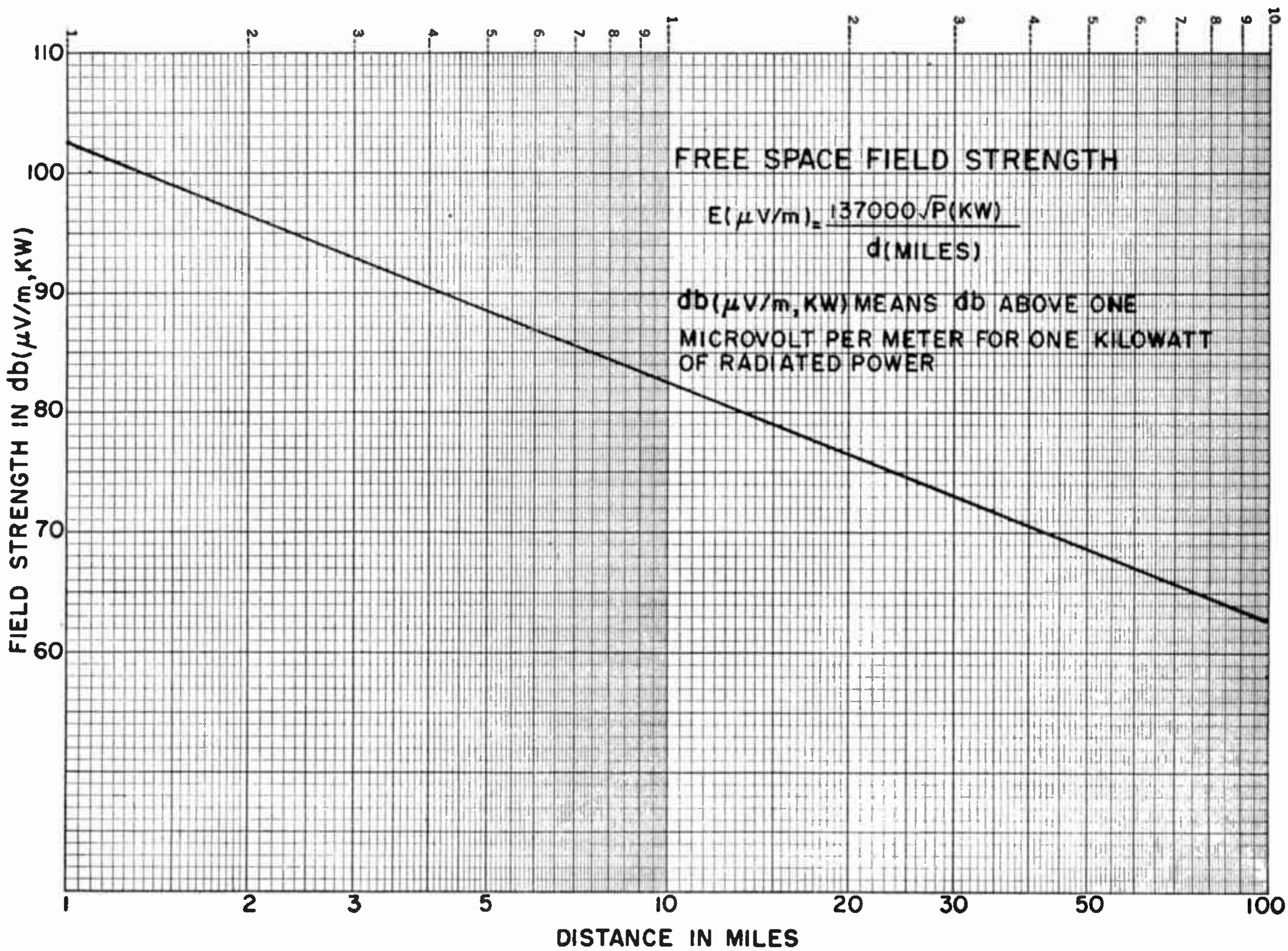


Fig. 1—The free space field strength of a one-half wavelength antenna as a function of distance for one kilowatt of radiated power.

We shall describe an 850 mc experimental project aimed at obtaining some of the needed information. Since propagation is related to the height of both transmitting and receiving antennas, arrangements were made to vary these factors. Provisions were made for measurements along two quite different radials, one flat and the other relatively hilly. The data obtained were analyzed statistically to indicate the trend of the field-strength median for a variety of receiving conditions. An effort was made to separate and measure (a) loss introduced by houses and trees, and (b) loss attributable to hills. A knowledge of the magnitude of these two effects under known experimental conditions is required to formulate a theoretical basis for calculating wave propagation for known topography and surface conditions.

Propagation Theory

The simplest conceivable propagation occurs in free space, either air of uniform density or vacuum. For a given radiated power the field strength, Fig. 1, is independent of frequency and varies inversely with distance. A simple mathematical analysis shows this.

The introduction of smooth earth and standard atmosphere complicates the situation but still leaves it in the realm of mathematical analysis. The complications may be described piecemeal by first looking at propagation over a patch of smooth earth small enough to be treated as planar. Horizontally polarized uhf radio waves are reflected by smooth earth. For our purpose, the reflection is like that which would be produced by a perfectly conducting surface. The field strength at any receiving point is thus a combination of two vector quantities, the field from a direct ray added to the field from an earth reflected ray. At 850 mc, this results in field strength which varies with distance as shown in Fig. 2 in which field strength is calculated for a receiving antenna height of 30 feet. The relation between field strength and antenna height at various distances is shown by the interference patterns of Fig. 3.

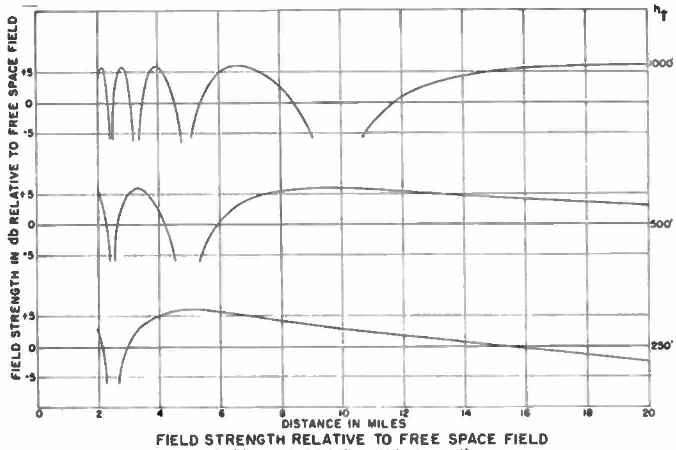


Fig. 2—The received field strength as a function of distance for a given transmitting and receiving antenna height.

For distances too great to be treated as planar it has seemed reasonable to treat the surface of the earth as a sphere and calculate the diffraction around the spherical surface. This has been done in a manner suitable for engineering use but will not be required in our analysis.

The presence of atmosphere introduces refraction since the air is more dense near the earth. This causes the wave to bend toward the earth. So-called "standard atmosphere" conditions prevail when the air is undisturbed by turbulence or stratification. It has been shown that use of an earth's radius equal to approximately 4/3

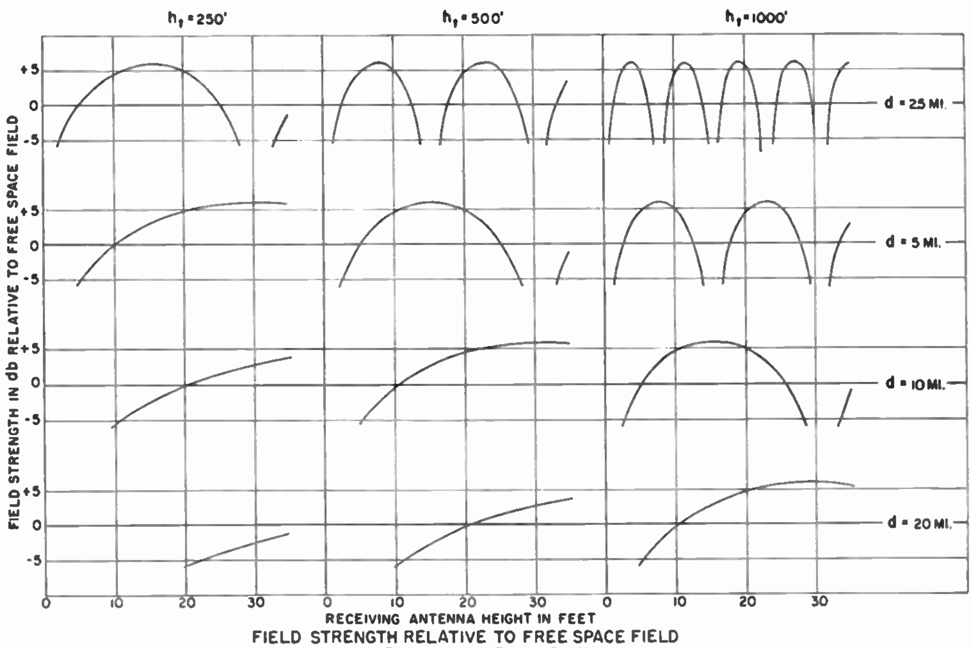


Fig. 3—The field strength as a function of receiving antenna height and distance between the two antennas for a given transmitting antenna height.

true radius, in plotting profiles of propagation paths, gives a satisfactory correction for this effect. It should be pointed out, however, that "standard atmosphere" may never exist along an entire propagation path.

Atmospheric complications include scattering, ducting, super-refraction and probably other important phenomena. All of these phenomena are functions of weather conditions and, consequently, resulting field strengths are subject to time variations. These variations will usually be most important at the greatest distances to be considered for broadcasting but may be important at any distance. Usually 30 miles may be taken as an average distance within which time variations can be ignored.

We have seen that there are three simple theories of propagation, namely, (a) free space, (b) smooth, perfectly conducting plane earth, and (c) smooth, perfectly conducting spherical earth. All assume sets of conditions fixed in time. All are useful for certain practical 850 mc propagation situations. They are not at all adequate, by themselves, for predicting TV broadcasting propagation at this or other frequencies in the uhf range.

The propagation theory which has been outlined omits two important factors, (a) the loss introduced by large obstacles in the propagation path and (b) the loss from relatively local clutter on the earth's surface near the receiving site. For lack of more rigorously applicable theory the large obstacles, that is hills, may be treated as perfectly absorbing knife edges.[3,4] The theory thus derived yields shadow loss which increases with frequency and fits experimental data well enough to be useful. The effect of local surface clutter can hardly be treated mathematically except on a statistical basis. Therefore, empirical "experience factors" will be derived for prediction of this loss.

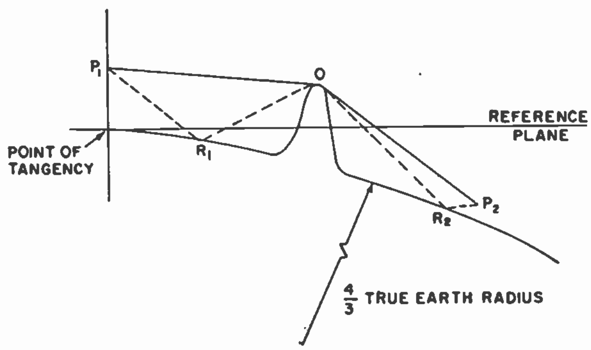


Fig. 4—An idealized picture illustrating the various possible transmission paths between P_1 and P_2.

An idealized picture of propagation which we shall use is illustrated in Fig. 4. Transmission from P_1 to P_2 may occur over ray paths P_1OP_2, $P_1R_1OP_2$, $P_1R_1OR_2P_2$ and $P_1OR_2P_2$. Surface conditions at R_1 and R_2 will determine the existence of all contributions to the field at P_2 except that which came via P_1OP_2. The proximity of a ray to the earth's surface will also be considered to be of importance in determining the weight given to the experience factor. It is apparent that surface roughness is extremely important in this propagation picture.

It will be helpful for our discussion to classify surface roughness into three broad classes: (a) Hill and valley features, (b) Buildings and trees, and (c) Fine degree roughness, such as grass, farm crops, waves, etc. These will be designated as first, second and third order roughness. Any or all degrees of roughness may be involved in a propagation problem.

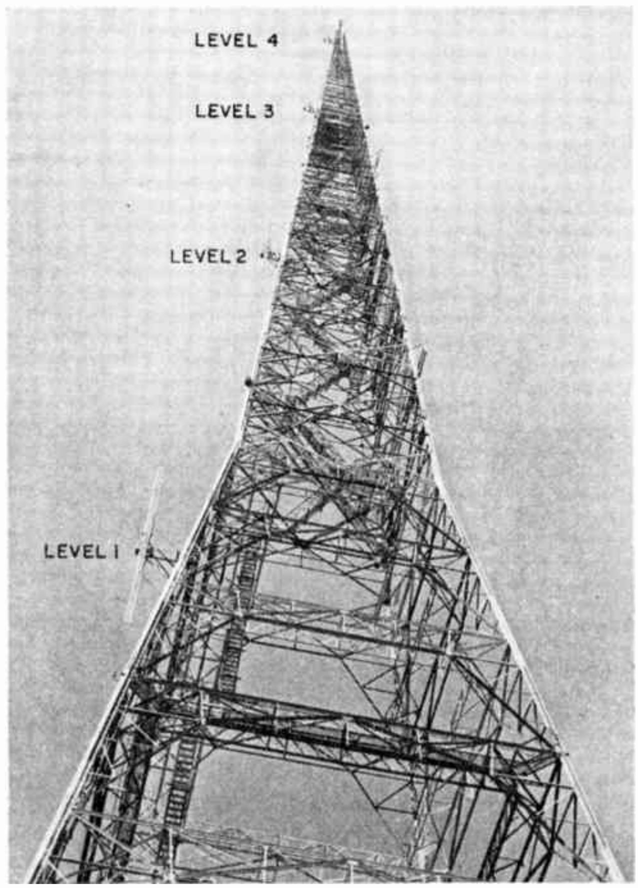


Fig. 5—A picture of the WOR 760-foot TV tower showing the location of the four transmitting antennas used in the propagation tests.

Field Survey

Field experiments were conducted in the summer of 1952 with apparatus installed on the WOR 760-foot TV tower on the Palisades, just west of New York City (see Figs. 5 and 6). Four horizontally polarized unidirectional antenna arrays designated as levels 1, 2, 3, and 4, were evenly spaced along the height of the tower. The antennas (Fig. 7) are 48-feet long, have vertical (half field) beam width of 1.6 degrees and horizontal beam width of 92 degrees (Fig. 8). They are zig-zag

[3] Kenneth Bullington, "Radio propagation above 30 mc," Proc. I.R.E., vol. 35, p. 1122; October 1947.
[4] Kenneth Bullington, "Radio propagation variations at vhf and uhf," Proc. I.R.E., vol. 40, p. 27; January 1950.

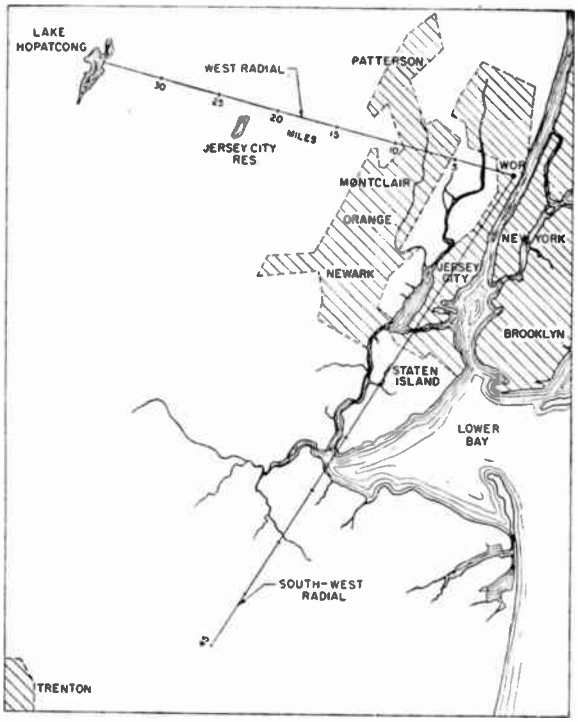


Fig. 6—A map showing the location of the two radials along which the measurements were made.

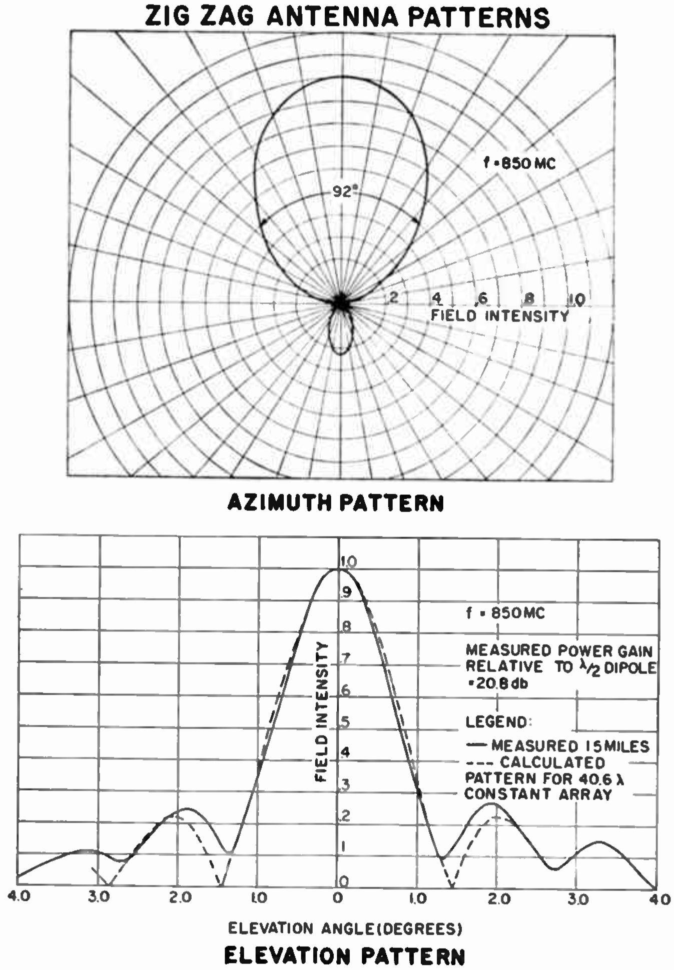


Fig. 8—The elevation and azimuth radiation patterns of the transmitting antennas.

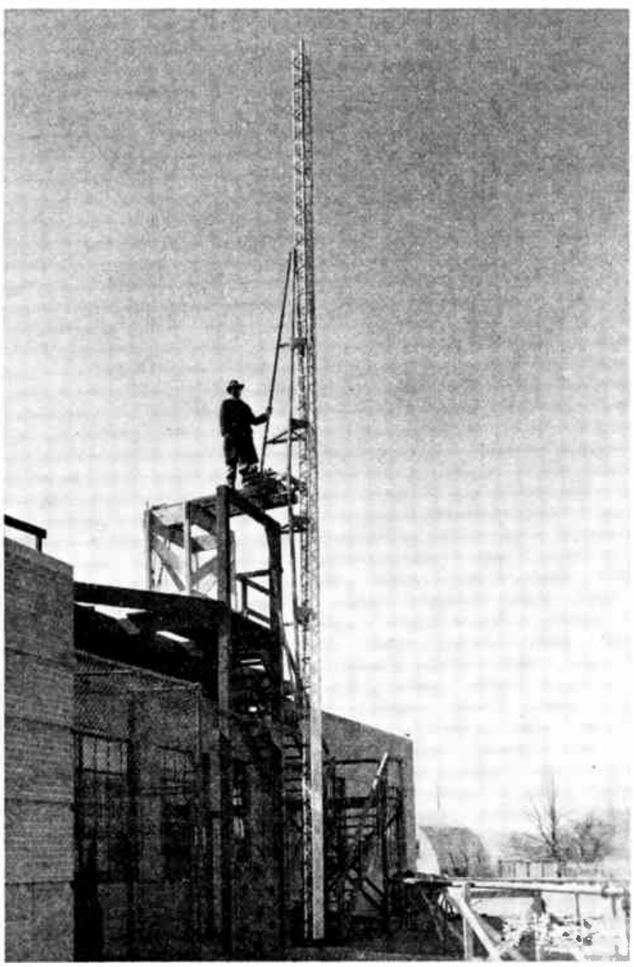


Fig. 7—A close-up of one of the uni-directional transmitting antennas

Fig. 9—(left) Making the field tests.

arrays with only six feed points per antenna. The resulting high gain of 20 db was necessary because of the low transmitter power available. The transmitters were 10 watt, self-excited oscillators remotely controlled and mounted at the antennas to eliminate costly and lossy transmission lines. Because of the narrow vertical patterns of the antennas they were tilted vertically to enable measurement on pattern maxima. This was also done by remote control, with selsyns to convey accurate angular position to the operator.

Field observations were made in a car, Fig. 9, in radio communication with the station operator. It was equipped with a motor-driven, telescoping, 35-foot mast installed periscope fashion in the roof. The mast could be raised in 30 seconds, so field strength versus height observations were easily and quickly made.

Two radial lines, which are shown in Fig. 6 and are designated as west and southwest radials, were laid out on the topographic map. The general character of first order roughness can be seen from profiles plotted for the radial lines, Fig. 10. The horizontal axis for these profiles is tangent to the sea level elevation at the transmitting antenna location. The sea level elevation was plotted below the horizontal axis using 4/3 the true earth radius to allow for the refraction of "standard atmosphere." Similar profiles, Figs. 11(a) and 11(b), for radial lines which pass through actual measuring sites were used for a study of the measured data given in Tables 1–7. When first order obstructions existed in direct-ray paths, these profiles were used to obtain obstruction heights.

In the course of the survey it was observed that the dependence of field strength upon receiving antenna height was extremely variable. Since no functional relationship could be established to relate the variations, it was decided to use the maximum field strength obtainable between ten and thirty feet rather than the value for a fixed receiving antenna height.

Second Order Roughness "Experience Factors"

An effort was made along the southwest radial to obtain a fairly homogeneous sampling of measurements in a congested residential area. All measurements were made with two story houses in the immediate foreground along a portion of the radial (0 to 13 miles) where first order roughness was negligible. There was almost always increase of field strength with receiving antenna height, because the antenna was elevated to a

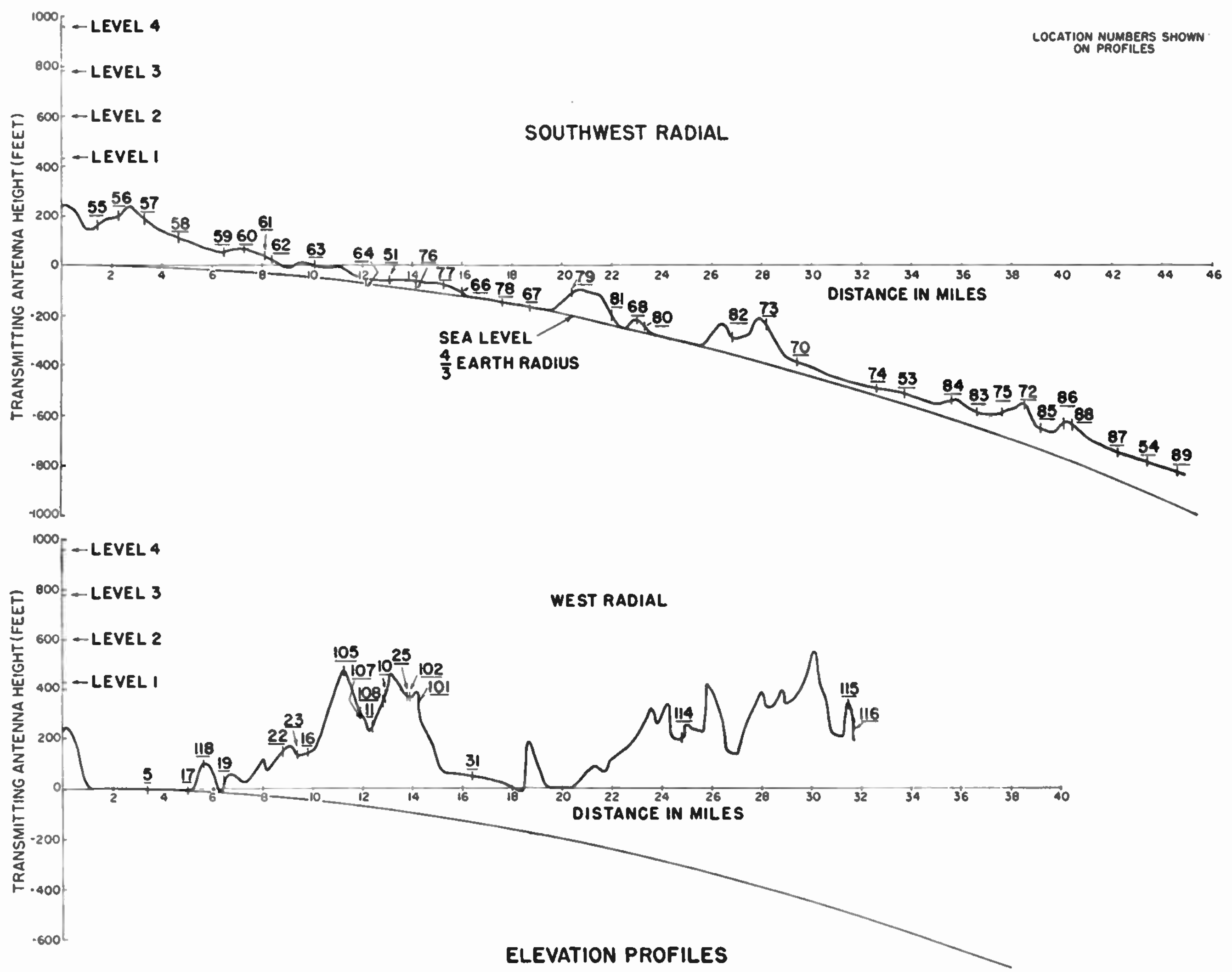


Fig. 10—A plot of the elevation profiles for the two radials shown on the map.

TABLE 1—LINE OF SIGHT LOCATIONS—SOUTHWEST RADIAL.

LOCATION	DESCRIPTION	DISTANCE MILES	FREE SPACE FIELD STRENGTH db(μv/M, KW)	LEVEL 1		LEVEL 2		LEVEL 3		LEVEL 4	
				MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db
55	C	1.5	99	101	−2	108	−9	105.5	−6.5	102.7	−3.7
56	C	2.3	95.2	87	8.2	91	4.2	95	6.2	94	1.2

Description—The letters A, B, and C are used to indicate the relative conditions of the immediate surroundings of the receiving point. A—Clear, B—Semi-clear, and C—Dense. * —Indicates that measurements were limited by sensitivity of receiver. *db(μv/m, KW)*—Decibels above one micro-volt per meter for one kilo-watt of radiated power.

TABLE 2—WEST RADIAL.

LOCATION	DESCRIPTION	DISTANCE MILES	FREE SPACE FIELD STRENGTH db(μv/M, KW)	LEVEL 1		LEVEL 2		LEVEL 3		LEVEL 4	
				MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db
5	A	3.4	92	95.7	−3.7	95	−3.0	95.7	−3.7	94.5	−2.5
17	A	5.0	88.5	86.4	2.1	86.1	2.4	87.5	1.0	87.5	1.0
18	A	5.5	88	83.1	4.9	83.1	0.9	89	−1.0	86.7	1.3
21	A	8.0	84	77	7	83.1	0.9	82	2.0	82.2	1.8
12	A	11.1	81.5	84.3	−2.8	82.5	−1.0	83.3	−1.8	83.5	−2.0
105	A	11.3	81.2	80.3	0.9	79.7	0.5	79.3	1.9	78.2	3.0

TABLE 3—CONGESTED RESIDENTIAL LOCATIONS—SOUTHWEST RADIAL.

LOCATION	DESCRIPTION	DISTANCE MILES	FREE SPACE FIELD STRENGTH db(μv/M, KW)	LEVEL 1		LEVEL 2		LEVEL 3		LEVEL 4	
				MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db
57	C	3.3	92	69	23	76	16	77.5	14.5	85.0	7
58	C	4.7	89	73	16	87	2	92.2	− 3.2	95.5	−6.5
59	C	6.5	86.2	52	34.2	62	24.2	65.9	20.3	67.7	18.5
60	C	7.3	85.2	58.6	26.6	71	14.2	75.2	10	77	8.2
61	C	8.1	84.3	51.6	32.7	55.6	28.7	63	21.6	61.7	27.6
62	C	8.8	83.6	58.6	25	58.6	25	59	24.6	59.7	23.7
63	C	10.1	82.5	56	26.5	66	16.5	69	13	69	13.5
64	C	12.1	81	54	27	60	21	61.5	19.5	59	22

TABLE 4—WEST RADIAL.

LOCATION	DESCRIPTION	DISTANCE MILES	FREE SPACE FIELD STRENGTH db(μv/M, KW)	LEVEL 1		LEVEL 2		LEVEL 3		LEVEL 4	
				MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db
20	B	7.5	85	61.2	23.8	58.6	26.4	63.9	21.1	60.1	24.9
22	C	8.8	83.5	57.7	25.8	64.2	19.3	66.3	17.2	66	17.5
16	B	9.75	82.5	42.5	40	42.9	39.6	46.4	36.1	48.7	33.8
15	C	10.3	82.2	66.1	16.1	69.1	13.1	73.6	8.6	76.2	6.0
14	C	10.7	82	61	21	60.4	21.6	61.7	20.3	62.3	19.7
13	C	11.2	81.5	61.7	19.8	59.6	21.9	61.8	19.7	62.3	19.2

TABLE 5—OPEN COUNTRY LOCATIONS—SOUTHWEST RADIAL

LOCATION	DESCRIPTION	DISTANCE MILES	FREE SPACE FIELD STRENGTH db(μv/M, KW)	LEVEL 1		LEVEL 2		LEVEL 3		LEVEL 4	
				MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db	MEASURED FIELD STRENGTH db(μv/M)	MEASURED LOSS db
76	A	14.1	79.5	67.3	12.2	74.9	4.6	80.2	−0.7	82.5	−3.0
77	A	15.3	78.7	62.1	16.6	67.4	11.3	75.1	3.6	78.3	0.4
66	B	16	78.4	63.1	15.3	64.5	13.9	67.1	11.3	70.0	8.4
78	B	17.6	77.5	53.5	24	64.1	13.4	69.2	8.3	71	6.5
67	A	18.7	77	66.9	10.1	70.7	6.3	76.2	0.8	77.8	−0.8
79	B	20.4	76	71.8	4.2	76.2	−0.2	81.2	−5.2	78.7	−2.7

TABLE 6

SHADOWED LOCATIONS—WEST RADIAL

LOCATION	DESCRIPTION	DISTANCE MILES	FREE SPACE FIELD STRENGTH db(μv/M, KW)	LEVEL 1 MEASURED FIELD STRENGTH db(μv/M)	LEVEL 1 MEASURED LOSS db	LEVEL 1 COMPUTED LOSS db	LEVEL 2 MEASURED FIELD STRENGTH db(μv/M, KW)	LEVEL 2 MEASURED LOSS db	LEVEL 2 COMPUTED LOSS db	LEVEL 3 MEASURED FIELD STRENGTH db(μv/M, KW)	LEVEL 3 MEASURED LOSS db	LEVEL 3 COMPUTED LOSS db	LEVEL 4 MEASURED FIELD STRENGTH db(μv/M, KW)	LEVEL 4 MEASURED LOSS db	LEVEL 4 COMPUTED LOSS db
19	B	6.4	86.5	58.9	27.6	24.5	65.9	20.6	20.5	70.6	15.9	18.5	73.5	13.0	12.5
23	C	9.3	83	47.1	35.9	33	49	34	29	56.6	26.4	28	56.3	26.7	27
117	A	9.9	82.5	77.1	5.4		81.8	0.7		85.6	−3.1		83	−0.5	
106	C	11.6	81	46.9	34.1	48	46.3	34.7	47	45.6	35.4	47	43.2	37.8	47
7	C	11.7	80.9	33.1	47.9	48	32.7	48.3	47.5	34.6	46.4	47	34	47.0	46
104	C	11.7	80.9	40.3	40.7	46	40.1	40.9	45.5	40.4	40.6	45.5	40.2	40.8	45
8	C	11.9	80.6	58	23.0	29	58.2	22.7	28.5	58.7	22.3	28.5	58.2	22.8	28
107	A	11.9	80.6	47.5	33.5	32	47.5	33.5	31.5	44	37	31	42.5	38.5	30
9	C	12.1	80.5	27.7	53.3	50	29	52	49.5	27	54	49.5	27	54	49
11	A	12.4	80.4	48.8	31.7	32	47.5	33	31	48.7	31.8	30	50.5	30	29
103	A	12.5	80.4	39	41.5	43	43.5	37	42	42.3	38.2	42	45	35.5	42
6	A	12.8	80.3	61.7	18.3	23	62.6	17.4	21.5	64.5	15.5	20	63.7	16.3	19
108	C	12.9	80.2	42.7	37.5	41	48.6	30.6	39.5	45.6	34.6	38	50	30.2	36
10	A	13.1	80.1	60.1	20.4	26	59.5	21	24.5	63	20.5	23	64	16.5	21
24	A	13.3	80.0	70.5	9.5	19	74.4	56	15	78.3	17	8	77.4	2.6	5
25	C	13.8	79.9	33.1	46.9	49	38.5	41.5	44	37.9	42.1	42	43	37	38
102	C	13.9	79.8	36	43.8	49	38	37.4	44	40.5	39.3	42	43	36.8	38
125	B	14.8	79.2	20*			20*			20*			20*		
32	B	15.3	79	23.2	55.8	49	25.4	53.6	48	28.9	50.1	48	30.3	48.7	48
31	B	16.4	78.5	44	34	38.5	48.1	29.9		50.9	27.1		52.7	25.3	
110	B	18.4	77.5	42.6	36.4		48.6	30.4	33.5	50.5	28.5	33.5	52.6	26.4	30.5
30	A	20.3	76.5	33.4	43.1		35.5	41.0		41	35.5		41.7	34.8	
111	A	21.4	76	49.1	26.9	31.0	53	23.0	29	56.3	19.7	25	57.3	18.7	19
29	B	22.1	75.5	25.5	50.0	37.5	28.4	47.1	35.5	33.6	41.9	31.5	36.7	33.8	27.5
112	B	23.7	75.2	51	24	25	55	20	21.5	60.3	44.7	12.5	64.9	10.1	12.5
113	C	24.5	75	20*			20*			20*			20*		
114	B	24.8	74.9	37.8	36.7	36	44	30.5	34	48.5	26	26	51.8	22.7	26
26	B	25.0	74.8	11.5	63	56.5	16.6	57.9	53.5	23.8	50.7	45.5	27.3	47.2	44.5
27	B	27	74.1	21	53	56	26.6	47.4	52	31	43	40	36.7	37.3	39
28	A	28.6	73.8	19.1	53.9	39	28.5	44.5	32	37	36	31	36.7	36.3	30
115	C	31.5	73.0	34.9	37.6	31	43.6	28.9	25	44.9	27.6	24.5	44.9	27.6	24
116	C	31.6	72.5	20*			20*			20*			20*		

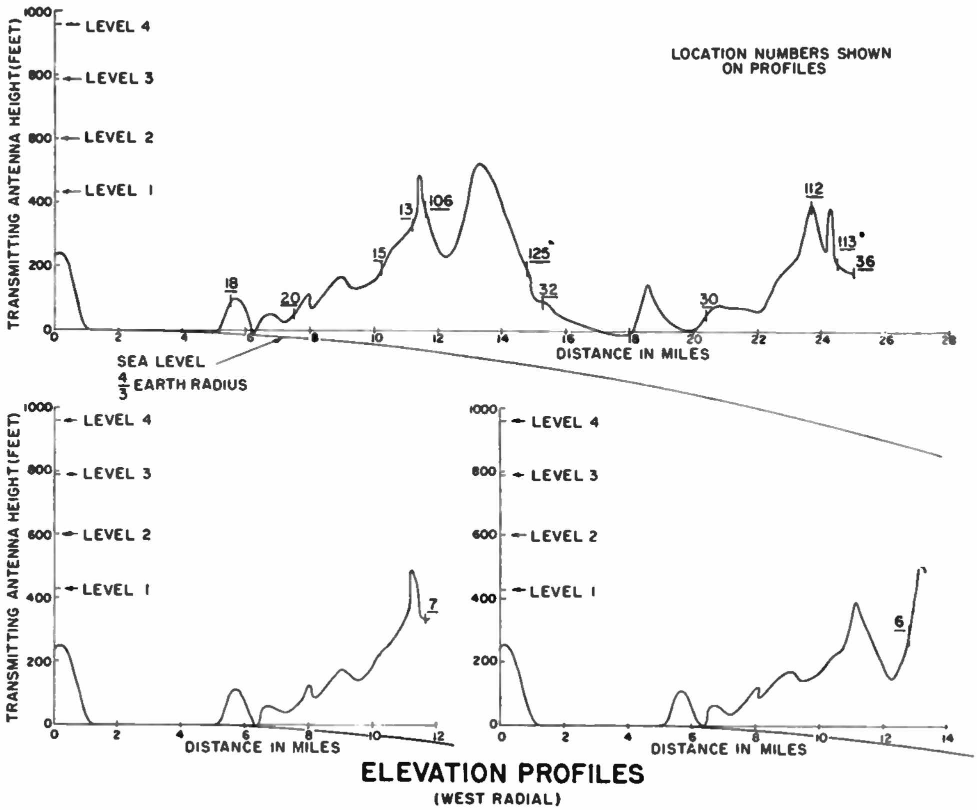
Note: No correlation between computed and measured loss for locations 117 and 130.

TABLE 7

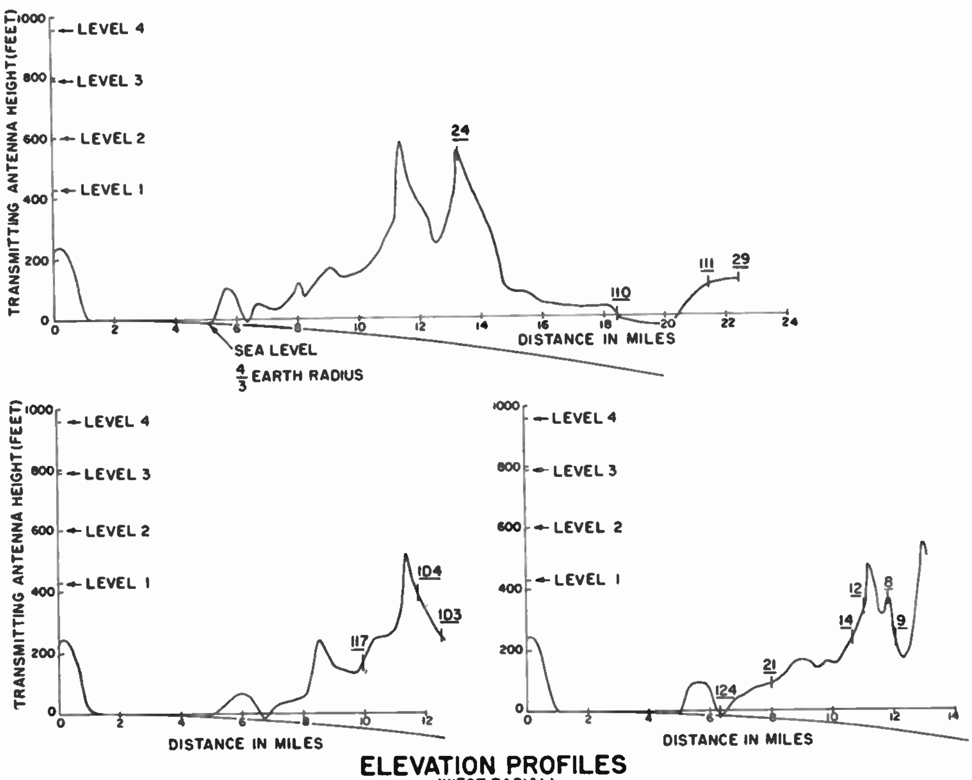
SHADOWED LOCATIONS—SOUTHWEST RADIAL

LOCATION	DESCRIPTION	DISTANCE MILES	FREE SPACE FIELD STRENGTH db(μv/M, KW)	LEVEL 1 MEASURED FIELD STRENGTH db(μv/M)	LEVEL 1 MEASURED LOSS db	LEVEL 1 COMPUTED LOSS db	LEVEL 2 MEASURED FIELD STRENGTH db(μv/M)	LEVEL 2 MEASURED LOSS db	LEVEL 2 COMPUTED LOSS db	LEVEL 3 MEASURED FIELD STRENGTH db(μv/M)	LEVEL 3 MEASURED LOSS db	LEVEL 3 COMPUTED LOSS db	LEVEL 4 MEASURED FIELD STRENGTH db(μv/M)	LEVEL 4 MEASURED LOSS db	LEVEL 4 COMPUTED LOSS db
81	B	22	75.7	63.4	12.3		64.7	11		68.6	7.1		70.6	5.1	
68	A	23	75.3	72.0	3.3		76.6	−1.3		76.5	−1.2		75.0	0.3	
80	B	23.3	75.0	54.2	20.8	27	50.7	25.3	25.5	59.7	15.3	23.5	60.1	14.9	22
82	A	26.8	74.0	71.5	3.5		73.8	0.2		79.3	−5.3		79.8	−5.8	
73	B	28.2	73.5	43	30.5	34	45.1	28.5	30	50.6	22.9	28	52.7	20.8	26
70	A	29.4	73.0	34	39	43	36.6	36.4	42	39.5	33.5	41	45.5	28.0	40
74	C	32.6	72.2	37.9	34.3	37	39.5	32.7	35.5	48	24.2	33.5	50.7	21.5	32
53	B	33.7	72.0	39	33	31	41.6	30.6	29.5	44.5	27.5	27.5	46.0	26.0	26
84	C	35.6	71.5	32.5	39	34	36	35.5	32	39.8	31.7	28	40.8	30.7	26
83	B	36.6	71.2	36.6	34.6	36.5	38.6	32.6	34.5	43.5	27.7	30.5	45.5	25.7	28.5
75	B	37.6	71	53.4	17.6	17	57.5	13.5	16	64.5	6.5	13.0	68.4	2.6	11.0
72	A	38.5	70.7	56	14.7	15	61.6	9.1	13	65.5	5.2	5	67.0	3.7	5.0
85	A	39.3	70.5	35.7	34.8	35	37.7	32.8	33	43.6	26.9	25.0	46.2	24.3	24
86	A	40.6	70.2	32.5	37.7	26	35.3	34.9	24	41.5	28.7	16.0	43.9	26.3	14
88	A	41.6	70.0	30.6	39.4	36	36.6	33.4	34	41.5	28.5	25.0	43.3	26.7	33
87	A	42.4	69.9	31.6	38.3	37	34.6	35.3	35	39.8	30.1	24.0	40.8	29.1	22
54	B	43.4	69.7	26	43.7	44	32.6	37.1	42	38.5	31.2	34.0	42.0	27.7	30
89	B	44.7	69.6	24.6	45	48	25.5	44.1	46	30.1	39.5	35.0	32.6	37	33

NOTE: No correlation between computed and measured loss for locations 81, 68, and 82.



(a)



(b)

Fig. 11—Similar profiles for radial lines which pass through the actual receiving sites.

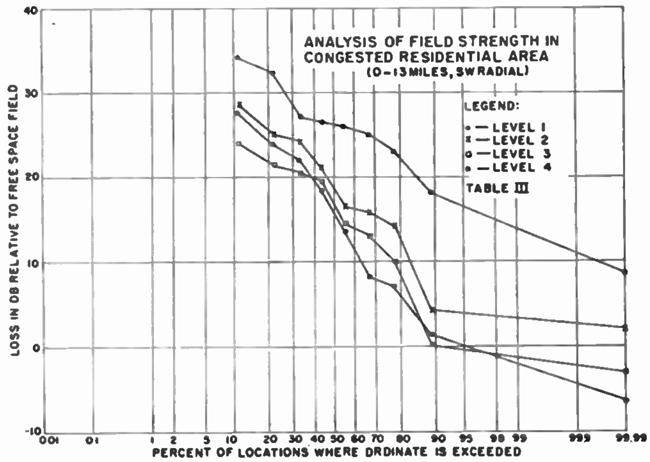


Fig. 12—A statistical analysis of a fairly homogenous sampling of measurements in a congested residential area along the southwest radial.

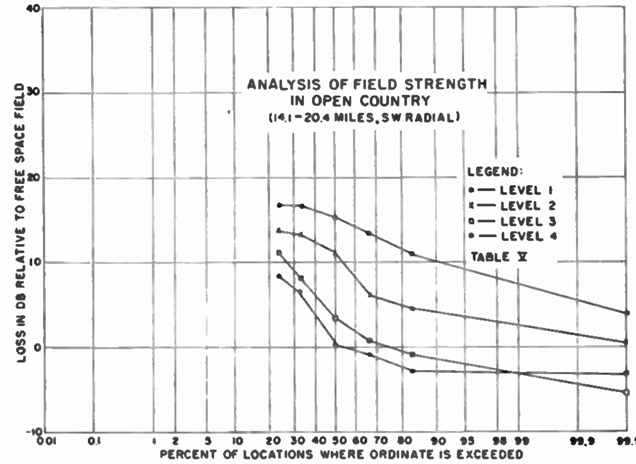


Fig. 14—A statistical analysis of the field strengths made in open country along the southwest radial.

position above the worst second order obstruction, namely the closest house. Maximum measured field strengths for receiving antenna heights between 10 and 30 feet were used in an analysis of the data shown in Fig. 12. The loss, which is plotted relative to free space field strength, may be seen to be an inverse function of transmitting antenna height. (Remember that level 4 denotes the highest antenna.) An analysis of similar locations along the west radial is shown in Fig. 13.

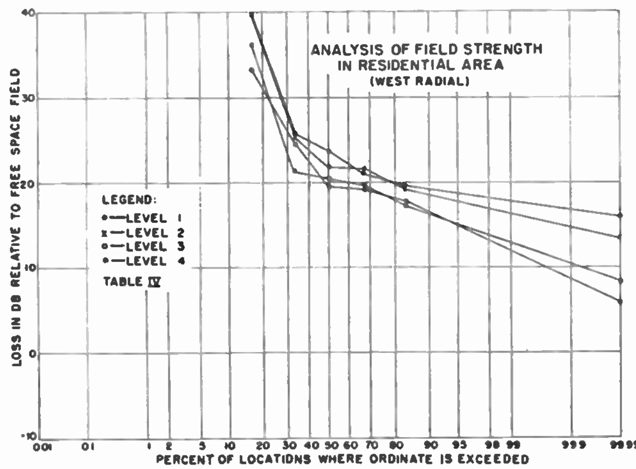


Fig. 13—A statistical analysis of residential area measurements made along the west radial.

Another reasonably homogeneous sampling of measurements was obtained in open country, again with negligible second order roughness. A similar analysis, made for these data, is shown in Fig. 14. This area was farm country with open fields and occasional wooded areas. Second order roughness was about the minimum that can be expected for practical receiving sites. Only scattered trees and very few buildings were visible.

From the analyses of Figs. 12 and 14 the "experience factor" curves of Fig. 15 were derived. These consist of median losses with respect to free-space field strength plotted as a function of transmitting antenna height.

No attempt was made to obtain "experience factors" for other possible second-order roughness conditions. For example, large apartment and office building areas could be expected to yield quite different data. There is also the possibility that other distance ranges would yield different "experience factors." This appears plausible since the amount of attenuation obtained at any site should depend upon the angle of approach at receiving antenna. This is the angle between a ray from the transmitting to receiving antenna and a tangent to the earth's surface at receiving site. Attenuation for low

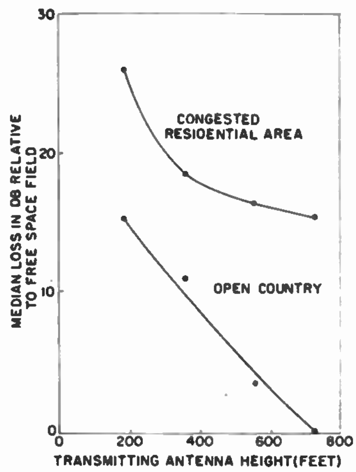


Fig. 15—A plot of the median losses obtained from Figs. 12 and 14. These will be referred to as "experience factors" and are used in prediction analysis.

angles of approach should be greater than for high angles since the path length through intervening obstacles is longer. On this basis the "experience factor" curves present the average effect of a range of angles of approach. It is believed that the data will apply to a large percentage of cases encountered.

Shadowed Area Propagation

Many shadowed receiving sites along both radials were included in the field strength measurements. These were grouped for analysis. The loss, referred to free space field strength for west radial shadowed sites

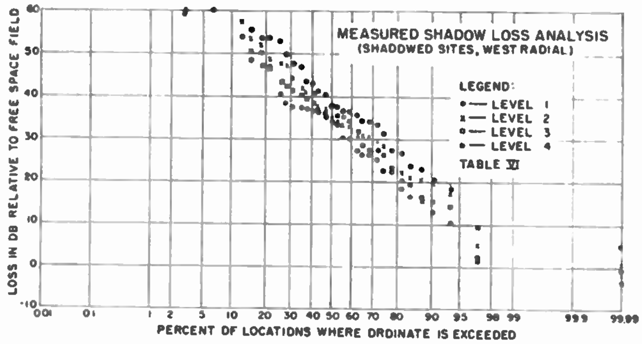


Fig. 16—A statistical analysis of the measured field strengths for shadowed receiving sites along the west radial.

is shown in Fig. 16 where it can be seen that transmitting antenna height was a relatively small factor. The shadow loss theory predicts that for deep shadowing there will be less dependence on transmitting antenna height than for light shadowing. With this in mind, the shadowed sites for the southwest radial were analyzed separately because the hills were smaller than west radial hills. The analysis of Fig. 17 does show more dependence on transmitting antenna height than the analysis of Fig. 16 which included many examples of deep shadowing.

Field strength prediction for each shadowed site was made with methods to be described under Field Strength Prediction. A comparison of predicted and measured fields for all shadowed area data is given in Fig. 18 in a prediction error analysis. This analysis indicates that the effect of transmitting antenna height has been quite well accounted for in the prediction, since the points for the four transmitting heights are grouped closely. Based on the Fig. 18 analysis for all shadowed area data, an "expected prediction error" curve was derived, Fig. 19. This shows the error of prediction as a function of the per cent of the measurements.

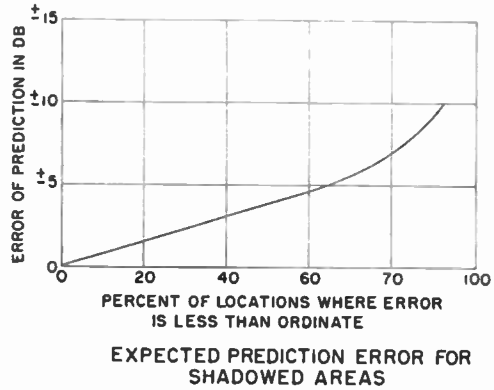


Fig. 19—The functional dependence of the error of prediction on the per cent of the number of measurements.

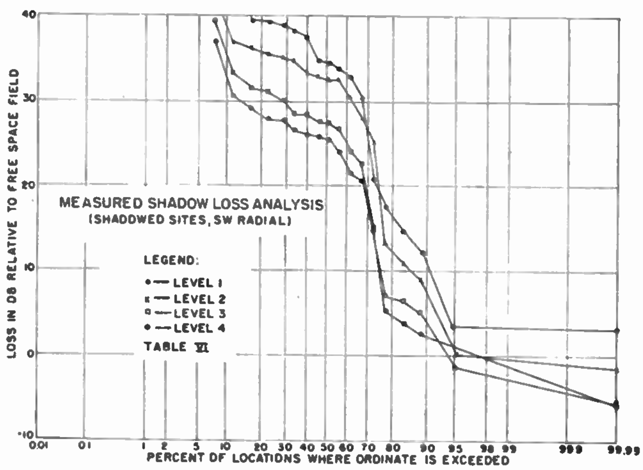


Fig. 17—A statistical analysis of the measured field strengths for shadowed receiving sites along the south-west radial.

The prediction error for shadowed sites along the Southwest radial is shown in Fig. 20. Comparison with Fig. 18 shows the prediction with moderate shadowing was as good as that including all shadow data.

Line-of-Sight Propagation

A selection of sites which topographic maps showed lacked first order obstructions was made along both radial lines. Visual examinations showed the sites to be without second order roughness so that direct-ray paths must have been completely line-of-sight. A tabulation of these data, Tables 1 and 2, shows values close to free space field strength. Included are sites in swamps, on hill sides and hill tops. In several cases visibility was not good enough to confirm that true line-of-sight existed.

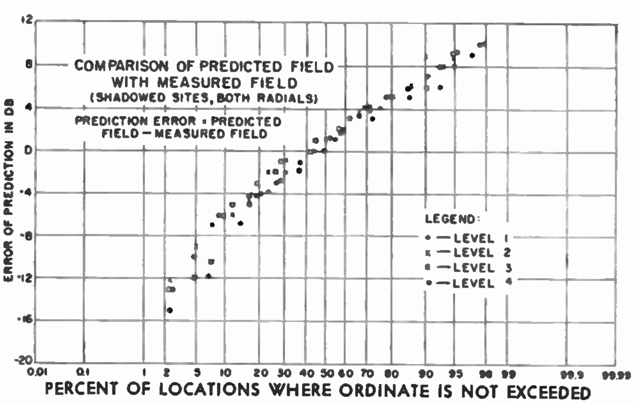


Fig. 18—A statistical analysis of the error of prediction for all shadowed receiving sites.

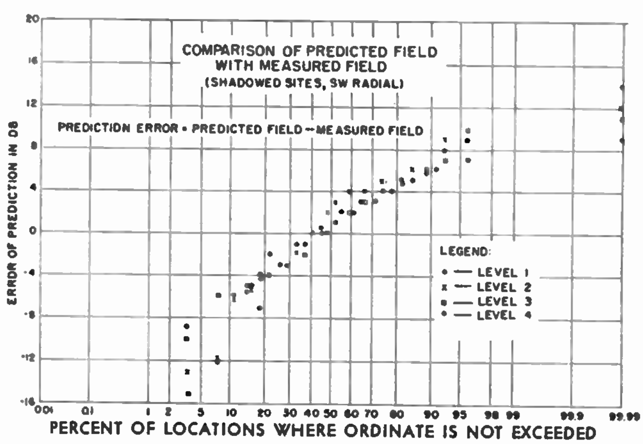


Fig. 20—A statistical analysis of the error of prediction for shadowed sites along the southwest radial.

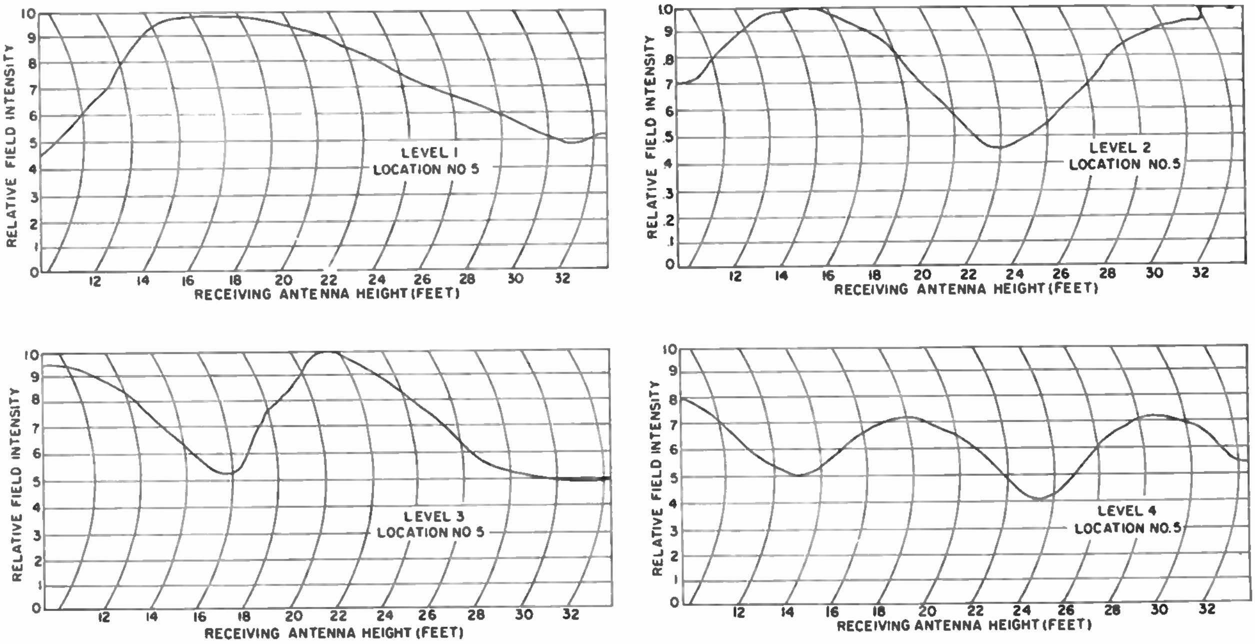


Fig. 21—Field strength recordings indicating existence of an interference pattern.

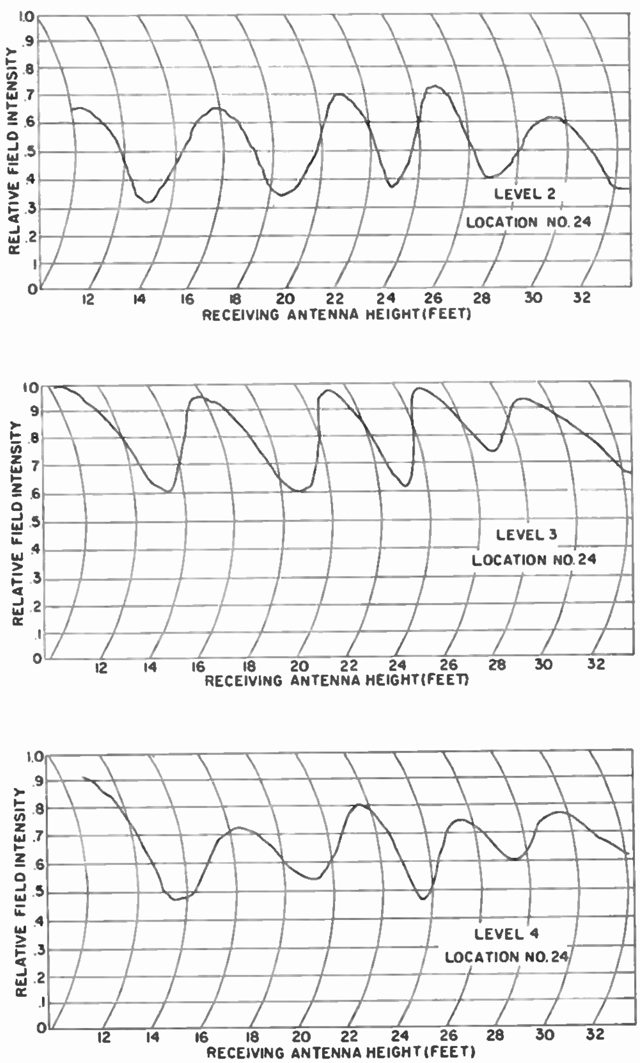


Fig. 22—Field strength versus height recordings made on Hill 1 (Fig. 23).

Surface Roughness Effects

There is experimental evidence from WOR tower measurements over swamp land on the west radial that 850-mc propagation can occur in nature in the manner described by the relatively simple smooth plane earth theory. Field strengths up to several db above the free space theoretical field were obtained. Most of the 3.5 mile swamp path was covered with about 6-foot grass and yielded interference patterns, Fig. 21, for the four transmitting antenna heights which bear marked resemblance to smooth, plane earth interference patterns. The nulls are not deep, indicating an effectively low reflection coefficient. There was no very significant difference (± 1 db) in maximum values for the four levels and the simple theory indicates none. The field strength at a fixed receiving antenna height varied about ± 1 db as the antenna was moved about. Two other facts are also noteworthy. Recordings of vertical transmitting antenna patterns at the swamp location looked like free space patterns when recorded on either maxima or minima of the interference patterns. Transmitting antenna maxima, when aimed by vertical tilting toward the receiving site, yielded maximum signal.

Interference patterns were observed beyond and atop hills. These patterns confirm existence of propagation as illustrated in Fig. 4. The interference patterns of Fig. 22 were recorded on a hilltop at 13.3 miles on the west radial. There was very little second order roughness for several hundred feet down the slope of the hill along the propagation path. The four transmitting antenna heights yielded patterns so similar that transmitting antenna height was obviously not a factor in determining the field strength versus height relationship, although it may have affected the magnitude of the maxima. The geometry along the propagation path was such that the earth reflection must have occurred within a few hundred feet of the receiving site, hill 3, Fig. 23. The slope of the hill made it possible for the incident and reflection angles of the reflected ray to be equal at a point much closer to the receiving site than would have been possible for a horizontal plane reflector. It is important to note that, in the absence of second order roughness, an earth reflection from the valley bottom could have produced an interference pattern at this same site. However, the evaluation of this factor for receiving sites is difficult, since the area responsible for the earth reflection is not always easily determined. Fortunately, the effect on propagation in comparison to shadow and other losses is usually negligible.

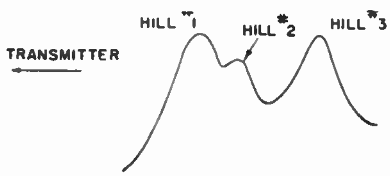


Fig. 23—Profile along west radial.

Another field distribution recording (Fig. 24) on hill 2 (Fig. 23) at 11.9 miles shows field strength along the radial line at a constant 10-foot height above a street.

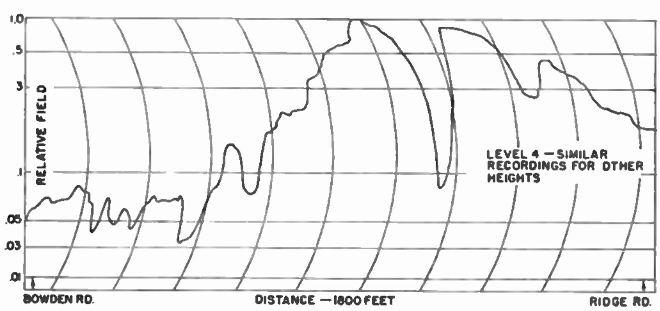


Fig. 24—Relative field strength on Hill 2 (Fig. 23) along a radial line at a constant 10-foot height above the street.

A rather deep null occurred on the hill top and again there was marked similarity of the recordings for all transmitting antenna heights. In Fig. 25, field strength versus height recordings on the hill top (at the location of the above-noted null) are seen to be similar to recordings from hill 3.

Hill number 1 (Fig. 23) also exhibited a vertical field strength distribution resembling an interference pattern. Again there was relatively little second order roughness along several hundred feet of propagation path.

Numerous recordings of field strength versus height, Fig. 26, throughout the valley between hills 1 and 3 showed distributions which seem to include interference pattern effects. These examples, by their irregularities, show evidence of second order roughness effects. Recordings were made in open areas to minimize second order roughness effects as much as possible.

An interesting example of a seasonal change of field distribution from a third order roughness effect was found adjoining a corn field. The field strength versus height relations before and after harvesting the corn are shown in Fig. 27. An observation of this kind was fortuitous since it suggests it may not be rare.

Two quite distinct reflection effects from objects beyond the receiving site are important. By far the most common is reflected signal from near-by buildings. This kind of reflected signal is almost always prominent in residential areas. TV picture observations in residential areas confirm the fact that such reflected signals usually involve short enough path differences as to not seriously impair the picture. These reflected signals are usually quite constant with time although seasonal variations of signal strength may result from nearby reflections if foliage is involved as either reflector or obstacle. Standing wave patterns from nearby reflecting objects can often be recognized. Reflected signals from hills beyond the receiving site also occur. Such reflected signals have been observed where direct signals were very low because of shadowing loss. There was quite rapid variation with time and considerable vertical polarization although the signal source was horizontally polarized. The hills involved were wooded. Field strength recordings along the radial line in the presence of this kind of reflected signal showed no standing wave where the average magnitude of reflected signal equaled the direct.

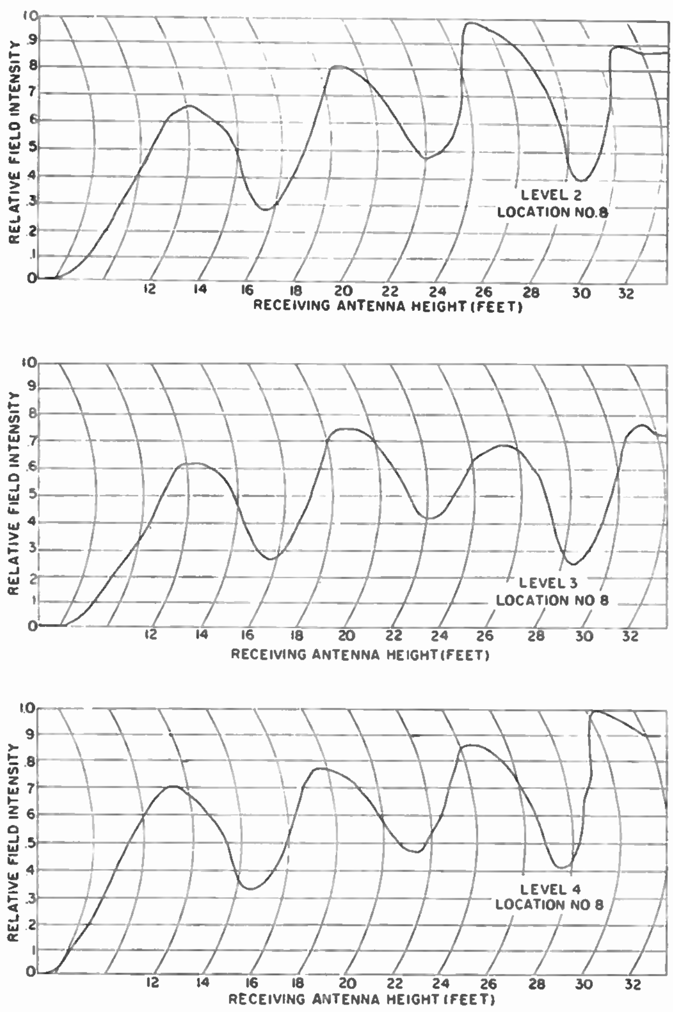


Fig. 25—Field strength versus height recordings at the location of null (Fig. 24).

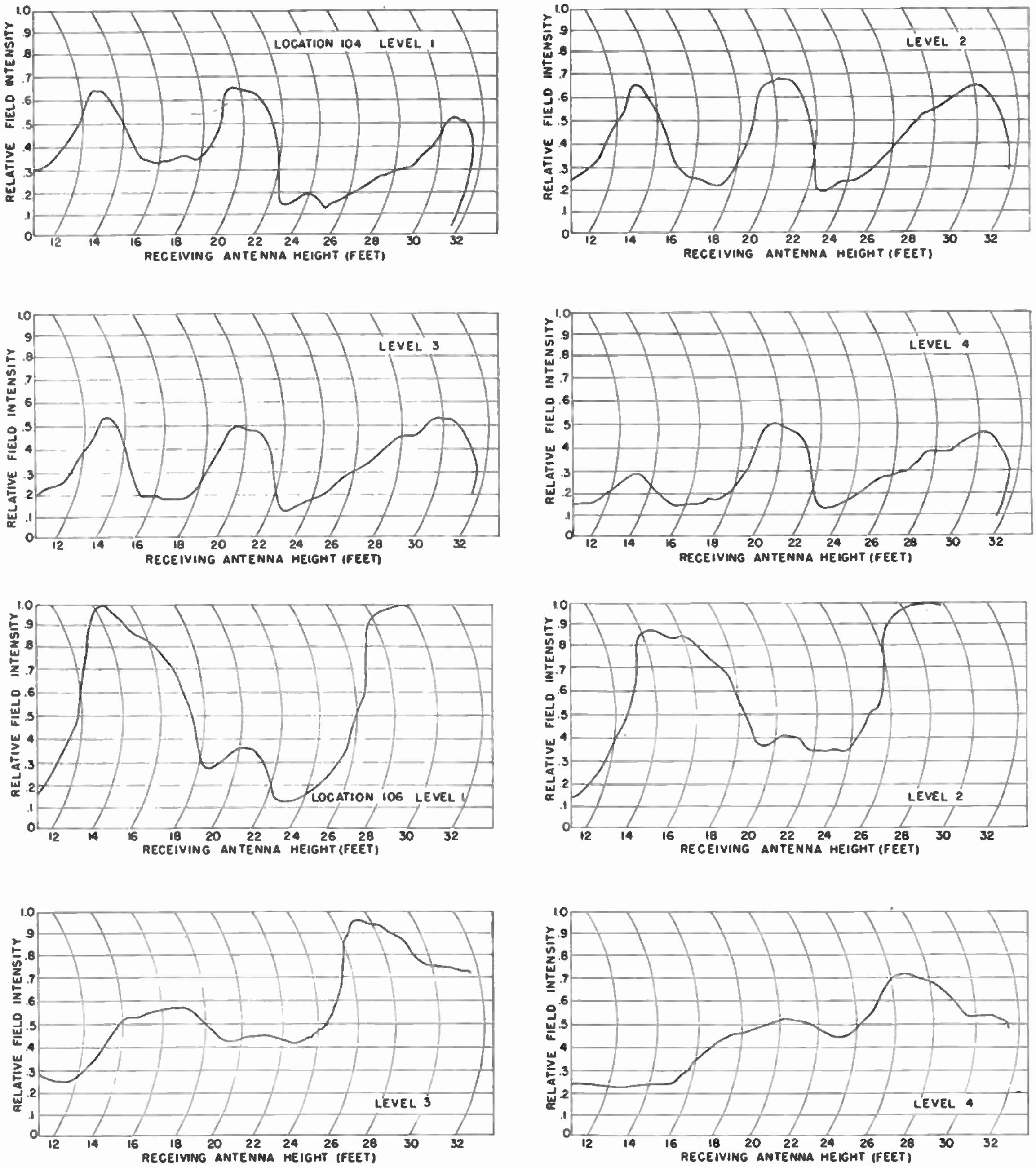


Fig. 26—Field strength recordings versus height indicating presence of interference effects.

The relative importance of the various propagation effects described cannot be accurately assessed. They have been described because they are believed to be components of a pattern which prevails for uhf propagation as pictured for TV broadcasting. So many factors are important that it seems safe to say that no terrain or surface condition can be described as *typical.*

It is, therefore, evident that field strength prediction for uhf TV broadcasting can hardly be either a precise science or a fine art. It can be greatly refined as experience is accumulated. Theoretical methods evolved for point-to-point communication can sometimes be used for first order roughness effects. Useful "experience factors" can be procured for second order roughness effects. In some cases reflection coefficient measurements may be desirable to determine the magnitude of third order roughness effects where examination of terrain indicates possible important earth reflection.

Field Strength Prediction

The first object of a coverage prediction is to aid in choosing the location for the transmitting antenna, both as to geographical location and height above terrain. Intimately involved in this choice are questions of antenna gain, horizontal pattern shape and vertical pattern shape. These are engineering aspects of what must be fundamentally an economic problem.

What are the preliminary facts which will be required to make a useful prediction of coverage for a uhf TV station? The best topographic information obtainable in the form of Geological Survey maps is the first requirement.[5] Three dimensional relief maps, when available, are also valuable.[6] Statistics on population distribution should be procured. Preferably, these data should give a dynamic picture of population which shows the trends so that an informed guess about future requirements can be made. A visual survey of surface conditions should be made to obtain a complete descriptive picture of second and third order roughness.

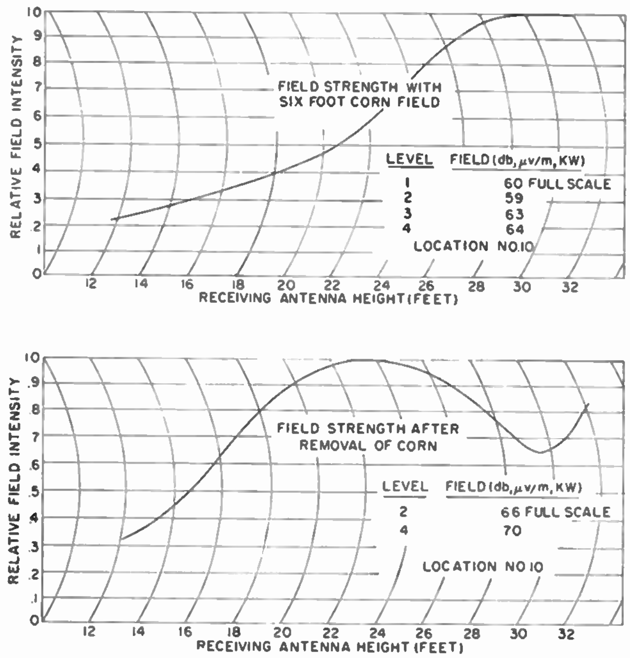


Fig. 27—Field strength recordings which indicate the effect of a field of corn.

Estimated median field strength curves for lines on the topographic map will be basic to the field strength prediction. Tentative choices of transmitting antenna location will be made after studying the topography and population distribution. In all probability there will be more than one likely location suggested by topography and population distribution. A relief map and a tiny light source[7] placed at possible locations will give rough indication of shadowed areas and materially assist in making good choices. Usually a few strategically chosen segments of radial lines for each choice of location will, upon examination of field strength predictions, make it possible to decide on the best choice.

After the transmitting antenna location has been chosen, a complete coverage prediction can be made where warranted. Careful choice of radial lines can minimize the labor required. The median field strength data will conveniently be presented as constant median field strength contours if the topography is relatively simple. The field strength exceeded in 50 per cent of the area within a contour, or the median, may not be the only desired picture of coverage. Field strengths exceeded in other percentages can be roughly estimated from the statistical analysis curves offered.

The field strength prediction will require the theoretical relations of Figs. 1, 28 and 29 and the "experience factor" curves of Fig. 15. Fig. 1 is the free space field strength used as a reference. Fig. 28 is a nomogram[8] from which to obtain shadow loss. Fig. 29 is a calculating chart for smooth plane earth theoretical field strength.

A sizable measure of good judgment will be invaluable in the use of theoretical relations and experience data for prediction. Guides rather than hard and fast rules are all that we can offer.

The shadow loss prediction is based on theory for a perfectly absorbing knife edge. Quite obviously ordinary hills are anything but knife edges, but experiments have shown that the theory provides a good guess. The nomogram of Fig. 28 can be applied to a profile to pre-

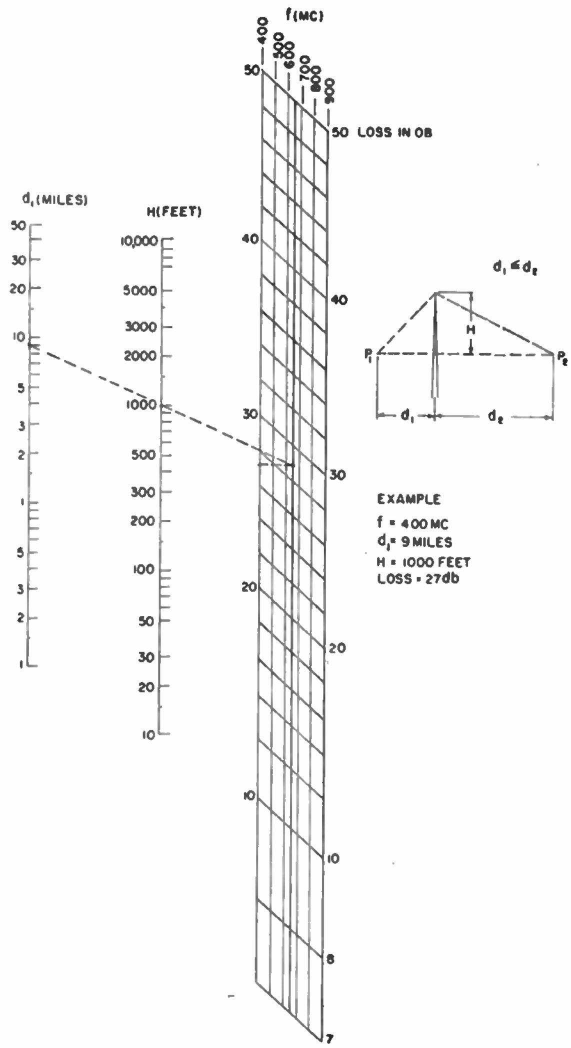


Fig. 28—A nomogram for computing the shadow loss due to a perfectly absorbing knife edge relative to free space field.

[5] A free folder describing topographic maps of the United States Geological Survey can be obtained from Geological Survey, Washington 25, D. C. Specify the area of interest when ordering.

[6] John Taylor, "UHF in Portland," *Broadcast News*, Special Edition, October 1952.

[7] A tiny surgical lamp, the "Grain of Wheat" bulb, can be obtained from George P. Tilling and Son Co., 3451 Walnut Street, Philadelphia, Pa.

[8] This nomogram is similar to one published by K. Bullington in the Proc. I.R.E., January 1950.

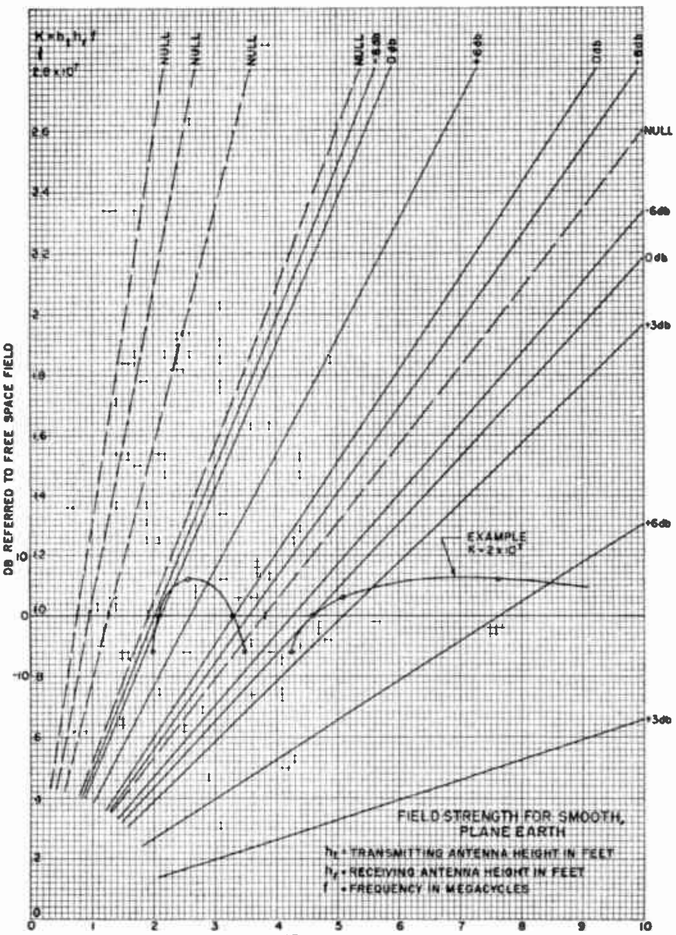


Fig. 29—A calculating chart for smooth-plane-earth theoretical field strength.

dict shadow loss between two points P_1 and P_2. H is the height of the shadowing obstruction above a line joining the points. By making $d_1 \leqq d_2$ the nomogram gives theoretical shadow loss that is in error less than ± 1.5 db. This error comes from an approximation used in constructing the nomogram.

If there are successive shadowing hills, the shadow loss for each hill can be estimated as shown in Fig. 30. First draw line P_1P_2 to obtain H_1 and the appropriate

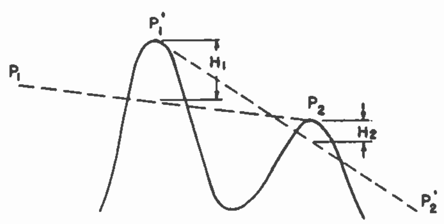


Fig. 30—Application of shadow loss to successive hills.

distances for the loss of Hill 1; then draw line $P_1'P_2'$ to obtain H_2 and the appropriate distances for the loss of Hill 2. The sum of the losses so obtained will be used for location P_2' with transmission from P_1.

Sometimes it will be necessary to decide whether or not surface roughness will permit an earth reflection. A useful approximate criterion, known as Rayleigh's criterion of roughness, can be borrowed from physical optics. If $h \sin \theta_i < \lambda/8$, Fig. 31, the surface will be considered to be smooth. It will be seen that the smaller the incidence angle, θ_i, the greater the height of roughness features, h, can be and the surface still be considered smooth. If the roughness criterion indicates "smooth," this means only that the surface *may* be smooth, a condition which can be verified only by field strength measurement. Because uhf wavelengths are very small compared with second order roughness di-

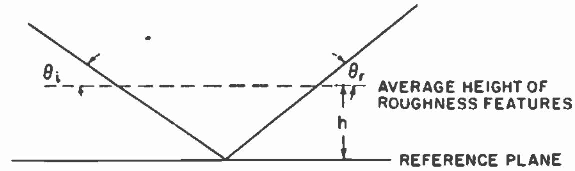


Fig. 31—A diagram showing the parameters involved in estimating Rayleigh's criterion roughness.

mensions, the presence of such features along the path will usually mean a "rough" surface and no earth reflection within the primary city area.

We will describe a prediction procedure for (a) a primary city, (b) a surrounding rural area, and (c) a secondary city. The transmitting antenna will be assumed to be not over 1000 feet above surrounding terrain and the distance will be limited to 30 or 40 miles. Median field strengths will be derived. Time variation will be assumed non-existent although this may be an invalid assumption for highly shadowed areas unless predictions are considered to be time medians.

(a) Most commonly the primary city area to be covered will have second order roughness over the entire length of radial lines to be examined. When this is true the prediction for a radial line will start with free space field strength of Fig. 1 as a reference. Along non-shadowed segments of the line the prediction field will be free space field strength reduced by the "experience factor" for congested residential area of Fig. 15. If a ray drawn from transmitting antenna to receiving site shows shadowing, the prediction field will be the free space field reduced by an "experience factor" and also reduced by the shadow loss of Fig. 28.

One should use judgment in applying the "experience factors." It is believed that most of the data serving as the basis for the "experience factor" curve for congested residential areas represent partly obstructed propagation even for the highest transmitting antenna. This comes about from the random building sizes found along the radial. With favorable terrain and more uniform building sizes the "experience factor" could be much lower. Also predictions using "experience factors" may be pessimistic for the lower uhf channels.

If the primary city area includes lakes or other large, flat areas unobstructed by second order roughness it will be desirable to determine if any important populated areas are in nulls of a possible interference pattern.

The field intensity over plane, smooth earth is

$$E = \frac{275,200}{d} \sin\left(\frac{6.92 \times 10^{-6}\, h_t h_r f}{d}\right)$$

where

h_t = height of transmitting antenna, ft.

d = distance between transmitting and receiving antennas, miles

h_r = height of receiving antenna, ft.

E = field intensity, μv/m, KW E.R.P.

f = frequency, mc.

The values of d for which E is zero (the distances of the nulls) can be found by equating

$$\frac{6.92 \times 10^{-5} h_t h_r f}{d} = 180°, 360°, 540°, \text{etc.}$$

If it appears likely that nulls will occur in undesirable places, this can usually be obviated by changing h_t or selecting another transmitting location. The entire field strength versus distance relationship for plane earth can be quickly obtained from Fig. 29. To use this chart, calculate K, which is the product of transmitting antenna height (feet), receiving antenna height (feet), and frequency (mc). Draw a horizontal line at the ordinate K and a Db ordinate scale as shown in the example. The field strength, referred to free space field, can be determined from the intersections of the sloping lines and the horizontal line.

(b) Rural coverage can be predicted by using free space field reduced by the open country "experience factor" of Fig. 15 where there are no first order obstructions.

Shadowed areas will have field strength reduced from the free space field by the shadow loss of Fig. 28. Experience indicates that the second order roughness loss will be proportional to the length the ray path traverses while in close proximity to second order roughness, from the obstruction top to the receiving site. The maximum value, obtained by extrapolating the open country "experience factor" to zero height, is about 20 db. This will range down to a minimum which is seldom below 5 db. The profile will enable one to determine how much of the ray path lies close to second order roughness.

(c) The case of secondary city coverage will require examination of intervening topography for prominent hills or ridges. A shadowing feature can usually be expected to reduce field strength from the free space value by the shadow loss of Fig. 28. The "experience factor" loss will tend toward the higher values, because of greater distances, than for the primary city. The improvement from transmitting antenna height that resulted from greater clearance with height over second order roughness probably diminishes as the ray approaching a receiving site becomes more nearly coincident with a tangent to the earth's surface.

A flat, smooth area between a shadowing obstruction and the secondary city calls for examination of the geometry for possible earth reflections. When their existence seems likely, the possible effects of such reflection can be anticipated from the plane earth field of Fig. 29. Third order roughness will determine the actual existence of such an earth reflection.

There is also the rather unlikely possibility of effects from earth reflecting surfaces before an obstruction. It appears prudent in such cases to rely on measurement rather than prediction, since this kind of earth reflection can theoretically account for field strengths which range between zero and twice the free space value.

Conclusion

It has been well stated by the Joint Technical Advisory Committee[9] that, "There are recurrent requests from well-meaning planners or users of radio frequencies that the outstanding facts of radio propagation be summarized in a brief statement of, say, one page or a few judicious sentences. This is impossible; nature did not so arrange the facts." We have examined some of the facts and have offered some interpretations.

A field survey has been described which involves the effects of wave refraction, earth reflection, diffraction and attenuation. The data have been analyzed to show the influence of these effects on the median field strength along two radial lines. Both measurement and theory show that the shadow loss for deep shadowing will be less dependent on transmitting height than for light shadowing. Experience factors have been obtained which bracket the loss introduced by houses and trees. A prediction of the field strength for the two radial lines has been made. Analysis of the prediction shows the amount of prediction error. Reduction of this error will depend upon refinement in the theory of shadow loss and the accumulation of more experience data. The results indicate that a useful degree of accuracy can be accomplished with the methods described in predicting the service area of a uhf broadcasting station.

Acknowledgments

In planning the field tests described in this paper, it soon became evident that the WOR tower and location at North Bergen, New Jersey, were ideally suited to our purposes. From the time we first approached the officials of The Mutual Broadcasting System with our tentative plans until the project was completed with the dismantling of the last antenna, we have enjoyed the complete and generous cooperation of the management and the operating staff of WOR.

Appendix I—Method of Determining Field Strength

The procedure for determining the field strength involved the use of a uhf receiver, a receiving antenna of known power gain with respect to a $\lambda/2$ dipole, and a signal generator of known input impedance. The block diagram of this setup is shown in Fig. 32(a). This consists of a calibrated signal generator having an input impedance of 75Ω, a corner-reflecting receiving antenna of known power gain and internal impedance of 75Ω, and a receiver. The procedure for determining the field

[9] "Radio Spectrum Conservation"—Joint Technical Advisory Committee, IRE-RTMA.

strength is first to tune the receiver to the incoming signal, and then obtain an identical receiver reading from the signal generator. For the conditions outlined the relation between the signal generator voltage and field strength is given by

$$E = \frac{\pi V}{\lambda\sqrt{G}} \tag{1}$$

where

E = Field strength in volts per meter
λ = Wavelength in meters
G = Gain of receiving antenna with respect to a $\lambda/2$ dipole
V = Internal voltage of signal generator.

Generally V must be determined from the known characteristics of a calibrated signal generator. Suppose the internal impedance of the signal generator is R_s and we wish to convert this to a generator having an impedance of R ohms by means of a perfect transformer. The equivalent diagram is shown in Fig. 32(b).

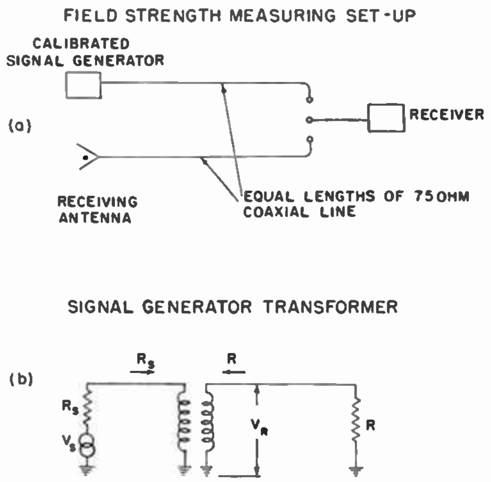


Fig. 32(a)—Block diagram illustrating calibration procedure. (b)—Equivalent circuit diagram of signal generator matching transformer.

Examining the power relations we derive the following equation

$$\frac{V_s^2}{4R_s} = \frac{V_R^2}{R},$$

so that

$$V_R = \frac{1}{2}\sqrt{\frac{R}{R_s}}\,V_s.$$

Hence the internal generated voltage of the modified generator, having an internal impedance of R ohms, is equal to 2 V_R volts. Substituting this for V in equation (1), we have

$$E = \frac{\pi}{\lambda}\sqrt{\frac{R}{R_s}}\,\frac{V_s}{\sqrt{G}}.$$

Appendix 2. Fresnel Diffraction at a Knife Edge

In computing the shadow losses due to hills we have assumed that the hill could be represented by a perfectly absorbing knife edge. Although this is a very rough approximation it does enable us to use the results that have been developed for the knife edge case in optics[10] and permits the gross evaluation of the effect of the variables involved. The geometry for the assumed problem is shown in Fig. 33. The ratio of the field

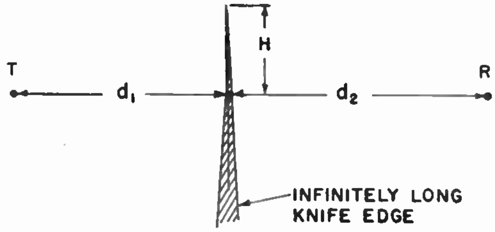


Fig. 33—Geometrical figure illustrating parameters involved in knife edge diffraction.

strength E at the receiving point R to its free-space value is given by the standard Fresnel integral

$$\frac{E}{E_0} = \frac{(1+j)}{2}\int_{v_0}^{\infty} e^{-j(\pi/2)v^2}dv,$$

where

$$v_0 = H\sqrt{\frac{2(d_1+d_2)}{\lambda d_1 d_2}}$$

$$d_1,\ d_2 \gg H$$

$$d_1,\ d_2 \gg \lambda.$$

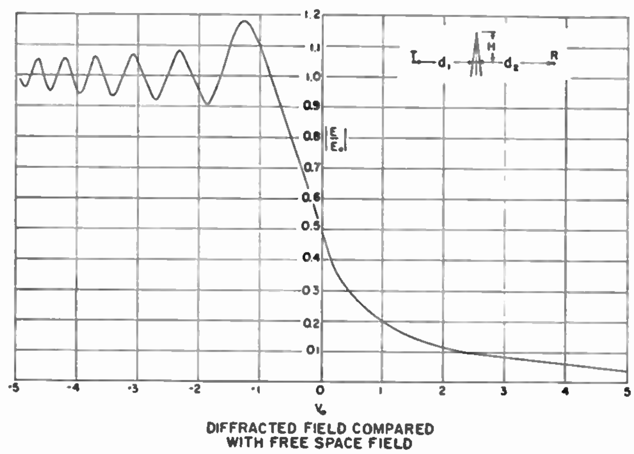


Fig. 34—Diffracted field as a function of v_0. (v_0 is a function of geometry and wavelength).

A plot of this ratio as a function of v_0 is shown in Fig. 34. A nomogram which shows the effect of varying any of the parameters has been shown in Fig. 28.

[10] E. C. Jordan, "Electro-magnetic waves and radiating systems," p. 572. Prentice-Hall, New York, N.Y.

Standards on Modulation Systems: Definitions of Terms, 1953*

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Amplitude Modulation (AM). Modulation in which the amplitude of a carrier is the characteristic varied.

Angle Modulation. Modulation in which the angle of a sine-wave carrier is the characteristic varied from its normal value.

Note—Phase and frequency modulation are particular forms of angle modulation.

Angle or Phase of a Sine Wave. The measure of the progression of the wave in time or space from a chosen instant or position.

Note 1—In the expression for a sine wave, the angle or phase is the value of the entire argument of the sine function.

Note 2—In the representation of a sine wave by a rotating vector, the angle or phase is the angle through which the vector has progressed.

Baud. A unit of signalling speed. The speed in bauds is the number of code elements per second.

Carrier. A wave suitable for being modulated.

Note—Examples of carriers are a sine wave, a recurring series of pulses, or a direct current.

Carrier Frequency. In a periodic carrier, the reciprocal of its period.

Note—The frequency of a periodic pulse carrier often is called the pulse-repetition frequency (prf).

Carrier-to-Noise Ratio. The ratio of the magnitude of the carrier to that of the noise after selection and before any nonlinear process such as amplitude limiting and detection.

Note—This ratio is expressed many different ways, for example, in terms of peak values in the case of impulse noise and in terms of root-mean-square values in the case of random noise. In special cases other measures of carrier and noise may be used, but their use should be clearly stated.

Companding. A process in which compression is followed by expansion.

Note—Companding is often used for noise reduction, in which case the compression is applied before the noise exposure and the expansion after the exposure.

Compression. A process in which the effective gain applied to a signal is varied as a function of the signal magnitude, the effective gain being greater for small than for large signals.

De-emphasis. A process complementary to pre-emphasis.

* Reprints of this Standard, 53 IRE 11.S1, may be purchased while available from The Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., at $0.50 per copy. A 20-per cent discount will be allowed for 100 or more copies mailed to one address.

Demodulation. The process of recovering the modulating wave from a modulated carrier.

Detection. See Demodulation.

Detector (in Receivers). (a) A device to effect the process of detection. (b) A mixer in a superheterodyne receiver.

Deviation Ratio. In a frequency-modulation system, the ratio of the maximum design frequency deviation to the maximum design modulating frequency of the system.

Differentiator (Differentiating Circuit, Differentiating Network). A transducer whose output wave form is the time derivative of its input wave form.

Note—Such a network preceding a frequency modulator makes the combination a phase-modulation modulator; or, following a phase-modulation detector, makes the combination a frequency-modulation detector. The ratio of output amplitude to input amplitude of a differentiator is proportional to frequency, and the output phase leads the input phase by 90°.

Discriminator. A device in which amplitude variations are derived in response to frequency variations.

Expansion. A process in which the effective gain applied to a signal is varied as a function of the signal magnitude, the effective gain being greater for large than small signals.

Frequency Deviation. In frequency modulation, the peak difference between the instantaneous frequency of the modulated wave and the carrier frequency.

Frequency-Division Multiplex. A device or process for the transmission of two or more signals over a common path by using a different frequency band for each signal.

Frequency Modulation (FM). Angle modulation of a sine-wave carrier in which the instantaneous frequency of the modulated wave differs from the carrier frequency by an amount proportional to the instantaneous value of the modulating wave.

Note—Combinations of phase and frequency modulation are commonly referred to as "frequency modulation."

Frequency-Shift Keying or FSK. That form of frequency modulation in which the modulating wave shifts the output frequency between predetermined values corresponding to the frequencies of correlated sources.

Frequency Swing. In frequency modulation, the difference between the maximum and minimum design values of the instantaneous frequency.

Gating. The process of selecting those portions of a wave which exist during one or more selected time intervals or which have magnitudes between selected limits.

Improvement Threshold. That value of carrier-to-noise ratio below which the signal-to-noise ratio decreases more rapidly than the carrier-to-noise ratio.

Instantaneous Companding. Companding in which the effective gain variations are made in response to instantaneous values of the signal wave.

Instantaneous Frequency. The time rate of change of the angle of an angle-modulated wave.

Instantaneous Sampling. The process for obtaining a sequence of instantaneous values of a wave. These values are called "instantaneous samples."

Integrating Network (Integrating Circuit, Integrator). A transducer whose output waveform is the time integral of its input waveform.

Note—Such a network preceding a phase modulator makes the combination a frequency modulator; or, following a frequency-modulation detector, makes the combination a phase-modulation detector. The ratio of output amplitude to input amplitude of an integrator is inversely proportional to frequency, and the output phase lags the input phase by 90°.

Intermediate Subcarrier. A carrier which may be modulated by one or more subcarriers and which is used as a modulating wave to modulate another carrier.

Inverse Limiter. A transducer, the output of which is constant for input of instantaneous values within a specified range and a linear or other prescribed function of the input for inputs above and below that range.

Note—This term describes a device used generally to remove the low-level portions of signals from an output wave. It is sometimes used to eliminate the annoying effects of cross talk in a system at the expense of some distortion.

Keying. The forming of signals (such as those employed in telegraph transmission) by the modulation of a direct-current or other carrier between discrete values of some characteristic.

Modulated Wave. A wave, some characteristic of which varies in accordance with the value of a modulating wave.

Modulating Wave. A wave which causes a variation of some characteristic of a carrier.

Modulation. The process or result of the process whereby some characteristic of one wave is varied in accordance with another wave.

Modulation Factor. The ratio of the peak variation actually used to the maximum design variation in a given type of modulation.

Note—In conventional amplitude modulation the maximum design variation is considered that for which the instantaneous amplitude of the modulated wave reaches zero.

Modulation Index. In frequency modulation with a sinusoidal modulating wave, the ratio of the frequency deviation to the frequency of the modulating wave.

Modulator. A device to effect the process of modulation.

Multiple Modulation. A succession of processes of modulation in which the modulated wave from one process becomes the modulating wave for the next.

Note—In designating multiple-modulation systems by their letter symbols, the processes are listed in the order in which the signal intelligence encounters them. For example, PPM-AM means a system in which one or more signals are used to position modulate their respective pulse subcarriers which are spaced in time and are used to amplitude modulate a carrier.

Percentage Modulation. The modulation factor expressed as a percentage.

Phase Deviation. The peak difference between the instantaneous angle of the modulated wave and the angle of the sine-wave carrier.

Phase Modulation (PM). Angle modulation in which the angle of a sine-wave carrier is caused to depart from the carrier angle by an amount proportional to the instantaneous value of the modulating wave.

Note—Combinations of phase and frequency modulation are commonly referred to as frequency modulation.

Pre-emphasis. A process in a system to emphasize the magnitude of some frequency components with respect to the magnitude of others.

Pulse Code. (a) A pulse train modulated so as to represent information. (b) Loosely, a code consisting of pulses, such as Morse code, Baudot code, binary code.

Pulse Code Modulation (PCM). Modulation which involves a pulse code.

Note—This is a generic term, and additional specification is required for a specific purpose.

Pulse-Duration Modulation (Pulse-Length Modulation) (Pulse-Width Modulation). A form of pulse-time modulation in which the duration of a pulse is varied.

Note—The terms "pulse-width modulation" and "pulse-length modulation" are also used to designate this system of modulation, but the term "pulse-duration modulation" is preferred.

Pulse Frequency Modulation (PFM). A form of pulse-time modulation in which the pulse repetition rate is the characteristic varied.

Note—A more precise term for "pulse frequency modulation" would be "pulse repetition-rate modulation."

Pulse-Interval Modulation. A form of pulse-time modulation in which the pulse spacing is varied.

Pulse Modulation. (a) Modulation of a carrier by a pulse train.

Note—In this sense, the term is used to describe the process of generating carrier-frequency pulses. (b) Modulation of one or more characteristics of a pulse carrier.

Note—In this sense, the term is used to describe methods of transmitting information on a pulse carrier.

Pulse Position Modulation (PPM). A form of pulse-time modulation in which the position in time of a pulse is varied.

Pulse-Time Modulation. Modulation in which the time of occurrence of some characteristic of a pulse carrier is varied from the unmodulated value.

Note—This is a general term which includes several forms of modulation, such as pulse-duration, pulse-position, pulse-interval modulation.

Quantization. A process in which the range of values of a wave is divided into a finite number of smaller subranges, each of which is represented by an assigned or "quantized") value within the subrange.

Note—"Quantized" may be used as an adjective modifying various forms of modulation, for example, quantized pulse-amplitude modulation.

Quantization Distortion (Quantization Noise). Inherent distortion introduced in process of quantization.

Quantization Level. In quantization a particular subrange, or a symbol designating it.

Quantization Noise. See **Quantization Distortion.**

Quantized Pulse Modulation. Pulse modulation which involves quantization.

Note—This is a generic term, including pulse numbers modulation and pulse code modulation as specific cases.

Regenerative Repeater. A repeater which performs pulse regeneration.

Note—Although this term carries the unfortunate connotation of a repeater employing a regenerative, or feedback, amplifier, its use in the literature has been wide and specifically as given in this definition.

Sidebands. (a) The frequency bands on both sides of the carrier frequency within which fall the frequencies of the wave produced by the process of modulation. (b) The wave components lying within such bands.

Note—In the process of amplitude modulation with a sine-wave carrier, the upper sideband includes the sum (carrier plus modulating) frequencies; the lower sideband includes the difference (carrier minus modulating) frequencies.

Side Frequency. One of the frequencies of a sideband.

Single-Sideband Modulation (SS, SSB). Modulation whereby the spectrum of the modulating wave is translated in frequency by a specified amount either with or without inversion.

Single-Tone Keying. That form of keying in which the modulating wave causes the carrier to be modulated with a single tone for one condition, which may be either "marking" or "spacing," and the carrier is unmodulated for the other condition.

Subcarrier. A carrier which is applied as a modulating wave to modulate another carrier.

Syllabic Companding. Companding in which the effective gain variations are made at speeds allowing response to the syllables of speech but not to individual cycles of the signal wave.

Synchronous Gate. A time gate wherein the output intervals are synchronized with an incoming signal.

Time-Division Multiplex. A device or process for the transmission of two or more signals over a common path by using successive time intervals for different signals.

Time Gate. A transducer which gives output only during chosen time intervals.

Two-Source Frequency Keying. That form of keying in which the modulating wave switches the output frequency between predetermined values corresponding to the frequencies of independent sources.

Two-Tone Keying. That form of keying in which the modulating wave causes the carrier to be modulated with one frequency for the "marking" condition and modulated with a different frequency for the "spacing" condition.

A Survey of the Limits in DC Amplification*

C. M. VERHAGEN†

Summary—**The influence of a cathode temperature change of planar diodes and triodes is calculated with the help of extended Langmuir tables. The possibility of compensation of this temperature effect is considered. Two units with a common cathode sleeve show no conclusive improvement in cancellation of the heater effect. The disturbance voltage as a result of anode voltage changes is investigated. Some new compensating circuits are presented, resulting in dynamic and in-phase balance simultaneously for a single adjustment. It is found that for all balanced circuits the stability demand for the supply or auxiliary voltage is nearly the same. The change of tube "constants" with time and the relation between anode current and flicker effect attenuation is investigated. It was observed that the poorly defined position of the heater in the cathode sleeve seriously limits tube stability.**

I. Introduction

OVER A PERIOD of some twenty years different circuits have been published showing methods for reducing the effect of supply-voltage changes. In order to investigate the operation of these, and other unpublished circuits, the influence of supply-voltage changes in these circuits will be investigated. To compare these most easily, the effect of supply-voltage changes will be expressed in an equivalent (′) input voltage v_x', the subscript (x) giving the source of the disturbance.

The Cathode Temperature

Its Influence

The investigation of the influence of the cathode temperature makes it necessary to consider first the action in a parallel-plane diode in connection with the space charge and the different escape voltages.[1]

In Fig. 1 these escape voltages together with the voltage minimum (Vm) caused by the space charge are shown for two different values of Ψk.

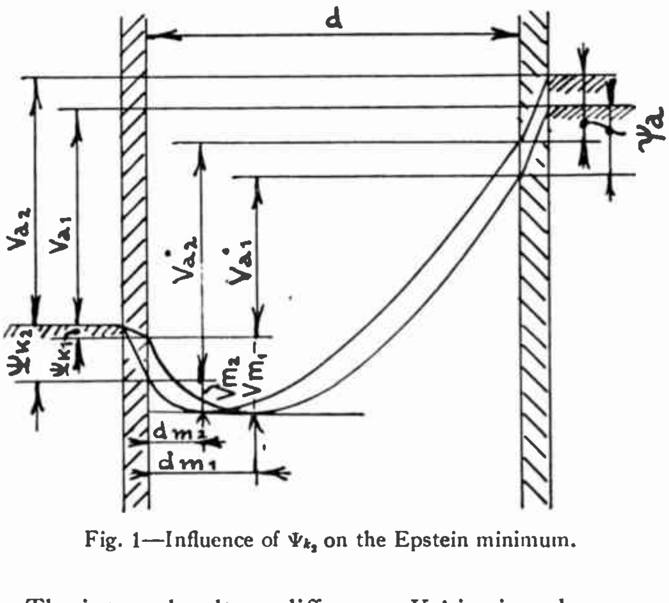

Fig. 1—Influence of Ψ_{k_2} on the Epstein minimum.

The internal-voltage difference Va' is given by

$$Va' = Va - \psi a + \Psi k. \tag{1}$$

In this, Ψk is the total escape voltage of the cathode.[2]

The influence of the space charge is given by the volt-

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[1] The escape voltage is equal to the work function divided by the electron charge.

[2] G. Herrmann and S. Wagener, "Die Oxydkathode II," Barth, Leipzig, Germany; 1943/1944.

age minimum (Epstein minimum) Vm:

$$Vm = -V_T \ln Is/Ia, \tag{2}$$

where $V_T = kT/e$.

Is can be given by

$$Is = AST^2 e \exp -\Psi k/V_T, \tag{3}$$

where S is the cathode surface (cm²).

The value of Ia can be found with the help of the Langmuir tables. Approximately, Ia can be given by

$$Ia = 2.33 \times 10^{-6} \times S \frac{(Va' - Vm)^{3/2}}{(d - dm)^2}. \tag{4}$$

The anode current is thus given by

$$Ia = 2.33 \times 10^{-6} \times S \frac{\left(Va - \psi a + V_T \ln \frac{AST^2}{Ia}\right)^{3/2}}{(d - dm)^2}. \tag{4a}$$

Hence the cathode-escape voltage (Ψk) does not appear[3] (see Fig. 1 for two values of Ψk). The cathode-escape voltage has an influence on d_m; this influence, however, is small (see IV, The Change in Tube "Constants" with Time).

When we neglect the influence of T on d_m we can find the equivalent input voltage for a cathode temperature change by differentiating the numerator of (4(a)) with respect to temperature. This voltage is expressed as an equivalent cathode-voltage change (v_T') giving the same change in anode current as the given temperature change.

$$v_T' = +\frac{dT}{11{,}600}\left(\ln \frac{AST^2}{Ia} + 2\right). \tag{5}$$

In this formula v_T' is the exact expression for the retarding-field region. Since in the space-charge region several approximations were used, v_T' was calculated accurately with the aid of the extended Langmuir tables of Kleynen.[4] The anode voltages are calculated for 970° and 1,000°K, and the difference divided by 30 (see Table I).

TABLE I

	Ia/S	v_T' (mv/degree centigrade)			
		$d=0.15$ mm	$d=0.25$ mm	$d=0.50$ mm	$d=0.80$ mm
$Is=3$ and $2A$/cm²	10 ma/cm²	1.92	2.00	2.53	2.94
	3 ma/cm²	1.81	1.94	2.20	2.33
	1 ma/cm²	1.76[5]	1.88	2.07	2.28
$Is=0.3$ and $0.2A$/cm²	10 ma/cm²	1.84	2.08	2.57	3.07
	3 ma/cm²	1.64	1.80	2.12	2.35
	1 ma/cm²	1.58	1.70	1.92	2.16

[3] B. Gysae and S. Wagener, "Der einfluss des kontaktpotentials auf die kennlinie von empfänger- und senderöhren," *Z. Tech. Phys.*, vol. 19, p. 264; September, 1938.

[4] P. H. J. A. Kleynen, "Extension of Langmuir's tables for a plane diode with a Maxwellian distribution of the electrons," *Philips Res. Rep.*, vol. 1, p. 81; January, 1946.

[5] In retarding-field region.

It is necessary to introduce values for A and Ψk to bring the results of Table I, in the retarding-field region, in agreement with (5).

There are, however, two lines of thought: one assuming A constant (120 amp/cm² degree²) and Ψk depending on temperature which makes for the chosen values of Is a temperature correction for Table I of about +0.7 mv necessary; the other assuming a much lower value for $\Psi k(\approx 1.0$ v) and A between 10^{-1} and 10^{-2}. By measuring v_T' on a diode we found results that were within 5 per cent of the calculations based on $\Psi k \approx 0.9$ v and $A = 0.10$ amp/cm² degree² ($Is = 3$ amp/cm²). By introducing $A = 120$ amp/cm² degree² and $\Psi k \approx 1.5$ v, however, our measurements deviated by more than 35 per cent which is difficult to explain by errors in the measurement. We therefore choose the former constants. Because this choice has an influence on v_T' that is much larger than the experimental error, it is not improbable that more insight can be obtained in this matter by accurate measurements of v_T'.

Introducing for A a value of 0.1 amp/cm² degree² in (5) we find

$$v_T' = \left\{1.76 - 0.2 \log \frac{Ia}{S}\right\} \text{mv/degree centigrade}, \tag{5a}$$

where Ia/S is in ma/cm².

Next we investigate the relation between a change in heater power and the resulting change in temperature.

Assuming a simplified tube structure where the conduction losses are given by $K_2(Tk - To)$ and the radiation losses by $K_1(Tk^4 - To^4)$ (To is the temperature of the surrounding air), we calculated $dWk/Wk\ Tk/dTk$ for several values of

$$\frac{K_2(Tk - To)}{K_1(Tk^4 - To^4)}$$

(see Table II). In these calculations K_1 is assumed to be independent of temperature.[6] The value of K_2 depends

TABLE II

$\frac{K_2(Tk-To)}{K_1(Tk^4-To^4)}$	$\frac{dWkTk}{WkdTk}$	$\frac{dTo}{dTk}$
0.4	3.26	6.7
0.5	3.14	5.8
0.6	3.04	5.1
0.7	2.94	4.5
0.8	2.85	4.1
1.0	2.71	3.5

strongly on the accuracy with which the cathode fits into the mica spacers. It is therefore not improbable that for small cathodes (Wk 1 to 2 w) the ratio

$$\frac{K_2(Tk - To)}{K_1(Tk^4 - To^4)}$$

[6] R. Champeix, "Pyrometrie optique des cathodes à oxyde mesure du pouvoir et massif spectrale," *Le Vide*, vol. 3, p. 469; July/September, 1948.

has a value between 0.5 and 1.0 with an average value of 0.7.[7] In accordance with the measurements of Liebold,[8] we assume for $dWk/Wk\ Tk/dTk$ an average value of 3.0.

To express a change of heater voltage as an equivalent input voltage we must lastly find the relation between changes in heater voltage and the subsequent changes of heating power. Generally the heater current source has an internal resistance Ri (see Fig. 2). Considering this and introducing $dRf/dW\ W/Rf = p$, we find that

$$dW/e_0 = 2W/Eo \frac{1}{1 - p\dfrac{Ri - Rf}{Ri + Rf}}. \tag{6}$$

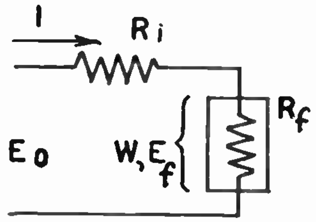


Fig. 2—Heater supply with internal resistance.

At the nominal heater voltage we found with a bridge method, $p = 0.24$. Let $T = 1{,}000°$K and insert for this temperature the value found for $W/dW\ dT/T$, then

$$dT = 2/3.0 e_0/Eo \frac{1{,}000}{1 - 0.24\dfrac{Ri - Rf}{Ri + Rf}}. \tag{7}$$

Inserting several values for Ri/R_f we can calculate dT (see Table III). Inserting an arbitrary constant k_0, the following results:

$$v_{f_0}' = v_T'{}_dT = k_0 e_0/Eo. \tag{8}$$

From (7) and (8) we find that

$$k_0 = v_T'2/3.0 \frac{1{,}000}{1 - 0.24\dfrac{Ri - Rf}{Ri + Rf}} \text{ volt.} \tag{8(a)}$$

Child[9] measured for $Ri \ll R_f$ a temperature change of 52.5° for a relative-heater-voltage change of 10 per cent whereas from Table III we find 54°.

TABLE III

	$Ri \gg Rf$	$Ri = Rf$	$Ri \ll Rf$
dT	0.88 $e_0/Eo\ 10^3$ degree	0.66 $e_0/Eo\ 10^3$ degree	0.54 $e_0/Eo\ 10^3$ degree

We must further consider the changes in voltage drop caused by $Ia\Delta Rox$, where ΔRox stands for the change in resistance of the oxide layer with temperature. This change ΔRox has two components, a bulk resistance change $\Delta R\dot{o}x$ and, if an interface layer is formed (see The Change in Tube "Constants" with Time), an interface resistance change $\Delta R\ddot{o}x$. With data from the measurements of Child,[9] and Metson[10] the resistance change of the oxide layer with interface, for an "old" tube (60,000 h) at a temperature of 1,000°K is found to be $Rox \approx -0.60$ ohm cm²/°C.

Assuming the same value for

$$\frac{\Delta Rox}{Rox}\frac{T}{\Delta T}$$

for new tubes, the resistance change will be much lower because both the bulk and interface resistance increase during life. The values in Table I have to be increased by the values found above, which means that the value of k_0 increases with life. An increase of at least 5 per cent has been observed in our measurements over a period of 200 hours.

We now have to find the equivalent input voltage for a cathode temperature change in a triode. The anode current in a triode, when we include the initial electron velocity, is given by

$$Ia = \frac{2.33\ 10^{-6}\ S\sigma^{3/2}}{(d - dm)^2}\left\{Vgk - \psi g + V_T \ln \frac{AST^2}{Ia} + \frac{Vak - \psi a + V_T \ln \dfrac{AST^2}{Ia}}{\mu}\right\}^{3/2}, \tag{9}$$

where[11]

$$\sigma = \frac{1}{1 + \dfrac{1}{\mu}\dfrac{d_{gk} + 4/3 d_{ak}}{d_{gk}}}.$$

This results in the equivalent circuit of Fig. 3. As a

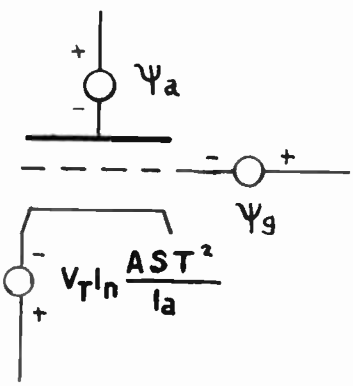


Fig. 3—Internal tube voltages.

change in cathode temperature causes only a variation in $-V_T \ln AST^2/Ia$ (compare (5(a)) with Table I)

[7] Information received from W. Nijenhuis of N. V. Philips' Gloeilampenfabrieken, Eindhoven.
[8] G. Herman and S. Wagener, "Die Oxydkathode II," p. 66, Barth, Leipzig, Germany; 1943–1944.
[9] M. R. Child, "The growth and properties of cathode interface layers in receiving valves," *P. O. Elec. Eng. Jour.*, vol. 44, p. 176; January, 1952.
[10] G. H. Metson, S. Wagener, M. F. Holmes and M. R. Child, "The life of oxide cathodes in modern receiving valves," *Proc. IEE*, part III, vol. 99, p. 69; March, 1952.
[11] W. Dahlke, "Gittereffektivpotential und kathodestromdichte einer ebenen triode unter berücksichtigung der inselbildung," *Telefunken Zeit.*, vol. 24, p. 213; December, 1951.

to a first approximation, the temperature change will, for the greater part, enter as a voltage change in series with the cathode. From (8) and (3(b)) follows

$$v_f' = 1/\sigma\ \mu/(\mu+1)v_{f_0}' = 1/\sigma\ \mu/(1+\mu)k_0e_f/Ef,$$

where $ef/Ef = e_0/Eo$ for $Ri \ll Rf$.

In this expression the "island" effect[12] has been neglected. This effect, however, is small for tubes with a sharp cut-off when the desired anode current is obtained with the lowest anode (screen) voltage permissible to prevent grid current. A detailed discussion and measurements of this effect on k_0 is given by Kessler.[13] In order to obtain an idea of the magnitude of k, from four different tube types, samples of 10 tubes each, the average k and the standard deviation, $S\Delta k$ were measured.[14]

$$S\Delta k = \sqrt{\sum_j^N \frac{(k_{j_1} - k_{j_2})^2}{N-1}}.$$

The first group of the ECC40 contained only new tubes, which were measured after an ageing period of 100 hours. All other tubes had been in use previous to this measurement. The small spread (6 per cent) found in the values of k of the ECC40 tubes is probably due to their same premeasurement treatment. To check this, the same measurement was done on a batch of used tubes of the same type ECC40, which showed a spread of the same magnitude (14 per cent) as found in the other types. A further confirmation was found in a second measurement of the spread from the type EF42 (see Table IV) in which new tubes were used, when the relative standard deviation $S\Delta k/k_{av}$ was observed to be 6.8 per cent.

TABLE IV

Tube type	$k_{average}$	$S\Delta k/k_{average}$	Condition
EF42	1.22 v	16%	$Vg_2k=150$ v, $Ia=5$ ma
EF40 (as triode)	1.29 v	16%	$Vak=150$ v, $Ia=1$ ma
ECC40	1.20 v	6%; 14%	$Vak=150$ v, $Ia=1$ ma
E90CC	1.44 v	19%	$Vak=150$ v, $Ia=1$ ma

These differences are probably caused by the increase of k due to the increase in resistance of the interface layer and oxide bulk with life.

Compensation of the Cathode Temperature Effect

The effect of a heater-voltage change can be partly compensated. The amount of compensation that can be obtained is directly dependent on the accuracy with which a compensating voltage, v_c', approaches the value of the disturbing voltage. The equivalent input voltage for a cathode temperature change and its compensating voltage are a function of time:

$$v_f'' - v_c'' = v_f'F_1(t) - v_c'F_2(t).$$

For $t=\infty$ both $F_1(t)$ and $F_2(t)$ are one. By adjusting $v_c = v_f'$, compensation for $r=\infty$ can be obtained

$$v_f'' - v_c'' = v_f'\{F_1(t) - F_2(t)\}.$$

The form

$$\{F_1(t) - F_2(t)\}$$

is zero for $t=0$ and $t=\infty$, and will have a maximum value β at $t=t_m$ (see Fig. 4).

An example of compensation is given by Ellenwood and Sorrows,[15] who obtain the time function $F_2(t)$ for the compensating voltage by an integrating RC network.

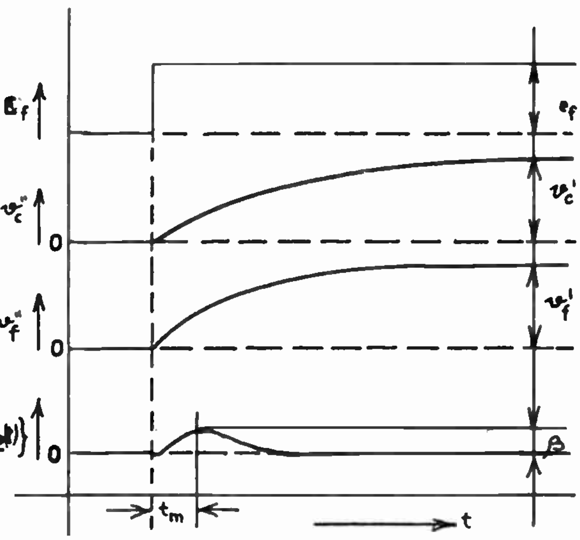


Fig. 4—Influence of different time constants in a compensation circuit.

For directly heated filaments a good compensation can be obtained, because the change of heater voltage appears as an immediate change of the average grid-cathode voltage and as the voltage v_f'. The sign and magnitude of the immediate voltage change depends on where the negative side of the anode supply is connected to the heater (see Fig. 5). The graph of Fig. 6 gives the relation between $(v_f' - v_c')$ and the position of the connection. When Ef and thus e_f are too low to compensate for v_f', the compensating voltage can be in-

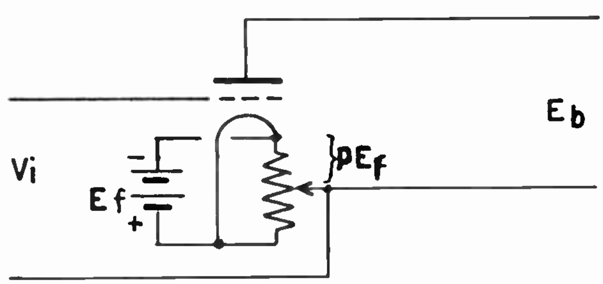


Fig. 5—Compensation circuit of a directly heated cathode.

[12] Ibid., p. 92.
[13] G. Kessler, "Der einflusz der heizung auf den kathodenstrom," *Arch. Elektrotech.*, vol. 39, p. 601; January, 1950.
[14] "Vacuum Tube Amplifiers," Rad. Lab. Series No. 18, McGraw Hill Book Co. Inc., New York, N. Y.; 1948.
[15] R. C. Ellenwood and H. E. Sorrows, "Cathode heater compensation for stabilized d.c. power supplies," *Jour. Nat. Bur. Stand.*, vol. 43, p. 251; September, 1949.

creased by increasing Ef and bringing a resistance in series with the heater.[16,17] This method is very effective,

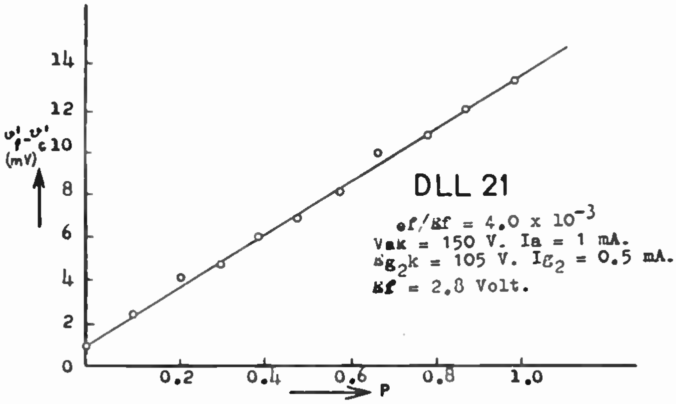


Fig. 6—Compensation as a function of the anode supply to filament connection.

especially when an accumulator is used for the heater supply, because only slow changes of Ef will occur, which makes the residual disturbance (β) negligible.

Miller[18] used as a compensating voltage the v_f' of an auxiliary tube in the circuit of Fig. 7. When the values of k of both tubes are different, the anode current Ia

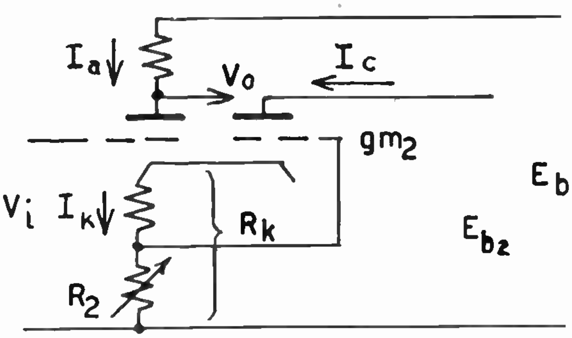


Fig. 7—Miller compensation circuit.

can be made independent of a change in heater voltage by adjusting R_2.

$\Delta R_2/Rk = \Delta k/k$ where $\Delta R_2 = R_2 - 1/gm_2$.

From our experience it was found that the adjustment of R_2 is time consuming and that after the adjustment is made compensation will be lost again due to changes in k with time. Though β will be small for this circuit, this will limit its performance when compensation is obtained. To find this limitation $\Delta k/k$ and the relative residual disturbance β were measured for two tube types (EF42 and 12SC7) in a modification of the Miller circuit. The results of measurements on 20 new tubes, type EF42, are given in Table V. During these measurements it was found that $\{F_1(t) - F_2(t)\}$ depends upon whether temperature is increasing or decreasing.

In calculating and measuring β a step function was used. In a practical design we must take into account the maximum occurring e_f/Ef. This maximum relative heater-voltage change, however, will seldom occur in a time short compared with t_m. In our laboratory we found that taking into account for β half of the values measured with a step function covered nearly all of the changes occurring in the supply line.

TABLE V

EF42 (as triode)	$Vak = 150$ v, $Ia = 1$ ma
$k_{av} = 1.22$ v $S\Delta k/k_{av} = 6.8\%$	$\beta_{av} = 4.8\%$ $t_{m_{av}} = 5.4$ sec
12SC7 (systems of different envelopes)	$Vak = 150$ v, $Ia = 1$ ma
$k_{av} = 1.39$ v $S\Delta k/k_{av} = 9.5\%$	$\beta_{av} = 3.1\%$ $t_{m_{av}} = 5.7$ sec
12SC7 (systems of the same envelope)	$Vak = 150$ v, $Ia = 1$ ma
$k_{av} = 1.39$ v $S\Delta k/k_{av} = 10.6\%$	$\beta_{av} = 2.3\%$ $t_{m_{av}} = 4.7$ sec

It was thought that by using double tubes with a common cathode sleeve, such as the 12SC7, it would be possible to find smaller values for β_{av} and $S\Delta k/k_{av}$, as compared with units with thermally separated cathodes. To investigate this, the same measurements done on the type EF42 tubes were done on 10 new tubes of the type 12SC7. The first measurement was on two similar units in different envelopes, in the second measurement two units of one envelope were compared. The results show that $S\Delta k$ for units in the same envelope, with the same cathode sleeve, is 12 per cent higher compared with units from different envelopes, while β_{av} is 25 per cent lower. Since number of tubes was limited, results are not conclusive, but significant improvement is improbable.

Negative Feedback of Cathode Changes

By using a current-dividing grid in a tube with two or more grids, voltage amplification can be obtained. The current to be divided can be kept constant by feedback; this feedback will to a first approximation have no effect on the mutual conductance of the current-dividing grid. Hence the anode current controlled in this way will only be slightly dependent on the cathode temperature. In a tetrode the screen grid has such a current dividing action, see Fig. 8.[19] We find for a small change

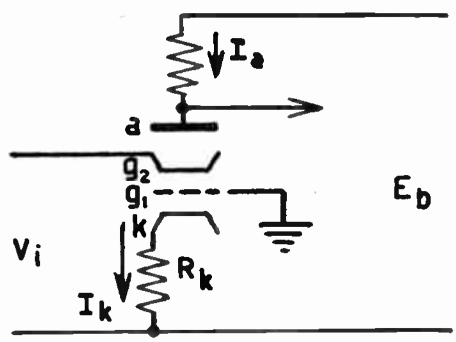


Fig. 8—Current feedback in current dividing tube.

[16] J. C. M. Brentano, "The measurement of ionisation currents by three electrode valves," *Nature*, vol. 108, p. 532; December, 1921.

[17] L. A. Turner and C. O. Siegelin, "An improved balanced circuit for use with electrometer tubes," *Rev. Sci. Instr.*, vol. 4, p. 429; August, 1933.

[18] S. E. Miller, "Sensitive d.c. amplifier with a.c. operation," *Electronics*, vol. 14, p. 27; November, 1941.

[19] See reference, footnote 15, p. 465

of the cathode voltage (v_f'), by neglecting the influence of the screen grid voltage on Ik, that

$$i_k = gm/(1 + gmRk)v_f'.$$

Let $i_a = i_k Ia/Ik(Vg_2g_1/Vag_1 \text{ constant})$ and $gmRk \gg 1$, then

$$i_a = 1/Rk\ Ia/Ik\ v_f'.$$

The same effect can be obtained by placing the feedback resistor in series with the space-charge grid. With an appropriate tube construction, an anode current can be drawn with a very small screen-grid voltage and small screen-grid current. The screen grid with its current-dividing action can then be used as a control grid with a relatively high mutual conductance

$$\left(\frac{\partial Ia}{\partial Vg_2g_1}\right)_{Ik} = gm_{\text{div}}.$$

The equivalent input voltage for a heater-voltage change will be

$$v_f'_{(\text{div})} = \frac{v_f'}{gm_{\text{div}}Rk}\,\frac{Ia}{Ik}.$$

This circuit will give the same attenuation for cathode-current changes regardless of the cause of the change (temperature, escape voltage or increase of the resistance of the oxide layer).

A cathode-voltage change also has a small influence on the current division. By this action the already attenuated cathode-voltage influence can be cancelled over a limited cathode-voltage region as shown by Müller and Dürichen,[20] Bousquet,[21] and Gray.[22]

Because no tubes are available with the above mentioned construction, and having the other desired properties for sensitive dc amplifiers, no measurements were done to test this arrangement. The construction of such a suitable tube will greatly diminish the difficulties encountered in sensitive dc amplifiers.

The Influence of Supply and In-Phase Voltages

After calculating the influence of cathode temperature changes we shall now investigate the influence of anode or screen-grid supply-voltage changes for several circuits.

Calculating the anode-voltage change v_{a_b} caused by supply voltage Eb for the circuit of Fig. 9 we find that

$$v_{a_b} = e_b\, Ri/(Ri + Ra).$$

Dividing v_{a_b} by the amplification, we find the equivalent input voltage v_b'.

$$v_b' = -e_b/\mu\ Ri/Ra = -e_b/Eb\ Ia/gm(Ra + Ro)/Ra, \quad (10)$$

where $Ro = Vak/Ia$.

[20] F. Müller and W. Dürichen, "Zur frage der nullpunktstabilität von elektrometerröhrengeräten," *Phys. Zeit.*, vol. 39, p. 657; September 15, 1938.

[21] A. G. Bousquet, "Random emission compensation," *Gen. Rad. Exp.*, vol. 48, p. 2; May, 1944.

[22] See reference, footnote 14, p. 465.

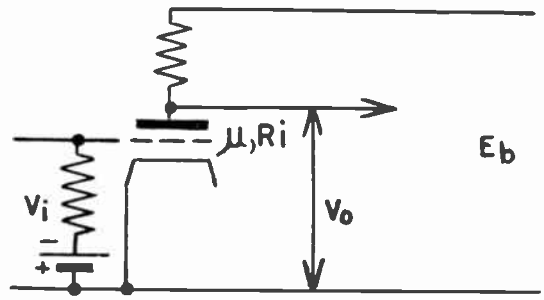


Fig. 9—Single-tube circuit.

To decrease the influence of e_b, a bridge circuit can be used, as in Fig. 10. When the circuit is dynamically balanced

$$[(Ra/Ri = R_1/R_2) \text{ and } Vi = 0],$$

we find that

$$Vo = Eb\ Ra/(Ri + Ra)\,\frac{Ro - Ri}{Ro + Ra}.$$

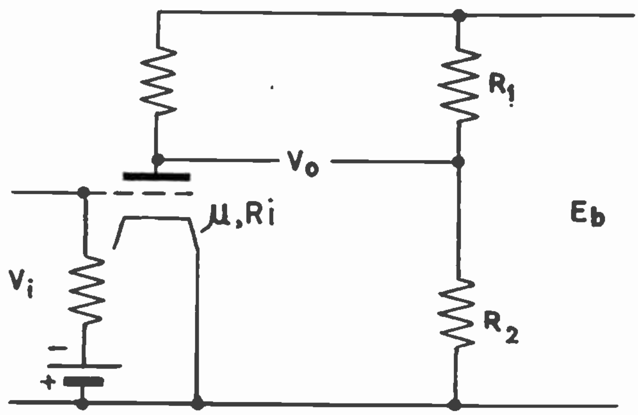


Fig. 10—Bridge circuit with single tube.

If a voltage source of the opposite sign is introduced in series with Vo, both balances are obtained.[23] If the bridge is statically balanced ($Ra/Ro = R_1/R_2$) we find that

$$v_b' = v_{0b}/A = e_b/Eb\ Ia/gm\{Ro/Ri - 1\}. \quad (11)$$

From this we see that the influence of the supply voltage decreases when Ri approaches the value of Ro. The ratio Ro/Ri, for a triode with fixed bias (battery), is always greater than 1. If a cathode resistor Rk_0 is used for bias without decoupling in a tube with a constant μ, the ratio

$$Ro'/Ri' = \frac{Ro + Rk_0}{Ri + (1 + \mu)Rk_0}$$

will be nearly 1. Ro' and Ri' represent the static and dynamic resistance between the anode and the negative side of the cathode resistor.[24,25] A disadvantage of this method is that a decrease in amplification results. This

[23] W. Soller, "One-tube balanced circuit for d.c. vacuum-tube amplifiers of very small currents," *Rev. Sci. Instr.*, v. 3, p. 416; Aug. 1932.

[24] M. M. Levy, "Balancing of Electrical Bridge Circuits Containing Nonlinear Elements," U. S. patent 2,440,282; Nov. 20, 1943.

[25] See reference, footnote 9, figs. 11, 13.

decrease is caused by an apparent increase in internal resistance equal to $(1+\mu)Rk_0$ (also see Fig. 23).

We will look further at two circuits in which the ratio Ro'/Ri' can be made exactly 1. For these circuits we will also investigate the possibility of making Ro'/Ri' much smaller than 1, because this will mean a simple method for obtaining voltage, or current, stabilization.

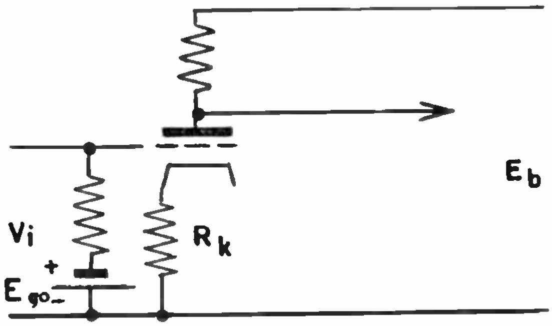


Fig. 11—Current feedback with positive grid-cathode voltage.

If a value of Rk greater than that necessary for giving the desired bias is selected, thus $Rk>Rk_0$, and if the extra negative voltage is opposed by a positive grid voltage by means of a battery (see Fig. 11), the ratio Ro'/Ri' can be made exactly 1, or smaller.

$$Ro'/Ri' = \frac{Ro + Rk}{Ri + (1+\mu)Rk}.$$

Letting $Ro+Rk_0=Ri+(1+\mu)Rk_0$, multiplying and dividing by Ia, and assuming $Vak+Eg_0 \gg Vgk$, we find that

$$\frac{Ro'}{Ri'} = \frac{1 + Vak/Eg_0}{(1+\mu) + Vak/Eg_0}. \qquad (12)$$

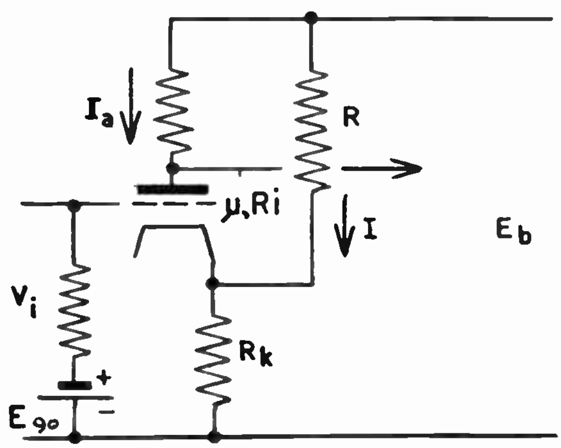


Fig. 12—Turner compensation circuit for a triode.

In the circuit of Turner[26] in Fig. 12 the internal resistance will have a negligible increase, resulting in normal amplification. In this circuit Ri'' can be made infinite and even negative. When we calculate the anode-current change for a change in supply voltage e_b and designate $(1-\mu\ Rk/R)=\Delta$, assuming this nearly zero, and further letting $\mu \gg 1$, we find

$$i_a = e_b\Delta/(Ri + Ra + \mu Rk) \quad \text{or}$$
$$Ri'' = e_b/i_a = (Ri + Ra + \mu Rk)/\Delta. \qquad (13)$$

By adjusting $R/Rk \leqq \mu$, Δ will become zero or negative, and thus the ratio Ro''/Ri'' also.

It is clear from the foregoing circuits that a bridge circuit with one tube can be statically and dynamically balanced simultaneously. Because of the decrease in mutual conductance of a tube with time, the dynamic balance must be checked occasionally. Another disadvantage of using one tube in a bridge circuit is that no decrease in the effect of a cathode-temperature change will be obtained. By applying two tubes in a bridge circuit, a partial cancellation of the cathode-temperature effect will be obtained, and there will be a chance that the dynamic resistance of both tubes will change by the same amount. If both tubes are exactly alike, there will be static and dynamic balance, furthermore the heater effect of both tubes will cancel out. It is, however, difficult to select such tubes because of spread in k, μ and gm (see Tables IV, V and VI). We therefore calculate the influence of anode supply voltage changes and derive from this result the spread to be allowed for a given equivalent input voltage.

TABLE VI

EF42 $Ia=2$ ma $-$ $Vak=140$ v, EF40 $Ia=1$ ma $-$ $Vak=80$ v

Type	μ	gm(ma/v)	$Ri(k\Omega)$
EF42 EF40	70.0 (14.2%) 35.6 (4.4%)	4.19 (15.5%) 1.56 (3.6%)	16.7 (3.1%) 22.9 (3.7%)
	$Ri'(k\Omega)$	Vg_0 (v)	Veff (v)
EF42 EF40	73.7 (6.7%) 81.3 (8.2%)	−1.36 (0.30) −1.63 (0.12)	+0.41 (0.17) +0.36 (0.14)

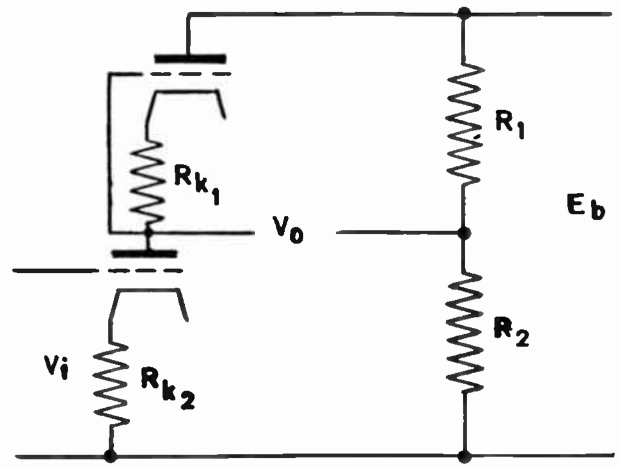


Fig. 13—Bridge circuit with two tubes in series.

Artzt[27] has published a bridge circuit in which the two tubes are connected in series (See Fig. 13). We denote $\mu_1-\mu_2=\Delta\mu$, $(\mu_1+\mu_2)/2=\mu$, and so forth. The dynamic

[26] L. A. Turner, "On balanced d.c. amplifying circuits," *Rev. Sci. Instr.*, vol. 4, p. 665; December, 1933.

[27] M. Artzt, "Survey of d.c. amplifiers," *Electronics*, vol. 18, p. 112; August, 1945.

resistance of a tube $Ri+(1+\mu)Rk$ we designate as Ri' (see Fig. 23) and $R_1=R_2+\Delta R$. For the influence of e_b we find

$$v_b' = e_b/2\mu\{\Delta Ri'/Ri' - \Delta R/R_1\}$$
$$= e_b/Eb\ Ia/gm'\{\Delta Ri'/Ri' - \Delta R/R_1\}(Ro'/Ri'), \quad (14)$$

where Ro'/Ri' is nearly 1. From Table VI we can read directly the value of $\Delta Ri'/Ri'$ to be expected for the given tube types, when the circuit is statically balanced by adjusting Rk. We thus find that for some selection (70 per cent acceptable) the relative difference $\Delta Ri'/Ri'$ is approximately 8 per cent. We can calculate that the influence of a heater-voltage change in this circuit is

$$v_f' = k\ e_f/Ef\{\Delta k/k + \Delta\mu/\mu - \Delta Ri'/Ri'\}. \quad (15)$$

Thus tube pairs with both a tolerable v_b' and v_f' must be selected, or v_f' can be made zero by adjusting $\Delta Ri'$ by means of Rk_1 or Rk_2, the resulting dynamic unbalance ($v_b'\neq 0$) being compensated for by adjusting ΔR as in (14). The static balance has then to be obtained by a third adjustment in the circuit following this stage.

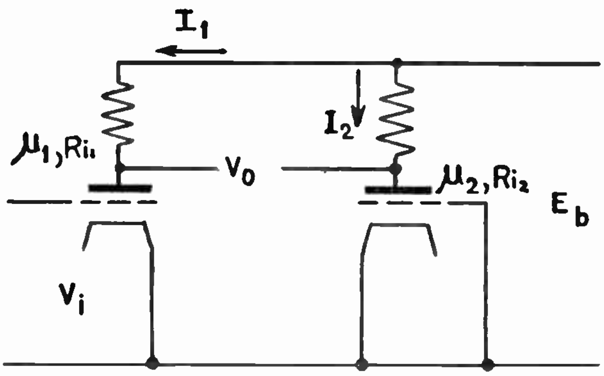


Fig. 14—Bridge circuit with two tubes in parallel.

A different circuit for two tubes in a bridge is given by Brentano[16] (see Fig. 14). The influence of e_b is given by

$$v_b'=e_b/\mu\ Ri/(Ri+Ra)\{\Delta Ra/Ra-\Delta Ri/Ri\}$$
$$=e_b/Eb\ Ia/gm\{\Delta Ra/Ra - \Delta Ri/Ri\}(Ro+Ra)/(Ri+Ra). \quad (16)$$

Contrary to the foregoing circuits, this one can amplify symmetrical input signals. Before we further investigate the sensitivity of such balanced circuits for supply-voltage changes, we will first consider a new requirement. This requirement is high discrimination, in case we want to measure floating voltages.

Floating Measurements

When a voltage between two points, A and B (Fig. 15), has to be measured, and neither of these points is grounded, there will generally be some difficulties due to ground- or third-electrode (C) currents when electron tubes are used for this measurement. Improvement can often be obtained by using an amplifier with a symmetrical input (Fig. 16); for generality we assume the amplifier is not grounded. Let the input voltage on each grid be v_{i_1} and v_{i_2} respectively, then we can separate these into two components,[28] an in-phase component v_{i_0}: $v_{i_0}=(v_{i_1}+v_{i_2})/2$, and an out-of-phase component v_i: $v_i=v_{i_1}-v_{i_2}$.

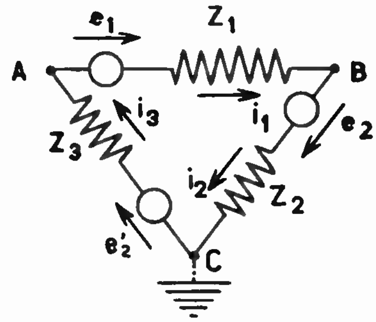


Fig. 15—Ground currents in a floating voltage source V_{AB}.

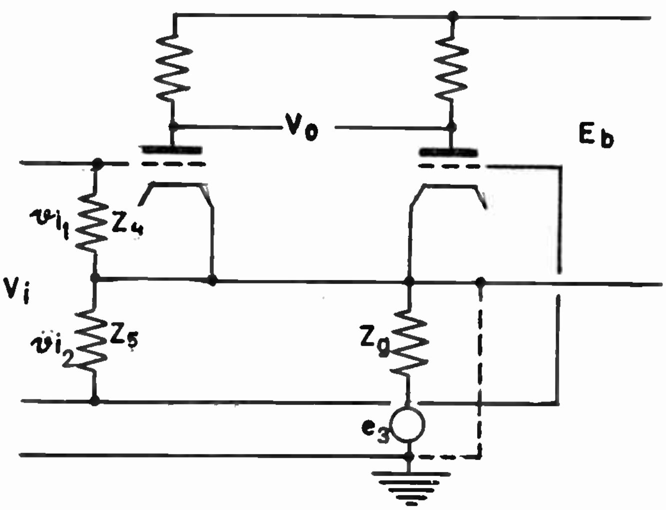


Fig. 16—Symmetrical amplifier (not grounded).

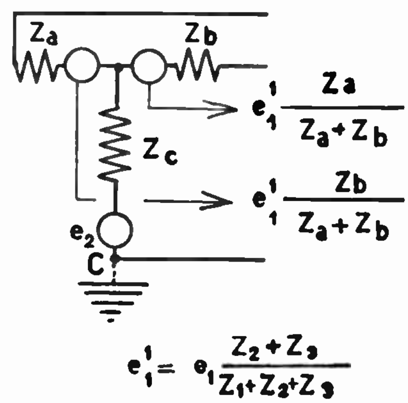


Fig. 17—Voltage source with symmetrical output.

When $e_2'=ne_2''$ in case $e_2'\neq -e_2''$, and/or $Z_2\neq Z_3$, balance can be obtained by adjustment of Z_4/Z_5. This leaves only the in-phase input voltage for which Fig. 15 may be changed in Fig. 17.

The sensitivity of the amplifier for in-phase grid voltages can be defined as the ratio of the peak to peak values of the inphase to out-of-phase input voltage, each voltage producing the same magnitude of output

[28] F. Offner, "Balanced amplifiers," Proc. I.R.E., vol. 33, p. 202; March, 1945.

voltage. We designate this ratio as the discrimination (D) of the amplifier, $D = v_{i_0}/v_0'$. The in-phase balance or discrimination of the circuit of Fig. 16 is given by

$$1/D = v_0'/v_{i_0} = \{\Delta\mu/\mu + Ri/(Ri+Ra)\{\Delta Ra/Ra - \Delta Ri/Ri)\}. \quad (17)$$

The discrimination can be made infinite, when the ratio Z_4/Z_5 (see Fig. 18) is adjusted in such a way as to cancel the result of the in-phase input voltage v_0' of the amplifier.[29,30] Another method is the adjustment of $\Delta\mu$ in (17), thus making the right-hand part zero. The grid cathode voltage range, for an infinite D, is, however, limited by the difference in curvature of both tube characteristics.

After some calculations we find for the difference voltage between the anodes when the curvature is taken into account

$$(i_{a_1} = a_{11}v_{gk_1} + a_{12}v^2{}_{gk_1} \cdots, v_{gk_1} = v_{gk_2} = V \sin \omega t, a_{11} - a_{21} = \Delta a_1, \text{ and so forth}).$$

$$v_0 = \Delta i_a Ra = Ra/4\, V\{2\Delta a_2 V + (4\Delta a_1 + 3\Delta a_3 V^2) \sin \omega t - 2\Delta a_2 V \cos 2\omega t \cdots \}.$$

Making (17) zero, the term $\{4\Delta a_1 + 3\Delta a_3 V^2\}$ vanishes. Neglecting the dc component and dividing by the amplification $(a_1 Ra)$,

$$v_0' = v_0/A = -1/2\, V/a_1\{\Delta a_2 V \cos 2\omega t \cdots \} \quad (18)$$

$$1/D = 1/2\, \Delta a_2/a_1\, V \cdots. \quad (18(a))$$

Hence the output voltage will contain only higher harmonics of v_{i_0}. When v_0' is still too large and point C in Fig. 15 is grounded, v_{i_0} can be decreased in case the current $(e_2+e_3)/(Zc+Zg)$ is smaller than e_2/Zc, by not

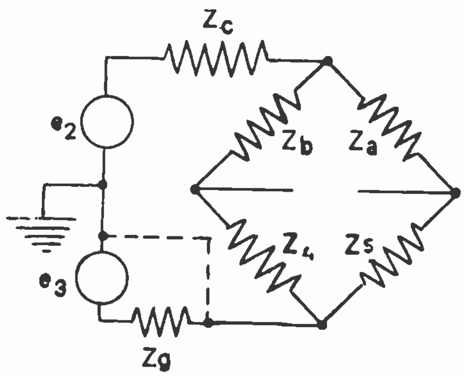


Fig. 18—Circuit for in-phase voltages.

grounding the amplifier (see Fig. 18). By application of negative feedback to the in-phase input voltage, v_0' can still further be decreased as will be shown.

For large values of D we find for the influence of heater-voltage changes that

$$v_f' = v_{f_1}' - v_{f_2}' = k\, e_f/Ef\{\Delta k/k\}. \quad (19)$$

[29] J. F. Toennies, "Differential amplifier," *Rev. Sci. Instr.*, vol. 9, p. 95; March, 1938.

[30] H. Goldberg, "Bio-electric research apparatus," Proc. I.R.E., vol. 32, p. 330; June, 1944.

The Feedback of In-phase Grid Voltages

By applying feedback the allowable in-phase input voltage increases for a given discrimination D. An in-phase feedback circuit has been given by Offner[31] (see Fig. 19). We find for D:

$$D = \frac{v_0'}{v_{i_0}} = -\frac{1}{AgmRa}\left\{\frac{\Delta\mu}{\mu}\,\frac{Ra+Ri}{Ri} + \frac{\Delta Ra}{Ra} - \frac{\Delta Ri}{Ri}\right\}. \quad (20)$$

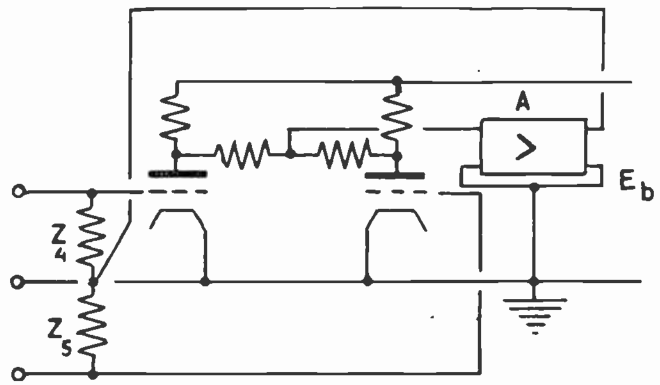


Fig. 19—Anode-gride in-phase feedback.

Hence by means of a suitable amount of amplification, A, of the feedback amplifier, the value of v_0' can be kept as low as desired. A feedback amplifier can be used, or the in-phase amplification of the first, and if necessary, further stages of the dc amplifier itself can perform this function. Disadvantages of this circuit are, that the grids are not at ground potential and, further, that the feedback of this circuit depends on the ratio $Z_4(Z_5)$ to $\{Zc+Za(Zb)\}$ (see Fig. 18), which makes the feedback dependent on the output resistance of the voltage source to be measured. The influence of the supply voltage in this circuit is given by

$$v_b' = e_b/gmRa\{\Delta\mu/\mu + \Delta Ra/Ra - \Delta Ri/Ri\}$$
$$= e_b/Eb\; Ia/gm\{\Delta gm/gm + \Delta Ra/Ra\}(Ra + Ro)/Ra. \quad (21)$$

Comparing this result with the value of v_b', found for a symmetrical amplifier without feedback (16), shows that no improvement is obtained.

By using a large common cathode resistor, together with an extra supply voltage (Ek) to counteract the dc drop across this resistor, we obtain a negative feedback (Fig. 20) that does not have the disadvantages of the circuit of Fig. 19.[29] The introduction of this negative voltage, Ek, however, gives rise to a new disturbance voltage caused by a change e_k. It can be shown, however, that by taking the correct precaution the voltage changes e_k and e_b (see Fig. 20) can be thought of as a single disturbance voltage, e_t, combined with an in-phase grid voltage. The latter is not harmful when the discrimination of the amplifier is made large. To show

[31] F. Offner, "Push-pull resistance coupled amplifiers," *Rev. Sci. Instr.*, vol. 8, p. 20; January, 1937.

this we separate e_b and e_k into two components:

$$e_k = (e_k + e_b)/2 + (e_k - e_b)/2 \quad \text{and}$$
$$e_b = (e_k + e_b)/2 - (e_k - e_b)/2.$$

We designate $(e_k - e_b)/2 = e_g$ where e_g has the same influence as v_{i_g}, so long as a change of e_g only changes the average grid voltage. This means that point O in Fig. 20 should only be used for fixing the grid potential. Random changes of Ek and Eb can thus be thought of as an in-phase voltage $e_g = (e_k - e_b)/2$, together with a supply-voltage change $e_t = e_k + e_b$.

We can calculate the equivalent input disturbances, v_b' and v_k', for changes in Eb and in Ek (see appendix):

$$v_b' = -\frac{e_b}{2gmRk}\left(\frac{\Delta\mu}{\mu} - \frac{2Rk}{Ri}\right) \tag{22}$$

$$v_k' = \frac{e_k}{2gmRk}\{\Delta\mu/\mu(Ri + Ra)/Ri - \Delta Ri/Ri + \Delta Ra/Ra\}. \tag{23}$$

The sum of v_k' and v_b', for $e_k = -e_b = e_g$, gives the influence of e_g or v_{i_g}:

$$v_g' = \frac{v_{ig}}{2gmRk}\{\Delta\mu/\mu(Ra + 2Rk + Ri)/Ri - \Delta Ri/Ri + \Delta Ra/Ra\}. \tag{24}$$

This formula we also find in "Vacuum Tube Amplifiers,"[32] but the influence of Ra is there neglected. The complete formula is given by Parnum.[33] In his paper the relation of (24) with those giving the influence of the supply voltages is not, however, presented. The influence of e_t is given by the sum of $v_k' + v_b'$, for $e_b = e_k = e_t/2$,

$$v_t' = \frac{e_t}{4gmRk}\{\Delta\mu/\mu(Ra - 2Rk + Ri)/Ri - \Delta Ri/Ri + \Delta Ra/Ra\}. \tag{25}$$

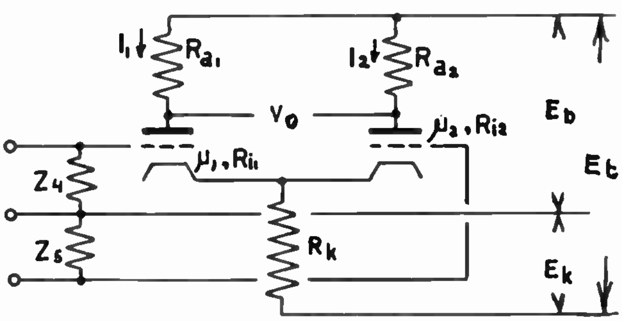


Fig. 20—Cathode-gride in-phase feedback.

With (19), (24), (25), influence of supply-voltage changes in circuit of Fig. 20 is completely described.

We shall now investigate the influence of the value of Rk on the discrimination. By increasing Rk the value of v_g' approaches a limited value, not zero, given by

$$\lim_{Rk\to\infty} v_g' = v_{i_g}\Delta\mu/\mu^2. \tag{24(a)}$$

This shows that (even by neglecting the nonlinearity of the characteristic) the discrimination can never be infinite when $\Delta\mu$ is not zero. This finite value is due to the influence of v_b' on v_g', because

$$\lim_{Rk\to\infty} v_b' = -e_b\Delta\mu/\mu^2. \tag{22(a)}$$

We see that the choice of a large value of Rk has only a limited effect in decreasing v_g'.

When we are able to make the discrimination infinite (v_g' zero) in the circuit of Fig. 20, by adjusting $\Delta\mu/\mu$ in the right-hand part of (24), a large value of Rk will become important in reducing the effect of supply-voltage changes. This can be shown by inserting the so found value of $\Delta\mu/\mu$ in (25):

$$v_t' = \frac{2e_t}{4gmRk}\{\Delta\mu/\mu(Ra + Ri)/Ri - \Delta Ri/Ri + \Delta Ra/Ra\}.$$

Since for large values of $gmRk\ \Delta\mu/\mu$ is very small, the first term can be neglected, giving

$$v_t' = e_t/Et\ Ik/2gm\{\Delta Ra/Ra - \Delta Ri/Ri\}Rk_s/Rk\ Et/Ek, \tag{25(b)}$$

where Rk_s is the static resistance of the cathode resistor and Rk its dynamic value.

By using a linear $Rk(Rk_s/Rk = 1)$, v_t' depends only on the ratio Et/Ek which has a minimum value of 1. When using a separate supply Ek together with a linear Rk, its influence is given by

$$v_k' = e_k/Ek\ Ik/2gm\ \{\Delta Ra/Ra - \Delta Ri/Ri\}. \tag{23(a)}$$

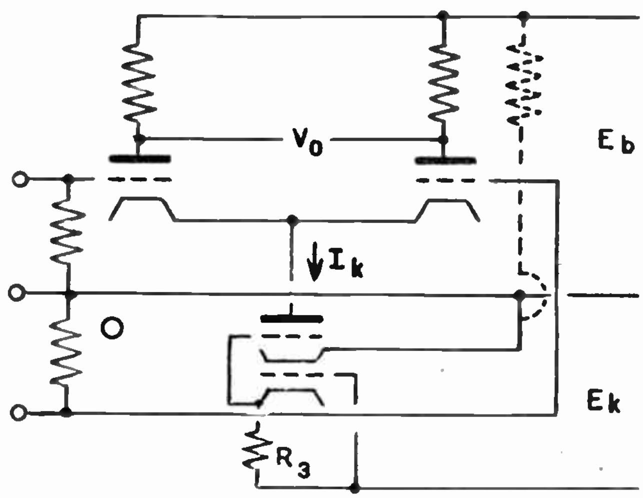


Fig. 21—Cathode-grid in-phase feedback with a pentode as cathode impedance.

The use of a pentode as a cathode resistor[30] decreases $Ek/Ik = Rk_s$ for a given dynamic $Rk(Rk \gg Rk_s)$, see Fig. 21. In this way we can obtain a value between 1/20 and 1/30 for Rk_s/Rk.

For the equivalent input voltage for a disturbance e_{g_2k} when the right-hand part of (21) is made zero, we find

[32] See reference, footnote 9, p. 448.
[33] D. H. Parnum, "Transmission factor of differential amplifiers," *Wireless Eng.*, vol. 27, p. 125; April, 1950.

after some calculations that

$$v_{g}'{}_{2k} = e_{g_2k}/Eg_2k\; Ik/2gm\{\Delta Ra/Ra - \Delta Ri/Ri\}. \quad (26)$$

We see that the stability requirement is the same as for (23(a)), but the voltage that must have this stability is lowered by a ratio Rk/Rk_s.

If the screen grid is supplied from point O (see Fig. 21), as given in most publications,[30] the desired effect for in-phase grid voltages is obtained. The screen-grid voltage, however, and thus Ik, will vary with Ek and for this reason the balance for in-phase grid voltages $(v_{i_1}+v_{i_2})/2$ will not coincide with the balance for $e_g=(e_k-e_b)/2$. When not a separate supply for Eg_2k is used, this would result in an extra balance condition except when the screen grid is connected directly to $+Et$ through a series resistor. The stability requirement for e_t/Et will then be given by (26), with e_t/Et instead of e_{g_2k}/Eg_2k.

In cases where the stability demand of (26) is difficult to attain, a modification of the circuit of Turner[26] can be used as shown in Fig. 22. The screen-grid current

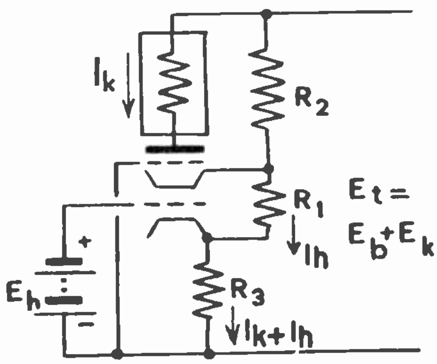


Fig. 22—Turner compensation circuit for a pentode.

of the cathode pentode must then be thought of as the anode current of the triode of Fig. 11. This screen-grid current can be made independent of supply-voltage changes e_t. The influence of changes in the screen-grid voltage on the current division can be compensated for by slightly readjusting the compensation condition $(\mu_{g_1g_2}=R_1/R_3)$. Then the anode current is independent of supply-voltage changes, e_t, and the tube behaves as a pentode for changes of Rk. In this circuit a new disturbance source, Eh, is introduced, however. For the influence of a change e_h we find after some calculations that

$$v_h' = e_h/Eh\; Ik/2gm\{\Delta Ra/Ra - \Delta Ri/Ri\}\,p \quad (27)$$

The value of p was measured on the EF40 and EF42 and found to be between 1.2 and 2.0 ($Vgk_0=-1.3$ v, $Ik=18$ ma, $Ia=6$ ma). The stability requirement for Eh is hence nearly the same as found for Ek (linear Rk) or Eg_2k. The value of Eh is, however, much smaller, and no current is drawn, making the application of Weston cells possible. The sensitivity of this circuit for changes in the cathode-grid circuit (microphonics, cathode temperature) is not negligible (e_h for a given v_h'); the same applies to the circuit of Fig. 21.

Another circuit in which a high dynamic resistance is obtained by means of current feedback was given in Figs. 11 and 12. This circuit can, for a tube with a high μ, have the same ratio of dynamic to static resistance as found with the pentode, but it allows a much higher disturbance level e_g for a given v_{g_0}'. The stability demand for Eg_0 is again the same as for Ek, $E_{g_2}k$, and Eh in the circuits of Figs. 19, 21, and 22.

From the foregoing calculations it is clear that an increase of Rk decreases v_g', but will only decrease the disturbance for a given relative supply-voltage stability e_t/Et if the voltages mentioned above (Ek, Eg_2k, Eh or Eg_0) are not taken from this supply voltage Et, but are taken from a separate voltage source. This separate source, however, must have a stability as given by (23(a)), (26), or (27).

We will finally calculate the increase in the portion of the tube characteristic over which the same value of D can be obtained when a common cathode resistor with a high dynamic value is used compared with the circuit where no in-phase feedback is applied (Fig. 16). For an in-phase grid voltage the circuit of Fig. 20 can be looked on as a cathode follower with both tubes in parallel. For such an amplifier we find when $2Rk \gg (Ri+Ra)$ that

$$v_{gk} = v_{i_g}/(1+\mu).$$

Combining this with (18) we find

$$v_g' = 1/2\,\Delta a_2/a_1\{[V/(1+\mu)]^2\cdots\}, \quad (18(b))$$

$$1/D = v_g'/v_{i_g} = 1/2\,\Delta a_2/a_1\{V/(1+\mu)^2\cdots\}, \quad (18(c))$$

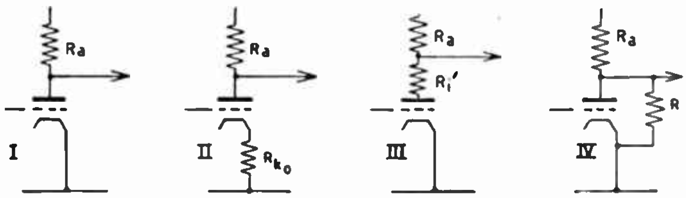
where V stands for the in-phase alternating voltage between (D, E) and O see Fig. 20). Hence the improvement for a large value of Rk will be $(1+\mu)^2$.[29] Because D depends on V, it is necessary to give the amplitude of the in-phase grid voltage by stating the discrimination.

In the circuit of Offner, shown in Fig. 19, the decrease of v_b' or v_g' by adjustment of $\Delta\mu$ or ΔRi is less effective because of the change in Ri with time.

We can now compare the sensitivity of the different circuits for supply-voltage changes. It is seen that one circuit is far inferior to the others (the single-tube circuit of Fig. 9) and another circuit is far superior to the others (25(b)). It must be remembered however, that for this case a $\Delta\mu$ adjustment is needed, and that, besides the supply voltage Et, an auxiliary voltage Ex is needed with the same stability demand as given for the other circuits. It is seen that a high value of specific conductance (gm/Ia) is important to keep the disturbance low. This value increases with decreasing Ia, especially in the retarding-field region if the grid does not lose control by over parts of the cathode surface "island effect."[12] Working in the retarding-field region means, however, a low mutual conductance which makes the attainment of a large bandwidth difficult, and gives a higher value of shot noise. Tubes with a high absolute and specific mutual conductance are nearly always microphonic. These limitations make the choice of suitable tubes difficult.

Compensation of Tube Differences

In Fig. 23 several ways are presented in which the triode tube constants can be altered externally, without making connections to the grid. The latter is important if the compensation is to be independent of resistance changes of the voltage source to be measured. We can look upon each of these circuits as being another tube with different constants as given under the circuit.



μ	μ	μ	$\frac{\mu}{1+Ri/R}$
Ri	$Ri+(1+\mu)Rk_0$	$Ri+Ri_1$	$\frac{Ri}{1+Ri/R}$
$9m$	$\frac{\mu}{Ri+(1+\mu)Rk_0}$	$\frac{\mu}{Ri+Ri_1}$	$3m$

Fig. 23—Change of tube constants.

Only in the last circuit is the amplification factor (μ) changed.[29,33,34] In Fig. 20 the important requirement for decreasing disturbance caused by supply-voltage changes and in-phase grid voltages is μ compensation.

A disadvantage of all known compensation circuits is the disturbance of static balance when the compensating resistors are adjusted (Rk_0, Ri', or R). This can be prevented by bringing these resistors in such a circuit that there is no dc drop across the resistor when there is static balance and no input voltage. A solution of the above problem for the circuit of section IV of Fig. 23 is given in Fig. 24. With the proper choice of R_2 and R_3,

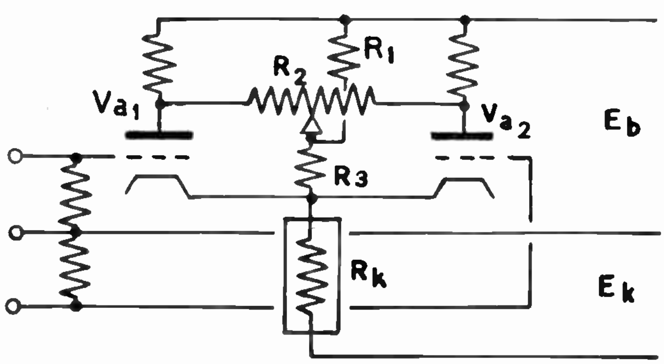


Fig. 24—μ balance without reaction on static balance.

the necessary range for maximum difference in μ can be obtained. By use of this circuit, in-phase balance ($v_g'=0$) can be obtained (see Fig. 24); by choosing a common cathode resistor, Rk, with a large dynamic resistance, dynamic balance is also obtained. Hence by adjusting R_2 the two balance conditions can be fulfilled simultaneously. A circuit that gives an external dynamic balance ($v_t'=0$) is given in Fig. 25, where R_4 is a nonlinear resistance ($\alpha\neq\beta$). For static balance ($Va_1=Va_2$) there will be no voltage drop across R_1+R_1' and R_2+R_2' if $Va=\alpha Et$. Thus the adjustment of dynamic balance will not disturb the static balance. Let the voltage change across the tubes, caused by a change e_t, be γe_t. Then we find a voltage change of $(\gamma-\beta)e_t$ across the voltage dividers R_1+R_1' or R_2+R_2'. By adjusting R_1 a voltage difference between both voltage dividers can be obtained that cancels the out-of-balance voltage between the anodes.

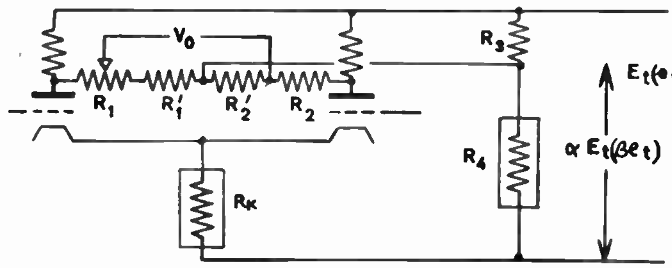


Fig. 25—Dynamic balance without reaction on static balance.

Tube Constants

Static Balance

It is important that the adjustment for static balance does not interfere with the other balances. The use of batteries is, however, unpractical (load of voltage divider). By adjusting a cathode or an anode resistor, static balance can be obtained. These adjustments, however, affect the other balances by changing the tube constants (Fig. 23). The effect can be decreased by introducing an auxiliary current through the static balancing resistors (Fig. 26) that is many times

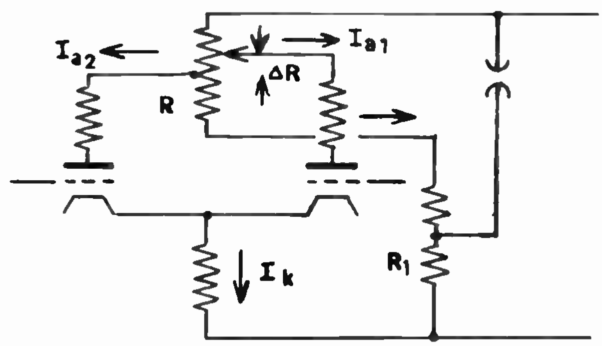


Fig. 26—Static balance for a triode bridge with decreased reaction on other balances.

greater than the cathode current Ik. With this arrangement only a small difference in the position of the taps, (ΔR), is needed to obtain static balance. If pentodes are used, the balancing voltage-difference can be applied between the screen grids while the anode current acts as the auxiliary current as shown in Fig. 27. The resistance changes are, in Fig. 26, $(I/Ia+1)$ or in Fig. 27, $(2Ia/Ig_2+1)$ times smaller than before. The influence on in-phase or dynamic balance will be decreased by the same amount. The stabilizer tube T_1 to keep the auxiliary current constant does not appear in the circuit of Fig. 27 because changes in Ik are already kept small for other reasons previously mentioned.

[34] C. Williamson and J. Nagy, "A push-pull-stabilized triode voltmeter," *Rev. Sci. Instr.*, vol. 9, p. 270; September, 1938.

The Spread in Tube Constants

From the investigation of the influence of supply-voltage changes it was seen that the disturbances depend mainly on the difference in tube constants. To obtain an idea of the differences that can be expected, the value x_{av} and the standard deviation ($S\Delta x/x_{av}$ in percentage or $S\Delta x$ in volts) were measured for two tube types. The internal resistances were measured with a fixed bias on the grid Ri or with undecoupled cathode resistor. The voltage Veff stands for $Vgk+Vak/\mu$. All tubes measured were aged for at least 200 hours.

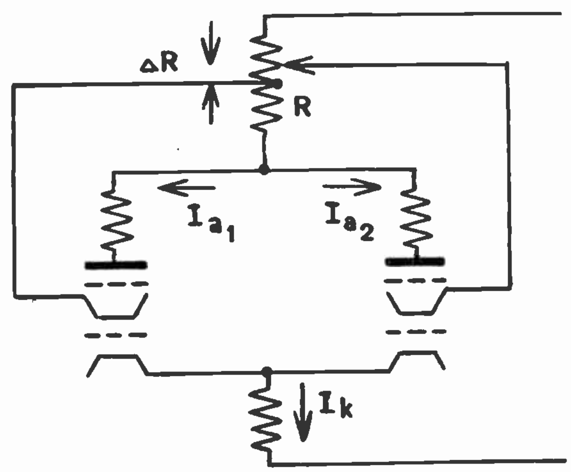


Fig. 27—Static balance for a pentode bridge with decreased reaction on other balances.

Keeping in mind that the results are not conclusive due to the limited number of sample tubes (10), we found a definite correlation between the spread in μ and Vgk when the spread in μ is large. For a constant value of $Ro(Vak/Ia)$ the spread in Ri will be smaller than it is for μ or gm. When tubes with a high value of gm are operated with a relatively low anode current we can expect a large spread in μ and gm. For obtaining static balance with a small anode current and bias an adjustment range of plus and minus 0.35 v will be sufficient.

Changes With Time

It is important to know how constant the tube characteristics remain during tube life. This is especially true when it is desired to compensate a balanced amplifier. We will investigate the possibility of anode current changes in a plane diode due to internal changes of the tube. The value of Ia is given by Langmuir[35] as an infinite series, which shows that Ia depends upon

anode-cathode voltage $V\dot{a}-Vm$,	cathode temperature Tk,
anode-cathode distance d,	emission current Is.

The escape voltage ψa for molybdenum is 4.2 v. If, however, barium is evaporated onto anode, escape voltage will be changed,[36] and thus value of $(V\dot{a}-Vm)$.

Rothe[37] found by measurement of $(\Psi k-\psi a)$ on 15 tubes type REN 904 (Telefunken), for a cathode temperature of about 500°K, values between -0.5 to -1.4 v. Assuming $\Psi k=1.0$ v we find $\psi a=1.9\pm0.5$ v. This is in good agreement with the values obtained by Lukirsky,[38] who found for a monomolecular Ba layer a value of 1.5 v, increasing to a constant value of 2.0 v when the layer thickness is more than a double-molecular layer.

The evaporation of Ba from the cathode depends strongly on its temperature, decreasing nearly ten times for a temperature decrease of 50 degrees C.

In our measurements (with temperature control) tubes were stable (drift <100 μv/10 hr) after the tubes were aged at the nominal heater voltage for about 100 hours. By ageing at the working-heater voltage, for which we chose 90 per cent of the nominal value, no change in drift (>100 μv/hr) was observed after an ageing period of 150 hours.

Except for tubes with a very small distance d between cathode and anode the value of d will stay constant during the life of a tube, if it is not severely shocked.

Changes of Tk can be caused by changes in surrounding temperature (To). By assuming a simplified tube structure (as done previously) we can obtain a relation between changes in surrounding air temperature (dTo) and subsequent changes of cathode temperature (dTk) (see Table II). Because $K_2(T_k-T_0)/K_1(T_k^4-T_0^4)$ may have a value between 0.5 and 1.0 the difference between dTk/dTo of two tubes can be $1/3.5-1/5.8=0.11$. Assuming an equivalent input voltage of 2.5 mv/°C we can expect a maximum equivalent input voltage in a balanced amplifier of 275 μv/degree. We measured on such a stage with tubes type EF40 values between 50 and 250 μv/degree. Peirson[39] found with common cathode tubes (12SC7) in a balanced amplifier a temperature dependence of 100 to 200 μv/degree.

Cathode-temperature changes can also be caused by spontaneous changes in heater resistance or position. By recording the output of a dc amplifier we observed changes (step function) probably caused by changes of the heater position or resistance. These changes observed with some tubes of the EF40 type have an equivalent input amplitude between 20 and 300 μv and a rise time of about 10 seconds. By simultaneously observing the output voltage and heater resistance, we found a definite relation between both changes. By using tubes with special heaters (as will be discussed) the above mentioned changes were never observed.

Besides sudden heater changes, some heaters show a slow change with time, perhaps caused by recrystallization of the tungsten heater wire. The influence of these changes can be kept small by supplying the heater

[35] I. Langmuir, "The effect of initial velocities on the potential distribution and thermionic current between parallel plane electrodes," *Phys. Rev.*, vol. 21, p. 419; April, 1923.

[36] H. B. Michaelson, "Variations of grid contact potential and associated grid currents," *Jour. Frank. Inst.*, vol. 249, p. 455; June, 1950.

[37] H. Rothe, "Untersuchungen über den gitterstromeinsatzpunkt bei verstärkerröhren," *Telefunken Zeit.*, vol. 13, p. 45; July 1932.

[38] P. I. Lukirsky, A. Sosina, S. Wekshinsky and T. Zarewa, "Versuche über die eigenschaften der atomschichten," *Zeit. fur Phys.*, vol. 71, p. 306; August 29, 1931.

[39] D. H. Peirson, "A D.C. amplifier using an electrometer valve," *Electronic Eng.*, vol. 22, p. 48; February, 1950.

from a voltage source with an internal resistance Ri equal to Rf:

$$dW/W = \Delta Rf/Rf \frac{Ri - Rf}{Ri + Rf - p(Ri - Rf)}, \quad (28)$$

where dW/W is the relative change in heater power, p as given in (6) and ΔRf the spontaneous resistance change.

A change in Is can occur for a change in Ψk. This can be caused by loss of Ba through evaporation or by poisoning. For cathode temperatures not higher than 1,000°K evaporation is small and will not limit the life of modern receiving tubes.[10] The increase of Ψk by poisoning depends on the amount of gas that is absorbed by the cathode and will be low if tubes are carefully manufactured and used.[10] The influence of a change in Ψk in a diode in the retarding-field region is zero because Ia does not depend on Ψk, except when the area sizes with constant Ψk are small and the interaction between the electrons of the different constant Ψk areas cannot be neglected.

If in the space-charge region the value of Ψk changes, a change in d_m (see Fig. 1) results. This will change Vak if Ia is kept constant. With the help of the extended Langmuir tables given by Kleynen[4] the equivalent input voltage was calculated, as accurately as the tables are given, for several values of d and Ia/S.

TABLE VII

$\{(V\dot{a}_1 - V\dot{a}_2) - (\Psi k_1 - \Psi k_2)\} = Vak_1 - Vak_2 = v\Psi_k{}^1$ in mv

d (mm)	Ia/S in ma/cm² 1		3		10	
	1,000°	970°	1,000°	970°	1,000°	970°
0.15	0.0¹	− 0.3²	− 5.3	− 6.3	−19.4	−23.1
0.25	−2.7	− 3.3	− 9.7	−11.5	−26.8	−32.1
0.50	−5.9	− 7.3	−15.6	−19.1	−39.6	−46.5
0.80	−8.6	−10.8	−20.7	−24.1	−51.7	−61.0

¹ $V\dot{a}_1$ and $V\dot{a}_2$ in retarding-field region.
² $V\dot{a}_1$ in retarding-field region.

The emission changes for which table VII was calculated were from 2 amp/cm² to 0.2 amp/cm² for $T = 970$°K ($\Delta\Psi k = 192.5$ mv), and from 3 amp/cm² to 0.3 amp/cm² for $T = 1{,}000$°K ($\Delta\Psi k = 198.5$ mv). The equivalent input voltage for a change in Ψk in the space-charge region is according to these calculations not zero—contrary to the statement of Gysae and Wagener.[3]

Due to chemical reaction between the oxide layer and the nickel sleeve an interface layer will sometimes result.[40,41] This layer seems to grow during the first half to two years of the life of a tube (continuous use),[42] and then stay constant at a value of about 40 Ωv cm². The shift which results from this resistance can be kept low by selecting a small value for the specific anode current.

Besides the slow changes in Ψk during the life of a tube, Ψk will change continuously around some average value. This appears at the output of the tube as a noise called flicker effect. It is probable that these changes can be treated as the very slow changes previously described. This perhaps explains the results obtained by van Wijngaarden.[43] He found that the flicker effect decreases greatly with a decrease in anode current, and approaches a constant low value in the retarding-field region.

Changes in d_m will not only change Ia but will also change the mutual conductance. The relative d_m change depends on the change in Is and the ratio Is/Ia. A low value of Ia and a high value of d will result in a relatively small change in gm for a given change in Is. In a well-designed tube, working under proper conditions, the change in Is during life will be small.

Nearly all the drop in gm observed in life tests for tubes with long life (>10,000 hr) can be explained by the increase of resistance of the interface layer in the oxide coating.[42,10] The drop in mutual conductance will be decreased when a low value of gm/S is chosen and electrode potentials are selected such that the "island effect"[12] is small.

Shock Stability

For sensitive amplifiers not working under laboratory conditions mechanical and acoustical insensitivity is an important characteristic. The results of shocking tubes can be divided into three effects.

The first effect is commonly referred to as *microphonics*. Tubes with low microphonics can be chosen from recommended types, or by selection (by testing) from a number of tubes. The latter method is unsatisfactory, however, for practical reasons. The equivalent input voltage v_m' of a given tube is, for relatively low anode currents, in most cases proportional to Vb/μ where Vb is the undecoupled part of the anode (screen) supply voltage and μ is $(\partial V_{ak}/\partial V_{gk})_{Ia}$ for triodes or $(\partial V_{g_2k}/\partial V_{g_1k})_{Ig_2}$ for pentodes.[44] Measuring the tube types EF40, E80F and CF50 (all low microphonic) EF42 (video amplifier) (all Philips) and VR56 (Rogers) we found the results outlined in Table VIII.

TABLE VIII

	EF40	E80F	CF50	EF42	VR56(EF6)
Number of tubes tested	50	13	8	9	420
Smaller than					
10. mv	50	13	8	7	
3. mv	46	13	7	2	
1.0 mv	46	10	5	0	166
0.3 mv	3	2	2		25
0.10 mv	0	0	0		8
0.03 mv					2

[40] W. Liebold, Thesis, University of Berlin; 1941.
[41] A. Eisenstein, "Some electrical propreties of an oxide cathode interface," *Phys. Rev.*, vol. 72, p. 531; September 15, 1947.
[42] C. C. Eaglesfield, "Life of valves with oxide-coated cathodes," *Elec. Comm.*, vol. 28, p. 95; June, 1951.
[43] J. G. van Wijngaarden, "Low Frequency noise in electron tubes," Thesis, Vrije Universiteit, Amsterdam; July, 1951.
[44] W. Graffunder und H. Rothe, "Prinzipielle Untersuchungen über das Klingen von Verstärkerröhren," *Telefunken Röhre 2*, vol. 36, p. 147; March 6 and November 8, 1936.

The measurement results are the equivalent input voltages (v_m') measured from peak to peak when the tubes are shocked. The shocking of the tubes in all the measurements to test shock stability was done by tapping the tubes with the fingernail. Though the shock given with this method is not clearly defined, better defined methods make the use of special apparatus necessary, which makes the checking of a tube outside the laboratory difficult. The VR56 is not a low microphonic tube type, but the results are given as an example of the possibilities of tube selection within a given type.

The effects in the second group are the results of more or less permanent changes in electrode positions. Changes in the anode current, produced by the above effect, are accompanied by changes of the mutual conductance and/or amplification factor. These changes were measured with a General Radio vacuum-tube bridge type 561-D. The results of this measurement on 3 tubes of the following types, connected as triodes are outlined in Table IX.

TABLE IX

	Vak	Ia(ma)	μ	$(\Delta\mu/\mu)_{max}$	$Ri(k\Omega)$	$(\Delta Ri/Ri)_{max}$
EF42	140 v	2	70	0.5%	17	1.2%
EF40	80 v	1	36	<0.15%	23	<0.2%

The anode-current changes falling in the third group are slow changes, as the result of a change in cathode temperature when the tube is shocked. Contrary to spontaneous changes, these are impulse-like. The change in cathode temperature can be the result of changes in the heater resistance by "cold" deformation, by recrystallization, or by a change in the heater position within the cathode sleeve. The latter cannot be the only cause, however, because changes of heater resistance when the tube was tapped were also observed in directly heated tubes and in metal filament incandescent lamps. For these measurements the tubes were directly soldered in the circuit, and as all heaters of the measured tubes were spot welded to the base pins, bad contacts were ruled out. By putting a tube of the type E80F in a sensitive dc amplifier, with the heater in a Wheatstone bridge, it was observed that by tapping the tube a change of heater resistance of 0.1 per cent was accompanied by a change of cathode temperature equivalent to about 1 mv input. Table X gives result of some measurements on the equivalent input voltage (slow changes) when the tubes were shocked.

In some tubes type E80F the heater wire was wound on a ceramic cylinder which gave, when tapped, changes of less than 100 μv. No spontaneous changes were observed with these special tubes.

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Appendix

Let

$$Ri_1 + (1 + \mu_1)Rk + Ra_1 = R_1, \qquad (1 + \mu_1)Rk = R_1',$$

$$Ri_2 + (1 + \mu_2)Rk + Ra_2 = R_2 \text{ and } (1 + \mu_2)Rk = R_2',$$

$$i_1 = \frac{R_2\{(\mu_1 + 1)e_k + e_b\} - R_1'\{(\mu_2 + 1)e_k + e_b\}}{R_1R_2 - R_1'R_2'},$$

$$i_2 = \frac{R_1\{(\mu_2 + 1)e_k + e_b\} - R_2'\{(\mu_1 + 1)e_k + e_b\}}{R_1R_2 - R_1'R_2'}.$$

Assume $e_k = 0$. Then

$$v_0 \approx \frac{e_b}{R_1R_2 - R_1'R_2'}\{\Delta\mu RkRa_2 + \Delta\mu RkRa_1 + \Delta RiRa - \Delta RaRi\}$$

$$R_1R_2 - R_1'R_2' \approx (Ri + Ra)\{Ri + Ra + 2(\mu + 1)Rk\},$$

$$v_b' = -e_b\frac{Ri}{Ri + Ra + 2(\mu + 1)Rk}\left\{\frac{\Delta\mu}{\mu}\,\frac{2Rk}{Ri} + \frac{1}{\mu}\left(\frac{\Delta Ri}{Ri} - \frac{\Delta Ra}{Ra}\right)\right\}.$$

TABLE X

Tube type	EF40	E80F	12SC7	UAF42
Number of tubes tested	12	5	8	4
Heater resistance cold	5Ω	3.3Ω	126Ω	126Ω
Thickness of complete heater with insulation	0.43 mm	0.76 mm	0.70 mm	0.31 mm
Internal diameter cathode sleeve	1.20 mm	0.95 mm	1.05 mm	0.72 mm
Average of maximum change of each tube	1.3 mv	0.7 mv	0.25 mv	0.4 mv
Maximum change of all tubes	2.5 mv	1.7 mv	0.5 mv	0.8 mv

Letting $2Rk \geqq Ri+Ra$ and $\mu \gg 1$ we find

$$v_b' = -\frac{e_b}{2gmRk}\left\{\frac{\Delta\mu}{\mu}\frac{2Rk}{Ri}\right\}. \tag{22}$$

Let $e_b=0$, then

$$v_0 \approx \frac{e_k}{R_1R_2 - R_1'R_2'}\{-\Delta\mu Ra^2 - \Delta\mu RaRi + \Delta RiRa(1+\mu) - \Delta RaRi(1+\mu)\},$$

$$v_k' = e_k\left(\frac{\mu+1}{\mu}\right)\frac{Ri}{Ri+Ra+2(\mu+1)Rk} \cdot \left\{\frac{\Delta\mu}{\mu+1}\frac{Ra+Ri}{Ri} + \frac{\Delta Ra}{Ra} - \frac{\Delta Ri}{Ri}\right\}.$$

Letting $2Rk \gg Ri+Ra$ and $\mu \gg 1$ we find

$$v_k' = \frac{e_k}{2gmRk}\left\{\frac{\Delta\mu}{\mu}\frac{Ra+Ri}{Ri} + \frac{\Delta Ra}{Ra} - \frac{\Delta Ri}{Ri}\right\}. \tag{23}$$

Theory of the Large Signal Behavior of Traveling-Wave Amplifiers*

A. NORDSIECK†

Summary—**We present results of calculations on the amplification, phase distortion, and harmonic content in an idealized traveling-wave amplifier when excited to levels beyond those for which the device behaves linearly. The idealization consists mainly of neglecting the action on the electron beam of radio-frequency fields due to the space charge in the beam. The limiting efficiency and the phase distortion are given for various values of the beam-to-circuit coupling parameter and of the electron injection velocity. The relative harmonic content of the beam is given to the fifth harmonic frequency, inclusive. The calculations were performed by numerically integrating the equations of motion of typical electrons entering the tube at equal intervals of initial phase. Most of the numerical work was done with IBM equipment, some with desk calculators, and in one case, on the Bell Laboratories relay computer. The results are presented almost entirely graphically, and the reader is referred to figures in main body of the paper for quantitative information.**

I. Introduction

THE "SMALL SIGNAL" or linear theory of the traveling wave amplifier has been treated by several authors.[1–6] They find that if the squares and products of ac quantities are neglected, so that the equations become linear in these quantities, there is a particular wave whose amplitude increases exponentially along the tube provided the electrons are injected into the tube with initial speed in a certain range near the phase velocity of the electromagnetic wave in the circuit. If the tube is long enough and the input signal strong enough, this wave is not a "small signal" everywhere in the tube and the linear theory does not apply. Nevertheless, it is interesting to know how the amplitude and phase of this "large-signal" wave change as one proceeds along the tube, since the tube may find important uses in the fields of broad-band power amplification, amplification plus limiting in frequency modulation, and harmonic generation. The calculations reported here were undertaken to answer the questions: (a) What is the limiting power level for various values of the beam-to-circuit coupling parameter and of the injection velocity? (b) How much phase distortion is introduced by driving the tube beyond its linear range of power output? (c) How rich in harmonics is the electron stream when the tube is driven hard? In all of the calculations it is assumed that there is only a single driving frequency present, so that the question of intermodulation is untouched. Probably the most serious approximation made in the work is the assumption that the electrons are accelerated by the electric fields of the circuit only, not by the fields of neighboring electrons (neglect of space-charge forces). No satisfactory way has been found to include these space-charge forces in the calculations without an unwarranted increase in labor. Thus the results are strictly applicable only to very small experimental beam current densities; however, it is felt that at all beam current densities used in practice the only serious effect of the space-charge forces is to reduce the high-harmonic content of the beam below that calculated.

The set of equations to be solved contains 1) the equations for the instantaneous fields in the circuit or transmission line and 2) the equations of motion of the electron stream. 1) Consists of ordinary transmission-line equations, except for a distributed generator which is the electron stream. The transmission line is taken to be loss-free, or more precisely, its loss is assumed small compared to the electronic small-signal gain, and is therefore neglected. Assuming a fixed finite loss would, it is believed, have made the results of less general

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1 Kompfer, Hatton, and Ashcroft, British Admiralty Report C.V.D. Cl. Misc. 40; January, 1945.

2 Bernier, *Ann. Radioelect.*, vol. 2, pp. 87–101; January, 1947.

3 Pierce, Proc. I.R.E., vol. 35, pp. 111–123; February, 1947.

4 Chu and Jackson, Report No. 38, M.I.T. Research Lab. of Electronics, Cambridge, Mass.; April, 1947.

5 Blank-Lapierre, Lapostolle, Vogue, and Wallauschek, *Onde elect.*, vol. 27, pp. 194–202; May, 1947.

6 Schulman and Heagy, *RCA Rev.*, vol. 8, pp. 585–611; Dec. 1947.

interest. In 2) the force on an electron is taken to be the instantaneous field at the electron due to the transmission line. The equations 2) may be written in two ways, either by regarding the electron stream as a fluid with a given density and velocity at every point along the tube, or by following a typical set of electrons in their motions. It was found very early in the course of the calculations that the second of these choices must be used because electrons overtake one another at or even considerably before the point along the tube where the limiting power level is attained. The first choice is clearly inadequate if overtaking occurs since the velocity of the fluid at a point is then not unique. This circumstance does not cause an increase in the labor of solving equations, but does complicate the problem of connecting our equations with published small-signal equations, for in these latter the fluid point of view is used.[3]

We use the same dimensionless coupling parameter C as defined and used by Pierce[7] and also the same dimensionless parameter b to specify the injection velocity.[8] Since all our results are functions of these two parameters, it may be well again to define them here.

$$C^3 \equiv (E^2u_0^2/2\omega^2P)(I_0/4V_0),$$

where E is the peak longitudinal electric field at the beam due to the wave propagating in the circuit, ω is the radian radio frequency, u_0 is the electron injection speed, P is the power carried by the wave, and V_0/I_0 is the direct-current beam impedance. Thus C is the real cube root of, essentially, a ratio of a circuit characteristic impedance to the beam impedance; $b\equiv(u_0-v_0)/Cu_0$, where u_0 and C are defined above and v_0 is the phase velocity of the wave in the absence of a beam. Thus a positive (respectively negative) value of b means average electron speed greater (respectively less) than the phase velocity of the circuit wave in the absence of a beam. Note, however, that the phase velocity of the *increasing* wave in the *presence* of the beam is always less than the average electron speed.

In the calculations the dependence of the behavior on C is handled as follows: We show that for $C\ll 1$ we may neglect certain terms in the equations and by introducing reduced variables make C disappear entirely from the equations. Thus we can treat all *small* values of C simultaneously. This suffices for comparison with most experimental work since it is very difficult in practice to make C as large as, for example, 0.1. The question of how small C must be for our "small C" results to apply is then answered by a separate single calculation in which the above-mentioned terms are not neglected. The dependence of tube behavior on b, i.e., injection speed, is handled by solving the equations six times over for as many chosen numerical values of b.

All but one of the calculations are performed on the assumption that all electrons of a given initial phase experience the same field, as would be the case if the beam were an infinitely thin pencil on the axis of the transmission line or an infinitely thin cylindrical shell coaxial with the line. Experimental practice is almost always far from this limiting case, but if the value of C used in the theory is the experimental value for the actual tube in question, determined from the small-signal gain or in some other equivalent way, we may say that the theory is treating the average electron of each initial phase, i.e., neither the most strongly nor the most weakly accelerated one. In any case a check on errors introduced by this assumption is provided by again making a separate single calculation in which a beam of finite diameter is simulated by dividing beam into two equal parts and assuming that one part experiences an accelerating field twice as large as the other.

For every integration of the equations a fixed numerical value of the power level had to be assumed at the point in the tube (called $z=0$) where the integration was started. This level was in all cases chosen just large enough so that the nonlinear terms in the equations were 5 per cent as large as the linear terms at $z=0$. The initial conditions at $z=0$ were taken to be those of the pure increasing wave of the small-signal solution. Conditions at points $z<0$, i.e., on the input side of the point of starting the integration, are given by the analytical formulas of the small-signal theory. For comparison with experimental operation of a tube of fixed length and variable input power, one must pick a segment of fixed length $l=z_2-z_1$ out of this mathematical model, with z_1 at the point corresponding to the experimental input power, and read off the graphs the voltage amplitude and phase at z_2. This interpretation is valid only if l is large enough so that when the tube is limiting at z_2 the small-signal theory still applies for some distance beyond z_1.

In Section II the working equations are derived, and in Section III the results on high-level amplification and phase distortion for various values of the injection speed are presented. Section IV deals with the two special calculations for finite C and finite beam diameter, respectively. These are done for the single value 1.5 of b. Section V presents the results concerning harmonic content of the beam.

II. Derivation of the Working Equations

Let the longitudinal radio-frequency electric field at the beam be $\partial V(z, t)/\partial z$, where z is the distance along the tube. Then the differential equation for V is

$$\frac{\partial^2 V}{\partial t^2} - v_0^2\frac{\partial^2 V}{\partial z^2} = v_0Z_0\frac{\partial^2\rho}{\partial t^2}, \tag{1}$$

where v_0 is as defined above, Z_0 is the circuit characteristic impedance $E^2u_0^2/2\omega^2P$ which enters the definition of C, and $\rho(z, t)$ is the instantaneous linear charge density of the beam. To determine the behavior of ρ, on the other hand, we have the equation of motion of an electron in the form

$$\frac{d^2z}{dt^2} = -\frac{e}{m}\frac{\partial V}{\partial z}. \tag{2}$$

[7] Pierce, *op. cit.*, eq. (19).
[8] *Ibid.*, eq. (44) and (48).

The connection between the orbits which are solutions of (2) and the linear charge density which appears on the right of (1) is the following: Let z_0 be the position of an electron at $t=0$ so that the solutions of (2) have the form $z=f(t, z_0)$. Then if we consider all those electrons whose initial positions lie in the range z_0 to z_0+dz_0, we have on account of conservation of charge

$$\rho(z, t)dz = \rho(z, t)\left|\frac{\partial f}{\partial z_0}\right| dz_0 = \rho(z_0, 0)dz_0; \tag{3}$$

or if z_0 is taken to be the initial end of the tube, where the beam enters unmodulated, and if the beam current there is I_0 so that $\rho(z_0, 0)=I_0/u_0$,

$$\rho(z, t) = \frac{I_0}{u_0}\left|\frac{\partial z_0}{\partial z}\right|_t . \tag{4}$$

Here the relation $z=f(t, z_0)$ is thought of as inverted or solved for $z_0: z_0=g(t, z)$. There is a mathematical complication in connection with (4) when electrons overtake one another since z_0 is then not a single-valued function of z. That is, electrons arriving at the point z at the time t may have had several different positions z_0 at $t=0$. Equation (4) has then to be modified by replacing the quantity $|\partial z_0/\partial z|$ by the sum of its values for all branches of the multivalued function z_0. Fortunately, the numerical calculations are most readily made in terms of z_0 as independent variable (see (6) and (11′) below). Since z is always a single-valued function of z_0, the complication, provided it is understood, is not at all troublesome.

At large-signal levels the quantity ρ will be rich in harmonics even if the tube is excited with a single frequency, as it is in all the cases treated. The transmission-line amplitude will however be essentially all at fundamental frequency because Z_0 becomes small at frequencies double the design frequency and higher ($E^2u_0^2/2\omega^2P$ decreases rapidly with increasing frequency)[9] and because v_0, which is independent of frequency over the fundamental band of frequencies, will generally depend on the frequency at harmonic frequencies. Therefore, we neglect all but the fundamental frequency component on the right of (1). This permits us to represent $V(z, t)$ in terms of just two functions of z, both slowly varying, namely an amplitude and a phase. Except for this circumstance, the calculations which extend over approximately 50 to 100 complete cycles, would have been much too long to undertake.

We introduce reduced independent variables,

$$y \equiv C\frac{\omega z}{u_0} \tag{5}$$

$$\Phi_0 \equiv \omega\frac{z_0}{u_0}, \tag{6}$$

and reduced dependent variables defined by

$$\Phi \equiv \omega\left(\frac{z}{u_0} - t\right) - \theta(y) \tag{7}$$

$$V(z, t) \equiv \frac{Z_0I_0}{C} A(y)\cos\Phi \tag{8}$$

$$z \equiv \frac{u_0}{\omega}[\Phi(y, \Phi_0) + \theta(y)] + u_0t \tag{9}$$

$$\frac{dz}{dt} \equiv u_0[1 + 2Cq(y, \Phi_0)]. \tag{10}$$

y is distance along the tube in electrical radians, times $C(v_0/u_0)$; Φ is the actual instantaneous phase of the fundamental frequency wave in the circuit, as we see from (8); but if z is the coordinate of an electron, then Φ is also the phase of that electron with respect to the wave. $A(y)$ is a reduced RF voltage amplitude; in terms of A the power in the circuit at the point y is $(V^2/Z_0)_{av}=2CA^2I_0V_0$.

In terms of these variables we may now rewrite (1) and (2). The right side of (1) is, in keeping with the discussion above, to be replaced by its fundamental frequency part, which is

$$\begin{aligned}\frac{\omega^2v_0Z_0}{\pi}\Bigg[&\sin\Phi\int_0^{2\pi}d\Phi'\sin\Phi'\frac{\partial^2\rho(y, \Phi')}{\partial\Phi'^2}\\ &+\cos\Phi\int_0^{2\pi}d\Phi'\cos\Phi'\frac{\partial^2\rho}{\partial\Phi'^2}\Bigg]\\ = -\frac{\omega^2v_0Z_0}{\pi}\Bigg[&\sin\Phi\int_0^{2\pi}d\Phi'\sin\Phi'\rho(y, \Phi')\\ &+\cos\Phi\int_0^{2\pi}d\Phi'\cos\Phi'\rho\Bigg].\end{aligned} \tag{11}$$

The last equality follows from two integrations by parts. Equation (11) can be further simplified by use of (4), which in the reduced variables reads

$$\rho(y, \Phi) = \frac{I_0}{u_0}\left|\frac{\partial\Phi_0}{\partial\Phi}\right|_y \frac{1}{1+2Cq}. \tag{4'}$$

Expression (11) finally becomes (neglecting $2Cq$ compared to 1)

$$-\frac{\omega^2Z_0I_0v_0}{\pi u_0}\Bigg[\sin\Phi\int_0^{2\pi}d\Phi_0\sin\Phi(y, \Phi_0) + \cos\Phi\int_0^{2\pi}d\Phi_0\cos\Phi\Bigg], \tag{11'}$$

and is valid in this form, whether or not there is a multiple stream condition (overtaking). The left side of (1) takes the form

$$\frac{\omega^2Z_0I_0v_0}{u_0}\left[\sin\Phi\cdot 2\frac{dA}{dy} - \cos\Phi\cdot 2\left(b+\frac{d\theta}{dy}\right)A\right][1+O(C)], \tag{12}$$

[9] *Ibid.*, fig. 10.

where terms of relative order C have not been written out. Neglecting these terms for the time being and equating coefficients of sin Φ, respectively cos Φ, in (11′) and (12) gives the two working equations for the amplitude and phase of the signal in the circuit (consequences of (1)):

$$\frac{dA(y)}{dy} = \frac{-1}{2\pi}\int_0^{2\pi} d\Phi_0 \sin \Phi(y, \Phi_0) \tag{13}$$

$$\frac{d\theta}{dy} + b = \frac{1}{2\pi A(y)}\int_0^{2\pi} d\Phi_0 \cos \Phi(y, \Phi_0). \tag{14}$$

In order to re-form (2) in terms of the reduced variables, take the time derivative of (10) and combine it with (2) and replace d/dt by $(dy/dt)(\partial/\partial y)$.

$$2Cu_0\frac{dy}{dt}\frac{\partial q(y, \Phi_0)}{\partial y} = -\frac{e}{m}\frac{\partial V}{\partial z} \tag{15}$$

$$2C^2\omega u_0[1+2Cq]\frac{\partial q}{\partial y} = +\frac{e}{m}\frac{Z_0 I_0 \omega}{Cu_0} A(y) \sin \Phi[1+O(C)]. \tag{16}$$

Now q never becomes as large as $0(1/C)$, even at limiting power level, so that for small C we may approximate this by

$$\frac{\partial q(y, \Phi_0)}{\partial y} = A(y) \sin \Phi(y, \Phi_0). \tag{17}$$

The factors on both sides of (16) cancel because of the definition of C^3 and the relation $+2eV_0 = mu_0^2$. In order to get the last of the four working equations, take the time derivative of (9) and combine it with (10):

$$u_0[1 + 2Cq] = \frac{u_0}{\omega}\frac{dy}{dt}\left[\frac{\partial \Phi}{\partial y} + \frac{d\theta}{dy}\right] + u_0 \tag{18}$$

$$2Cu_0 q = Cu_0[1 + 2Cq]\left[\frac{\partial \Phi}{\partial y} + \frac{d\theta}{dy}\right] \tag{19}$$

$$\frac{\partial \Phi(y, \Phi_0)}{\partial y} = -\frac{d\theta}{dy} + 2q(y, \Phi_0). \tag{20}$$

To summarize, (13), (14), (17), and (20) are used to calculate $A(y)$, $\theta(y)$, $q(y, \Phi_0)$ and $\Phi(y, \Phi_0)$. They do not contain the parameter C and are approximations valid for $C \ll 1$ and for all values of y which are of interest (namely as far as the value of y at which A is a maximum and somewhat beyond). At $y=0$, the point of the tube where the integration is started, A, $d\theta/dy$, $q(0, \Phi_0)$, and $\Phi(0, \Phi_0)$ are taken from the small-signal increasing wave and the equations are integrated numerically in the direction of y increasing. The interval in Φ_0 is taken to be $\pi/8$, i.e., sixteen electrons of equally spaced initial phase are treated. The interval in y is taken to be 0.2 and third differences are effectively taken into account. $A(y)$ reaches its maximum at $y=6$ to $y=11$, depending on the value of b.

The small-signal solution of our equations, which is needed to determine the initial conditions for the integration, is found as follows: Assume

$$A(y) = A_0 e^{\alpha y}; \qquad \frac{d\theta}{dy} = \beta$$
$$q = -\gamma A_0 e^{\alpha y} \cos(\Phi_0 - \beta y + \delta) \tag{21}$$
$$\Phi = \Phi_0 + \beta y - 2\epsilon A_0 e^{\alpha y} \sin(\Phi_0 - \beta y + \zeta),$$

where, $\alpha, \beta, \cdots, \zeta$ are constants, then substitute into the working equations and retain only terms of the lowest order in A_0. This gives just enough relations to determine the six constants $\alpha, \beta, \cdots, \zeta$ in terms of the parameter b ((17) and (20) each provide two relations since they must hold identically in Φ_0). After some calculation these relations take the form $\zeta = 2\delta = 2\tan^{-1}\alpha/(-\beta)$; $\epsilon = \gamma^2 = 1/(\alpha^2+\beta^2)$, where α and β are roots of

$$(j\alpha + \beta)^2(j\alpha + \beta + b) = 1. \tag{22}$$

The real and imaginary parts of this equation are Pierce's equations (49) and (50); our α, β, b are his x, y, b, and we have zero circuit loss so that his $d=0$. We contemplate only that solution of (22) for which $\alpha > 0$ (increasing wave). Then for any chosen b the expressions (21) are explicitly calculable, except only that the value of A_0 must be decided upon. This value is chosen so as to limit the term $2q$, on the right of (20), to 5 per cent of the value of the other term $d\theta/dy$, at $y=0$. The largest nonlinear term in any of our equations is then limited to 5 per cent or less of the corresponding linear terms.

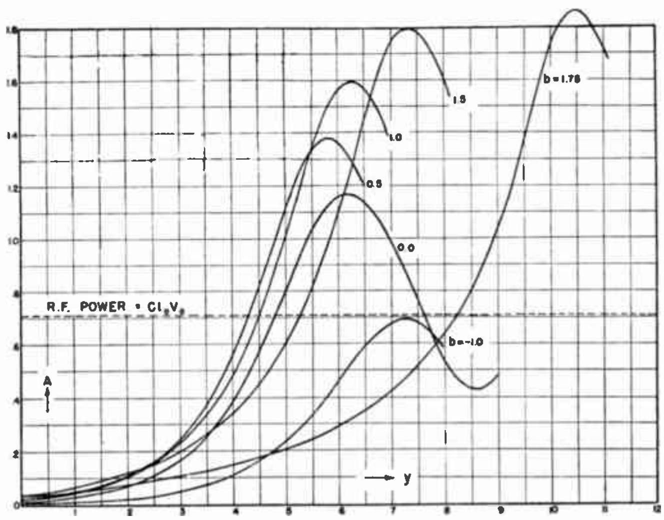

Fig. 1—RF voltage amplitude versus distance along tube. Parameter: injection speed.

III. Results for Amplitude and Phase for Small C

The amplitude $A(y)$ as a function of y for the six values of b: $b = -1.0$, 0, 0.5, 1.0, 1.5, and 1.75 is shown in Fig. 1. Since the RF power in the circuit at the point y is given by $2CA^2I_0V_0$, the physical meaning of the ordinate scale is evident. The numerical values of b were chosen for the following reasons: By reference to Pierce's Fig. 2,[10] we see that the small-signal gain per unit y

[10] *Ibid.*, fig. 2.

(our α, his x) increases slowly as b increases from negative values, becomes a maximum at $b=0$, and goes to zero at $b=1.87$. The limiting power for negative b is small, as expected, because the electrons are injected too slowly; therefore, this region is of less practical interest and only one negative value of b is treated. The region $b=0$ to $b=1.87$ is of great interest since α decreases there while the limiting power is expected to increase with increasing b. Therefore, this region is covered by five calculations. The one calculation for $b=0$ is carried far enough beyond the point of maximum A to exhibit a subsequent minimum and the beginning of a second region of increasing A. This oscillatory behavior of A after its first maximum, while probably of little practical interest, is predicted quite generally by the theory, as one can verify by studying the behavior, with increasing y, of the curve of q versus Φ.

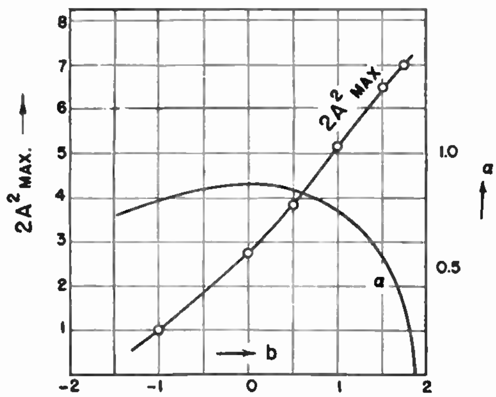

Fig. 2—Maximum efficiency versus injection versus injection speed.

The results of the calculations substantiate the guess that the limiting power increases with b as far as $b=1.75$. In Fig. 2, $2A^2_{\max}$ is plotted versus b in order to illustrate this point, with the curve of α versus b included for comparison. The small circles indicate the calculated points. Fig. 2 suggests that *large*-signal amplification may take place for $b \geqq 1.87$ even though small-signal amplification is not possible ($\alpha=0$).

The information of Fig. 1 is displayed over again in Fig. 3 on a decibel scale, with the zero of the scale fixed at CI_0V_0. These graphs illustrate a point relating to the initial departure of the radio-frequency voltage from the small-signal law of exponential increase, i.e., relating to initial compression or expansion. We see that for $b<1.0$ there is initial compression while for $b>1.0$ there is initial expansion. It can be shown analytically that if the nonlinear terms in the equations are taken into account approximately they predict initial compression for $b<1.06$ and initial expansion for $b>1.06$.[11] Therefore, we have a cross-check between the analytical and the numerical predictions.

[11] L. R. Walker, unpublished Bell Telephone Laboratories Memorandum No. 47-150-16.

The phase of the radio-frequency signal at various points in the tube is given by the graphs of Fig. 4. We plot not θ but the amount by which θ differs from its small-signal value βy. The actual signal lags behind a corresponding small signal by $\Delta\theta \equiv \beta y - \theta(y)$ radians at the point y.

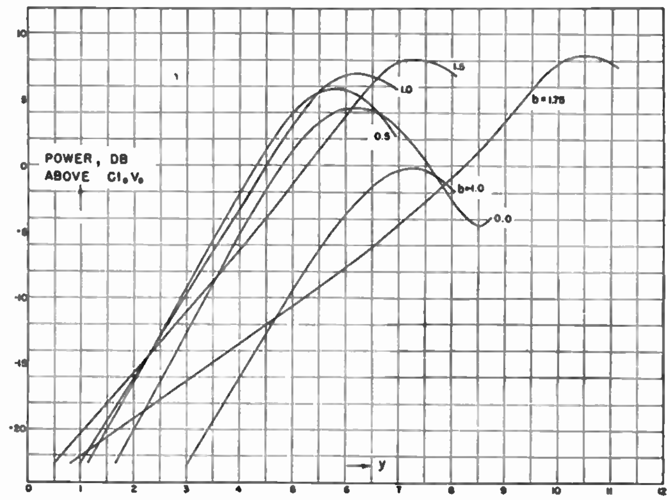

Fig. 3—RF power versus distance along tube. Parameter: injection speed.

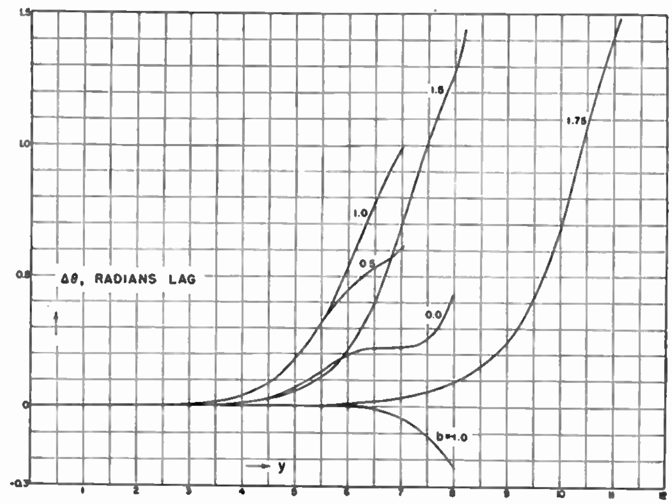

Fig. 4—RF phase (relative to small signal phase) versus distance along tube. Parameter: injection speed.

The curves of Fig. 3 and Fig. 4 are somewhat deceptive in one respect: they give a false idea of how far along in the tube the electron stream behaves linearly. One may describe the situation the other way round and say that the voltage amplitude increases exponentially and the phase retains its small-signal value relatively far beyond the point where the electron stream ceases to conform to the small-signal solution. To illustrate this point we give the values of y at which overtaking sets in: they are $y=6.0$, 4.8, 4.5, 4.9, 5.9, and 8.9 at $b=-1.0$, 0, 0.5, 1.0, 1.5, and 1.75, respectively. The compression and phase distortion at these values of y are seen to be generally very small. The author believes that intermodulation disturbances would set in much earlier, i.e., at smaller y or lower power output.

IV. Finite Beam Diameter and Finite C for $b=1.5$

The single value 1.5 of b was chosen for the more laborious calculations designed to check the effect of finite beam diameter and finite C because the limiting power is quite high at this value of b and is attained, at the same time, with a reasonably short tube (α is also large).

In order to treat the case of a beam of finite diameter, in which some electrons are nearer the circuit than others and hence experience stronger radio-frequency fields, we imagine the beam divided into two parts of equal direct current $I_0/2$, the radially inner part or core (called beam number 1) experiencing, on the average, a field half as large as the outer shell (called beam number 2). The ratio of the fields acting on the two beams could of course have been assumed to have any value; the value $\frac{1}{2}$ was arrived at by taking the ratio of the two average fields in a typical helical circuit used in practice, on the assumption that the total beam completely filled the interior of the helix. We treat the two partial beams just as the whole beam was treated in Section II, by integrating the equations of motion of 16 typical electrons of equally spaced initial phase in each beam. Thus the working equations corresponding to (13) and (14) are modified by each having two terms on the right, one term for each beam. The working equations (17) and (20), on the other hand, become double in number, holding for $q_1(y, \Phi_0)$ and $\Phi_1(y, \Phi_0)$ where the subscript 1 refers to beam 1 and again for q_2 and Φ_2 referring to beam 2.

A complication occurs in the definition of the reduced variables since C enters these definitions, and we now have two values of C, one for each beam. The C chosen for defining the reduced variables is

$$C \equiv \left[(Z_{01} + Z_{02}) \frac{I_0/2}{4V_0} \right]^{1/3}, \tag{23}$$

where Z_{01} is the characteristic impedance, defined as before, using the field acting on beam 1, Z_{02} similarly, and I_0 is the total beam current. This definition of C makes the small-signal gain the same as that of a single beam tube with the same C.

The working equations read

$$\left.\begin{aligned} \frac{dA}{dy} &= -\frac{1}{2\pi}\int_0^{2\pi} d\Phi_0\left[\tfrac{1}{2}f_1 \sin \Phi_1(y, \Phi_0) + \tfrac{1}{2}f_2 \sin \Phi_2\right] \\ \frac{d\theta}{dy} + b &= \frac{1}{2\pi A}\int_0^{2\pi} d\Phi_0\left[\tfrac{1}{2}f_1 \cos \Phi_1 + \tfrac{1}{2}f_2 \cos \Phi_2\right] \end{aligned}\right\} \tag{24}$$

$$\left.\begin{aligned} \frac{\partial}{\partial y} q_i(y, \Phi_0) &= f_i A(y) \sin \Phi_i(y, \Phi_0) \\ \frac{\partial}{\partial y} \Phi_i(y, \Phi_0) &= -\frac{d\theta}{dy} + 2q_i(y, \Phi_0) \end{aligned}\right\} i = 1, 2, \tag{25}$$

where $f_1 = \frac{1}{2}f_2$ and $f_1^2 + f_2^2 = 2$.

The results of this calculation are given in Fig. 5, together with the single-beam results for $b=1.5$ for comparison. The effect of having the beam spread out radially is evidently a considerable one, reducing the limiting power by a factor of 0.66 and broadening the maximum. The effects of reducing and broadening the maximum are explainable on the basis that beam 2 stops supplying power to the circuit and starts absorbing power from it at a different point y than does beam 1. From the calculations these two values of y are 6.9 and >8.3, respectively. The phase lag is seen to be comparatively little changed.

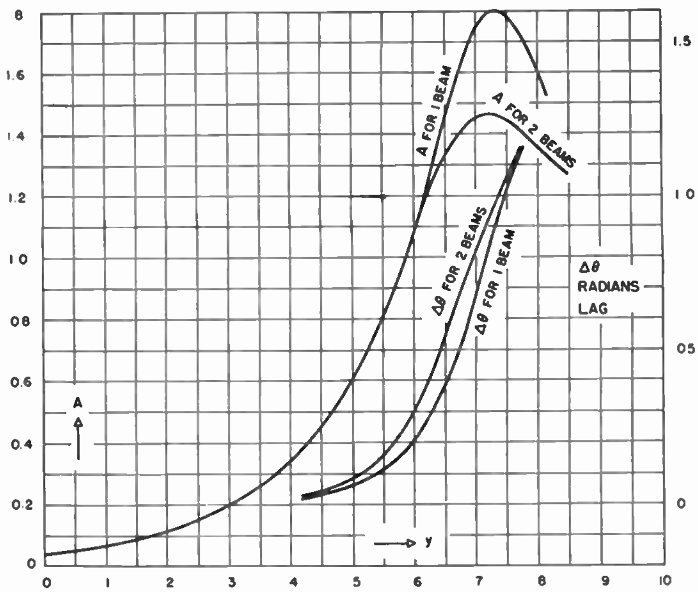

Fig. 5—Comparison of RF amplitude and phase for 1-beam and 2-beam approximations. Injection speed $b=1.5$.

We turn now to the calculation for finite C, which is again made for $b=1.5$, and which was stimulated by the question, What is the maximum efficiency (i.e., ratio of maximum radio-frequency power to I_0V_0) at a value of C which is attainable in practice but not small compared to unity? We choose $C=0.1$ as such a value. According to the "small C" calculations of Sections II and III, the maximum efficiency should be 6.5 $C=65$ per cent (see Fig. 2). This is a surprisingly high figure for such a weak coupling between beam and circuit, and we want to know how much this figure must be corrected on account of C being finite.

If one derives the working equations for finite C following exactly the steps of Section II, the correction terms in the resulting working equations are numerous and complicated. It is found worth while in this case to use a somewhat different set of variables, as follows: Equation (9) constitutes a transformation to a co-ordinate system moving with the actual wave; for the present purposes we transform instead to a co-ordinate system moving at the constant speed v_0, which is the phase velocity in the circuit in the absence of electrons.

In place of (5) to (10) we have

$$y \equiv C \frac{\omega z}{v_0} \tag{5'}$$

$$\phi_0 \equiv \frac{\omega z}{u_0} \quad (6')$$

$$\phi \equiv \omega\left(\frac{z}{v_0} - t\right) \quad (7')$$

$$V(z, t) \equiv \frac{Z_0 I_0}{4C}\left[a_1(y)\cos\phi - a_2(y)\sin\phi\right] \quad (8')$$

$$z \equiv \frac{v_0}{\omega}\phi(y, \phi_0) + v_0 t \quad (9')$$

$$\frac{dz}{dt} \equiv v_0\left[1 + Cw(y, \phi_0)\right], \quad (10')$$

and in place of (4′) we have

$$\rho(y, \phi) = \frac{I_0}{v_0}\left|\frac{\partial\phi_0}{\partial\phi}\right|_y \frac{1}{1 + Cw}. \quad (4'')$$

After considerable calculation one finds the following working equations:

$$\frac{da_1(y)}{dy} = -\frac{2}{\pi}\int_0^{2\pi} d\phi_0 \frac{\sin\phi(y, \phi_0)}{1 + Cw(y, \phi_0)} \quad (13')$$

$$\frac{da_2(y)}{dy} = -\frac{2}{\pi}\int_0^{2\pi} d\phi_0 \frac{\cos\phi(y, \phi_0)}{1 + Cw(y, \phi_0)} \quad (14')$$

$$\frac{\partial}{\partial y}\left[1 + Cw(y, \phi_0)\right]^2 = C\left[\left(a_1 + C\frac{da_2}{dy}\right)\sin\phi + \left(a_2 - C\frac{da_1}{dy}\right)\cos\phi\right] \quad (17')$$

$$\frac{\partial}{\partial y}\phi(y, \phi_0) = \frac{w(y, \phi_0)}{1 + Cw}. \quad (20')$$

This is the simplest form of the equations valid for arbitrary C. There is evidently still substantially more labor involved in integrating these than in integrating the "small C" equations. The quantities $A(y)$ and $\theta(y)$ of Section II are related to $a_1(y)$ and $a_2(y)$ by

$$4A(y) = \left[a_1^2(y) + a_2^2(y)\right]^{1/2} \quad (26)$$

$$\tan\left(-\theta(y) - by\right) = a_2(y)/a_1(y). \quad (27)$$

Results of the integration of the "finite C" equations (13′), (14′), (17′), and (20′) for $C = 0.1$ and $b = 1.5$ are shown in Fig. 6, together with the infinitesimal C results for the same injection speed for comparison. The limiting efficiency turns out to be 53 per cent as compared with the 65 per cent extrapolated from the infinitestimal C calculation, a gratifyingly small decrease. The behavior of the phase is actually improved by the increase in C.

A rough analytical formula may now be given for the maximum efficiency as a function of b and C. For small C the maximum efficiency is quite closely $(2.75 + 2.42b)C$ from Fig. 2. For finite C we multiply this by a factor $(1 - g(b)C)$, where $g(1.50) = 1.85$ as determined by the calculation just described. If we ignore the unknown dependence of g on b, we get

$$\text{max efficiency} \equiv \eta(b, C) = (2.75 + 2.42b)(C - 1.85C^2). \quad (28)$$

This formula represents a rough upper limit to practically attainable efficiencies because, among other things, the finite beam diameter effect (beginning of this section) is neglected in it.

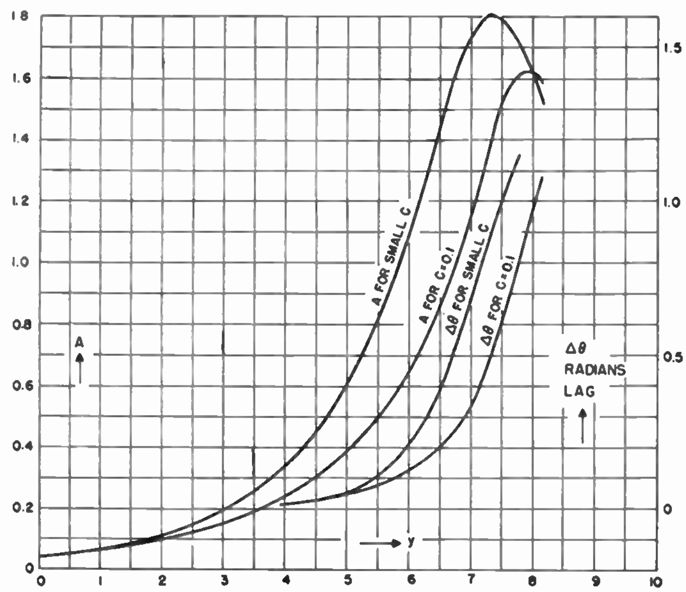

Fig. 6—Comparison of RF amplitude and phase for very small C and for $C = 0.1$.

V. Harmonic Content of the Electron Stream

We may expect the harmonic content of the electron current to be considerable in the traveling-wave amplifier at large-signal levels, as in all velocity-modulation tubes. Rather crude information on this harmonic content is available as a by-product of the calculations described above. The information is crude for at least two reasons: The subdivision of the range of Φ_0 into only sixteen parts makes the computation of the higher Fourier coefficients inaccurate; and space-charge forces, which we have neglected entirely, may be expected to smooth out the high concentrations of charge which occur at points of "overtaking" and which contribute considerably to the large values of the higher-order Fourier coefficients in our treatment. Therefore, the results presented in Fig. 7 must be regarded as order-of-magnitude results only and as overestimates.

The waveform of the electron current in the tube at the point y is given by

$$I_y(t) = \rho(z, t)\frac{dz}{dt} = I_0\left|\frac{\partial\Phi_0}{\partial\Phi}\right|. \quad (29)$$

The Fourier expansion of this current wave is

$$I_y(t) = \sum_{n=0}^{\infty} i_n \cos n\Phi + \sum_{n=1}^{\infty} j_n \sin n\Phi. \quad (30)$$

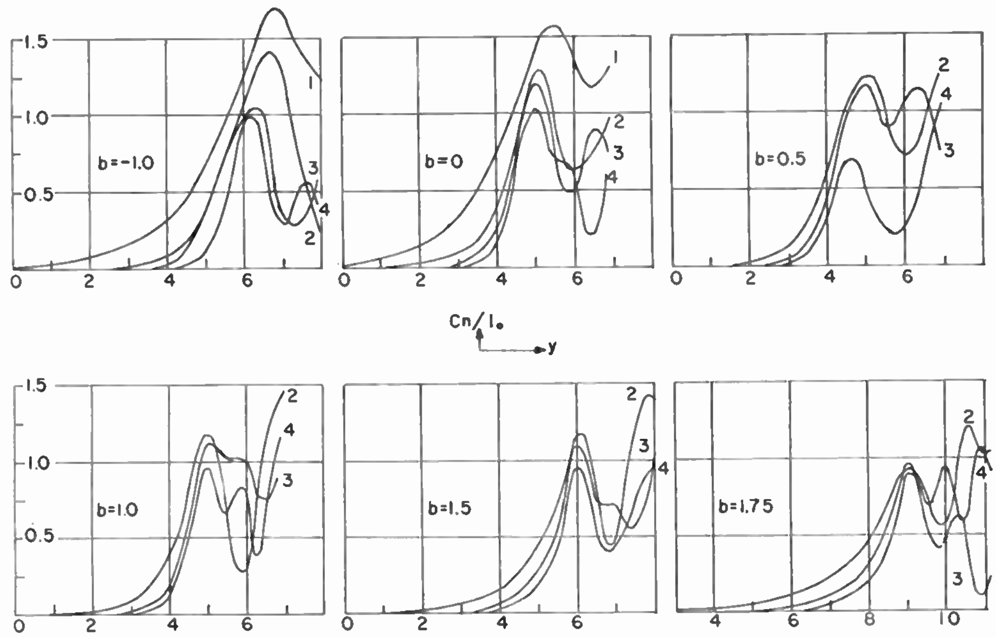

Fig. 7—Harmonic content of electron stream. Curves are labelled with n.

Hence the magnitude of the n^{th} harmonic component of the beam current at the point y in the tube is (for $n>0$).

$$c_n(y) = \sqrt{i_n^2 + j_n^2} = I_0\left[\left(\frac{1}{\pi}\int_0^{2\pi} d\Phi_0 \cos n\Phi\right)^2 + \left(\frac{1}{\pi}\int_0^{2\pi} d\Phi_0 \sin n\Phi\right)^2\right]^{1/2}. \qquad (31)$$

Since $\Phi(y, \Phi_0)$ was already calculated it was a comparatively minor chore to calculate the c_n. In formula (29) we must again understand that if Φ_0 is a multivalued function of Φ the sum of the magnitudes of $\partial\Phi_0/\partial\Phi$ for all branches of the function must be taken. Formula (31) on the other band is valid without modification even where overtaking has occurred.

In Fig. 7 we plot c_n/I_0 versus y for $n=1, 2, 3, 4$ and for the six injection velocities treated. c_1 is shown for only two values of the injection velocity since it is not of special interest. c_5 was also calculated but has not been included in Fig. 7 in order to avoid too much confusion of the curves; it behaves qualitatively like c_4.

By way of general qualitative conclusions from the data of Fig. 7, we observe that all the harmonics up to quite high order (including $n=5$ at least) have first amplitude peaks, of the order of unity, together near the point along the tube where overtaking begins (compare the values of y for initial overtaking as given near the end of Section III). Thereafter, the harmonics fluctuate irregularly between zero and approximately unity.

Acknowledgments

The author gratefully acknowledges the aid of Mrs. C. E. Shannon, Misses F. M. Metz and R. A. Weiss, and Dr. R. W. Hamming, all of the Bell Laboratories Mathematics Department, who carried out the numerical computations.

Correction

A. J. Simmons, author of the paper, "A Compact Broad-Band Microwave Quarter-Wave Plate," which appeared on pages 1089–1090 of the September, 1952 issue of the Proceedings of the I.R.E., has brought the following correction to the attention of the editors:

A term is missing in equation (2a) which appears in column two on page 1089. Correct equation should be

$$t_{11} = \cos 2\beta l - \frac{B_1'}{Y_0}\sin 2\beta l - \frac{B}{Y_0}\sin \beta l_1 \cos \beta l_1' + \frac{B_1'B}{Y_0^2}\sin \beta l_1 \sin \beta l_1'.$$

The Germanium Diode as Video Detector*

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Summary—**Methods are discussed for measuring those characteristics of a germanium diode which are important in video-detector application. Some of the measuring techniques described were designed for rapid production-line checks. A theoretical analysis of the operation of the crystal diode as a video detector is presented. The forward and back conductances have been assumed constant over the range of operation, and loads with both small and large time constants have been considered.**

Introduction

During the research on fundamentals of television receivers and associated circuit design the many advantages of using germanium crystal diodes as video detectors were once again realized.[1] The germanium diode is particularly suited to video-detector application because of its high forward conductance, implying low losses, and its low intrinsic capacitance and high back resistance, giving good wide-band operation. The absence of a filament in this type of diode eliminates one source of ripple voltage, and since the element is small, it can be wired into a circuit as easily as an ordinary resistor.

Measurements of Germanium Diode Properties

To determine the suitability of a germanium diode for video-detector application, measurement may be made of the following.

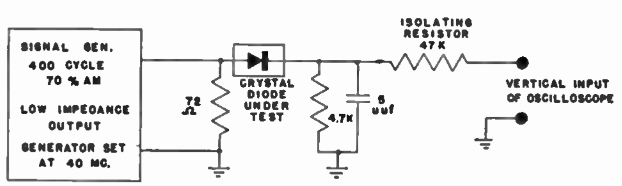

Fig. 1—Circuit for measurement of dynamic efficiency.

A. Dynamic Efficiency as Video Detector

The arrangement shown in Fig. 1 is used to measure the "dynamic efficiency" of the diodes for video detection. A measurement made in this manner is the most important specification of the 1N60 diode since it simulates actual use as a video detector. A constant input signal, 1.8 rms, is fed into the crystal under test. This signal is 70 per cent amplitude modulated at 400 cps, and the output measured on an oscilloscope. If the output voltage does not fall within certain prescribed limits, the crystal is rejected.

B. Current Values at Specified Voltages

A measurement of the back current, I_B, which corresponds to a fixed voltage across the crystal may be made with the circuit shown in Fig. 2. When the voltmeter, V, is connected as shown, the voltage drop across the ammeter is negligible compared with the high voltage drop across the crystal. Acceptance limits for I_B may be marked directly on the ammeter.

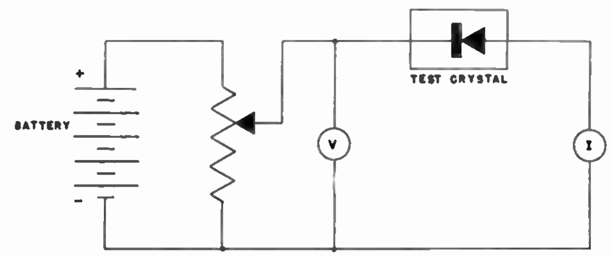

Fig. 2—Circuit for measurement of back current.

The forward current, I_F, (again at some fixed voltage) may be measured with a similar circuit with the crystal reversed. In this case, however, the voltmeter is connected directly across the crystal since the current flow in the meter is small compared with that in the high-conductance crystal.

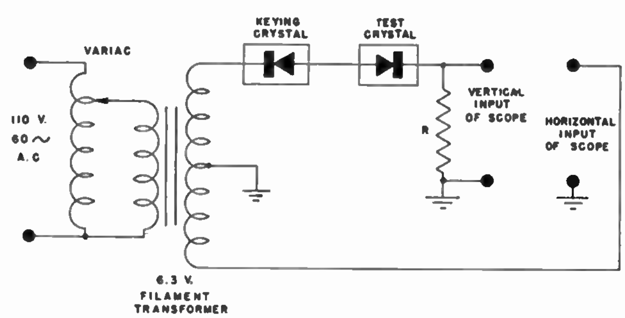

Fig. 3—Circuit for measurement of forward and back characteristics.

C. Back and Forward I-E Characteristics

For a rapid check of the static crystal characteristics a 60-cycle sweep test, using the circuit shown in Fig. 3, was devised. The test crystal is shown in position for checking the linearity of the back characteristic. For checking the forward characteristic the keying crystal is short-circuited and a much smaller load resistance is used.

Oscilloscope pictures would be similar to curves in Figs. 4 and 5. Voltage applied and gain of oscilloscope

* Decimal classification: R362.1×R538. Original manuscript received by the Institute, May 26, 1951; revised manuscript received October 29, 1952.

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[1] W. B. Whalley, "Simplification of television receivers," *Sylvania Tech.*, vol. III, pp. 9–12; January, 1950; Proc. I.R.E., vol. 38, pp. 1404–1408; December, 1950; Proc. IRE (Australia), vol. 13, pp. 99–103; April, 1952.

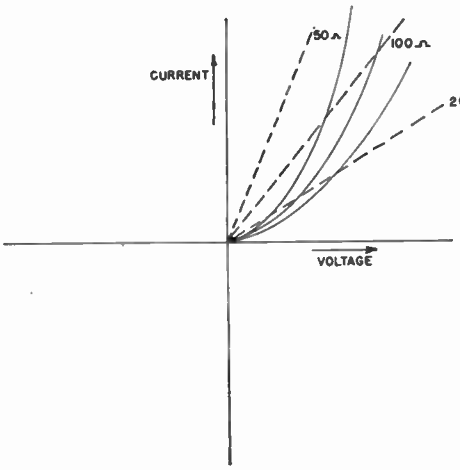

Fig. 4—Typical oscillogram sketch of dc back characteristics.

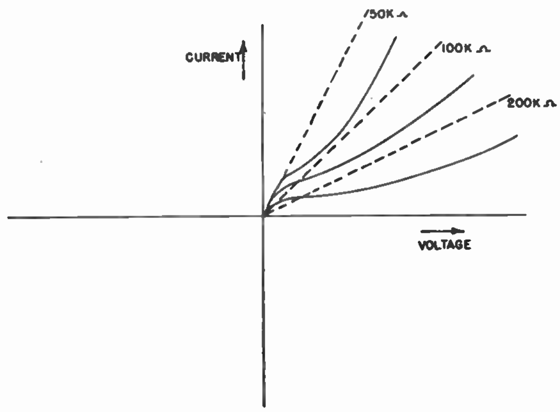

Fig. 5—Typical oscillogram sketch of dc forward characteristics.

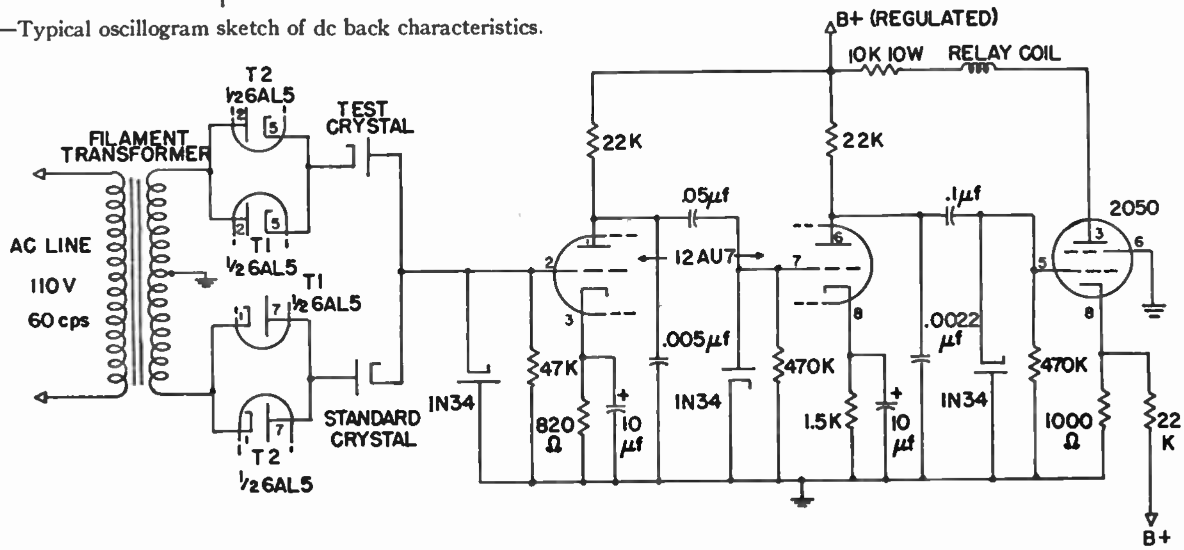

Fig. 6—Germanium-diode back characteristic comparator.

are set for portion of static characteristic which is of interest.

For rapid automatic checking of static crystal characteristics, the diode to be tested may be inserted into a balanced amplifier bridge for comparison with a standard diode and suitable acceptance—rejection limits set. A circuit for this type of measurement is shown in Fig. 6.

D. Diode Capacitance

The static crystal capacitance can be measured with the circuit shown in Fig. 7. The purpose of the battery with its necessary by-pass is to maintain a measurable value of Q. The procedure is as follows: With the point A disconnected, the coil is resonated and the capacitance reading recorded; point A is reconnected and the coil resonated again by means of the variable capacitor of the Q meter. The difference in capacitance readings is the effective cold (nonconductive) capacitance of the crystal under test. For a production check, acceptance limits on samples may be set for this quantity.

E. Demodulation Linearity of Diode as Video Detector

To obtain a measure of the linearity of the crystal as a demodulator, the circuit shown in Fig. 8 was designed. The resultant picture on the oscilloscope would be a curve varying in linearity, depending on the characteristics of the crystal under test.

Study of Dynamic Performance

When crystal diodes are used as second detectors in TV receivers, experience has shown that, under the condition of constant input signal level, a normal over-all response characteristic of the IF amplifier will be obtained only with crystals having the proper characteristics. A qualitative measure of the effect of different crystal characteristics on an IF amplitude response curve can be obtained with the circuit shown in Fig. 9.

The IF amplifier is aligned to give the standard response curve with a bogey crystal. This response curve is shown in Fig. 10(a). Fig. 10(b) and (c) show response curves obtained with crystals having, for example, an undesirable characteristic in the back direction resulting in improper loading and detuning of the last IF stage.

Fig. 11 is a circuit arrangement for determining quantitatively the effect of varying signal voltage[2] upon the

[2] It should be noted that in a well-designed receiver having good AGC the signal level at the input to the second detector is constant.

response curve of a single IF tank circuit with a germanium diode and usual video load. The signal generator is initially set at 40 mc. The tank circuit is tuned to resonance at this frequency, with the voltage at point A maintained at 2.5 volts rms. When this voltage is changed to 0.6 volt rms the tank circuit is observed to be no longer in tune. The frequency and output level of the signal generator are now varied until the tank circuit is again in resonance with the voltage at point A maintained at 0.6 volt.

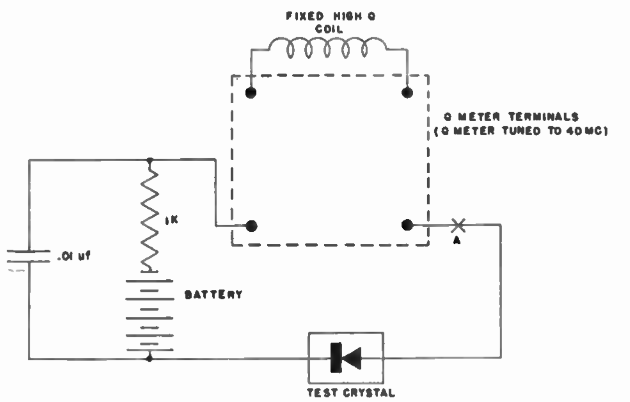

Fig. 7—Circuit for measurement of diode capacitance.

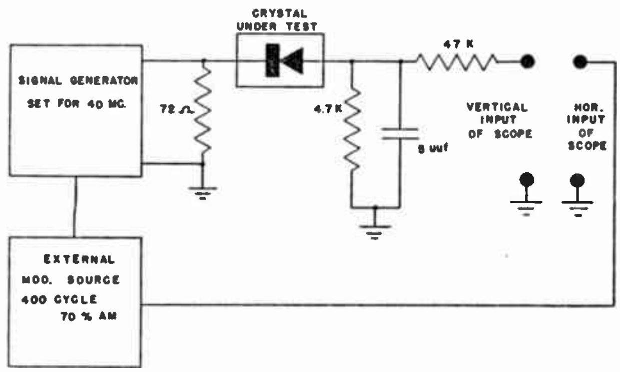

Fig. 8—Circuit for measurement of demodulation linearity.

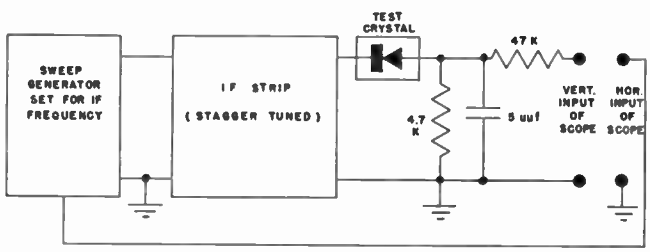

Fig. 9—Circuit for measurement of "IF" response curve.

This change in the resonant frequency of the tank circuit with signal level is caused by a change in the input capacitance of the detector circuit. This is primarily due to the effective ac resistance of the crystal diode changing with level and coupling more or less capacitance back to the detector input circuit.

A variation of the above method would be to replace the fixed-frequency generator by a sweep generator set at 40 mc with a 10-mc sweep.

A marker generator is used to mark the center frequency of the response curve. With this circuit arrangement the generator center frequency is held constant. If a measure of the capacitance change in the detector circuit is desired, C may be varied to bring the center frequency of the tank back to its original frequency as the signal level is changed. The change in C represents

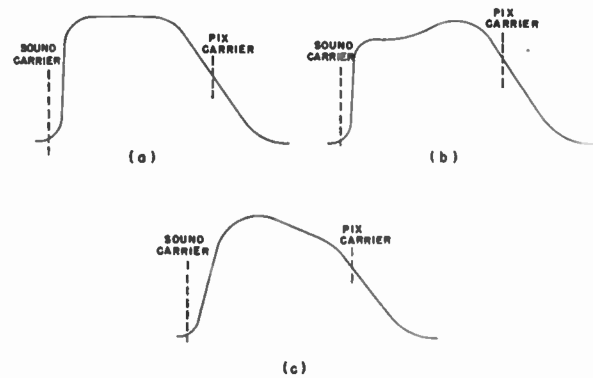

Fig. 10—Some "IF" response curves.

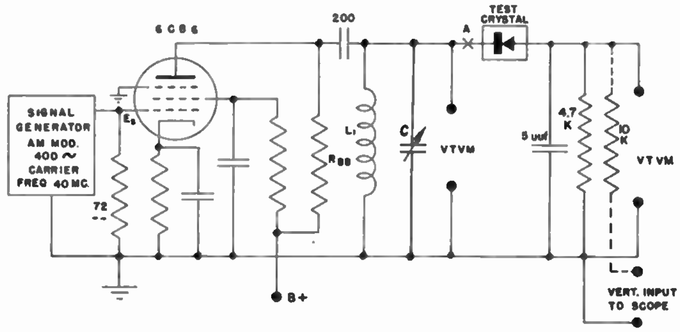

Fig. 11—Circuit for determination of input impedance of detector circuit.

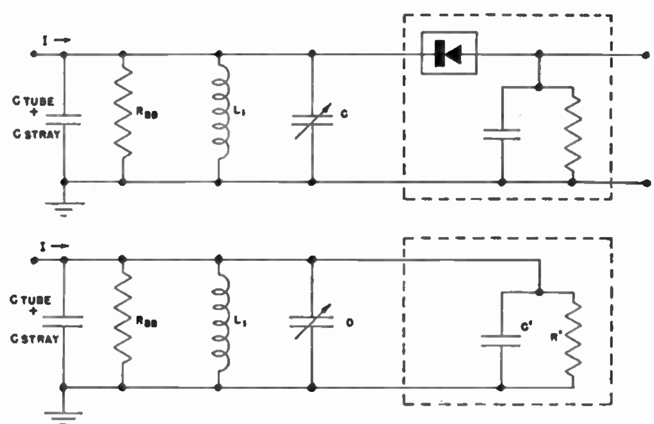

Fig. 12—Equivalent circuits for Fig. 11.

the change in the input capacitance of the detector circuit with change in signal level. In addition, the actual values of C_1 and R_1, the equivalent input capacitance and input resistance as shown in Fig. 12, may be determined by the following procedure:

Disconnect the detector from point A in Fig. 11 and tune C for maximum voltage. Set the voltage at point A at the desired level for the measurement. The grid voltage which corresponds to this level is recorded as e_{s1}. Point A is then reconnected and the capacitance C across L_1 is decreased until the tank is in resonance. The voltage at point A is held at the desired level by changing the grid voltage to e_{s2} while the tank is kept in resonance. The resultant change in C is equal to C_1. R_1 can be determined by means of the following formula:

$$R_1 = \frac{R_s}{\dfrac{e_{s2}}{e_{s1}} - 1},$$

where R_s is the resonant impedance of the tank with point A disconnected. If R_s is determined by the usual methods, the grid voltage may be calibrated in terms of R_1. Once R_1 and C_1 are known, the effective values of R and C of the crystal itself may be determined if desired.

Theoretical Analysis

In the analysis which follows a solution will be presented of the idealized case of a crystal of constant forward and constant back resistance operating into a known dc load resistance which is by-passed for the IF frequency. This solution is presented in universal curves which specify the rectification efficiency and the detector-circuit input resistance as functions of the constants of the crystal and its load. A figure of merit is developed for the crystal as a video detector which is indicative of the performance obtainable with a given crystal. Universal curves are presented which give this figure of merit as a function of the constants of the crystal and its load. The influences of the capacitance of the diode and the incomplete by-passing of the fundamental of current are discussed under the assumption that voltages of frequencies higher than first harmonic do not appear at the output.

A. Rectification Efficiency and Input Resistance with Loads Having Large Time Constants

The first approach to the video detector problem uses the circuit of Fig. 13 and assumes:

1. The driving voltage is a cosine wave from a zero impedance source.
2. The crystal forward conductance, G_F, and the back conductance, G_B, *are constant over the range of operation.*
3. The load voltage is pure dc. The efficiency of rectification, ρ, is defined by

$$\rho \equiv \frac{V_0}{V_1} = \frac{I_0 R_0}{I_1 R_1} = \cos\theta, \qquad (1)$$

so that

$$\frac{R_1}{R_0} = \frac{1}{\rho}\frac{I_0}{I_1} = \frac{1}{\cos\theta}\frac{I_0}{I_1}. \qquad (2)$$

In Fig. 13, the voltage V_x across the crystal is

$$V_x = V_1(\cos x - \cos\theta). \qquad (3)$$

The dc and the fundamental frequency relationships

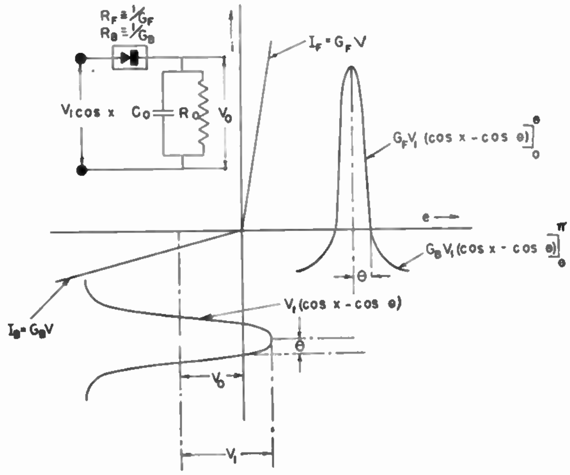

Fig. 13—Operation of video-detector-circuit large time constant load.

are easily found from the usual Fourier spectral analysis to be as follows:

$$I_0 = \frac{V_1}{\pi}\left[G_F \int_0^\theta (\cos x - \cos\theta)dx + G_B \int_\theta^\pi (\cos x - \cos\theta)dx\right]$$

$$= \frac{V_1}{\pi}\left[(G_F - G_B)(\sin\theta - \theta\cos\theta) - G_B\pi\cos\theta\right],$$

and, since $I_0 = G_0 V_0$,

$$\frac{G_0 + G_B}{G_F - G_B} = \frac{\tan\theta - \theta}{\pi}. \qquad (4)$$

Turning now to consideration of fundamental frequency,

$$I_1 = \frac{2V_1}{\pi}\left[G_F \int_0^\theta (\cos x - \cos\theta)\cos x\,dx + G_B \int_\theta^\pi (\cos x - \cos\theta)\cos x\,dx\right]$$

$$= \frac{V_1}{\pi}\left[(G_F - G_B)(\theta - \cos\theta\sin\theta) + G_B\pi\right].$$

Whence

$$G_1 \equiv \frac{I_1}{V_1} = \frac{(G_F - G_B)(2\theta - \sin 2\theta) + 2\pi G_B}{2\pi}. \qquad (5)$$

Equation (4) is an expression[3] which implicitly determines θ (and thus ρ) as a function of the circuit constants G_0, G_F and G_B. From a knowledge of θ and these

[3] A form of this equation first appeared in Sylvania Memorandum "On the Rectification Efficiency of, and the Circuit Loading Due to a Crystal Detector," by A. A. Grometstein; April, 1950.

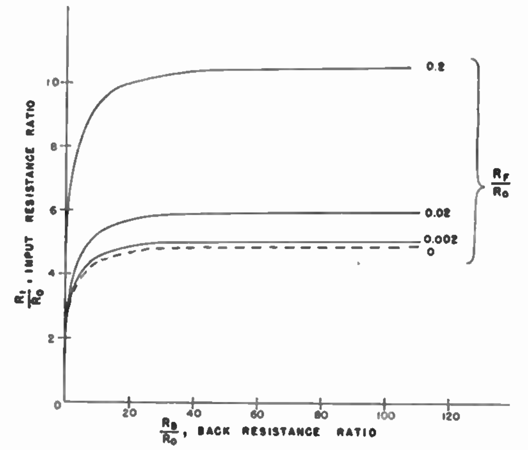

Fig. 14—Detector-circuit input resistance as a function of back resistance, for constant-load resistance.

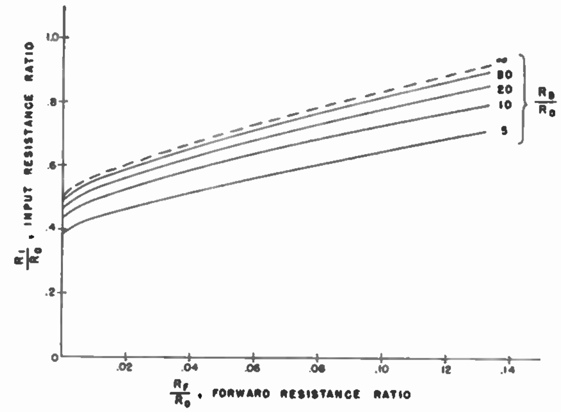

Fig. 16—Detector-circuit input resistance as a function of forward resistance, for constant-load resistance.

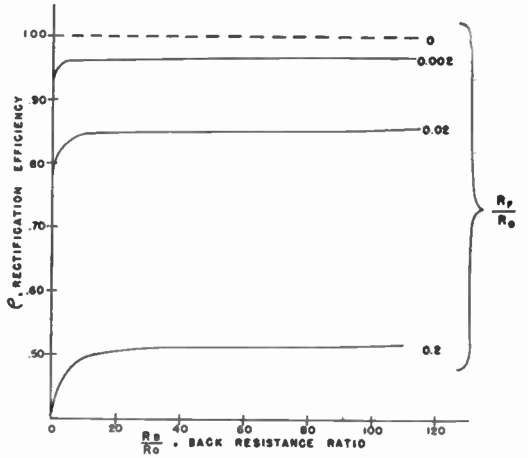

Fig. 15—Efficiency of rectification as a function of back resistance, for constant-load resistance.

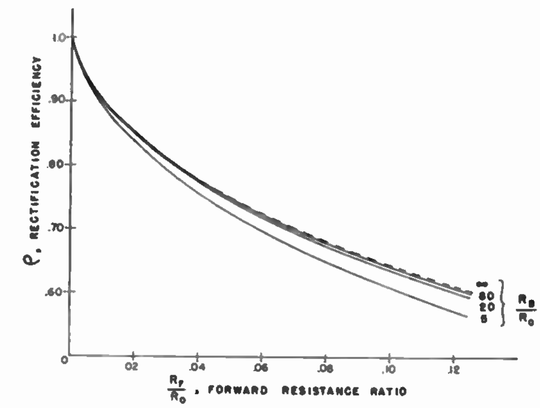

Fig. 17—Efficiency of rectification as a function of forward resistance, for constant-load resistance.

circuit constants the detector-circuit input resistance, $R_1 \equiv 1/G_1$, may be determined by (5).

To determine the relative importance of the circuit parameters under ordinary circuit operation, the above equations may be rewritten in several forms, two of which are presented below.

SET I

Normalized with Respect to Load Resistance (for Varying Back Resistance)

$$\frac{R_1}{R_0} = \frac{2(\tan\theta - \theta + \pi)}{(2\tan\theta - \sin 2\theta)(R_0/R_F) + (2\theta - \sin 2\theta - 2\pi)}$$

$$\rho = \cos\theta$$

$$\frac{R_B}{R_0} = \frac{(R_F/R_0)(\tan\theta - \theta + \pi)}{\tan\theta - \theta - \pi R_F/R_0}.$$

R_1/R_0 and the rectification efficiency p can be plotted as functions of R_b/R_0, with R_f/R_0 as a parameter. The two families of curves are shown in Figs. 14 and 15. If the forward resistance is taken as variable, then the following equation results.

SET II

Normalized with Load Resistance (for Varying Forward Resistance)

$$\frac{R_1}{R_0} = \frac{2(\tan\theta - \theta)}{(R_0/R_B)(2\tan\theta - \sin 2\theta) + (2\theta - \sin 2\theta)}$$

$$\rho = \cos\theta$$

$$\frac{R_F}{R_0} = \frac{(R_B/R_0)(\tan\theta - \theta)}{(\tan\theta - \theta) + \pi(1 + R_B/R_0)}.$$

Figs. 16 and 17 show the input-resistance ratio and the rectification efficiency as functions of the forward-resistance ratio with R_b/R_0 as parameter.

B. *Figure of Merit—Pure Resistance Loading of Tank Circuit*

The relative importance of the factors derived will be considered in the application of the crystal to the common detector circuit.

It is assumed that the input tank circuit to the last IF stage has been designed and that a suitable last IF tube has been chosen. By virtue of amplifier action, a given voltage appears at the IF amplifier grid. The desired bandwidth of the last IF plate circuit is assumed, known from prior considerations. The best crystal for

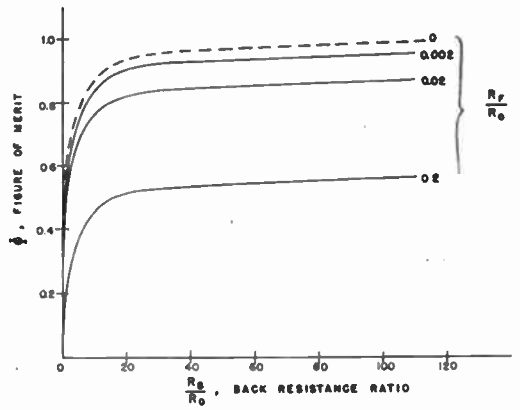

Fig. 18—Germanium-diode figure of merit as a function of back resistance.

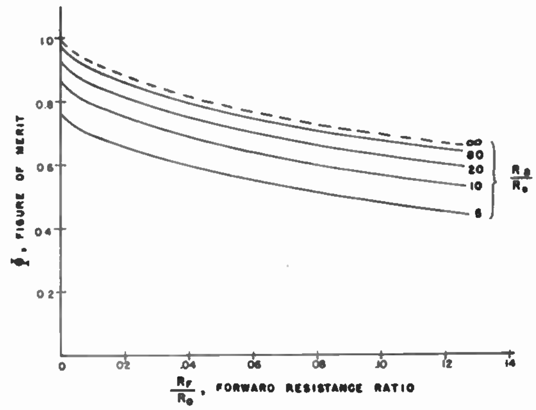

Fig. 19—Germanium-diode figure of merit as a function of forward resistance.

the application, all other things being equal, is the one which will produce the maximum output voltage in a given load over the required bandwidth. The output capacitance of the last IF tube, C_0 dictates the choice of coil. In general, to obtain the desired IF bandwidth it would be necessary to "tap in," so that the detector circuit is responsible for the entire loading.

The impedance of the last IF tank circuit at resonance is n^2R_1, where n is the turns ratio and R_1 the detector-circuit input resistance. The plate voltage developed is $G_mE_{g1}n^2R_1$ and the voltage across the detector input circuit is $G_mE_{g1}nR_1$. Hence the dc output voltage is $G_mE_{g1}nR_1\rho$, and grid-grid gain =

$$\frac{\text{dc output at video amp grid}}{\text{peak ac input to last IF grid}} = G_mnR_1\rho. \qquad (6)$$

The turns ratio is given by

$$n = \sqrt{\frac{1}{R_1C_0\Delta\omega}}\,.$$

Substituting in (6),

$$\text{grid-grid gain} = \frac{\left(\dfrac{G_m}{\sqrt{C_0}}\right)(\rho\sqrt{R_1})}{\sqrt{\Delta\omega}}$$

$$= \sqrt{\frac{\text{(tube merit)(crystal merit)}}{\text{bandwidth}}}\,.$$

Since $\rho\sqrt{R_1}$ is a positive quantity and the load resistance is assumed to be fixed, it is more convenient to define the crystal figure of merit as

$$\phi \equiv 2\rho^2\,\frac{R_1}{R_0}\,. \qquad (7)$$

Physically, the crystal figure of merit

$$= \frac{\text{dc power out of diode}}{\text{ac power into diode}}\,.$$

This figure of merit is plotted for varying back and varying forward resistance in Figs. 18 and 19, respectively.

C. Rectification Efficiency and Input Resistance with Diode Capacitance and Small Time Constant

The above concepts must be extended if the input impedance of the detector circuit is not a pure resistance i.e., if the capacitance of the crystal diode is included or the load capacitor is not a perfect by-pass. The following assumptions lead to a first-order solution of the problem.

1. The driving voltage is a cosine wave from a zero impedance source.
2. The forward conductance, G_f, and the back conductance, G_b, are constant over the range of operation.
3. The load voltage is dc plus the fundamental frequency component; the load capacitor (unrealistically) is assumed to be a perfect by-pass to all higher frequency components.

The circuit of Fig. 13 may be used if the capacitance C_x is considered to be across the diode and $V_f\cos(x-\phi)$ appears with the dc, V_{01}, as the voltage across the load. The voltage across the diode is then given by

$$\begin{aligned} V_x &= V_1\cos x - V_f\cos(x-\phi) - V_{01} \\ &= \overline{V}[\cos(x+\beta) - \cos\theta'], \end{aligned} \qquad (8)$$

where

$$\overline{V} = V_1\sqrt{1+\left(\frac{V_f}{V_1}\right)^2 - 2\,\frac{V_f}{V_1}\cos\phi},$$

$$\beta = \tan^{-1}\left[\frac{V_f\sin\phi}{V_1 - V_f\cos\phi}\right],$$

$$\frac{V_{01}}{\overline{V}} = \cos\phi' = \frac{V_{01}/V_1}{\sqrt{1+\left(\dfrac{V_f}{V_1}\right)^2 - 2\,\dfrac{V_f}{V_1}\cos\phi}}$$

$$\rho = \frac{V_{01}}{V_1}\,.$$

It will be observed that (8) differs from (3) in the following respects:

(1) the amplitude has been multiplied by a factor,
(2) there is an advance in phase, and
(3) the angle of flow is now θ' instead of θ.

Neither the difference in amplitude nor the difference in phase will have any bearing on the effective ac diode resistance under the assumptions of linear forward and back conductance. Therefore,

$$\frac{G_0 + G_B}{G_F - G_B} = \frac{\tan\theta' - \theta'}{\pi},$$

$$R_1' = \frac{2\pi}{(G_F - G_B)(2\theta' - \sin 2\theta') + 2\pi G_B}.$$

A comparison of these equations with (4) and (5) shows that R_1', the effective ac resistance of the diode, is equivalent to the input resistance R_1 in the case of a completely by-passed load. Thus, for the case in point, once R_1 is known, the input resistance and input capacitance are easily calculated by standard ac methods. The rectification efficiency then becomes,

$$\rho = \cos\theta' \sqrt{1 + \left(\frac{V_f}{V_1}\right)^2 - 2\left(\frac{V_f}{V_1}\right)\cos\phi} \qquad (1)'$$

where V_f/V_1 is the ac voltage transfer characteristic and ϕ is the phase angle of the load.

D. *Figure of Merit Tank Circuit Loading Not Purely Resistive*

If the detector circuit represents capacitive loading on the tank circuit, the figure of merit given above must be modified. Once the detector-circuit input resistance R_1 and the input capacitance C_1 have been determined, then:

$$\frac{n^2 R_1}{X_{L0}} = Q = \frac{\omega_0}{\Delta\omega},$$

$$L\left(C_0 + \frac{C_1}{n^2}\right) = \frac{1}{\omega_0^2},$$

where C_1 is the input capacitance of the detector circuit. Thus the turns ratio is given by

$$n_c = \sqrt{\frac{1 - \Delta\omega R_1 C_1}{\Delta\omega R_1 C_0}} = n\sqrt{1 - \Delta\omega R_1 C_1}.$$

Substituting in (6), as above,

$$\text{grid-grid gain} = \frac{\left(\frac{G_m}{\sqrt{C_0}}\right)(\rho\sqrt{R_1}\sqrt{1 - \Delta\omega R_1 C_1})}{\sqrt{\Delta\omega}}.$$

Thus the crystal figure of merit with a capacitive detector input circuit is $\phi_c = \phi(1 - \Delta\omega R_1 C_1)$. Therefore, under normal circuit operation, we cannot develop as high a voltage at the detector-circuit input in the presence of input capacitance as in its absence.

Conclusion

A number of experimental methods have been described for measuring the basic crystal-diode parameters. In addition, methods were described for measuring the detector-circuit input parameters which may be of interest in television receiver engineering.

Also, it was found that for a constant input signal level, the shape of the IF response characteristic will not depend on the back resistance of the crystal if the latter is above a certain minimum value. If the signal level changes, the effective ac resistance of the crystal will change because of the curvature of the diode characteristic. As a result of this change in resistance the detector-circuit input impedance will vary, and may detune the last IF stage and change its bandwidth.

A crystal figure of merit was developed for the specific application of video detection. Universal curves were presented which define germanium diode performance under the assumptions of constant forward and back conductance, zero-impedance driving voltage generator and high susceptance load.

Acknowledgment

The authors are grateful to E. H. Ulm of the Electronics Division for encouragement and support of this project. They are also glad to acknowledge the valuable assistance of A. Grometstein of the Electronics Division and E. J. Cuddy, E. F. Prufer, Shirley W. Harrison, and M. Fischman of the Physics Laboratories.

Correction

Aaron D. Bresler, author of the paper, "On the Approximation Problem in Network Synthesis," which appeared on pages 1724–1728 of the December, 1952 issue of the Proceedings of the I.R.E., has brought the following correction to the attention of the editors:

Equation (6), which appears in column one on page 1725 and which reads

$$\beta_1^2 = \beta_2\beta_3 = (\alpha^2 - \tfrac{1}{4})\frac{\beta_3}{\beta_2} = \left[\frac{\alpha + \frac{1}{2}}{\alpha - \frac{1}{2}}\right]^{2\alpha}, \qquad (6)$$

should be replaced by (6a) and (6b) as follows:

$$\beta_1^2 = \beta_2\beta_3 = (\alpha^2 - \tfrac{1}{4}) \qquad (6a)$$

$$\frac{\beta_3}{\beta_2} = \left[\frac{\alpha + \frac{1}{2}}{\alpha - \frac{1}{2}}\right]^{2\alpha}. \qquad (6b)$$

An Experimental Investigation of the Corner-Reflector Antenna*

EDWARD F. HARRIS†, ASSOCIATE, IRE

Summary—**This paper describes an experimental investigation, the purpose of which is to provide a fund of radiation-pattern design data for the corner-reflector antenna. The configuration is varied both in the angle between the plane sheets and the spacing from apex to dipole.**

Measurements are shown for corner angles from 30 to 270 degrees and for corner to dipole spacings of 0.1 to 3 wavelengths. Both *H*-plane and *E*-plane radiation patterns are shown for each configuration employing semi-infinite sheets. Certain specific corners are investigated for the effects on pattern of finite side lengths, side heights, and of the substitution of spines and grid construction for the solid sheets. Good correlation is found between calculated and measured patterns for the cases that may be handled by the method of images. Sets of integrable patterns are shown further to correlate measurements with gain calculations.

Introduction

THE USE OF SURFACES to reflect and focus electromagnetic energy seems to have had general application early in the radio art.[1,2] Work of a theoretical nature on radiation pattern and impedance characteristics of the corner-reflector configuration has appeared in the literature. Of particular interest is the work done by Kraus.[3] He has applied the method of images and has shown calculated data for several corner angles. Moullin has run additional experimental data on wide angles and large spacings.[4]

The majority of design data given in the literature concerns only beam width and power gain, with little information given on actual radiation patterns. There is little concerning applications that require patterns having special characteristic shapes. The principal use of the corner reflector has been to produce a maximum of radiation in the direction of the bisector of the corner angle. This single-lobe condition is termed the first order mode, but as the spacing S is increased the directional pattern begins to show multiple lobes, the condition which determines the higher order modes of operation. The higher radiation modes may exhibit large forward gain with narrow main lobe and small side lobes, as well as zero forward gain and two or more secondary lobes in the structure. However, these higher modes having special characteristic shapes may find good applications in fields such as direction finding and multiple coverage needs.

It is the purpose of this paper to present a series of experimentally determined patterns to show in a systematic manner the variations in radiation patterns of the corner-reflector antenna. Small increments in parameters are taken to bracket the entire range of practical operation of the device. The patterns shown represent relative electric field intensity at a large distance from the antenna for any direction in the plane of measurement, and should be useful in the preliminary design of a corner-reflector antenna having specific pattern characteristics.

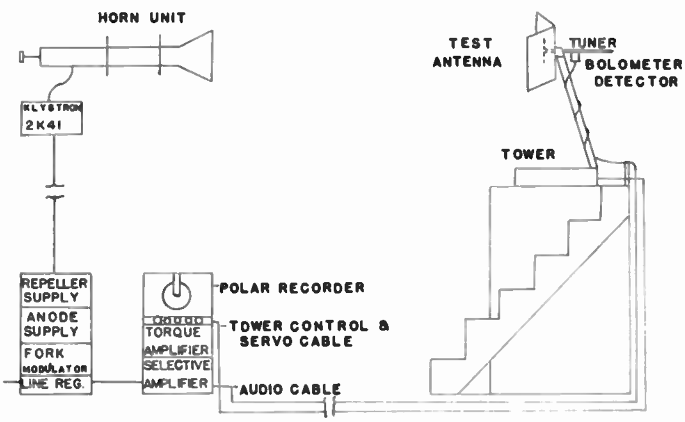

Fig. 1—Experimental setup for measuring antenna patterns.

The measuring techniques were straightforward; the experimental setup is shown in Fig. 1. It was desired to make these measurements in a frequency range for which the antenna dimensions would be of convenient size. A wavelength of 10 centimeters corresponds to approximately 4 inches. The mounting fixture and aluminum sheets had side dimensions of 16 by 24 inches, thus making the sides 4 by 6 wavelengths—values which approximate semi-infinite sheets for the measurements involved. The mounting fixture was constructed of wood and arranged to index the aluminum sheets at 10-degree intervals.

In an investigation of this nature it is imperative that automatic means for recording the polar radiation diagram be employed since the quantity of data taken would make manual plotting impractical. In addition it is highly desirable to obtain the increased pattern resolution which only an automatic polar recorder can provide.

* Decimal classification: R325.71. Original manuscript received by the Institute, May 23, 1952. Presented at the 1952 IRE National Convention, New York, N. Y.

† President, Mark Products Company, Chicago, Ill.

[1] S. G. Brown, "System of Wireless Telegraphy," U. S. Patent No. 741,622; October 20, 1903.

[2] F. Schroter, "Short Wave Aerial," U. S. Patent No. 1,830,176; November 3, 1931.

[3] J. D. Kraus, "The corner reflector antenna," Proc. I.R.E., vol. 28, p. 513; November, 1940.

[4] E. B. Moullin, "Radio Aerials," Oxford at the Clarendon Press, London, England; 1949.

Results and Conclusions

The experimental pattern data appear in Figs. 8 to 24 toward the end of this paper. Fig. 2 shows the antenna configuration and co-ordinate system. As in the horn type of antenna, the most useful radiation patterns of the corner-reflector antenna are those in the two principal planes, usually designated by the terms E-plane and H-plane because they are parallel to the electric and magnetic field intensities, respectively. The patterns shown are the original patterns taken on the polar recorder and assembled and photographed to provide the plates.

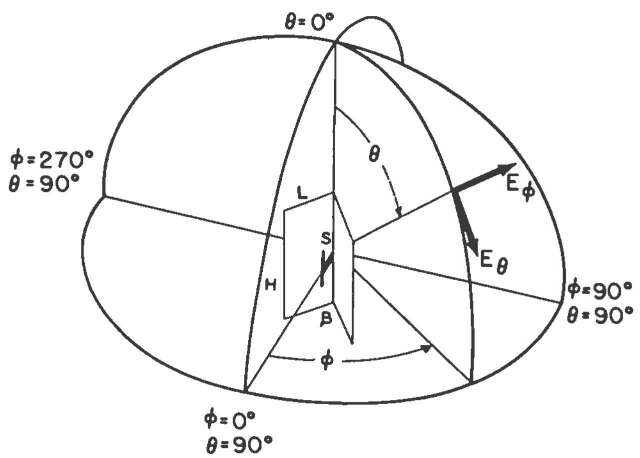

Fig. 2—Antenna and co-ordinate system.

Several interesting trends may be observed by close examination of the H-plane patterns. For any particular corner angle, for example, $\beta=60°$, the magnitude of any particular set of minor lobes relative to a major lobe may be traced from its origin. At a spacing $S=1.4$ wavelengths the single-lobe condition is still in existence, but at 1.5 the first set of minor lobes begins to appear. Continuing, somewhere between 1.5 and 1.6 wavelengths spacing the amplitudes of the two minor lobes and the major lobes become equal, and through a series of spacings during which the minor lobes grow larger in relation to the major lobe, at $S=1.8$ wavelengths the center lobe has entirely disappeared. It is this growth of secondary lobe structure that represents the higher order modes of operation and is responsible for the divided characteristic of the beam at larger spacings. Fig. 3 is an empirical set of curves which depicts the region over which the spacing may be carried for any particular corner angle in order to maintain first order mode of operation. The data for the preparation of these curves were obtained from direct observation of the measured patterns.

Figs. 4 and 5 are curves of the radiation resistance of a driven half-wavelength dipole versus spacing for the 180, 90, 60, and 45 degree corners. Figs. 6 and 7 are curves of the forward gain versus spacing for the same cases. These curves are the result of calculations. The information contained in these curves is intended for general evaluation of the corner-reflector antenna in that it complements the pattern data and provides gain and impedance values as a guide in the selection of patterns.

In an effort to extend the method of the corner reflector beyond the case of the flat sheet, reflector angles of 210, 240, and 270 degrees were investigated. An immediate application of this material may be seen from inspection of the patterns of the 270° reflector with spacings around 0.3 wavelength. This configuration provides a perfectly uniform distribution of radiated field

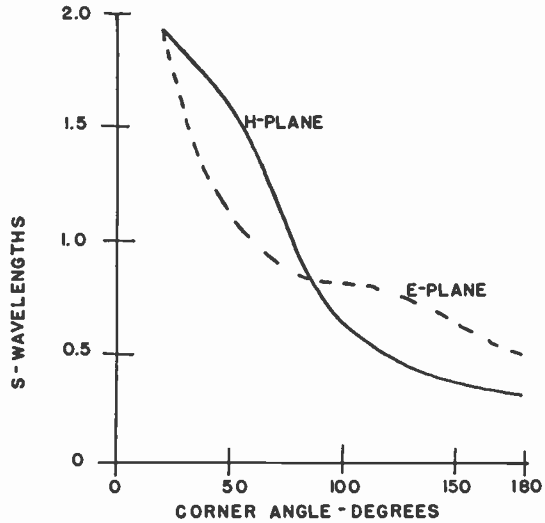

Fig. 3—Region of first order mode of operation.

over 180 degrees of azimuth. Beyond this the field drops rapidly to zero. At greater spacings the pattern becomes elongated, and finally multilobes appear. For services requiring coverage diagrams of 180 degrees this is a useful device and might conceivably be mounted at the corner of a building in order to take full advantage of structural conditions.

It is of interest to investigate methods of tilting the main beam off the center line of the corner angle. Fig. 21 shows the effects on pattern of setting the dipole 15 and 25 degrees off the center line as the apex to dipole spacing is varied. In the 15° case the beam is single lobed for small spacings and the tilt condition begins to appear at 0.5. This tilt is in a direction opposite to the offset of the dipole, and the maximum angle of about 15° occurs at $S=0.6$ wavelength. Beyond this the beam becomes multilobed. At a spacing of 1.0 wavelength a single tilted lobe reappears at an angle of about 20 degrees in the direction of the dipole offset. At greater spacings the break up becomes severe, and the operation is no longer practical. In the 25° offset operation is similar, except that the second single lobe condition occurs around 1.5 wavelengths, with a deflection somewhat greater than before. Designs utilizing two dipoles in lobe switching applications are possibly using this property of the 90° corner reflector. However, the effects of the passive dipole were not investigated.

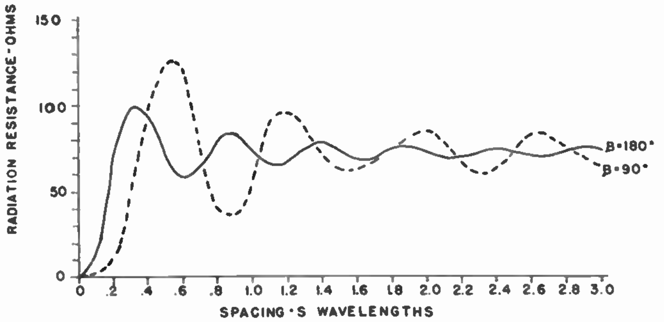

Fig. 4—Radiation resistance of driven λ/2 dipole versus spacing for $\beta=90°$ and $\beta=180°$.

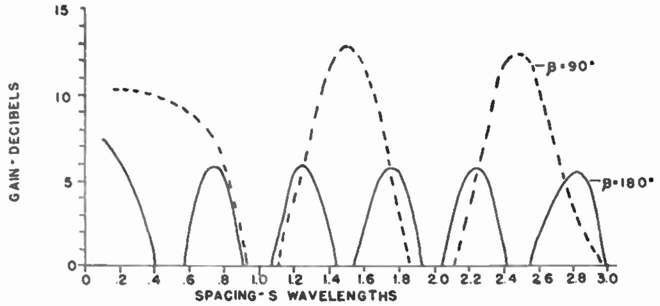

Fig. 6—Gain versus spacing for $\beta=90°$ and $\beta=180°$, direction $\phi=0°$, $\theta=90°$.

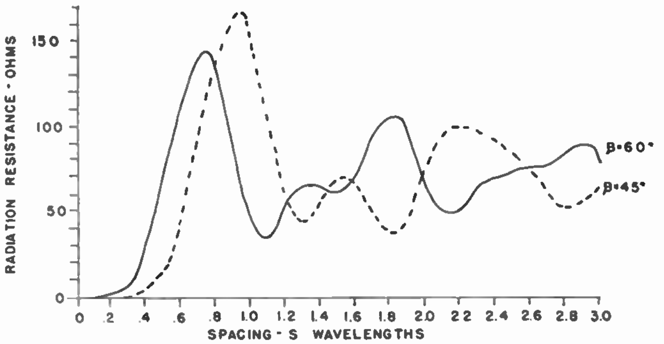

Fig. 5—Radiation resistance of driven λ/2 dipole versus spacing for $\beta=45°$ and $\beta=60°$.

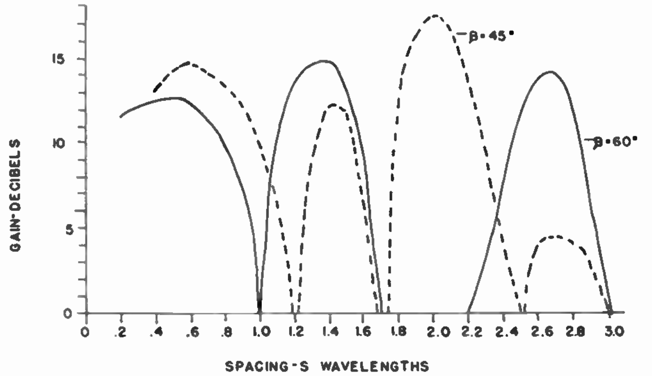

Fig. 7—Gain versus spacing for $\beta=45°$ and $\beta=60°$, direction $\phi=0°$ $\theta=90°$.

(Below)—Experimental Pattern Data Reproduced Directly from the Originals as Taken on the Automatic Polar Antenna Pattern Recorder

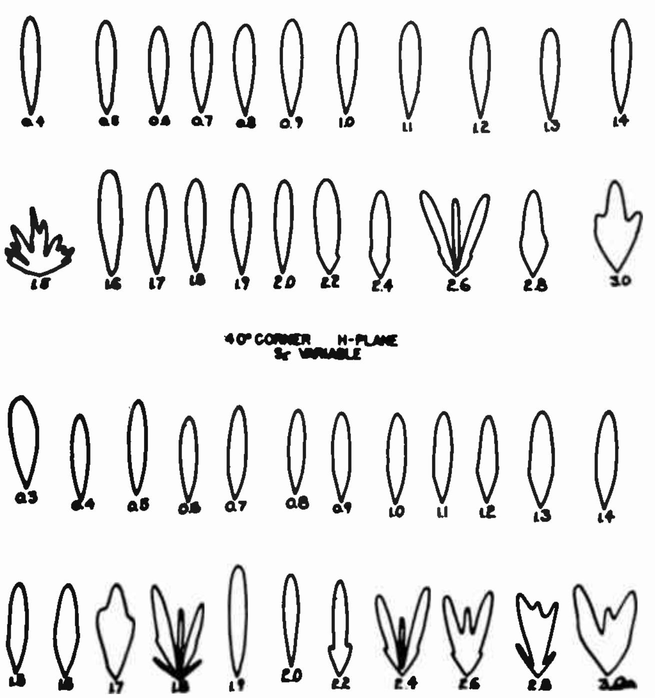

Fig. 8—30° and 40° corner, *H*-plane, S_λ-variable.

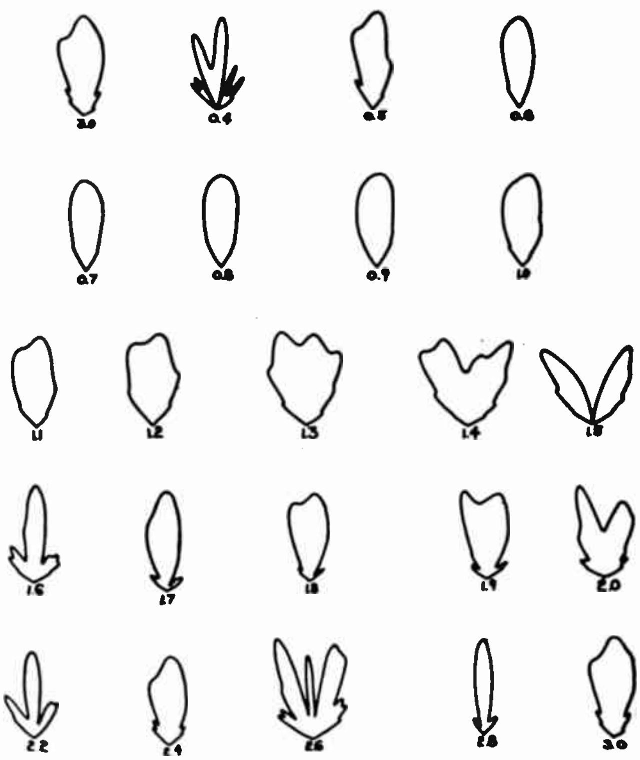

Fig. 9—30° corner, *E*-plane, S_λ-variable.

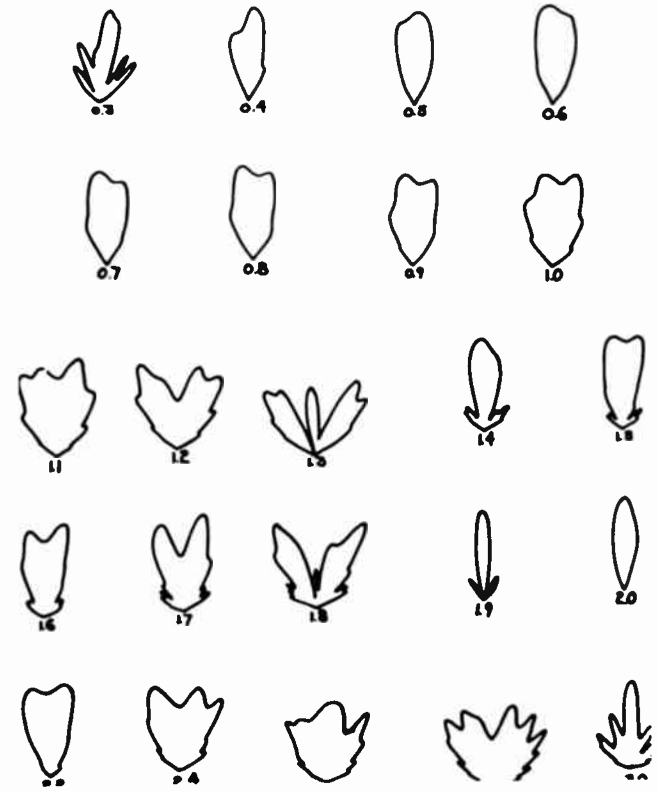

Fig. 10—40° corner, E-plane, S_λ-variable.

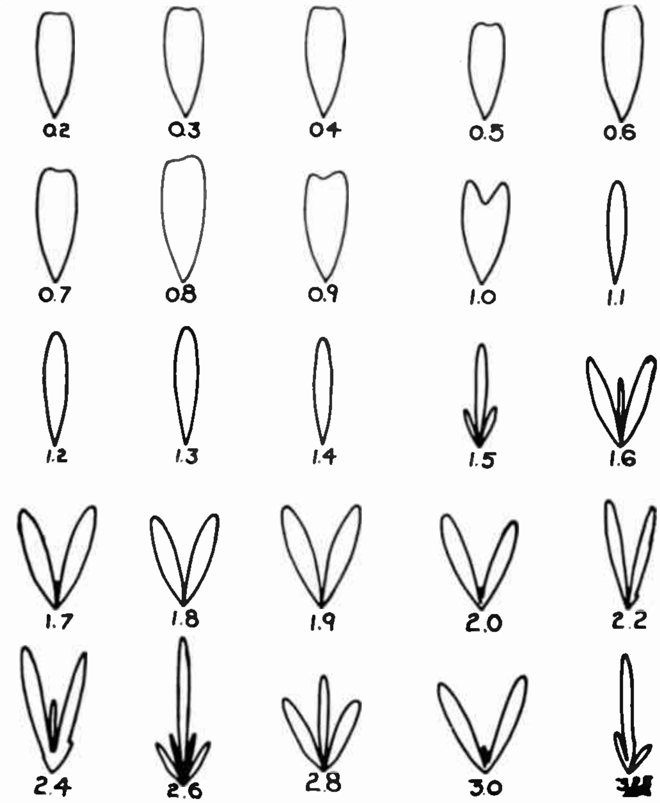

Fig. 11—60° corner, H-plane, S_λ-variable.

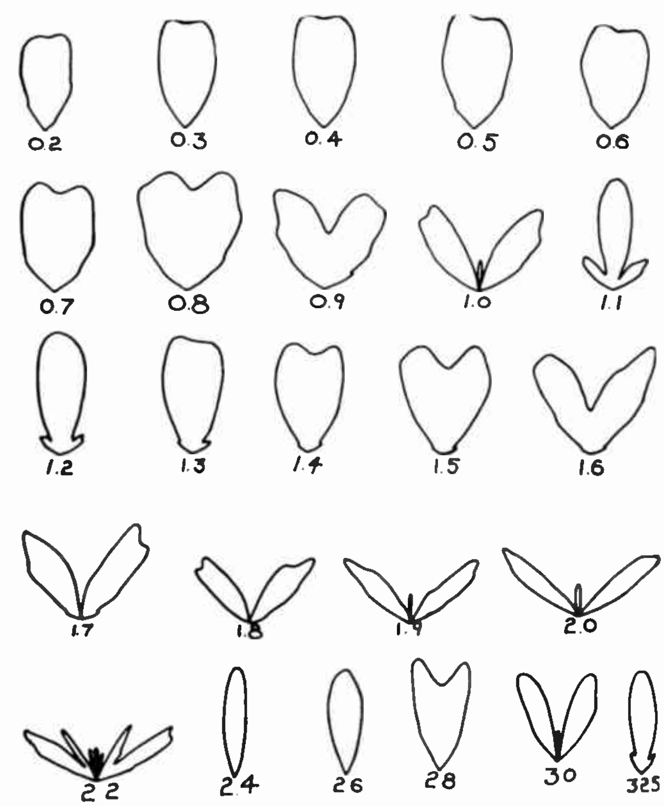

Fig. 12—60° corner, E-plane, S_λ-variable.

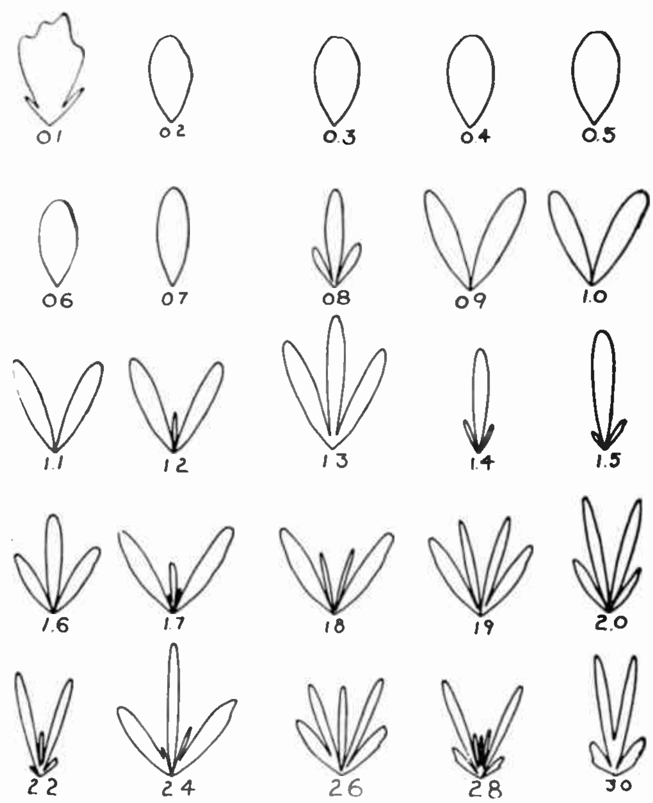

Fig. 13—90° corner, H-plane, S_λ-variable.

Of practical interest is the operation of the array using spine construction in place of solid sheet reflectors. Fig. 22 shows several cases of spine construction. The first set using 0.1 wavelength spine spacing and a 6-wavelength reflector substantially reproduces those patterns as taken with semi-infinite sheets. When the length of the reflector is reduced to 2 wavelengths there is a tendency to fatten some of the lobes and to lose the definite minimum ($S-1.0$). However, the agreement is still quite good. When the spine spacing is increased to 0.2

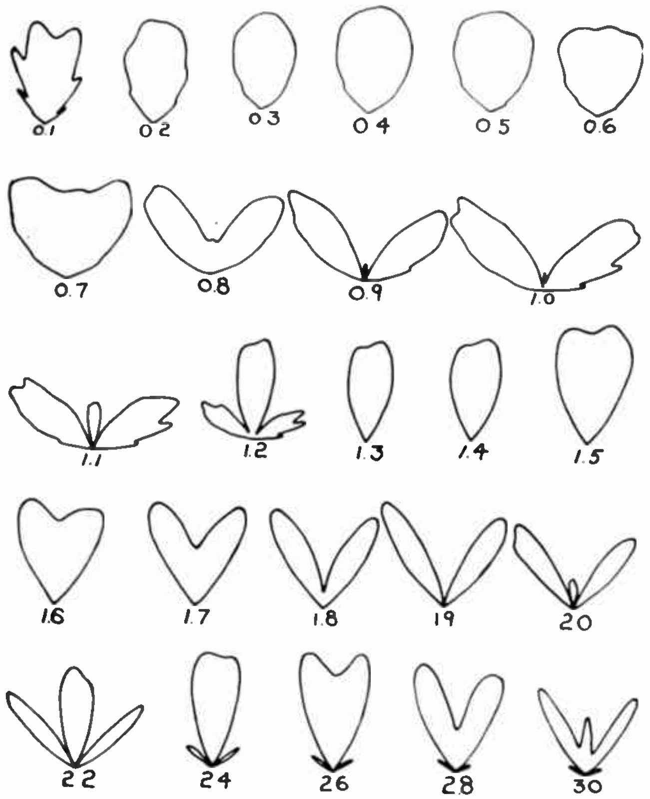

Fig. 14—90° corner, *E*-plane, S_λ-variable.

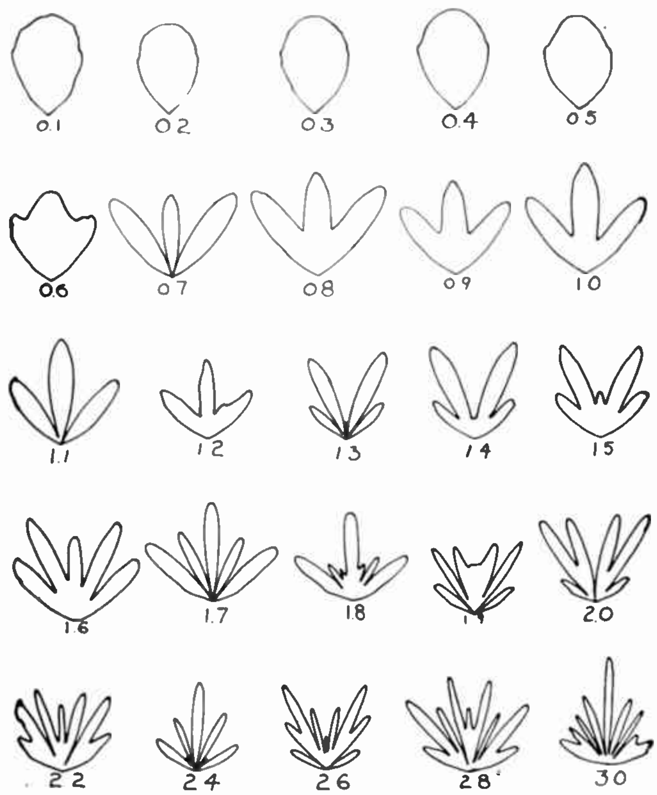

Fig. 15—120° corner, *H*-plane, S_λ-variable.

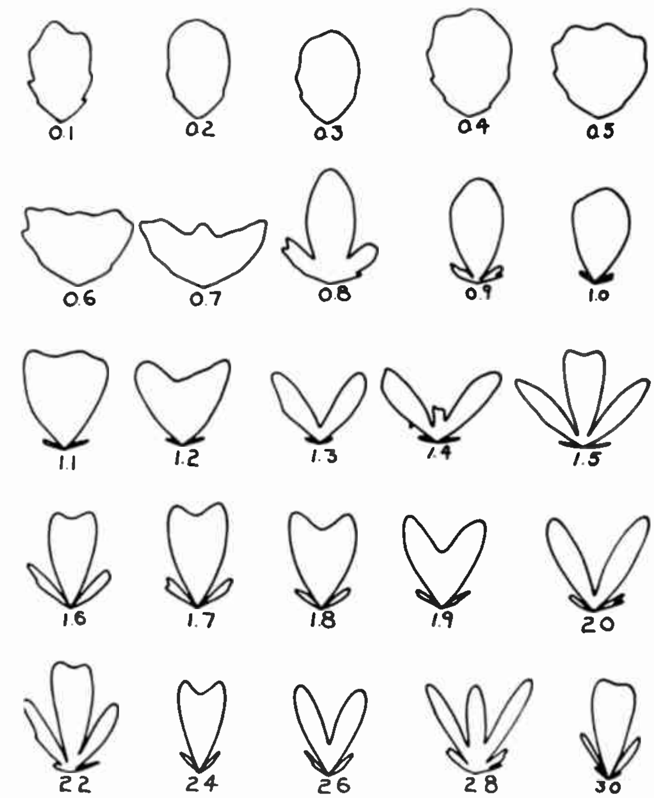

Fig. 16—120° corner, *E*-plane, S_λ-variable.

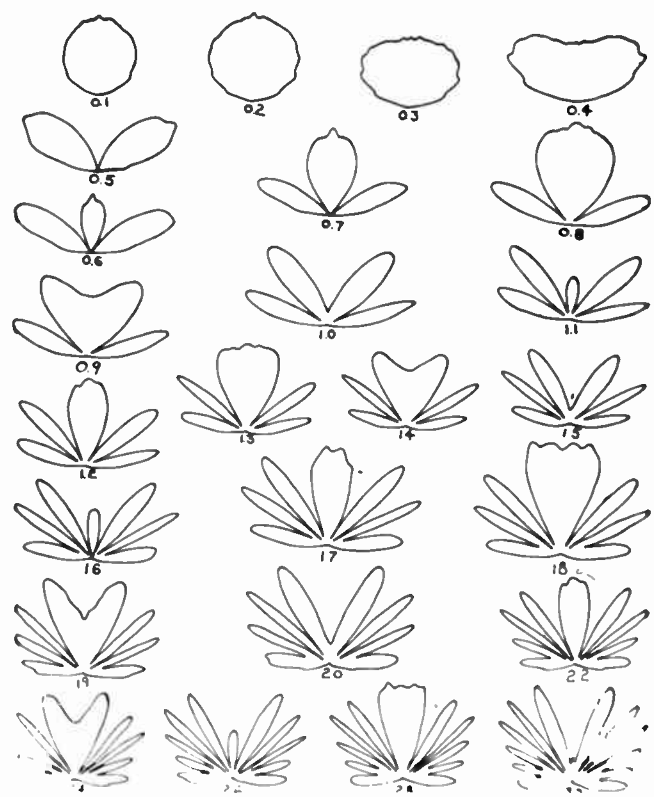

Fig. 17—180° corner (flat sheet), *H*-plane, S_λ-variable.

wavelength the patterns hold up quite well, but when it is further increased to 0.3 wavelength there is quite an increase in radiation through the reflector sheets, although the main lobes hold up well. The last set of patterns shows the condition of the 0.1 spine spacing and reflector length of 1 wavelength. The 1-wavelength reflector seems capable of maintaining the pattern at spacings as large as 0.6 wavelength, and although the general shape of the pattern is recognizable at greater spacings, the diffraction around the reflector is so great as to make for very poor operation.

The sets of space patterns shown on Figs. 23 and 24

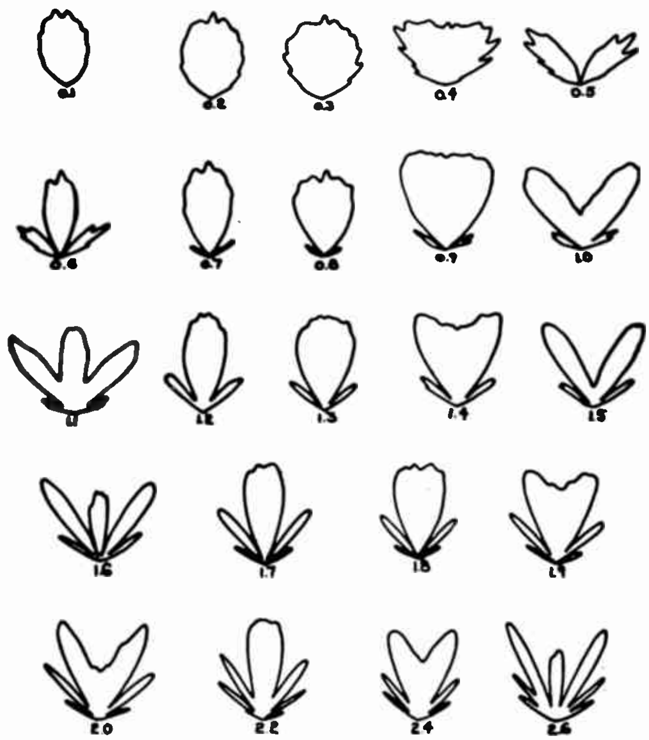

Fig. 18—180° corner (flat sheet), E-plane, S_λ-variable.

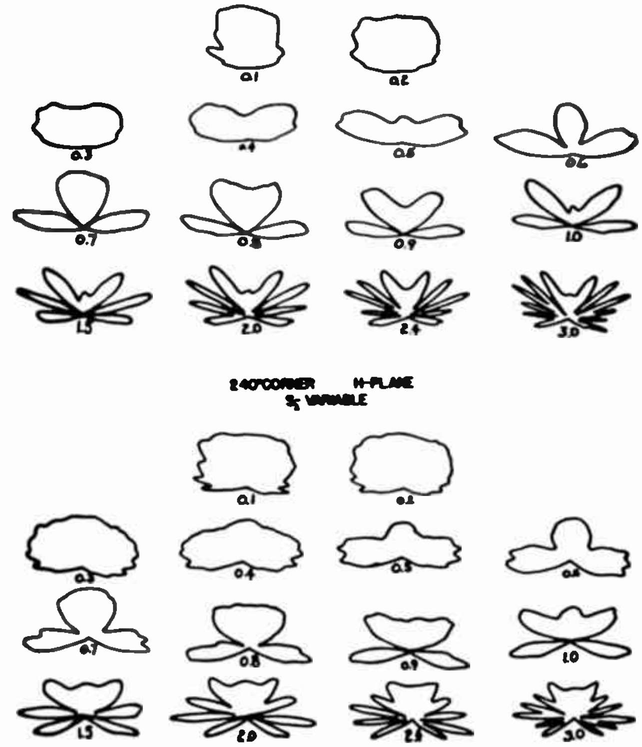

Fig. 19—210° and 240° corner, H-plane, S_λ-variable.

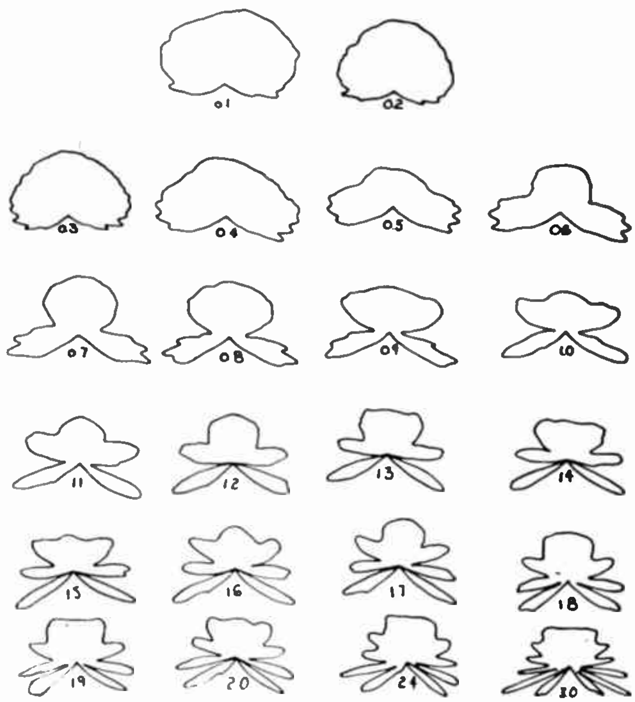

Fig. 20—270° corner, H-plane, S_λ-variable.

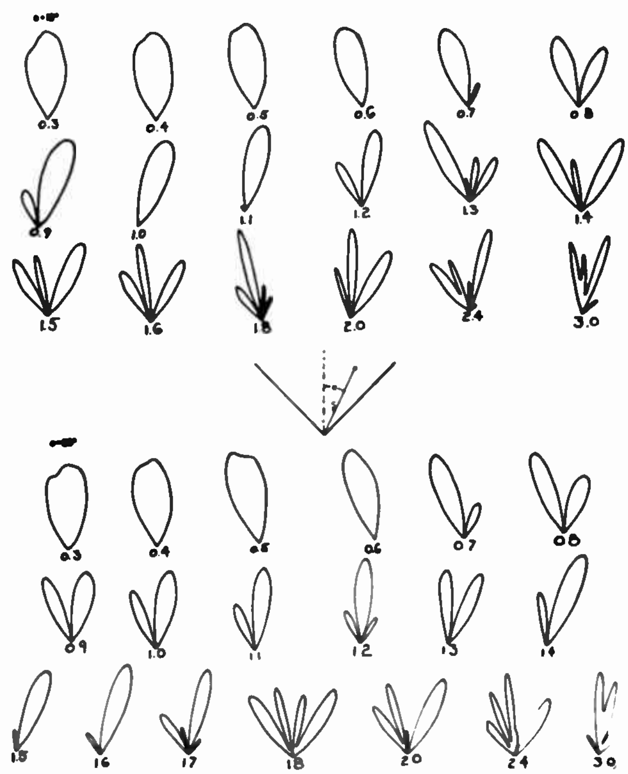

Fig. 21—90° corner, H-plane, $\lambda/2$ dipole located off bisector, S_λ-variable.

afford a check on the accuracy of measurements. The calculated gain values for the 90° corner fed with a half-wavelength dipole are 9.8 db for $S=0.5$ wavelength spacing and 12.8 db for the 1.5 wavelength spacing. From planimeter readings of pattern areas and the proper summation the gains as calculated from measured space patterns are 9.88 db and 13.1 db. respectively. The correlation is good considering that the equipment must remain perfectly stable over the period of a run of the set of space patterns. The measured gain of a half-wave dipole at 1.0 wavelength from the apex in a 90° corner is 12.36 db in the direction of the main lobes. A full-wavelength dipole was substituted for the half-wave driven unit in the 90° corner and gains of 10.46 db and 14.8 db were measured using 0.5 and 1.5

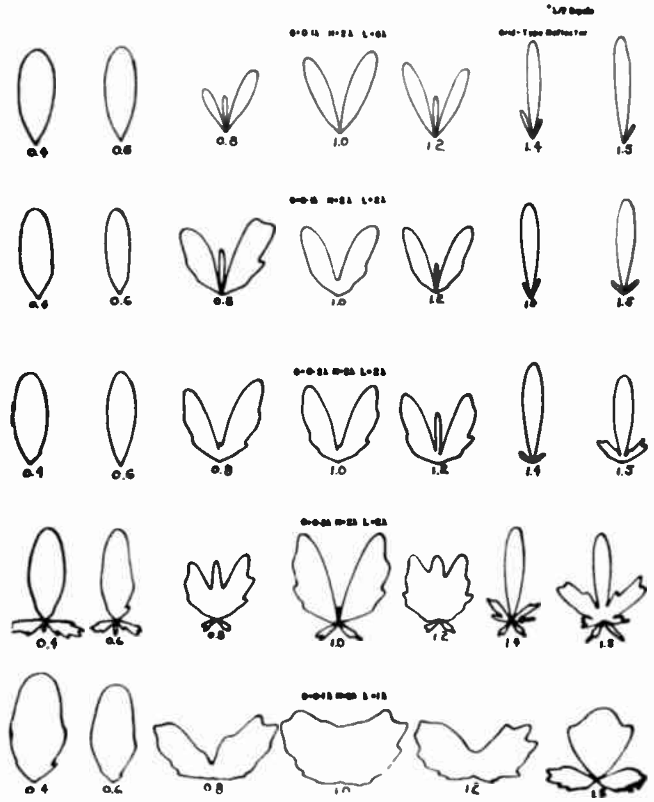

Fig. 22—90° corner, H-plane, spine construction, S_λ-variable.

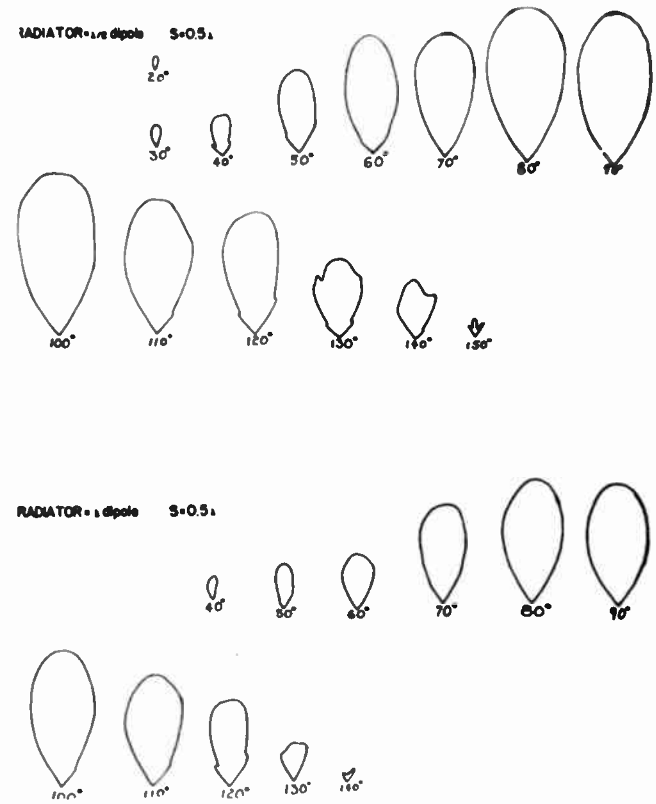

Fig. 23—90° corner, E_θ-component; space patterns as functions of ϕ, taken out at 10° intervals.

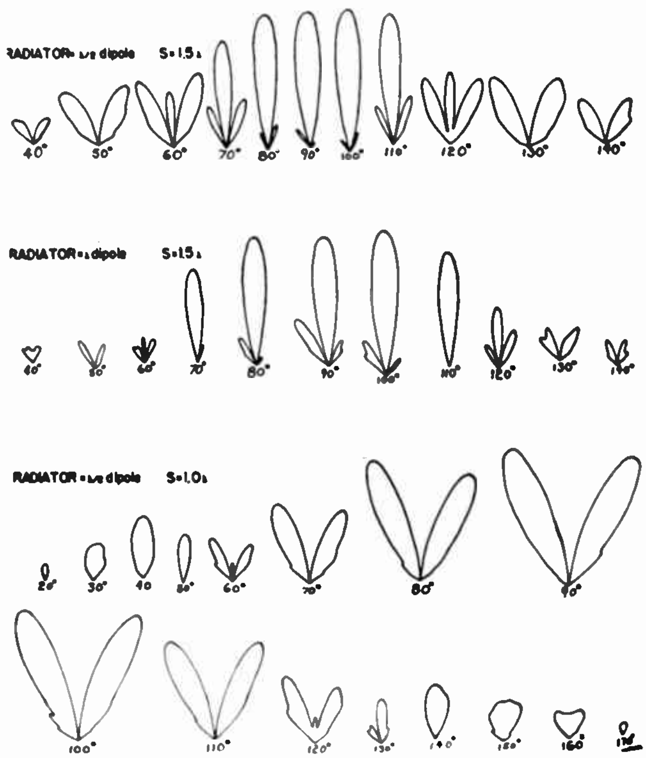

Fig. 24—90° corner, E_θ-component; space patterns as a function of ϕ, θ taken at 10° intervals.

wavelength spacings, respectively. All gain figures are expressed relative to a half-wave dipole.

Acknowledgment

The author wishes to express appreciation to the members of the Antenna Research Laboratory, Columbus, Ohio, for their co-operation and assistance, and to Dr. R. E. Beam of Northwestern University, Evanston, Illinois, for the invaluable aid of his many helpful suggestions.

Optimum Patterns for Endfire Arrays*

R. H. DUHAMEL†, ASSOCIATE, IRE

Summary—**The optimum design methods of Dolph and Riblet for the broadside array with an odd number of elements have been modified so that a common design procedure may be used. This procedure was then extended to the endfire array with an odd number of elements. A comparison of the optimum and other designs for endfire arrays is then given.**

Introduction

IN THE DESIGN of a linear array to produce a beam pattern, it is quite important that the beam be as narrow as possible and the side lobe level be quite low. During recent years many papers have appeared on the problem of designing a linear array to produce a specified pattern. Specifically, the synthesis of the optimum pattern for a broadside array has received considerable attention. Dolph[1] devised a method of synthesizing an optimum pattern for a broadside array of isotropic elements spaced a half-wavelength or greater. Riblet[2] later extended Dolph's method to include an array with an odd number of isotropic elements for which the spacing is less than a half wavelength. An optimum pattern is defined as a pattern for which the beam width is a minimum for a given side lobe level or, on the other hand, the side lobe level is a minimum for a given beam width. The optimum patterns are obtained from the Tchebycheff polynomials and hence will be referred to as either an optimum or a Tchebycheff pattern. An important and necessary property of the Tchebycheff pattern is that the side lobes are all equal. In a recent paper, Sinclair and Cairns[3] have considered broad side arrays of non-isotropic sources. They established the mathematical conditions which the optimum polynomial must satisfy and gave approximate methods for deriving this polynomial. They also outlined a method of synthesizing the optimum pattern for a broadside array with an even number of isotropic sources when the element spacing is less than a half wavelength. Thus, the broadside array has received a rather complete treatment.

Although various designs for endfire arrays have been published, the optimum design has not been given. The currents in a uniform endfire array are phased so that the waves from each of the sources arrive in phase in the direction of the main beam at large distances. Hansen and Woodyard[4] demonstrated that the gain of an endfire array could be increased by phasing the currents so that the waves do not arrive strictly in phase in the direction of the main beam. Later, Schelkunoff[5] obtained an improved design by utilizing the correspondence between the pattern of an array and the value of a complex polynomial on the unit circle. The improved patterns were obtained by equi-spacing the nulls of the polynomial on an appropriate arc of the unit circle. In the following, the optimum pattern for an endfire array with an odd number of isotropic sources spaced less than a half wavelength will be derived in a manner similar to that used by Riblet. This design yields a slightly better pattern than Schelkunoff's with the added advantage that direct control of the side lobe level is obtained. In order to obtain the optimum pattern for an endfire array with non-isotropic elements or with an even number of isotropic elements it would be necessary to resort to the techniques set forth by Sinclair and Cairns. Although there is a great difference in the mathematical techniques between the optimum designs for an odd and an even number of elements the differences between the resulting patterns and current distributions will be minor.

The optimum design method for the broadside array will be reviewed, then extended to the endfire array, and finally, the optimum design will be compared with other designs for endfire arrays.

Broadside Array

The optimum design methods of Dolph and Riblet for the broadside array will be modified so that a common design method may be used for either a broadside or an endfire array with an odd number of elements.

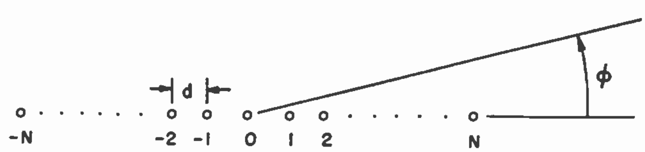

Fig. 1—Schematic broadside array of $2N+1$ elements having a spacing of d.

Consider the array shown in Fig. 1 with $2N+1$ elements with a spacing of d. Let the current in the n'th element be given by

$$I_n = A_n e^{-jn\alpha} \tag{1}$$

* Decimal classification: R125.1×R325.113. Original manuscript received by the Institute, June 25, 1952; revised manuscript received October 17, 1952.

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[1] C. L. Dolph, "A current distribution for broadside arrays which optimizes the relationship between beam width and side lobe level," Proc. I.R.E., vol. 34, no. 6, pp. 335–348; June, 1946.

[2] H. J. Riblet, "Discussion of Dolph's paper," Proc. I.R.E., vol. 35, pp. 489–492; May, 1947.

[3] G. Sinclair and F. V. Cairns, "Optimum patterns for arrays of non-isotropic sources," Transactions IRE, PGAP-1, pp. 50–61; February, 1952.

[4] W. W. Hansen and J. R. Woodyard, "A new principle in directional antenna design," Proc. I.R.E., vol. 26, pp. 33–346; March, 1938.

[5] S. A. Schelkunoff, "A mathematical theory of linear arrays," *B.S.T.J.*, vol. 22, pp. 88–107; January, 1943.

where $A_n = A_{-n}$ and the A_n's are all real. The progressive phase delay from left to right is given by α. For a broadside array α is set equal to zero. This results in an in-phase symmetrical current distribution. However, the quantity α will be retained in the following since the equations will then hold for both the broadside and endfire arrays.

The space factor of the array is given by

$$S_{2n+1} = \sum_{n=0}^{N} \epsilon_n A_n \cos n\psi \tag{2}$$

where ϵ_n is equal to one for $n=0$ and is equal to two for $n \neq 0$ and

$$\psi = \beta d \cos \phi - \alpha \tag{3}$$

This is simply a finite Fourier series with only cosine terms. It will be necessary at this point to review some of the properties of the Tchebycheff polynomials, from which the optimum patterns are derived. A polynomial of order N is defined by

$$T_N(z) = \cos (N \text{ arc} \cos z), \qquad |z| \leqq 1 \tag{4}$$

$$T_N(z) = \cosh (N \text{ arc} \cosh z), \qquad |z| > 1 \tag{5}$$

To write as a polynomial in z, it is only necessary to let $\theta = \text{arc} \cos z$, substitute in (4) and expand $\cos N\theta$ in terms of powers of $\cos \theta$. Although (4) is valid only for $|z| \leqq 1$, the resulting polynomial in z holds for all z. Using Dwight,[6] (4) becomes for $n > 1$

$$T_{2n}(z) = \sum_{q=0}^{n} A_{2q}{}^{2n} z^{2q} \tag{6}$$

$$T_{2n-1}(z) = \sum_{q=1}^{n} A_{2q-1}{}^{2n-1} z^{2q-1} \tag{7}$$

where

$$A_{2q}{}^{2n} = \frac{(-1)^{n-q} 2n(n+q-1)! 2^{2q-1}}{(2q)!(n-q)!}$$

$$A_{2q-1}{}^{2n-1} = \frac{(-1)^{n-q}(2n-1)(n+q-2)! 2^{2q-1}}{(2q-1)!(n-q)!}.$$

$T_4(z)$ is shown in Fig. 2. Notice that $T_4(z)$ crosses the z axis four times and that it oscillates between 1 and -1 for $|z| < 1$. Also for $z > 1$, $T_4(z)$ eventually increases at a rate proportional to z^4. In general $T_N(z)$ would increase at a rate proportional to z^N for $z > 1$ and would cross the z axis N times.

The optimum properties of the Tchebycheff polynomial are applied to the broadside array in the following manner. The $T_N(z)$ polynomial is used with $2N+1$ elements and the transformation

$$z = a \cos \psi + b \tag{8}$$

is used so that as ϕ varies from 0 to $\pi/2$ to π, z varies from -1 to z_0 to -1. For $d < \lambda/2$, it follows that

[6] H. B. Dwight, "Tables of Integrals and Other Mathematical Data," equation 403.3., MacMillan Co., New York, 1947.

$$a = \frac{z_0 + 1}{1 - \cos \beta d}, \qquad b = \frac{z_0 \cos \beta d + 1}{\cos \beta d - 1}. \tag{9}$$

if $d \geqq \lambda/2$, then the minimum value of $\cos \psi$ (which is equal to -1) is substituted in (9) in place of $\cos \beta d$. We then have for $d \geqq \lambda/2$

$$a = \frac{z_0 + 1}{2}, \qquad b = \frac{z_0 - 1}{2}. \tag{10}$$

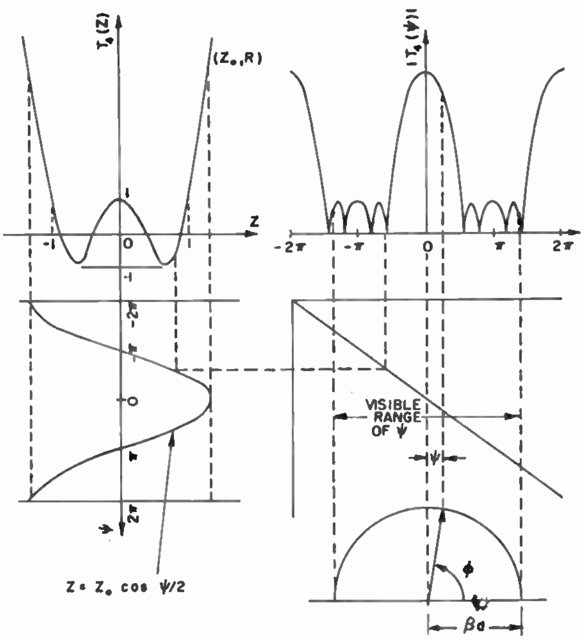

Fig. 2—Graphical construction of optimum pattern from T_4 polynomial for a five-element array with $d \geqq \lambda/2$ using $z = z_0 \cos\psi/2$.

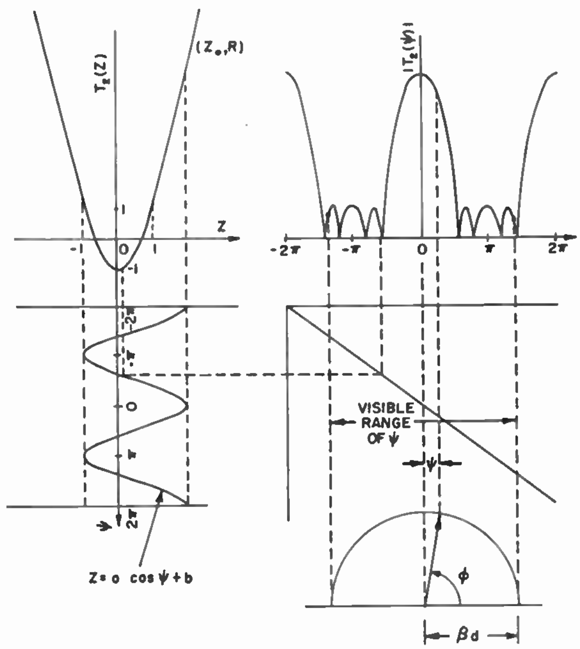

Fig. 3—Graphical construction of optimum pattern from T_2 polynomial for a five-element array with $d > \lambda/2$ using $z = a \cos \psi + b$.

Figs. 3 and 4 give a graphical representation of the transformation from the Tchebycheff polynomial to the Tchebycheff pattern for the two cases. Although the above transformation for $d>\lambda/2$ is different than that given by Dolph, it leads to the same result. Dolph applied the $T_{2N}(z)$ polynomial for $2N+1$ elements and used the transformation $z=z_0 \cos(\psi/2)$. Dolph's transformation is illustrated in Fig. 2. The equivalence of the two methods may be understood by comparing Figs. 2 and 3, both of which are for five element arrays. The advantage of using the above transformation, equation (8), is that it is not necessary to distinguish between the design methods for $d\geqq\lambda/2$ and $d<\lambda/2$. Also the same type of transformation will be used for the endfire array.

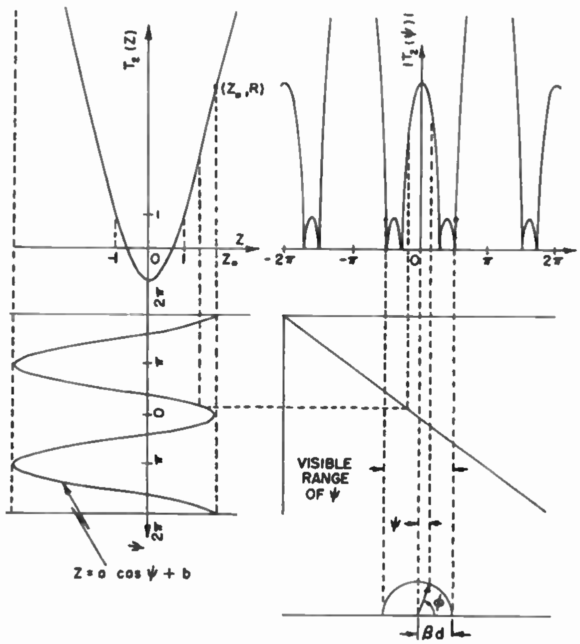

Fig. 4—Graphical construction of optimum pattern from T_2 polynomial for a five-element array with $d<\lambda/2$ using $z=a\cos\psi+b$.

Notice in Fig. 4 for $d<\lambda/2$ that there are very large lobes just outside the visible range of ψ. These large invisible lobes represent large quantities of stored energy in comparison with the radiated energy which in turn signifies that the antenna efficiency will be reduced. This type of array is termed a super-gain array. If spacing is much less than a quarter wavelength, the low efficiency may then limit practical applications of the array.

The value of z_0 determines the maximum value, R, of the main beam which in turn determines the side lobe level. If either the side lobe level or the position of the first null is specified then z_0 may be found as shown below. For a constant side lobe level the beam width is decreased and the directivity gain is increased by increasing the order of the Tchebycheff pattern, N.

From (4), it is seen that nulls of $T_N(z)$ occur when

$$\text{arc cos } z_k = \frac{(2k-1)\pi}{2N}, \qquad k = 1, 2, \cdots, N$$

Then using $z=a\cos\psi+b$, the nulls of the Tchebycheff pattern are found to occur when

$$\psi_k = \pm \text{ arc cos}\left[\frac{\cos \pi(2k-1)/2N - b}{a}\right] \qquad k = 1, 2, \cdots, N. \quad (11)$$

In a similar manner, $T_N'(\psi)=0$ whenever

$$\psi_k = \pm \text{ arc cos}\left[\frac{\cos \pi K/N - b}{a}\right] \qquad k = 1, 2, \cdots, N-1. \quad (12)$$

For the broadside array $\psi=\beta d\cos\phi$.

If the side lobe to main beam ratio is chosen as $1/R$ then $T_N(z_0)=R$ and from (4)

$$z_0 = \cosh\left[(1/N) \text{ arc cosh } R\right]. \quad (13)$$

Thus a, b, and the positions of the side lobes and nulls may all be determined once the value of z_0 is known.

If, on the other hand, the first null ϕ_1 is specified, then from (3), (9), (10) and (11) it follows that

$$z_0 = \frac{1+\cos(\pi/2N)(1-\cos\beta d) - \cos(\beta d\cos\phi_1)}{\cos(\beta d\cos\phi_1) - \cos\beta d} \quad (14)$$

where $\cos\beta d$ is replaced by -1 for $d>\lambda/2$.

From the above considerations it is evident that once the values of N and z_0 are specified, the Tchebycheff pattern is uniquely determined. The problem of determining the current coefficients A_n may be accomplished easily after expanding the Tchebycheff polynomial into a finite Fourier series. This is done in Appendix I and the result is

$$T_N(a\cos\psi+b) = \sum_{n=0}^{N} C_n^N \cos n\psi \quad (15)$$

where the C_n^N's are functions of a and b. Then by equating (2) to (15) it is seen that[7]

$$A_n = C_n^N/\epsilon_n. \quad (16)$$

For $d\geqq\lambda/2$, it will be noticed that a and b are independent of βd. This means that the required element currents are independent of frequency as long as $d\geqq\lambda/2$. As βd approaches $2\pi(d\rightarrow\lambda)$ major lobes will appear in the pattern at $\phi=0$ and $\phi=\pi$. Thus, the element spacing is usually restricted to be less than a wavelength for broadside arrays.

For $d<\lambda/2$ the required element currents are functions of βd. Hence, the pattern for an array with fixed currents is optimum only at the frequency for which it is designed.

Riblet[2] proved in his paper that of all the polynomials (whose coefficients may be real or complex) of degree N, T_N alone possesses the following optimum properties: (a) if the side lobe level is specified, the distance to the first null is minimized; (b) if the first null is specified, the side lobe level is minimized. Let us now apply these polynomials to the endfire array.

Endfire Array

The optimum design procedure for an endfire array with an odd number of elements is carried through in the same manner as that for the broadside array with a few

[7] This method of evaluating the A_n's is equivalent to, but different than that used by Dolph.

minor changes. The most important difference is that the progressive phase delay is no longer zero. As before, the space factor is given by

$$S_{2N+1} = \sum_{n=0}^{N} \epsilon_n A_n \cos n\psi \tag{17}$$

where $\psi = \beta d \cos \phi - \alpha$. The A_n's must always be real for the optimum patterns because the coefficients of the Tchebycheff polynomials are real.

Now the element spacing for an endfire array must be somewhat less than a half wavelength to insure that a second major lobe does not appear. For $2N+1$ elements, the optimum patterns will again be derived from the Tchebycheff polynomial of order N. Thus it will be necessary to use a transformation of the type $z = a \cos \psi + b$. The required transformation is illustrated in Fig. 5. The unknown parameters may be determined from the following conditions. At $\psi = 0$, $z = -1$ and at $\phi = 0$, $z = z_0$. At $\phi = 180°$, we must have $z = 1$. These conditions yield the following three equations

$$-1 = a + b \tag{18}$$

$$z_0 = a \cos (\beta d - \alpha) + b \tag{19}$$

$$1 = a \cos (\beta d + \alpha) + b \tag{20}$$

from which a, b, and α may be determined. Doing this, we have

$$a = \frac{-(z_0 + 3) - 2 \cos \beta d \sqrt{2(z_0 + 1)}}{2 \sin^2 \beta d} \tag{21}$$

$$b = -1 - a \tag{22}$$

$$\alpha = \sin^{-1} \frac{z_0 - 1}{2a \sin \beta d} \tag{23}$$

$$\alpha = \cos^{-1} \frac{z_0 + 3 + 2a}{2a \cos \beta d}. \tag{24}$$

The angle α is usually located in the fourth quadrant. If the side lobe level is specified, then the parameter z_0 is determined according to equation (13). The positions of the nulls and the secondary maxima may be calculated from (11) and (12). The inverse problem of determining z_0 when the first null is specified is not so simple. It is a rather complicated procedure to solve for z_0 for a given value of the first null, ϕ_1. It is felt that it would be simpler to determine z_0 by a "cut and try process." That is, choose several side lobe levels and then find the positions of the corresponding first nulls by means of (11). Then an accurate guess could be made in choosing a side lobe level which would give the desired position of the first null.

The source currents are determined in the same manner as for the broadside array. As an example, consider a seven element array with $d = \lambda/4$ for which the T_3 polynomial will be used. Let the side lobe level be 20 db down. Then from (13)

$$z_0 = \cosh (1/3 \text{ arc cosh } 10) = 1.5408.$$

Then substituting in (21) to (24) we obtain

$$a = -2.2705, \quad b = 1.2705, \quad \alpha = -6.8°.$$

Using Appendix II, the pattern as a function of ψ is given by

$$T_3(\psi) = -43.69 + 72.28 \cos \psi - 39.30 \cos 2\psi + 11.70 \cos 3\psi.$$

By equating this to (2) and using (1), the source currents are

$$I_0 = -43.69$$
$$I_1 = I_{-1}^* = 36.14\underline{/6.8°}$$
$$I_2 = I_{-2}^* = -19.65\underline{/13.3°}$$
$$I_3 = I_{-3}^* = 5.85\underline{/20.4°}$$

where the * denotes the complex conjugate. The pattern for the array is shown in Fig. 6. Also included are patterns for other designs which will be discussed in the next section.

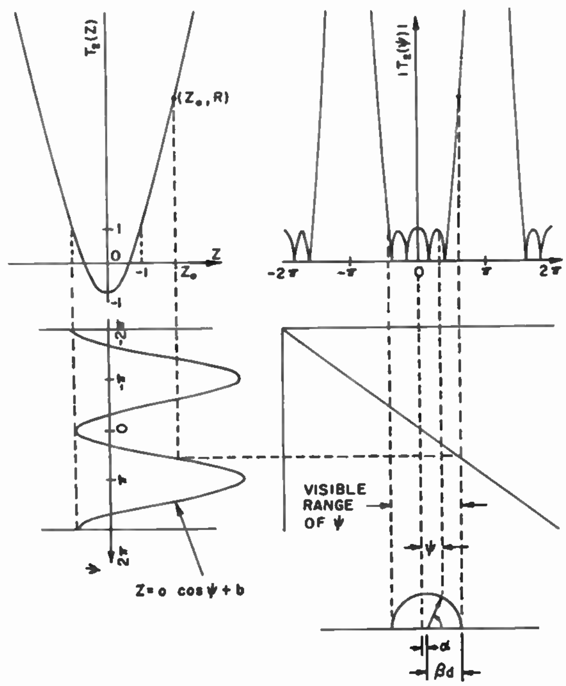

Fig. 5—Graphical construction of optimum pattern from T_3 polynomial for a five-element endfire array with $d = \lambda/4$.

A bi-directional endfire pattern may be easily obtained by a slight modification. From Fig. 5 it is apparent that a major lobe would occur at both $\phi = 0$ and $\phi = 180°$ if the progressive phase delay is set equal to zero. The required transformation for an optimum bi-directional endfire pattern may be determined from the following conditions. At $\psi = 0$, $z = -1$ and at $\phi = 0$, $z = z_0$. Then with the transformation $z = a \cos \psi + b$, where $\psi = \beta d \cos \phi$, it is found that we must have

$$a = \frac{z_0 + 1}{\cos \beta d - 1}, \quad b = \frac{z_0 + \cos \beta d}{1 - \cos \beta d}.$$

The remainder of the design procedure is identical to that for the broadside array.

Comparison of Endfire Designs

In the following we shall compare the patterns and source currents for several different designs for a seven element array with quarter wavelength spacing. Consider first the uniform endfire array for which the magnitudes of all the source currents are equal and the progressive phase delay α is equal to βd. The space factor is then

$$S_{2N+1} = \sum_{n=0}^{N} \epsilon_n \cos n\psi = \frac{\sin (2N+1)\psi/2}{\sin \psi/2}$$

or

$$S_7 = \frac{\sin (7\psi/2)}{\sin (\psi/2)} \tag{25}$$

where $\psi=\pi(\cos \phi-1)/2$ for quarter wavelength spacing. This pattern is shown in Fig. 6. The nulls of this expression occur when

$$\psi = -\frac{2k\pi}{7} \qquad k = 1, 2, \cdots .$$

It will be noticed that only three of the nulls appear in the visible range of ψ whereas the optimum pattern has six nulls.

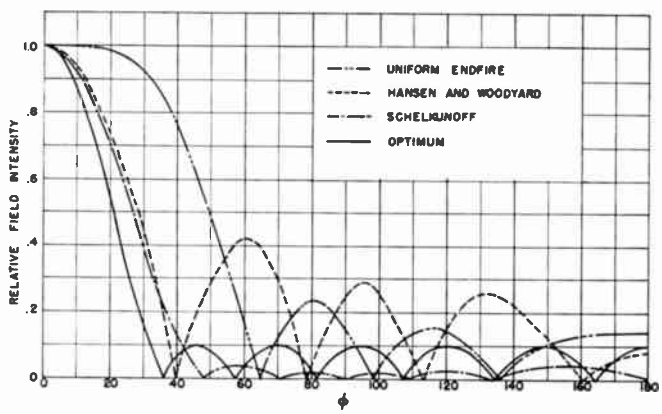

Fig. 6—Comparison of endfire designs for seven-element array with quarter wavelength spacing.

Hansen and Woodyard[4] devised an improved endfire design for an aperture by letting the magnitude of the current distribution across the aperture be constant and then choosing the progressive phase delay constant, α, such that the gain of the aperture was maximized. They found that if the total phase shift across the aperture was about π plus 2π times the width of the aperture in wavelengths, then the gain is maximized and is increased by a factor of about 1.8 over that for the usual endfire design. For an array, the corresponding phase difference between adjacent radiators would be about $\pi/(M-1)$ radians greater than βd where M is the total number of radiators. Thus, for the seven element array we have

$$\alpha = \beta d + \frac{\pi}{M - I} = 120°$$

The space factor is again given by (23) except we now have $\psi=\pi(\cos \phi-4/3)/2$. The resulting pattern is shown in Fig. 6. The first side lobe has a relative magnitude of 0.42 which may be undesirable for many applications. However the first null occurs at 39° whereas the first null was at 65° for the uniform endfire array. Notice that four nulls appear in the pattern now. Goward,[8] using a pattern synthesis method introduced by Woodward,[9] proposed a method for reducing the side lobe level of the Hansen and Woodyard design for the continuous aperture and thus increasing the gain. However, the application of this technique to an array is rather nebulous.

Let us now consider Schelkunoff's design for the endfire array. It will be convenient to move the origin in Fig. 1 to the last element on the left and renumber the elements from 0 to $2N$. The space factor is then

$$S_{2N+1} = \sum_{n=0}^{2N} A_n e^{jn\psi}. \tag{26}$$

Then by letting $\alpha=\beta d$, we have

$$S_{2N+1} = \sum_{n=0}^{2N} A_n z^n \tag{27}$$

where

$$z = e^{j\psi} = e^{j\beta d(\cos\phi - 1)}.$$

Thus, the space factor is given by the value of the complex polynomial on the unit circle defined by $|z|=1$. The polynomial may also be written as the product of its factors.

$$S_{2N+1} = A_{2N} \prod_{n=1}^{2N} (z - a_n) \tag{28}$$

The a_n's are the zeros of the polynomial. As ϕ varies from 0 to π, ψ varies from 0 to $2\beta d$. Viewed in the complex plane (see Fig. 6) z moves in the clockwise direction from $z=1$ to $z=-1$ for quarter wavelength spacing. Schelkunoff's design consists of equi-spacing the nulls of (28) in the active range z. Thus, for the seven element array, the space factor is

$$S_7 = A_6 \prod_{n=1}^{6} (z - e^{-j(n\pi/6)}). \tag{29}$$

The positions of the zeros are shown in Fig. 6. The resulting pattern is shown in Fig. 7. Notice that the side lobes are quite low and that the beam width is approximately the same as that for the Hansen and Woodyard design. The source currents are obtained by expanding (29) into the series form given by (27).

[8] F. K. Goward, "An improvement in end fire arrays," *Proc. I.E.E.*, part III, vol. 94, pp. 415–418; November, 1947.

[9] P. M. Woodward, "A method of calculating the field over a plane aperture required to produce a given polar diagram," *Proc. I.E.E.*, part IIIA, vol. 93, pp. 1554–1558; March–May, 1946.

For the uniform endfire array, the zeros are equispaced on the complete unit circle, which accounts for the presence of only three nulls in the visible portion of the space factor.

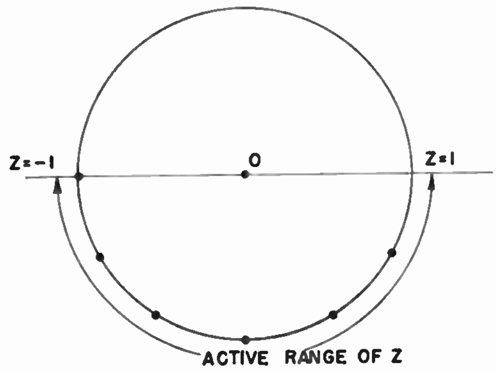

Fig. 7—Disposition of null points for an endfire array with nulls equispaced in active range of z for $d=\lambda/4$.

The optimum pattern is also shown in Fig. 7. Since the side lobes are of different magnitudes, it is difficult to compare the optimum pattern with the pattern obtained using Schelkunoff's method. This latter pattern has its first null at 48 degrees and a maximum side lobe 28 db down. If the optimum pattern had been chosen to have a side lobe level of 28 db then the first null would be at 44.5 degrees. On the other hand, if the optimum pattern had been specified to have the first null at 48 degrees, then the side lobe level would be 31 db down. Thus the optimum design offers a slight improvement over Schelkunoff's design. The improvement is due to two factors. First, in contrast to Schelkunoff's design, the side lobes of the Tchebycheff pattern are all equal. Secondly, an extra side lobe appears in the Tchebycheff pattern. That is, the sixth null occurs at $\phi=164°$ whereas the sixth null in Schelkunoff's design occurs at $\phi=180°$. Thus it is apparent that an improvement of Schelkunoff's technique may be obtained by rotating the zeros on the unit circle a slight amount in the counterclockwise direction. This would introduce an extra side lobe in the visible pattern. An important advantage of the optimum design is that direct control of the side level is easily obtained.

So far, we have compared only the patterns for the various endfire designs. Table I gives the relative source currents required for the different designs. In each case the currents are chosen so that the field strength in the direction of the main beam is the same for each design. The currents for Schelkunoff's and the optimum designs are quite large in comparison to the currents for the uniform endfire design. Thus, the increase in gain will be accompanied by a reduction in the antenna efficiency, bandwidth, and radiation resistance. For the case where the beam width for the optimum pattern is the same as that for the equi-spaced null pattern, there is only a slight difference in the source currents. Note that the "taper" of the current magnitudes for each of the optimum designs is roughly the same and is given approximately by 1.0: 0.8: 0.4: 0.1.

As the gain is increased, it is also necessary to establish the source currents with greater precision. This becomes a serious problem in the design and construction of the array. For example, with the optimum design No. 1 shown in Table I, the source currents would have to be accurate to within 0.1 per cent of the maximum current in the array in order to keep the side lobe level within one db of the desired value. On the other hand, it is difficult to obtain accuracies better than about 10 per cent in actual designs. Thus, at first thought, the optimum design procedure appears to be of academic interest only. However, it may be utilized as described below.

The major effect of errors in the current distribution is an increase in the side lobe level. Use was made of an electro-mechanical antenna pattern calculator at the RCA Laboratories at Princeton, N. J. in order to illustrate this effect. With this calculator it is possible to establish the currents with an accuracy of roughly five per cent of the maximum current in the array. When the calculator was set up for the optimum design No. 2, the pattern shown in Fig. 8, was obtained. Notice that the maximum side lobe is only 12 db down whereas the theoretical level is 31 db down. With the optimum design No. 1 set up on the calculator, the main lobe was barely distinguishable from the side lobes. Also included in Fig. 8 are the patterns for the optimum designs No. 3 and No. 4 as obtained from the calculator. The maximum side lobes are 20 and 26 db down for the No. 3 and No. 4 designs as compared to the theoretical values of 52 and 63 db respectively. Thus, the errors in the

TABLE I

RELATIVE SOURCE CURRENTS AND PATTERN CHARACTERISTICS FOR VARIOUS ENDFIRE DESIGNS FOR A LINEAR ARRAY OF SEVEN ELEMENTS WITH QUARTER WAVELENGTH SPACING. THE BEAM WIDTH IS THE ANGLE BETWEEN THE HALF-FIELD INTENSITY POINTS ON THE PATTERNS. THE OTHER CURRENTS ARE OBTAINED BY THE RELATION $I_{-n}=I_n^*$.

	Beam width	Side lobe level in db	Progressive phase delay in degrees	Current magnitudes			
				I_0	I_1	I_2	I_3
Uniform	99	13	90	1.0	1.0	1.0	1.0
Hansen and Woodyard	57	7.4	120	1.82	1.82	1.82	1.82
Schelkunoff	53	28	165	13.05	10.63	5.50	1.42
Optimum No 1	42	20	173.2	30.58	25.30	13.75	4.10
Optimum No 2	52	31	166.3	14.31	11.58	6.05	1.63
Optimum No 3	67	52	150.8	5.751	4.554	2.153	.483
Optimum No 4	78	63	141.6	4.272	3.370	1.512	.313

source currents result in an increase in the side lobe level. Hence, for applications where a reduction of antenna bandwidth and efficiency could be tolerated, an array could be "over designed" in order to avoid the stringent requirements on the precision of the source currents. That is, if a 20 db side lobe level was desired, then the array would be designed to have, maybe, a 50 db side lobe level. The resulting pattern although not optimum would still be a decided improvement over the uniform endfire pattern. To illustrate the difference, a pattern for the same physical array as above with a progressive phase delay of ninety degrees and the same taper as above for the optimum design is included in Fig. 8.

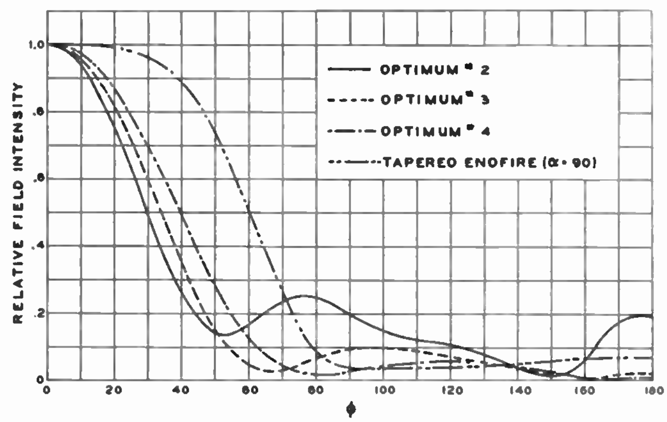

Fig. 8—Patterns plotted by an electromechanical pattern calculator for the various optimum designs listed in Table I.

It should be realized that as the side lobe level is decreased, the beam width increases and hence the gain and the supergain effects are diminished.

Acknowledgments

The major portion of this work was performed at the University of Illinois while the author was a member of the Radio Direction Finding Research Group which is sponsored by the Office of Naval Research. The author is indebted to Dr. E. C. Jordan of the University of Illinois and Dr. G. H. Brown of the RCA Laboratories Division for valuable comments on the problem.

Appendix I

If $z=ax+b(x=\cos\psi)$ is substituted in (6) and (7) and use is made of the binomial expansion

$$(ax+b)^m=\sum_{k=0}^{m}\binom{m}{k}(ax)^{m-k}b^k$$

then

$$T_{2n}(x)=\sum_{q=0}^{n}\sum_{k=0}^{2q}A_{2q}{}^{2n}\binom{2q}{k}(ax)^{2q-k}b^k$$

$$T_{2n-1}(x)=\sum_{q=1}^{n}\sum_{k=0}^{2q-1}A_{2q-1}{}^{2n-1}\binom{2q-1}{k}(ax)^{2q-k-1}b^k$$

where

$$\binom{n}{k}=\frac{n!}{k!(n-k)!}. \tag{30}$$

By regrouping the terms and substituting for the A's the following equations are obtained.

$$\begin{aligned}T_{2n}(x)&=\sum_{q=0}^{n}B_{2q}{}^{2n}x^{2q}+\sum_{q=1}^{n}B_{2q-1}{}^{2n}x^{2q-1}\\ T_{2n-1}(x)&=\sum_{q=1}^{n}B_{2q-1}{}^{2n-1}x^{2q-1}+\sum_{q=0}^{n-1}B_{2q}{}^{2n-1}x^{2q}\end{aligned} \tag{31}$$

where

$$\begin{aligned}B_{2q}{}^{2n}&=\sum_{p=q}^{n}\frac{(-1)^{n-p}2n(n+p-1)!2^{2p-1}b^{2(p-q)}a^{2q}}{(n-p)!(2p-2q)!(2q)!}\\ B_{2q-1}{}^{2n}&=\sum_{p=q}^{n}\frac{(-1)^{n-p}2n(n+p-1)!2^{2p-1}b^{2(p-q)+1}}{(n-p)!(2p-2q+1)!(2q-1)!}\\ &\qquad\times a^{2q-1}\\ B_{2q-1}{}^{2n-1}&=\sum_{p=q}^{n}\frac{(-1)^{n-p}(2n-1)(n+p-2)!2^{2p-2}b^{2(p-q)}}{(n-p)!(2p-2q)!(2q-1)!}\\ &\qquad\times a^{2q-1}\\ B_{2q}{}^{2n-1}&=\sum_{p=q+1}^{n}\frac{(-1)^{n-p}(2n-1)(n+p-2)!2^{2p-2}}{(n-p)!(2p-2q-1)!(2q)!}\\ &\qquad\times b^{2(p-q)-1}a^{2q}.\end{aligned} \tag{32}$$

Thus, the Tchebycheff polynomial is expressed as a power series in $x=\cos\psi$. Now substitute the expansions[10]

$$\cos^{2p}\psi=\frac{1}{2^{2p-1}}\sum_{m=0}^{p}\frac{\epsilon_{p-m}}{2}\binom{2p}{m}\cos(2p-2m)\psi$$

$$\cos^{2p-1}\psi=\frac{1}{2^{2p-2}}\sum_{m=0}^{p-1}\binom{2p-1}{m}\cos(2p-2m-1)\psi$$

$$\text{where}\quad \epsilon_{p-m}\ \begin{aligned}&=1\quad\text{for}\quad p-m=0\\ &=2\quad\text{for}\quad p-m\neq 0\end{aligned}$$

into (31), regroup terms and obtain

$$T_N(\psi)=\sum_{n=0}^{N}C_n{}^N\cos n\psi \tag{33}$$

where

$$\begin{aligned}C_{2q}{}^{2n}&=\frac{\epsilon_q}{2}\sum_{s=q}^{n}\frac{B_{2s}{}^{2n}}{2^{2s-1}}\binom{2s}{s-q}\\ C_{2q-1}{}^{2n}&=\sum_{s=q}^{n}\frac{B_{2s-1}{}^{2n}}{2^{2s-2}}\binom{2s-1}{s-q}\\ C_{2q-1}{}^{2n-1}&=\sum_{s=q}^{n}\frac{B_{2s-1}{}^{2n-1}}{2^{2s-2}}\binom{2s-1}{s-q}\\ C_{2q}{}^{2n-1}&=\frac{\epsilon_q}{2}\sum_{s=q}^{n-1}\frac{B_{2s}{}^{2n-1}}{2^{2s-1}}\binom{2s}{s-q}.\end{aligned} \tag{34}$$

[10] Hobson, "Plane Trigonometry," Cambridge University Press, pp. 280; 1918.

The expressions (32) may be substituted for the B's above if desired. Thus, the coefficients of the Fourier series for T_N may be evaluated by using (32), (33) and (34). This is an exact finite Fourier series representation of T_N.

For small values of N, it is a simple task to evaluate the C_n^N's in terms of a and b. However, if N is doubled the labor is quadrupled so that for large N (such as 20 or more) the procedure is quite laborious. For large N it would probably be easier to use numerical methods of Fourier analysis.

Appendix II gives the C_n^N's in terms of a and b up to $N=8$.

Appendix II

The Fourier coefficients for the Tchebycheff patterns, T_2 to T_8 are given below.

T_2 pattern

$$C_0^2 = -1 + a^2 + 2b^2$$
$$C_1^2 = 4ab$$
$$C_2^2 = a^2$$

T_3 pattern

$$C_0^3 = -3b + 4b^3 + 6ba^2$$
$$C_1^3 = -3a + 12b^2a + 3a^3$$
$$C_2^3 = 6ba^2$$
$$C_3^3 = a^3$$

T_4 pattern

$$C_0^4 = 1 - 8b^2 + 8b^4 + 24a^2b^2 + 3a^4 - 4a^2$$
$$C_1^4 = -16ab + 32ab^3 + 24a^3b$$
$$C_2^4 = -4a^2 + 24a^2b^2 + 4a^4$$
$$C_3^4 = 8a^3b$$
$$C_4^4 = a^4$$

T_5 pattern

$$C_0^5 = 5b - 20b^3 + 16b^5 - 30ba^2 + 8b^3a^2 + 30a^4b$$
$$C_1^5 = 5a - 60b^2a + 80b^4a - 15a^3 + 120b^2a^3 + 10a^5$$
$$C_2^5 = -30ba^2 + 80b^3a^2 + 40a^4b$$
$$C_3^5 = -5a^3 + 40b^2a^3 + 5a^5$$
$$C_4^5 = 10a^4b$$
$$C_5^5 = a^5$$

T_6 pattern

$$C_0^6 = -1 + 18b^2 - 48b^4 + 32b^6 + 9a^2 - 144b^2a^2 + 240b^4a^2 - 18a^4 + 180b^2a^4 + 10a^6$$
$$C_1^6 = 36ba - 192b^3a + 192b^5a - 144ba^3 + 480b^3a^3 + 120ba^5$$
$$C_2^6 = 9a^2 - 144b^2a^2 + 240b^4a^2 - 24a^4 + 240b^2a^4 + 15a^6$$
$$C_3^6 = -48ba^3 + 160b^3a^3 + 60ba^5$$
$$C_4^6 = -6a^4 + 60b^2a^4 + 6a^6$$
$$C_5^6 = 12ba^5$$
$$C_6^6 = a^6$$

T_7 pattern

$$C_0^7 = -7b + 56b^3 - 112b^5 + 64b^7 + 84ba^2 - 560b^3a^2 + 672b^5a^2 - 210ba^4 + 840b^3a^4 + 140ba^6$$
$$C_1^7 = -7a + 168b^2a - 560b^4a + 448b^6a + 42a^3 - 840b^2a^3 + 1680b^4a^3 - 70a^5 + 840b^2a^5 + 35a^7$$
$$C_2^7 = 84ba^2 - 560b^3a^2 + 672b^5a^2 - 280ba^4 + 1120b^3a^4 + 210ba^6$$
$$C_3^7 = 14a^3 - 280b^2a^3 + 560b^4a^3 - 35a^5 + 420b^2a^5 + 21a^7$$
$$C_4^7 = -70ba^4 + 280b^3a^4 + 84ba^6$$
$$C_5^7 = -7a^5 + 84b^2a^5 + 7a^7$$
$$C_6^7 = 14ba^6$$
$$C_7^7 = a^7$$

T_8 pattern

$$C_0^8 = 1 - 32b^2 + 160b^4 - 256b^6 + 128b^8 - 16a^2 + 480b^2a^2 - 1920b^4a^2 + 1792b^6a^2 + 60a^4 - 1440b^2a^4 + 3360b^4a^4 - 80a^6 + 1120b^2a^6 + 35a^8$$
$$C_1^8 = -64ba + 640b^3a - 1536b^5a + 1024b^7a + 480ba^3 - 3840b^3a^3 + 5376b^5a^3 - 960ba^5 + 4480b^3a^5 + 560ba^7$$
$$C_2^8 = -16a^2 + 480b^2a^2 - 1920b^4a^2 + 1792b^6a^2 + 80a^4 - 1920b^2a^4 + 4480b^4a^4 - 120a^6 + 1680b^2a^6 + 56a^8$$
$$C_3^8 = 160ba^3 - 1280b^3a^3 + 1792b^5a^3 - 480ba^5 + 2240b^3a^5 + 336ba^7$$
$$C_4^8 = 20a^4 - 480b^2a^4 + 1120b^4a^4 - 48a^6 + 672b^2a^6 + 28a^8$$
$$C_5^8 = -96ba^5 + 448b^3a^5 + 112ba^7$$
$$C_6^8 = -8a^6 + 112b^2a^6 + 8a^8$$
$$C_7^8 = 16ba^7$$
$$C_8^8 = a^8$$

The Response of a Tuned Circuit to a Ramp Function*

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Summary—The responses of a series-loaded parallel-tuned circuit to a linearly increasing driving force and to ramp functions with various times of rise have been computed. Curves are given to show the effect of various rise times of the ramp function on circuits with Q's of 0.5, 1, 2, 4, 8, and ∞.

Introduction

THE RESPONSE of a tuned circuit to the unit-step function is known and can be readily calculated by the usual operational methods. In many practical problems it is of interest to know how much the transient is changed if the driving force rises to the final value in a finite time rather than in zero time. This driving force will be called a "ramp function," and is shown by the line OAB of Fig. 1. It can be obtained by the addition of straight lines OAC and DE. Line DE, which starts at the time axis, has the same slope as OAC but the sign is negative. Since the two slopes are equal and opposite, the sum, AB, will have zero slope beyond point A.

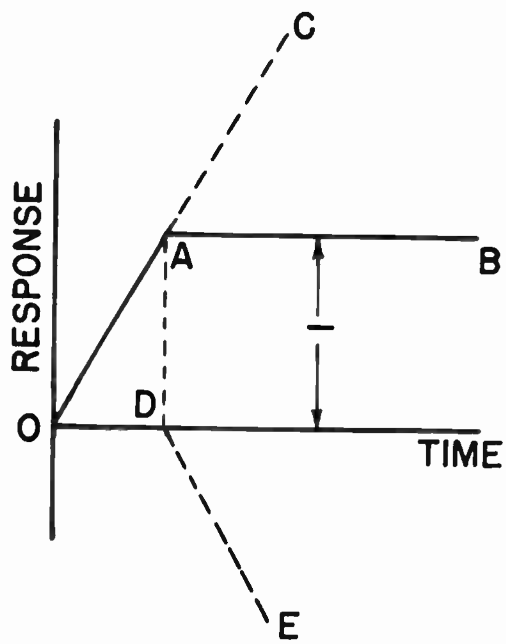

Fig. 1—Ramp function.

Ramp functions are fairly common in electronic circuits. The high voltage power supplies in television receivers are usually driven by a sawtooth with a short-return time. This waveform can be analyzed as the sum of a slow rise with periodic ramp functions superimposed. The effect of the return time on the peak voltage developed across the tuned circuit is determined by the transient response of the tuned circuit to a ramp function. Inductance-controlled deflection circuits use similar waveforms, and the same theory applies.

When a rectangular aperture is used to scan a sudden transition from black to white, the output is a ramp function. Other applications occur in dual standard color receivers where the retrace time of the horizontal sweep must be different for the two line frequencies. Similar problems occur in pulse delay networks and in blocking oscillators.

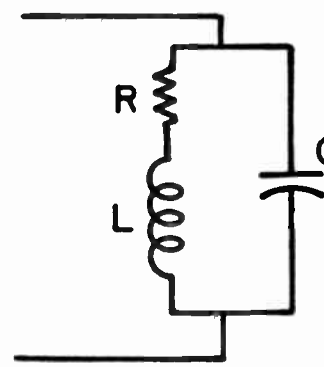

Fig. 2—Tuned circuit.

Theory

The impedance of the tuned circuit of Fig. 2 is given by

$$Z = \frac{1}{C} \frac{p + 2\alpha}{(p+\alpha)^2 + \beta^2}, \tag{1}$$

where

$$p = i\omega, \qquad \alpha = \frac{R}{2L} = \frac{\omega_0}{2Q},$$

$$\beta^2 = \frac{1}{LC} - \frac{R^2}{4L^2} = \omega_0^2 \left\{ 1 - \frac{1}{4Q^2} \right\}$$

and

$$\omega_0^2 = \frac{1}{LC} = \alpha^2 + \beta^2.$$

* Decimal classification: R141.2. Original manuscript received by the Institute, August 18, 1952.

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Let the driving force be $\omega_0 t U(t)$, where $U(t)$ is the unit-step function. The Laplace transform of the driving force is ω_0/s^2. The transform of the system response to a linear increase of current is

$$\frac{\omega_0}{C}\,\frac{s+2\alpha}{s^2\{(s+\alpha)^2+\beta^2\}} = \frac{\omega_0}{C}\left\{\frac{\beta^2-3\alpha^2}{(\alpha^2+\beta^2)^2}\,\frac{1}{s} + \frac{2\alpha}{\alpha^2+\beta^2}\,\frac{1}{s^2} - \frac{\beta^2-3\alpha^2}{(\alpha^2+\beta^2)^2}\,\frac{s+\alpha}{(s+\alpha)^2+\beta^2} + \frac{\alpha^3-3\alpha\beta^2}{(\alpha^2+\beta^2)^2}\,\frac{1}{(s+\alpha)^2+\beta^2}\right\}. \tag{2}$$

Using transform pairs 1.101, 2.101, 1.3031, and 1.301, respectively, of Gardner and Barnes,[1] the system response becomes

$$\text{response} = \sqrt{\frac{L}{C}}\left\{1-\frac{1}{Q^2}+\frac{\omega_0 t}{Q} - \left(1-\frac{1}{Q^2}\right)e^{-\omega_0 t/2Q}\cos\left(1-\frac{1}{4Q^2}\right)^{1/2}\omega_0 t - \frac{3-1/Q^2}{(4Q^2-1)^{1/2}}\,e^{-\omega_0 t/2Q}\sin\left(1-\frac{1}{4Q^2}\right)^{1/2}\omega_0 t\right\}. \tag{3}$$

The cases $Q=1$ and $Q=\frac{1}{2}$ must be evaluated separately.

$$\text{Response} = \sqrt{\frac{L}{C}}\left\{\omega_0 t - \frac{2}{\sqrt{3}}\,e^{-\omega_0 t/2}\sin\frac{\sqrt{3}}{2}\,\omega_0 t\right\} \tag{4}$$

when $Q=1$.

$$\text{Response} = \sqrt{\frac{L}{C}}\,\{2\omega_0 t - 3 + (\omega_0 t + 3)e^{-\omega_0 t}\} \tag{5}$$

when $Q=\frac{1}{2}$.

Let the time of rise, OD of Fig. 1, be $\tau = t/\omega_0$. This means that the rise time is expressed in radians of a cycle at the resonant frequency of the tuned circuit. For example, $\omega_0\tau=\pi$ means that the ramp function reaches the end of the rise, point A, in a period of time corresponding to one-half cycle at the resonant frequency of the tuned circuit.

The response of the system to the linear function OC of Fig. 1 is given by (3), (4), and (5). Similarly, the response to the function DE is given by the same equations but delayed in time by an amount τ. When the two results are subtracted, and the difference divided by $\omega_0\tau$ to normalize to unity, the response to the ramp function is obtained. The response of the system to a unit step of current, $\tau=0$, can be obtained by differentiating equations (3), (4), and (5) with respect to $\omega_0 t$.

[1] M. F. Gardner and J. L. Barnes, "Transients in Linear Systems," John Wiley and Sons, New York, N. Y., vol. 1, pp. 338 ff; 1942.

The curves of Fig. 3 show the response of the tuned circuit to the linearly increasing function $\omega_0 t U(t)$. For low values of Q the transient approaches the dotted steady-state curve very rapidly. All curves start out as a parabola at the origin. The critically damped case corresponds to $Q=\frac{1}{2}$, and in this case the response approaches the steady-state curve asymptotically without oscillation. When $Q\rightarrow\infty$, there is no damping and the response is an undamped cosine wave.

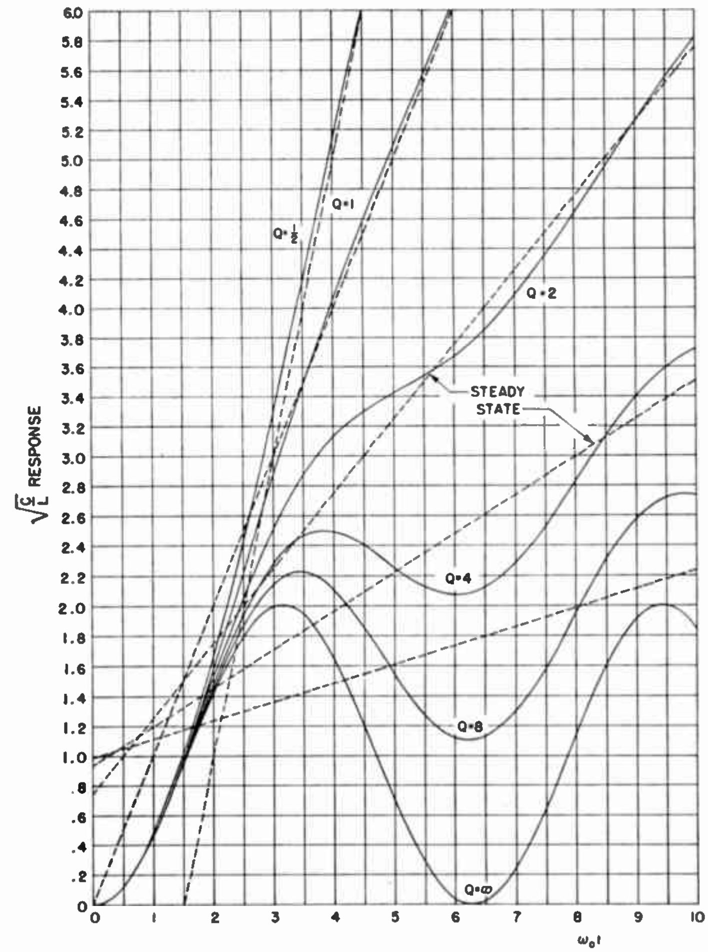

Fig. 3—Response to linearly increasing function.

The response to a unit-step is shown by Fig. 4. The critically damped case, $Q=\frac{1}{2}$, is the highest Q which does not oscillate about the steady-state value. In each case the steady-state value is $1/Q$. All curves start out with unit slope at the origin. As $Q\rightarrow\infty$, the response becomes an undamped sine wave.

The curves of Figs. 5 to 10, inclusive, show the effect of various times of rise of the ramp function on the tuned circuit, as a function of the circuit Q. The parameter $\omega_0\tau$ is the time of rise of the ramp function in radians at the resonant frequency of the tuned circuit. In all of these curves it will be noted that the curve shape is not appreciably affected if the time of rise is less than one radian.

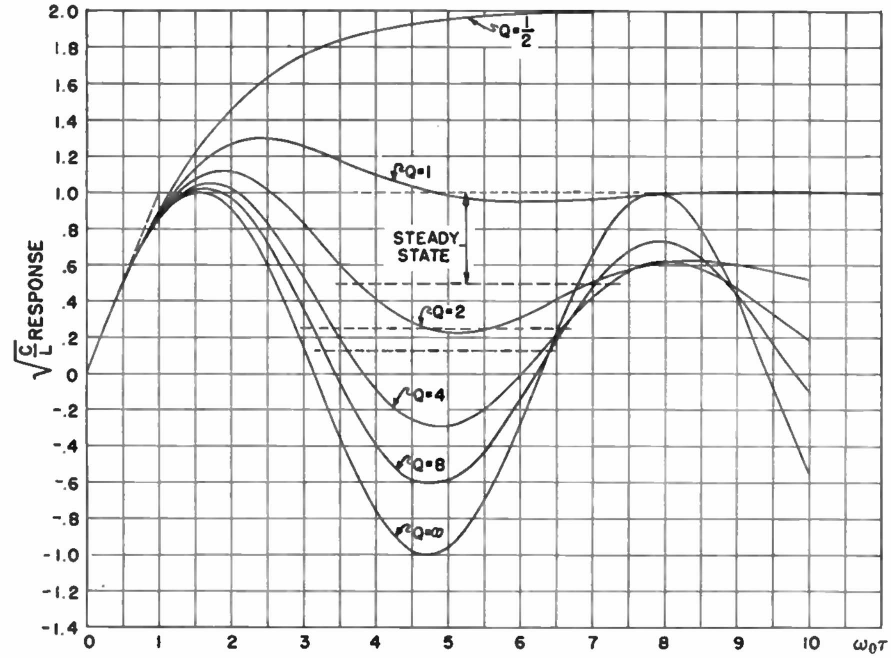

Fig. 4—Unit-step response.

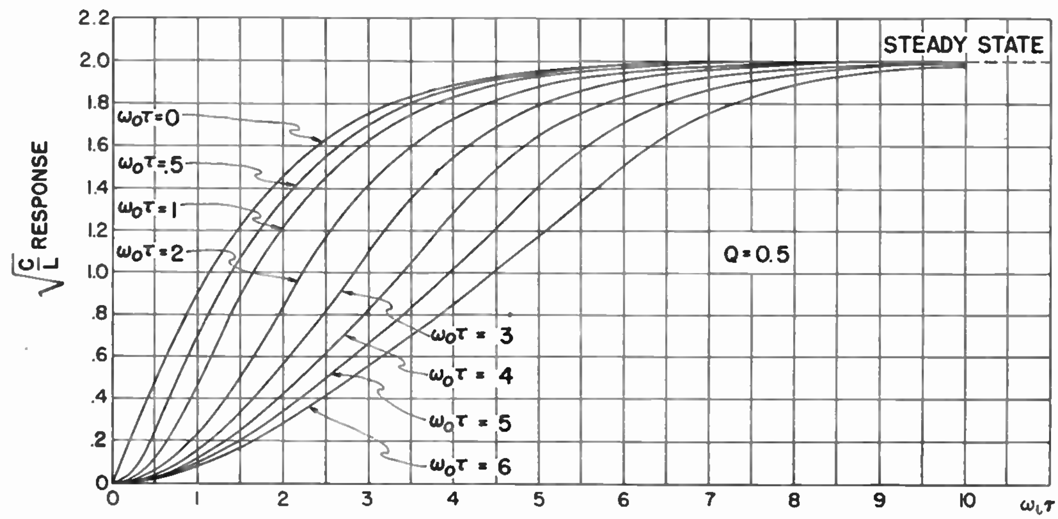

Fig. 5—Response to ramp function, $Q=0.5$.

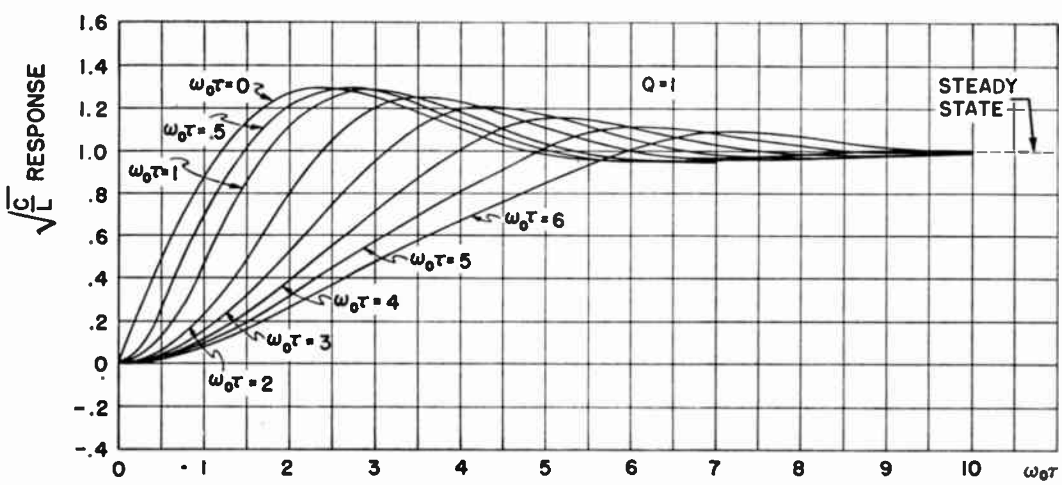

Fig. 6—Response to ramp function, $Q=1$.

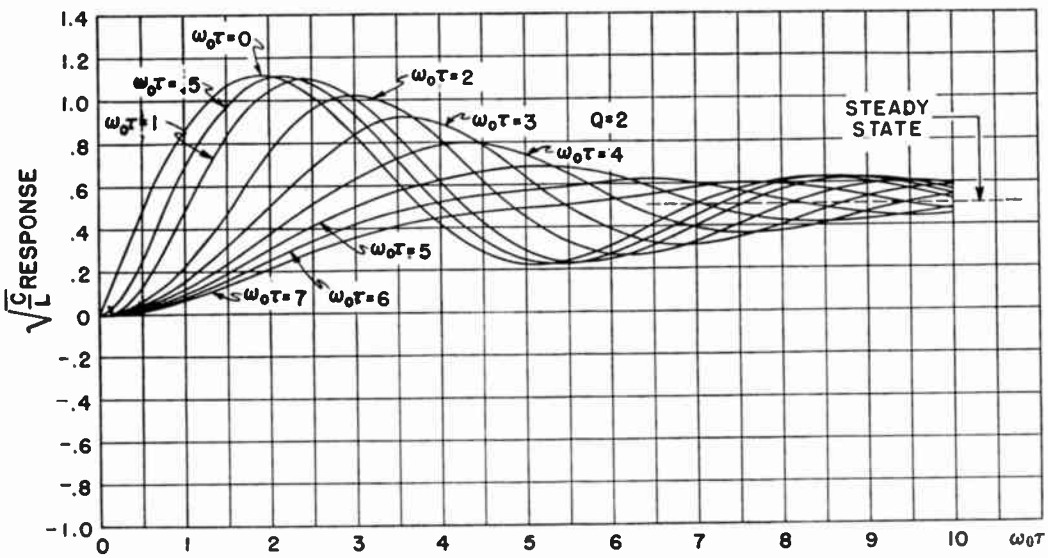

Fig. 7—Response to ramp function, $Q=2$.

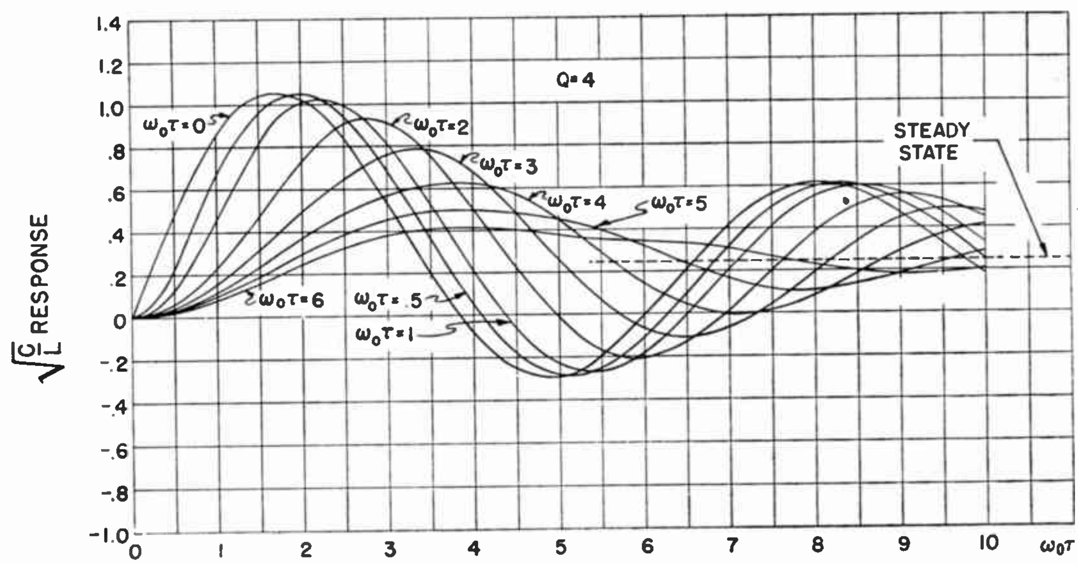

Fig. 8—Response to ramp function, $Q=4$.

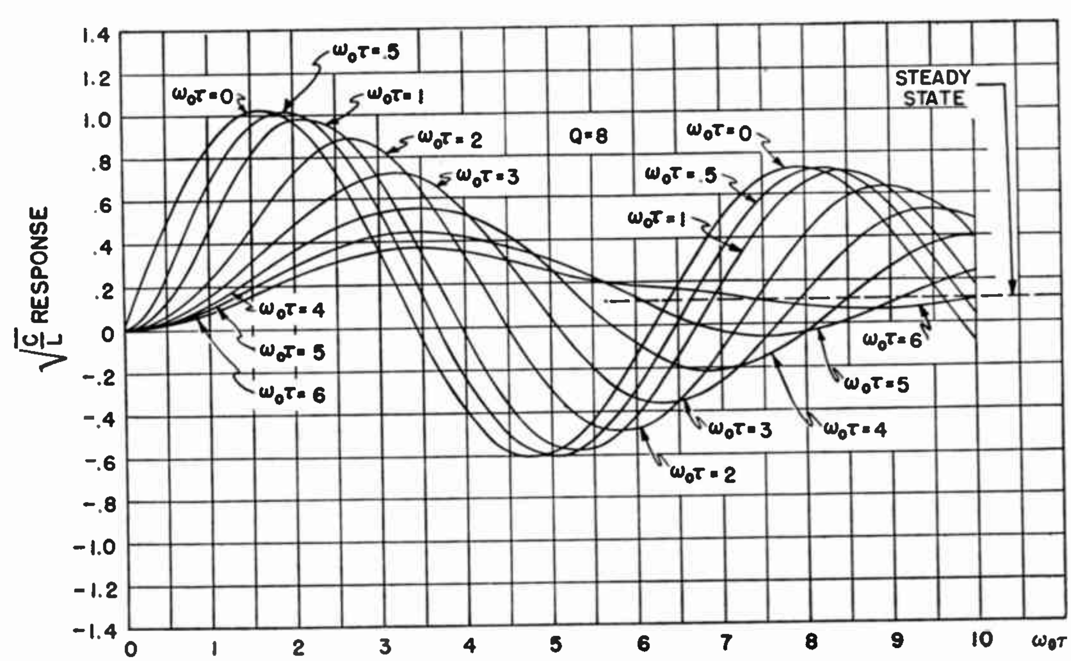

Fig. 9—Response to ramp function, $Q=8$.

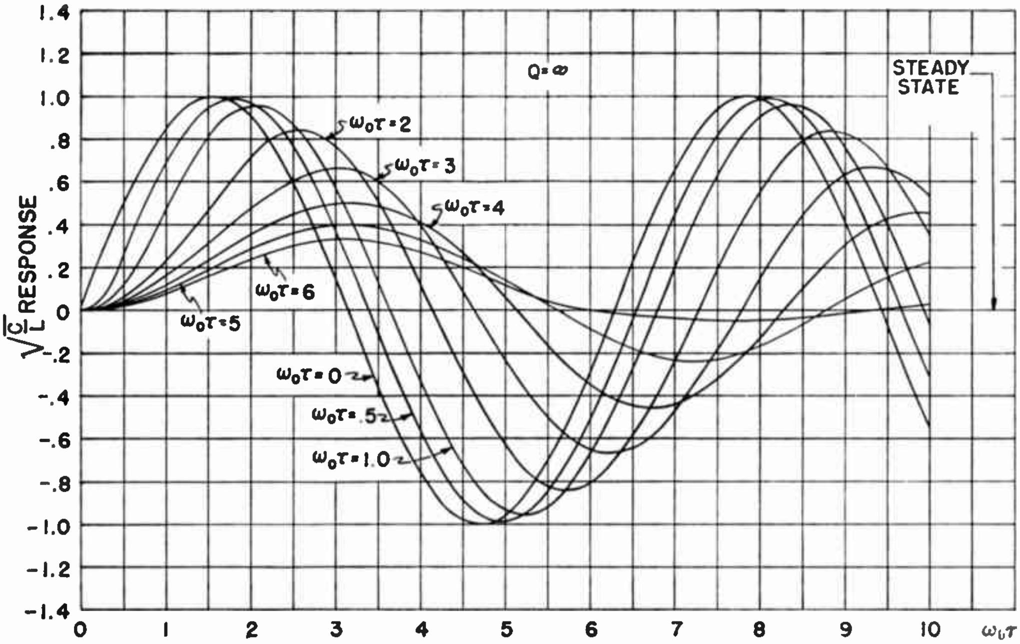

Fig. 10—Response to ramp function, $Q = \infty$.

Conclusions

The responses of a tuned circuit to a linearly increasing driving force, to a unit-step function, and to ramp functions with various rise times have been derived as a function of the circuit Q. In most applications the rise time need not be much less than that corresponding to one radian at the resonant frequency of the tuned circuit. In a high-voltage power supply in a television receiver, which is excited by the part of the horizontal-sweep waveform corresponding to the retrace time, the maximum voltage will be developed when the retrace time is less than a quarter cycle at the resonant frequency of the tuned circuit in the power supply.

In circuits which use a damper tube across the tuned circuit the first positive swing of voltage will be as shown by these curves, but the rest of the curve, starting with the negative excursion, will be removed by the rectifier.

Contributors to Proceedings of the I.R.E.

Percy W. Bridgman was born on April 21, 1882, and educated in the public schools of Newton, Mass., and at Harvard University from which he received the Ph.D. degree in physics in 1908.

P. W. Bridgman

Dr. Bridgman has been connected with the Physics Department of Harvard University continuously from graduation to the present time. His principal scientific work has been with the production of high pressure and the measurement of the various physical effects of high pressure.

Dr. Bridgman's work has been described in a number of technical publications and in the book, "The Physics of High Pressure." In addition, some attention has been given to critical analysis of the fundamental concepts of physics. Essential ideas of this analysis are set forth in the book, "The Logic of Modern Physics."

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From 1951 to 1952 Dr. DuHamel worked on television antennas at RCA Laboratories, Princeton, N. J. Since 1952, he has been a research assistant professor in Electrical Engineering at the University of Illinois, where he is in charge of the Antenna Laboratory.

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EDWARD F. HARRIS

From 1944 to 1946 as a member of the Electronics Training Group of the Signal Corps, he served as Signal Officer of the Public Safety Division GHQ in Japan, Prior to 1944 he was engaged in an NDRC antenna research program at Ohio State University.

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M. G. MORGAN

Dr. Morgan served as a staff engineer at the Navy Electronics Research Group at the University of California during 1946–1947. He was an assistant professor of electrical engineering and assistant dean of the Thayer School of Engineering, Dartmouth College, from 1947 to June, 1950, when he was appointed director of research, his present position. Since October, 1948, he has been chief investigator on a program of ionospheric research sponsored by the Office of Naval Research. He is a member of the American Society for Engineering Education, Sigma Xi, and past-president of the Intercollegiate Ski Union. He is a member of the IRE Wave Propagation Committee.

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Arnold T. Nordsieck was born on January 5, 1911 in Marysville, Ohio. He received his M.S. degree in physics from Ohio State University in 1932, and Ph.D. from the University of California in 1935. From 1935 to 1937 he was National Research Fellow in physics at Leipzig, Germany and at Stanford University, California.

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From 1937 to 1941 Dr. Nordsieck taught physics and did research in theoretical physics and in microwaves at Columbia University. From 1942 to 1947 he was a member of the technical staff of Bell Telephone Laboratories. In 1945–46 he was associate professor of physics at Columbia University. Since 1947 he has been professor of physics at the University of Illinois. His main interests are in theoretical physics.

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C. M. VERHAGEN

From 1942 to 1945 Mr. Verhagen worked for Philips Incandescent Lamp Works of Eindhoven as a research assistant. He joined the Netherlands Telegraph Company "Radio Holland" in 1946, where he supervised the construction and service of radar equipment on board merchant ships.

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❖

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W. B. WHALLEY

In 1936 Mr. Whalley was development engineer at the Radio Valve Company in Toronto. From 1937 to 1940 he was with RCA Manufacturing Company in Harrison, N. J., and in 1940 transferred to war work on radar systems and cathode-ray tubes in Canada. In 1943 he returned to the Radio Corporation of America, doing research on radar and Television transmitting tubes and circuits at the RCA Laboratories Division, Princeton. In 1947 he was appointed Assistant Professor of Engineering Physics at Cornell University. At present he is head of the Applications Research Laboratory of Sylvania Electric Products Inc., Bayside, N. Y.

Mr. Whalley is a member of Sigma Xi, American Association of University Professors, and American Physical Society, and American Institute of Electrical Engineers.

Correspondence

Revised Specifications for Field Test of NTSC Compatible Color Television*

Listed below is the full text of "Revised Specifications for Field Test of NTSC Compatible Color Television" recently approved by the National Television System Committee. These specifications differ in certain particulars from those announced by the NTSC on Nov. 26, 1951. The changes were made to improve the performance of the signal with respect to compatibility and quality of color reproduction.

These specifications will be used by Panels 15, 16, and 17 of the NTSC in a comprehensive field test of the signal which will be inaugurated shortly.

It is hoped that technical personnel will want to comment on the field test transmissions. Comments on reception of the signal should be sent to the undersigned.

Revised Specifications

Test Specifications—Group I

1. The image is scanned at uniform velocities from left to right and from top to bottom with 525 lines per frame and nominally 60 fields per second, interlaced 2-to-1.

2. The aspect ratio of the image is 4 units horizontally and 3 units vertically.

3. The blanking level is fixed at 75 per cent (± 2.5 per cent) of the peak amplitude of the carrier envelope. The maximum white (luminance) level is not more than 15 per cent nor less than 10 per cent of the peak carrier amplitude.

4. The horizontal and vertical synchronizing pulses are those specified in Section 3.682 of Subpart E of Part 3 of the FCC Rules Governing Radio Broadcast Services (as amended April 11, 1952; effective June 2, 1952), modified to provide the color-synchronizing signal described in Spec. 21 (Group II of these specifications).

5. Increase in initial light intensity corresponds to a decrease in amplitude of the carrier envelope (negative modulation).

6. The television channel occupies a total width of 6 mc. Vestigial-sideband amplitude-modulation transmission is used for the picture signal in accordance with the FCC Rules cited in Spec. 4, above.

7. The sound transmission is by frequency modulation, with maximum deviation ± 25 kc, and with pre-emphasis in accordance with a 75-μsec time constant. The frequency of the unmodulated sound carrier is 4.5 mc$\pm$1000 cycles above the frequency of the main picture carrier actually in use at the transmitter.

8. The radiated signals are horizontally polarized.

9. The power of the aural-signal transmitter is not less than 50 per cent nor more than 70 per cent of the peak power of the visual-signal transmitter.

Test Specifications—Group II

10. The color picture signal has the following composition:

$$E_m = E_Y' + \{E_Q' \sin(\omega t + 33°) + E_I' \cos(\omega t + 33°)\}$$

where

$$E_Q' = 0.41(E_B' - E_Y') + 0.48(E_R' - E_Y')$$

$$E_I' = -0.27(E_B' - E_Y') + 0.74(E_R' - E_Y')$$

$$E_Y' = 0.30E_R' + 0.59E_G' + 0.11E_B'$$

The phase of the color burst is

$$\sin(\omega t + 180°).$$

Notes: For color-difference frequencies below 500 kc, signal can be represented by

$$E_m = E_Y' + \left\{\frac{1}{1.14}\left[\frac{1}{1.78}(E_B' - E_Y')\sin\omega t + (E_R' - E_Y')\cos\omega t\right]\right\}.$$

In these expressions the symbols have the following significance:

E_m is the total video voltage, corresponding to the scanning of a particular picture element, applied to the modulator of the picture transmitter.

E_Y' is the gamma-corrected voltage of the monochrome (black-and-white) portion of the color picture signal, corresponding to the given picture element.

E_R', E_G', and E_B' are the gamma-corrected voltages corresponding to the red, green, and blue signals intended for the color picture tube, during the scanning of the given picture element.

E_Q' and E_I' are the two gamma-corrected orthogonal components of the chrominance signal corresponding respectively to the narrow-band and wide-band axes.

ω is 2π times the frequency of the chrominance subcarrier. The phase reference of this frequency is the color synchronizing signal (see Spec. 21 below) which corresponds to amplitude modulation of a continuous sine wave of the form $\sin(\omega t + 180°)$ where t is the time.

The portion of each expression between brackets represents the chrominance subcarrier signal which carries the chrominance information.

It is recommended that field-test receivers incorporate a reserve of 10-db gain in the chrominance channel over the gain required by the above expressions.

11. The primary colors referred to by E_R', E_G', and E_B' have the following chromaticities in the CIE system of specification:

	x	y
Red (R)	0.67	0.33
Green (G)	0.21	0.71
Blue (B)	0.14	0.08

12. Color signal is so proportioned that when chrominance subcarrier vanishes, chromaticity reproduced corresponds to Illuminant $C(x = 0.310,\ y = 0.316)$.

13. Gamma correction is such that the desired pictorial result shall be obtained on a display device having a transfer gradient (gamma exponent) of 2.75. The equipment used shall be capable of an overall-transfer gradient of unity with a display device having a transfer gradient of 2.75. The voltages E_Y', E_R', E_G', E_B', E_Q', and E_I' in the expression of Spec. 10, above, refer to the gamma-corrected signals.

14. The color subcarrier frequency is 3.579545 mc$\pm$0.0003 per cent with a maximum rate of change not to exceed 1/10 cycle per second per second.

15. The horizontal-scanning frequency is 2/455 times the color-subcarrier frequency. This corresponds nominally to 15,750 cycles per second (the actual value is 15,734.264$\pm$0.047 cycles per second).

16. The bandwidth assigned to the monochrome signal E_Y' is in accordance with the FCC standard for black-and-white transmissions, as noted in Spec. 6.

17. The bandwidth assigned prior to modulation to the color-difference signals E_Q' and E_I' is given by Table I.

TABLE I

Q-channel bandwith
- at 400 kc less than 2 db down
- at 500 kc less than 6 db down
- at 600 kc at least 6 db down

I-channel bandwidth
- at 1.3 mc less than 2 db down
- at 3.6 mc at least 20 db down

18. E_Y', E_R', E_G', E_B', E_Q', and E_I' are all matched to each other in time to within ± 0.05 μsec. This is a tentative tolerance to be established definitely later.

19. The overall-transmission bandwidth assigned to the modulated-chrominance subcarrier shall extend to at least 1.5 mc below chrominance-subcarrier frequency and to at least 0.6 mc above chrominance-subcarrier frequency, at an attenuation of 2 db.

20. A sinewave, introduced at those terminals of the transmitter which are normally fed the color picture signal, shall produce a radiated signal having an envelope time delay, relative to 0.1 mc, of zero μ sec up to a frequency of 2.5 mc; and then linearly decreasing to 4.3 mc so as to be equal to -0.26 μsec at 3.579545 mc. The tolerance on all these delays shall be ± 0.05 μsec relative to the delay at 0.1 mc.

21. The color synchronizing signal is that specified in Fig. 1.

22. The field strength measured at any frequency beyond the limits of the assigned channel shall be at least 60 db below the peak carrier level.

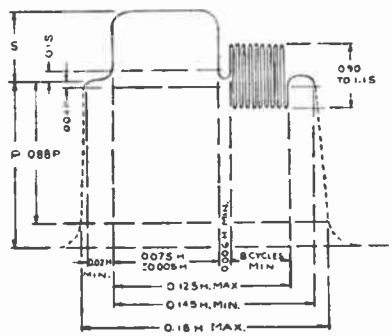

Fig. 1—Revised specifications for field test of NTSC compatible color television.

Notes

1. Radiated-signal envelope shall correspond to the modulating signal of above figure, as modified by transmission characteristics of specification number 6.
2. The burst frequency shall be the frequency specified for the chrominance subcarrier. The tolerance on the frequency shall be ± 0.0003 per cent with a maximum rate of change of frequency not to exceed 1/10 cycle per second per second

Correspondence

3. The horizontal-scanning frequency shall be 2/455 times the burst frequency.
4. Burst follows each horizontal pulse, but is omitted following the equalizing pulses and during the broad vertical pulses.
5. Vertical blanking 0.07 to 0.08V.
6. The dimensions specified for the burst determine the times of starting and stopping the burst, but not its phase.
7. Dimension "P" represents the peak-to-peak excursion of the luminance signal, but does not include the chrominance signal.

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* Received by the Institute, February 11, 1953.

Note on the Iterated Network and Its Application to Differentiators*

The author would like to offer the following alternative proof of the formula for a power of a matrix given recently by Pease,[1] and to point out some variations in the literature in the definitions of the functions used.

This proof is based on Sylvester's theorem[2] giving for a polynomial $P(A)$ of a matrix A, of order m and with eigenvalues $\lambda_1 \cdots \lambda_m$,

$$P(A) = \sum_{r=1}^{m} P(\lambda_r) \frac{\prod_{s\neq r} (\lambda_s I - A)}{\prod_{s\neq r} (\lambda_s - \lambda_r)}. \quad (1)$$

In the case in which $P(A) = A^n$,

$$A^n = \sum_{r=1}^{m} \lambda_r^n \frac{\prod_{s\neq r} (\lambda_s I - A)}{\prod_{s\neq r} (\lambda_s - \lambda_r)}. \quad (2)$$

I, of course, is the unit matrix of order m.

Since, in the present case, the matrix is that of a 4-terminal, $m=2$, and the eigenvalues are

$$\lambda = \tfrac{1}{2}\{a_{11} + a_{22} + \sqrt{(a_{11}+a_{22})^2/4 - (a_{11}a_{22} - a_{21}a_{12})}\}$$
$$= x + \sqrt{x^2 - 1}, \quad (3)$$

and

$$1/\lambda = x - \sqrt{x^2 - 1}. \quad (4)$$

The a's are elements of A, $x = a_{11} + a_{22}/2$, and, since A is unitary, $a_{11}a_{22} - a_{21}a_{12} = 1$. Inserting (3) and (4) into (2),

$$A^n = \lambda^n \frac{\lambda^{-1}I - A}{\lambda^{-1} - \lambda} + \lambda^{-n} \frac{\lambda I - A}{\lambda - \lambda^{-1}}$$
$$= \frac{(\lambda^n - \lambda^{-n})A - (\lambda^{n-1} - \lambda^{1-n})I}{\lambda - \lambda^{-1}}. \quad (5)$$

Equations (3) and (4) make this

$$A^n = \frac{(x+\sqrt{x^2-1})^n - (x-\sqrt{x^2-1})^n}{2j\sqrt{1-x^2}} A$$
$$- \frac{(x+\sqrt{x^2-1})^{n-1} - (x-\sqrt{x^2-1})^{n-1}}{2j\sqrt{x^2-1}} I. \quad (6)$$

The Tchebycheff function $U_n(X)$ has been defined in at least two different ways.[3,4] If one takes the "equal ripple" function

$$U_n(x) = \sin n \arccos x$$
$$= \frac{1}{2j}\{(x + \sqrt{x^2-1})^n - (x - \sqrt{x^2-1})^n\} \quad (7)$$

and its modification

$$U_{n-1}^*(x) = \frac{U_n(x)}{\sqrt{1-x^2}}, \quad (8)$$

then

$$A^n = AU_{n-1}^*(x) - IU_{n-2}^*(x). \quad (9)$$

On the other hand, the notation

$$U_n(x) = \frac{\sin n \arccos x}{\sqrt{1-x^2}}$$
$$= \frac{(x+\sqrt{x^2-1})^n - (x-\sqrt{x^2-1})^n}{2j\sqrt{1-x^2}} \quad (10)$$

makes the formula

$$A^n = AU_n(x) - IU_{n-1}(x) \quad (11)$$

in agreement with Pease's formula (11).

H. L. Armstrong
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* Received by the Institute, September 15, 1952.

[1] M. C. Pease, "The iterated network and its application to differentiators," Proc. I.R.E., vol. 40, p. 709; June, 1952.

[2] R. A. Frazer, W. J. Duncan and A. R. Collar, "Elementary Matrices," Cambridge University Press, p. 78 et seq.; 1947.

[3] W. Klein, "Tschebyscheffsche Funktionen," *Arch. für Elektrotech.*, vol. 39, p. 647; 1950.

[4] B. van der Pol and T. J. Weijers, "Tchebycheff polynomials and their relation to circular functions, bessel functions, and Lissajous figures," *Physica*, vol. 1, p. 78; 1933.

Electronics—How Shall It Be Defined?*

Words and terms require definitions. A single word is desirable, such as house, locomotive, mountain. But often a definite thing or concept cannot be expressed by one word, such as machine screw, connecting rod, drill press. Each such thing requires two words.

The object of any definition is to define. Webster's Unabridged Dictionary has the following to say:

"*Define:* to state explicitly, clearly.

Definition: limitation, setting of limits, a word or group of words expressing the essential nature of a person or thing."

Hence, the object of any definition is to separate the term from all other terms so that the reader obtains a clear, definite idea of the thing which the term covers. A certain amount of narrowness is a necessary part of any good, clear definition—as narrow as needed to avoid confusion.

Other examples from Webster's Unabridged Dictionary are:

Locomotive: "A self-propelled engine or vehicle, specifically, a steam engine mounted with its boiler and accessories on a truck or trucks, designed to run on gaged rails, for hauling cars, wagons, etc., and includes electric, gasoline, compressed air and steam."

Note. I notice that it does not include all self-propelled vehicles, such as automobiles, bicycles, motorcycles, motorboats.

Wagon: "A kind of four-wheeled vehicle, especially one used for carrying freight or merchandise."

Razor: "A keen-edged cutting instrument used in shaving one's hair."

General use is not always the true criterion. Witness the word "potentiometer" as applied by newcomers into the electrical field. Fortunately, we were able to standardize on the word "voltage divider" because that device was not a meter, it did not measure anything, but did enable the user to adjust a voltage as desired.

Thus it follows that in defining "electronics" we should favor the narrow region of wording, rather than the broad, inclusive wording. The latter usage spreads the meaning over such a wide field that the reader will not know what "electronics" really means.

A glossary is not for the man who has grown up in the business and is familiar with all its terminology. It is for the young man just entering engineering who really desires to learn.

"Electronics" should not be defined broadly by the ultimate service which it renders to mankind, but by what it really is. We define a machine screw, a nail, a data file, an iron pipe, by what it really is, not by its large usefulness in our civilization.

Though electronic devices are used in telephony, yet the overwhelming percentage of telephone connections uses no electronics. The granular carbon transmitter, local battery and common battery transmission of the voice, the manual and automatic switching of telephone lines all use electricity as it was known before electronics arrived.

The above are the reasons why I urge that there be adopted some definition like the following:

Electronics is that branch of science and technology in which the electron nature of electricity is important, such as the conduction of electricity through gases, semiconductors, or in vacuo.

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* Received by the Institute November 17, 1952.

Electronics*

Herein is a definition of electronics proposed by Dr. Imre Molnar to replace 70.05.005.

Electronics is that branch of science and technology which relates to the propagation of electricity when the electronic theory of matter is required to explain phenomena satisfactorily; and which relates to devices and systems where the application of such principles is predominant.

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Redefining Electronics*

Everitt's intention was to orient the direction of thought connected with the term "electronics," and he succeeded in doing this.[1] But when he attempted to delimit the field of reference, his proposition suffered a curious transposition in concept, and thereby becomes difficult to justify.

"Fundamentally," he says, "electronics is interested primarily in extending man's senses in space. . . ." And he later qualifies this concept by stating formally, "Electronics . . . deals primarily with the supplementing of man's senses and his brain power. . . ."

The technique of television or any other electronic device does not involve any integration of man's sensing equipment with the device. Certainly there is extension involved in what is achieved with electronic equipment. Any supplementation which occurs, however, is related to the information, rather than the sense organs to which it is presented.

The concept of "the supplementing of man's senses" is the point of transposition. It involves an inference which leaps over natural logic, and perhaps was intended to do so because the case of the word "chivalry" is cited. Conceptual continuity is logical even in this case when one realizes that chivalry is a symbol chosen to represent a mode of human behavior.

It is therefore suggested that definition of electronics might follow these lines:

> Electronics is the science of free electron behavior, particularly as it refers to the control of electron movement in conductive materials when used to implement specific purposes, such as telecommunication, servomechanisms, computers, and generally in contrivances which utilize phenomena associated with movement of free electrons in a conductive medium.

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* Received by the Institute, August 25, 1952.
[1] W. L. Everitt, "Let us redefine electronics," Proc. I.R.E., vol. 40, p. 899; August, 1952.

Comment on Spencer's Redefinition of Electronics*

Spencer[1] has raised questions which are most appropriate in connection with my suggestion that a broader concept of electronics, common in the thinking of most engineers, be recognized in some form such as my proposed definition. At most I should be accused only of verbal extrapolation, of taking into account not only our present concept in the use of the word "electronics," but also the rate of change of this concept during the past ten years. If my definition is believed generally to be too broad for the word "electronics," then we will need to coin a new word to fit the important concept described by my definition. Personally, I would have preferred the word "communication," but I must admit that common usage has developed this word in still another direction, including, as it does, such media as newspapers, books, and even public transportation, and not being generally accepted as applying to the transmission of information for the use of machines.

I cannot agree that "the technique of television . . . does not involve any integration of man's sensing equipment with the device." Television standards must, and do, fit conditions dictated by the acuity of the eye, its persistence of vision, its response to flicker, and in the case of more recent proposals, the way in which the eye observes the whole gamut of colors in nature. Only thus can the maximum useful information be transmitted in a given bandwidth. Similarly, other electronic systems, presenting information to the human senses or to machines, must consider in their design the nature of the receptors which make use of the signals. The control of electron movement is only a part of the complex picture.

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* Received by the Institute, September 9, 1952.
[1] R. E. Spencer, "Redefining electronics," Proc. I.R.E., vol. 41, p. 668, this issue.

Redefining Electronics*

In a recent issue Everitt attempts a redefinition of the word *electronics*.[1] It is advisable, in lexicography, to attempt the difficult task of world-wide standardization, so perhaps a comment may be ventured from this country.

It is a pity that the examples Everitt chooses of words with changing meaning both fall outside the scientific field. The effort should always be made from the outset to achieve precision and permanence in scientific terminology. If "gravity" meant something different to us from what it meant to Newton, or if Benjamin Franklin's "lightning" was now taken to include, say, cosmic rays and the aurora borealis, mutual understanding would be greatly hindered.

The proposed new definition of *electronics* is altogether too wide. It is, in fact, almost synonymous with *instrumentation*, and if adopted, some new word would be needed when it was required to distinguish between pneumatic instrumentation and what is at present called electronic instrumentation. Moreover, such a definition would seem to exclude any thermionic tube intended for power generation rather than information handling.

While it may be granted that the old definition quoted is too narrow, I would contend that the only broadening necessary concerns the state of the electrons involved. If, instead of limiting them to flow through a gas, we include in *electronics* any device in which electrons travel otherwise than along normal conductors, then I maintain that the definition has been brought up to date without stretching it beyond current usage.

B. E. Noltingk
Wraystone, Wray Park Road
Reigate, Surrey, England

* Received by the Institute, October 15, 1952.
[1] W. L. Everitt, "Let us redefine electronics," Proc. I.R.E., vol. 40, p. 899; August, 1952.

More on the Delay Multivibrator*

I had analyzed the delay multivibrator circuit[1] and my results were essentially similar to those obtained by Keith Glegg.[2] There is a difference, however, in the expression for the plate current of the first tube during the pulse, and consequently in the one for the duration of the pulse. My result is

$$I_{p1} = V_b\left[\frac{1+\mu x}{\alpha}\right] \tag{1}$$

and

$$\frac{T}{CRg} = \log_e \frac{1-\dfrac{R_k}{R_2'}+R_{L_1}\dfrac{1+\mu x}{\alpha}}{\left(1+\dfrac{1}{\mu}\right)\left(1-R_k\dfrac{1+\mu x}{\alpha}\right)}, \tag{2}$$

where

x = the ratio of the potential of the first grid to the supply potential of V_b,

$R_2' = R_K + R_{p2} + R_{L_2}$,

$\alpha = R_p + R_{L_1} + (\mu+1)R_k$,

and other symbols have the same meaning as in Glegg's paper. On expanding (2) in power series and equating the coefficient of x^2 to zero, we obtain the condition for linearity as

$$R_{L_1} = R_k\left(1+\frac{2R_k}{\alpha}-\frac{R_k}{R_2'}\right). \tag{3}$$

It is interesting to note that the coefficient of x^2 as obtained in the above expansion becomes identical with that obtained by Glegg, if

$$R_{L_1} = R_k\left(\frac{R_k}{R_2'}-1\right)$$

or

$$= \frac{(R_k - R_2')\left[(2\mu+1)R_k + 2R_p\right]}{3R_2' - 2R_k}.$$

Both these conditions, however, require R_{L_1} to be negative.

On the assumption that the change in the grid potential of the second tube during the pulse is small compared to V_b (an essential condition for linearity), I was able to obtain an additional relation between R_{L_1} and R_k, *viz.*,

$$R_{L_1} = \frac{\mu+1}{\mu}R_k \tag{4}$$

as a condition for greatest linearity.

A full derivation of these expressions may be found in footnote reference 1.

Y. D. Altekar
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* Received by the Institute, January 11, 1953.
[1] Y. D. Altekar, "A method for obtaining the products of two variables or squares of variables electronically," *Jour. Sci. Ind. Res.*, vol. 10B, pp. 237–243; 1951.
[2] K. Glegg, "Cathode coupled multivibrated operation," Proc. I.R.E., vol. 38, pp. 655–658; June, 1950.

Correspondence

Helical Winding Slow-Wave Structures in Exponential Line Pulse Transformers*

Pulse transformers using sections of exponential transmission line have been described by Schatz and Williams.[1,2] Where appreciable transformation ratios or pulse lengths of more than a few mμsec are involved, the line sections required have resulted in structures which are frequently unwieldy and difficult to fabricate. Attempts to increase the ratio of electrical length to physical length by using solid or liquid dielectric materials have met some success,[2] but the improvements attainable are limited, and, furthermore, the slow-wave system in which capacitance per unit length is increased is one of undesirably low impedance level. Studies have recently been made of pulse transformers suitable for use at the high impedance levels required in magnetron pulsing service; slow-wave structures using shielded helices of varying pitch have been found suitable for this service. This note describes an experimental pulse transformer of this type.

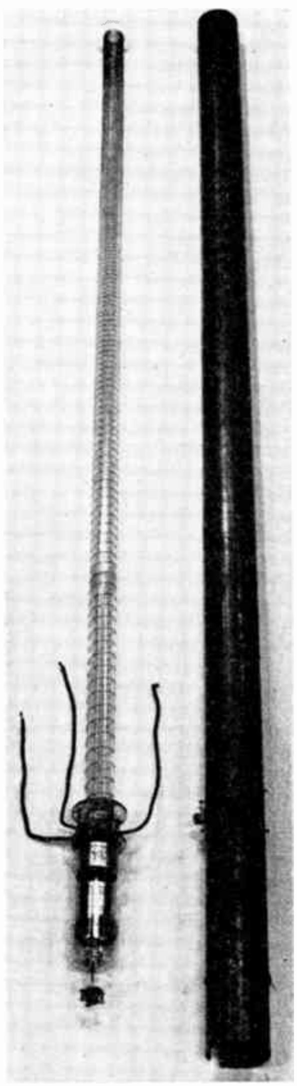

Fig. 1—Exponential-line section pulse transformer consisting of tapered helix and cylindrical outer shell, together with integral hydrogen-thyratron pulse generator.

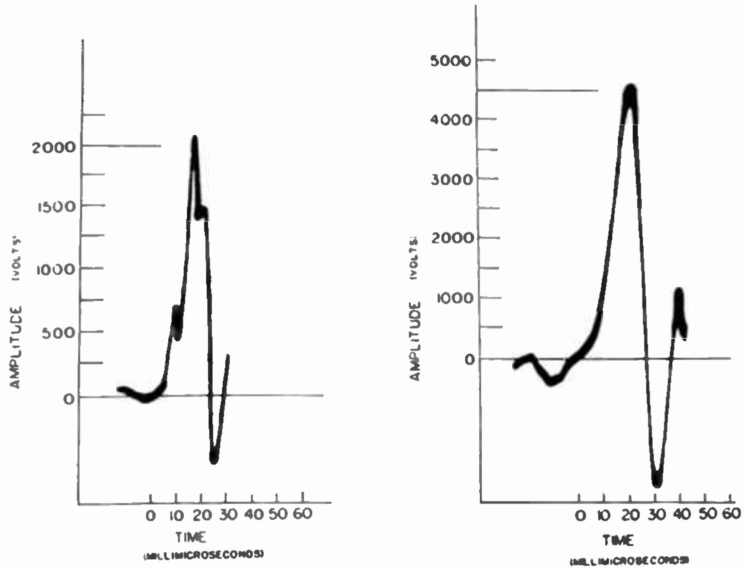

Fig. 2—Left, input pulses; right, output pulses; for test of experimental helical-line pulse transformer. The transformer is terminated in a resistive load of 1590 ohms, its nominal output impedance.

The helical winding exponential line pulse transformer consists of a coaxial structure comprising a constant diameter cylindrical outer shell and helical inside conductor. The inside conductor is of constant diameter and varying turn density. Fig. 1 shows the disassembled parts of an experimental unit together with the pulse generator, a hydrogen thyratron, mounted at the low impedance end of the pulse transformer. An integral assembly of pulse generator and transformer is generally necessary in mμsec systems to insure sufficiently broad-band characteristics in the coupling.[3]

The effect of transformer parameters on performance is readily computed,[2] taking into account the variation in ratio of electrical length to physical length along the tapered helix. The accurate determination or design of the helix parameters is, however, very difficult.

Expressions for inductance and capacitance per unit length of a uniform shielded helix are available,[4] to be sure, but these expressions have been derived without inclusion of such disturbing factors as wire size, end effects, and the dielectric constant of supporting forms. Susskind has made experimental measurements of shielded helix parameters over a very limited impedance range.[5] The experimental transformer described was, therefore, designed to approximate specifications using Susskind's data.

Since the wave velocity along the helix axis is a function of helix turn density, it is necessary to vary continuously the physical impedance flare coefficient in order to obtain the uniform electrical flare coefficient necessary for application of earlier analytical expressions for pulse response. Calculations for this purpose are most readily carried out for a transformer using a series of uniform physical flare sections; the effect of small variations in flare coefficient along the line length should not be marked.

The physical constants of the experimental transformer of Fig. 1 are as follows:

Physical length	1.81 meters
Electrical length	21.7 meters
Impedance ratio	290/1590 ohms

Shape factor $ct/e = 1 + 0.014\Delta t$
Δt = pulse duration in mμsec
with $e^{\gamma l/2} = 2.33$.

The factor ct/e permits the calculation of the output voltage with specified input voltages. For instance, with an input pulse of 10 mμsec rise time the ratio of output to input maximum voltage, computed by superposition methods,[1,2] is $0.96 \times \sqrt{1590/290}$, or 2.22. Fig. 2 shows test input and output voltage traces for the experimental transformer under this condition. The experimental results are obviously in good accord with predicted performance.

In the experimental transformer described there has been achieved a reduction in physical length to approximately 8 per cent of that required for a simple air dielectric coaxial structure. Further work is underway to determine the maximum realizable size reduction in a pulse transformer of this type.

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* Received by the Institute, September 26, 1952. The work described was supported in part by the Office of Naval Research under Contracts NR-200(00) and N7 onr 30306.

[1] E. R. Schatz and E. M. Williams, "Pulse transients in exponential transmission lines," Proc. I.R.E., vol. 38, pp. 1208–1212; October, 1950.

[2] E. M. Williams and E. R. Schatz, "Design of exponential line pulse transformers," Proc. I.R.E., vol. 39, pp. 84–86; January, 1951.

[3] J. B. Woodford, Jr., and E. M. Williams, "The initial conduction interval in high-speed thyratrons," Jour. Appl. Phys., vol. 23, pp. 722–724; July, 1952.

[4] F. M. Phillips, "A note on the inductance of screened single-layer solenoids," Proc. I.E.E., vol. 96, Pt. 3, pp. 138–140; 1947.

[5] "Characteristic impedance of shielded coils," TV Eng., vol. 2, p. 26; March, 1951.

Below—President James W. McRae (center) presents special plaques at the Annual Meeting to two founders of the IRE, Alfred N. Goldsmith and John V. L. Hogan (left and right). Seated is Donald B. Sinclair, 1952 president. The third founder, Robert H. Marriott, is deceased.

Above—Dr. A. M. Zarem, wittily performs the duties of toastmaster at the Annual Banquet.

IRE CONVENTION AND

LARGEST

35,642

New York City was the Mecca of the radio engineering world March 23 to 26 when 35,642 engineers and scientists gathered from all parts of the world for the 1953 IRE National Convention, making it the largest engineering meeting ever held. The Waldorf-Astoria Hotel, Grand Central Palace, and Belmont Plaza were filled with members and guests who had come to see a "Preview of Progress," theme of the convention's technical sessions and exhibits.

The record-shattering attendance (7,000 more than in 1952), the smoothness with which the convention was run, and the wealth of technical information presented in the 221 technical papers and 405 exhibits—all added up to the most successful national convention ever held by the Institute.

The convention opened on Monday morning, March 23, with the Annual Meeting of the Institute at which William R. Hewlett gave the principal address. In special ceremonies commemorating the birth of the Institute, bronze plaques were presented to Alfred N. Goldsmith and John V. L. Hogan, who together with Robert H. Marriott (deceased) founded the IRE in 1912. This was followed by the presentation of special pins to the nine surviving Charter Members of the IRE: P. B. Collison, Lee deForest, Lloyd Espensheid, Alfred N. Goldsmith, Frank Hinners, John V. L. Hogan, Greenleaf W. Pickard, Emil J. Simon, and Arthur F. Van Dyck; all of whom were present except Lee deForest.

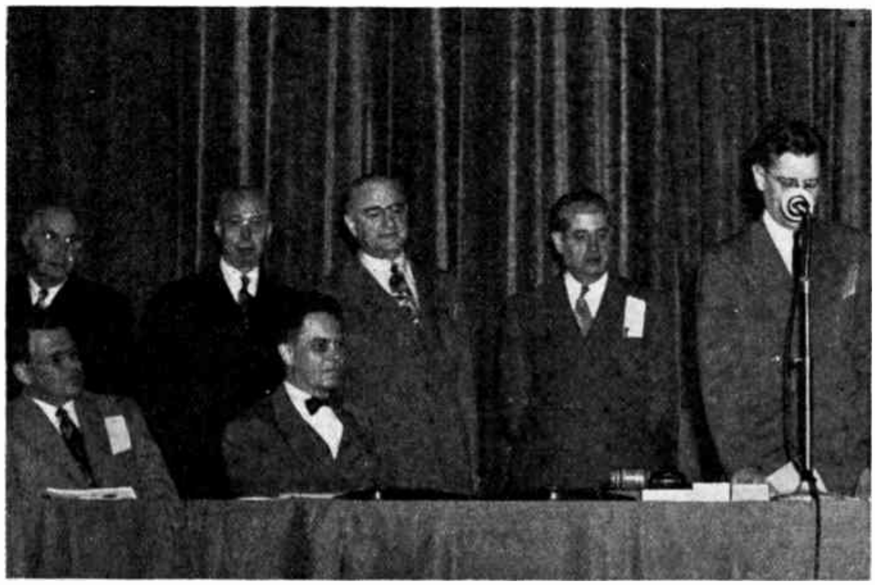
Technical Sessions

The four-day program of technical papers, held at the Waldorf-Astoria, Grand Central Palace, and Belmont Plaza, covered virtually every phase of endeavor in the radio-electronics field (as reported in the March issue of Proceedings). Of the 43 technical sessions, all but three were organized by IRE Professional Groups, reflecting the rapid growth of this important activity.

Highlight of the program was the Tuesday evening symposium on "Electronics in Flight," sponsored jointly with the Institute of the Aeronautical Sciences, during which a distinguished panel of experts discussed the mutual problems of electronic engineers, aerodynamic engineers, and operational people in the aircraft industry.

Group-sponsored symposia were held on

IRE Charter Members



Eight of the nine surviving Charter Members of the IRE honored during special ceremonies at the Annual Meeting of the Institute. *Standing* (left to right): Charter Members P. B. Collison, Lloyd Espensheid, Alfred N. Goldsmith, and Frank Hinners; President James W. McRae; Charter Members John V. L. Hogan,

General scene of the Annual IRE Banquet in the Grand Ballroom of the Waldorf-Astoria Hotel. Recipients of Fellow Awards for 1953, lined up behind speakers' table, are being introduced by Toastmaster Zarem.

Left to right—Dr. A. N. Goldsmith, Editor; C. A. Wiese, R. F. Gehner and C. A Peerenboom, George Banta Publishing Co.; G. W. Bailey, Executive Secretary; E. K. Gannett, Administrative Editor; and J. W. McRae (seated), President.

ENGINEERING SHOW

EVER HELD

ATTEND

Above—William R. Hewlett, who acted as the principal speaker at the Annual Meeting of the Institute.

Below—Brig. Gen. David Sarnoff (center), chairman of the board of RCA and principal speaker at the Annual Banquet, receives the newly established Founders Award from President James W. McRae, while John M. Miller (right), winner of the Medal of Honor, looks on.

a wide variety of topics, such as nucleonics, digital computers, wide-band amplifiers, television broadcasting, instrumentation, microwave equipment, mobile communications, and acoustics. The symposia were complemented by regular sessions on such timely subjects as transistors, information theory, engineering management, and quality control. The program was rounded out with sessions on antennas, electron devices, circuit theory, communication systems, medical electronics, airborne electronics, radio telemetry, remote control systems, audio, and broadcast receivers. All told, the comprehensive program covered the fields of activity of 18 Professional Groups.

AT ANNUAL MEETING

Greenleaf W. Pickard, Emil J. Simon, and Arthur F. Van Dyck. *Seated:* William R. Hewlett, principal speaker; Donald B. Sinclair, 1952 IRE President; W. R. G. Baker, Treasurer; and George W. Bailey, Executive Secretary. The ninth Charter Member, not present, is Lee deForest.

CONVENTION RECORD

An important innovation of this year's convention is the publication of the CONVENTION RECORD OF THE I.R.E., containing all available papers presented at the convention. As reported in detail in the April issue (page 554), the Record will be published in 10 parts, divided according to subjects, approximately two months after the convention. Paid members of Professional Groups (as of April 30) will receive the Part pertaining to the field of their Group free of charge. Copies may be purchased, while available, from the Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y.

EXHIBITS

A high point of the convention was the Radio Engineering Show at Grand Central Palace. Four hundred and five exhibitors filled all four floors of the Palace with $10,000,000 worth of their newest products, making it the largest showing of communications and electronic apparatus ever to be assembled.

Transistors, television, and computers, highlights of the technical sessions, were also much in evidence among the exhibits. Point-contact and junction transistors, some of which are now commercially available, were displayed by several manufacturers. Featured were power transistors and transistorized transmitters, radio receivers, and hearing aids.

Industrial television received a boost with the disclosure of lighter and more compact television cameras. An improved kinescope recording unit, television picture tubes, high-power uhf transmitting tubes,

Some of the more than 35,000 interested spectators at the 1953 IRE Convention crowd around exhibitors' booths at Grand Central Palace.

Birds'-eye view of some of the first-floor exhibits at the 1953 IRE Radio Engineering Show, Grand Central Palace. The show's 405 exhibits occupied 4 floors.

A Vidicon camera which can be plugged into a home television receiver is demonstrated during a technical session by (left to right) G. W. Gray, W. S. Pike, L. E. Flory, and V. K. Zworykin of RCA Laboratories. The closed-circuit system is designed for home, school, and industrial use. Power requirements and the signal pulses for the camera are taken from the receiver through adaptors placed between the tubes and their sockets.

Grant Fairbanks (left) and W. L. Everitt, Univ. of Illinois, demonstrate a speech compressor during a session. The device speeds up words and music without affecting pitch or understandability.

and uhf studio equipment received considerbale attention. Color television was in evidence with the display of equipment for generating NTSC color signals.

The subject of transistors was given further attention at the opening press conference of the convention on Monday afternoon when a special display of transistors and transistorized equipment was shown to the press, accompanied with discussions by leading authorities of Bell Telephone Laboratories, RCA, General Electric Co., Raytheon Manufacturing Co., and the Signal Corps Engineering Laboratories.

A noteworthy feature of the Radio Engineering Show was the manner in which many of the exhibits pertaining to the same fields of interest were grouped together. Thus the visitor who was interested in such subjects as audio, nuclear, computers, instruments, components, and airborne electronics was able to find the displays of particular interest to him with maximum convenience. These centers were representative of only some of the many systems, apparatus, parts, materials, and techniques which went to make up the most informative and comprehensive exhibition in the radio industry.

Social Events

The social activities of the convention received a send-off on the first evening of the convention when a "get-together" cocktail party was held in the spacious Grand Ballroom of the Waldorf-Astoria. The success of the affair is attested to by the fact that 1,800 members and guests attended, providing an excellent opportunity for visitors from all parts of the country to renew old acquaintances and make new ones.

A sell-out crowd of 1,500 attended the Annual IRE Banquet, held in the Grand Ballroom on Wednesday evening, to hear an address by Gen. David Sarnoff, chairman of the board of RCA, on the subject "Electronics and the Engineer." Dr. A. M. Zarem, Stanford Research Institute, ably performed the duties of toastmaster. The occasion was highlighted by the presentation of the first IRE Founders Award to Gen. Sarnoff by President James W. McRae. The recently established Founders Award is a non-annual award for outstanding leadership in the radio engineering profession.

The annual IRE awards for 1953 were presented by President McRae to the recipients, with John M. Miller, Naval Research Laboratory, receiving the Medal of Honor, the Institute's highest annual technical award; and Frank G. Kear responding on behalf of the 49 newly elected Fellows whose photographs appear on pages 674 through 681 of this issue.

The Morris Liebmann Memorial Prize was awarded to John A. Pierce, Harvard University, and the Zworykin Television Prize Award was presented to Frank Gray, Bell Telephone Laboratories. The Harry Diamond Memorial Award went to Robert M. Page, Naval Research Laboratory. Awards were presented for outstanding PROCEEDINGS papers, with Richard C. Booton, MIT, receiving the Browder J. Thompson Memorial Award, and E. O. Johnson and W. M. Webster, RCA Laboratories, receiving the Editor's Award.

The social program was rounded out with an entertaining schedule of fashion shows, tours, and matinees for the wives of members attending the Convention.

President James W. McRae addresses IRE members during Annual Meeting.

W. R. G. Baker presents the annual report of the Treasurer at the Annual Meeting.

Frank G. Kear responds on behalf of the newly elected Fellows at the Banquet.

George Sziklai, RCA Laboratories, demonstrates a transistor audio amplifier.

HIGHLIGHTS OF IRE INDUSTRY-WIDE TRANSISTOR EXHIBIT AND PRESS CONFERENCE

Top—Conventional equipment compared with smaller transistorized versions. From left to right: A wire repeater, Geiger counter, and frequency meter.
Bottom—Transistor equipment including a radio receiver, test oscillator, "phone-plug" amplifier, and several hearing aids.

Left—R. L. Wallace, Jr., Bell Telephone Laboratories, displays a transistor amplifier in his right hand which plugs into the coaxial cable in his left.

Right—James Sweeny, General Electric Company, operates a transistor transmitter in a pot of boiling water to demonstrate hermetically sealed transistor.

Above—E. W. Herold of RCA Laboratories discussses transistor developments at the first industry-wide exhibit held at the opening of the 1953 IRE Convention.

Right—Bell Telephone Laboratories' exhibit of transistors and their application to telephone systems, including the original point-contact transistor and several cut-away models.

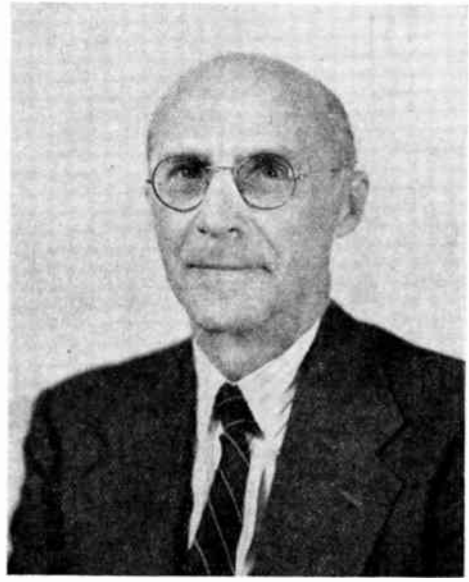
IRE Awards, 1953

Founders Award

David Sarnoff

"For outstanding contributions to the radio engineering profession through wise and courageous leadership in the planning and administration of technical developments which have greatly increased the impact of electronics on the public welfare."

Medal of Honor—1953



John M. Miller

"In recognition of his pioneering contributions to our basic knowledge of electron tube theory, of radio instruments and measurements, and of crystal controlled oscillators."

Morris Liebmann Memorial Prize—1953

John A. Pierce

"For his pioneering and sustained outstanding contributions to radio navigation, and his related fundamental studies of radio wave propagation."

Vladimir K. Zworykin Television Prize Award—1953

Frank Gray

"In recognition of the fundamental importance to color television of his early studies of the television signal spectrum."

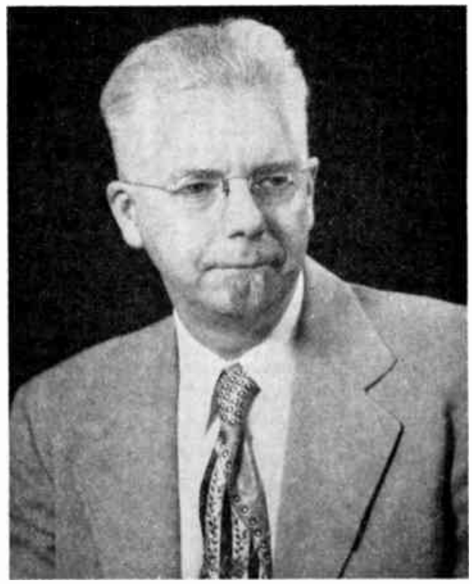
IRE Awards, 1953

Browder J. Thompson Memorial Award—1953

RICHARD C. BOOTON, JR.

"For his paper entitled, 'An Optimization Theory for Time-Varying Linear Systems with Nonstationary Statistical Inputs,' which appeared in the August, 1952 issue of the PROCEEDINGS OF THE I.R.E."

Harry Diamond Memorial Award—1953



ROBERT M. PAGE

"For outstanding contributions to the development of radar through pioneering work and through sustained efforts over the years."

Editor's Award—1953

E. O. JOHNSON

WILLIAM M. WEBSTER

"For their paper entitled, 'The Plasmatron, A Continuously Controllable Gas-Discharge Developmental Tube,' which appeared in the June, 1952 issue of the PROCEEDINGS OF THE I.R.E."

New Fellows

EDWARD W. ALLEN, JR.

"For his technical and administrative contributions in the fields of radio wave propagation and radio noise."

JEAN P. ARNAUD

"For his outstanding work in telecommunication development, and educational activity in South America."

BENJAMIN B. BAUER

"For his important contributions to the development of microphone and other audio devices."

JOHN W. BELL

"For outstanding contributions to the design and development of military radar and cathode-ray direction finding equipment in Canada."

LEONARD J. BLACK

"For research in the field of electromagnetic radiation, and a distinguished record in teaching radio engineering."

HENRY G. BOOKER

"For his theoretical research in electromagnetism and radio wave propagation."

WILLIAM E. BRADLEY

"In recognition of his technical contributions and original analytical methods in the fields of television and frequency modulation detection."

JOHN L. CALLAHAN

"For his contributions to international radio communications and especially to radiophoto and multiplex techniques."

KENNETH A. CHITTICK

"For his contributions to radio and television engineering, and for his meritorious work on many important Institute and industry technical committees."

New Fellows

Arthur A. Collins

"For his contributions to the art of transmitter design and precision remote tuning systems, and for his far-sighted engineering direction."

Edward U. Condon

"In recognition of his outstanding contributions and distinguished leadership in the development of microwave radar and in the atomic energy program."

William W. Eitel

"For his pioneering contributions to power tube design."

Harry Faulkner

"For his engineering achievements in the field of world-wide radio communication, and his contributions to international agreement on tele-communication practices."

Enoch B. Ferrell

"For his original contributions and research leadership in high-frequency power amplifiers, servo-mechanisms, telephone switching, and quality control."

Warren R. Ferris

"For early contributions in the high-frequency operation of vacuum tubes."

Lyman R. Fink

"For his contributions to the development of armament radar."

Lawrence R. Hafstad

"In recognition of his basic research work in electronics and nuclear science, and his administration of important scientific programs."

Ferdinand Hamburger, Jr.

"For his leadership as a teacher and author in the electronics and electrical engineering fields."

New Fellows

Lewis B. Headrick

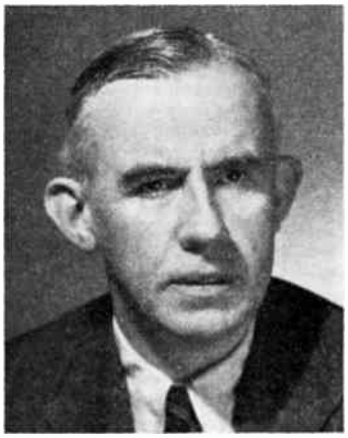
"For contributions and leadership in the development of cathode-ray tubes and television camera tubes."

Philip J. Herbst

"For his contributions to numerous branches of radio engineering, particularly television and air navigation."

John Hessel

"For his contributions to military radio communications, particularly in the fields of man-carried combat radios and radio relay equipment."



Hans E. Hollmann

"For his fundamental analyses and inventions in the fields of electron tubes and high frequency circuits."

Theodore A. Hunter

"For his contributions to the design of stable tunable oscillators, and for his able promotion of the activities of the Institute."

Eric J. Isbister

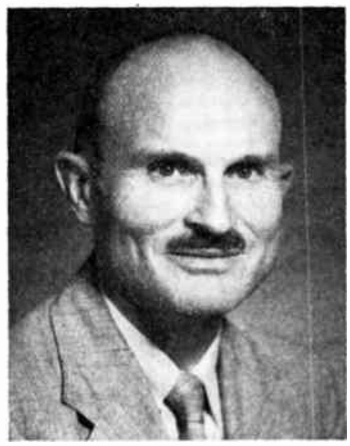
"For his contributions in the development of equipment for radio navigation, aircraft radar, and marine radar."

Edward C. Jordan

"For his work in radio direction-finding and antenna developments."

Frank G. Kear

"In recognition of his contributions in the fields of radio aids to air navigation and radio and television broadcasting."



Ronold W. P. King

"For his many contributions to the theory of radiating systems."

New Fellows

Royce G. Kloeffler

"In recognition of his achievements as an educator and author in the fields of electrical and electronic engineering."

Edmund A. Laport

"For his contributions to transmitter development, international communication systems, and to antenna engineering."

Russell R. Law

"For his contributions to vacuum tube and television development."

William A. MacDonald

"For his engineering contributions and his sustained leadership in electronic development."

Jack A. McCullough

"For his pioneering contributions to power tube design."

Jack A. Morton

"For his leadership and contributions to the engineering physics and development of transistors and of wide-band microwave electron tubes."

Allan B. Oxley

"For his contributions and sustained leadership in radio engineering and manufacturing in Canada."

Albert Preisman

"For contributions to the application of graphical constructions, video amplifier theory and technical education."

J. C. R. Punchard

"For his contributions to the design and development of communication and radar equipment, particularly for the Canadian Armed Services."

New Fellows

JAN A. RAJCHMAN

"For outstanding contributions in secondary emission multiplication and digital storage devices."

JOHN A. RATCLIFFE

"For his outstanding contributions to radio propagation research and to education."

STEPHEN O. RICE

"For his mathematical investigations of communication problems of noise, nonlinear circuits and efficient coding of information."

WALTHER RICHTER

"In recognition of his contributions in the field of industrial electronics."

ALFRED A. ROETKEN

"For his contributions to the development of the transcontinental microwave radio relay system and the development of single-sideband radio telephone receivers."

WILLIAM M. RUST, JR.

"In recognition of his outstanding contributions to geophysics and communications engineering pertaining to the petroleum industry."

JORGEN RYBNER

"For outstanding contributions to research and teaching in the field of telecommunications."

DANIEL SILVERMAN

"In recognition of his application of electronics to geophysical exploration."

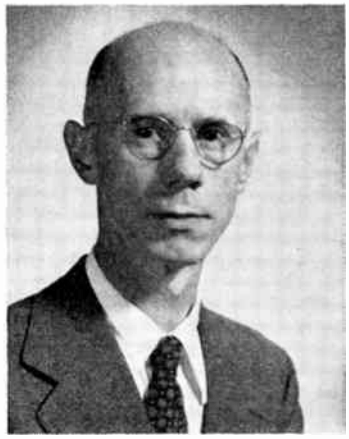
ARCHIE W. STRAITON

"For his many technical contributions in the field of radio propagation."

New Fellows

Irven Travis

"For his contributions to the fields of computing devices and of gunfire-control equipment and to engineering education."



Bertram A. Trevor

"For contributions in the field of pulse multiplex communication systems and long-range receiving equipment."

Charles J. Young

"In recognition of his many contributions in the fields of facsimile and electronic timing equipment."

Harry W. Wells

"In recognition of his contributions to ionospheric research and the organization of a world-wide network of ionospheric stations."

Institute News and Radio Notes

Calendar of COMING EVENTS

1953 National Conference on Airborne Electronics, Dayton, Ohio, May 11–14

1953 Electronics Parts Show, Conrad Hilton Hotel, Chicago, Ill., May 18–21

National Telemetering Conference, Edgewater Beach Hotel, Chicago, Ill., May 20–22

AIEE Conference on Electric Heating, Detroit-Leland Hotel, Detroit, Mich., May 26–27

High Frequency Communication Symposium, IRE Professional Group on Communications Systems, Long Lines Auditorium, 32 Avenue of the Americas, New York, N. Y., June 11–12

Symposium on Microwave Optics, McGill University, Montreal, Canada, June 22–24

American Society for Engineering Education Convention, University of Florida, Gainesville, Fla., June 22–26

IRE Western Convention and Electronic Show, Civic Auditorium, San Francisco, Calif., August 19–21

National Electronics Conference, Hotel Sherman, Chicago, Ill., September 28–30

1953 IRE-RTMA Radio Fall Meeting, Toronto, Ont., October 26–28

1954 Sixth Southwestern IRE Conference and Electronics Show, Tulsa, Okla., February 4–6

Technical Committee Notes

The **Standards** Committee met on February 19 under the Chairmanship of A. G. Jensen. The list of "industrial electronics terms" was discussed briefly and the recommendation made that the material be sent on Grand Tour. The Committee reviewed the proposed modifications to the Receivers Committee's scope. Jack Avins was asked to prepare a revised scope incorporating the proposed changes. Considerable time was devoted to an informal discussion on improving the over-all procedure of the Standards Committee and an Ad Hoc Committee was set up to study all recommendations made. R. F. Shea, J. G. Brainerd, Ernst Weber and M. W. Baldwin, Jr., were appointed to the committee; Professor Weber serving as Chairman. Initial consideration was given to the Proposed Standards on Antennas and Wave Guides: Definitions of Terms, as presented by D. C. Ports and A. G. Fox. The committee will continue its review of the terms at the April meeting.

Under the Chairmansip of D. C. Ports, the **Antennas and Wave Guides** Committee met on February 11. John Ruze has accepted an invitation to be a permanent member of this committee and P. H. Smith has been appointed Vice Chairman for 1953. The Chairman announced that the latest list of Definitions of Waveguide Terms (51 IRE 2. PS1), including those revisions made at the meeting of the Antennas and Wave Guides Committee on January 7, 1953, has been submitted to members of the Standards Committee for approval at their next meeting. A number of wave guide component terms were approved.

On January 22 the **Circuits** Committee convened under the Chairmanship of C. H. Page. There was considerable discussion on the terms *unilateral network* and *bilateral network*. It was decided to delete these two terms from the list (52 IRE 4. PS1) awaiting action by the Standards Committee. It was also decided these terms could best be defined in terms of various types of portpairs.

On January 20, the **Feedback Control Systems** Committee met under the Chairmanship of W. M. Pease. Chairman Pease reported that the Standards Committee on January 8 has approved the change in name for this committee from the Technical Committee on Servo Systems to the Technical Committee on Feedback Control Systems. A draft of the committee report, 53 IRE 26. R1, Recommended Symbols for Feedback Control Systems, Part I, was discussed and a final version adopted. The report will be transmitted to the Symbols Committee for comment before further distribution.

The **Video Techniques** Committee met on February 6 under the Chairmanship of W. J. Poch. A. J. Baracket, unable to attend this meeting, telephoned in his report on the Subcommittee on Video Systems and Components. He reported that two meetings have been held since the last meeting of the Video Techniques Committee. Most of the time was spent in reviewing the proposed standard on the Methods of Measurement of Geometric Distortion. It is planned to submit the corrected report to the Subcommittee within several weeks and if it is approved, it may reach the Video Techniques Committee by next month. The Subcommittee has also agreed to consider the problem of establishing a "Method of Measurement of Overshoot and Percent Tilt of a Rectangular Wave Form." S. K. Athey indicated that work on definitions in the field of video recording would be initiated in the near future. K. B. Benson suggested that consideration be given to a "Method of Measuring Interlace" which includes the photographic image on the film. Considerable time was spent in reviewing the proposed definitions sent to the committee members prior to the meeting. A number of revisions were made and nine of these definitions were tentatively approved subject to an over-all review when the remaining terms under immediate consideration have been studied by the committee.

The **Information Theory & Modulation Systems** Committee met on February 17 under the Chairmanship of W. G. Tuller. The contents of a Bureau of Ships letter *re* the desirability of 4–6 weeks advance notice to Navy personnel when meetings are planned was noted. J. G. Kreer, Jr., reported on the status of interchanging information with foreign groups. The committee reviewed in detail the Information Theory definitions, recommending a number of revisions. Discussion ensued on the term "Bandwidth," which is to be referred to the Basic Terms Committee.

The **Electron Devices** Committee convened on February 13 under the Chairmanship of G. D. O'Neill. Subcommittee 7.2 on Storage Tubes and the analogous AIEE subcommittee have formulated a number of mutually acceptable definitions. It is felt that the wide interest already evidenced in these terms makes it advisable to circulate the preliminary draft among users of the Williams tube. The proposed Phototube definitions have been revised by R. S.Burnap and R. B. Janes. Discussion ensued on the definition for "phototube." Dr. Janes is to give further consideration to the matter before final presentation of the list.

On February 16 the **Navigation Aids** Committee met under the chairmanship of P. C. Sandretto. Discussion ensued on the terms defined at the last meeting. Further modifications were suggested. These terms were then approved by the committee. The committee then proceeded with consideration of the remaining terms on Harry Davis' list (submitted by McLean).

A. Hoyt Taylor addressing inaugural meeting of Orange Belt Subsection.

Inaugural Meeting of Orange Belt Subsection

Over two hundred attended the inaugural meeting of the Orange Belt Subsection of the Los Angeles Section held in February at the National Bureau of Standards, Corona Laboratories, Corona, Calif.

Feature speaker of the meeting was Dr. A. Hoyt Taylor, who reminisced on the pioneer days of electronics in a speech entitled "From Cats Whisker to Transistor—My Fifty Years in Electronics."

Other speakers included R. D. Huntoon, director of Corona Laboratories, C. D. Perrine, assistant manager of the Guided Missile Division of Consolidated-Vultee Aircraft Corporation, and W. G. Hodson, Chairman of the Los Angeles Section of the IRE.

Professional Group News

Airborne Electronics

The Dayton (Ohio) Chapter of the Professional Group on Airborne Electronics has been approved officially by the IRE Executive Committee. Maurice Jacobs has been appointed chairman. He will also serve as chairman of the executive committee, and the following officers will also serve: Paul Wiegert, vice-chairman; George Glan, secretary-treasurer; Robert Siff, chairman of arrangements committee; Harold King, chairman of papers committee; Merrill Barkley, chairman of publicity and membership committees.

At the first meeting of the Chapter, at the Biltmore Hotel, D. W. Webber, Collins Radio Company, Cedar Rapids, Ia., presented a paper on "Modular Design of Airborne Communications Equipment." Additional meetings have been planned.

The Philadelphia Chapter held two meetings recently at the Franklin Institute. C. E. Dolberg was Chairman of the earlier meeting, at which H. A. Affel, Jr., of the Philco Corporation, gave a paper on "Transistors for Military Applications." He discussed mainly the ability of the transistor to meet environmental conditions for military equipment. Besides presenting charts giving the approximate value of the various parameters of a transistor, he also explained briefly the functioning of the contact and junction type transistors. A group discussion followed his talk. At another meeting K. C. Black served as the Chairman and R. R. Freas, of the Radio Corporation of America, spoke about "Electronic Timers for Loran Receivers." Mr. Freas explained the Loran system in general terms before discussing in detail the methods of obtaining accurate timing in the Loran receiver. His presentation was followed by a group discussion.

Antennas and Propagations

The Professional Group on Antennas and Propagation has achieved a growth where it can now undertake a quarterly publication. Starting as soon as possible, this new organ, *Transactions*, will serve to promote as well as co-ordinate and integrate all professional activities of the group.

J. B. Smyth, U. S. Naval Electronics Laboratory, San Diego, Calif, has agreed to take the editorship, and H. A. Finke, Director of Engineering of Polytechnic Research and Development Co., Inc., 55 Johnson Street, Brooklyn 1, N. Y., has accepted the position of news editor. *Transactions* will include not only group news and views, but also papers covering current scientific investigations in the field of antennas and propagation, summaries of relevant meetings, conferences, and symposia, and invited papers dealing with pertinent current knowledge. Communications, primarily from members of the group, as well as reviews of books, papers, and so on, are planned. *Transactions* in content, format, and appearance is to be a thoroughly professional publication.

It is anticipated that *Transactions* will become an accepted vehicle for final publication of papers in the field on interest of the Professional Group on Antennas and Propagation. The organization of committees to carry out the above program is now under way, and announcements regarding procedures will appear in future issues of *Transactions*. In the interim, papers for publication in the *Transactions* are now being invited and should be submitted to the editor. The co-operation and support of all IRE members whose professional interest lies in this field is desired for the success of this publication.

The Professional Group on Communication Systems extends a special invitation to members of the Group to attend their High Frequency Symposium to be held June 11 and 12 at the Long Lines Auditorium, 32 Avenue of the Americas, New York, N. Y. For information about the program see "Communication Systems" on this page.

The Chicago Chapter met recently at the Western Society of Engineers Auditorium, Vice-Chairman George Kearse presiding. E. F. Harris, president of Mark Products, Inc., Chicago, Ill., gave "A Study of the Corner Reflector." He presented a number of patterns of corner reflectors, using corner angle and distance out to dipole as variables showing the effects of each over wide ranges. An election of officers was held. New officers of the Chapter are as follows: Chairman, G. Kearse; Vice-Chairman, L. R. Krahe; Secretary-Treasurer, R. E. Jensen.

Audio

The Chicago chapter of the Professional Group on Audio met recently at the Western Society of Engineers Auditorium with R. E. Troxel as Chairman. D. W. Martin, research supervisor of the Baldwin Company, Cincinnati, Ohio, presented a paper entitled the "Enhancement of Music by Reverberation."

The Milwaukee Chapter met recently at the Engineers Society of Milwaukee Building under the chairmanship of D. E. Mereon. A large number of members heard Gene Fiebich, engineer with the Heath Company, Benton Harbor, Mich., speak on "Heath Kits, Instruments and Measurements." Mr. Fiebich, assisted by Mr. Turner, also of the Heath Company, demonstrated the latest complete line of assembled Heath Kit instruments. Gain, response, intermodulation and distortion were measured on audio amplifiers.

The Philadelphia Chapter met at the Edison Building Auditorium recently with H. K. Neuber as Chairman. About 350 members heard Emery Cook, president of Cook Industries, discuss the problems encountered in the recording and reproduction of music from discs. He placed special emphasis on the significance of binaural recordings. In this connection, he covered the technical problems of binaural recordings on discs from both the recording and play-back viewpoint. The aspect of microphone spacing as related to loudspeaker spacing was also covered. Mr. Cook has received wide recognition for his work in the binaural field.

Broadcast Transmission Systems

The Boston Chapter of the Professional Group on Broadcast Transmission Systems met recently in the Auditorium Studio of WBZ, under the chairmanship of P. K. Baldwin. Speaker of the evening was R. D. Chipp, Director of Engineering, DuMont Television Network, who spoke on "Technical Records for the TV Broadcaster." He explained the purpose of technical records, tube records, maintenance records, maintenance tests, and studio and field equipment tests which his organization has found necessary in the accomplishment of smooth and straightforward TV operations.

The Chicago Chapter of the Group held a meeting recently at the Western Society of Engineers Auditorium with Stephen Bushman as Chairman. R. M. Cohen, of RCA Laboratories, spoke on "The Application of RCA Point-Contact Transistors." He described the characteristics of the problems encountered in transistor circuit design, and showed working circuits for audio, rf and switching circuits using point-contact transistors. The audience was supplied with folders outling characteristics and circuits.

Communication Systems

The program of the high frequency symposium sponsored by the Professional Group on Communications Systems on June 11 and 12 follows. The symposium will be held in the Long Lines Auditorium, 32 Avenue of the Americas, New York, N. Y.

Seven papers will be presented on June 11, followed by an inspection of the American Telephone and Telegraph overseas telephone terminal at the place of meeting. On June 12 a trip to the overseas radio telephone transmitter and receiver plants in New Jersey will be made.

The first four papers will deal with the topic of "Solar Research and its Application to High Frequency Communications Systems." W. A. Miller, Radio Corporation of America, RCA Laboratories Division, will establish the correlation between solar activity and high frequency communication. A. B. Moulton, also of RCA Laboratories, and J. H. Rush, High Altitude Observatory of Harvard University and the University of Colorado, Boulder, Colo., will cover the theoretical aspects and touch on research work under way. J. H. Nelson, Operations Department, RCA Communications, Inc., will describe a practical application to high frequency communication.

C. G. Dietsch, Engineering Department, RCA Communications, Inc., in his paper, "The Tangier Relay System," will describe a high frequency radio relay station organized to handle telephone, telegraph, radiophoto and international programs.

In the "Development of the LD-T2 Transmitter," N. F. Schlaack, Bell Telephone Laboratories, Inc., will describe the most modern version of the single sideband radio transmitter used for overseas telephone service.

K. P. Stiles, Long Lines Department, AT&T Company, will present an over-all description of the facilities which will be inspected at the overseas telephone terminal and the New Jersey transmitter and receiver plant.

Detailed registration information will be mailed by May 1. Special invitation is extended to members of the Professional Group on Antennas and Propagation.

Professional Group News *(Cont.)*

1951 Computer Proceedings Again Available

Due to the heavy demand for the Proceedings of the 1951 Joint Computer Conference held in Philadelphia December 10–12, 1951, 500 copies will be reprinted and made available at $3.50 per copy from IRE Headquarters. The Proceedings were entitled "Review of Electronic Digital Computers."

Electronic Computers

The Los Angeles Chapter of the Professional Group on Electronic Computers met recently at the UCLA Institute for Numerical Analysis. W. F. Gunning was Chairman. A. R. Piatt, design engineer for Librascope, Inc., spoke on "Mechanical Analog Computer Techniques." He described many of the basic building blocks used in rotating and linkage type mechanical analog computers.

The Philadelphia Chapter has been meeting at the Franklin Institute regularly under the Chairmanship of I. L. Auerbach. Recent papers presented to the Chapter have included "Static Magnetic Memory Systems," by W. D. Woo, "Automation—Today's Industrial Revolution," by J. T. Diebold of Harbridge House, and "Comparison and Evaluation of Digital Computer Memory Systems." Mr. Diebold covered the business problems associated with automation as distinct from the technological aspects, and the social and economic consequences of automation. In discussing the redesign of a product and its process he anticipates not only the simple addition of automatic controls to our present machinery in both factory and office, but a whole new family of machines completely divorced from the limitations of the human operator. The latter paper reviewed and compared various types of memory systems available for use with digital computers. Many of the types were illustrated, and some of the underlying philosophy behind memory systems given.

UCLA Computer Proceedings Sold Out

The Professional Group on Electronic Computers announces that the Proceedings of the UCLA Computer Conference held in May 1952 have all been sold and are no longer available.

Engineering Management

The Chicago Chapter of the Professional Group on Engineering Management met recently at the Western Society of Engineers Auditorium with F. W. Schor as Chairman. Gordon Sargent, staff member of the Dale Carnegie Institute, spoke on "Human Ventures." Aided by ten members of the Chapter, he discussed the Dale Carnegie principles of human relations as applied to management.

Information Theory

The Albuquerque, New Mexico, Chapter of the Professional Group on Information Theory met recently at Mitchell Hall, University of New Mexico, Y. M. Hill was the Chairman. Lt. Col. L. V. Skinner gave a paper titled, "An Introduction to Correlation Analysis."

Vehicular Communications

The Chicago Chapter of the Professional Group on Vehicular Communications met at the Western Society of Engineers Auditorium recently with R. V. Dondanville as Chairman. A. A. MacDonald, assistant chief engineer of Motorola Two-Way Communication Equipment Department, spoke about "450 Megacycle Two-Way Communication Equipment." A question and answer period followed the presentation.

Symposium on Microwave Optics

Members of the IRE are invited to participate in a Symposium on Microwave Optics, to be held at McGill University, Montreal, Canada, June 22–24.

The symposium is sponsored jointly by Commission 6 of the Canadian National Committee and Commission 6A of the United States National Committee of the International Scientific Radio Union. It will mark opening of the Eaton Electronics Research Laboratory at McGill University.

Radio Meteorology Conference

A conference on radio meteorology will be held at the University of Texas, in Austin, on November 9–12. Papers are invited on cloud physics and precipitation mechanisms, radar rainfall determination, operational use of weather radar, tropospheric propagation and attenuation of radio signals, refractive index meteorology and climatology, thunderstorm and tornado atmospherics, fading characteristics of atmospheric and surface reflected radio signals, atmospheric turbulence and scattering, radar analysis and meteorological structure of tropical and subtropical disturbances. Abstracts of 100 words should be sent to J. R. Gerhardt, Box F, University Station, Austin, Tex., by July 1. Two copies of a 1500–2000 word summary, one suitable for reproduction should be submitted by Sept. 1.

IRE Chairmen Named for WESCON

The IRE project-committee chairmen have been announced for the 1953 Western Electronic Show and Convention, August 19–21, Civic Auditorium, San Francisco, Calif., jointly sponsored by the IRE Seventh Region and West Coast Electronic Manufacturers' Association.

B. M. Oliver, Hewlett Packard Company, will be chairman of the papers committee and W. S. Pritchett, University of California, will serve as chairman of the arrangements committee. Both chairmen will work with Walter Noller, Remler Company Limited, Wescon vice president and IRE representative.

National Telemetering Conference

The first National Telemetering Conference is to take place in Chicago, Ill., May 20–22, 1953, at the Edgewater Beach Hotel. It will be sponsored by the Institute of Aeronautical Sciences, the Institute of Radio Engineers, the American Institute of Electrical Engineers, and the Instrument Society of America.

This first Conference is significant in that it represents the first exchange of instrumentation information between commercial organizations and research aircraft and missile test groups.

The keynote speech will be delivered at the opening night banquet by J. E. Hobson of the Stanford Research Institute. A comprehensive program of papers has been arranged, including representative instrumentation reports from other countries.

Western Computer Conference

A bright future for electronic computers in a large variety of uses in science and industry was painted at the Western Computer Conference held in Los Angeles, Calif. Speakers included Simon Ramo, vice president for operations, Hughes Aircraft Co.; R. C. Huntoon, chief, Cornoa guided missile laboratories, National Bureau of Standards; L. A. Dubridge, president, California Institute of Technology; and J. E. Hobson, director, Stanford Research Institute.

Panel at the Western Computer Conference evaluating analog and digital computers. Left to right are J. L. Barnes, North American Aviation; Floyd Steele, Digital Control Systems, Inc.; G. D. McCann, moderator and Conference Committee Chairman, California Institute of Technology; Louis Redenour, International Telemeter Corp.; and A. W. Vance, RCA Laboratories, Princeton, N.J.

IRE People

Emil E. Mayer (F'34), consulting engineer, and chairman of the board of directors of Magnetic Amplifiers, Inc., of New York, N. Y., died recently.

EMIL E. MAYER

Dr. Mayer was born in Elberfeld, Germany on January 7, 1885. He received the E.E. degree from the Technical University of Berlin in 1906, and the Dr. of Eng. degree from the University of Munich in 1919.

Starting as a test engineer with Allgemeine Elektrizitaets Gesellschaft of Berlin in 1906, he rose to the position of managing engineer, in charge of design and manufacture. In 1911 he became a commercial engineer with Siemens-Schuckert of Berlin, and in 1913 a radio engineer with Hochfrequenzmaschinen Aktiengesellschaft of Berlin, in charge of development of transatlantic radio telegraphy by means of high frequency alternators. Joining the Telefunken Company of Berlin in 1919, as chief engineer he was in charge of all technical and research activities. He served as vice president for two years, then became the American representative of the German General Electric Company in 1924. He handled contract relations and the exchange of technical information with that company and the Int. General Electric Co. In 1928 he returned to Germany to become vice president in charge of engineering. In 1931 he became president of the Telefunken Company.

Dr. Mayer came to this country in 1933, and has been a consulting engineer and director of many companies since that time. These include the Teca Corporation, Locke Steel Chain Corporation, American Measuring Instruments Corporation, General Ceramics and Steatite Corporation (from 1943 to 1953), and the Wilcox-Gay Corporation of Charlotte, Mich. (1937 to 1950), of which he became president in 1949. He also served as president and consultant for the Selectar Manufacturing Corporation during World War II.

Dr. Mayer negotiated numerous patent license agreements in the electronic field for European and American concerns, besides holding numerous patents of his own.

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Maxwell Ratner (A'52) has been appointed sales manager of the Federated Semi-Conductor Company of New York, N. Y.

Mr. Ratner was born in New York, N. Y. on January 21, 1915. He attended the College of the City of New York, and received the M.S. degree from Fordham University in 1947.

Formerly with the Research and Development Board of the Department of Defense, Mr. Ratner was a member of the Secretariat of its Panel on Electron Tubes at New York University. He was concerned with co-ordinating government programs in transistors and semi-conductors.

Prior to this time Mr. Ratner did research in semi-conductor materials towards a Ph.D. degree at the Polytechnic Institute, and microwave development with the Polytechnic Research and Development Co., Inc., both of Brooklyn, N. Y. During World War II he served five years in the U. S. Army Air Force as a radar officer, studying at Harvard and M.I.T. part of this time.

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Karl B. Hoffman (A'34–SM'52), formerly technical director, has been elevated to the office of vice president of the WGR Broadcasting Corporation. He will be in charge of TV planning and operation.

KARL B. HOFFMAN

After graduating from the Bliss Electrical Engineering School of Washington, D. C., in 1921, Mr. Hoffman became associated with the Esca Electrical Supply Co. of Albany, N. Y., where he was in charge of organizing and managing an amateur radio supply department. He worked for the Shotton Radio Manufacturing Co. and the Raven Radio Co. before joining the engineering staff of WGY of the General Electric Company in 1923.

In 1927 Mr. Hoffman joined the group directed by Dr. Alexanderson to design and build television transmitters and receivers. As maintenance engineer, he was in charge of installing the first remote television pickup, which telecast Governor Alfred E. Smith's acceptance of the Democratic party's nomination for president in 1928, and of regular broadcasting of picture signals by WGY in 1929.

Mr. Hoffman became associated with the Buffalo Broadcasting Corporation (predecessor of the WGR Broadcasting Corporation) as technical director in 1933. He has designed and supervised construction of various transmitting facilities of the Corporation, and also served three years as manager of the program department.

Mr. Hoffman is a past chairman of the Buffalo Chapter of the I.R.E., and past director of the Engineering Society of Buffalo. Since 1952 he has been director of communications for Erie County Civil Defense. He is a member of the American Radio Relay League.

Cledo Brunetti (A'37–SM'46–F'49) formerly associate director of the Stanford Research Institute, Stanford, Calif., has accepted an executive research post with the Mechanical Division of General Mills. He will be concerned with setting up a new general research laboratory and industrial development of the division.

Dr. Brunetti was born on April 1, 1910 in Virginia, Minn. He received the B.S. degree in electrical engineering from the University of Minnesota in 1932 and the first Ph.D. in electrical engineering granted by the University, in 1937. That same year he was awarded the National Scholarship Key by the National I. A. Civic League.

In 1937 he joined the faculty of Lehigh University as assistant professor of electrical engineering, and also directed research in radio, electronics, and electrical engineering. As a consulting engineer during this period, he worked on the New York World's Fair in 1938.

He became associated with the National Bureau of Standards radio section in 1941, where he worked on the development of the radio proximity fuze. He became chief of the production engineering section of the Ordnance Development Division in 1941, then of the engineering electronics section.

Since 1949 Dr. Brunetti has been associate director of the Stanford Research Institute in Stanford, Calif.

Dr. Brunetti has served on various IRE committees, and has written numerous technical papers. He has received many awards from the government in recognition of his services, and in 1941 was named by Eta Kappa Nu as "America's Outstanding Young Electrical Engineer."

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Marion E. Bond (A'29–SM'44), chief engineer of the two-way radio development section of the Communications and Electronics Division of Motorola, Inc., died recently at his home in Elmhurst, Ill., after an extended illness.

One of the pioneers in the development of two-way radio for police, fire, and similar applications, Mr. Bond was born May 30, 1903 in Chittenango, N. Y. He received the B.S. degree in electrical engineering from Ohio State University in 1927.

After graduation Mr. Bond joined the Radio Division of the American Bosch Magneto Corp. in Springfield, Mass. As head of the components section he was in charge of design and development of loudspeakers, magnetic pickups, phonograph and audio frequency amplifiers. In 1938 he joined the Colonial Radio Corp. of Buffalo, N. Y., designing auto-radio equipment. In 1938 he became associated with the Galvin Mfg. Corp. of Chicago, where he was concerned with the design and development of communication equipment for the armed forces.

Books

New Books

Proceedings of the National Electronics Conference Volume 8. Published (1952) by National Electronics Conference, 852 E. 83rd St., Chicago 19, Ill. 835 pages, charts, diagrams, tables. $9\frac{1}{4}\times 6\frac{1}{4}$. Case bound. $5.00.

This book contains all of the technical papers presented at the 1952 conference. The ninety-seven papers cover electronic research, development, and application in antennas, the assembly and measurement of components, audio, circuits, coding and recording techniques, computers, delay lines and hf-test equipment, electronic instruments, engineering managment, industrial measurements, magnetic amplifiers, memory tubes, radar, radio navigation, reliability of components and equipment, semiconductors, servomechanism, television, transistors, and waveguides.

Volumes 2, 4, 5, 6, and 7 may also be obtained for $5.00 per copy.

Radio Interference Suppression by G. L. Stephens

Published (1952) for *Wireless World* by Iliffe and Sons Ltd., Dorset House, Stamford St., London, S.2.1, England. 128 pages+2-page index+2-page bibliography. 65 figures. $8\frac{1}{2}\times 5\frac{1}{2}$. 10s. 6d.

The book describes the principles and practical circuitry involved in the suppression of radio-frequency noise. It is written for the practical engineer confronted with the problem of reducing interference noise in radio-telephone or television reception. This interference may be produced by electrical generators, motors, household appliances, medical apparatus, and lighting signs.

The text describes the type of electrical circuits which will be the most effective for suppression of each type of interference and the limitations of component elements used in these circuits. A calculation of the reactance of a few typical values of "L" and "C" at radio frequencies might add weight to the text and be in keeping with the quantitative estimates given of line and load impedances.

The subject matter is well covered and reasonably complete. The book reads smoothly and is couched in nonmathematical language. Also included are the essential features of the earlier edition written by G. W. Ingram in 1939; a small amount of new material on filter component elements and interference in television receivers is added.

C. T. Grant
Bell Telephone Labs., Inc.
Murray Hill, N. J.

TV Factbook No. 16

Published (1953) by Television Digest, Wyatt Bldg., Washington, D. C. 268 pages. $3.00.

A directory of all authorized post-freeze new TV stations showing which ones are televising and when the others expect to begin highlights the new edition of the TV Factbook. There were 175 such grants as of January 3, 1953, of which 48 were vhf and 127 were uhf.

A new feature is an analysis of the 162 most important markets of the United States, estimating households and number of TV sets in each market at the end of 1952. Other new features are directories of community antenna systems, theatre-TV installations, and FCC channel allocation tables with priority lists. Some standard features brought up to date are directories of TV equipment manufacturers (including receivers, transmitters, tubes, theatre-TV, community antennas), program sources, FCC personnel, attorneys and engineers, station representatives, trade associations, unions, and publications.

The Factbook also tabulates all 748 applications for new TV stations pending before the FCC as of January 3 (465 vhf, 283 uhf), together with names, addresses, facilities, and equipment. It contains personnel and facilities data, with digests of rate cards, of the 125 United States and the two Canadian and one Mexican border stations now on the air, plus 19 others due to begin. Rate cards of the networks, and data on actual and projected stations in Mexico, Cuba, and South America are also included.

A 34×22-inch wall map in color goes with each Factbook, showing present TV cities, all cities with more than 10,000 population, and routes of all actual and projected coaxial-microwave network interconnections.

Fundamentals of Engineering Electronics by William G. Dow

Published (1952) by John Wiley & Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 538 pages+20-page index+24-page bibliography+xviii pages. 242 figures. $5\frac{1}{2}\times 9\frac{1}{4}$. $8.50.

William G. Dow is professor of electrical engineering, University of Michigan, Ann Arbor, Mich.

There are few texts which can compare with Professor Dow's "Fundamentals of Engineering Electronics" in quality and quantity of figures, as this second edition reminds us. The printing is improved by limited changes in the contents, and by use of the mks system of units, thus bringing one of the few important nonconformists into line. Furthermore, "Engineering Electronics" has lived to see the meaning of words change so that the title with which it was baptized is no longer appropriate. It is yet sufficiently young to announce the imminent arrival of "Fundamentals of Physical Electronics" and "Microwave Electron Tubes." Their presence gives Professor Dow the answer to any criticism of this volume's content; what is not found here is probably elsewhere in the trilogy.

As in the first edition, the text contains an introduction to physical electronics, discussing potential distributions, electron ballistics, space charge, thermionic cathodes, vacuum tubes, microwave tubes, cathode-ray devices, kinetic theory, gaseous conduction, and so forth. The reviewer, never having read the first edition, gives his impressions considering that much of the material may have appeared previously and stood the test of time.

The care and diligence devoted to the figures carries over to the text. The reader appreciates the author's efforts to make the presentation simple. However, it appears that these efforts have been carried considerably beyond the optimum point, resulting in a failure to recognize the facts of life, unusual approaches which might be personal eccentricity, and a general isolation of the material from collateral work which a student may meet.

For example, the force equation, usually considered important in physical electronics, is not named and comes in indirectly, giving an awkward effect. As for vectors, it almost seems the author avoids meeting a vector face to face, since we have the same equation labeled "vector" that rearranged is labeled "scalar." The student has learned there is no division by vectors. Elements of vector analysis which could be profitably used enter by the back door, and remain unnamed. Refusal to recognize the component of a vector, normal to a surface, results in surface density of charge becoming a vector. We find volume density a scalar, surface density a vector, and lineal density apparently unspecified, possibly because the equation which follows the first use of lineal density requires that it be a vector, whereas the second equation which follows requires it be a scalar.

The stated objective of the text is "to make it relatively easy for any student of electrical engineering, and more particularly those planning a life work in electronic circuit engineering, to achieve a reasonably complete and satisfactory understanding of the internal functioning of the electron devices that serve as the active elements in electronic circuits." This reviewer is skeptical about the phrase "relatively easy." Too many complications result from a difficult style of writing and a continual endeavor to see things in a simple light.

Consider the following: "in rotary motion, such as is here being implied, the term acceleration is defined as a ratio of force to mass, *not* as a velocity component's time derivative." Since in vector analysis or mechanics it is a customary exercise to show that the result is what the author says it is not, the statement might have been omitted.

The reviewer regrets that the text is not as easy to follow as the figures are to understand, that the student may be substantially fenced off from corresponding material to which he should be relating his work, and that the avoiding of direct mathematical derivations and the indirectness of some approaches result in confusion for the reader. Nevertheless, the reader will want to consider that the first edition, despite its shortcomings, compared well with others in its field.

J. G. Brainerd
Moore School of Electrical Engineering
Philadelphia, Pa.

The International Telecommunication Union an Experiment in International Cooperation by George Arthur Codding, Jr.

Published (1952) by E. J. Brill, Leiden, Netherlands. 496 pages+9-page index+xvii pages. $6\frac{1}{2}\times 9\frac{1}{2}$. Price: 25 Dutch Florins.

This is a story of the development of the International Telecommunication Union. The book begins with the founding, organization, and development of its predecessor, the International Telegraph Union, in 1865. Chapter two deals with the International

Books

Radiotelegraph Union and the International Conferences during the period from 1903 (Berlin Preliminary Radio Conference) to 1927 (Washington Radio Conference).

Chapter three describes the merger of the telegraph and radio conventions or treaties into the single Telecommunication Convention formulated at Madrid in 1932, and reviews the subsequent meetings at Cairo and elsewhere prior to World War II. The following chapter discusses the effects of the war on international communication activities and refers to the Moscow Preparatory Conference of 1946.

The next three chapters relate to the Atlantic City Conferences procedures, languages used at the Conferences, the radio-frequency allocation problems which resulted from the new techniques, and expanded uses of radio, particularly for aeronautical service and high-frequency broadcasting. There is extensive treatment of the formal and legal aspects of the Plenipotentiary and Administrative Conferences, arrangements for the settlement of disputes, and relationships between ITU and other international organizations.

Another chapter describes the International Telegraph and Telephone Conference held in Paris in 1949. It also presents in considerable detail the work of the Provisional Frequency Board and the eleven conferences, both general and regional, which have been endeavoring to find a solution to the extremely complex problems involved in the international registration of frequency assignments. This chapter is of special value to those who are now dealing with this phase of international radio regulation. The latter part of this material discusses the organization and recent meetings of the three International Consultative Committees (Telegraph, Telephone, and Radio) which are known, respectively, as the CCIT, CCIF, and CCIR.

Chapter nine relates to the present organization of the International Telecommunication Union, its administrative council, secretariat, and finances.

A thirteen-page section entitled, "Summary Conclusions," gives the following statement: "The International Telecommunication Union is a symbol of almost a century of international co-operation aimed at the regulation of three important elements of modern society: telegraph, telephone, and radio. The manner in which this regulation evolved is exemplary in its progressiveness and continuity. One must always keep in mind that, except in the case of a few of the early conferences, this work has been carried out almost exclusively by telecommunication experts. Their disregard for legal formalities has at times led them into temporary difficulties, but at the same time it has prevented them from becoming so deeply involved in many of the ever present political machinations as to neglect their essential work."

The book includes one hundred and ninety-one footnote references, principally to the official documents of the various international conferences and committees and other documents issued by the ITU, its predecessors or their subsidiary organizations. The Appendices include the agreement between the United Nations and the International Telecommunication Union and a list of the countries of the world showing their relationship to the "Acts" of the Union (Atlantic City Convention, 1947, Regulations and Geneva Agreement, 1951). One 18-page Appendix is a bibliography which lists about five hundred books, magazine articles, and other publications relating to the regulation of international communication services.

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New Zealand Radio Meteorological Investigation, The Canterbury Project Report of Factual Data Volume I

Published (1952) by the Department of Scientific and Industrial Research, Wellington, New Zealand. 858 pages+xi pages. 11 figures. XVII tables. 5¾×9.

During 1946 and 1947, the government of New Zealand sponsored a comprehensive investigation of low-level atmospheric ducts resulting from the passage of warm dry air from land to over the ocean. The volume under review is the first of a three-volume report of the results of this investigation; Volumes II and III are to be devoted to graphs of the meterological and radio data, respectively.

About 100 pages are devoted to a discussion of the radio-meteorological problem, the equipment used, the site of the experiments, and the weather situation. The remainder of the volume is devoted to tabulations of meteorological data from surface and upper air weather reports, wired sonde observations, and aircraft meteorological observations. The tabular data for each day are preceded by comments on the observations and conditions prevailing, and an over-all discussion of the resulting meteorological and radio data. The discussion of the accuracies obtained in the various measurements is unusually well done, and should be very helpful in the analysis of the results, as well as serving as an excellent guide to those planning similar experiments. The Report will be of value to those engaged in exploring the formation of ducts and their effects on propagation conditions.

Martin Katzin
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Washington, D. C.

Molecular Microwave Spectra Tables, by Paul Kisliuk and Charles H. Townes, National Bureau of Standards Circular 518

Published (1952) by the Government Printing Office, Washington 25, D. C. 127 pages+vi pages. 10¼×7¾. $0.65, with additional ¼ the cost for foreign mailing.

This group of tables gives the frequencies, assignment of quantum number and intensities of approximately 1,800 microwave absorption lines. It also includes the best available values of other pertinent molecular data, such as rotational constants, dipole moments, quadrupole coupling constants, and rotation-vibration interaction constants. In addition to listing the frequencies once for each molecule, the circular lists them again in consecutive ascending order of frequency.

The tables constitute a modernization and revision of these previously published and contain a considerable amount of otherwise unpublished information. Revisions are planned for publication from time to time as intensive work in microwave spectroscopy makes the tables obsolescent.

Only molecular lines of frequency greater than 1,000 mc are listed. This excludes resonances found by molecular beam techniques rather than the usual microwave absorption measurements as well as lines such as those of atomic hydrogen and cesium that fall in the microwave region. Microwave absorption in paramagnetic gases due to transitions between Zeeman components is not included.

References are given for all data in the tables. For easy calculation of quadrupole hyperfine structure, a tabulation of Casimir's function is included. The introduction contains explanations of the tables and a short discussion of microwave spectra and important formulas. Indices give an alphabetical list of authors and lists of molecules by name and by chemical symbol.

High Energy Particles by Bruno Rossi

Published (1952) by Prentice-Hall, Inc., Publishers, 70 Fifth Ave., New York 11, N. Y. 525 pages +15-page bibliography+20-page appendix+7-page index+xii pages. 239 figures. 5¾×9. $12.50.

Bruno Rossi is a professor of physics at Massachusetts Institute of Technology, Cambridge, Mass.

This unusually well-written book was intended by Professor Rossi to serve as a reference handbook for students and research workers in the fields of cosmic rays and artificially accelerated high-energy particles (a fact attested by the extensive bibliography); however, it should have a much broader appeal. The same clarity of style, completeness and timeliness that will make the book indispensable to the active worker in the high-energy field make it invaluable also to the engineer or physicist (not in the field) who wishes to bring himself up to date in the bewildering realm of mesons and other elementary particles.

The book is divided into eight chapters. The first is an introduction discussing the role of elementary particles in physics. Chapter two is devoted to the theory of electromagnetic interactions, and the next chapter treats experimental methods. Elementary particles are discussed in chapter four, with the complex subject of cascade showers following. Chapter six summarizes the experimental results on the electromagnetic interactions of high-energy particles, while the nuclear interactions of artificially accelerated particles and of cosmic rays are covered in chapters seven and eight.

Writing a book on a rapidly developing subject is a formidable task, and one that could be attempted only by an active contributor to the field, such as Professor Rossi. He is to be congratulated on the excellence of the work he has done.

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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. I.R.E., from February, 1952 through January, 1953, is published by the *Wireless Engineer* and may be purchased for $0.68 (including postage) from the Institute of Radio Engineers, 1 East 79th St., New York 21, N. Y. As supplies are limited, the publishers ask us to stress the need for early application for copies. 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

016:534 910

References to Contemporary Papers on Acoustics—R. T. Beyer. (*Jour. Acoust. Soc. Amer.*, vol. 24, no. 6, pp. 790–798; November, 1952.) Continuation of 1 of 1953.

534.+621.3].001.362 911

Acoustics in Relation to Radio Engineering—E. G. Richardson. (*Jour. Brit. I.R.E.*, vol. 12, pp. 577–584; November, 1952.) Analogies between acoustic and em phenomena are demonstrated and, in particular, the relations between acoustic and em impedors, resonators, filters, transmission lines, waveguides and radiators are discussed. Phenomena in the application of sound ranging are compared with ionospheric effects in the propagation of em waves.

534.23:534.374 912

Acoustical Interference Filters—K. K. Curtis & L. N. Hadley. (*Jour acoust. Soc. Amer.*, vol. 24, pp. 721–725; November, 1952.) The propagation of normally incident waves through a system comprising five parallel layers of distinct media is discussed. Analysis indicates that such a system passes a number of narrow bands whose mid-frequency spacings depend on the thickness of the third layer. The observed response of experimental filters comprising alternate layers of air and 0.02 mm Al foil agrees fairly well with theoretical predictions over the investigated frequency range of 5–20 kc.

534.231 913

An Exact Solution of the Acoustical Field near a Circular Transducer—E. W. Guptill. (*Jour. Acoust. Soc. Amer.*, vol. 24, p. 784; November, 1952.)

534.24 914

The Reflexion of a Spherical Acoustic Pulse by an Absorbent Infinite Plane, and Related Problems—P. E. Doak. (*Proc. Roy. Soc. A*, vol. 215, pp. 233–254; November, 25, 1952.) A formal integral solution of the problem is given for the case of a plane whose impedance depends in any way on the frequency and angle of incidence of the pulse. In many practical cases the impedance can be assumed to be independent of the angle of incidence; in this case the integral is relatively easy to evaluate, and a simple exact expression, in closed form, is obtained for the reflected pulse when the wall impedance is purely resistive, and consequently independent of frequency. The formal integral solution is evaluated approximately for wall impedances of the types: (*a*) resistance and mass; (*b*) resistance and stiffness; (*c*) resistance, mass and stiffness. The solutions are compared with the corresponding solutions for plane incident waves.

534.321.9:621.315.616.9 915

The Effectiveness of Plastic Focusing Lenses with High-Intensity Ultrasonic Radiation—J. A. Bronzo & J. M. Anderson. (*Jour. Acous. Soc. Amer.*, vol. 24, pp. 718–720; November, 1952.) Plexiglas and polystyrene lenses were investigated, using a frequency of 1 mc. Plexiglas showed signs of breakdown at intensities above about 2 W/cm², while polystyrene showed no signs of damage at intensities up to 6.6 W/cm².

534.614 916

Systematic Errors in Indirect Measurements of the Velocity of Sound—P. W. Smith, Jr. (*Jour. Acoust. Soc. Amer.*, vol. 24, pp. 687–695; November, 1952.)

534.64 917

Apparatus for Absolute Measurement of Analogous Impedance of Acoustic Elements—G. B. Thurston. (*Jour. Acoust. Soc. Amer.*, vol. 24, pp. 649–652; November, 1952.) Description of hydrodynamic test equipment for measuring pressure and flow in liquid-filled tubes, at frequencies up to 700 cps.

534.78+534.756 918

A Phase Principle for Complex-Frequency Analysis and its Implications in Auditory Theory—W. H. Huggins. (*Jour. Acoust. Soc. Amer.*, vol. 24, pp. 582–589; November, 1952.) A filtering scheme based on the phase-frequency characteristic of a filter is useful for analyzing signals, such as speech, which are produced by excitation of a system having one or more resonances. The phase principle is particularly well suited to neural mechanisms of inhibition and facilitation, and may be the principle on which the ear actually operates.

534.78 919

Some Experiments on the Perception of Synthetic Speech Sounds—F. S. Cooper, P. C. Delattre, A. M. Liberman, J. M. Borst & L. J. Gerstman. (*Jour. Acoust. Soc. Amer.*, vol. 24, pp. 597–606; November, 1952.) An investigation of the characteristics of some consonant sounds. Methods, results and working hypotheses are discussed.

534.78:534.756 920

On the Process of Speech Perception—J.C.R. Licklider. (*Jour. Acoust. Soc. Amer.*, vol., 24 pp. 590–594; November, 1952.)

534.843.001.57 921

Analysis of Transient Acoustic Phenomena by means of Reduced-Scale Models—T. Korn and F. Kirschner. (*Ann. Télécommun.*, vol. 7, pp. 414–420; Oct., 1952.) In determinations of speech intelligibility in halls, transient phenomena must be taken into account. Methods of investigating such effects, using models, are described in detail. The materials used in the construction of the models must be chosen so that the coefficient of absorption at a frequency $f' = kf$ is equal to that of the corresponding hall materials at the test frequency f, k being the scale reduction factor for the model. Tests made at a frequency of 15.36 kc on 1/30-scale models, using pulses of duration 20 ms, illustrate the importance of hall geometry as regards intelligibility.

534.844/.845 922

A Reverberation Chamber with Polycylindrical Walls—J. H. Botsford, R. N. Lane and R. B. Watson. (*Jour. Acoust. Soc. Amer.*, vol. 24, pp. 742–744; Nov., 1952.) A chamber designed for low energy loss is described; it is constructed of reinforced concrete, with diffusing cylinders of random size cast in the ceiling and walls, and the whole of the interior surface is heavily enamelled. The chamber is used for determining the absorption coefficients of materials in the frequency range 250 cps–12 kc.

534.845 923

Acoustic Propagation in Granular Media—R. W. Morse. (*Jour. Acoust. Soc. Amer.*, vol. 24, pp. 696–700; Nov., 1952.) Formulas derived for the propagation velocity and attenuation in a medium composed of closely packed solid particles immersed in a fluid give values in good agreement with experimental results obtained by other workers for the case of uniform-size particles.

534.845/.846] 621.396.712.3 924

Membrane Sound Absorbers and their Application to Broadcasting Studios—C. L. S. Gilford. (*BBC Quart.*, vol. 7, no. 4, pp. 246–256; Winter 1952–53.) A membrane of ordinary bitumen roofing felt closes the front of a wooden framework with side walls of inexpensive sheet material. The membrane may be backed by a blanket of rock-wool or other absorptive material; this prevents a rapid reduction of absorption as the resonance frequency rises. A protective cover in front of the membrane consists normally of perforated hardboard which acts as an acoustic filter. The air space behind the membrane is divided by partitions. The resonance frequency of such membranes is calculated, the theory of the absorption is discussed, and experimentally determined absorption characteristics of single and double membranes are shown. The use of devices of this type for low-frequency absorption in broadcasting studios is illustrated.

534.845.1 925
Tube Method of Measuring Sound Absorption—H. O. Taylor. (*Jour. Acoust. Soc. Amer.*, vol. 24, pp. 701–704; Nov., 1952.) Kennelly and Kurokawa (*Proc. Amer. Acad. Arts Sci.*, vol. 56, no. 1; 1921) measured acoustic impedance by a method in which a telephone receiver at one end of a tube served both as sound source and detector, the other end of the tube being closed by a sliding piston. A similar technique is now used to determine the absorption of the material forming the face of the piston.

621.395.61:546.431.824-31 926
Theoretical Sensitivity of a Transversely Loaded Circular Bimorph Transducer—E. G. Thurston. (*Jour. Acoust. Soc. Amer.*, vol. 24, pp. 656–659; Nov., 1952.) A calculation of the sensitivity is made for centrally supported $BaTiO_3$ transducers, on the basis of elasticity theory. A value of 0.5 mV/microbar can easily be attained for low-frequency applications. Experimental results show reasonable agreement.

ANTENNAS AND TRANSMISSION LINES

621.315.14.027.8:621.396.82 927
Radio Noise in Relation to the Design of High-Voltage Transmission Lines—H. L. Rorden and R. S. Gens. (*Elec. Eng., N. Y.*, vol. 71, pp. 873–878; Oct., 1952.) Condensed text of paper presented at the A.I.E.E. Winter General Meeting, January 1952, giving an account of tests carried out by the Bonneville Power Administration on 230-kV power lines. The results obtained indicate that objectionable interference will not be experienced in a radio receiver with antenna 200 ft away from a hv transmission line if the fair-weather rf field strength at that distance is $\not>15$ μV/m. In rainy weather the radio noise level may be 30 or more times the fair-weather noise level, drops of water on a conductor being far more effective in causing radio noise than conductor irregularities. Radio noise in lv lines crossing or parallel to hv lines may, owing to coupling between the lv and hv systems, travel considerable distances along the hv lines. For interference levels to be tolerable, conductor diameters should be at least 0.95 in. for 230 kV, 1.25 in. for 300 kV and 1.45 in. for 345 kV.

621.315.212:621.397.24:621.3.018.78 928
Some Factors affecting the Performance of Coaxial Cables for Permanent Television Links—H. Ashcroft, W. W. H. Clarke and J. D. S. Hinchliffe. (*Proc. IEE* (London), Part IIIA, vol. 99, pp. 350–356. Discussion pp. 472–478; April/May, 1952.)

621.392.21:621.396.67:621.397.6 929
Suspended Locked-Coil-Rope Television Feeder Systems—E. C. Cork. (*Proc. IEE* (London), Part IIIA, vol. 99, no. 18, pp. 243–252. Discussion pp. 264–269; 1952.) A detailed description of the feeder system noted in 244 of January.

621.392.21.09 930
A U.H.F. Surface-Wave Transmission Line—C. E. Sharp and G. Goubau. (Proc. I.R.E., vol. 41, pp. 107–109; Jan., 1953.) Description of an antenna-feed unit for operation in the frequency range 1.7–2.4 kmc and designed to fulfil the requirements of insertion loss <3 db for 150-ft length and swr $\not>1.5$. The wire used is 0.102-in. diameter soft-drawn Cu covered with a 0.014-in. layer of extruded polyethylene. The launching device is described and illustrated.

621.392.26:538.221 931
Ferrites at Microwaves—N. G. Sakiotis and H. N. Chait. (Proc. I.R.E., vol. 41, pp. 87–93; Jan., 1953.) Propagation in an unbounded ferromagnetic medium and in waveguides containing ferrites is discussed; results are reported of an experimental study on circular and rectangular waveguides enclosing commercially available ferrites, using a frequency of 9.375 mc. Applications to switches, aerials etc. are described.

621.317.352:621.392.26 932
Experimental Study and First Results of Measurements of the Attenuation of TE_{01} (H_{01}) Waves in Short Sections of Straight Circular Waveguide—Comte and Ponthus. (See 1068.)

621.396.67:621.316.54 933
A Radio Transmission-Line Exchange—(*Engineer* (London), vol. 194, pp. 451–452; Oct. 3, 1952.) Another account, with fuller details, of the equipment noted in 33 of January.

621.396.67:621.397.6 934
Television Transmitting Aerials—H. Cafferata, C. Gillam and J. F. Ramsay. (*Proc. IEE* (London), Part IIIA, vol. 99, no. 18, pp. 215–230. Discussion, pp. 264–269; 1952.) Factors affecting the design of a transmitting antenna are reviewed, with particular reference to common-antenna working for vision and sound. The properties of 2-phase antenna systems are discussed and curves are presented which enable the horizontal radiation patterns of such antennas to be interpreted in terms of impedance measurements on the distribution-feeder system. The development of the Sutton Coldfield antenna is described and modifications adopted for Holme Moss and the later high-power stations are explained.

621.396.677 935
Gain of Electromagnetic Horns—E. H. Braun. (Proc. I.R.E., vol. 41, pp. 109–115; Jan., 1953.) Results obtained by Jakes (1847 of 1951) are verified by further experiments. Theory is developed which gives results in good agreement with experimental data, and explains why the $2D^2/\lambda$ far-field criterion is invalid, D being the largest horn dimension. Curves are presented from which the error in gain measured at any distance may be obtained.

621.396.677 936
Plane Reflectors for Ray Deflection with Directive Antennas—H. G. Unger. (*Frequenz*, vol. 6, pp. 272–278; Sept., 1952.) Systems are considered in which a parabolic or other antenna located on the ground directs radiation on to an elevated reflector which reflects the beam in the required direction of propagation. Using the Kirchhoff-Huyghens diffraction-theory approximation a determination is made of the field strength at the receiver with and without the reflector. The conditions are derived for which the introduction of the reflector causes no reduction of field strength. For large reflector areas there is a gain which has a maximum value for a certain geometrical arrangement; an explanation is given in terms of interference theory.

621.396.677 937
A New Electromagnetic Lens—G. von Trentini. (*Rev. Telegr. Electronica* (Buenos Aires), vol. 40, pp. 341–345; June, 1952.) The artificial dielectric used consists of two or more linear arrays of vertical wires arranged parallel to the electric vector, each wire being reactively loaded at regular intervals. Tests on various arrays gave results in good agreement with calculation. A 26-cm wide plano-concave lens is described, with loading varying from capacitive at the center to inductive at the outer edges. This results in a refractive index varying discontinuously from 0.2 to 2.1. At λ 3.2 cm, the gain was about 7, half-power beam-width 7° and nearest side-lobes 8%, comparing favorably with the performance of a parallel-plate metal lens of width 30 cm.

621.396.677:621.396.9 938
Wide-Angle-Scan Radar Antenna—H. N. Chait. (*Electronics*, vol. 26, pp. 128–132; Jan., 1953.) A description is given of the construction and the method of design of an antenna which is the microwave analogue of the Schmidt optical system used in projection-type television receivers. A feed horn pivots about the center of curvature of a semicylindrical reflector and the plane wave front is restored by means of a specially shaped polystyrene lens. An exact solution for the shape of the correcting lens was obtained by use of the zero-phase method, which is described in detail. Curves are given showing the variation of side-lobe level, beam width and relative gain with change of the scanning angle, and another set of curves shows the improvement resulting from off-axis correction of the lens shape. A scan of 20 beam-widths is obtained with side lobes 20 db below the peak intensity.

621.396.677.5:621.318.132 939
Ferroxcube Aerial Rods—H. van Suchtelen. (*Electronic Appl. Bull.*, vol. 13, pp. 88–100; June 1952.) The size of loop antennas can be greatly reduced by using a core of some suitable magnetic material. The design is discussed of antennas using small coils wound on rods of ferroxcube. The losses and Q-factors of such antennas are investigated and experimental results are quoted. Temperature effects on the permeability of the core material are analyzed.

621.396.67.001.11 940
Advanced Antenna Theory [Book Review]—S. A. Schelkunoff. Publishers: J. Wiley & Sons, New York, 216 pp., $6.50; 1952. (*Electronics*, vol. 26, pp. 353–354; Jan., 1953.) "The book will be of interest mainly to those engineers who have an adequate mathematical background."

CIRCUITS AND CIRCUIT ELEMENTS

621.3.015.7:621.318.57 941
Compensation against Effects of Grid-Cathode Capacitance in Pulse-Height Selectors—W. M. Grim and A. B. Van Rennes. (*Rev. Sci. Instr.*, vol. 23, p. 563; Oct., 1952.) The precision of the selector may be reduced by an effect due to the grid/cathode capacitance of the input tube. When a "constant-current" tube is used in the cathode circuit, this effect can be compensated by applying the input signal, appropriately modified, to the grid of the "constant-current" tube.

621.314.2 942
The Design of Thermocouple Transformers for Infrared Chopped Beam Systems—T. S. Robinson. (*Jour. Sci. Instr.*, vol. 29, pp. 311–313; Oct., 1952.) Description of transformer for operation at 10 cps, to be inserted between a thermocouple detector and an amplifier; the design is such as to give at the amplifier output a signal/noise ratio approaching that at the thermocouple terminals.

621.314.3† 943
Magnetic Amplifiers—M. Latour. (*Rev. gén. élect.*, vol. 61, pp. 423–424; Oct., 1952.) Comment on 2132 of 1952 (Pistoulet) pointing out the existence of several French publications and patents on the subject, one paper, by Léonard and Weber, dating back to 1906 and thus being considerably prior to the American paper, by Alexanderson and Nixdorff, quoted by Pistoulet as the first publication on magnetic amplifiers.

621.314.3† 944
A Mathematical Analysis of Parallel-Connected Magnetic Amplifiers with Resistive Loads—H. S. Kirschbaum: L. A. Pipes. (*Jour. Appl. Phys.*, vol. 23, pp. 1278–1279; Nov., 1952.) Comment on 3014 of 1952 and author's reply.

621.314.3† 945
Magnetic Amplifiers for Industry—W. F. Horton. (*Radio & Telev. News, Radio-Elec-*

tronic Eng. Section, vol. 48, pp. 10–12; Nov., 1952.) Description, with outline circuit diagrams, of a series of Westinghouse "Magamp" instruments with power ratings from 40 mW to 180 W, and examples of their application.

621.316.84/.86 946

New Developments in Fixed Resistors for Electronic Applications—R. A. Osche. (*Elec. Mfg.*, vol. 47, pp. 118–123, 256; April, 1951.) A review of the characteristics of deposited-carbon, metal-film, and borocarbon resistors, and comparison with carbon-composition and wire-wound resistors.

621.316.86 947

High-Temperature Carbon-Film Resistors—(*Electronics*, vol. 26, pp. 148, 154; Jan., 1953.) Details are given of the method of pyrolysis used in the production of stable carbon-film resistors with values from 10Ω to 5MΩ and capable of operation at temperatures up to 200°C. The carbon film is deposited on a porcelain base made from alumina mixed with other ingredients such as SiO_2 and CaO in an alcohol binder.

621.316.86:537.312.6 948

Industrial Applications of Semiconductors. Part 8—Thermistors—R. E. Burgess. (*Research* (London), vol. 5, pp. 469–474; Oct., 1952.) Description of the characteristics of thermistors, typical forms of construction, and various applications.

621.318.435.3.025.3:621.314.3† 949

Three-Phase Transductor Circuits for Magnetic Amplifiers—A. G. Milnes. (*Proc. IEE* (London), Part IV, vol. 99, pp. 336–357; Dec., 1952. Digest, *ibid.*, Part II, vol. 99, pp. 615–619; Dec., 1952.) Various transductor connections for 3-phase operation are described; a 3-element circuit with separate excitation and a 6-element circuit with auto-excitation are discussed in detail. A theoretical and experimental comparison is made between single-phase and 3-phase circuits. The latter should be used where balanced loading of the supply is sufficiently desirable to compensate for the increased circuit complexity.

621.318.57:621.396.615.18 950

The Coupling of "Scale-of-Two" Circuits—R. Ascoli. (*Nuovo Cim.*, vol. 8, pp. 584–585; Aug. 1, 1951.) A simple and reliable RC divider circuit is described and illustrated.

621.318.57+621.392.52]:621.397.26.029.63 951

Ultra-High-Frequency Switches and Filters—G. F. Small. (*Proc. IEE* (London), Part IIIA, vol. 99, pp. 464–472. Discussion, pp. 472–478; April/May 1952.) The design of filters and switches for the London-Birmingham television relay link is described. The type of switch adopted uses resonant stub lines in shunt with the main transmission lines; typical performance details are given. The filters are of the band-stop type and consist of stub lines, composed of a series of nominal λ/4 line sections, shunted across the main transmission line. Design curves are given and typical filter-attenuation curves shown.

621.318.57:681.142 952

Further Data on the Design of Eccles-Jordan Flip-Flops—M. Rubinoff. (*Elec. Eng., N. Y.*, vol. 71, pp. 905–910; Oct. 1952.) Full text of paper presented at the A.I.E.E. Summer General Meeting, June 1952. Analysis, under specified conditions, of the grounded-cathode type of flip-flop circuit including no inductors. A graphical design technique is described which should facilitate the design of switching circuits using a large number of identical direct-coupled circuits.

621.319.4 953

Fixed Capacitors for Electronic Circuits—P. S. Schmidt. (*Elec. Mfg.*, vol. 47, pp. 100–105, 248; May 1951.) A review of the development of new types, involving both new materials and new manufacturing techniques.

621.319.4:621.396.615 954

Tuning Stability Nomograph—J. T. Hogan. (*Rev. Sci. Instr.*, vol. 23, p. 566; Oct. 1952.) A chart is given for facilitating calculation of the parallel capacitors required to give an over-all temperature coefficient of zero in circuits using inductance tuning.

621.392:517.63 955

The Calculation of Energy Flow using the Laplace Transformation—A. C. Sim. (*Proc. IEE* (London), Part IV, vol. 99, no. 4, pp. 376–382; Dec. 1952. Digest, *ibid.*, Part II, vol. 99, no. 72, p. 624; Dec. 1952.) The Laplace-Parseval integral is applied to determine energy flow directly from the Laplace transforms of the applied force and the resultant response. The method is applicable to both transient and steady states, and leads directly to the most simple form of solution. The method is most advantageous when the applied force possesses several discontinuities. As an example, the classical problem of eddy-current losses in linear sheet conductors is solved in general for arbitrary excitations.

621.392.5 956

On Transformations of Linear Active Networks with Applications at Ultra-High Frequencies—H. Hsu. (PROC. I.R.E., vol. 41, no. 1, pp. 59–67; Jan. 1953.) A relation similar to the star-delta transformation is developed for linear circuits including tubes. Formulas are derived for triode circuits under both negative-grid and positive-grid conditions; the results are extended to tetrodes and pentodes. The method is useful for analyzing the operation of uhf triodes and for determining amplifier admittances and gain.

621.392.5 957

Response of a Quadripole to Discontinuous Signals—The Phenomenon of Gibbs and Methods of Generalized Summation—T. Vogel. (*Ann. Télécommun.*, vol. 7, pp. 421–428; Oct. 1952.) Mathematical analysis with application to the design of quadripoles with good transmission characteristics for square-wave signals.

621.392.5:[621.396.615+621.396.645 958

Tunable *RC*-Bridge Network with only One Variable Element—W. Götze. (*Funk u. Ton*, vol. 6, pp. 393–399; Aug. 1952.) A circuit is described with which an output of required phase is obtained by varying either one resistance or one capacitance, the magnitude of the output voltage remaining constant. Details are given of a stable tube generator with frequency ranging from a very low value to 10^7 cps. The circuit can also be used in phase shifters, frequency meters and selective amplifiers.

621.392.5.018.7 959

A Theory of Time Series for Waveform-Transmission Systems—W. E. Thomson. (*Proc. IEE* (London), Part IV, vol. 99, pp. 397–409; Dec. 1952.) A theoretical study is made of time series suitable for application to low-pass systems with a fixed upper limit to the pass band. For "division" and "inversion" of time series, algebraic division is not, in general, legitimate for time series related to bandwidth limitation, although it is legitimate for certain other forms of time series. More general methods of inversion are treated and an iterative method, which can always be used, is discussed in detail and a practical computing scheme is outlined. See also 55 of January (Lewis), which also deals with the time-series method.

621.392.5.029.63 960

A Wide-Band Hybrid Ring for U.H.F.—W. V. Tyminski and A. E. Hylas. (PROC. I.R.E., vol. 41, pp. 81–87; Jan. 1953.) The bandwidth of a 6λ/4 hybrid ring is increased by replacing the 3λ/4 long arm with a λ/4 line and introducing a frequency-insensitive reversal of phase. Calculations are made of input and transfer admittances, insertion loss and the effect of capacitance across the loads. Some experimental results are given, and applications to power splitters, harmonic generators, mixers etc. are indicated.

621.392.52 961

Wave-Filter Characteristics by a Direct Method—R. C. Taylor and C. U. Watts. (*Elec. Eng., N. Y.*, vol. 71, p. 911; Oct. 1952.) Digest of paper presented at the A.I.E.E. Winter General Meeting, January 1952. A chart is given from which the attenuation (in db) can be directly determined for seven commonly used ladder-filter sections, the abscissae used being the number of bandwidths from the cut-off frequency.

621.392.52 962

Formulas for calculating Filter Circuits with Flattened Attenuation Curves—G. Bosse. (*Funk u. Ton*, vol. 6, pp. 416–425 and 493; Aug. and Sept. 1952.) Simple formulas are given for calculating Tchebycheff-type low-pass filters with attenuation values falling between prescribed limits either inside or outside the pass band. Attenuation curves obtainable with filters thus calculated are shown.

621.392.52:621.3.015.3 963

The Transient Response of R.F. and I.F. Filters to a Wave Packet—A. W. Gent. (*Proc. IEE* (London), Part IV, vol. 99, pp. 326–335; Dec. 1952.) Full paper. See 350 of February.

621.392.54:538.69 964

Characteristics of the Magnetic Attenuator at U.H.F.—F. Reggia and R. W. Beatty. (PROC. I.R.E., vol. 41, pp. 93–100; Jan. 1953.) The design of the field-controlled attenuator previously described by Reggia (1237 of 1952) is discussed, and performance data are given for attenuators using various ferromagnetic materials.

621.395.665.1:534.86 965

New Principle for Electronic Volume Compression—H. E. Haynes. (*Radio & Telev. News, Radio-Electronic Eng. Section*, vol. 48, pp. 7–9, 29; Nov. 1952.) Reprint. See 2152 of 1952.

621.396.611.21 966

Loading of Quartz Oscillator Plates—L. T. Sogn and P. A. Simpson. (*Jour. Res. Nat. Bur. Stand.*, vol. 49, pp. 325–327; Nov. 1952.) An investigation was made of the effectiveness of mechanical loading for eliminating unwanted modes of vibration and adjusting the oscillation frequency towards the value corresponding to greatest activity. The loading was done by applying various amounts of Wood's metal to different parts of the plate surfaces. For concave thickness-shear-mode and X-cut plates, improvement was obtained when the loading was applied at or near the active central area; for flat or convex X-cut plates with an extensional mode of vibration improvement was obtained when the loading was applied near the periphery. For both types, loading increased the Q factor.

621.396.611.4 967

Microwave Cavity Resonators as Circuit Elements—S. K. Chatterjee. (*Jour. Indian Inst. Sci.*, Section B, vol. 34, pp. 99–112; Oct. 1952.) The differential equation for the equivalent circuit of a double-loop-coupled cavity resonator is derived by using Lagrange's equation and Maxwell's equations. The losses in the walls and the Q of the cavity operating in the TE_{11n} mode are evaluated.

621.396.615 968

Power Spectrum of a Nonlinear Oscillator with a Frequency/Amplitude Law, Perturbed by Noise—A. Blaquière. (*C. R. Acad. Sci.* (Paris), vol. 235, pp. 1201–1203; Nov. 17, 1952.) Discussion for the case where the dependence of the frequency ν on the amplitude a is represented by the formula $\nu = \nu_0(1 + 3\mu a^2/8)$. The power spectrum is determined and can be represented as a single line with a superposed noise spectrum resulting from the amplitude and phase fluctuations. The methods previously developed (2162 and 2486 of 1952) enable the noise spectrum to be resolved into a continuous band and a periodic line component.

621.396.615 969
Multiple-Feedback Oscillators—(*Electronics*, vol. 26, pp. 200, 208; Jan. 1953.) The use of several feedback circuits in connection with a single tank circuit makes it possible to construct oscillators capable of operation over a wide frequency range, the *LC* value of the tank circuit remaining fixed. Such oscillators, and the theory on which they are based, are described by M. Morrison in U.S.A. patent No. 2587 750. Important points of the theory are here given and two typical oscillator circuits are described. In the second of these a special current transformer, with a 10:1 ratio, is used to effect phase alteration of the feedback current. The transformer, with a variable capacitor across the output terminals, is adjusted to provide feedback current in phase with the ac component of the anode load current, so that the proper grid-voltage angle is obtained for maintaining oscillation at the desired frequency.

621.396.615 970
Constant-Amplitude Oscillator—(*Tech. News Bull. Nat. Bur. Stand.*, vol. 36, pp. 172–173; Nov. 1952.) A circuit developed by N. C. Hekimian comprises a conventional oscillator whose rectified output is applied as a positive voltage to the grid of a clamp tube which shares the voltage-dropping resistance in the oscillator anode-supply lead. For a more detailed account see *Electronics*, vol. 24, p. 164; July 1951.

621.396.615:517.941.91 971
On an Equation connected with the Theory of Triode Oscillations—R. A. Smith. (*Proc. Camb. Phil. Soc.*, vol. 48, part 4, pp. 698–717; Oct. 1952.) Detailed discussion of the periodic solutions of the differential equation

$$\ddot{x}+k\dot{x}f(x)+x=pk\lambda\cos(\lambda t+\alpha),$$

in which the parameter k is small. The equation is concerned with the forced oscillations in a simple electrical circuit containing a triode. Van der Pol has treated the case where $f(x)=x^2-1$, and an extensive theory of triode oscillations has been built up for this case. In practice, however, the graph of $f(x)$ has a much flatter bottom than that of the function (x^2-1), and Cartwright (2740 of 1948 and 1417 of 1951) has raised the question whether this flattening will alter the performance of the circuit near resonance. The main part of this paper is an attempt to answer this question. An estimate is made of the permissible degree of flattening of the $f(x)$ curve without the detuning diagram losing any of its significant properties. Various pertinent theorems are established and the possible periodic solutions of the equation are discussed in relation to the detuning diagram. Particular results include the following: —(*a*) if k is small enough and the value of λ lies in a certain small range near 2, forced oscillations with half the applied frequency are possible; two of these have amplitudes very large compared with that of the applied EMF one mode being stable and the other unstable, (*b*) only when the graph of $f(x)$ is asymmetrical can there be an appreciable range of values of λ for which subharmonic oscillations of order $\frac{1}{2}$ take place.

621.396.615:621.316.729 972
Pull-In of Nonlinear Oscillators with Filtered Output—G. Cahen and J. Loeb. (*Ann. Télécommun.*, vol. 7, pp. 411–413; Oct. 1952.) The effect is considered of applying a small voltage δV, of frequency $f+\delta f$, to an oscillator whose fundamental frequency f has been freed from harmonics by filtering. A general formula is derived which gives the maximum value of δf for pull-in to occur. For the particular case of an oscillator with a single oscillatory circuit, the formula reduces to the known expression $\delta f/f=1/(2Q\cdot\delta V/V)$, Q being the quality factor of the circuit and V the voltage amplitude of the free oscillations. A formula is also given for the threshold value of δV below which pull-in is not possible.

621.396.615.142.2:621.392.54:538.69 973
An X-Band Sweep Oscillator—I. D. Olin. (Proc. I.R.E., vol. 41, pp. 10–13; Jan. 1953.) A generator which sweeps through the frequency band 8.5–9.5 kmc in 1.5 sec is described; a mechanically tuned klystron is used. The output level is maintained constant to within ±0.1 db by means of an automatic control circuit including a variable attenuator with ferrite rotator unit (above), which also provides am at 1 kc.

621.396.615.17 974
Production of Standard Waves with a 3000-kV Impulse Generator—N. Narayan and K. S. Prabhu. (*Jour. Indian Inst. Sci.*, Section B, vol. 34, pp. 113–122; Oct. 1952.) Equipment installed at Bangalore is discussed. The characteristics of the generator were determined experimentally; a method of setting it up to produce a desired standard waveform is described.

621.396.615.17 975
Multichannel Crystal Control of V.H.F. and U.H.F. Oscillators—A. Hahnel. (Proc. I.R.E., vol. 41, pp. 79–81; Jan., 1953.) The frequencies of a number of channels are controlled by a single crystal, using an oscillator whose phase is varied periodically by application of pulses; the oscillator output comprises a spectrum of harmonically related frequencies whose fundamental is the crystal-controlled pulse frequency. A particular triode oscillator covers the range 250–850 mc with fundamental in the range 1–10 mc. By appropriately tuning the tank circuit any desired harmonic can be emphasized by an amount up to 40 db.

621.396.645:621.396.822 976
Methods of Reducing the Ratio of Noise to Signal at the Output Terminals of Amplifiers—M. J. O. Strutt. (*Elektrotech. Z.*, vol. 73, pp. 649–653; Oct. 11, 1952.) A summarized account of earlier work by the author alone or in collaboration with A. van der Ziel. See 3041 of 1942, 1075, 2088 and 2089 of 1943, 749 and 750 of 1945, 2843, 2844 and 3740 of 1946, 1573 and 3067 of 1947, 1599, 2074 and 3358 of 1948, 1332, 1487, 1631 and 2312 of 1949.

621.396.645:621.397.24 977
The Design of Amplifiers for the Birmingham-Manchester Coaxial Cable—W. T. Duerdoth. (*Proc. IEE*, Part IIIA, vol. 99, pp. 385–393. Discussion, pp. 472–478; April/May, 1952.) Description, with schematic circuit diagrams, of (*a*) a line amplifier suitable for the transmission of 405-line television signals on a 1-mc carrier, or for 16 telephony channel groups, on $\frac{3}{8}$-in. diameter coaxial cables, (*b*) a constant-gain amplifier for the transmitting terminal. Practical difficulties which limit the performance of the amplifiers and associated transformers are discussed and possible improvements for future designs are suggested.

621.396.645.029.3 978
The Maestro—a Power Amplifier—D. Sarser and M. C. Sprinkle. (*Audio Eng.*, vol. 36, pp. 19–21, 89; Nov., 1952.) Description, with complete circuit details, of a new version of the "musician's amplifier" (70 of 1950) with an output of 90 W. A pair of R.C.A. Type-6146 beam power tubes, operated in class AB, are used with a specially designed Peerless output transformer, Type S-268-Q. Power-supply circuits are also described.

621.396.645.029.3 979
New Medium-Cost Amplifier of Unusual Performance—G. L. Werner and H. Berlin. (*Audio Eng.*, vol. 36, pp. 30–31, 71; Nov., 1952.) Description, with circuit details, of an amplifier with the necessary flexibility and performance for most high-fidelity applications; only standard, readily available tubes are used.

621.396.645.029.33:621.3.018.78 980
Essential Similarity and Relations between Amplitude and Phase Distortions in Video Amplifiers—F. J. Tischer. (*Arch. elekt. Übertragung*, vol. 6, pp. 452–459; Nov., 1952.) In the complex method of representation, the transmission function of a video amplifier for steady oscillations can be regarded as the sum of an ideal transmission function and an error function. The error function is applied in investigating the relations between an arbitrary transmission function with superposed amplitude and phase errors and the associated distortions, for the minimum-phase-shift condition. The investigation gives an insight into the physical relation between the distortions. The formulas derived are applied to determine the distortions for a system with a Tchebycheff-type transmission function; these distortions are then compared with those for a system with a steady fall of amplitude near the cut-off frequency.

621.396.645.35 981
The Direct-Voltage Tube Amplifier—G. Kessler. (*Arch. Tech. Messen*, nos. 198, 200 and 204, pp. 163–166, 211–214 and 19–22; July and Sept., 1952, and Jan. 1953.) Effects of tube and circuit noise, fluctuations of the current source and temperature changes are evaluated. A review of basic amplifier circuits for measurement purposes illustrates different directly-coupled interstage and output arrangements and methods of compensating the above effects. Methods of amplification involving modulation are noted. 107 references.

621.396.645.35 982
The Parallel-T D.C. Amplifier: A Low-Drift Amplifier witb Wide Frequency Response—P. S. T. Buckerfield (*Proc. IEE* (London), Part II, vol. 99, pp. 497–506; Oct. 1952.) The frequency response of a modulated dc amplifier is broadened by the parallel addition of a conventional ac amplifier, which has inherently a response capable of being made complementary to that of the dc amplifier. The result is a homogeneous design without excessive phase shift. Circuit details are given of an amplifier designed for use with a high-speed pen recorder in the frequency range 0–90 cps. The avoidance of direct coupling renders the use of electrometer tubes unnecessary.

621.396.645.37 983
Feedback-Amplifier Design—R. J. D. Reeves. (*Proc. IEE* (London), Part IV, vol. 99, pp. 383–389; Dec., 1952.) Discussion of the pole pattern required for a low-pass amplifier with over-all feedback. Analysis is presented for an amplifier whose gain function contains no internal zeros and for which the feedback fraction is independent of frequency. A general solution, applicable to any number of poles, is obtained for the case of critical damping. For amplifiers with maximally flat response curves [3013 of 1941 (Landon)] graphical solutions are given for systems with n poles, where n ranges from 1 to 6. The synthesis of a prescribed pole pattern is considered and the method of realizing conjugate poles by feedback pairs is described.

621.396.645.371:621.397.24 984
Two Simple Types of Feedback Amplifier for the Relaying of Television Signals over Coaxial Cables—F. G. Clifford. (*Proc. IEE* (London), Part IIIA, vol. 99, pp. 367–373. Discussion, pp. 472–478; April/May, 1952.) A method is described for applying negative feedback to flatten the gain/frequency characteristic of 3-stage amplifiers using tuned-anode intervalve couplings resonant at the mid-band frequency. Two typical amplifiers are described.

GENERAL PHYSICS

061.3:538.56.029.6 985
Summarized Proceedings of a Conference

on Microwave Physics—Oxford, July 1952—D. A. Wright. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 337–341; Nov., 1952.) Summaries are given of the papers presented and of the subsequent discussions.

534.26+535.42 986
Diffraction of a Shock or an Electromagnetic Pulse by a Right-Angled Wedge—J. B. Keller. (*Jour. Appl. Phys.*, vol. 23, pp. 1267–1268; Nov., 1952.) Solutions given in a previous paper [2963 of 1951 (Keller and Blank)] are used as a basis for calculating the field distribution; comparison with measurements made for the acoustic case shows satisfactory agreement.

53.5.223 987
Precision Determination of the Velocity of Light derived from a Band-Spectrum Method—D. H. Rank, R. P. Ruth and K. L. Vander Sluis. (*Jour. Opt. Soc. Amer.*, vol. 42, pp. 693–698; Oct., 1952.) Detailed report of the work noted in 3053 of 1952. The value obtained for the velocity is 299 776±6 km/sec.

537:531].001.362 988
Extension and More Exact Statement of the Analogies between Electrical and Mechanical Devices—G. Lander. (*Frequenz*, vol. 6, pp. 235–246 and 257–266; Aug. and Sept., 1952.)

537:531].001.362 989
Electrical and Mechanical Analogies—F. Raymond. (*Rev. gén. élect.*, vol. 61, pp. 465–475; Oct., 1952.) An introduction to the study of general problems concerning electrical networks and mechanical systems whose operation is governed by linear laws. Matrix symbolism is explained by examples and a table of elementary electrical and mechanical analogues is given. Application of the theory is made to discussion of the characteristics of transducers such as loudspeakers and piezoelectric or magnetostrictive generators.

537.122 990
Dirac's New Classical Theory of the Electron—G. Höhler. (*Ann. Phys., Lpz.*, vol. 10, pp. 196–200; Feb. 15, 1952.) Dirac's new theory (1574 and 3059 of 1952) is shown to arise as a limiting case from both the particle and wave theories; for continuously distributed charge it is equivalent to the Maxwell-Lorentz theory.

537.122:538.21:511.61 991
Number Theory and the Magnetic Properties of an Electron Gas—M. F. M. Osborne. (*Phys. Rev.*, vol. 88, pp. 438–451; Nov. 1, 1952.)

537.122:538.21:511.61 992
Application of the Theory of Numbers to the Magnetic Properties of a Free Electron Gas—M. C. Steele. (*Phys. Rev.*, vol. 88, pp. 451–464; Nov. 1, 1952.)

537.221 993
A Direct Comparison of the Kelvin and Electron-Beam Methods of Contact Potential Measurement—P. A. Anderson. (*Phys. Rev.*, vol. 88, pp. 655–658; Nov. 1, 1952.)

537.226:537.1 994
The Poisson-Kelvin Hypothesis and the Theory of Dielectrics—W. B. Smith-White. (*Jour. Roy. Soc. N.S.W.*, vol. 85, Part 3, pp. 82–112; May 23, 1952.) A comprehensive critical review of fundamental mathematical theory. Electrostatics only is considered, but the treatment indicates that a complete reconstruction of electrodynamic theory is essential.

537.226.1+538.213 995
An Optical Method for determining the Complex Dielectric Constant ϵ and the Magnetic Permeability μ—H. Falkenhagen and G. Kelbg. (*Ann. Phys., Lpz.*, vol. 10, pp. 170–176; Feb. 15, 1952.) A method applicable in the cm and mm wave ranges is described which depends on determination of the normal-incidence reflection and transmission factors of a plane-parallel plate. The calculation is based on the general matrix equations of the field as derived from Maxwell's equations.

537.311.1:546.87-1 996
The Mean Free Path of Conduction Electrons in Bismuth—A. B. Pippard and R. G. Chambers. (*Proc. Phys. Soc.*, vol. 65, pp. 955–956; Nov. 1, 1952.) Experimental results for Bi single crystals are given from which it is concluded that, in agreement with the suggestion of Sondheimer (2190 of 1952), the mean free path of the conduction electrons in Bi is much longer than in normal metals, and that, for electrons traveling perpendicular to the triad axis, the path length is probably in the range 2–4 μ at room temperature.

537.523/.525 997
Secondary Processes Active in the Electrical Breakdown of Gases—L. B. Loeb. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 341–349; Nov., 1952.) Results of recent investigations indicate the complexity of gas-breakdown mechanisms. Three cathode mechanisms and two anode mechanisms are analysed and their relative importance and the conditions for their relative importance and the conditions for their occurrence are discussed. They lead to the same type of generalized threshold breakdown condition, subject to statistical fluctuations and to space-charge effects. This condition masks the active processes and renders analysis difficult. The various factors affecting the discharge are considered, and proper methods of discharge investigation are indicated. 49 references.

537.533.8 998
The Mechanism of Field-Dependent Secondary Emission—H. Jacobs, J. Freely and F. A. Brand. (*Phys. Rev.*, vol. 88, pp. 492–499; Nov. 1, 1952.) An account of experiments made to test the theory that the mechanism of field-dependent secondary emission is similar to that of the "Townsend avalanche" in gas discharges. Special tubes were used having dynodes with a porous MgO coating. The results indicate that the high yield of secondaries is independent of the base material. The variation of secondary current with field strength was in accordance with the gas-discharge equations; this was confirmed by retarding-field measurements of the energies and mean free paths of the secondaries. The time required for the surface to become charged was determined by using pulsed bombarding currents, and was found to be consistent with the theory.

537.568 999
Recombination of Gaseous Ions—H. S. W. Massey. (*Advances Phys.*, vol. 1, pp. 395–426; Oct., 1952.) Recombination between positive ions and electrons and between positive ions and negative ions are considered separately, since the nature of the processes involved and the experimental techniques for observing them are quite different in the two cases. Although there is still little information about the rate of recombination of positive and negative ions at low pressure, there has been substantial clarification of the processes of electron-ion recombination during the last two years due to the application of microwave techniques. The knowledge of the mechanism of electron-ion recombination derived from both theory and experiment is reviewed in detail, and a shorter account given of present knowledge of ion-ion recombination at pressures less than that of the atmosphere. Recombination phenomena in the ionosphere, in the solar corona, and in interstellar gas, are also discussed briefly. 80 references.

538.113 1000
Antiferromagnetism and Ferrimagnetism—L. Néel. (*Proc. Phys. Soc.*, vol. 65, pp. 869–885; Nov. 1, 1952.) The 7th Holweck Lecture, May 1952, reviewing the present state of knowledge of antiferromagnetism, including ferrimagnetism, and describing interesting phenomena related to the magnetic properties of certain ferrites and of pyrrhotite.

538.114 1001
Spin Degeneracy and the Theory of Collective Electron Ferromagnetism—A. B. Lidiard. (*Proc. Phys. Soc.*, vol. 65, pp. 885–893; Nov. 1, 1952.) A model is discussed for which the results obtained by neglecting or taking account of spin degeneracy are identical. The model can be treated exactly and leads to equations for the free energy, magnetization, etc., which are a generalization of those of Stoner's theory of collective electron ferromagnetism (2548 of 1939).

538.114 1002
Zener's Treatment of Ferromagnetism—A. Teviotdale. (*Proc. Phys. Soc.*, vol. 65, pp. 957–958; Nov. 1, 1952.) A critical review of Zener's theory.

538.114 1003
The Theory of Ferromagnetism and Heisenberg's Model—J. Yvon. (*Jour. Phys. Radium*, vol. 13, pp. 488–489; Oct., 1952.)

538.221 1004
Some Magnetic Properties of Metals: Part 3—Diamagnetic Resonance—R. B. Dingle. (*Proc. Roy. Soc. A*, vol. 212, pp. 38–47; April 8, 1952.) EM radiation incident on a large system of electrons moving in a constant magnetic field H in a metal is strongly absorbed near a frequency $\nu = eH/2\pi mc$, where m is the effective mass. The resonance absorption is of the same order of magnitude as the absorption due to skin effect. Part 2: 2490 of 1952.

538.221 1005
Some Magnetic Properties of Metals: Part 4—Properties of Small Systems of Electrons—R. B. Dingle. (*Proc. Roy. Soc. A*, vol. 212, pp. 47–65; April 8, 1952.) A calculation is made of the magnetic properties of a system of electrons within a cylinder of very small radius with its axis parallel to the field direction. The expressions obtained for magnetic susceptibility, thermodynamic potential and specific heat contain a steady term which remains of significant magnitude at all temperatures, together with terms periodic in the field which are significant only at very low temperatures. The influence of electron spin is discussed. Similar calculations are made for a small spherical system.

538.566 1006
The Limits of Total Reflection: Part 2—Rigorous Wave-Theory Calculation—H. Maecker. (*Ann. Phys., Lpz.*, vol. 10, pp. 153–160; Feb. 15, 1952.) The variation of intensity in the region of the critical reflection angle is calculated numerically for a particular case previously considered by Ott (18 of 1943). The curve obtained indicates a continuous variation from the region of partial to that of total reflection.

538.566 1007
Propagation of a Wave Front in Anisotropic Dispersive Media—M. Marziani. (*R. C. Accad. Naz. Lincei*, vol. 13, pp. 127–131; Sept./Oct., 1952.) Extension of work previously noted (390 of February) to the case of anisotropic media. Expressions are derived for the velocity of propagation and for the field in the vicinity of the wave front.

539.152.2 1008
Determination of Nuclear Moments from Hertzian Spectra—G. J. Béné. (*Jour. Phys. Radium*, vol. 13, pp. 473–479; Oct., 1952.) Discussion of the interaction between nucleus and electron, and an account of the principal methods, apart from resonance methods, of applying microwave spectroscopy to the determination of nuclear moments.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.5 1009
The Velocity Distribution of Sporadic Meteors: Part 2—M. Almond, J. G. Davies and A. C. B. Lovell. (*Mon. Not. R. Astr. Soc.*, vol. 112, pp. 21–39; 1952.) Extension of the work noted in part 1 (1899 of 1952) to the investigation of fainter meteors; similar results are obtained.

523.5:621.396.9 1010
The Velocity Distribution of Sporadic Meteors: Part 3—Calculation of the Theoretical Distributions—J. A. Clegg. (*Mon. Not. R. Astr. Soc.*, vol. 112, pp. 399–413; 1952.) A general method is described for estimating the field of view of any antenna system, and hence of predicting the ratio of true to observed hourly numbers for particles of different velocity and flight direction. Part 2: 1009 above.

523.72:550.38 1011
Relation between Geomagnetic Activity and Solar Radio Activity—J. F. Denisse. (*Ann. Géophys.*, vol. 8, pp. 55–64; Jan./March, 1952.) A fuller account of work noted previously [2983 of 1951 (Denisse, Steinberg and Zisler)].

550.384.3 1012
The Representation of the Main Geomagnetic Field and of its Secular Variation by Means of Two Eccentric Dipoles—H. G. Macht. (*Trans. Amer. Geophys. Union*, vol. 32, pp. 555–562; Aug., 1951.) A model comprising two dipoles is suggested to account for the second-order terms in the expansion representing the geomagnetic potential. The systematic displacement of the dipoles with time explains certain features of the geomagnetic secular variation. The model provides a generally improved approximation to the actual surface geomagnetic field.

550.385+551.594.5 1013
Theories of Magnetic Storms and Aurorae—J. W. Dungey. (*Nature* (London), vol. 170, p. 795; Nov. 8, 1952.) Discussion of theories advanced by Alfvén (2777 of 1950) and Martyn (1374 of 1951).

551.510.3:535.325:538.56 1014
Slide Rule computes Radio Refractive Index of Air—S. Weintraub. (*Electronics*, vol. 26, pp. 182–198; Jan., 1953.) Details are given of methods developed at the National Bureau of Standards for computing atmospheric r.f. refractive indexes from radiosonde meteorological data, using special types of straight or circular slide-rule. The methods give adequate accuracy for almost any conditions likely to be encountered in the troposphere.

551.510.41:546.214 1015
Ozone in the Earth's Atmosphere—G. M. B. Dobson. (*Endeavour*, vol. 11, pp. 215–219; Oct., 1952.) General discussion of processes involved in the formation of ozone, methods of measuring ozone content and its vertical distribution in the atmosphere, and variations of ozone content with meteorological conditions.

551.510.535 "1952":621.396.11 1016
Ionosphere Review: 1952—T. W. Bennington. (*Wireless World*, vol. 59, pp. 59–60; Feb., 1953.) Curves are given for the period 1947–1952 showing (*a*) monthly mean values, (*b*) 12-month running averages of sunspot numbers and noon and midnight F_2 critical frequencies. From these it appears probable that both sunspot numbers and critical frequencies will continue to decrease slightly during 1953.

551.594.5:551.510.535 1017
Recent Advances in Auroral Spectroscopy and in our Knowledge of the Upper Atmosphere—L. Vegard. (*Ann. Géophys.*, vol. 8, pp. 91–99; Jan./March, 1952.)

551.594.5:621.397.8 1018
Auroral Effects on Television—Thayer. (See 1173.)

551.594.6:621.396.11 1019
The Propagation of a Radio Atmospheric: Part 2—Budden. (See 1108.)

LOCATION AND AIDS TO NAVIGATION

621.396.9:551.594.221 1020
Recent Developments in Radio Location of Thunderstorm Centers—W. J. Kessler and S. P. Hersperger. (*Bull. Amer. Mt. Soc.*, vol. 33, pp. 153–157; April, 1952.) Short account of the triangulation method used by the U. S. air forces during the war, and of the single-station technique being investigated at the University of Florida.

621.396.9:621.396.677 1021
Wide-Angle-Scan Radar Antenna—Chait. (See 938.)

621.396.932 1022
Radar Chart-Matching Devices—J. H. Dickson. (*Jour. Inst. Nav.*, vol. 5, pp. 331–341. Discussion, pp. 341–344; Oct., 1952.) For navigation at sea, chart-matching devices are used enabling a p.p.i. presentation and a chart to be viewed in coincidence; three main classes of device are distinguished, viz., (*a*) reflecting devices without magnification, (*b*) optical instruments involving lenses, and (*c*) projection devices. Descriptions are given of the U. S. Navy virtual-position reflectoscope and of a simplified reflectoscope, both using a 45° semi-reflecting mirror; these have the disadvantage of requiring specially prepared charts. Standard navigational charts can be used with the Admiralty Research Laboratory chart-comparison unit, the construction and performance of which are described in some detail. This has a telescopic optical system and is viewed through an eyepiece, the p.p.i. face being screened from external light. Experimental projection methods using forms of the Schmidt optical system or CR tubes of special construction are also briefly described. For a shorter account see *J. R. Soc. Arts*, vol. 100, pp. 825–830; Oct. 31, 1952.

621.396.932 1023
Navigational Work in Ocean Weather Ships—C. E. N. Frankcom. (*Jour. Inst. Nav.*, vol. 5, pp. 351–361; Oct., 1952.) Equipment carried by the British weather ships includes loran (type DA52), and radar for taking upper-wind observations, for obtaining fixes of aircraft and for navigation. An ordinary radio receiver is used for consol bearings. Navigation aids between ship and aircraft include MF beacon, VHF and MF DF and a eureka responder. Facilities for MF, HF and VHF communications are provided.

621.396.933:656.7 1024
The Use of Radio in the Navigation and Operation of Civil Aircraft—D. H. C. Scholes. (*Jour. Brit. Inst. Radio Eng.*, vol. 12, pp. 595–623; Dec., 1952.) "Systems of navigation and communication for landing, traffic control and en-route communication and navigation are described in sufficient technical detail to enable their functions and capabilities to be appreciated, but the general purpose of the paper is to indicate the contributions made by radio to the solution of operational problems."

621.396.933.088 1025
An Examination of Some Site and Transmission-Path Errors of the Decca Navigator System when used over Land—L. G. Reynolds. (*Proc. IEE* (London), part III, vol. 100, pp. 29–35; Jan., 1953.) The results of measurements of the Decometer errors near obstacles such as trees and telegraph wires are described and discussed. Vertical obstacles were found to show some uniformity as regards their effect, but the effects of long horizontal conductors were very variable. Errors determined at good sites on the base-line extensions of the red and green lattices are shown graphically and exhibit the effects of the near-field components and the finite conductivity of the ground. Good agreement is obtained between the mean of the observations and a composite theoretical curve based on Norton's plane-earth theory and Bremmer's curved-earth theory of ground-wave propagation.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.37:546.482.21 1026
The Electrical Conductivity connected with the Phosphorescence of Cadmium-Sulphide Crystals—I. Broser and R. Warminsky. (*Z. Phys.*, vol. 133, pp. 340–361; Oct. 2, 1952.)

537.311.31:669-124.2 1027
The Effect of Temperature of Deformation on the Electrical Resistivity of Cold-Worked Metals and Alloys—T. Broom. (*Proc. Phys. Soc.*, vol. 65, pp. 871–881; Nov. 1, 1952.) Special apparatus was used to draw wires of various metals at temperatures between −183°C and +100°C, and to measure their resistivities at the temperature of drawing. The increase of resistivity with deformation was found to depend on the difference between the deformation and recrystallization temperatures. A unified theory of the effect of deformation on the resistivity of both pure metals and alloys can possibly be based on stacking faults in the crystal lattice.

537.311.32/.33:546.561.221 1028
Electrical and Optical Properties of Copper Sulphides—L. Eisenmann. (*Ann. Phys., Lpz.*, vol. 10, pp. 129–152; Feb. 15, 1952.) The sulphur content of thin copper-sulphide layers was varied and measurements were made of the resistance at temperatures down to 14°K, and the absorption for wavelengths from 0.4 to 3 μ. The structure of the layers was examined. Cu_2S can take up S until CuS is formed. Cu_2S, the two-phase Cu_2S-$Cu_{1.8}S$ and the compound $Cu_{1.8}S$ have semiconductor properties; CuS has the properties of a metal, its conductivity being comparable with that of Hg. Two alternative interpretations of the observations are discussed.

537.311.33:538.21 1029
Magnetodynamics of Semiconductors—P. M. Prache and H. Billottet. (*Câbles & Transm.*, vol. 6, pp. 317–332; Oct., 1952.) The variation with frequency of the apparent permeability of ferrite cores is very different from that of ferromagnetic metal cores. Experimental work in this field is reviewed and test methods used by the authors for the range 0–40 mc are described. In this frequency range the conductivity can be considered, to a first approximation, as the sum of a constant term and one proportional to the frequency. The permittivity increases with frequency very rapidly at first and then more slowly. A surprising result is the rapid rise of the parameter $\lambda/\omega\mu_0$ (λ being related to surface resistivity), accompanied by a permeability decrease, for frequencies >1 mc. Results of measurements of the variation with temperature of the permeability, conductivity and permittivity of the ferrite samples investigated are presented and discussed.

537.311.33:546.289-1 1030
Resistance of Germanium Contacts—J. B. Gunn. (*Proc. Phys. Soc.*, vol. 65, pp. 908–909; Nov. 1, 1952.) Previous attempts to explain the HV I/V characteristics of Ge point contacts have not taken account of the very high current densities which normally occur near the contact, and which give rise to electric fields far greater than those which Ryder and Shockley (1398 of 1951) found to cause non-linear mobility effects of the type discussed by Shockley (1011 of 1952). A tentative theory is presented which shows that these effects, together with the Zener current observed by McAfee *et al.* (164 of 1952), can account for the observed I/V characteristics, and in particular for the high resistances recently noted in *p-p* Ge-Ge contacts by Granville *et al.* (117 of January).

537.311.33:546.47-31:539.231 1031
Effect of Dissolved Oxygen on the Electrical Conductivity of Zinc Oxide—H. Fritzsche. (*Z. Phys.*, vol. 133, pp. 422–437; Oct. 2, 1952.) ZnO layers with reproducible characteristics are obtained by cathodic sputtering of the metal in an atmosphere containing oxygen. Measurements of the conductivity of such layers in vacuo give results which can be explained by the assumption that dissolved oxygen acts as an electron trap, and at higher temperatures diffuses out of the ZnO lattice.

538.221 1032
Results of Measurements on High-Permeability Ferrite Cores: Part 2—M. Kornetzki, J. Brackmann, J. Frey and W. Gieseke. (*Z. angew. Phys.*, vol. 4, pp. 371–374; Oct., 1952.) Continuation of work noted in 2736 of 1951 (Kornetzki). Measurements of after-effect losses are reported; the loss factor is nearly constant at low frequencies and increases steeply near a frequency related to the gyromagnetic critical frequency. For a given value of permeability there exists an optimum ferrite composition; for Ni-Zn optimum ferrites the loss tangent at low frequencies is about 7%, for Mn-Zn optimum ferrites it is about 2.5%.

538.221:538.24.096 1033
The Temperature Variation of the Magnetization of Nickel in Low and Moderate Fields—R. S. Tebble, J. E. Wood and J. J. Florentin. (*Proc. Phys. Soc.*, vol. 65, pp. 858–871; Nov. 1, 1952.) An account of measurements on the reversible changes of magnetization accompanying change of temperature of annealed Ni, with discussion of the results in relation to the work of Bates (914 of 1951), Stoner (2689 of 1951) and others on the magneto-caloric effect.

538.221:546.72-1 1034
Effect of Oxygen in Solid Solution in High-Purity Iron on Certain Magnetic Properties in Weak Alternating Fields—J. Bourrat, G. Chaudron and I. Épelboin. (*Compt. Rend. Acad. Sci.* (Paris), vol. 235, pp. 1290–1292; Nov. 24, 1952.) Measurements on four samples of iron, of different degrees of chemical purity, show that metallic impurities have a preponderating effect on initial permeability when the quantity of dissolved oxygen is small. This effect decreases rapidly with increase of oxygen content and practically disappears for an oxygen content of about 0.013%.

538.221:621.318.2 1035
A New Permanent Magnet from Powdered Manganese Bismuthide—E. Adams, W. M. Hubbard and A. M. Syeles. (*Jour. Appl. Phys.*, vol. 23, pp. 1207–1211; Nov., 1952.) A method is described for preparing the magnet material, known as "bismanol," by hot-pressing in the presence of a strong magnetic field finely powdered MnBi crystals obtained from a mixture of Mn (16.65%) and Bi (83.35%). Specimens have been obtained with a BH_{max} value as high as 4.3×10^6 gauss-oersted, coercive force of 3400 oersted and remanence of 4300 gauss. In addition to its importance as a substitute for Co alloys, the material is particularly useful for applications where a high coercive force is required.

538.221:621.318.2 1036
Preferred Crystal Orientation in Ferromagnetic Ceramics—A. L. Stuytis, G. W. Rathenau and E. W. Gorter. (*Jour. Appl. Phys.*, vol. 23, p. 1282; Nov., 1952.) The use of $BaFe_{12}O_{19}$ in permanent-magnet materials has been described previously [2824 of 1952 (Went *et al.*)]. By orienting the crystals of this oxide in a magnetic field before sintering, it is possible to obtain a material with a BH_{max} value as high as $3\times10_6$ gauss-oersted.

538.221:669.14.018.58:539.431 1037
Effect of Fatigue on the Magnetic Properties of Steels—A. Kovacs and P. Laurent. (*C. R. Acad. Sci.* (Paris), vol. 235, pp. 1224–1226; Nov. 17, 1952.) Permeability measurements were made at intervals during vibratory fatigue tests of Ni-Cr steels. The results obtained are shown in curves and discussed.

538.652:546.74-1 1038
Magnetostrictive Effect and the Δ-E Effect in Nickel—H. Nödtvedt. (*Nature* (London), vol. 170, pp. 884–885; Nov. 22, 1952.) An account of experiments demonstrating the connection between the two effects.

539.234:546.72 1039
Influence of a Magnetic Field on the Electrical Resistance of Iron Films—B. Franken, A. van Itterbeek, G. J. van den Berg and D. A. Lockhorst. (*Physica*, vol. 18, pp. 771–779; Oct., 1952.) Continuation of work noted in 2828 of 1952 (van Itterbeek *et al.*). It has been previously assumed that the coercive force is given by the value of field strength for which the resistance/magnetizing-field curve shows a turning point. Low-temperature measurements to test the validity of this assumption are reported and discussed.

546.212+621.315.613.1]:537.226.2 1040
Dielectric Constant of Water Films—L. S. Palmer, A. Cunliffe and J. M. Hough. (*Nature*, (London), vol. 170, p. 796; Nov. 8, 1952.) Continuation of the investigations previously reported [2793 of 1952 (Cownie and Palmer) and 99 of January (Palmer)]. Measurements were made at 2 and 3 mc, using thin mica plates separated by films of water; the dielectric constant of the water varied from >20 for films about 5μ thick to <10 for films about 2μ thick. Measurements on composite mica/water blocks at 2 mc and at 2.5 kc indicated that the water films have the properties of "liquid ice" rather than of solid water.

546.26-1 1041
Electrical Resistivity of Artificial Graphite—J. Okada and T. Ikegawa. (*Jour. Appl. Phys.*, vol. 23, pp. 1282–1283; Nov., 1952.) Report of an investigation of the influence on the graphite resistivity of the temperature at which the raw coke is calcined.

546.431.824-31:537.228.1 1042
Electromechanical Properties of $BaTiO_3$ Compositions showing Substantial Shifts in Phase Transition Points—D. A. Berlincourt and F. Kulcsar. (*Jour. Acoust. Soc. Amer.*, vol. 24, pp. 709–713; Nov., 1952.) Measurements of the dielectric and piezoelectric properties over the temperature range −40° to 120°C are reported for $BaTiO_3$ ceramics containing $CaTiO_3$, Y_2O_3 and $PbTiO_3$. The composition containing 5% $CaTiO_3$ combines high piezoelectric response with good temperature stability.

546.431.824-31:621.3.011.5.001.572 1043
Statistical Model of Barium Titanate at Room Temperature—R. Hagedorn. (*Z. Phys.*, vol. 133, no. 3, pp. 394–421; 1952.) A simple model of the structure of a $BaTiO_3$ single crystal is discussed which may serve to explain the properties of the material at ordinary temperatures.

548.0:53:546.391.85 1044
The Piezoelectric, Dielectric, and Elastic Properties of $ND_4D_2PO_4$ (Deuterated ADP)—W. P. Mason and B. T. Matthias. (*Phys. Rev.*, vol. 88, pp. 477–479; Nov. 1, 1952.) Heavy-water ADP crystallizes in the same form as normal ADP but has a transition at −31°C, whereas normal ADP has a transition at −125°C. At temperatures below the transition the crystals are antiferroelectric, with one of the "*a*" crystal axes as the antiferroelectric axis. Measurements were made on the material at temperatures from 80°C down to the transition temperature; results are shown in graphs. The crystal has zero temperature coefficient of frequency at 0°C, and may be useful for transducers and mechanical filters.

548.0:53:546.391.85 1045
Properties of a Tetragonal Antiferroelectric Crystal—W. P. Mason. (*Phys. Rev.*, vol. 88, pp. 480–484; Nov. 1, 1952.) Theory explaining the properties of deuterated ADP is developed in terms of thermodynamics. Observed changes of crystal structure and dielectric constant at the transition temperature are related to the occurrence of spontaneous polarization. Use is made of the results given in 1044 above.

548.0:534/539].001.8 1046
Recent Applications of Synthetic Crystals—J. Chapelle. (*Ann. Télécommun.*, vol. 7, pp. 398–407; Oct., 1952.) Review of applications in microphones, ultrasonic transducers, optical and electrical filters, Kerr cells, etc.

621.314.6:621.396.822 1047
Flicker Effect in Crystal Detectors—N. Nifontoff. (*Comp. Rend. Acad. Sci.* (Paris), vol. 235, pp. 1117–1118; Nov. 10, 1952.) Continuing work previously reported (2827 of 1952 and back references), results are now given for the resistance variations and flicker effect in Si and Ge crystal rectifiers as functions of the current in the forward and reverse directions. Similar curves were obtained in all cases. The results were identical for increasing or decreasing current except for Ge, which exhibits a little hysteresis.

621.315.612.4 1048
A New Dielectric Material—L. Nicolini. (*Nature* (London), vol. 170, p. 938; Nov. 29, 1952.) Short note on a material prepared by heating pure TiO_2 to 1400°C. X-ray analysis gives the same Debye-Scherrer pattern as for rutile, with an axis ratio 1:0.911. The dielectric constant varies from over 10^4 at 10 cps to an asymptotic value of about 100 for frequencies above about 100 kc. Tan δ has a pronounced maximum at about 30 kc.

MATHEMATICS

681.142 1049
The C.S.I.R.O. Differential Analyser—D. M. Myers and W. R. Blunden. (*Jour. Inst. Eng.* (Australia), vol. 24, pp. 195–204; Oct./Nov., 1952.) Description of the analyser installed by the Commonwealth Scientific and Industrial Research Organization in the University of Sydney, and of its use in solving the differential equations relating to various technical problems.

681.142 1050
An Electronic Computer—M. Beard and T. Pearcey. (*Jour. Sci. Instr.*, vol. 29, pp. 305–311; Oct., 1952.) Description of the C.S.I.R.O. Mark I binary digital computer. Mercury-filled ultrasonic delay tubes are used for the main high-speed store and a magnetic drum for the low-speed intermediate store; the total capacity is 40 960 binary digits. The command code specifies a source and a destination. Programmes are placed in the main store and commands are adopted sequentially at a rate of 500 per sec. Manual controls provide for variation of operation.

681.142 1051
Multiplication in the Manchester University High-Speed Digital Computor—A. A. Robinson. (*Electronic Eng.*, vol. 25, pp. 6–10; Jan., 1953.)

681.142:003.62 1052
On Optimum Relations between Circuit Elements and Logical Symbols in the Design of Electronic Calculators—A. D. Booth. (*Jour. Brit. Inst. Radio Eng.*, vol. 12, pp. 587–594; Dec., 1952.) The functions of the main units of a high-speed calculator are discussed, possible means of representing typical units by logical symbols are considered, and methods by which such symbols can be transformed into engineering details are examined. A logical notation for computer elements is suggested which is such that a small number of basic standard "building blocks" can be combined to form units of any desired functional complexity that will operate in a reliable manner.

681.142:621.315.612.4 1053
Ferroelectric Storage Elements for Digital Computers and Switching Systems—J. R. Anderson. (*Elec. Eng.* (N. Y.), vol. 71, pp. 916–922; Oct., 1952.) Revised text of paper presented at the A.I.E.E. Fall General Meeting October 1952. The advantages of ferroelectric materials, $BaTiO_3$ in particular, for information storage are discussed and practical storage devices are described which are capable of storing 2500 "bits" of information per square inch on the surface of a material only a few thousandths of an inch thick, using pulses of duration <1 μs.

681.142:621-526 1054
The Design and Testing of an Electronic Simulator for a Hydraulic Remote-Position-Control Servomechanism—F. J. U. Ritson and P. H. Hammond. (*Proc. IEE* (London), Part II, vol. 99, pp. 533–548. Discussion, pp. 549–552; Dec., 1952.)

681.142:621-526 1055
An Analogue Computer for Use in the Design of Servo Systems—E. E. Ward. (*Proc. IEE* (London), Part II, vol. 99, pp. 521–532. Discussion, pp. 549–552; Dec., 1952.)

681.142:621.396.645 1056
A Source of Computing Voltage with Continuously Variable Output—R. W. Williams and G. M. Parker. (*Jour. Sci. Instr.*, vol. 29, pp. 322–324; Oct., 1952.) For use as an ac voltage source in an analogue computer, the resistance potentiometer has the disadvantage that its output does not vary smoothly. As an alternative in which this difficulty is overcome, a circuit is described using a magslip feedback resolver [235 of 1952 (Bell)].

MEASUREMENTS AND TEST GEAR

536.632:537.311.33 1057
Sensitive Recording Alternating-Current Hall-Effect Apparatus—N. M. Pell and R. L. Sproull. (*Rev. Sci. Instr.*, vol. 23, pp. 548–552; Oct., 1952.) Apparatus is described capable of measuring mobilities of 10^{-2} cm/sec per V/cm or less in semiconductors; it is especially suitable for measuring Hall effect in samples of very low conductivity. Independent units are used for the magnetic field, current source and detector circuit. An amplidyne control circuit is used for regulating the magnet current and reducing the time required to reverse the magnetic field.

621.3.018.41(083.74):621.317.361:529.77 1058
The Estimation of Absolute Frequency in 1950–51—H. M. Smith. (*Proc. IEE* (London), part IV, vol. 99, pp. 273–278; Dec., 1952.) Full paper. See 153 of January.

621.317(083.74)+537.71 1059
Units and Standards of Electrical Measurement—A Review of Progress—L. Hartshorn. (*Proc. IEE* (London), part I, vol. 99, pp. 271–279; Nov., 1952.) The functions of the various international organizations concerned with electrical standards are noted and an account is given of their work since 1942 and of the consequent redefinition of British legal standards. Experimental work has shown that over a 10-year period standard manganin resistors are stable to within ±1 part in 10^6, while Weston cells may vary by as little as ±5 parts in 10^6, with capacitors and inductors showing relatively greater drift. The precision and convenience of Wenner's method for the absolute determination of the ohm is pointed out. Progress in frequency standards is noted, and consideration given to the possible choice of the magnetic moment of the proton, or other atomic constant, as the primary electrical unit.

621.317.3:621.396.615.141.2:621.396.619.11 1060
Measurements of Some Operational Characteristics of an Amplitude-Modulated Injection-Locked U.H.F. Magnetron Transmitter—L. L. Koros. (Proc. I.R.E., vol. 41, pp. 4–10; Jan., 1953.) The loop impedance looking towards the magnetron, which is important for determining the position of the injection junction, is computed from the voltage SWR and the position of the minimum-voltage plane of the injection current reflected from the magnetron output loop. The loop-impedance changes are observed as the anode voltage is varied. The phenomena occurring in the output transmission line are described, loads of various types being used. The load impedance as seen from the magnetron varies with the power level during the modulation cycle; this is shown on a special circular diagram. At the low-power end of the modulation cycle the value of the load, for a given type of magnetron and setting of tuner, depends only on the carrier frequency.

621.317.328:621.396.615(083.74) 1061
Signal-Generator System for Low Output Levels—J. W. Herbstreit. (*Electronics*, vol. 26, pp. 218–224; Jan., 1953.) For measuring signal levels far below that useful for communications, a method similar to that described by Gainsborough (3202 of 1947) was adopted in which two standard signal generators were used with a crystal mixer, the desired frequency being either the sum or difference frequency of the two generators. Details are given of the procedure, and other applications of the heterodyne principle are suggested.

621.317.332 1062
A Modified Wide-Range Shunted-T Circuit for the Measurement of Impedance in the AF, RF and VHF Ranges—D. Karo. (*Proc. IEE* (London), part III, vol. 100, pp. 25–28; Jan., 1953.) The T circuit is connected across the lower-voltage section of a voltage divider connected across the source, the shunt also being across the source. A variable multiplication ratio is thus obtained and hence a much greater range for a given set of standards. Tests are described at frequencies from 1 kc to 50 mc. By using an additional standard the frequency can be eliminated from the balance equations.

621.317.335.029.64 1063
Dielectric Constants of Gases in the Microwave Region—A. Gozzini. (*Nuovo Cim.*, vol. 8, pp. 361–368; June 1, 1951.) A method of measurement is described which makes use of reflections in mismatched waveguides. Results are given for H_2, air and CO_2.

621.317.34:621.397.2 1064
Apparatus for the Measurement of Non-linearity in Television Trunk Systems—G. W. S. Griffith. (*Proc. IEE* (London), part IIIA, vol. 99, pp. 394–397. Discussion, pp. 472–478; April/May, 1952.) Defects of methods of measurement using sawtooth or stepped waveforms are noted and two methods are described which avoid such difficulties. In the 2-frequency method used for the London-Birmingham television radio-relay link, a 250-cps signal, with an amplitude covering almost the whole amplitude characteristic of the system, is transmitted together with a 50-kc signal whose amplitude is about one-fiftieth of that of the 250-cps signal. The small signal is thus made to examine, 250 times per second, the whole amplitude characteristic of the transmission system. At the receiving end the signals are separated and used respectively for the horizontal and vertical deflection of a CRO beam. For a perfectly linear system a rectangular display pattern is obtained. Use of a clamp circuit for the vertical (50 kc) deflection results in a pattern with a horizontal base line and with ordinates representing the slope of the amplitude characteristic of the system under test. For investigation of the effect of the dc component in a television waveform, a test signal with waveform essentially of the television type must be used. Such a waveform can also be used to measure synchronization-pulse compression.

621.317.34:621.397.6 1065
The Transient Testing of Television Apparatus—V. J. Cooper. (*Marconi Rev.*, vol. 16, pp. 1–7; 1st Quarter, 1953.) The relative merits of spike and step waveforms for estimating the response characteristics of television equipment are briefly considered. A limited-spectrum test waveform approximately represented by a sine-squared step function has been found amenable to mathematical analysis. Results thus obtained by Skwirzynski (1066 below) are summarized.

621.317.34:621.397.6 1066
The Response of a Vestigial Side-Band System to a "Sine-Squared" Step Transition—J. K. Skwirzynski. (*Marconi Rev.*, vol. 16, pp. 8–24; 1st Quarter, 1953.) The transient response to a "sine-squared" step function is determined for a transmission system which can be represented by a triply tuned circuit with a maximally flat admittance function and arbitrarily placed carrier frequency. The problem is solved by means of the zero-frequency carrier method due to Wheeler (3297 of 1941). Response curves for various transition times of the sine-squared step function are shown for (*a*) the double-sideband system with 100% modulation, (*b*) the vestigial-sideband system with 100% modulation, (*c*) the vestigial-sideband system with modulation from 30% to 100% of the carrier level. The results obtained are discussed; they can be used to estimate the change of transient distortion resulting from slight changes in the position of the carrier frequency in relation to the filter characteristics (EG temperature drift, etc.).

621.317.35:519.272.1 1067
A Microwave Correlator—R. M. Page, A. Brodzinsky and R. R. Zirm. (Proc. I.R.E., vol. 41, pp. 128–131; Jan. 1953.) The spectrum of a wide-band UHF signal is converted to a form more convenient for correlation computations by a method which involves local generation of a line spectrum and summation of the resulting frequency differences into a narrower band while retaining the original correlative characteristics. Analysis is presented together with experimental results obtained on a noise signal of bandwidth 200 mc centered at 1.1 kmc.

621.317.352:621.392.26 1068
Experimental Study and First Results of Measurements of the Attenuation of TE_{01} (H_{01}) Waves in Short Sections of Straight Circular Waveguide—G. Comte and A. Ponthus. (*Câbles & Transm.*, vol. 6, pp. 333–352; Oct., 1952.) The propagation of TE_{01} waves in an infinitely long circular waveguide and their reflection at a conducting piston are analyzed, and resonance phenomena in a section limited by two conducting pistons are discussed. Measurements of the propagation constants, in particular the attenuation, are described which are based on such resonance effects. The attenuation produced by coupling a receiver to the waveguide is calculated. The construction of a differential wavemeter for wavelengths around 3 cm is described, and attenuation and *Q*-factor measurements are reported in detail for the following waveguide materials: (*a*) red Cu, Cu polished with emery, Cu electrolytically polished, (*b*) Al alloy polished with emery, (*c*) mild steel, well polished after turning.

621.317.361 1069
Precision Frequency Measurements—K. Gosslau. (*Frequenz*, vol. 6, pp. 249–255; Sept., 1952.) Precision methods depend on a comparison between the unknown frequency and a standard. A standard-frequency quartz-oscillator circuit is described in which the crystal is only lightly loaded; this is incorporated in an equipment which covers the range 500 cps–70 mc, making use of harmonics. The relative uncertainty is ±1 part in 10^8 and the absolute uncertainty ±0.2 cps. Operation is simple compared with other instruments of similar precision.

621.317.37 1070
The Measurement of High Standing-Wave Ratios—T. J. Buchanan. (*Proc. IEE* (London), part IV, vol. 99, pp. 372–375; Dec., 1952.) The rate of change of phase at a minimum in a standing-wave pattern increases as the SWR increases, becoming infinite in the case of a full standing wave. The apparatus described enables measurements by a phase and amplitude balance method to be made of the phase difference between two points close to and equidistant from a minimum. A comparison is made with the Roberts-von Hippel method and a modification of this method is described.

621.317.382.029.6 1071
A General Method for the Absolute Measurement of Microwave Power—A. L. Cullen. (*Proc. IEE* (London), part IV, vol. 99, pp. 429–430; Dec., 1952.) Discussion on 2850 of 1952.

621.317.444.029.64:538.614 1072
A Microwave Magnetometer—P. J. Allen. (PROC. I.R.E., vol. 41, pp. 100–104; Jan. 1953.) The phenomenon of Faraday rotation in ferrites is used to obtain a magnetometer of high sensitivity giving a voltage output proportional to the applied magnetic field. Dominant-mode rectangular waveguides, aligned axially and with crossed electric planes, are used as polarizer and analyzer; the rotator is a ferrite cylinder enclosed in a circular waveguide. Response curves obtained with rotators of two different lengths are shown. Variations as small as 1 gamma have been detected.

621.392.43:621.317.7.029.63/.64 1073
A U.H.F. and Microwave Matching Termination—R. C. Ellenwood and W. E. Ryan. (PROC. I.R.E., vol. 41, pp. 104–107; Jan., 1953.) The unit comprises two low-loss dielectric slugs and a lossy dielectric load which are slid inside a waveguide or coaxial line by means of bakelite rods parallel to the axis. Matching is accomplished by altering the distances between the dielectrics. Wide frequency ranges are covered.

621.317.7.088 1074
The Error of a Measurement Instrument, and Its Probability—W. Hasselbeck. (*Funk u. Ton*, vol. 6, pp. 400–405; Aug., 1952.) The greater the number of possible causes of error, the less is the probability of their superposition to produce the maximum possible error. Known statistical methods are used to calculate the probability of an error of given magnitude.

621.317.714.029.62/63 1075
Calibrating Ammeters above 100 mc—H. R. Meahl and C. C. Allen. (PROC. I.R.E., vol. 41, pp. 152–159; Jan., 1953.) Methods of measuring currents at frequencies >100 mc are reviewed, types of vacuum thermocouple available for low-current work are described, and methods of calibration are compared. The electrodynamic method (2832 of 1950) is suitable for large currents, the calorimeter method for medium currents, and the thermistor-bridge for small currents. Attention is drawn to the importance of obtaining agreement between at least two methods based on different principles.

621.317.74:621.397.2 1076
Group-Delay Distortion-Measuring Equipment—G. J. Hunt and L. G. Kemp. (*Proc. IEE* (London), part IIIA, vol. 99, pp. 411–414. Discussion, pp. 472–478; April/May, 1952.) Equipment is described for measuring changes in the slope of the phase/frequency characteristic of a transmission system in the range 200 kc–5 mc. A continuously variable am test frequency is used, the sidebands serving for measurement of the slope of the phase-distortion characteristic relative to the slope at a datum frequency. Direct readings of group-delay distortion are given by a CRO display calibrated in intervals of 10 mμs.

621.317.755 1077
The Presentation of Frequency Markers and Frequency Scales for Visible-Indication Instruments—A. Scholz and G. Stuwe. (*Funk u. Ton*, vol. 6, pp. 470–480; Sept., 1952.) A description is given of a multipurpose CRO using a sine-wave deflection voltage with blanking of the return stroke. The frequency scale is provided by dark spots on a line parallel to that on which the required curve is based; a single-beam tube is used with beam switching.

621.317.757 1078
Theory and Construction of a Harmonic-Distortion Meter—G. E. Jones, Jr. (*Audio Eng.*, vol. 36, pp. 22–23, 79; Nov. 1952.) Description, with circuit details, of a relatively simple meter for measuring harmonics of fundamental frequencies in the range 16 cps–20 kc. The instrument is only suitable for measuring harmonics with amplitudes $\nless 1\%$ of that of the fundamental.

621.317.761.029.424/.45 1079
Reed-Type Frequency Meter—H. Behrmann. (*Z. Ver. Dtsch. Ing.*, vol. 94, pp. 992–994; Oct. 2, 1952.) Details are given of a meter using a reed of variable length for the measurement of the frequency of mechanical vibrations or electrical oscillations in the ranges 5–45 and 30–270 cps. For measurements of ac frequency a small electromagnet is used to excite the reed.

621.317.772 1080
A Precision Phase Comparator for Use at Low Radio Frequencies—B. G. Pressey, C. S. Fowler and R. W. Mason. (*Proc. IEE* (London), part IV, vol. 99, pp. 318–325; Dec. 1952.) Full paper. See 483 of February.

621.317.784:621.319 1081
The Design and Application of a Portable Electrostatic Wattmeter—F. R. Axworthy and J. K. Choudhury. (*Trans. Soc. Instr. Technol.*, vol. 4, pp. 57–63. Discussion, pp. 64–66; June 1952.) Description of an instrument of the quadrant-electrometer type, using a mica vane coated with Al by evaporation and suspended by a phosphor-bronze strip. Methods of eliminating the error due to power loss in the shunt are described and applications of the wattmeter to the measurement of iron losses are discussed, with illustrative examples.

621.317.784.029.64 1082
A Torque-Operated Wattmeter for 3-cm Microwaves—A. L. Cullen and I. M. Stephenson. (*Proc. IEE* (London), part IV, vol. 99, no. 4, pp. 294–301; Dec. 1952. Digest, *ibid.*, part II, vol. 99, pp. 516–517; Oct. 1952.) Power measurements are made in a waveguide by observation of the torque exerted by the em field in the waveguide on a small vane suspended by a quartz fiber, the vane deflection being measured optically. Pulsed microwave power in the range 10–60 W can be measured to within about $\pm 1.5\%$. Theory of a standing-wave method of calibration has previously been given [1052 and 2850 of 1952 (Cullen)].

621.396.611.21(083.74) 1083
High-Stability 100-kc Crystal Units for Frequency Standards—J. P. Griffin. (*Bell Lab. Rec.*, vol. 30, pp. 433–437; Nov. 1952.) Short account of the development of the latest type of wire-supported quartz-crystal units, such as those used for the loran system of the U. S. Navy.

621.396.615.029.55/.63(083.74) 1084
Three Generators of Standard Signals in the Frequency Band 20–1200 mc—P. Herreng, G. Couanault and G. Plottin. (*Câbles & Transm.*, vol. 6, pp. 353–367; Oct. 1952.) A detailed description is given of generators with ranges of 20–250 mc, 150–500 mc and 500–1200 mc respectively. The general characteristics are the same for all three, but different techniques are used in the oscillatory circuits, attenuator networks, modulators, etc. Schematic circuit diagrams and photographs of the complete instruments are given.

621.396.615.14:535.325:546.217 1085
A Highly Stable Microwave Oscillator and its Application to the Measurement of the Spatial Variations of Refractive Index in the Atmosphere—L. Essen. (*Proc. IEE* (London), part III, vol. 100, pp. 19–24; Jan. 1953.) For frequencies >3 kmc the type of oscillator described by Pound (2865 of 1948, in which please read "Volume 11" instead of "Volume 2") is particularly suitable. An oscillator of this type, using an invar cavity resonator as the control element, is described and its performance discussed. The factors affecting frequency stability are analysed. It is possible to obtain a bandwidth of 1 part in 10^{10} at a frequency of 9.2 kmc, with a stability to within 1 part in 10^8 per hour. A heterodyne method was used to demonstrate the variation of the refractive index of air drawn through the resonator. Erratic variations were observed when the indrawn air passed over a dish of water, thus changing its moisture content. Variations of 1 part in 10^6 were found fairly general, even inside a laboratory.

621.396.615.16/.17].015.7 1086
High-Power Square-Pulse Generator—W. E. Williams, Jr. (*Electronics*, vol. 25, pp. 144–145; Oct. 1952.) Short description, with detailed circuit diagram, of a generator providing current pulses up to several amperes at voltages adjustable up to 1.2 kV. The pulse recurrence frequency is variable in steps of 10 from 10 to 60/sec, the duty cycle being constant at 1%. Designed primarily for testing cathode emission under pulsed operation, the generator may have other applications. Overshoot at the beginning and end of each pulse is negligible. See also *Tech. News Bull. Nat. Bur. Stand.*, vol. 36, p. 181; Dec. 1952.

621.396.615.17:621.317.34:621.397.2 1087
A Waveform Generator and Display Unit for the Testing of a Television Channel—P. E. Ackland-Snow and G. A. Gledhill. (*Proc. IEE* (London), part IIIA, vol. 99, pp. 332–337. Discussion, pp. 472–478; April/May 1952.) Description, with block diagrams, of a generator providing six different television-type waveforms for testing a communication channel, together with a CR display unit for examination of the channel output waveform.

621.396.615.17.015.33.027.4 1088
Generators of Repeated Surge Waves—M. Teissié-Solier and J. Lagasse. (*Rev. Gén. Élect.*, vol. 61, pp. 425–429; Oct. 1952.) Description of a generator suitable for tests on electrical material, particularly transformers, and furnishing 200–400-V steep-fronted pulses with repetition frequency equal to the mains frequency and with variable rise time. Various applications are suggested.

621.396.615.17.018.75:621.397.62 1089
A Testing Pulse for Television Links—I. F. Macdiarmid. (*Proc. IEE* (London), part IIIA, vol. 99, pp. 436–444. Discussion, pp. 472–478; April/May 1952.) The advantages of the "sine-squared" pulse waveform for testing the characteristics of a television link are discussed and a pulse generator is described which produces such pulses having a half-amplitude width of 0.17 μs, the normal repetition frequency being 10^4/sec. The very short pulses required for driving the generator are produced by a triggered blocking oscillator and have an amplitude of about 40 V, with a half-amplitude width of 0.035 μs. The circuits described are stable and easily reproducible.

621.396.645.35:621.314.5 1090
High-Speed DC Amplifier—F. Y. Masson. (*Elect. Mfg.*, vol. 47, pp. 118, 120; May 1951.) Description of an instrument using a moving-coil galvanometer, with 200-kc alternating field superposed on a permanent-magnet field, for converting dc to ac for high-gain amplification. An energy gain of 10^8 is achieved and the amplifier develops sufficient power to operate a dc moving-coil recorder. Response time is <0.1

sec, so that the instrument is very suitable for rapid production testing of components such as resistors.

621.396.65.001.4:621.396.11.029.64 1091

A Technique for 4000-Mc Propagation Testing for Radio-Relay Systems—W. J. Bray and R. L. Corke. (*Proc. IEE* (London), part IIIA, vol. 99, no. 18, pp. 281–289. Discussion, pp. 310–312; 1952.) Simple technique is described for measurement of the path attenuation of radio-relay systems operating at frequencies of the order of 4 kmc. A coaxial-line oscillator, with AM at 1 kc, is used as transmitter. The receiver comprises a Si-crystal detector followed by a 1-kc high-gain narrow-band amplifier and diode rectifier operating a meter. With a transmitter output of 1W, path attenuations up to 95 db can be measured to within 0.5 db. A method of measuring the path attenuation is described. The results of tests on typical paths over land and over sea are shown graphically.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

526.956:534.321.9:538.652 1092

Echo Sounding Equipment with New Indicating Instrument—W. Krzikalla. (*Arch. elekt. Übertragung*, vol. 6, pp. 473–477; Nov., 1952.) Description of magnetostrictive equipment operated at a frequency of 30 kc by a pulse of current from a capacitor discharge. Depth measurement is effected by measurement of the time interval between the transmitted ultrasonic pulse and the echo, using a relaxation-circuit timer with a direct-reading indicator, which can be compensated for measurements in either sea or fresh water.

538.569.2.047 1093

Possible Industrial Hazards in the Use of Microwave Radiation—H. M. Hines and J. E. Randall. (*Elect. Eng. N. Y.*, vol. 71, pp. 879–881; Oct., 1952.) Revised text of paper presented at the A.I.E.E. Summer General Meeting, June 1952.

621–578:538.3 1094

A "Loudspeaker" Clutch—(*Tech. News Bull. Nat. Bur. Stand.*, vol. 36, pp. 161–163; Nov., 1952.) A fast-acting clutch developed by J. Rabinow is actuated by passing dc through a coil located in a constant magnetic field. Possible uses in magnetic recorders, electronic computers, etc. are indicated.

621.314.3.001.8† 1095

Industrial Applications of Magnetic Amplifiers—R. W. Moore. (*Elect. Eng. N. Y.*, vol. 71, pp. 912–916; Oct., 1952.) Typical examples of the use of Westinghouse "Magamps" are described.

621.316.71:621.383:67 1096

Photoelectric Register-Control Devices—R. Kretzmann. (*Electronic Appl. Bull.*, vol. 13, pp. 81–87; June, 1952.) Description of the application of photocell equipment for register control (*a*) in a wrapping machine, (*b*) of the position of a strip of paper or fabric passing through a processing machine.

621.317.083.7:621.394.324 1097

Multipoint Telemetering System using Teletype Transmission—A. J. Hornfeck and G. R. Markow. (*Elect. Eng. N. Y.*, vol. 71, pp. 929–935; Oct., 1952.) Essentially full text of paper presented at the A.I.E.E. Summer General Meeting, June 1952. Description of a system developed for remote monitoring of pumping stations on a pipe line, to permit automatic operation from a central office. The measured variable quantities are converted to pulses of proportionate duration, and the teletype equipment prints a succession of identical letters (coded for each variable), whose number indicates the measure of the particular variable in question.

621.383.001.8:621.386 1098

Electronic Intensification of Fluoroscopic Images—M. C. Teves and T. Tol. (*Philips Tech. Rev.*, vol. 14, pp. 33–43; Aug., 1952.) The intensifier unit consists essentially of an evacuated tube having a fluoroscopic screen mounted in contact with a photocathode of suitably chosen material. A potential of 25 kV applied between the photocathode and the perforated anode produces an image about 1000 times brighter on a viewing screen of linear dimensions 1/9 those of the fluoroscopic screen. This image is optically enlarged to its original size.

621.384.621.1 1099

A Versatile Focusing System for Van de Graaff Accelerating Tubes—P. Howard-Flanders. (*Nature* (London), vol. 170, pp. 744–745; Nov. 1, 1952.) By means of a resistor-type voltage divider, the ES field is kept low in a short initial section of the accelerating tube and is maintained at a uniform high value throughout the remaining length. This arrangement acts like a lens situated at the point where the field changes. Suitable adjustment of the potentials of the accelerating electrodes brings injected particles to a focus on the target at the end of the tube. Good focusing has been obtained over a wide range of voltage and current for both electrons and positive ions.

621.385.833 1100

Calculation of Spherical Aberration for the Electrostatic Electron Lens—D. W. Shipley. (*Sylvania Technologist*, vol. 5, pp. 87–93; Oct., 1952.) Methods of successive approximation and numerical integration are applied to check electrolyte-tank measurements of the spherical aberration of an ES unipotential lens consisting of three coaxial cylinders.

621.385.833 1101

Spherical Aberration of Grid-Type Electron Lenses—M. Bernard. (*Compt. Rend. Acad. Sci.* (Paris), vol. 235, pp. 1115–1117; Nov. 10, 1952.) Analysis for the case where the grid can be regarded as a continuous membrane, transparent to electrons.

621.385.833 1102

Calculation of the Potential Distribution on the Axis of an Electrostatic Immersion Objective with Thick Electrodes—A. Septier. (*Compt. Rend. Acad. Sci.* (Paris), vol. 235, pp. 1203–1206; Nov. 17, 1952.)

621.387.4:535.37 1103

The Phosphor-Phototube Radiation Detector—D. P. Cole, P. A. Duffy, M. E. Hayes, W. S. Lusby and E. L. Webb. (*Elect. Eng., N. Y.*, vol. 71, pp. 935–939; Oct., 1952.) Description of the development of a portable photocell-phosphor combination Type AN/PDR-18, which uses the average current from the photocell to indicate radiation intensities from about 0.05 to 500 röntgen/hr in four linear-scale decades.

PROPAGATION OF WAVES

538.566.3:535.51 1104

The Theory of the Limiting Polarization of Radio Waves Reflected from the Ionosphere—K. G. Budden. (*Proc. Roy. Soc. A*, vol. 215, pp. 215–233; Nov. 25, 1952.) The coupling between the ordinary and extraordinary waves which occurs in the lower part of the ionosphere is discussed. There is a "limiting" region where a downcoming characteristic wave acquires the limiting polarization observed at ground level. Booker (3277 of 1936) has given an approximate specification for the level of the limiting region. A more exact specification is here given and a method is developed for calculating the limiting polarization of a downcoming characteristic wave. The theory is based on Försterling's coupled-wave equations (1884 of 1942), which apply only to vertical incidence. These equations contain a coupling parameter Ψ which depends on the gradients of electron density and collision frequency. The level of the limiting region is specified in terms of Ψ and the refractive indexes for the characteristic waves. For frequencies greater than about 1 Mc the limiting polarization, in the case of a specific model of the ionosphere, is that given by the magneto-ionic theory for a certain limiting point at a definite height. This height, in practical cases, corresponds to low values of electron density and collision frequency, so that at high frequencies the limiting polarization is determined only by the magnitude and direction of the geomagnetic field in the ionosphere.

621.396.11+621.396.812.3 1105

Waves and Fluctuations—E. C. S. Megaw. (*Proc. IEE* (London), part III, vol. 100, pp. 1–8; Jan., 1953.) Chairman's address to the Radio Section of the I.E.E., discussing the effects of turbulence and fluctuations of refractive index on the transmission of waves through the earth's atmosphere, wave propagation through slightly inhomogeneous media, intensity fluctuation or scintillation, and long-range propagation of radio waves by atmospheric scattering.

621.396.11 1106

Toward a Theory of Reflection by a Rough Surface—W. S. Ament. (PROC. I.R.E., vol. 41, pp. 142–146; Jan., 1953.) The reflection of EM waves by the surface of the sea is investigated theoretically by combining a statistical description of the surface with Maxwell's equations. Formulas are thus derived from which the average specular-reflection coefficient can be determined from the average surface currents when methods of evaluating certain integrals have been found.

621.396.11 1107

Can Prediction be Simplified?—R. Gea Sacasa. (*Rev. Telecomunicación* (Madrid), vol. 8, pp. 6–21; Sept., 1952.) N.B.S. prediction procedure, described in Ebert's paper (1476 of 1951) on the reception in Switzerland of signals from WWV, is discussed in detail and compared with the "Spanish method" (1404 of 1952). Graphs are given showing predicted optimum working frequency against hour of day for the two methods. The "Spanish method" gives results in good agreement with observations, and is much simpler than the N.B.S. method.

621.396.11:551.594.6 1108

The Propagation of a Radio Atmospheric: Part 2—K. G. Budden. (*Phil. Mag.*, vol. 43, pp. 1179–1200; Nov., 1952.) In part 1 (1652 of 1951) one of the modes of propagation of atmospherics considered was by means of a duct formed between the ionosphere and the earth, for which a theory was outlined. Part 2 gives the full mathematical theory and shows how the characteristics of the waveguide modes can be determined in the most general case, provided the reflecting properties of the earth and the ionosphere are known as functions of the angle of incidence. The waveguide modes are found to be of two types, which are termed quasitransverse magnetic and quasitransverse electric. The particular case is discussed in which the earth is perfectly conducting and the ionosphere a homogeneous ionized medium. The results of numerical calculations for a few special cases are presented in curves which show (*a*) the attenuation in the various modes as a function of frequency, (*b*) the polarization characteristics of the wave in typical modes, (*c*) the amplitudes of the waves excited in typical modes by a vertical electric-dipole source.

621.396.11:621.397.8 1109

A Survey of British Research on Wave Propagation with Particular Reference to Television—R. L. Smith-Rose. (*Proc. IEE* (London), part IIIA, vol. 99, no. 18, pp. 270–280. Discussion, pp. 310–312; 1952.) Theoretical and experimental investigations of EM wave propagation over distances up to about 100 km, and the effects of terrain irregularities, atmospheric refraction, diffraction and scattering, are discussed. Investigations for the meter waveband for distances of a few hundred kilometers show the preponderating effect of meteorological conditions. Apart from the use of radio links at frequencies in the region of 900 mc, comparatively little has been done regarding the study of propagation in the 470–960 mc television band, but at frequencies between 3 and 10 kmc considerable work has been done in connection with radar developments. It seems likely that the use of such high frequencies in a television service will be confined to radio links for point-to-point distribution. 55 references.

621.396.11:621.397.8 1110

Long-Distance Propagation in Relation to Television in the United Kingdom.—J. A. Saxton. (*Proc. IEE* (London), part IIIA, vol. 99, no. 18, pp. 294–299. Discussion, pp. 310–312; 1952.) A review is given of the effects of irregularities in the terrain and of non-standard refraction in the troposphere on radio field-strength characteristics in the VHF band. The results of an analysis of existing data on long-range transmission in this band in the United Kingdom are then applied to the problem of determining the spacing required between two transmitters working on a common frequency when the degree of interference in each local service area caused by signals from the more distant transmitter must not exceed various specified limits. The effects of non-standard tropospheric refraction, in particular, are such as to make this spacing much greater than it would be if such departures from standard did not exist.

621.396.8 1111

Transmission Loss in Radio Propagation—K. A. Norton. (Proc. I.R.E., vol. 41, pp. 146–152; Jan., 1953.) "Transmission loss" is defined as the ratio of radiated power to signal power available from a loss-free receiving antenna, thus including only losses occurring in the antennas and the intervening medium. The measurement and calculation of this quantity are discussed. The time variations to be expected in cases of ionospheric or tropospheric propagation are investigated with reference to Rayleigh's distribution theory of random vibrations. The transmission-loss concept is useful for estimating the maximum effective range of a radio system subject to interference from both noise and unwanted signals.

621.396.81 1112

Prediction of the Nocturnal Duct and its Effect on UHF—L. J. Anderson and E. E. Gossard. (Proc. I.R.E., vol. 41, pp. 136–139; Jan., 1953.) Predictions, based on meteorological data covering 3½ years, are made of the probability distribution of diurnal variations in field strength for two 98-mile CRPL links in Colorado, operating respectively on 100 mc and 1 kmc. Agreement between predicted and measured values is encouraging.

621.396.81 1113

Field Strength of KC2XAK, 534.75 Mc/s, recorded at Riverhead, N. Y.—G. S. Wickizer. (Proc. I.R.E., vol. 41, pp. 140–142; Jan., 1953.) Analysis of field strengths recorded during 22 months for transmission over a nonoptical path of length 33 miles. Over-all variation was about 12 db in winter and 33 db in summer; variation of the median level was relatively small. Fading 10 db or more below the median level occurred during summer.

621.396.81.029.62/64 1114

Radio Transmission Beyond the Horizon in the 40- to 4000-Mc/s Band—K. Bullington. (Proc. I.R.E., vol. 41, pp. 132–135; Jan., 1953.) Tests using frequencies of 3.7 kmc, 535 mc and 460 mc are reported; reliable signals have been received at distances of several hundred miles. Median signal levels are 50–90 db below the free-space values but are hundreds of decibels higher than values predicted from classical theory. At points far beyond the horizon the received power is relatively independent of antenna height, meteorological factors and frequency over the range 40 mc–3.7 kmc. No long-delay echoes have been observed. A shorter account was noted in 3216 of 1952.

RECEPTION

621.396.621:517.432.1 1115

Theory of the Impulse Response of Receivers—R. Kitai. (*Proc. IEE* (London), part IV, vol. 99, pp. 279–288. Discussion, p. 430; Dec., 1952.) Full paper. See 2879 of 1952.

621.396.82+621.397.82 1116

Current Radio-Interference Problems—Lee. (See 1179.)

621.396.82:621.315.14.027.8 1117

Radio Noise in Relation to the Design of High-Voltage Transmission Lines—Rorden and Gens. (See 927.)

STATIONS AND COMMUNICATION SYSTEMS

621.39:358.236(494) 1118

The Evolution of Electrical Communication in the [Swiss] Army—M. Nüscheler. (*Bull. schweiz. elektrotech. Ver.*, vol. 43, pp. 820–824; Oct. 4, 1952.) History of developments since 1852.

621.39(689) 1119

Telecommunications in Nyasaland—C. R. Dickenson. (*Overseas Eng.*, vol. 26, pp. 76–77; Oct., 1952.) An outline description of present facilities and proposed developments, including multichannel R/T links operating in the VHF bands.

621.395+621.396.5]:061.31 1120

The 16th Plenary Assembly of the Comité Consultatif International Téléphonique (C.C.I.F.), October, 1951, Florence—R. Sueur. (*Câbles & Transm.*, vol. 6, pp. 281–284; Oct., 1952.) Review of the proceedings.

621.395.44 1121

Terminal Equipment of Modern Carrier-Frequency Telephony Systems—J. Bauer. (*Bull. schweiz. elektrotech. Ver.*, vol. 43, pp. 824–829; Oct. 4, 1952.) Description of basic groups and of equipment suitable for symmetrical lines and for coaxial cables.

621.396:656.2(54) 1122

Radio on Indian Railways—H. C. Towers. (*Elect. Jour.*, formerly *Electrician*, vol. 149, pp. 1241–1243; Oct. 24, 1952.) An outline of recent developments in the installation of 20-W and 500-W radiocommunication sets, and of experiments on various applications of VHF mobile equipment.

621.396(494):061.2 1123

Growth, Organization and Activities of the Swiss Radiocommunication and Broadcasting Organizations—F. Rothen. (*Bull. schweiz. elektrotech. Ver.*, vol. 43, pp. 815–820; Oct. 4, 1952.)

621.396.1 1124

Recent International Conferences and the Radiofrequency Spectrum—(*Rev. Telegr. Electronica* (Buenos Aires), vol. 40, pp. 415–419, 424; July, 1952.) A report of work done from 1946 to 1951, with particular reference to frequency allocations for the American continent.

621.396.619.16 1125

Experimental System of Multifrequency Code Modulation—A. Pinet. (*Câbles & Transm.*, vol. 6, pp. 285–300; Oct., 1952.) The basic principles of code modulation are outlined and a detailed description is given of the first experimental system of this type developed in the research laboratories of the French P.T.T. The coding equipment is of the parallel type previously considered [2331 of 1952 (Libois)]. Somewhat complex circuits were found necessary to obtain a satisfactory signal/noise ratio, particularly at low modulation levels, so that the number of valves for each of the 12 channels is about 15, whereas for a similar delta-modulation system [2330 of 1952 (Libois)] only about 6 valves per channel are required. Control equipment for adjusting and testing the various sections of the system is noted.

621.396.65.029.63/.64 1126

UHF Radio-Relay-System Engineering—J. J. Egli. (Proc. I.R.E., vol. 41, pp. 115–124; Jan., 1953.) Abacs are presented for finding required path clearances from path parameters and operating frequency. Methods of minimizing fading are discussed; formulas are derived for determining for a given path profile the required spacing of receiving aerials for space-diversity reception. The power levels at various points of a single-hop system are evaluated, and the performance and reliability of a multi-hop system are discussed.

621.396.65.029.64:621.396.619.13 1127

An FM Microwave Radio Relay—R. E. Lacy and C. E. Sharp. (Proc. I.R.E., vol. 41, pp. 125–128; Jan., 1953.) 1952 I.R.E. National Convention paper. A transportable unit for military purposes, operating with 50 channels in the frequency range 8–8.5 kmc, uses a common antenna for transmitter and receiver, with duplexing by means of a waveguide magic T. The antenna system comprises a paraboloidal reflector fed by an off-center horn. The tuned CW magnetron gives >50 W carrier power. Frequency is stabilized by the method previously described by Bruck (1569 of 1948).

621.396.712 1128

Transmitting Equipment at the Wave-Overijse Center—G. Hansen. (*HF, Brussels*, vol. 2, pp. 81–96; 1952.) A general description is given of the high-power transmitters, antenna systems, antenna-switching arrangements and control consoles at this new center for broadcasting in French and Flemish the national programmes on medium wavelengths, and for SW transmission of the "colonial" programme to the Belgian Congo and of the "world" programme to other parts of the world. Detailed descriptions are included of the output stages, HV rectifier system, SW antenna-switching system and heat-recovery system.

621.396.931 1129

Synchronous FM System—(*Wireless World*, vol. 59, pp. 77–78; Feb., 1953.) Operation of a VHF communication system in a hilly, and in some parts mountainous, area such as Ayrshire is found impractical with a single central station. A master station controlling two slave stations has consequently been adopted. The master station broadcasts at 97.5 mc, and the link transmitters feeding the satellites operate at 146.25 mc, both frequencies being controlled by a master crystal with a frequency of 1.35416 mc. A process of frequency multiplication and division restores the frequency of 97.5 mc for transmission by the satellites, time delay circuits ensuring simultaneous speech radiation from the master and slave stations.

SUBSIDIARY APPARATUS

621-526:621.3.015.3 1130

Transients in Nonlinear Servomechanisms with Filters—J. Loeb. (*Ann. Télécommun.*, vol. 7, pp. 408–410; Oct., 1952.) Continuing previous work (2041 of 1952), equations are derived for amplitude x and elongation y (a complex function of time t) for the case of small damping, and a new construction for point-to-point tracing of the x/t curve is indicated which furnishes a new criterion of stability of steady states of oscillation.

621-526:621.3.015.7 1131

The Pulse Transfer Function and its Application to Sampling Servo Systems—R. H. Barker. (*Proc. IEE* (London), part IV, vol. 99, pp. 302–317; Dec., 1952. Digest, *ibid.*, part II, vol. 99, pp. 302–317; Oct., 1952.) A method of analysis of sampling in a linear system is described which is based on a sequence transformation closely analogous to the Laplace transform. The pulse transfer function relates a sequence of samples at the output of the system to the input sequence of pulses producing it. Servo systems with a finite time-delay in the feedback loop are particularly considered. A list of transforms is provided in an appendix.

621-526:681.142 1132

The Design and Testing of an Electronic Simulator for a Hydraulic Remote-Position-Control Servomechanism—F. J. U. Ritson and P. H. Hammond. (*Proc. IEE* (London), part II, vol. 99, pp. 533–548. Discussion, pp. 549–552; Dec., 1952.)

621.311.6:621.316.72 1133

Exceptionally Stable Regulated Power Supply for Electrometer Tubes—W. P. Senett and R. W. Pierce. (*Rev. Sci. Instr.*, vol. 23, pp. 534–537; Oct., 1952.) Description of a two-stage regulator circuit delivering about 100 mA at 80 V with the output regulated to within several parts per million for line-voltage changes of ±10%. Drift of the output voltage is about 5 parts per million per hour. Theory, construction details and results of performance tests are given.

621.311.6:621.316.72 1134

Radiofrequency Power Supply—E. M. Reilley, R. S. Bender and H. J. Hausman. (*Rev. Sci. Instr.*, vol. 23, pp. 572–573; Oct., 1952.) Brief description of a circuit designed for a maximum current output of 500 μA, an output voltage of 500–3000 V being provided by means of a chain of wire-wound resistors. The variation of the dc output voltage, measured over a period of 48 hours, was <2 parts in 104.

621.316.7.076.12 1135

Compensation of Feedback-Control Systems subject to Saturation—G. C. Newton, Jr. (*Jour. Franklin Inst.*, vol. 254, pp. 281–296 and 391–413; Oct. & Nov., 1952.) A theory for the design of compensating networks for feedback-control systems and filters is developed. . . The novel feature of this theory is its consideration of saturation and transient performance in addition to the usual steady-state behavior. This theory is essentially an extension of the researches of Wiener and Lee in statistical methods for filter design. Saturation is handled by limiting the RMS signal levels at critical points in the linear model used as the design basis for the physical system. Transient performance is handled by limiting the integral-square errors to a set of transient test signals."

621.316.722.1:621.387 1136

Corona Discharge Tubes for Voltage Stabilization—E. E. Shelton and F. Wade. (*Electronic Eng.*, vol. 25, pp. 18–21; Jan., 1953.) The mechanism of the corona discharge is briefly described by reference to the current/voltage characteristic of the glow-discharge tube. The construction, processing and performance of experimental coaxial types of tube are discussed. The data presented are sufficient for the design of tubes with regulating voltages between 400 and 1000 V operating at currents up to at least 100 μA. Such tubes can be made with characteristics stable for over 1000 hours.

621.316.726.087.9 1137

The Importance of Frequency Rate Indication in the Control of the National Grid—(*Muirhead Technique*, vol. 6, pp. 27–29; Oct., 1952.) Description of the construction and use of equipment indicating time error, mains-frequency error and rate of change of mains frequency. The unit comprises a mains-driven and a tuning-fork controlled motor, both coupled to a differential gear whose output shaft moves only when the mains frequency differs from 50 cps. Systems of differentiators and geared counters operate the indicating instruments.

TELEVISION AND PHOTOTELEGRAPHY

621.397.2 1138

Radio Facsimile Weather Map Transmissions—(*Muirhead Technique*, vol. 6, p. 29; Oct., 1952.) Details of transmission times and frequencies for the stations Washington, Port Lyautey, Balboa and Frankfort.

621.397.2:621.317.34 1139

Apparatus for the Measurement of Non-linearity in Television Trunk Systems—Griffith. (See 1064.)

621.397.2:621.317.74 1140

Group-Delay Distortion-Measuring Equipment—Hunt and Kemp. (See 1076.)

621.397.2:621.396.615.17:621.317.34 1141

A Waveform Generator and Display Unit for the Testing of a Television Channel—Ackland-Snow and Gledhill. (See 1087.)

621.397.2:621.396.65 1142

Permanent Point-to-Point Links for Relaying Television—H. Faulkner. (*Proc. IEE* (London), part IIIA, vol. 99, no. 18, pp. 313–322. Discussion, pp. 472–478; 1952.) An account of the various cable and radio relay links for the transmission of television signals, with a video-frequency bandwidth of at least 3 Mc, between London and other stations in the United Kingdom network. See also 820 (Clayton et al.), 2342 and 2625 (Kilvington et al.) of 1952.

621.397.42/.26 1143

Temporary Linkages for Outside Broadcasting Purposes—A. R. A. Rendall and W. N. Anderson. (*Proc. IEE* (London), part IIIA, vol. 99, pp. 323–331. Discussion, pp. 472–478; April/May, 1952.) Discussion of the relative merits of meter-wave and centimeter-wave equipment for use on television outside-broadcasting links shows that both can be usefully employed. A general description is given of the arrangements adopted by the BBC in conjunction with the GPO to provide links between pickup point and transmitter giving satisfactory picture quality. Examples are given of typical links, with performance data.

621.397.24/.26 1144

Operating Television O.B. Units—J. F. Hartwright. (*Wireless World*, vol. 59, pp. 74–76; Feb., 1953.) A short account of the function of the various members of the team required to operate the different units of the equipment for an outside television broadcast, and of the procedures adopted under various operating conditions.

621.397.24 1145

Cable Links for Television Outside Broadcasts—T. Kilvington. (*Proc. IEE* (London), Part IIIA, vol. 99, pp. 415–420. Discussion, pp. 472–478; April/May, 1952.) An outline of the methods used for the transmission of television signals over (*a*) special low-loss balanced-pair cable, (*b*) coaxial cable, or (*c*) pairs in the ordinary telephone network. For (*a*) and (*c*) transmission is direct; for (*b*) a carrier system is used.

621.397.24:621.3.018.78:621.315.212 1146

Some Factors affecting the Performance of Coaxial Cables for Permanent Television Links—H. Ashcroft, W. W. H. Clarke and J. D. S. Hinchliffe. (*Proc. IEE* (London), part IIIA, vol. 99, pp. 350–356. Discussion pp. 472–478;) April/May, 1952.

621.397.24:621.315.212.4 1147

The Birmingham-Manchester-Holme-Moss Television-Cable System—R. J. Halsey and H. Williams. (*Proc. IEE* (London), part IIIA, vol. 99, pp. 398–410. Discussion, pp. 472–478; April/May, 1952.) Description of the specification for, and the construction and performance characteristics of the system. See also 2342 of 1952.

621.397.24:621.395.521.3 1148

The Delay Equalization of the London-Birmingham Television Cable System—J. W. Allnatt. (*Proc. IEE* (London), part IIIA, vol. 99, pp. 338–349. Discussion, pp. 472–478; April/May, 1952.) A systematic method of equalizer design is outlined and a description is given of the units developed for the London-Birmingham cable link, together with performance figures. See also 2625 of 1952 (Kilvington et al.).

621.397.24:621.396.619.24 1149

Television Frequency-Translating Terminal Equipment for the Birmingham-Holme-Moss Coaxial Cable—A. H. Roche and L. E. Weaver. (*Proc. IEE* (London), part IIIA, vol. 99, pp. 455–463. Discussion, pp. 472–478; April/May, 1952.) Factors affecting the performance of vestigial-sideband systems are discussed and reasons are given for the choice of carrier frequency and sideband width for the Birmingham-Manchester television link, which uses standard 3/8-in. diameter coaxial cable. A description is given of the double-modulation method by which the vision signal is converted into a modulated wave with a carrier frequency of 1.056 mc and a vestigial-sideband width of 500 kc. At the receiving terminal the procedure is reversed and the vision signal restored to its original place in the frequency spectrum. Electrical and mechanical details of the equipment are described, with special attention to the design of the modulators and band-shaping filters.

621.397.24:621.396.645 1150

The Design of Amplifiers for the Birmingham-Manchester Coaxial Cable—Duerdoth. (See 977.)

621.397.24:621.396.645.371 1151

Two Simple Types of Feedback Amplifier for the Relaying of Television Signals over Coaxial Cables—Clifford. (See 984.)

1152

621.397.26.029.63:621.318.57+621.392.52

Ultra-High-Frequency Switches and Filters—Small. (See 951.)

621.397.335 1153

Synchronization and Pulse Technique in Television—J. Günther. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 4, pp. 161–175; Sept./Oct., 1952.) The proposed C.C.I.R. standard composite signal for 625-line television is discussed and possible synchronization faults and the corresponding necessary precautions are indicated. Technical terms which have newly gained acceptance in the German television sphere are listed together with their English equivalents.

621.397.5 1154

Some Possibilities for the Compression of Television Signals by Recoding—E. C. Cherry and G. G. Gouriet. (*Proc. IEE* (London), part III, vol. 100, pp. 9–18; Jan., 1953.) Discussion with reference to communication theory.

621.397.5:535.88 1155

An Experimental System for Slightly-Delayed Projection of Television Pictures—P. Mandel. (*Jour. Brit. I.R.E.*, vol. 12, pp. 567–575; Nov., 1952.) For another account of the system see 241 of January.

621.397.6:621.317.34 1156

The Transient Testing of Television Apparatus—Cooper. (See 1065.)

621.397.6:621.317.34 1157

The Response of a Vestigial Side-Band System to a "Sine-Squared" Step Transition—Skwirzynski. (See 1066.)

621.397.6:621.396.65 1158

Portable Equipment for a Microwave Television Link—G. Dawson. (*Proc. IEE* (London), part IIIA, vol. 99, pp. 379–384. Discussion, pp. 472–478; April/May, 1952.) Description of transmitting and receiving equipment for a link operating in either the 3.6–4.2 or the 4.4–4.8-kmc band. A coaxial-line VM oscillator tube [1214 below (Lambert)] was found so stable in operation that no AFC system was required. With paraboloidal aerials 4 ft. in diameter, satisfactory operation is obtained under normal conditions over paths of length up to 40 miles.

621.397.6:621.396.65 1159

A Mobile High-Power Microwave Link for Vision and Sound—F. W. Cutts. (*Proc. IEE* (London), part IIIA, vol. 99, no. 18, pp. 374–378. Discussion pp. 472–478; April/May, 1952.) Description of equipment for transmitting one video and two audio signals from an outside-broadcast site back to base. Paraboloid reflectors 4 ft. in diameter, with waveguide feed from a type-R5081 reflex klystron for the 3.9–4.2-kMc band or from a Type-R6010 klystron for the 4.4–4.8-kMc band, provide an output power of 4W. Double f.m. is used for the audio signals to reduce cross-modulation. Performance test results are in good agreement with theory. For a description of the Type-R5081 tube see 1219 below (Pearce and Mayo).

621.397.6:621.396.67.029.63 1160

Hinged Tubular Mast for Decimeter Television Mobile Transmitter—H. Käding. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 4, pp. 181–183; Sept./Oct. 1952.) Details are given of a four-section mast which when erected carries the aerial at a height of 17 m, and of the trailer on which it is supported.

621.397.61 1161

High-Power Television-Transmitter Technique, with Particular Reference to the Transmitter at Holme Moss—V. J. Cooper. (*Proc. IEE* (London), part IIIA, vol. 99, no. 18, pp. 231–242. Discussion, pp. 264–269; 1952.) Performance requirements for television transmitters are discussed. Comparison of high-level and low-level modulating systems indicates that high-level systems are preferable. The design features of a B.B.C. 50-kw transmitter are discussed, with special reference to shunt-regulated amplifiers, pulse-stabilization methods and black-level control. Transmitter testing methods are considered briefly. See also 567 of February.

621.397.61:621.392.52:621.396.67 1162

A Combining Filter for Vision and Sound Transmission—B. M. Sosin. (*Proc. IEE* (London), part IIIA, vol. 99, no. 18, pp. 253–264. Discussion, pp. 264–269; 1952.) Possible methods of combining sound and vision signals and feeding them into a common antenna are reviewed. Use of the Maxwell bridge leads to a compact design with good performance. Detailed analysis of this type of combining filter is presented. A coaxial unit is described and its equivalent circuit given, performance measurements being shown graphically. Two types have been designed, one for the high-power transmitter at Holme Moss (50-kw video and 12-kw sound), the other for medium-power transmitters (5-kw video and 2-kw sound).

621.397.61:621.396.619.2 1163

Television Modulation—F. C. McLean and E. Green. (*Wireless World*, vol. 59, p. 63; Feb., 1953.) Critical comments on paper noted in 568 of February.

621.397.611:628.9 1164

Illumination in Television Studios—J. Sánchez Cordovés. (*Rev. Telecomunicación (Madrid)*, vol. 8, pp. 39–43; Sept., 1952.) Discussion of the factors to be considered in determining correct illumination in relation to the characteristics of the camera tube used.

621.397.611.2 1165

Some Problems of Television Pickup Technique—H. Hewel. (*Tech. Hausmitt. Nordw-Dtsch. Rdfunks*, vol. 4, pp. 176–180; Sept./Oct., 1952.) The principles and design are discussed of three items of auxiliary equipment for the portable pickup set described in 2056 of 1952; they are: (*a*) automatic aperature control of the optical system; (*b*) synchronized black-level control in the video amplifier; (*c*) electronic picture-signal mixer.

621.397.62 1166

Tuner for Complete UHF-TV Coverage without Moving Contacts—R. J. Lindeman and C. E. Dean. (Proc. I.R.E., vol. 41, pp. 67–72; Jan., 1953.) The circuit comprises two-stage preselector, crystal mixer, triode oscillator and cascode-type first IF stage; the inupt matches a 300-Ω aerial and the output is suitable for feeding the IF amplifier of a conventional television receiver. The tuned circuits use balanced parallel-rod constructions, and are screened. Methods of achieving low noise figure, adequate rejection of unwanted signals, and low oscillator radiation are discussed and measurements of the performance are reported.

621.397.62 1167

Comparative Survey of Input Stages of Television Receivers—W. Reichel. (*Funk u. Ton*, vol. 6, pp. 406–415; Aug., 1952.) HF amplifier, oscillator, mixer and tuning arrangements used in various U. S. and European receivers are compared; details are given of some interesting examples from German models.

621.397.62:535.88 1168

Special Problems in Television Large-Picture Installations—E. Schwartz. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 4, pp. 184–194; Sept./Oct., 1952.) Arrangements using CR tubes with optical projection systems are surveyed, and their limitations indicated; the possibilities of systems using light tubes, as in the eidophor process, are noted.

621.397.62:621.396.615.17.018.75 1169

A Testing Pulse for Television Links—Macdiarmid. (See 1089.)

621.397.7 1170

Television Facilities of the Canadian Broadcasting Corporation—J. E. Hayes. (*Jour. Soc. Mot. Pict. Telev. Engrs.*, vol. 59, pp. 398–405; Nov., 1952.) Description of television broadcasting stations established in Montreal and Toronto. Operation at Montreal is on channel 2 (54–60 mc) with an effective radiated power of 16 kw, and at Toronto is on channel 9 (186–192 mc) with an output of 26 kw.

621.397.7 1171

Television Broadcasting Stations—P. A. T. Bevan. (*Proc. IEE* (London), part IIIA, vol. 99, no. 18, pp. 179–214. Discussion, pp. 264–269; 1952.) A survey is made of the distribution scheme by which the B.B.C. hopes to provide the greater part of the United Kingdom with a 405-line television service. The basic factors underlying the choice of scheme are discussed, together with the arrangements for dividing the available 41–68 mc frequency band into five separate operating channels using the receiver-attenuation system of vestigial-sideband transmission. The proposals for operating two geographically separated transmitters, one of high power and one of low power, on each channel to complete a 10-station plan, and the precautions needed to minimize co-channel interference, are explained. Problems of site selection for transmitting stations are mentioned and a brief account is given of propagation phenomena which affect the estimation of service area. The scheme for linking the stations with the London studio center is briefly described. The general planning and design of the complete transmitting equipment used at the new B.B.C. high-power television stations, and of that intended for use at the future low-power stations, are surveyed, with particular reference to the different types of vision and sound transmitter, monitoring equipment, air- and water-cooling installations, vision/sound combining circuits, transmission lines and aerial systems. For a shorter version, see *B.B.C. Quart.*, vol. 7, pp. 235–245; Winter, 1952–1953.

621.397.7 1172

The Selection and Testing of Sites for Television Transmitters in the United Kingdom—L. F. Tagholm and G. I. Ross. (*Proc. IEE* (London), part IIIA, vol. 99, no. 18, pp. 300–309. Discussion, pp. 310–312; 1952.) Discussion of the problem of providing a television service for the whole of the United Kingdom, using five high-power transmitters linked with five low-power transmitters operating at different frequencies in the 41–68-mc band. Site testing and determination of service areas were carried out at various places, using an aerial supported by a balloon. Reasons for the final choice of sites are given. Experiments were made at two sites to determine the effects of re-radiation from neighbouring structures which might give rise to ghost images.

621.397.8:551.594.5 1173

Auroral Effects on Television—R. E. Thayer. Proc. I.R.E., vol. 41, pp. 161; Jan., 1953.) Brief illustrated report of a particular form of television interference observed in the Ithaca, New York, fringe area and attributed to auroral propagation phenomena of the type reported previously, eg by Moore (2002 of 1951).

621.397.8:621.396.11 1174

A Survey of British Research on Wave Propagation with Particular Reference to Television—Smith-Rose. (See 1109.)

621.397.8:621.396.11 1175

Long-Distance Propagation in Relation to Television in the United Kingdom—Saxton. (See 1110.)

621.397.812 1176

Ionospheric Influences in Television Reception—F. A. Kitchen and K. W. Tremellen. (*Proc. IEE* (London), part IIIA, vol. 99, No. 18, pp. 290–293. Discussion, pp. 310–312; 1952.) The effects of ionospheric reflections of VHF waves on television reception are reviewed. Sustained propagation via the F_2 region may occur for limited periods during years of maximum solar activity, whilst irregular and intermittent propagation via sporadic E-region clouds may occur during certain months of every year. The range of frequencies concerned includes those now used for television services in the United Kingdom and elsewhere. A summary is presented of the circumstances under which interference may be expected with the reception of television transmissions, owing to the use of common frequencies for remotely situated short-range VHF systems.

621.397.813:621.3.018.78 1177

Effects of Amplitude and Phase Distortion on Limited-Spectrum Television Signals—M. Jouguet. (*Câbles & Transm.*, vol. 6, pp. 301–316; October, 1952.) A concise exposition is given of the principles of the method of orthogonal signals developed by Oswald (210 of 1951) and Ville (2933 of 1951), and of the analogous method of "paired echoes" described by Wheeler (3642 of 1939), for studying the effects of amplitude and phase distortion on the transmission of signals. Relations between these two methods are noted and Wheeler's method is applied to discussion of (*a*) signals of the $\cos^2\theta$ type, (*b*) simple amplitude distortion, (*c*) simple phase distortion, (*d*) a system with both amplitude and phase distortion.

621.397.813:621.395.521.3 1178

A Variable Time-Equalizer for Video-Frequency Waveform Correction—J. M. Linke. (*Proc. IEE* (London), part IIIA, vol. 99, pp. 427–435. Discussion, pp. 472–478; April/May, 1952.) Description of equipment operating on the time-equalizer principle, whereby a wide variety of transmission characteristics can be obtained by combining the input signal with series of (*a*) advanced, (*b*) retarded echoes (attenuated replicas) of the input signal, each echo being individually adjustable in amplitude and sign. The equipment consists essentially of a delay line with adjustable tapping devices which include arrangements for combining and amplifying the various signal components.

621.397.82+621.396.82 1179

Current Radio-Interference Problems—E. M. Lee. (*Jour. Brit. I.R.E.*, vol. 12, pp. 551–

564; Nov., 1952.) A comprehensive review of the many different sources of interference with radio and particularly television reception, with special reference to G.P.O. statistics for 1951 and methods recommended for reducing such interference. Evidence is given of the improved performance of petrol engines resulting from the fitting of suppressors in the ignition-system leads, particularly as regards starting in cold weather and increase of sparking-plug life. Legislation on the subject and specifications for suppressor units are discussed, and an outline is given of methods for the measurement of interference.

621.397.82:621.396.67:656.13 1180
Sensitivity of a Television Antenna to Interference from Motor Vehicles as dependent on its Vertical or Horizontal Polarization—F. de Clerck. (*HF, Brussels*, vol. 2, no. 4, pp. 97–98; 1952.) Tests carried out on a dipole antenna, tuned to 180 mc which could be arranged either vertically or horizontally at the top of a mast 10 m high, indicate that interference from the ignition system of a motor car is less with vertical polarization, the ratio of the interference effects in the two cases being of the order of 8 db.

TRANSMISSION

621.396.619.11 1181
The "Rothman" Modulation System—A. J. A. Coghlan. (*Rev. Telegr. Electronica (Buenos Aires)*, vol. 40, pp. 420–422, 436; July, 1952.) A screen-grid modulation system is described in which a small portion of the transmitter output is rectified, demodulated and reapplied to the output stage in such a way that the change in valve operating angle due to the modulation is compensated. Greatly increased efficiency results. The power requirements of the screen grid in the modulator valve must not be an appreciable fraction of the total output power. The circuit is unsuitable for frequency-multiplier stages.

621.396.931:621.396.619.23 1182
Carrier Control with Self-Biased Clamp-Tube Modulator—(*QST*, vol. 36, pp. 41–44; Nov., 1952.) Discussion of effects obtained in screen-grid modulation of the RF amplifier tube by a clamp-valve modulator with a Se rectifier connected between the control grid and the microphone transformer. A certain amount of carrier control is obtained, but distortion results from the clipping of the positive half of the AF cycle by the rectifier. Substitution of a grid-blocking capacitor for the rectifier reduces the distortion without impairnig carrier control.

TUBES AND THERMIONICS

537.533:621.385 1183
Low-Noise Electron Streams—H. W. König. (*Arch. elekt. Übertragung*, vol. 6, pp. 445–452; Nov., 1952.) The sinusoidal fluctuations of field strength and electron velocity in the planes of two consecutive grids, $(n-1)$ and n, are linearly related. The determinant of the system of equations is given by the ratio of the velocities in the grid planes, IE, $D_n=v_{n-1}/v_n$. In the limiting case $D_n\rightarrow 0$, both noise components in the plane of grid n vanish. Analysis is presented for the case of an acceleration space between cathode and grid-2, with an intermediate grid-1, assuming pure velocity fluctuations at the cathode. By suitable choice of the electrode separations, the field-strength fluctuations or the velocity fluctuations in the grid-2 plane can be made to vanish. The velocity ratios can also be chosen so that both noise components in the grid-2 plane can be reduced to a fraction σ of the corresponding values for the accelerating space without the intermediate grid-1, the current density and the end velocity v_2 being the same in both cases. If Z is the noise factor of a klystron with a divided pre-acceleration space, $Z-1$ is reduced by a factor $\sigma 2$ which for $v_1 \ll v_2$, is approximately $v_1/4v_2$. The maximum reduction of $Z-1$ attainable by this method is estimated as about 20 db.

621.314.7 1184
High-Frequency Transistor Tetrode—R. L. Wallace, Jr., L. G. Schimpf and E. Dickten. (*Electronics*, vol. 26, pp. 112–113; Jan., 1953.) See 677 of March.

621.385:621.396.822 1185
Low-Frequency Noise in Electron Tubes. C. Electrometer Tubes—J. G. van Wijngaarden and E. F. de Haan. (*Physica*, vol. 18, pp. 705–713; Oct., 1952.) Noise measurements are reported on Philips Type-4060, Ferranti Type-BM4A and Victoreen Type-VX41 electrometer valves and on a normal Type-AF7 pentode. Flicker noise is important in all the electrometer types at low frequencies. There is satisfactory agreement between values of grid current measured directly and values derived from noise measurements.

621.385.029.64 1186
Travelling-Wave Tube with Sinuous Rectangular Waveguide—H. Kleinwächter. (*Arch. elekt. Übertragung*, vol. 6, p. 460; Nov., 1952.) Theory and a sketch are given of a travelling-wave UHF amplifier tube with very great bandwidth. This embodies a section of rectangular waveguide, the middle portion of which is much reduced in width and corrugated. H_{01} waves are fed through a side branch and travel through the waveguide in the same direction as the electron beam.

621.385.029.65 1187
Spatial Harmonic Traveling-Wave Amplifier—S. Millman. (*Bell Lab. Rec.*, vol. 30, pp. 413–416; Nov., 1952.) Description of the construction of a new type of travelling-wave tube operating at 50 km, in which the usual helix is replaced by a Cu block with three longitudinal slots, down which the main stream of electrons travels, and 100 transverse resonator slots, of width 0.0065 in. and depth 0.056 in. The dimensions of the Cu block are $2\frac{1}{4}\times\frac{3}{8}\times\frac{1}{4}$in. Amplification results from interaction between the electrons and the travelling wave at the successive transverse slots, the action being effectively the same as if the speeds of wave and electrons were equal, although the electrons are actually travelling more slowly, the ratio of velocities being $d/(d+c)$, where d is the slot spacing and λ the wavelength of the travelling wave. The bandwidth of the amplifier is about 1.5 kmc/s, gain >20 db, and estimated output power about 25 mw.

621.385.032.216 1188
Spectroscopic Investigations of Oxide Cathodes in Gas Discharges—E. Krautz. (*Z. Naturf.*, vol. 6a, pp. 16–24; Jan., 1951.) Investigations were made of cathodes with pure alkaline-earth oxides, mixed crystals, and compounds including other metal oxides, over the wavelength range 2200–7600 Å. Spectrograms show that the destruction of the oxide layer in a low-pressure discharge is largely an atomic process. The method affords a ready means of assessing not only the composition of the surface layers but also the resistance of the cathodes to evaporation, disintegration and dissociation.

621.385.032.216 1189
Bariated Tungsten Emitters—R. C. Hughes and P. P. Coppola. (*Jour. Appl. Phys.*, vol. 23, pp. 1261–1262; Nov., 1952.) Methods are described for the preparation of cathodes in which BaO is dispersed in W without simultaneous oxidation of the W. Cathodes with a BaO content of 5–10% can give continuous emission >100 A/cm² and are highly resistant to damage. A typical specimen gave an emission of about 8A/cm² at a temperature of 1000°C, with a life >650 hours at 1100°C. Richardson-equation constants for a moderately active cathode are $\phi=1.56$ eV and $A=0.6$ A/cm²/deg².

621.385.032.216:537.533 1190
Poisoning of Oxide-Cathode Emission by Oxygen—A. A. Shepherd. (*Nature* (London), vol. 170, pp. 839–890; Nov. 15, 1952.) Experiments are described which indicate the probability that the main cause of oxygen poisoning is the adsorption of thin films of oxygen on both the outer cathode surface and the interior crystallite surfaces, with consequent reduction of both emission and coating conductivity at high temperatures. At low temperatures, such adsorption will reduce the emission from the outer surface without appreciably affecting the conductivity of the coating, since conduction at temperatures below 700°K is mainly a direct crystal-to-crystal process.

621.385.2/.3 1191
The Triple-Diode-Triode E/U/PABC 80—H. te Gude. (*Funk u. Ton*, vol. 6, pp. 449–458; Sept., 1952.) The construction and characteristics of the tube are described and its use in the demodulator and AF stages of an am/fm receiver is illustrated.

621.385.2/.3]:621.396.822 1192
Low-Frequency Noise in Electron Tubes. A. Space-Charge Reduction of Flicker Effect—J. G. van Wijngaarden and K. M. van Vliet. (*Physica*, vol. 18, pp. 683–688; Oct., 1952.) Theoretical discussion of the flicker effect in diodes and triodes under normal operating conditions. This effect can be represented by an equivalent input intensity (or input noise resistance) which is a nearly constant part of the intensity of the fluctuations in the total emission current for a given cathode temperature.

621.385.2 1193
Influence of Initial Velocities on Electron Transit Time in Diodes—J. T. Wallmark. (*Jour. Appl. Phys.*, vol. 23, pp. 1096–1099; Oct., 1952.) The transit-time spread due to the spread of initial velocities is calculated for the case of nonuniform velocity distribution, and is given as a function of the current for (*a*) a normal diode, and (*b*) an inverted diode, IE, one in which the beam is reflected. The transit-time spread is also given as a function of anode voltage within certain limits. The theory is used to explain the properties of the tube previously described (1477 of 1952.)

621.385.2 1194
Effect of Filament Voltage on the Plate Current of a Diode—H. F. Ivey. (*Jour. Appl. Phys.*, vol. 23, pp. 1254–1256; Nov., 1952.) Analysis applicable to both planar and cylindrical arrangements is given for a diode with filament directly heated, either by dc or by ac. Graphs show the variation of I/I_0 with V_a/V_f for space-charge-limited conditions and with V_f/T for retarding-field conditions ($I=$actual anode current and I_0 the corresponding value with equipotential cathode).

621.385.2 1195
Calculation of the Efficiency and Damping of Diode Rectifiers—H. H. van Abbe. (*Funk u. Ton*, vol. 6, pp. 459–469; Sept., 1952.) Disadvantages of usual methods of calculation, based on the exponential initial portion of the diode characteristic, are indicated and a method is developed involving successive approximations. Divergences between calculated and measured values are $<5\%$. The effect of variations of the diode characteristic is investigated.

621.385.2:537.525.92 1196
The Space-Charge Smoothing Factor: Part 2—C. S. Bull. (*Proc. IEE* (London), part IV, vol. 99, pp. 289–293; Dec., 1952.) Full paper. See 291 of January.

621.385.2:[621.317.3+621.316.72 1197
Saturated-Diode Operation of Miniature Tubes—V. H. Attree. (*Electronic Eng.*, vol. 25, pp. 27–29; Jan., 1953.) Characteristics of a typical subminiature pentode and triode operated as diodes are discussed; the filament power required for temperature-limited operation is only a few milliwatts. A small change of

filament heating current produces a relatively large change of anode current. The tubes are useful as amplitude-sensing devices in low-power stabilizer circuits and as substitutes for thermojunctions in RF measurements.

621.385.2:621.396.822 1198
On the Relation between the Conductance and the Noise Power Spectrum of Certain Electronic Streams—J. J. Freeman. (*Jour. Appl. Phys.*, vol. 23, pp. 1223–1225; Nov., 1952.) Analysis is given for diodes with temperature-limited current and for diodes with retarding field. The noise-power spectrum can be considered as the sum of two terms, viz. that due to the pure shot noise which would be produced if the electrons were emitted with zero velocity, and that due to the thermal noise, which is related to the conductance by Nyquist's law.

621.385.2.032.216 1199
Electrical Measurement of the Cathode Temperature of Diodes—D. A. Bell, J. C. Cluley and H. O. Berktay. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 322–323; Oct., 1952.) The slope of the retarding-field characteristic of the diode is measured in terms of the resistance, and the temperature T is found from the relation $i.v_a = i.dV/di = kT/e$. Comparison with the results of pyrometer measurements indicates that the method is reliable for oxide-coated cathodes and may be useful at temperatures below those suitable for pyrometer measurements.

621.385.3 1200
Transit-Time Oscillations in Triodes—I. A. Harris; M. R. Gavin. (*Brit. Jour. Appl. Phys.*, vol. 3, pp. 363–364; Nov., 1952.) Discussion on paper abstracted in 2664 of 1952 (Critchley and Gavin).

621.385.3:621.396.615.14 1201
Limiting Frequency of Triode Oscillator—Y. Koike and S. Yamanaka. (*Tech. Rep. Tohoku Univ.*, vol. 16, no. 1, pp. 8–16; 1951.) Report of experiments made with two types of disk-seal triodes to verify theory previously advanced (871 of 1952). Values of limiting frequency were noted for different values of anode voltage. The influence of electrode configuration is indicated. An unexpectedly high value of electron transit angle was observed; a satisfactory explanation is not yet available.

621.385.3:621.396.822 1202
Low-Frequency Noise in Electron Tubes. B. Measurements on Triodes under Normal Operating Conditions—J. G. van Wijngaarden, K. M. van Vliet and C. J. van Leeuwen. (*Physica*, vol. 18, pp. 689–704; Oct., 1952.) Physical theories of the flicker effect are reviewed and measurements are reported covering the frequency range 1 cps–5 kc and various types of cathode. For W cathodes, deviations from the white shot-noise spectrum were observed only at very low frequencies. W-Th cathodes exhibited flicker effect of intensity inversely proportional to frequency, ν. With oxide cathodes, flicker effect was observed with a component proportional to ν^{-1} (attributed to changes inside the oxide layer) and in some cases with a second component proportional to ν^{-2} (attributed to changes at the surface of the cathode).

621.385.3:621.396.822 1203
The Limiting Sensitivity of Amplifier Tubes: Part 1—Theory of the Triode—H. Rothe. (*Arch. elekt. Übertragung*, vol. 6, pp. 461–468; Nov., 1952.) Correction, *ibid.*, vol. 6, p. 498; Dec., 1952. Tube noise consists of a coherent part, principally due to the electrons passing through the tube, and an incoherent part mainly due to the effect of the space charge between the cathode and the point of minimum potential. The effects in the different sections of a triode are analysed, making use of equivalent circuits, and a very simple expression for the noise figure is derived which differs from the expression hitherto given.

621.385.3.029.63 1204
R.F. Performance of a UHF Triode—H. W. A. Chalberg. (Proc. I.R.E., vol. 41, pp. 46–50; Jan., 1953.) 1952 I.R.E. National Convention paper. Techniques used in evaluating the performance of the Type Z-2103 tube (the development number corresponding to production type 6AJ4) are described; both coaxial-line measuring equipment and lumped-constant test circuits were used. Values of gain and noise figure are compared with values for available VHF tubes; results indicate the suitability of the Z-2103 valve for use as RF amplifier in VHF-UHF television tuners.

621.385.3.029.63 1205
Development of a UHF Grounded-Grid Amplifier—C. E. Horton. (Proc. I.R.E., vol. 41, pp. 73–79; Jan., 1953.) Account of the development of the 9-pin miniature triode Type 6AJ4 for operation in television receivers at frequencies up to 900 mc. Methods of reducing coupling between input and output circuits are discussed, and measurement techniques are indicated.

621.385.4.029.63 1206
One-Kilowatt Tetrode for UHF Transmitters—W. P. Bennett and H. F. Kazanowski. (Proc. I.R.E., vol. 41, pp. 13–19; Jan., 1953.) A forced-air-cooled tetrode capable of delivering 1.2 kW output in television service at frequencies up to 900 mc has a coaxial electrode structure with metal/ceramic seals, and an indirectly heated matrix-type cathode permitting application of high anode voltages without sparking. Production techniques are described for obtaining uniform close spacings. A suitable circuit is illustrated, and performance figures are given for operation as a cathode-driven amplifier.

621.396.615.14 1207
Theory of the Reflex Resnatron—M. Garbuny. (Proc. I.R.E., vol. 41, pp. 37–42; Jan., 1953.) An account is given of the mechanism of energy interchange between the density-modulated electron beam and the HF field in the tube previously described [3617 of 1952 (Sheppard et al.)]. The method of Lagrangian parameters is used to determine conditions for maximum energy transfer. Efficiency, amplitude modulation and bandwidth are discussed. Theoretical and experimental results are compared for operation in the neighbourhood of 600 mc; satisfactory agreement is obtained.

621.396.615.14 1208
An Axial-Flow Resnatron for UHF—R. L. McCreary, W. J. Armstrong and S. C. McNees. (Proc. I.R.E., vol. 41, pp. 42–46; Jan., 1953.) A design in which the electron stream is directed axially rather than radially leads to constructional simplicity and operational stability. Mechanical features are described and performance figures are tabulated and shown in graphs. The specification calls for a power output of 30 kW cw, a tuning range of 300–600 Mc, a power gain of at least 10 db with a 4-Mc bandwidth and an anode efficiency of at least 50%, and easy replacement or adjustment of electrodes.

621.396.615.141.2 1209
Magnetrons—J. Verweel. (*Philips tech. Rev.*, vol. 14, pp. 44–58; Aug., 1952.) An introductory article covering electron motion, oscillations, resonators, output systems and manufacturing problems.

621.396.615.141.2 1210
The Magnetron in the Static Cut-Off State. Experimental Study: Part 2—The Influence of the Inclination of the Magnetic Field to the Magnetron Axis—J. L. Delcroix. (*Compt. Rend. Acad. Sci.*, (Paris), vol. 235, pp. 1018–1020; Nov. 3, 1952.) Measurements were made of the residual current in the cut-off state for varying inclinations of the magnetic field to the magnetron axis, the inclination being produced by means of auxiliary coils providing a field perpendicular to the main magnetic field. The results confirm the existence, previously observed, of three regimes, B_0, B_1 and B_2, of which B_1 is only slightly influenced by the inclination, B_2 is critically influenced, and B_3 has intermediate properties. B_0 is the Brillouin regime, B_1 and B_2 are the first two bidromic regimes. Part 1: 3620 of 1952.

621.396.615.142:621.396.65 1211
A Coaxial-Line Velocity-Modulated Oscillator for Use in Frequency-Modulated Radio Links—D. E. Lambert. (*Proc. IEE* (London), part IIIA, vol. 99, pp. 421–426. Discussion, pp. 472–478; April/May, 1952.) "An outline is given of the tube requirements for frequency-modulated microwave radio links, and the design and development of suitable tubes for the 3.6–5.0-kmc band are described."

621.396.615.142.2 1212
High-Power U.H.F. Klystron—(*Tele, Tech*, vol. 11, pp. 60–61; Oct., 1952.) Description of the construction of the Type-V42 tube, a linear amplifier with a gain of over 23 db and an output of 15 kw in the 470–890-mc band. A demountable cathode assembly permits inexpensive replacement. Tuning is effected by flexure of the thin walls of the resonators. The concentric construction gives rigidity, and freedom from vibration and shock effects. The maximum beam current is 3 A at 17 kv, and bombarder voltage 2.4 kv between cathode and filament. The overall length is 50 in., and the weight 150 lb without the 100-lb beam-focusing magnets.

621.396.615.142.2 1213
High-Power Klystrons at UHF—D. H. Preist, C. E. Murdock and J. J. Woerner. (Proc. I.R.E., vol. 41, pp. 20–25; Jan., 1953.) The three-cavity gridless magnetically focused klystron is discussed and a particular type, the Eimac 5-kw television klystron, is described in detail. This has the cavities partly outside the vacuum system, thus facilitating tuning. The advantages of this type of tube over conventional grid-control types are indicated.

621.396.615.142.2:621.396.619.13 1214
The Design of a Reflex-Klystron Oscillator for Frequency Modulation at Centimeter Wavelengths—A. F. Pearce and B. J. Mayo. (*Proc. IEE* (London), part IIIA, vol. 99, pp. 445–454. Discussion, pp. 472–478; April/May, 1952.) The factors governing the design of a reflex klystron suitable for fm are discussed, and the conditions are considered which give maximum linearity between the oscillation frequency and the modulating voltage applied to the reflector. Practical design problems are mentioned and a description is given of the design and construction of a new tube, Type R5081, which gave an output of about 4 W at 4 kmc when used as the fm transmitting valve in a television relay link.

MISCELLANEOUS

061.4:621.396 1215
Sixth R.S.G.B. Radio Show—(*Wireless World*, vol. 59, pp. 6–8; Jan., 1953.) Brief descriptions of commercial and home-constructed amateur equipment.

621.3:629.13:061.4 1216
S.B.A.C. [Society of British Aircraft Constructors] Farnborough. Details of Electrical Exhibits—(*Electrician*, vol. 149, pp. 675–679; Sept. 5, 1952.) Short descriptions of a wide variety of electrical and electronic equipment on view at the annual show, September, 1952.

621.317.7:061.3 1217
"Electronic Instruments" Symposium—(*Electrician*, vol. 149, pp. 689–690; Sept. 5, 1952.) Brief summaries are given of the papers read at the symposium on "Electronic instruments in research and industry," September 1952, arranged by the Scientific Instrument Manufacturers' Association.