Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

November, 1951

Volume 39 Number 11

Chicago Telephone Supply Corporation

MINIATURIZED "ALL-WEATHER" COMPONENTS

Pictured above is a miniaturized variable resistor designed for stable operation over extreme ranges of temperature and humidity. With the advent of jet planes, guided missiles, and the like, the restrictions imposed on the size, weight, and performance requirements of electronic equipment become more exacting.

PROCEEDINGS OF THE I.R.E.

TABLE OF CONTENTS, INDICATED BY BLACK-AND-WHITE MARGIN, FOLLOWS PAGE 32A

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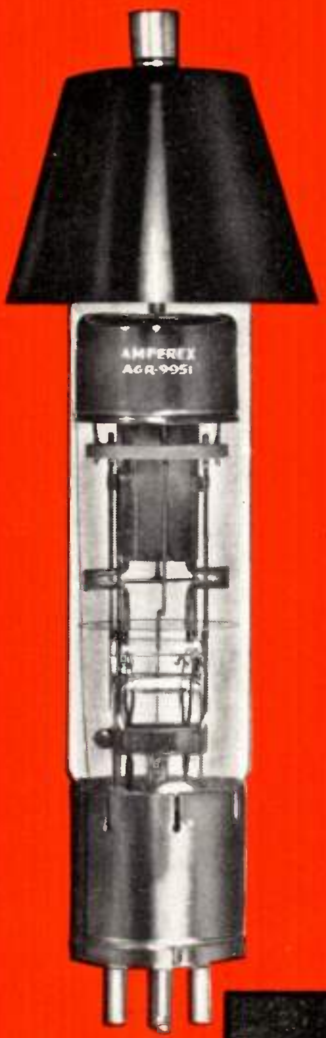
The Institute of Radio Engineers

New AMPEREX tubes

Specifically designed for grid-control operation at peak anode voltages as high as **27,000 v.** for heavy duty INDUSTRIAL uses and high power TRANSMITTERS with outputs to **150 KW.** (3 phase full wave)

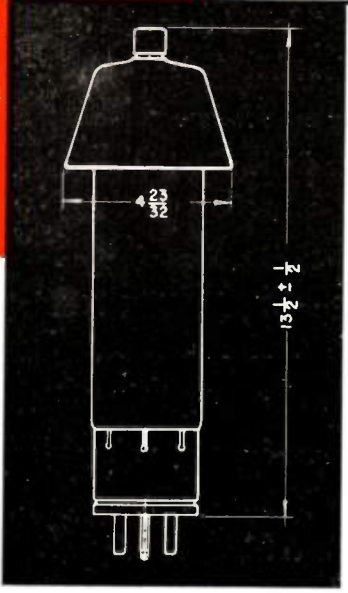
	AGR-9951/5870		AGR-9950/5869	
CATHODE Directly Heated, Oxide Coated				
MAXIMUM PEAK ANODE VOLTAGE				
Inverse	27,000	10,000	13,000	10,000
Forward	27,000	10,000	13,000	10,000
CONDENSED MERCURY TEMPERATURE LIMITS (centigrade)	+30° to +40°	+25° to +60°	+25° to +55°	+25° to +60°
MAXIMUM PLATE CURRENT (Amperes)				
Peak	10		4	
Average	2.5		1	
FREQUENCY RANGE (cps)	25 to 150		25 to 150	
FILAMENT VOLTAGE	5.0		5.0	
FILAMENT CURRENT (amperes)	15		6.5	
TUBE VOLTAGE DROP (volts, approx.)	14 (Ib = 10 amperes)		15 (Ib = 4 amperes)	



PROVEN LIFE

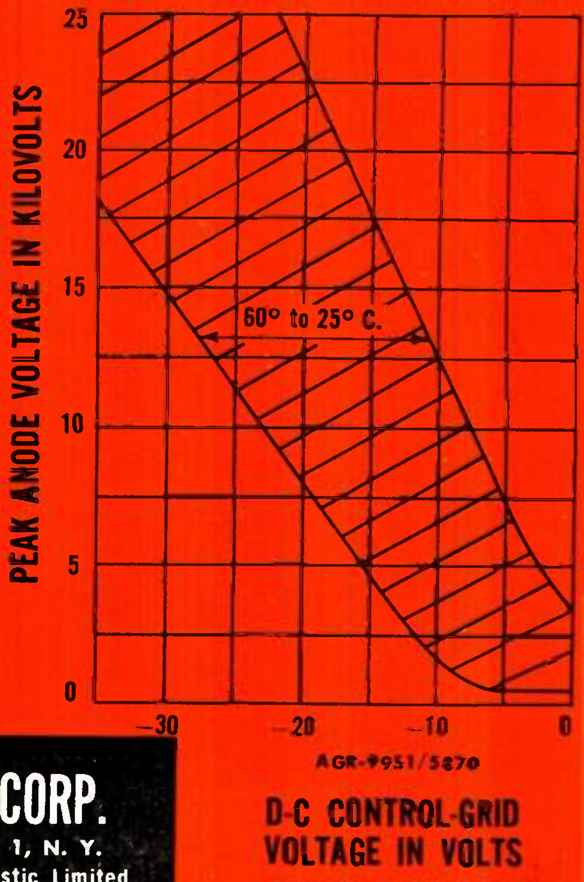
AGR-9951/5870

$110.00



Re-tube with AMPEREX

GENERAL CONTROL CHARACTERISTICS



Data sheets and charts available on request

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RECTIFYING TUBES

with NEGATIVE CONTROL characteristics

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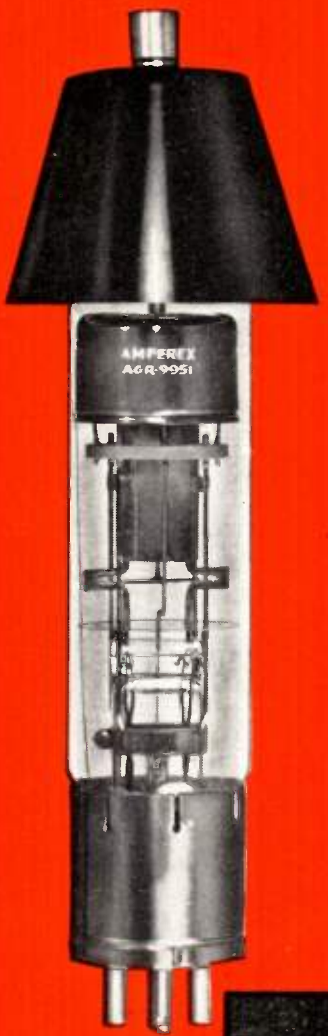

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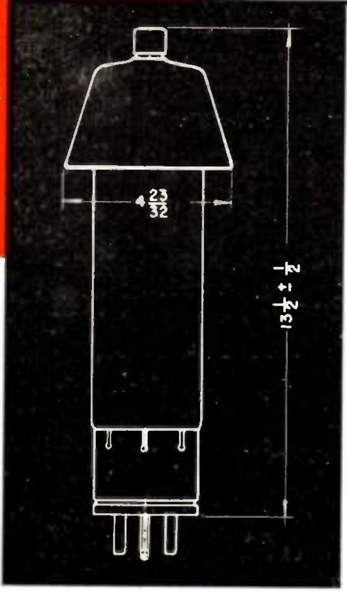

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RECTIFYING TUBES
with NEGATIVE CONTROL characteristics

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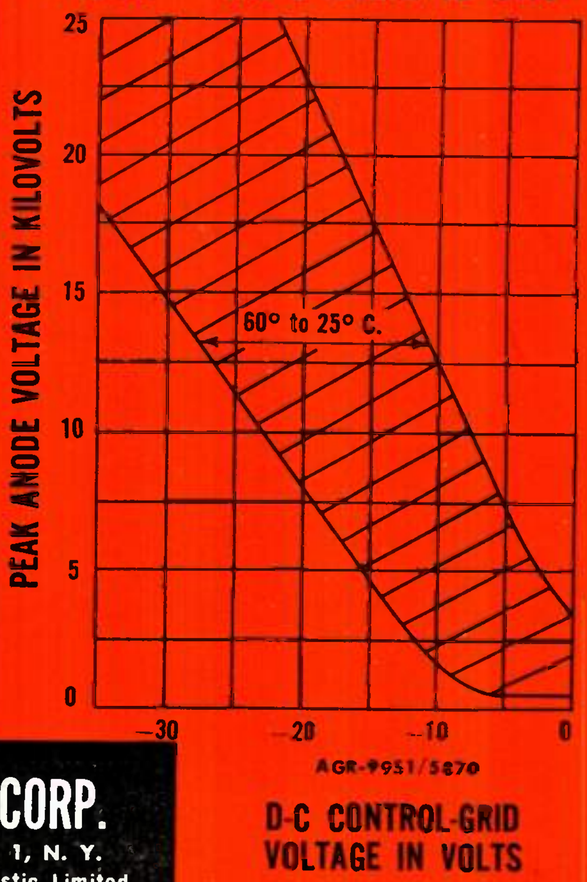

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MEMBER AUDIT BUREAU OF CIRCULATIONS

W. M. Rust, Jr.

BOARD OF DIRECTORS, 1951-1952

William Rust, Jr. attended the Rice Institute in Houston, Texas, where he received the degree of Doctor of Philosophy in 1931. His fields of study were in mathematics of physics and electrical engineering.

After a year of post-graduate study at the Charlottenburg Polytechnic Institute in Berlin, Germany, Dr. Rust spent a year as an instructor in mathematics at Harvard University. He was then employed as a research geophysicist by the Humble Oil and Refining Company in 1934, and was made Chief of their Geophysics Research Section in 1937. His work with this group was concerned with the problem of improving or devising new equipment and techniques for obtaining information of the deeper formations of the earth's crust, as a guide to the drilling of wells to discover new deposits of oil and gas.

During World War II, Dr. Rust was a consultant to Division II, NDRC, dealing with problems related to the effect of subterranean explosions. As Humble's representative under contracts with the Radiation Laboratory, MIT, Committee on Propagation of NDRC, Texas University, and other organizations, he was responsible for work ranging from radar components to plane-to-plane fire control.

Dr. Rust was President of the Society of Exploration Geophysicists in 1944, and as such represented the geophysical industry in the subsequent allocation proceedings of the Federal Communications Commission. Since the establishment, by the FCC, of the Petroleum Radio Service, he has been active in the industry committees concerned with the problems of this service. He has participated in numerous allocation and rule-making procedures of the FCC. He is a member of the American Petroleum Institute-Control Committee on Radio Facilities and assisted in the organization of the National Petroleum Radio Frequency Co-ordinating Association.

Dr. Rust holds a number of U. S. and foreign patents in the fields of seismic prospecting and electrical well logging. He is a member of a number of technical societies including the American Mathematical Society, American Physical Society, American Institute of Mining and Metallurgical Engineers, American Association of Petroleum Geologists, and the American Geophysical Union. He became a Senior Member of the IRE in 1947. Dr. Rust is a charter member of the Houston Section of the IRE and is at present the Regional Director for Region Six.

The IRE Professional Groups and the Institute

BENJAMIN B. BAUER

In order further to increase the value to IRE members of The Institute of Radio Engineers, the establishment of the IRE Professional Groups was authorized by the Board of Directors. The results of this action have, up to the present, been encouraging. The number of the Professional Groups steadily grows, and their activities are being thoughtfully and constructively expanded.

Since the success of the Groups is of vital interest to the IRE members, the following significant guest editorial from the Chairman of the IRE Professional Group on Audio, who is as well vice-president and chief engineer of Shure Brothers, Inc., in Chicago, is of particular and timely interest.—*The Editor.*

The rapid expansion of the sciences of communication and electronics into new fields has confronted the Institute with numerous problems. Inevitably, the professional interests of one or more groups of members have become temporarily overlooked. Consider, for example, the field of Audio Technology: To convey intelligence audibly to the sense of hearing is one of the most important end objects of electronic instrumentation. Nevertheless, of the 177 papers published in the PROCEEDINGS during the calendar year of 1950, only three titles relate to Audio. Similarly, on the Institute Section level, the tendency has been to pay court to the technology predominant in the particular locality, resulting in neglect of the interests of a smaller, though important segment of local membership.

Thus it is seen that occasional gaps have taken place in the technical services rendered by the Institute to various groups of members. The IRE Professional Groups are intended to bridge these gaps. The formation and evolution of Professional Groups is a tribute to the power of adaptability which the Institute shares with the industry which bears its name. In recognition of the growing complexity and extension of the fields encompassed by the Institute, the Professional Groups have been created within the framework of the IRE to bring together and serve members with common interests in specialized aspects of the Profession.

The Professional Group on Audio, for example, has been formed for the purpose of serving the professional interests of the IRE members concerned with Audio Technology. In the beginning, the Group pursued this objective by sponsoring sessions on Audio at various IRE meetings and conventions. Soon after its inception, the Group started issuing a NEWSLETTER to keep the members up to date regarding the activities of the Group, and including news and program notes. Beginning with the Fall of 1950, the Group began mailing to its members reprints of technical papers presented at the Audio Sessions which could not have been published promptly in the PROCEEDINGS.

Recently the NEWSLETTER has been improved and expanded. With the July, 1951 issue, a new series of technical editorials covering Sound Reproduction has been started. Each editorial is written especially for the NEWSLETTER by some outstanding authority. The first of the series, by L. L. Beranek of the Massachusetts Institute of Technology, is entitled, "Design of Loudspeaker Grilles." The September issue carries an Editorial by H. F. Olson of the RCA Laboratories dealing with, "Selection of a Loudspeaker." It is contemplated that this series will be followed by another, dealing with studio acoustics and sound transmission. A third series will deal with the problems in sound reinforcement and public address. The NEWSLETTER has also opened its pages to the Sectional Professional Groups on Audio, helping them to voice and solve their mutual problems. The first article dealing with this subject, in the July, 1951 issue, is by S. L. Almas, of the K.L.A. Laboratories Incorporated, Chairman of the Detroit Section PGA. Plans are afoot, stemming from a suggestion by A. B. Jacobson, Chairman of the Seattle Section PGA, to record on tape and distribute "tapescripts" of important talks on Audio to Sectional Audio Groups.

Thus, the Professional Group on Audio, along with other Professional Groups is rapidly becoming an integrated part of the whole which constitutes the IRE. In future years, the IRE historians may well agree that the creation of Professional Groups has been one of the most significant events in the history of the Institute.

Using Tests to Select Engineers*

WARREN G. FINDLEY†

Summary—**Experience in selecting students for admission to undergraduate engineering colleges provides a clear outline for a program of tests and related procedures that should prove helpful in identifying potential engineering talent early in high school. Qualified students may then be guided toward adequate preparation for engineering training. Such a program would include the following as a basic minimum: the students' average grades, tests of mathematical aptitude, reading comprehension, spatial visualization, and interest inventories.**

A recently revised program of examinations are now available for selecting students for graduate study in engineering. Research is being undertaken which gives promise of the development, in the measurable future, of a means of detecting creative talent for scientific research.

WHEN ARE TESTS HELPFUL?

THE VALUE of standardized tests for selecting engineers must be judged by the extent to which they improve selection over what can be done without tests by using other information routinely available or readily obtainable. That standardized tests will distinguish between superior and inferior applicants for engineering training or employment is not sufficient. Such tests must do the job better or add to what can be done by other methods (i.e., ready-made tests or evaluation of previous school records).

We may illustrate this point graphically by reference to Fig. 1. In the cross-hatched bottom line of each set of three show how well the five engineering colleges, by using the averages of high-school grades alone, can predict whether their students will make average grades or better. Thus, enrolled engineering freshmen whose high-school averages were in the top 4 per cent of the averages of all their classmates had 84 chances in 100 of doing average or better work in the first term of engineering college; those who stood in the next 12 per cent with respect to high-school averages had 75 chances in 100 of doing average work or better; and those who stood in the lowest 4 per cent with respect to high-school averages had only 16 chances in 100 of doing average work or better in the first semester of engineering college. The added effectiveness of prediction attained by using two mathematical tests is shown by the black lines in the same charts. Those who were in the top 4 per cent on a weighted composite based on high-school averages and scores on the two tests had 97 chances in 100 of doing average work or better in first-semester engineering; those who were in the next 12 per cent on the weighted composite had 88 chances in 100 of doing above average work in the first semester; while the bottom 4 per cent with respect to the composite had only 3 chances in 100 of doing average or better work during the first semester.

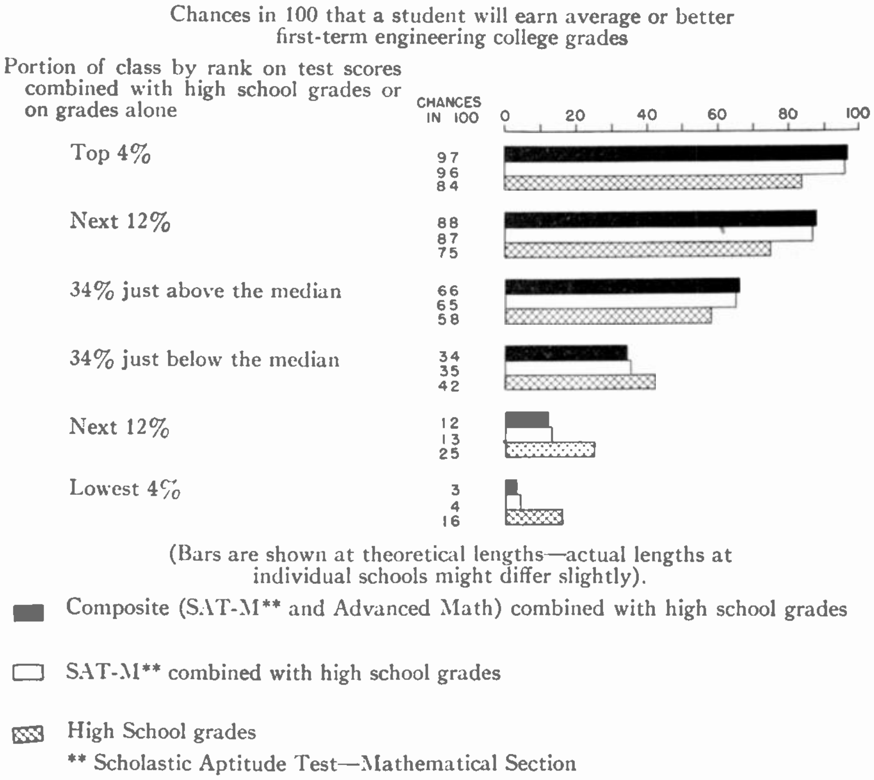

Fig. 1—Prediction of scholastic success in a group of five engineering colleges by college board test scores combined with high school grades.[1]

A corollary to the preceding statements is that standardized tests are particularly helpful where other sources of information about applicants for engineering training or employment, such as school records, do not afford comparable data on all applicants. This would apply especially to applications for graduate study in engineering at a university or applications for junior professional employment in a large engineering organization which are received from undergraduate colleges of all sorts with varied grading systems.

A second general proposition, applicable especially in these days of shortages of engineers and engineering students, is that tests and other information about applicants are most useful when the number of applicants is large relative to the number of openings for training or employment. In other times than these the author might be expected to devote a relatively large part of his remarks to selection for graduate training and employment because of their more natural place in the thinking of practicing engineers. These problems will be mentioned and discussed in their place. But the problem of the moment in selecting engineers is in identifying, at an early stage, potential engineering ability in high-school boys so that they may be guided into pre-engineering study. And so, confident that relations prevailing at the college entrance level of selection are most pertinent to the problem of selection (identification) in early high school, we turn to recent recorded experience in selecting students for undergraduate engineering colleges.

WHAT TESTS HELP MOST?

Over and above what can be ascertained by scrutiny of an applicant's previous school record, the most significant single factor to measure by tests is mathematical aptitude. This has been found consistently in studies of the Pre-Engineering Inventory (produced for the Measurement and Guidance Project in Engineering Education, sponsored jointly, since 1943, by the Engineers' Council for Professional Development and the American Society for Engineering Education), the American Council on Education Psychological Examination, the Engineering and Physical Science Aptitude Test, and the examinations of the College Entrance Examination Board.

A set of difficult arithmetic reasoning problems will serve very well. Additional problems in the fundamentals of algebra and geometry, though helpful, are not strictly necessary. The greatest contribution of the latter is that they require less reading time than corresponding arithmetic reasoning exercises. More advanced mathematics should be tested only if all applicants may be presumed to have studied approximately

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* Decimal classification: R070. Original manuscript received by the Institute, March 29, 1951.

† Director of Test Development for the Educational Testing Service, Princeton, N. J.

[1] Data, courtesy of Dr. W. B. Schrader, for 721 enrolled engineering freshmen tested during their first week in the Fall of 1948 at Carnegie Institute of Technology, Cornell University, Lehigh University, Rutgers University, and the University of Pennsylvania.

the same subjects. But in all cases the problems should be kept short, be based on the major topics of instruction, and require ingenuity in solution rather than memory of formulas or standard processes of solution. One writer has put it this way: "(A mathematical aptitude test) provides a measure of the slope of the student's learning curve on the subject, and success in freshman mathematics appears to be predictable from a knowledge of this slope regardless of the actual position on the curve." Of course, insofar as school records are lacking and certain formal mathematics training is presumed as a base for more advanced instruction, the above statement would have to be modified to include assessing the position as well as the slope. But when exposure to certain mathematical concepts is clearly indicated by academic records, the slope is the determining factor. In the study summarized in Fig. 1, the mathematical section of the Scholastic Aptitude Test, consisting of short arithmetic, algebra, and geometry problems, showed as good or slightly better correlation with mathematics grades and grade averages in engineering than did the test of corresponding length in advanced mathematics.

The second most generally significant factor to test in potential engineers is the ability to comprehend and interpret scientific reading material and data. Tests of this nature appear in most programs used in selecting or guiding potential engineers. They vary from tests of general reading comprehension in the humanities and social sciences as well as in the natural sciences to tests designed especially to include scientific passages and material presented in tables, graphs, and diagrams. Tests of ability to interpret quantitative data accurately and logically and to make reasonable inferences, deductions, and generalizations from data are in increasing demand.

Beyond these two major aptitudes a variety of other measures are useful for selective purposes whenever certain subjects may be presumed to have been studied by all applicants. For example, if a year of physics has been required of all applicants for entrance to an undergraduate engineering course, basic problems in that subject may well be set as part of the selective process.

Success in descriptive geometry, engineering drawing, surveying, and the like is especially dependent on abilities reflected in spatial-relations tests. Tests of comprehension of mechanical principles or movements in pictorially presented situations also add to our understanding of the applicants' background abilities, and in some measure improve the selection for engineering training.

If the task of selection is conceived more broadly as that of selecting intellectually and socially well-rounded persons for eventual admission to the engineering profession, tests like the one in the Pre-Engineering Inventory on Understanding Modern Society should be considered. At the graduate entrance level the so-called Profile Tests of the Graduate Record Examination and the Cooperative General Culture Test serve a similar purpose in measuring general understanding outside of one's immediate professional field.

It is perhaps worthy of special mention here that the Medical College Admission Test, required of candidates for admission to institutions in the Association of American Medical Colleges, includes a special section on Understanding Modern Society. Although this section will not approve prediction of success in medical training, the medical colleges support it strongly because it will indicate a breadth of viewpoint important in professional practice.

Further extension of the concept of desirable general characteristics should lead to the inclusion of measures of skill in human relations. At present no established group tests of emotional stability or personal adjustment can be recommended for general use although many promising instruments are effectively used by clinical psychologists in individual diagnosis. Studies of how well individual students are accepted by their contemporaries and chosen for leadership may well lead to the development of procedures that would permit objective statements about personal adjustment and leadership qualities. Ability to direct or supervise others would be especially worth exploring.

Advanced Tests

Recently, under the stimulus of the National Research Council's need for an advanced test in engineering for candidates for Atomic Energy Commission Fellowships, the so-called Advanced Test in Engineering of the Graduate Record Examination has been revised and expanded by a committee of five professors of engineering, nominated by the American Society for Engineering Education. Now a 3-hour examination, it consists of 100 general engineering problems followed by four groups of 25 difficult problems in the four chief traditional branches of engineering training; these latter sections are clearly labeled so that each examinee can start in the section in which he feels most competent. The more limited previous test proved useful in predicting success in graduate study of engineering; however, the revised test should provide a more searching, better-balanced instrument of selection and guidance.

The Advanced Test in Engineering is only one segment of the program of examinations developed for the National Research Council's program under the auspices of the Educational Testing Service. Corresponding tests in mathematics, physics, chemistry, biology, and geology have been developed by committees of outstanding graduate professors. A high-level aptitude test yielding separate verbal and mathematical aptitude scores is included. All these tests and the Profile Tests are available in the Graduate Record Examination programs for selection, appraisal, or guidance of graduate students in engineering.

Interests

For many years interest inventories have been widely accepted for use in guiding students and graduates into the broad areas of curriculum or employment. Such inventories are helpful when responses are given in a spirit of unbiased self-exploration. If used for selection in a single field, they are readily faked. Interest inventories, in conjunction with aptitude tests, should prove especially useful in identifying potential engineering talent early in secondary school.

The most widely used of these tests are the Strong Vocational Interest Blank and the Kuder Preference Record. The Strong Vocational Interest Blank, based on a careful study of the interests of men well established in their professions, has been in use since 1927 and has stood up well. The base group in engineering involved over 500 members of the engineering societies. The Kuder Preference Record is a relative newcomer, first published about 1940. It has been widely used in secondary schools, and an increasing number of studies show that its more general categories of interest (mechanical, computational, scientific, persuasive, literary, artistic, musical, social service, clerical, and outdoor activity) lend themselves to interpretations similar to those derived from the Strong Vocational Interest Blank. In addition, it has a handy self-scoring format. Both inventories suffer somewhat from the limitation of requiring an adult vocabulary. It is also true that interests change during adolescence, and become relatively more fixed at 18 or so. Experimental study of such variations might well be made an integral part of any major guidance program in secondary schools.

The effectiveness of these interest measures in counseling will depend considerably on the breadth of the counselor's background and the amount of additional information he has available on each student. As one example, a student who shows strong interest in science and has adequate mathematical ability might show secondary interest and ability in art. An alert counselor could point out the prospect of combining these talents in industrial design or architectural engineering.

New Approaches

Much is being said nowadays, and rightly, about the importance of identifying creative talent, or competence for scientific research. Instruments and procedures for this purpose have just begun to be developed. Work during World War II on selecting pilots for the United States Air Force led to confidence that certain elements of biographical information can predict success in training and leadership. This approach is being applied to certain kinds of scientific personnel in special studies. It depends on finding enough discriminating items of information from the testing of large numbers thought to be possibly related. In the case of Air Force pilots certain items about participation in sports and hobbies, items of specialized information that could only be gained from active participation in such activities as flying, driving an automobile, hunting, and the like, proved useful, when enough were brought together, in predicting pilot success.

Another approach is being made in a study by the American Institute for Research under the auspices of the Office of Naval Research. Descriptions have been secured from research personnel of critical incidents which led them to judge other research workers either outstandingly effective or downright inept. Aptitude tests have now been built based on 36 categories of types of critical incidents, and these may be found under the following major headings: formulating problems and hypotheses, planning and designing the investigation, conducting the investigation, interpreting research results, preparing reports, administering research projects, accepting organizational responsibility, and accepting personal responsibility. In the course of a few years' time these tests will be validated against measures of productive effectiveness.

Other tests being experimented with include one in which the examinees are asked to describe "what would happen if" some fundamental change in the physical or biological laws were to transpire. Another requires the examinee to show a fluency of ideas by classifying 25 items of a list into as many sets of four as can be identified with any basic principle in science. This latter test is based on the theory that creative research demands, among other talents, an ability to conceive and consider great numbers of possible relationships very rapidly. Still another type of test poses the question, "Do you have enough data to solve?," rather than merely "Solve the following," as evidence of ability to grasp the essential features of problems and of willingness to pronounce tough-minded judgments; this is in contrast to the usual requirement of ingenuity in operations.

Our greatest hopes are that these and other research studies will reveal basic intellectual and motivational factors which are not measured in current tests but are useful in identifying outstanding scientific and engineering talent.

ACKNOWLEDGMENTS

The author wishes to express his indebtedness to A. P. Johnson of the Educational Testing Service for data and references used in preparing this paper and to E. M. Rickard for aid in digesting the references and preparing the bibliography.

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1. H. S. Rogers and G. W. Holcomb, "Inventory of engineering motives," *Jour. Eng. Ed.*, vol. 22, pp. 851–860; June, 1932.

   This early biographical inventory was designed to reveal the "intensity of motives" in respect to:
   1. "Constructive Instinct"
   2. Scientific Curiosity in Mechanisms
   3. Abstract Ideas of a Quantitative Nature
   4. Precise Observation and Description of Physical Things
   5. Practical Application
   6. Planning and Organization
   7. Manual Activity and Craftsmanship.

2. W. V. Bingham, "Aptitudes and Aptitude Testing," Harper and Bros., New York, N. Y.; 1937.

   Part I includes a broad discussion of aptitude, intelligence, and interest. In Part II, in discussing the field of engineering, attention is focused on the following:
   1. General Scholastic Aptitude
   2. Aptitude for Mathematics
   3. Aptitude for Thinking about Space Relations
   4. Aptitude for Understanding Mechanisms
   5. Aptitude for Mastering Physical Sciences.

   The statement is made that superiority in tests known to be indicative of these aptitudes, as well as a liking for engineering work, the necessary health, and constancy of purpose, indicate a high probability of success. Low scores should not be construed as barring an engineering career but as warning signals.

   The book contains a very complete appendix of representative tests and interest schedules.

3. I. L. Kandel, "Professional aptitude tests in medicine, law, and engineering," Bureau of Publications, Teachers College, Columbia University, New York, N. Y.; 1940.

   This book reviews the early development of tests for engineering ability, and the part taken by the Society for the Promotion of Engineering Education and the Engineers Council for Professional Development.

4. E. N. Brush, "Mechanical ability as a factor in engineering aptitude," *Jour. Appl. Psych.*, vol. 25, pp. 300–312; 1941.

   The conclusion is reached that mechanical ability, as measured by several tests studied, may be regarded as a component of engineering aptitude, but the actual predictive power of most single tests of mechanical ability is not great.

5. C. H. Griffin and H. Borow, "An engineering and physical-science aptitude test," *Jour. Appl. Psych.*, vol. 28, pp. 376–387; 1944.

   This describes a battery of 6 tests developed to assess suitability for training in technical work on the college leve. It states that multiple correlations between the weighted battery of test parts and course achievement were as high as 0.79.

6. J. P. Guilford and J. I. Lacey, Ed., "Printed Classification Tests," Report No. 5 of AAF Aviation Psychology Program Research Reports, U. S. Government Printing Office, Washington, D. C.; 1947.

   Chapter 27 on Biographical Data summarizes the most significant research prior to United States Air Force experimentation, including aeronautical and industrial sources.

7. H. A. Edgerton, S. H. Britt, and R. D. Norman, "Later achievements of male contestants in the first annual science talent search," *Amer. Scien.*, vol. 36, pp. 403–413; 1948.

   The article describes briefly the first annual science talent search sponsored by the Westinghouse Electric Corporation, a program that has been carried on each year since, and it discusses the follow-up study of the later achievements of the contestants. Other pertinent references to the program are given at the end of the article.

8. E. C. Koerper, "Milwaukee plan of aptitude testing," *Mech. Eng.*, vol. 70, p. 750; September, 1948.

   This outlines a plan to develop the following:
   1. Test Batteries
   2. Interview Forms
   3. Job Specification System
   4. Employee Progress Appraisal Methods.

9. D. B. Stuit, G. S. Dickson, T. F. Jordan, and L. Schloerb, "Predicting Success in Professional Schools," Amer. Council on Ed., Washington, D. C., 1949.

   Chapter II summarizes detailed research findings about predicting success in engineering schools.

10. D. E. Super, "Appraising Vocational Fitness," Harper and Bros., New York, N. Y.; 1949.

    This is a general manual of tests and testing. The discussion of interest inventories, personality measures, and biographical data blanks is particularly good because of the author's experimental studies in these areas.

11. W. T. Donahue, C. H. Coombs, and R. M. W. Travers, Eds., "The Measurement of Student Adjustment and Achievement" (A collection of papers presented at a conference of guidance workers), University of Michigan Press, Ann Arbor, Michigan; 1949.

    A paper, entitled "Significant research on the prediction of academic success," by R. M. W. Travers points out that tests to predict success should be given as early in the student's career as possible. Studies summarized indicate that predicted grades in the first 2 years of engineering school may be expected to correlate between 0.6 and 0.7 with actual grades if predictions are made from tests in mathematical and scientific fields. A bibliography of 272 studies is given.

    This volume also contains pertinent material in a paper entitled "Problems in guidance in vocational and technical training," by G. K. Bennett.

12. L. W. Guth, "Discovering and developing creative engineers," *Mach. Design*, vol. 21, pp. 89–94; March, 1949.

    This article describes tests used in a survey conducted at the General Electric Company. The tests, designed for selection of creative engineering talent, attempt to appraise adaptive inventiveness, design sense and perception, and reasoning power. Solution of the problems is dependent on application and ingenuity rather than on memory of formulas, which are supplied where needed.

13. J. C. Flanagan, "Critical Requirements for Research Personnel," Amer. Instit. for Research, Pittsburgh, Pa.; March, 1949.

    This describes the first phase in a broad program aiming to develop tests and evaluation procedures for research personnel for the Office of Naval Research. (The same program is referred to under M. H. Weislogel in reference 16.)

14. E. J. Riegel, "Evolution of math tests," *College Board Rev.*, vol. 1, p. 95, New York, N. Y.; May, 1949.

    This includes a description of a validity study made in cooperation with five large engineering schools.

15. O. W. Eschbach, "Report of the committee on selection and guidance," *Jour. Eng. Ed.*, vol. 40, pp. 112–119; October, 1949.

    This is a fairly recent report on the testing program of the Measurement and Guidance Project in Engineering Education, sponsored jointly, since 1943, by the Engineers' Council for Professional Development and the American Society for Engineering Education. During the year 1946–1947, 32,098 students were tested in this program.

16. M. H. Weislogel, "The Development of a Test for Selecting Research Personnel," Amer. Instit. for Research, Pittsburgh, Pa.; January, 1950.

    This report describes test items that have been pretested and put into final form available for use in the later phases of a long-range research program for the Office of Naval Research. (The same program is referred to under J. C. Flanagan in reference 13.)

17. F. Lord, J. T. Cowles, and M. Cyna-

mon, "The pre-engineering inventory as a predictor of success in engineering colleges," *Jour. Appl. Psych.*, vol. 34, pp. 30–39; February, 1950.
The analyses indicate that the Pre-Engineering Inventory Composite Score performs very satisfactorily, on the whole, as a predictor of engineering school success.

18. R. F. Berdie and N. A. Sutter, "Predicting success of engineering students," *Jour. Ed. Psych.*, vol. 41, pp. 184–190; March, 1950.
Report is made on a study of 1,019 freshmen in the Institute of Technology of the University of Minnesota.
19. A. P. Johnson, "Tests and testing programs of interest to engineering educators," *Jour. Eng. Ed.*, vol. 41, pp. 277–283; January, 1951.
The author reviews recent activities under the Measurement and Guidance Project in Engineering Education and the current problems in finding prospective engineers and selecting engineering students, including selection at the graduate level.

Fundamentals of Secondary Electron Emission*

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This paper has been secured by the Tutorial Papers Subcommittee of the IRE Committee on Education as a part of a planned program of publication of valuable tutorial material. It is here presented with the approval of that Subcommittee.—*The Editor.*

Summary—**Secondary electron emission is of great importance to the physicist because of its bearing upon the problem of the interactions between fundamental particles and to the radio engineer because of its applications, as well as its effect, upon the operation of electronic tubes. A complete theoretical picture capable of accounting quantitatively for all observed phenomena does not exist. Secondary electron emission differs from other modes of emission in many respects. The essential characteristics can best be evaluated by considering a typical experimental arrangement for investigating the phenomenon. Three categories of emitted electrons are recognized. The yield may depend upon various factors, such as the primary energy, collector voltage, target temperature, time, angle of incidence, atomic properties of target, and the composition of the target.**

The difficulties of propounding a satisfactory theory are evident from an individual consideration of each of the various processes involved. The primary interaction, primary energy loss, escape of secondaries, and integration over the range of the primary must each be treated to arrive at a final solution. In several previous attempts at formulating a theory, only the most loosely bound electrons in the solid have been regarded as constituting the source of secondary electrons. Normalizations are required for comparison of the results with existing experimental data. There are cogent reasons for regarding bound electrons as a very important source of secondaries. The probability of ionization in gases exhibits the same general dependence upon primary energy as secondary electron emission, and this resemblance suggests a possible model for secondary emission based upon detailed considerations of primary ionization probabilities.

I. INTRODUCTION

WHEN A SOLID BODY is subjected to bombardment by electrically charged particles, some electrons which may be detectable under suitable circumstances are always emitted. Although this process, commonly designated "secondary electron emission," has been observed to occur in various forms, by far the most widely investigated type is that in which an electron beam falling upon the surface of a target in a vacuum causes the emission of a stream of electrons from the surface upon which it impinges. It should be emphasized, however, that this variety of secondary emission is not endowed with any intrinsically greater significance than any other. Rather, the distinction arises solely from the geometrical and practical circumstances that in this case the phenomenon is readily observable and is, in fact, involved in the operation of common electronic devices.

In the field of radio engineering, secondary electron emission originally manifested itself only as a source of annoyance which seriously interfered with the satisfactory functioning of vacuum tubes. The problem was solved by the addition of the suppressor grid or its equivalent to the tetrode; after the advent of the pentode, the effect received relatively little attention. In recent years, however, successful attempts, both unintentional and conscious, have been made to utilize the phenomenon to some advantage. It is now generally realized that secondary electron emission is inherent in the operation of a cathode-ray tube. Devices such as the magnetron and certain types of reflex klystron depend upon secondary electron emission for their high output capabilities. In dynatron and photomultiplier tubes, secondary electron emission constitutes the fundamental principle of operation.

Electron emission may be divided into four principal categories: (1) thermionic emission, (2) photoelectric emission, (3) field emission, and (4) secondary emission. Despite the radical differences among these modes of emission, certain similarities exist among the first three. In general, satisfactory theories have been formulated (although complicated systems, such as activated

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barium-strontium oxide-coated cathodes, have been the subject of considerable controversy) despite the fact that they may be unable to account for all observed phenomena quantitatively (as in the case of the photoelectric effect). Nothing resembling a complete theoretical picture of secondary electron emission exists up to the present time, nor is it likely that an all-inclusive model can be developed in the light of the limitations of our present grasp of physics of the solid state.

The most striking characteristic of secondary emission as contrasted with the other types is the similarity of behavior over a wide variety of materials. The range of values of yields encountered is quite limited as compared with thermionic or photoelectric yields. The nature of the dependence of secondary emission upon the work function of the surface differs appreciably from that in the other cases, and the yield is actually quite insensitive to the nature of the barrier. At first sight it might be expected that secondary emission is more closely related to photoelectric emission than to thermionic emission. This resemblance, which happens to be valid in a certain sense for rather subtle reasons, cannot be accorded much significance upon closer scrutiny. Whereas photoelectric emission from a solid is primarily a surface effect, this is certainly not the case for secondary emission. A single photoelectron absorbs all of the energy of the incident quantum *hν*, and hence the situation is dominated by the work function *φ*. On the other hand, each secondary electron absorbs only a small fraction of the energy of the primary which may penetrate a considerable distance into the target material. Thus, we are here concerned with a combination of volume and surface effects.

From the standpoint of tube engineering, the operational problems associated with secondary electron emitters are to some extent somewhat less complicated than those encountered in applications utilizing thermionic emission. It is appropriate to emphasize at this time that the generalizations to which the discussion in this paper will be confined are occasionally subject to exceptions. In the event that these are not specifically mentioned, it should not be tacitly assumed that the broad statements are necessarily all-inclusive. For example, there are certain specific applications for which the requirements of reproducibility and long-time stability of specially prepared targets present very difficult technical problems. Except for a few composite surfaces with certain desired characteristics, activation procedures are not involved. As in the case of thermionic emitters, there are two general classes of materials which are useful as a source of electrons, namely, metals and semiconductors. A third class, broadly termed "insulators" for lack of a less ambiguous designation (note that semiconductors under some conditions are included in this group), is also of importance primarily because of practical applications, although the first two present greater theoretical interest.

II. Basic Experimental Concepts

The most direct approach to an evaluation of the significant characteristics of secondary electron emission is to consider a typical experimental arrangement for investigating this phenomenon. Fig. 1 is a schematic diagram of such an apparatus. It consists essentially of an electron gun G serving as the source of a beam of primary electrons I_p which, after acceleration through a difference of potential V_p, bombards the target T. The secondary electrons I_s leaving the target are then attracted to the collector C, owing to the presence of the positive voltage V_c applied to the collector with respect to the target.

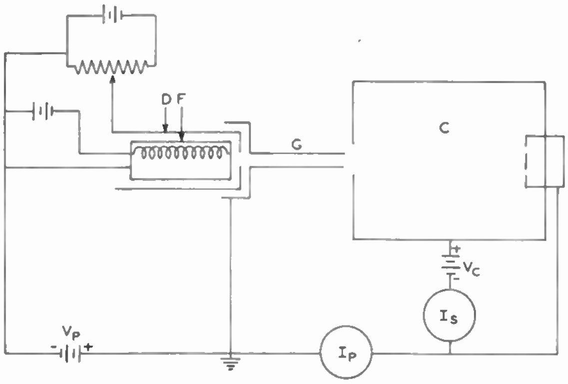

Fig. 1—Schematic diagram of a typical experimental arrangement for measuring secondary electron emission.

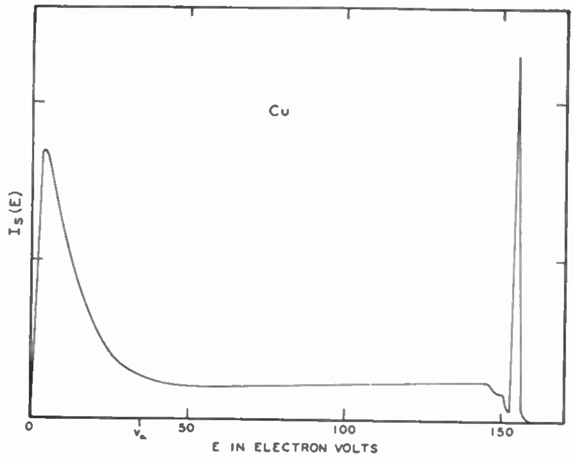

Fig. 2—Typical energy distribution of secondary electrons. This particular curve, obtained by a method utilizing a transverse magnetic analyzer, refers to a Cu target. (See E. Rudberg, *Phys. Rev.*, vol. 50, p. 138; 1936.)

The yield δ is defined as the number of secondaries *collected* per incident primary, given by the ratio I_s/I_p. It should be noted that this is a purely empirical definition, but nevertheless one of practical interest in any event. Some of the primary electrons are directly reflected, whereas others are scattered with some loss of energy. These do not constitute true secondary electrons which have actually been knocked out of the target by an impinging primary. It is practically impossible to distinguish between true secondaries and primaries reflected after suffering energy losses, and this is indeed unfortunate from the point of view of obtaining theoretically useful experimental data. However, the distinction is unnecessary as regards obtaining "emission" from a target regardless of the origin of the "emitted" electrons.

It has been stated that the secondary electrons emitted by a material bombarded by a primary beam can be ascribed to three different mechanisms. This classification is made on the basis of the energy distribution of collected electrons, which is obtained by applying retarding potentials $-V_c$ to the collector with respect to the target, taking into account the contact difference in potential between the surfaces of these two electrodes. As is evident in Fig. 2, there is a sharp distinguishable peak at the energy of the incident beam corresponding to eV_p, and these electrons are, of course, to be identified as elastically reflected primaries. At the other end of the spectrum a group of slow electrons may be observed. The average energy of these electrons is only a few ev when the primary energy is of the order of hundreds of ev. Between these electrons and the elastically reflected primaries there is an intermediate group which results principally from inelastic scattering in the lattice. The slow electrons arise from collisions between the primaries and the atomic electrons of the target in which sufficient energy is transferred to the latter so that they can penetrate to the surface and emerge from the solid material. These electrons are therefore true secondaries; although most possess low energy, it is certainly not warranted, however tempting, to assign an arbitrary upper limit, such as V_s, above which no true secondaries appear and below which scattered primaries are prohibited.

Various factors in addition to the nature of the target material may affect the magnitude of the yield of secondary electrons. These will be discussed briefly in a general manner, and certain special situations will be mentioned.

A. Primary Voltage V_p

All known secondary emitters manifest the same qualitative dependence of yield upon primary energy. Starting at low voltage, the yield rises smoothly until a maximum value, often in the neighborhood of 400 to 600 volts, is attained. Thereafter, the yield decreases slowly and may approach a more or less constant value at very high energy. A typical curve is shown in Fig. 3. The maximum value of the yield, $\delta_{\max}$, is often cited, probably because it is convenient, at least from the practical point of view and for some purposes, to know the highest multiplication which could be expected under optimum conditions. The corresponding voltage is designated $V_{p_{\max}}$. Although $\delta_{\max}$ and $V_{p_{\max}}$ alone are not necessarily of any fundamental theoretical significance, it is interesting that at least in the case of metals a universal curve which fits the available data within the experimental errors is obtained by applying a normalization in which the ratio $\delta/\delta_{\max}$ is plotted as a function of $V_p/V_{p_{\max}}$. This is shown in Fig. 4. The shape of the *yield versus energy* relationship can be accounted for, at least qualitatively, as will be described in the following section.

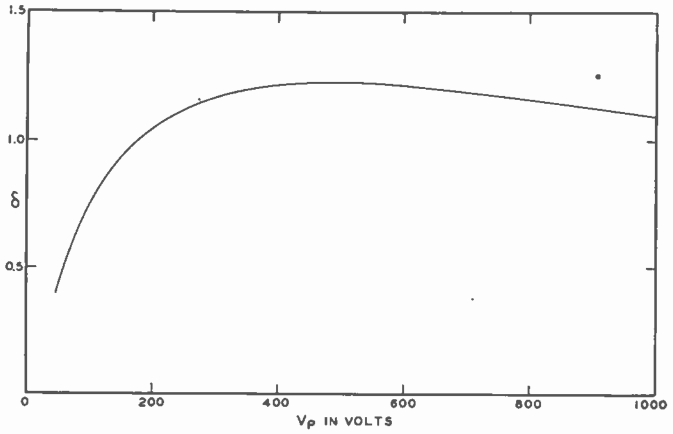

Fig. 3—Dependence of secondary electron emission upon primary bombarding energy for a typical metal, in this case Ni.

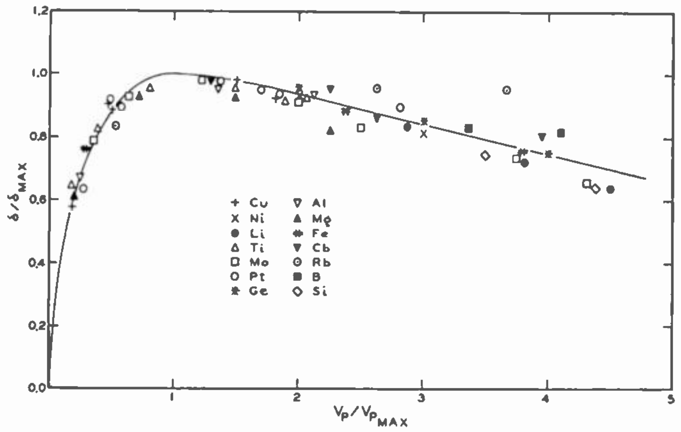

Fig. 4—Normalized yield, $\delta/\delta_{\max}$, plotted as a function of normalized primary energy, $V_p/V_{p_{\max}}$, for various elements. This is the so-called universal curve[5] of secondary electron emission.

Other values of primary voltage sometimes cited because of practical considerations, particularly in the case of insulators, are the two crossover points, in the δ versus V_p plot, at which the yield attains the value unity. It is evident that if an insulator is subjected to electron bombardment the surface will charge negatively, as long as the yield is less than one secondary per incident primary, until it approaches the cathode potential, thereby effectively reducing the primary bombarding energy. When the yield exceeds unity, between the so-called "lower and upper sticking potentials" a positive charge is acquired by the surface, thereby reducing the effective collector voltage V_c until the measured yield approaches one. Above the upper crossover voltage, the surface becomes negatively charged until the yield again approaches unity.

B. Collector Voltage V_c

In the case of metals, the secondary current is independent of the collector voltage as long as this is positive. For substances with lower conductivity, such as certain oxides, on the contrary, an increase of yield with increasing collector voltage is sometimes observed, as is illustrated in Fig. 5. This effect may be qualitatively understood when it is recognized that by virtue of a yield larger than unity more electrons leave the emitter than enter. As a consequence of the high resistance of the oxide layer, a positive charge will appear on the outer surface and a negative charge on the inner surface, thereby forming a double layer. The resulting field promotes a phenomenon analogous to cold emission. The dependence upon collector voltage arises from the fact that the potential on the outer surface of the oxide layer is limited to values smaller than the collector voltage. An increase in the latter is accompanied by an increase in the internal field.

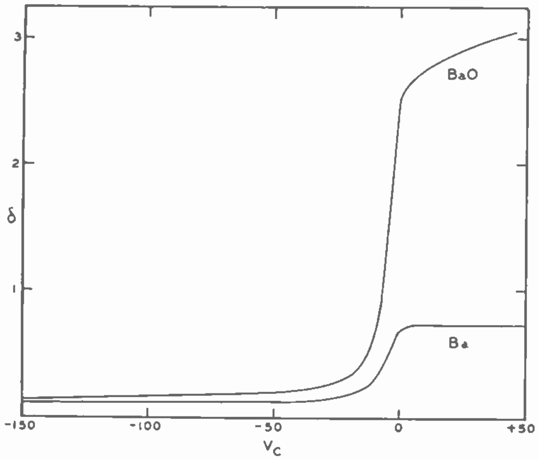

Fig. 5—Dependence of yield upon collector voltage, illustrating lack of saturation[3] in the case of BaO. (See H. Bruining and J. H. De Boer, *Physica* (The Hague), vol. 6, p. 823; 1939.)

C. Temperature of Target

The secondary electron emission from metallic materials is not ostensibly dependent upon the temperature of the target, at least in any fundamental manner. Actually, changes in the nature of the surface layer or in the crystal structure may be introduced by heat treatment, thereby producing only small changes in secondary emission. In the case of at least one semiconducting medium, which in particular has been investigated rather extensively because of its great practical importance—the so-called oxide-coated cathode—, appreciable variations with temperature do occur.[1,2]

D. Time

At least in the case of metals, there are no essential changes of secondary electron emission with time, except for obvious consequences of structural changes which the target surface may undergo during the lifetime of the tube. Specialized emitters consisting, for example, of thin films of aluminum oxide on an aluminum base, with an outer layer of Cs, operating by virtue of the so-called "Malter Effect,"[3] are capable of emitting thousands of electrons per bombarding primary. This phenomenon, also termed "thin-film field emission," is produced by an extreme manifestation of the mechanism already described in Section II B. It is evident that it might be anticipated from the nature of the process that the maximum emission is not attained until the primary beam has been on for some time and, furthermore, that emission may persist after the beam has been turned off; this is quite contrary to the situation with ordinary secondary electron emission which displays no detectable time delays.

E. Angle of Incidence

The data plotted in Fig. 3 were obtained with the beam striking the target surface perpendicularly. As might be expected from the general nature of the processes involved, the yields for oblique incidence are somewhat larger inasmuch as the secondary electrons are formed closer to the surface and are consequently absorbed to a lesser extent before reaching the surface barrier.

For nearly grazing incidence, the yield may be increased by a factor of as much as three, depending upon the primary voltage and the composition of the target. In general, the maximum value of the yield and the energy at which it occurs are both higher, which is consistent at least with qualitative expectations.

F. Atomic Properties of Target

No simple correlation between the secondary yield and the known atomic properties of the target exists, as in the case of other types of electron emission. In some instances, trends which are certainly suggestive are revealed, although the relationships are evidently indirect. For example, the curve in Fig. 6 shows the correlation between δ_{max} and ϕ. The positive slope is opposite that which would be expected at first sight, and the suggestion of a correlation is a consequence of the fact that the work function changes along with some other atomic property which really predominates the second-

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[1] M. A. Pomerantz, "Secondary electron emission from oxide-coated cathodes," *Jour. Frank. Inst.*, vol. 241, p. 415; vol. 242, p. 41; 1946.

[2] J. B. Johnson, "Secondary electron emission from targets of barium-strontium oxide," *Phys. Rev.*, vol. 73, p. 1058; 1948.

[3] L. Malter, "Thin-film field emission," *Phys. Rev.*, vol. 50, p. 48; 1936.

ary emission process. Actually, a reduction in ϕ, introduced without alteration of the bulk material, would result in an increase in the yield.

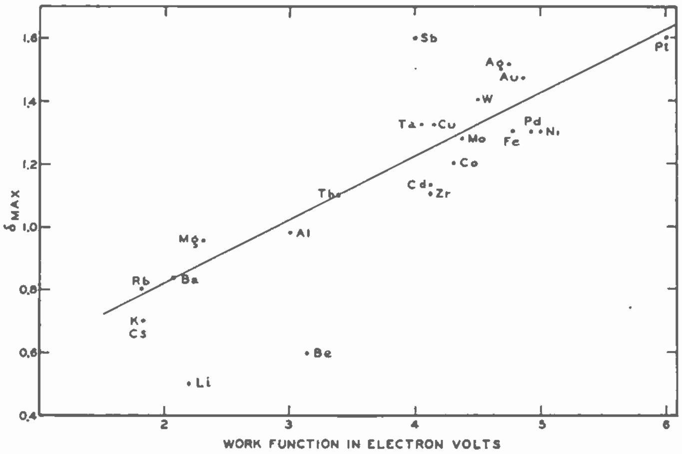

Fig. 6—Correlation between maximum secondary electron emission and work function for various metals.[3a]

Another illustration of the sort of trend which may be exhibited is shown in Fig. 7, where the yield is plotted as a function of the first atomic ionization potential of the target element. Now obviously, this characteristic itself retains no significance when the atoms are brought together to form a solid. However, it is not unreasonable to assume that it may reflect the effective number of conduction electrons in the metal (high-ionization potentials corresponding to small effective numbers) and hence the number of electrons which will impede the progress of an outgoing secondary. Thus, the addition of an electron to the outer shell should decrease the yield. It has not yet been possible to test this hypothesis experimentally because of the inavailability of the requisite data.

G. Composition of Target

It was noted in the introduction that secondary electron emission is unique as contrasted with the other types, in view of the relative insensitivity to the nature of the emitting material. It is indeed remarkable that values of δ_{max} (includingspecial unstable cases of high yields) vary only by a factor somewhat greater than one order of magnitude. It is appropriate to include herewith at least a general statement regarding the values of yield which are encountered in practice.

Secondary electron emitters may be classified into four categories as follows:

(1) Elements: This group includes all chemical elements normally in the solid state for which measurements have been reported. The maximum values of the yield range from about 0.5 to 1.6 for clean surfaces, regardless of the specific nature of the conductivity of the solid.

(2) Compounds: This group includes all chemical compounds which are designated either as semiconductors or as insulators at room temperature. The maximum values of the yield range from aproximately 1.0 to 7.5.

(3) Composite surfaces: These are various complicated systems, sometimes designated "photocathodes" because they are characterized by very high photoelectric sensitivity, usually prepared by evaporating layer upon layer of different materials in a vacuum and by performing other special operations. For example, the notation [Ag]—Cs_2O, Ag—Cs refers to an electrode consisting of a silver base covered with a Cs_2O layer (also containing Ag atoms), on the surface of which Cs atoms are absorbed. The yields range between 3 and 10, in general.

(4) Activated alloys: Certain alloys, for example several per cent Mg with Ag, when "activated" by what appears to be an oxidation procedure, produce yields as high as 18, without any stability. Yields as high as 4 to 5 can be maintained.

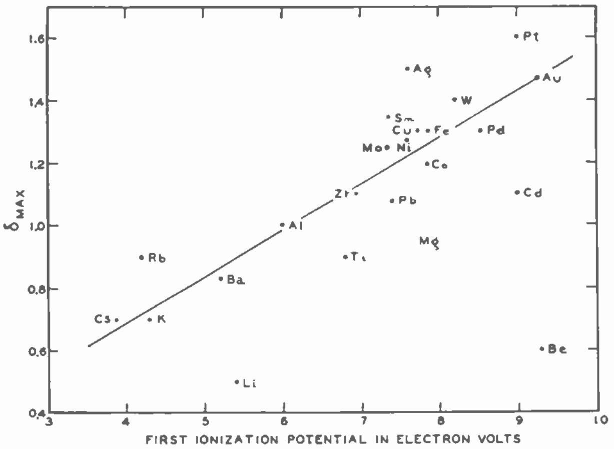

Fig. 7—Correlation between maximum secondary electron emission and first atomic ionization potential for various elements.

III. THEORY OF SECONDARY EMISSION

In contrast to the theory of thermionic emission, the theory of secondary emission is in a very unsatisfactory state. This is a consequence of the fact that whereas the problem of thermionic emission can be treated in a fairly straightforward manner by standard techniques of statistical mechanics[4] that of secondary emission is highly complex, involving several processes none of of which is well understood. The difficulties involved become obvious from an examination of the following outline which lists the steps in the computation of the secondary yield from a solid.

A. The Primary Interaction

The problem here is to compute the probability that a primary electron of a given energy will interact with one of the electrons in the solid to produce a sec-

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[3a] K. G. McKay, "Advances in Electronics," Academic Press, Inc., New York, N. Y., vol. I; 1948.

[4] W. E. Danforth, "Elements of thermionics," PROC. I.R.E., vol. 39, p. 485; May, 1951.

ondary electron with sufficient energy to emerge. Our present state of knowledge of solids is such that it is certainly impossible to solve even this very fundamental problem in any exact way. It is not even clear, for example, whether the sources of secondary electrons in a metal are the conduction electrons or the more tightly bound inner-shell electrons, a complication that does not exist in thermionic emission since it is known in that case that only the conduction electrons can be involved. Any attempt to discuss the question of primary interaction thus involves the introduction of certain simplifying assumptions about the solid, and even then the problem is generally extremely complicated.

B. Primary Energy Loss

The probability of producing a secondary electron of a given energy certainly depends upon the energy of the primary that produces it. Consequently, one must have some knowledge of the energy of the primary as a function of the depth to which it has penetrated. Since there is no experimental evidence available on the rate of energy loss of low-energy primaries, it must be computed theoretically. Inasmuch as the primary can lose energy in several ways (production of secondaries, excitation of bound electrons, and the like), such a computation is very difficult.

C. Escape of Secondaries

In order for a secondary to be observed, it must escape from the surface of the solid. To accomplish this, it must move through the body of the solid from the point at which it was created to the surface, retaining sufficient energy to penetrate the surface potential barrier. Although the problem of the penetration of the barrier can be treated quite adequately, the motion of a slow secondary through the solid is not at all well understood. Several possible assumptions can be made, such as exponential absorption, random diffusion, uniform energy loss, and the like, but whether any of these assumptions are valid is not at all clear.

D. Integration over the Range of the Primary

The solution of the above problems will furnish information regarding the number of emergent secondaries which are produced at a given depth in the solid. In order to obtain the total yield, it is necessary to integrate this result over all possible depths, i.e., from zero to the total range of the primary. If steps (A), (B), and (C) have been solved, this process can always be accomplished numerically if necessary, and consequently this does not represent an essential difficulty.

As may be concluded from the above discussion, the general problem of secondary emission is very complicated, and at present the only feasible approach involves adopting various simplifying assumptions and investigating the agreement between theories based on these approximations and experiment. Several theories of secondary emission from metals have been developed in this manner. Of these we shall discuss the theories of Baroody and Wooldridge which, though not the only attempts, are quite typical of the usual approach, since both assume that the loosely bound valence electrons constitute the principal source of secondaries.

Baroody's theory,[5] which is somewhat simpler, employs the Sommerfeld model of the metal, as is done in the theory of thermionic emission. The interaction between the primary and secondary electrons is treated in a purely classical manner, and the secondaries are assumed to be absorbed exponentially in their passage to the surface. This theory, in common with all existing theories of secondary emission, involves several parameters whose magnitudes are unknown, and consequently the values of the secondary yields to be expected from metals cannot be computed in absolute terms. A relation between yield and energy is obtained, however. Baroody points out the existence of the universal curve shown in Fig. 4, relating experimental values of the secondary yield and energy, and compares his theoretical results with it. Unfortunately, although the theoretical curves have the same general form, the quantitative agreement is very poor. For example, for $V_p/V_{p_{\max}}=4.5$, the experimental value of $\delta/\delta_{\max}$ is three times the computed value. This lack of agreement should not cause great concern, however, inasmuch as this application of the Fermi gas model is exceedingly questionable, and the aim of the investigation was primarily to demonstrate certain qualitative features of secondary emission.

Wooldridge's theory[6] is similar to Baroody's in that it considers only the valence electrons as potential secondary electrons, and assumes an exponential law for their absorption. It differs from Baroody's theory in treating the primary interaction quantum mechanically and taking into account the interaction of the valence electrons with the lattice. Again, absolute values of yields cannot be determined, and for comparison with experiment, the theoretical value of $\delta_{\max}$ is set equal to the empirical value. Within the expected range of validity of the formulas, the yield curves thus obtained agree with experiment quite well for the dense metals, such as silver and copper; but for less dense substances, such as lithium and aluminum, the agreement is rather poor. Wooldridge assumes that the primary loses energy only by the production of secondaries, and attributes the disparity with experiment in the case of light elements to the neglect of other types of energy loss. It is quite possible, however, that this disagreement arises from errors inherent in the basic assumptions.

Both of the aforementioned theories are based upon the hypothesis that valence electrons are the principal source of secondaries, and at sufficiently low primary

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[5] E. M. Baroody, "A theory of secondary electron emission from metals," *Phys. Rev.*, vol. 78, p. 780; 1950.

[6] D. E. Wooldridge, "Theory of secondary emission," *Phys. Rev.*, vol. 56, p. 562; 1949.

energies, this is undoubtedly the case. At very high primary energy, on the other hand, bound inner-shell electrons are the principal source since their binding is then very small compared to the energy of the incident primary and since they are much more numerous than the valence electrons. At intermediate energies (approximately 100 to 2,000 volts) it is not at all obvious that one group of electrons or the other necessarily plays the dominant role. The work of Baroody, Wooldridge, and others shows that by proper choice of parameters, the yield versus energy curves for metals can be explained qualitatively on the valence electron assumption, and for some metals, as mentioned above, good quantitative agreement is even obtained.

The hypothesis that the bound electrons may be the principal source of secondaries has not as yet been investigated quantitatively; however, there are cogent qualitative arguments indicating that this may be the case. (1) Bound electrons which could be available as secondaries are much more numerous than the valence electrons in most metals. (2) The ionization of gases has been investigated experimentally, and it has many features in common with secondary emission from metals. In particular, if the probability of ionization σ of a gas molecule is plotted as a function of the energy W of the incident electron, a curve having the same general form as the secondary yield curve shown in Fig. 3 is obtained. Furthermore, if this curve is normalized by plotting $\sigma/\sigma_{\max}$ against $W/W_{\max}$, a universal curve which is very similar to the universal emission curve results, as is seen in Fig. 8.

This resemblance suggests a possible model for secondary emission. If the bound electrons in a solid are the principal source of secondaries, the production of an internal secondary will be essentially an ionization process. Hence the probability of production should have much the same dependence on energy as the probability of ionization of gases. Although many factors affect the shape of the secondary yield versus energy relationship, it is quite possible that the primary interaction is the dominant factor, in which case the universal secondary emission curve and the universal ionization curve should be very similar. An examination of Fig. 8 reveals that this is indeed the case. Furthermore, rough calculations indicate that if one modifies the gas curve to take into account the fact that only those electrons with sufficient energy to penetrate the surface barrier can be observed as secondaries, the resultant curve will be in even better agreement with the secondary emission curve.

Consequently, it seems important to investigate in more detail the possibility that the bound electrons may be an important source of secondaries. It should be pointed out that whether or not the conduction electrons in a metal are a copious source of secondaries they are certainly of great importance in the emission process since it is very likely that interactions between internal secondaries and the conduction electrons are responsible for most of the secondary absorption which so drastically limits the secondary yield of metals. It might be expected that, in the case of metals, filling an inner-shell in progressing through the periodic system should increase the secondary electron emission, whereas the addition of an electron to the outer-shell should decrease the yield. The absence of conduction electrons in insulators thus accounts for their high secondary yields.

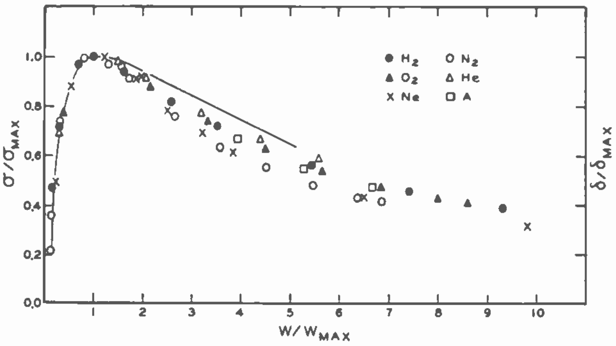

Fig. 8—Normalized primary ionization probabilities, $\sigma/\sigma_{\max}$; plotted as a function of normalized primary energy, $W/W_{\max}$, for various gases. The solid curve is the universal curve of secondary electron shown in Fig. 4.

IV. CONCLUSION

Many specialized features of secondary electron emission have not been mentioned in this paper, and it has been possible to include only a brief summary of the salient features of the phenomenon, in conformity with the primary purpose of this review. Frequent references and inclusion of detailed descriptions of experiments have necessarily been avoided in the interests of stressing general principles rather than presenting an encyclopedic survey of the literature. For this reason, descriptions of applications of the phenomenon have likewise been omitted. For additional facts, the reader is referred to the lengthier articles containing rather comprehensive bibliographies.[7,8]

In conclusion it seems appropriate to hazard a prediction regarding the course of future progress in this field. Certainly the known engineering goals are well defined; efforts to obtain high yields with good stability will undoubtedly be continued, together with the search for new applications of the effect. From the scientific point of view, there is need for a self-consistent series of reliable measurements on all chemical elements in the periodic system which can be made into suitable targets. The disagreement among the results of different experiments is such as to preclude many crucial comparisons which could cast light upon the nature of the mechanisms involved in the process of secondary emission.

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Propagation at 412 Megacycles from a High-Power Transmitter*

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Summary—**Extended measurements are reported which indicate the existence of pronounced nocturnal superrefraction during an appreciable percentage of the summer and of very persistent scattering by atmospheric turbulence near the surface in all seasons. The measurements were taken over rolling midwestern terrain at a distance of about 100 miles. Mobile road tests were made to supplement the fixed-point measurements and to provide an approximate indication of the relation between field strength and distance. Aerial tests were made to show the effects of antenna height at large distance. Graphs are provided which show the effects of distance, terrain, antenna height, and time upon the field strength. The practical significance of the results in the broadcast and communication fields is indicated.**

INTRODUCTION

THE CALCULATION of tropospheric field strength was first satisfactorily treated by making the simplifying assumptions of a smooth, spherical earth and a linear variation of refractive index of the medium with height above the surface.[1,2] As a result of numerous short-wave propagation measurements made prior to about 1940, it was recognized that atmospheric irregularities may produce major anomalies in the field strength determined by the above methods.[3–6] During World War II both experimental and theoretical studies were made which revealed the extensive occurrence of nonstandard refraction, particularly over ocean areas, and provided improved mathematical methods for treatment of the simpler cases of nonlinear gradient of refractive index near the surface.[7–11] More recent studies of propagation, mainly on overland paths, have illustrated the complicated nature of atmospheric refraction in the presence of turbulence and the effects of terrain irregularities.[12–16] Frequent instances of abnormally strong fields in the diffraction region have been observed, and the mechanism of scattering from atmospheric irregularities, or "blobs," has been postulated to explain such fields.[17]

It is the purpose of this paper to report the results of a series of tests made at 412 mc with a transmitter having sufficient power output to permit measurements of field strength somewhat farther into the nonoptical region than has been possible customarily.

EXPERIMENTAL CONDITIONS

The tests to be described were conducted during the summer of 1948 and during the fall, winter, and spring of 1949–1950. A small amount of work was done at short range to determine the effect of terrain. The great bulk of the work was done at ranges in excess of 80 miles. Since low antennas were used for surface work at both ends of the path and since the terrain was relatively smooth, the receiving antenna was located several thousand feet below line-of-sight at each of the remote receiver sites. Reception of the ground wave was therefore impossible except during the rather rare occurrence of strong superrefraction. Thus, most of the results represent reception of tropospheric waves, presumably resulting from scattering by air masses in the lower troposphere having a refractive index differing slightly from the average.

The transmitter was located at the Cedar Rapids Airport, and the transmitting antenna was mounted at a height of about 40 feet on the roof of the Collins Radio Company hangar. Two types of transmitting antennas were used, one a biconical horn with omnidirectional radiation in the horizontal plane and a power gain of about 5 relative to an isotropic source, the other a pyramidal horn with an approximately square aperture and a power gain of 28. The former was arranged so as

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† Collins Radio Company, Research Div., Cedar Rapids, Iowa.

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[16] J. P. Day and L. G. Trolese, "Propagation of short radio waves over desert terrain," Proc. I.R.E., vol. 38, pp. 165–175; February, 1950.

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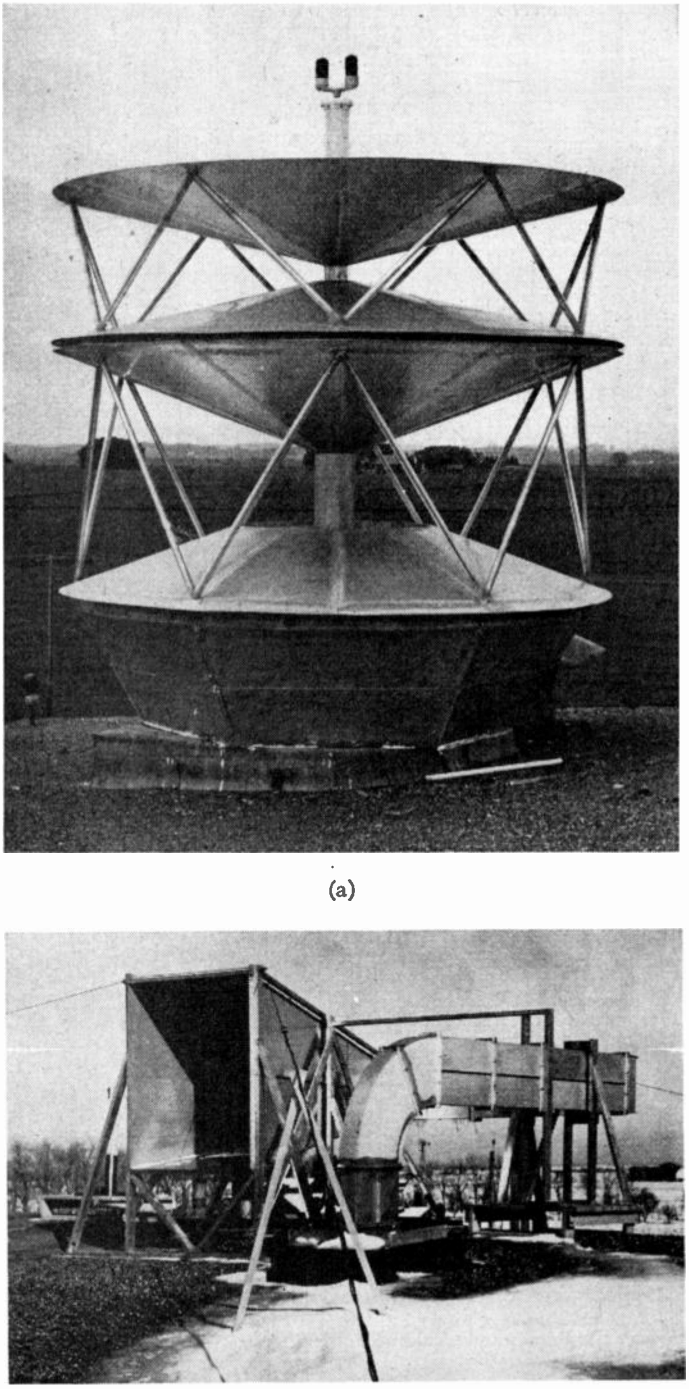
to produce either vertically or horizontally polarized radiation. The latter produced only horizontally polarized waves (see Fig. 1).



(a)

(b)

Fig. 1 (a) and (b)—View of two types of transmitting antennas. Radiation from upper biconical horn is vertically polarized, that from lower is horizontally polarized. Elevation is approximately 40 feet.

The transmitter consisted of a 50-kw resnatron oscillator coupled by means of special waveguide gear to the antenna. The transmitter output power was unmodulated and was held at a value of approximately 30 kw. A frequency in the range of 406 to 420 mc was used, with the bulk of the measurements being made at 412 mc.

The receiving antenna used for all of the fixed-point measurements was a 10-foot paraboloid, with a power gain of 140. A corner-reflector antenna with a power gain of about 10 was used for the surface mobile and exploratory tests (see Fig. 2). The receiving antenna was located about 10 feet above the surface. For aerial measurements, horizontal and vertical dipoles backed by a ground screen were mounted on the nose of a DC-3 airplane. The receiver was originally connected to an Esterline-Angus 1-ma recorder, operated at a speed of 12 inches per hour. In later tests, this record was supplemented by hourly photographs of a special counter panel indicating the time in minutes during which each of several levels was exceeded. The latter record greatly reduced the effort and time required for subsequent analysis.

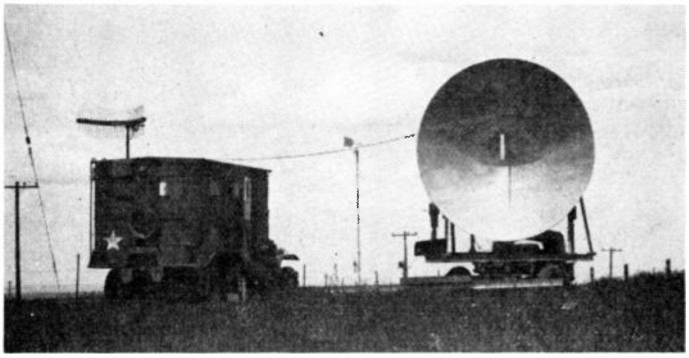


Fig. 2—View of receiving station showing 10-foot parabolic antenna and corner reflector antenna mounted on truck. Elevation of antennas is approximately 10 feet.

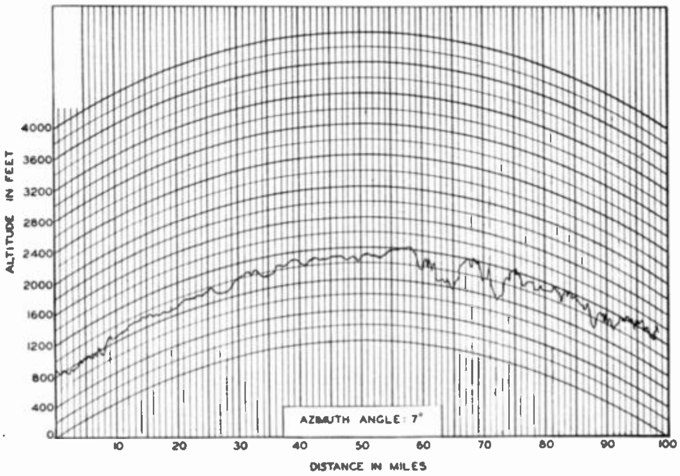


Fig. 3—Terrain profile of propagation path from Cedar Rapids to Waukon.

Extended measurements were made at three receiving sites, as follows: Waukon, Iowa, distance 98 miles, summer 1948 and fall 1949; Mitchellville, Iowa, distance 86 miles, winter 1949–1950; Quincy, Illinois, distance 134 miles, spring 1950. The terrain is gently rolling, with few wooded areas. A typical profile, that for the Waukon path, is shown in Fig. 3. The co-ordinate system is so chosen that a straight line represents a ray path in a standard atmosphere. The terrain on the other two paths was generally similar to that shown from 0 to 50 miles in Fig. 3. In addition to these fixed measurements, several mobile measurements were made with a truck and an airplane to provide more complete information regarding the effects of terrain and distance.

MOBILE TESTS

To determine the effects of terrain and distance upon the field strength, a truck was driven along an approximately radial road and a recording was made relating field strength with road distance. Portions of this record were analyzed to determine the statistical relation between field strength and location. The results are shown in Fig. 4. In each interval, an average distance was selected, as shown at the right of the curve, and field strength values for other distances in this interval were subjected to an appropriate correction for distance. Thus, these curves show only the effect of terrain, buildings, trees, and the like.

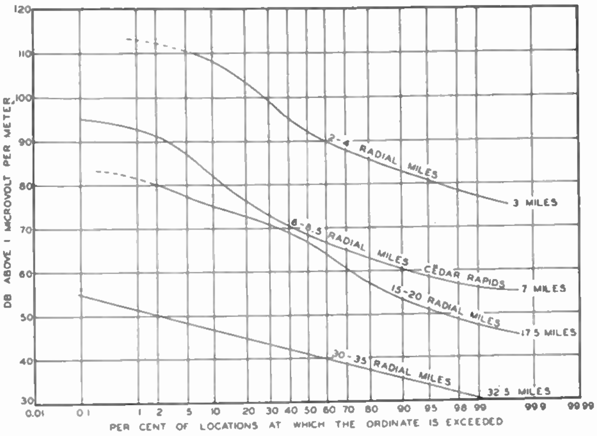


Fig. 4—Statistical relation between field strength and percentage of locations at which this field strength is exceeded. Measurements were made along an approximately radial road. Frequency 410 mc. Power 30 kw. Transmitting antenna power gain 5. Height 40 feet. Vertical polarization. Receiving antenna height 10 feet.

Though the data are too limited to warrant firm conclusions, several results are striking: Each of the upper three curves shows a range between the 1 and the 99 per cent values of about 36 db. It appears that the variation of field strength due to change of location in any one limited area is on the order of ±20 db within a distance of about 20 miles and with a receiving antenna about 10 feet above road level. A second noteworthy feature of Fig. 4 is the moderate depression of the curve obtained for the Cedar Rapids area below the position to be expected for a corresponding range in open country. This is only 5 to 10 db, indicating that the field strength in urban areas, even with low receiving antennas, is only moderately lower than in more open, rural areas. A third feature of Fig. 4 is that the terrain effect is only a little over 20 db for the curve corresponding to 32.5-miles distance. This could be only a coincidence if it were not corroborated by other tests at greater distances. It appears that the terrain effect is reduced with increasing distance, attaining a value of about ±10 db at a range where the tropospheric wave is strongly predominant.

The road trip was continued to a distance somewhat in excess of 200 miles. A plot of field strength versus distance is shown in Fig. 5 for that portion of the distance where reasonably reliable measurements could be made. In this, each vertical line represents the range of field strength measured in 3/16 mile of road distance, and the dot indicates the estimated median value. The dashed line represents an inverse-square variation of field strength with distance. This curve appears to fit the data reasonably well, considering terrain effects, out to 25 or 30 miles, except at 25 miles, where a deep river valley caused a depression of field strength. Between 30 and 50 miles the measured medians drop well below the inverse-square curve as we should expect when diffraction loss is considered. However, beyond 50 miles the measured field strength again appears to drop no faster than the inverse-square curve if we make allowances for various river valleys. Since diffraction of the ground wave is such as to produce quite a rapid decrease in field strength with increasing distance beyond the radio horizon, we must suppose that a tropospheric wave becomes effective beyond 40 or 50 miles so as to extend the effective range. Beyond about 125 miles the field strength was too low to permit effective measurement with the corner-reflector antenna used for the mobile test.

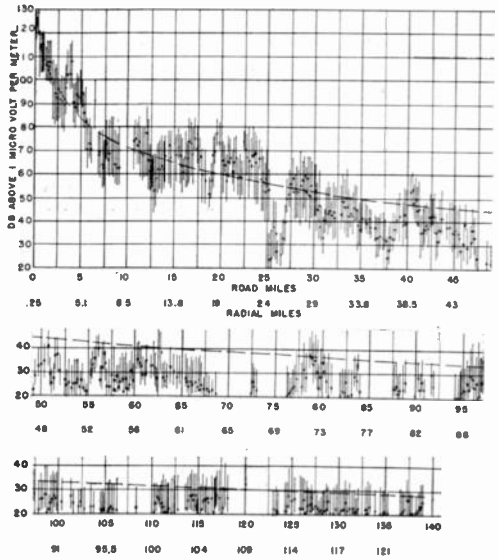


Fig. 5—Relation between field strength and distance. Measurements were made along an approximately radial road. Each line represents range of field strength observed in 3/16 mile of road distance and each dot represents estimated median. Dashed line is an inverse-square curve applying theoretically at short distances. Frequency 410 mc. Power 30 kw. Transmitting antenna power gain 5. Height 40 feet. Vertical polarization. Receiving antenna height 10 feet.

In Fig. 6 we see the results of measurements made with an airplane flown on a radial course at three different elevations. Very roughly, the elevation above terrain may be taken as 1,000 feet less than the elevation above sea level although terrain clearance varied as much as 500 feet during the flight. Each vertical line represents the range of field strength in a radial distance of 4 miles and the dot again indicates the median value. The dashed curves represent simple analytical approximations, in general, an inverse-distance relation at short distance and an exponential relation at longer distance. Within the range where the direct ray and ground-reflected ray, or rays, interfere to produce a lobe structure, the results are somewhat erratic, but show generally that the measured field strength approximates the free-space field strength. Beyond this range there is a diffraction region where the field strength drops rapidly with increasing distance. This extends 30 to 40 miles beyond the radio horizon. Near the outer limit of this region, the fluctuation range of the signal gradually increases, indicating the onset of contribution from the tropospheric wave. Beyond this region, the curves show a reduction in slope, which is characteristic of the region where the tropospheric wave is predominant. This third region was never reached at 15,000 feet within the limits of the flight. The constancy of the exponential coefficient in both the diffraction region and the tropospheric-wave region is a rather striking effect which facilitates moderate extrapolation. It will be observed, however, that the variation becomes somewhat more complicated at high altitudes.

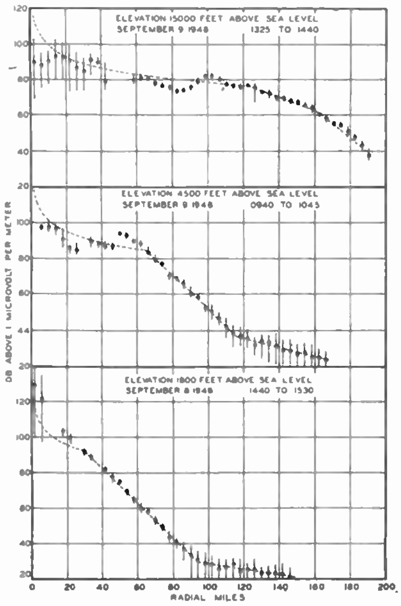


Fig. 6—Relation between field strength and distance as measured at three altitudes in an airplane. Each line represents range of field strength observed in 4 miles of radial distance and each dot represents estimated median. Dashed lines represent simple functions of distance chosen for best fit with measurements. For terrain elevation, see Fig. 3. Frequency 410 mc. Power 30 kw. Transmitting antenna power gain 5. Horizontal polarization. Azimuth angle 7 degrees.

The results just described are assembled for comparison with each other and with the theoretical results in Fig. 7. Here a logarithmic distance scale is used to permit extension to large distances. Also, the field strength has been converted to an equivalent value obtained with 1 kw radiated from a half-wave dipole by subtracting 10 log $(5/1.64\times 30)=19.6$ db from all measured values. Curves A and C were computed from the theory applying in the case of a smooth spherical earth, with the radius increased by a factor of 4/3 to take into account average atmospheric refraction. Curves B, D, E, and F were drawn as smooth curves representing the estimated best fit with experimental data. Extension of the curves beyond a distance of about 150 miles was accomplished partly by extrapolating the curves in Figs. 5 and 6 and partly by utilizing the results of limited surface measurements made at a distance of 225 miles. The optical range for a spherical earth and standard refraction as calculated by the formula

$$d_0 = \sqrt{2h_t} + \sqrt{2h_r} \tag{1}$$

is shown on each of the curves. The median experimental values taken from Fig. 4 are shown by circles. The scatter of these points about curve B illustrates the prominence of the terrain factor even after purely local fluctuations have been removed. No effort was made to show lobe structure in Curves D, E, and F because no regular lobe structure could be observed. We should note finally that all data for Fig. 7 were obtained during

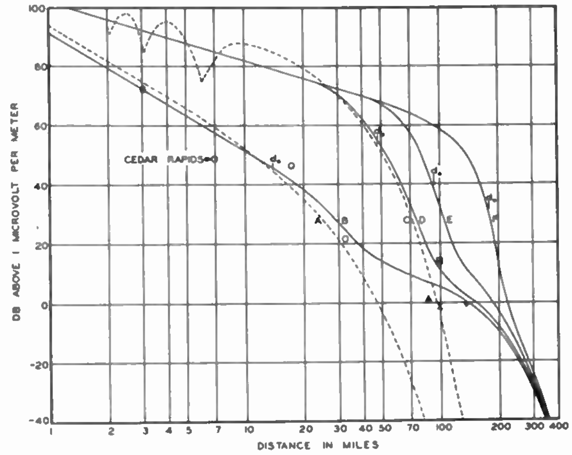


Fig. 7—Variation of median field strength with distance for various receiving antenna heights. Frequency 410 mc. Effective radiated power 1 kw. Transmitting antenna height 40 feet.

A Theoretical curve for receiving antenna height 10 feet.
B Experimental curve for receiving antenna height 10 feet.
C Theoretical curve for receiving antenna height 1,000 feet.
D Experimental curve for receiving antenna height 1,800 feet msl.
E Experimental curve for receiving antenna height 4,500 feet msl.
F Experimental curve for receiving antenna height 15,000 feet msl.
d_0 Radio optical distance.
o Median values taken from Fig. 4.
x Median value from Curve A, Fig. 12 (Waukon, Fall, 98 miles).
△ Median value from Curve B, Fig. 12 (Mitchellville, Winter, 86 miles).
+ Median value from Curve C, Fig. 12 (Quincy, Spring, 134 miles).
⊡ Median value from Curve D, Fig. 12 (Waukon, Summer, 98 miles).

those daylight hours which were found to yield minimum anomalies due to superrefraction.

It is observed that the experimental results, except for terrain effects, agree rather well with calculations made for a standard refractive index gradient in the atmosphere out to a distance beyond optical distance varying from about 20 to about 40 miles. The most striking feature of the experimental curves is the point of inflection 20 to 30 miles beyond optical distance. Here the tropospheric wave becomes rapidly more prominent with increasing distance, the attenuation becomes less, and fading begins. The decreased attenuation of the tropospheric wave leads to an amazing increase in range, particularly for low antennas. This increase in range may be as much as several hundred miles for low antennas and a high-power transmitter. The increase in range becomes relatively small when one antenna is sufficiently elevated, probably because waves scattered at low levels, where turbulence is most intense, do not greatly increase the optical distance, and because scattering at high levels is relatively poor. With both antennas at an elevation of several thousand feet, the effect of scattering is probably negligible. A secondary effect, shown by Fig. 7, is the convergence of the curves for various elevations at large distances. This indicates a pronounced decrease in the height gain of the antenna over certain ranges of height. Since height gain is intimately associated with diffraction and since the tropospheric wave can reach the receiving antenna with less diffraction around the curve of the earth than the ground wave, this effect is to be expected. The larger grazing angle of the tropospheric wave at the receiver also tends to account for the reduced terrain effect at large distances mentioned earlier.

The median values obtained at various fixed sites and at various seasons are also shown in Fig. 7 by the symbols x, △, +, and ⊡. The points representing summer results at Waukon lie above Curve B, probably because the site, which was on a carefully selected high point, was better than a median site. The greater prevalence of nocturnal superrefraction in the summer must also be considered. The point representing fall results at Waukon is based on measurements during one week only when rather persistent stormy weather caused abnormal depression of the field strength. The point representing winter results at Mitchellville lies below Curve B, probably because the receiving site in this case was poorer than the median. The receiving antenna was partly shadowed by a gentle rise in the ground facing the transmitter. The point representing results at Quincy is seen to lie very close to Curve B. Since these points are based on about 2,300 hours of data, their rather close grouping about Curve B serves to lend considerable validity to this curve in the region near 100 miles. In fact, when the results secured at the three sites are suitably corrected for distance and are combined in a manner which will be explained later, a median value of 4.5 db above 1 μv per meter is obtained at 100 miles. This is only 0.5 db less than the value shown by Curve B.

MEASUREMENTS AT A FIXED POINT

An extensive series of field-strength recordings was made at three sites as follows:

Location	*Distance*	*Period*	*No. of Hours*
Waukon, Iowa	98 miles	June, July, August, 1948	500 approx.
Waukon, Iowa	98 miles	October, November 1949	156
Mitchellville, Iowa (WHO)	86 miles	November, 1949 to March, 1950	1,162
Quincy, Illinois (WTAD-FM)	134 miles	April, May, 1950	517

In each case, measurements were made with a receiving antenna only about 10 feet above the surface. In general, the records cover operation for 24 hours a day for intervals of about 6-days duration spaced fairly uniformly throughout the periods mentioned. Thus, reasonably representative samples are available for all seasons of the year.

During the tests made at Waukon and during the mobile tests, a pair of biconical transmitting antennas was used which were capable of radiating either vertically polarized or horizontally polarized waves. It was found that the strength of received signals at large distances varied so little when the direction of polarization was changed that this variation could not be detected in the presence of the strong fluctuations characterizing the signal at nearly all times. During the tests made at Mitchellville and Quincy, a pyramidal horn was used at the transmitter to improve the signal level at the receiver, to provide greater purity of polarization when horizontal polarization was used, and to assure a more reliable figure for the power gain. In all tests made with the pyramidal horn, horizontal polarization was used.

The field strength measured at the fixed sites was usually characterized by a rapid scintillation of large amplitude. The rapidity of the scintillation was frequently so great that the record produced by an Esterline-Angus recording milliammeter, operated with a chart speed of 12 inches per hour, showed a solid band of ink with irregular upper and lower borders. During summer nights, superrefraction was quite common. This had the effect of increasing the field strength quite markedly and reducing the rapidity of fading. Occasionally, the superrefraction became so pronounced that the fading almost disappeared and the field strength resembled that within a few miles of the transmitter. On a few occasions, peak levels of about 80 db above 1 μv per meter were recorded, roughly 40 db above the daytime median value. Such anomalies were associated, as one might expect, with relatively clear skies and strong radiation cooling of the surface. However, no simple correlation between field strength and meteorological conditions could be determined, except in the case of the sunrise maximum. This effect was a characteristic duct of unusual strength appearing

1 to 3 hours after sunrise. The explanation for this appears to be the creation of high-surface humidity by the evaporation of dew. In all cases, the duct disappeared with amazing rapidity when the evaporation was essentially complete and turbulent mixing of the air became well-established. Duct conditions during the winter were rare or nonexistent. Storms along or near the propagation path generally caused pronounced depression of the field strength, probably because of better mixing of the air at all heights.

No conclusive results are available to indicate the angle of arrival of the wave at the receiver. The rapid fluctuation of field strength and the relatively broad antenna beam (approximately 15 degrees) made it difficult to determine the angle of arrival with any accuracy from a test in which the antenna was slowly rotated in a vertical plane. The wave appeared to arrive so nearly in a horizontal direction that all tests were made with the antenna directed horizontally. However, it was quite evident that the antenna beam was appreciably broader horizontally in the presence of a scattered field, indicating scattering from a considerable volume of space. Also, the reduced shadowing effect of terrain obstacles at large distances, as well as the reduced height gain at the receiver, indicated an effective source at least several degrees above the horizon. It may be argued that the effective power gain of the receiving antenna may be much less with a scattered signal than with a single plane wave. Hence, the true field strength at the receiver may be appreciably greater than that measured with an antenna having a beam width as narrow as that used in these tests. This argument is justified to a certain extent. However, successive measurements made with a corner reflector (power gain = 10) and a 10-foot paraboloid (power gain = 140) indicated no significant differences either in the measured field strength or in the character of the fading. We must conclude that the scattered waves in this case had a sufficiently small deviation from the general direction of arrival so that the 10-foot paraboloid gave results differing by only a very few db from the true field strength. The best estimate is that scattered waves arrived principally from directions varying from 0 to about 10 degrees vertically and from −10 to +10 degrees horizontally. This narrow beam of the scattered field appears to indicate atmospheric irregularities of dimensions generally large compared with the wavelength.

The statistical distribution of field strength for each hour of recording was determined in a manner described elsewhere.[18] The observed values taken from the recorder chart or the time totalizer were plotted on Rayleigh graph paper, and a smooth curve was drawn through these points. In the majority of cases, the hourly distribution curve drawn on Rayleigh paper was nearly linear, especially at the low field-strength end. This fact was utilized to justify extrapolation of the curve to values of field strength lower than those which could be reliably measured. It proved of great value for periods of unusually weak fields when the received signal dropped below the receiver noise level during an appreciable fraction of the time. Two sets of readings were taken from these smoothed curves: The first set consisted of the values of field strength exceeded during an arbitrary percentage of the time, such as 1 per cent, 10 per cent, 50 per cent, and so on. These values could then be plotted against time to show the variation of the hourly field strength for the period of measurement. Space does not permit inclusion of these curves here. The second set consisted of the percentage of time during each hour that each of a number of arbitrary levels was exceeded. This tabulation was subsequently utilized in determining the percentage of time during which each of these levels was exceeded for a particular hour of the day for an entire period of measurement.

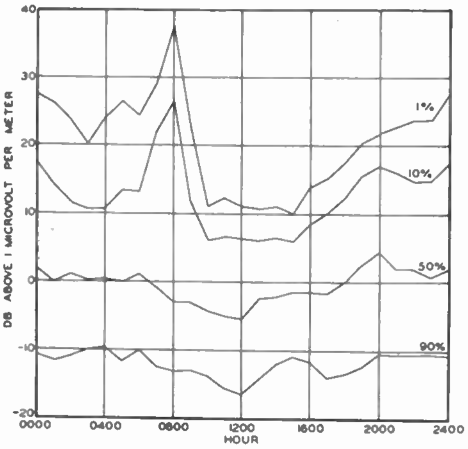


Fig. 8—Diurnal variation of field strength at Waukon, Iowa, averaged over 156 hours during October and November, 1949. Number opposite curve represents percentage of time ordinate value is exceeded. See Fig. 12 for other conditions. Effective radiated power 1 kw.

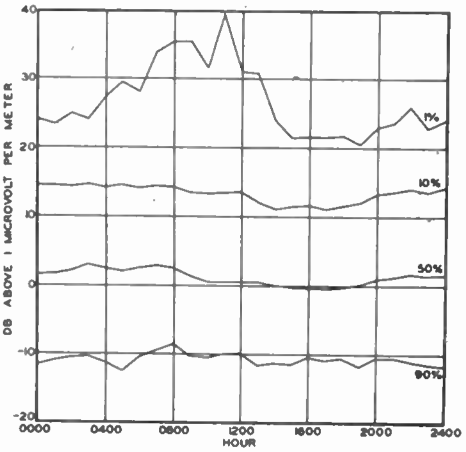


Fig. 9—Diurnal variation of field strength at Mitchellville, Iowa, averaged over 1,162 hours during November–March, 1949–1950. Number opposite curve represents percentage of time for which ordinate value is exceeded. See Fig. 12 for other conditions. Effective radiated power 1 kw.

These results, showing the diurnal variation of field

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[18] R. P. Decker, "Notes on the analysis of radio propagation data," Proc. I.R.E., pp. 1382–1388; this issue.

strength at four locations at various times of the year, are plotted in Figs. 8 to 11. Fig. 8 is based upon a very limited amount of data obtained in the fall, during a period when frequent storms caused abnormal depression of the field strength. However, sufficient fair, mild weather occurred to produce characteristic nocturnal superrefraction during a significant percentage of the time, as indicated by the two upper curves. The sunrise maximum occurring between 0,700 and 0,900 is quite evident. It should be noted that the hour indicated on the abscissa scale represents the beginning of the hour for which the ordinate value applies. A somewhat less pronounced rise of field strength occurred during the first half of the night while radiation cooling was most prominent. Fig. 9 indicates an almost complete absence of diurnal variation during the winter months. The forenoon rise in the 1-per cent curve may indicate an effect similar to that observed shortly after sunrise during the summer. However, the absence of this rise in the 10-per cent curve indicates that superrefraction is quite rare in the winter. Fig. 10 shows the reappearance of the sunrise maximum in the spring, with only very moderate improvement in field strength during the other nighttime hours. Fig. 11 shows the strong diurnal variations occurring during summer months. The sunrise maximum is very pronounced. Another somewhat less prominent maximum occurs during the evening hours while radiation cooling is most rapid. Because of the extreme anomalies occurring during summer nights, and the rather limited period of measurement, the results shown in Fig. 11 are necessarily subject to greater uncertainty than those shown for the winter months in Fig. 9.

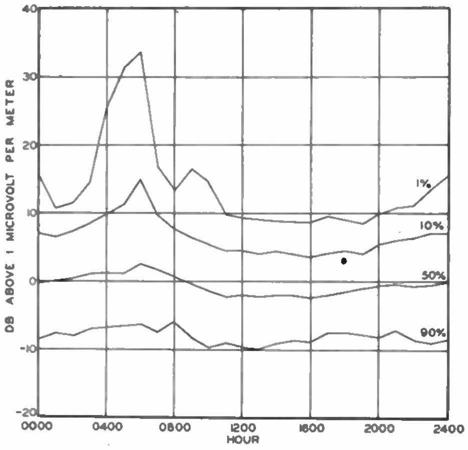

Fig. 10—Diurnal variation of field strength at Quincy, Illinois, averaged over 517 hours during April and May, 1950. Number opposite curve represents percentage of time for which ordinate value is exceeded. See Fig. 12 for other conditions. Effective radiated power 1 kw.

When the results for an entire period of measurement are summarized, we obtain the results shown in Fig. 12. We see that Curves A and B are nearly coincident in spite of different locations and seasons. The favorable location and the moderate superrefraction at site A were largely offset by an unusual incidence of stormy weather so that the over-all results resemble closely those secured during the winter months at site B. Curve C, representing spring results at an increased distance, has a shape which is not readily explained. Except for the portion of the curve below an abscissa of 2 per cent, which can be explained on the basis of nocturnal superrefraction, this curve shows an abnormally small fading range. Curve D indicates a generally high field strength, partly because of a favorable receiving location and partly because of rather persistent favorable propaga-

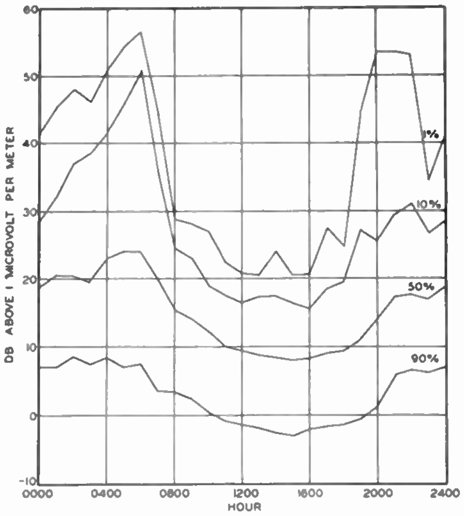

Fig. 11—Diurnal variation of field strength at Waukon, Iowa, averaged over approximately 500 hours during June, July, and August, 1948. Number opposite curve represents percentage of time for which ordinate value is exceeded. See Fig. 12 for other conditions. Effective radiated power 1 kw.

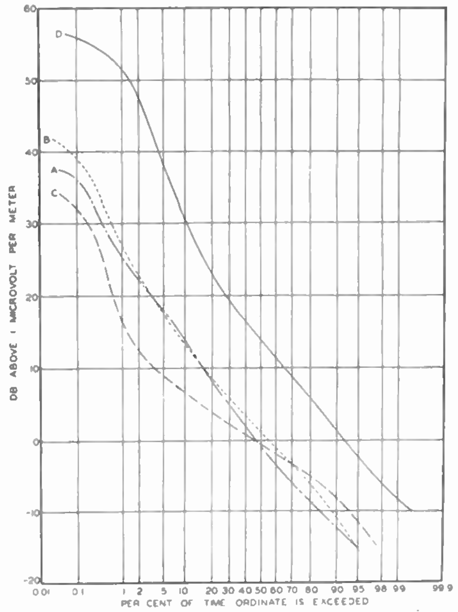

Fig. 12—Statistical variation of field strength at large distances. Frequency 410–412 mc. Effective radiated power 1 kw. Transmitting antenna height 40 feet. Vertical polarization, Curve D, horizontal polarization, Curves A, B, C. Receiving antenna height 10 feet.

A Waukon, Iowa, 98 miles, 156 hours, October–November, 1949.
B Mitchellville, Iowa, 86 miles, 1162 hours, November–March, 1949–1950.
C Quincy, Illinois, 134 miles, 517 hours, April–May, 1950.
D Waukon, Iowa, 98 miles, approximately 500 hours, June–August, 1948.

tion conditions occurring during the summer. It is interesting to note that the field strength range included between the 1-and 90-per cent levels varies from 38 db in the winter to 50 db in the summer. We should note also the very considerable interference of a 400-mc signal during a summer evening. Fig. 11 shows a level of about 54 db above 1 μv per meter exceeded 1 per cent of the time during the hours 2,000–2,300, whereas Curve D (Fig. 12) shows a level of 51.5 db exceeded 1 per cent of the time for the whole day.

In Fig. 13, an effort has been made to combine the four curves of Fig. 12 to show the field strength to be expected at a distance of 100 miles. The data for the various locations were weighted generally in accordance with the number of hours of observation, the number of months covered, and the estimated reliability. Arbitrary weighting factors were assigned as follows:

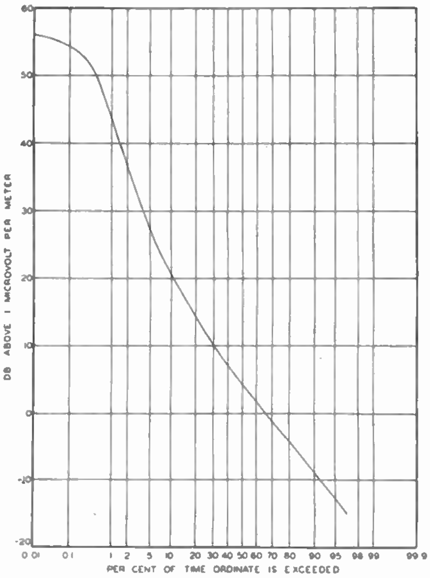

Fig. 13—Statistical variation of field strength at 100 miles. The results of Fig. 12 have been combined as explained in the text. See Fig. 12 for conditions. Effective radiated power 1 kw. Transmitting antenna height 40 feet. Receiving antenna height 10 feet.

Data for Site	*Season*	*Weighting Factor*	*Distance*	*Add db*
Waukon	Fall	1	98 miles	0
Mitchellville	Winter	5	86 miles	−2
Quincy	Spring	2	134 miles	4.5
Waukon	Summer	4	98 miles	0

Distance corrections were taken from Curve B of Fig. 7. The weighting factor may be considered also as the number of months to which the data are assumed to apply. In this curve, representing all-year results, we observe a range between the 1-and 90-per cent levels of 52 db, a median value of 4.5 db (agreeing well with Fig. 7, Curve B), and a nuisance signal of 43 db exceeded 1 per cent of the time. This figure also embodies the effects of favorable and less favorable terrain in that the highest values of field strength are based entirely on the favorable site at Waukon, whereas the lowest values of field strength are considerably influenced by the results at the less favorable sites at Mitchellville and Quincy, with combined weighting factors of 7.

The diurnal variation taken from these composite results is shown in Fig. 14. We observe again the morning and evening maxima, the former being more persistent as shown in the 10-per cent curve. The median curve shows a diurnal variation of only about 6 db, whereas the 90-per cent curve shows a negligible diurnal variation.

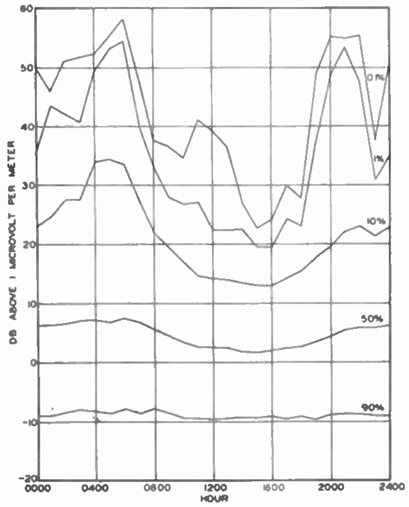

Fig. 14—Diurnal variation of field strength at 100 miles averaged over approximately 2,340 hours during all seasons. See Fig. 12 for conditions. Effective radiated power 1 kw. Low ground-based antennas.

On numerous occasions, when airplanes were observed to be flying over the propagation path, the observed field strength fluctuated rhythmically in amplitude, sometimes over a range of more than 10 db, with a period varying from a short value of less than a second to a long value of several seconds and then again to a short value. The duration of such a disturbance varied from a few seconds to more than a minute. At times, this effect caused a buildup as high as 20 db in the field strength. Though the net effect on observed field strength was slight on the propagation paths chosen in these tests, it is believed that a significant increase in the interference value of the signal could occur on a path where air traffic was heavy and the routes were generally parallel to the path.

CONCLUSIONS

We find that at 400 mc the field strength at a median location may be computed with reasonable accuracy in the conventional manner, that is, by assuming a spherical earth and standard atmospheric refraction, provided that the distance does not exceed the radio-optical distance by more than about 20 miles. Even within this range, nocturnal radiation cooling and morning surface evaporation may increase the field strength quite markedly for a small percentage of the time during the warm season. At greater distances, the diurnal variation due to nocturnal superrefraction is even more prominent. However, the field strength also develops strong, rapid fluctuations and decreases more slowly with increasing distance as a result of wave scattering in the lower troposphere. This scattering is largely independent of the time of day or the season. It is greatly reduced only by certain frontal conditions which induce more thorough mixing of the air. Pronounced superrefraction occurs only during summer nights, but this effect is sufficiently persistent to cause a significant field strength increase during as much as 5 per cent of the time for the entire year.

The degree of the anomaly produced by atmospheric turbulence and superrefraction for very low antennas at 100-miles distance can be appreciated by comparing a field strength of about -60 db above 1 μv per meter computed for a spherical earth and standard refraction with a value of about 5 db representing the median measured value. This discrepancy is reduced when either antenna is raised. With one antenna at 1,000 feet, the corresponding values are about -5 db and 11 db.

Since a high transmitting antenna may produce a greatly extended service range but little more interfering field strength at great distances, it is evident that a transmitting antenna height of about 1,000 feet is desirable in any proposed high-quality uhf broadcasting network. For communication grade of service, it appears entirely feasible to operate a 100-mile link with low, directional antennas and about 10 kw of transmitter power with a probability of satisfactory field strength more than 90 per cent of the time. In fact, quite effective use of the scattered wave can be made to a distance of 200 miles, as indicated by Fig. 7. It must be recognized, however, that the results reported here are strictly applicable only to Iowa and adjacent regions, and that differing terrain and meteorological conditions in other regions may modify the results materially.

Because of multipath propagation occurring on such long paths, a certain degree of distortion must be expected in the reception of modulated waves. Since the relative delay on tropospheric paths is probably small compared with a modulation-frequency period, such distortion should be much less severe than on ionospheric paths.[19] Very brief tests with speech modulation tend to confirm this conclusion. The signal was almost always more readable than that received over a low-power link operating in the 5,000-kc range.

ACKNOWLEDGMENT

The work reported here was supported in part by the Central Radio Propagation Laboratory under Contract CST-10783.

The assistance of members of the Research Division of the Collins Radio Company, who conducted the tests and analyzed the data, and the generous co-operation of the personnel of Radio Station WHO and WTAD-FM are gratefully acknowledged.

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[19] Irvin H. Gerks, "An analysis of distortion resulting from two-path propagation," PROC. I.R.E., vol. 37, pp. 1272–1277; November, 1949.

Notes on the Analysis of Radio-Propagation Data*

R. P. DECKER†

Summary—**This paper deals with the reduction of radio-propagation data. The primary aim is to present a clear picture of signal-strength variation with a minimum amount of computational work. A newly developed recording device is described and illustrated, together with an effective method of data analysis.**

I. INTRODUCTION

AN INCREASING number of ultra-high-frequency and very-high-frequency radio-propagation studies have been carried out lately by investigators in different countries. This work is certain to continue, and a proportionate amount of data will have to be recorded and analyzed. Many hours of tedious work have, so far, been required to present a statistical picture of the variation of field strength over a given period of time. Thus, it may be of some value to describe the newer recording techniques together with an effective method of data reduction and presentation. This paper is based chiefly on the author's experience with high-power uhf propagation at 410 mc.[1] Modifications would have to be made in a similar treatment for lower frequencies.

The problem is then to analyze the data with a minimum amount of work and to summarize the results by means of curves and charts. The amplitude variation of the signal is translated into data which will have the most significance for statistical analysis. The representation will be most effective when the percentage of time is graphed during which various signal strengths are exceeded. The basic interval is taken to be the hour, since it is short enough to preclude, in general, a change in propagation conditions, and long enough to smooth out the random variations which are of little or no interest.

A recording instrument (usually of the Esterline-Angus type), is used to record continuously the output

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[1] Measurements were made under the auspices of the Central Radio Propagation Laboratory during the winter and spring of 1949–1950. The transmitter, a 400-mc resnatron with an average power output of 24 kw, was located at the airport in Cedar Rapids, Iowa. The receiving sites were at station WHO near Mitchellville, Iowa (86.1 mi.) and at station WTAD-FM, Quincy, Illinois (133.9 mi.).

or avc voltage of a field-intensity meter. Additional equipment consists of a signal level versus time indicator or time totalizer, a counter panel with camera, and a signal generator with a suitable attenuator for calibration purposes. The signal level versus time indicator and the counter panel, which have been developed recently, will be described in some detail later.

The calibration procedure, which is of primary importance, will be considered first.

II. CALIBRATION PROCEDURE

The calibration of the receiver relates output meter reading to signal strength. The relation between field strength and receiver input voltage is given by the following equation:

$$E^2 = \frac{1.64(480\pi^2)e^{2\alpha l}}{P_t G_t \lambda^2 G_r Z_0} E_r^2 \left(\frac{\text{volts}}{\text{meter}}\right)^2, \qquad (1)$$

where

E = field strength in volts/meter,

E_r = received input voltage,

α = attenuation in nepers per axial foot of the cable connecting the receiving exciter to the receiver input,

l = length of the cable in feet,

G_r = power gain of the receiving aperture referred to an isotropic source,

λ = wavelength in meters,

G_t = power gain of the transmitting aperture referred to an isotropic source,

P_t = power flow in kw through the transmitting aperture, and

Z_0 = surge impedance in ohms of the cable.

In decibel notation (1) may be written as follows:

$$\begin{aligned} 20 \log E = & - 10 \log P_t - 10 \log G_t + 38.91 - 20 \log \lambda \\ & - 10 \log G_r - 10 \log Z_0 \\ & + 8.69\alpha l + 20 \log E_r \text{ (decibels)}. \end{aligned} \qquad (2)$$

Once the constants have been evaluated, (2) becomes

$$E' = 20 \log E_r - 20 \log \lambda - 10 \log P_t + \text{constant}. \qquad (3)$$

E' is then expressed in db above 1μv per meter referred to a kw radiated from a half-wave dipole. The voltage required of the signal generator to produce a given output meter deflection can immediately be translated into field strength by (3).

As a rule, the frequency is stable and can be included as a constant in (3). The transmitted power, however, will possibly vary, especially when the resnatron is used in high-power uhf work.

With the aid of (3) a calibration curve of output meter reading versus db above 1 μv per meter is plotted. When a change in power level occurs, it is only necessary to add the appropriate amount of db for each meter reading.

The question which frequently arises is how often receiver calibrations should be made. It is evident that the accuracy of the analysis increases with the number of receiver calibrations. However, too much time spent in calibration could possibly make the data for the particular hour unreliable if abrupt changes in propagation occur, such as presented by an early morning duct, for example. It is therefore advisable to calibrate as infrequently as possible when such conditions prevail, since the shape of the distribution curve (see Section V) could be seriously affected, especially in the low- or high-percentage region. The minimum number of calibrations permissible must be determined by the stability of the receiver. The exact length of time during which the receiving and recording equipment were inoperative must be known in order that corrections for the percentage of time can be made.

III. ESTERLINE-ANGUS ANALYSIS

The percentage of time during which various preselected levels were exceeded may be obtained by an analysis of the Esterline-Angus chart. The chart speed should be fast enough to observe normal fading; 16 divisions or 12 inches per hour is usually adequate. The analysis has been accomplished by estimating the percentage for each division. The average of the 16 divisions yields the average for the hour.

Accurate plotting of the distribution curves becomes easier when the levels are chosen so that there is a constant db difference between them. The range of the receiver in db divided by 9 results in the db increment for 10 levels. Recourse to the calibration curve will give the 10 corresponding output meter readings to be used in the analysis.

Under normal propagation conditions, an hour of the Esterline-Angus chart can be completely analyzed in 50 minutes by a person with some experience. However, some practice is necessary before reasonable accuracy (within 2 per cent) can be obtained. Levels which were exceeded from 0.01 to 2 per cent and from 98 to 99 per cent of the time under normal conditions are not easily or accurately detected by this method. At these levels, the width of a pen line for a very short peak or dip in signal is not an accurate indication of the percentage.

IV. THE SIGNAL LEVEL VERSUS TIME INDICATOR (SLVTI)

In the preceding section it was noted that the errors involved in a visual analysis of the Esterline-Angus charts were about 2 per cent. When the results of this analysis are plotted on log log or on probability paper, an extreme scattering of points could occur, especially in the low- or high-percentage regions. It also requires an excessive amount of time for analysis, and results in eye fatigue. To overcome these disadvantages, a device called the "Signal Level Versus Time Indicator"[2] (Fig. 1) has been developed which indicates the time or per-

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[2] For a description of a similar unit, see R. W. George, "Signal-strength analyser," *Electronics;* January, 1951.

centages of time during which a number of preselected levels were exceeded. The indicator shown in Fig. 1 works in conjunction with a counter panel (Fig. 3), upon which synchronous timing motors and revolution counters are mounted.

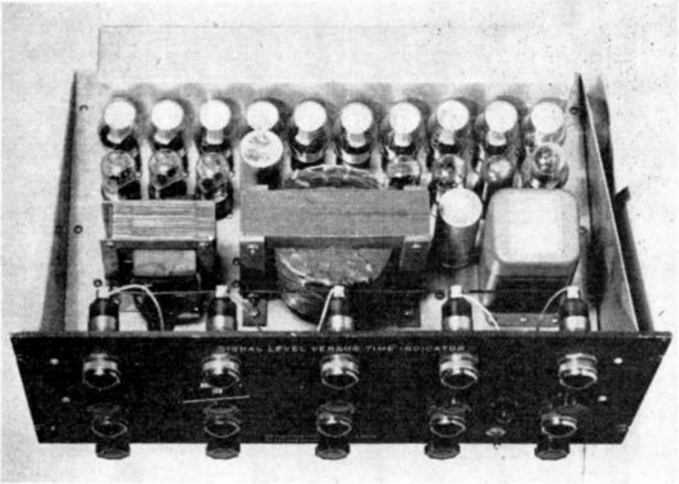

Fig. 1—Signal level versus time indicator.

The circuit diagram for such a 10-channel device is shown in Fig. 2. The 6SN7 stages are so biased by the selenium rectifier and voltage-divider circuit that an increasingly negative avc voltage applied to the grid causes the voltage across the 100,000 ohm plate resistor to go positive with respect to the thyratron cathode. As soon as the critical voltage is reached, which is determined by the 400-ohm voltage dividers, the thyratron will fire and close its plate-circuit relay. This, in turn, energizes a timing motor which drives a revolution counter. In order to make the thyratrons self-extinguishing, ac voltage is used on the plate. A panel light is shunted across each motor for level adjustments.

Under conditions of rapid fading, a relay is energized as often as several times a second. Thus the synchronous timing motor which drives the revolution counter must start and stop instantly to prevent cumulative errors. If the speed of the counter is 10 rpm, the counter will indicate the time in tenths of minutes for which its level was exceeded. However, if the counter speed is $166\frac{2}{3}$ rpm, the counter will indicate hundredths of a per cent. The latter speed eliminates the process of division in the calculation of percentages. This represents a saving of about 2 man hours per hundred hours of data. It also means increased accuracy in plotting the distribution curves, since a fraction of a per cent error in the low- or high-percentage region of Rayleigh paper

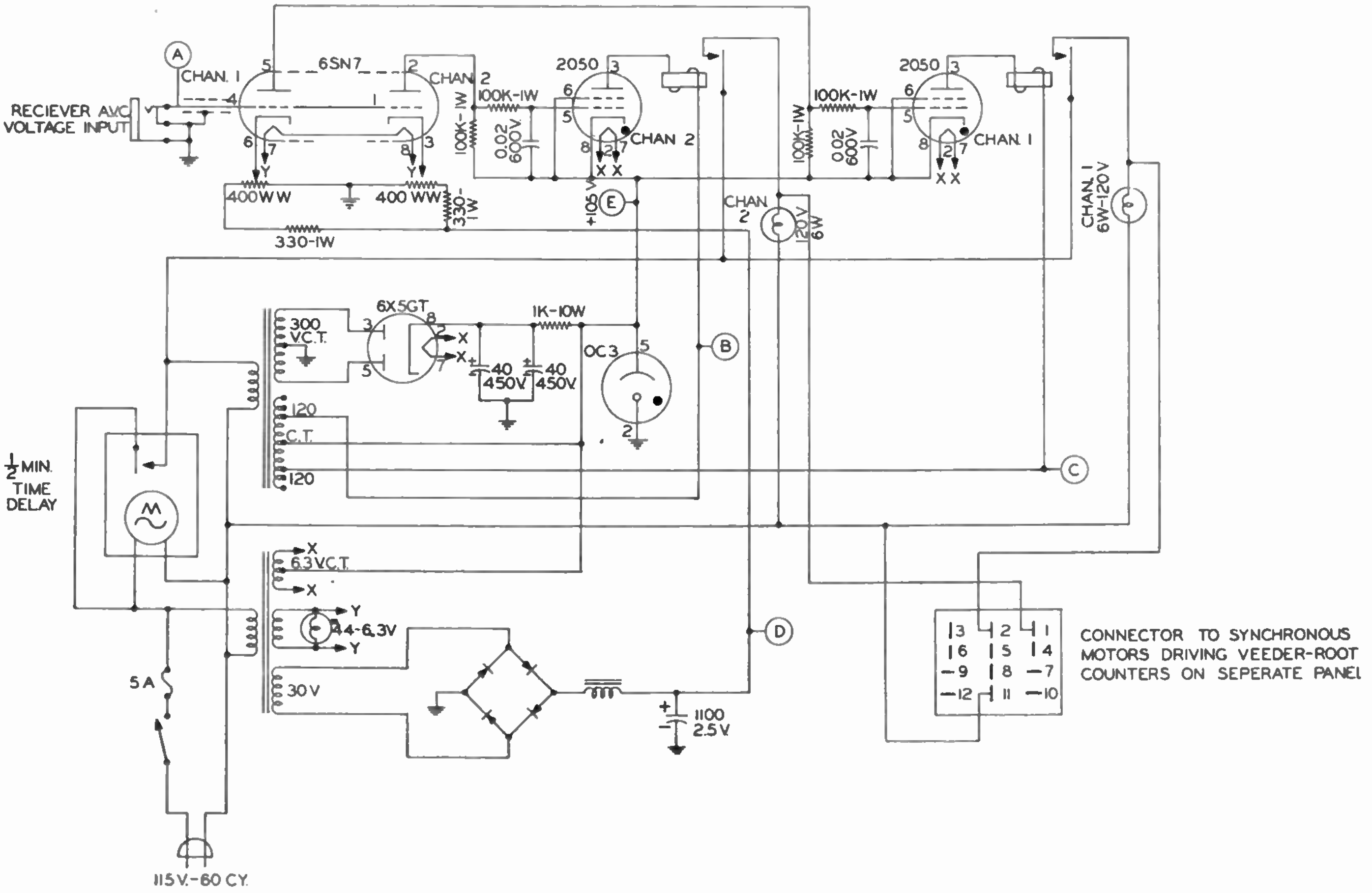

Fig. 2—Circuit diagram of the signal level versus time indicator. All resistances in ohms; all capacitances in μf. Note: Only 2 of 10 channels are shown. A=Input to all channels. B=Only to channels 2, 4, 6, 8, and 10. C=Only to channels 1, 3, 5, 7, and 9. E and B=All channels.

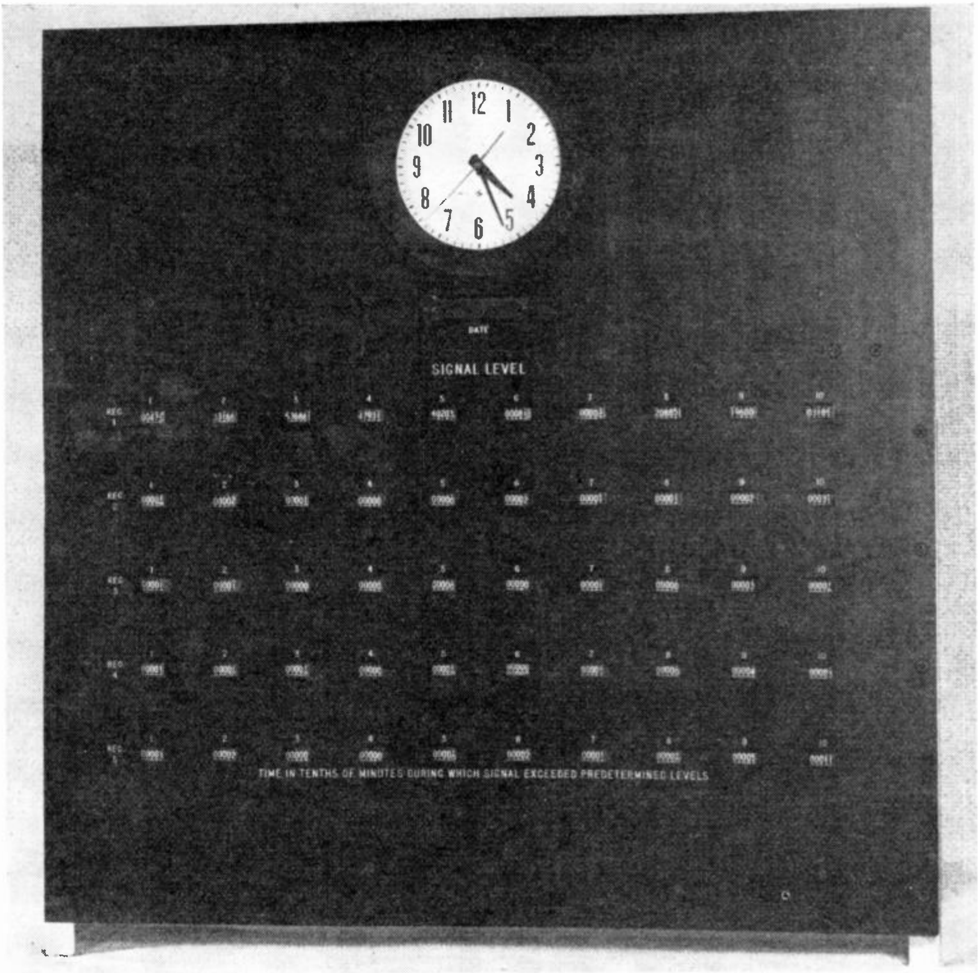

Fig. 3—Counter panel.

could distort the shape of the curve (as is possible with a slowly driven counter). As the data will eventually be plotted on Rayleigh paper, in which the abscissa distance from the origin is proportional to $\log 4 - \log [2 - \log P]$ (where P is percentage), it is easily seen that percentage increments near 0 or 100 represent relatively large distances.

The model shown in Fig. 2 must be externally regulated on account of the bias and thyratron plate voltages. Good regulation is required so that any voltage changes occurring within the SLVTI are negligible with respect to the 0–6v input for which this unit is designed.

The threshold of potential of the 2050 thyratrons has been found to be about 0.02 volt wide. This, as well as imperfect regulation, will cause the thyratron to fire and extinguish within ± 0.05 volt of the intended level. The circuit would, therefore, be improved by a well-regulated bias voltage and an additional stage of amplification ahead of the 6SN7's for greater firing accuracy with respect to the input voltage.

The circuit has been field tested and has proved to be entirely satisfactory. One indication of its performance is the exactness with which the resulting points lie on a smooth curve. A recent analysis of about 1,500 hours of data has shown that the points fit exceptionally well in each instance.

The ten counters with white on black numerals, which are mounted on a separate black panel (Fig. 3), are photographed automatically every hour on the hour. When the field run is over, the film (35 mm.) is inserted in a suitable viewer and the counter reading is recorded for each level and hour. Subtracting the counter readings of two successive hours yields the time or percentage of time (depending on the counter speed) during which the particular level has been exceeded.

As in the case of an Esterline-Angus analysis, the levels are best chosen in such a manner that there is an equal db increment between them. In calibrating, it is only necessary to adjust the signal generator until the desired level just clicks in, and to record the number of microvolts required.

The advantages of this method lie in the saving of tedious work in analysis, and in the increased accuracy. With the crude method of the Esterline-Angus chart analysis a full hour was required to analyze and draw the curve for an hour's data; but with the SLVTI the analysis may be done much more accurately in 6 minutes—a speed-up of ten to one.

V. CURVE PLOTTING

The percentage of time during which a number of preselected levels was exceeded according to the SLVTI and the value of these levels in db above 1 μv/m/kw, as determined by the calibration, have now been established. For purposes of condensation and interpolation, the points are plotted on suitable paper. It is well known that log log or Rayleigh paper is customary for this purpose. Norton,[3] starting with Lord Rayleigh's original work,[4] has shown that the percentage of time during which a field strength will theoretically be exceeded is given by

$$P = 100e^{-E^2/(E_t^2+E_g^2)} \qquad (E_g \ll E_t), \qquad (4)$$

where

P = percentage of time,

E = field intensity of resultant of sky wave plus ground wave, which is exceeded P per cent of the time,

E_t = rms of tropospheric wave, and

E_g = rms of ground wave.

The Rayleigh distribution holds normally at large distances in which the ground wave E_g is small compared to the tropospheric wave E_t. In addition, duct or layer-type propagation conditions would not be "Rayleigh-distributed" with time.

If k log log $(100/P)$ is represented as the abscissa distance and k' log $E/\sqrt{E_t^2+E_g^2}$ as the ordinate distance, it is easily seen that (4) will plot as a straight line. Since the field strength is in db above 1 μv/m/kw, it follows that on a paper with a k log log $(100/P)$ abscissa for percentage of time during which the field strength is exceeded, and with a linear ordinate for the db values, the theoretical distribution curve plots as a straight line. This graph paper is generally known as "Rayleigh distribution paper."

The points for each hour are successively plotted on Rayleigh paper. If a SLVTI is used, the points should lie on a perfectly smooth curve.

The point farthest into the 99-per cent region is determined by the setting of the lowest level on the SLVTI and by the sensitivity of the receiver. It is, of course, desirable to have a receiver sensitive enough under all conditions to run the lowest level counter almost continuously so that a 100-per cent point is always available to shape the right side of the curve correctly. This level also serves as a time check for interruptions in transmitter operation. When no points are available in this region, the extrapolation is best accomplished by drawing a line tangent at the last point of the curve. Evidence supporting this method of extrapolation was obtained in the analysis of about 1,500 hours of SLVTI data recorded at stations WHO and WTAD-FM during the winter and spring of 1949–1950.

Once the curve has been drawn, the percentages are tabulated for equally spaced db values by grouping according to the hour of the day (Table I). Equal db increments result in an easier and more accurate drawing of the average distribution curves and in greater uniformity of the tabulated data. For a picture of the inverse variation, the field strength values, which were exceeded by 1 per cent, 10 per cent, 50 per cent, 90 per cent, and 99 per cent of the time, are tabulated (Table II, see page 1387).

At WHO, for example, the lowest field intensity used in the analysis was chosen as −20 db. The successive points were spaced 2.5 db apart. It is interesting to consider the possibility of setting the SLVTI levels exactly at these values. This procedure would reduce the time spent in data reduction by about 30 per cent,

TABLE I

TABULATION OF DATA BY HOURS.

Percentage of Time that the Signal Exceeded Each of a Number of Levels Expressed in db above One Microvolt per Meter for One Kilowatt Radiated from a Half-Wave Dipole.

	LEVEL IN DB																
Date	−20	−17.5	−15	−12.5	−10	−7.5	−5	−2.5	0	2.5	5	7.5	10	12.5	15	17.5	20
0000																	
May																	
12	98.5	97.4	95.3	92	86	77	62	42	22	7.0	0.5						
13	100	99.83	99.6	99	97.4	94	85	67	39	18	5.3	0.7	0.02				
14	99.83	99.68	99.31	98.6	97	94	87.7	77	55	30	10	3.1	1	0.35	00.14	0.03	
15	99	98.3	97	95	91.5	86	76	65	46	28	10	0.3					
16	100	100	100	100	100	99.75	99.1	97.2	93	87.5	78	66	53	40	25	11	2
17	100	100	99.9	99.73	99.22	97.8	93.5	83	56	31	11	3.2	1	0.4	0.16	0.04	
0100																	
12	99.62	99.3	98.6	97.2	94.8	90	81	67	41	15	3.2	0.45	0.02				
13	100	100	100	99.85	99.3	97	86	65	43	22	7.5	2.0	0.5	0.25			
14	99.83	99.67	99.33	98.7	97.3	95	90	81	65	46	25	10	1.3				
15	98.6	97.7	96.2	94	90.2	85	76	65	49.5	31	15	5.5	0.8	0.01			
16	100	100	100	99.9	99.67	99	96.8	90.5	75	57	36	18	5	0.4			
17	100	100	100	100	99.7	98.8	95.1	84	64	44	22	7.5	1.1	0.03			
18																	

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[3] K. A. Norton, "Advances in Electronics," Academic Press Inc. New York, N. Y., vol. 1, pp. 381–423; 1948.

[4] J. W. S. Rayleigh, *Phil. Mag.*, vol. 10, pp. 73–78; 1880.

as it would eliminate the interpolation of percentages for the equally-spaced db values. However, this can be done only if the sensitivity of the receiver is exceptionally stable, and if the transmitter power is constant to within ±0.25 db. (This value of 0.25 db results from the estimation of the probable error in curve drawing.) However, the interpolation of the db values corresponding to 1 per cent, 10 per cent, 50 per cent, 90 per cent, and 99 per cent, would not be eliminated.

VI. Condensation of Data

The tabular data of db above 1 μv/m for 1 per cent, 10 per cent, 50 per cent, 90 per cent, and 99 per cent, are used to draw a graph of the continuous hourly variation of these percentages versus field intensity, during the length of time for which the data were recorded (Fig. 4). Thus, one may see at a glance the exact behavior of the field strength. The difference in db between the 1 and 99 per cent values is indicative of the fading range and is included in Table II.

TABLE II

Tabulation of db Values for Given Percentages.

Signal Level Expressed in db above One Microvolt per Meter for One Kilowatt Radiated from a Half-Wave Dipole as a Function of the Percentage of Time that This Signal Level Was Exceeded.

Date	Hour	E_1	E_{10}	E_{50}	E_{90}	E_{99}	E_1-E_{99}
May 11	1500	5.0	0.5	−4.9	− 6.8	−10.4	15.4
	1600	4.0	0	−3.0	− 5.7	− 8.9	12.9
	1700	8.0	1.9	−2.4	− 3.9	− 4.5	12.5
	1800	3.2	−0.9	−4.0	− 6.6	− 9.8	13.0
	1900	3.8	0.3	−1.9	− 5.1	− 9.0	12.8
	2000	5.0	−0.2	−1.6	− 3.8	− 6.2	11.2
	2100	11.0	0	−4.2	− 8.8	−14.3	25.3
	2200	2.0	−1.6	−4.2	− 6.7	− 9.5	11.5
	2300	1.0	−2.0	−5.8	−10.7	−16.9	17.9
May 12	0000	4.5	2.0	−3.4	−11.5	−21.6	26.1

The percentages for the equally spaced db values, which were previously grouped, are now averaged for each hour of the day, and are retabulated. This procedure results in an average distribution curve for each hour of the day; Fig. 5 (see page 1388) is a typical example. When the values for these 24 curves are weighted and averaged, the grand average distribution curve of the entire propagation run (Table III, below) is obtained.

To obtain the average diurnal variation of field strength, the 24 hourly distribution curves are examined and a graph is made of db versus hour of the day for 1 per cent, 10 per cent, 50 per cent, 90 per cent, and 99 per cent (Fig. 6, see page 1388).

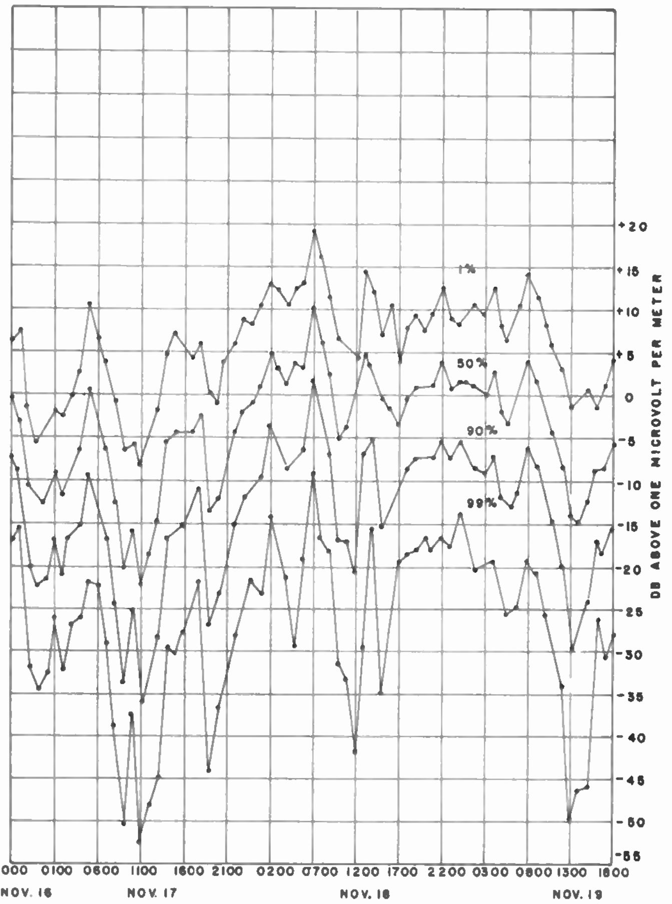

Fig. 4—An example of the continuous hour-by-hour variation of field strength. Field strength measured at WHO tower near Mitchellville, Iowa, November 16–23, 1949. One kw radiated from half-wave dipole. $h_t=41$ ft, $h_r=10$ ft, $d=86.1$ mi, and $f=410$ mc. Figure on curve represents percentage of time ordinate value is exceeded.

TABLE III

Typical Summary Table.

Hourly Summary of Field Strength at WTAD-FM Tower Near Quincy, Illinois, May 22–28, 1950. Percentage of Time that the Signal Exceeded Each of a Number of Levels Expressed in db above One Microvolt per Meter for One Kilowatt Radiated from a Half-Wave Dipole.

	LEVEL IN DB								
Hour	−20	−17.5	−15	−12.5	−10	−7.5	−5	−2.5	0
0000	98.77	98.05	96.85	94.93	91.53	86.75	78.50	66.92	50.83
0100	98.98	98.37	97.45	96.00	93.43	89.30	81.17	69.02	52.75
0200	99.02	98.42	97.50	95.88	92.93	88.30	80.35	68.92	53.33
0300	99.41	99.03	98.42	97.29	95.28	91.50	84.58	72.58	55.33
0400	99.26	98.76	98.02	96.67	94.32	90.17	82.30	70.17	51.67
1900	99.80	99.59	99.16	98.28	96.35	92.77	84.80	71.83	49.33
2000	99.71	99.46	98.97	97.90	95.85	91.80	83.33	68.58	47.67
2100	99.91	99.80	99.53	98.92	97.44	93.87	85.52	70.63	49.50
2200	98.62	97.78	96.58	94.51	91.47	87.12	79.77	69.88	55.33
2300	98.83	98.05	96.87	94.86	91.64	86.63	78.63	68.37	52.83
A	99.49	99.13	98.52	97.38	95.25	91.35	83.01	68.05	47.58

It is seen that the hour-by-hour variation, the average hourly distribution curves, the summary curve, and the diurnal variation, present a most effective picture of field-strength variation. The probable over-all accuracy of the analysis may be calculated by finding the standard deviation of the various sources of errors. A representative list of such sources is the following:

1. Transmitter power
2. Signal generator attenuator
3. SLVTI
4. Curve drawing and interpolation

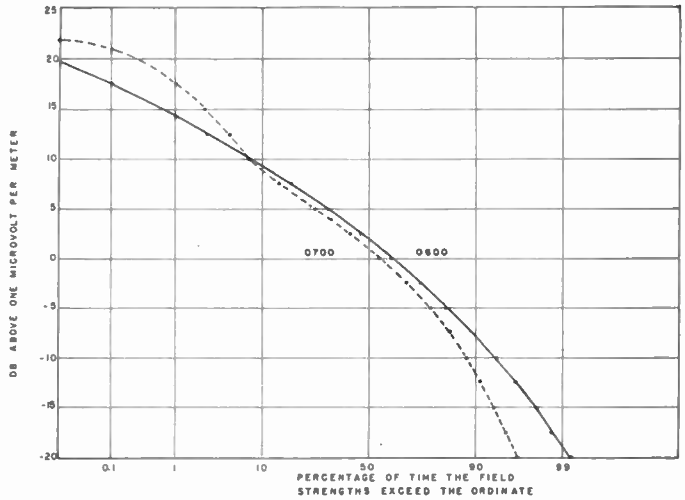

Fig. 5—Typical average hourly distribution curve. Field strength measured at WTAD-FM tower near Quincy, Illinois, April, 1950. One kw radiated from half-wave dipole. $h_t=41$ ft, $h_r=10$ ft, $d=133.9$ mi, and $f=412$ mc. Dotted and solid lines show average for ten days.

VII. Conclusion

The work in question relates to the fundamental processes involved in the analysis of propagation data. The primary aim in data reduction is to present an effective picture of field-intensity variation with a minimum amount of interpolation and tabulation. The methods and equipment described above are believed to be accurate and reliable in ascertaining the percentage of time for which various field strengths were exceeded.

If the power output of the transmitter and the sensitivity of the receiver are constant, a considerable amount of time may be saved in analysis. The Esterline-Angus analysis should, of course, be avoided, unless the SLVTI or counter panel should fail. It is seen that they afford a 10-to-1 speed-up in analysis time.

Although the SLVTI and counter-panel system, with the suggested improvements in the circuit, would seem ideal, two additional improvements are possible. The first is the use of counters which can be reset electrically every hour on the hour. This would eliminate the process of subtracting the counter readings for successive hours. Unfortunately, a small, accurate revolution counter, which can be reset electrically, is still not available on the market. A unit could conceivably be constructed using standard resetable counters, although it would be extremely cumbersome and unreliable. The second improvement is an interpolation computer in which the db and percentage values are inserted and the intermediate values electronically determined. Using, for instance, the Gregory-Newton interpolation formula, weighted for log log $(100/P)$, such a device could be designed with a few hundred vacuum tubes. The initial cost, however, would probably outweigh its advantages.

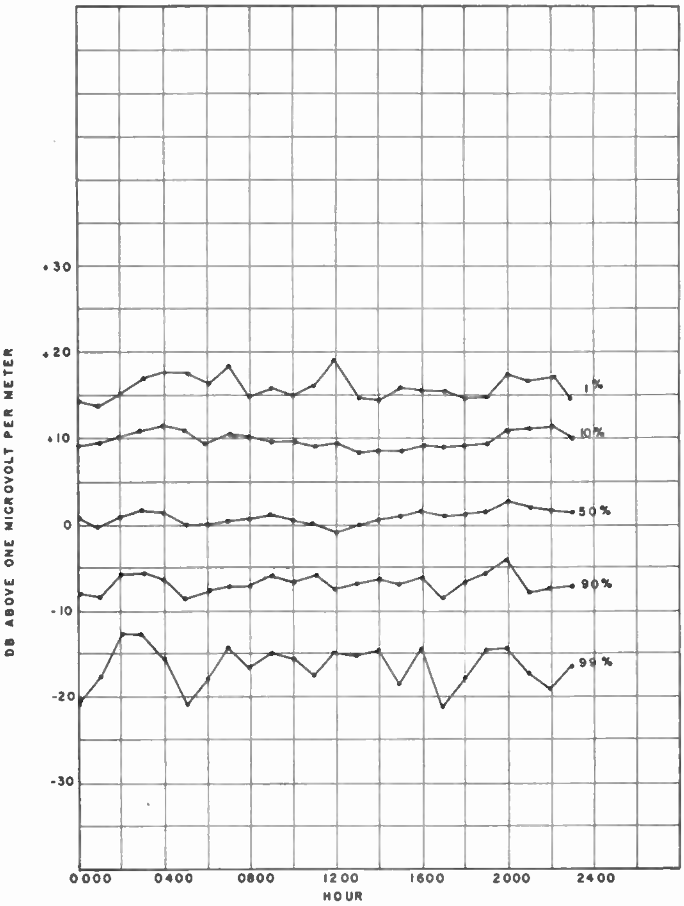

Fig. 6—An example of diurnal variation. Field strength measured at WHO tower near Mitchellville, Iowa, January, 1950. One kw radiated from half-wave dipole. $h_t=41$ ft, $h_r=10$ ft, $d=86.1$ mi, and $f=412$ mc. Figure on curve represents percentage of time ordinate value is exceeded.

Acknowledgment

Many thanks are due I. H. Gerks for his encouragement and advice. Mr. Gerks, who originally suggested that this paper be written, has prepared a discussion of the results of the 400-mc field-intensity measurements recorded at stations WHO and WTAD-FM.

Artificial Dielectrics for Microwaves*

W. M. SHARPLESS†, SENIOR MEMBER, IRE

Summary—**This paper presents a procedure for measuring the dielectric properties of metal-loaded artificial dielectrics in the microwave region by the use of the short-circuited line method. Formulas, based on transmission-line theory, are included and serve as guides in predicting the approximate dielectric properties of certain loading configurations.**

INTRODUCTION

FOLLOWING THE PUBLICATION of a recent paper on metallic delay lenses,[1] interest has been stimulated in the use of metal-loaded artificial dielectrics in the microwave field. Several papers[2–4] dealing with the theoretical analysis of metal-strip delay structures may be found in the recent technical literature. One paper[5] suggests a method of measuring the dielectric properties of such materials by the use of a microwave interferometer.

It is the purpose of this paper to present a procedure for measuring the dielectric properties of metal-loaded artificial dielectrics in the microwave region by the short-circuited coaxial-line method.[6] The use of an interferometer, a lens, or a prism structure to determine the properties of artificial dielectrics was discarded because these methods require the construction of a large sample for measurement.[7] By the short-circuited line method it is possible to measure small samples of artificial dielectric material over an extended frequency range in a single coaxial cavity, provided that the sample is prepared in such a way as to eliminate edge effects.

Simple formulas, based on transmission-line theory, are also presented and may be used to estimate proximity and frequency effects in strip-loaded artificial dielectric material.

Calculations and measurements of the dielectric constant and transmission loss are given for a few typical metal-loaded dielectric configurations. Fig. 1 shows a sketch of a strip-loaded type of dielectric material that is useful for microwaves. The measurements and calculations are based principally on this type. The measurements were all made in the wavelength range between 3 and 8 cm.

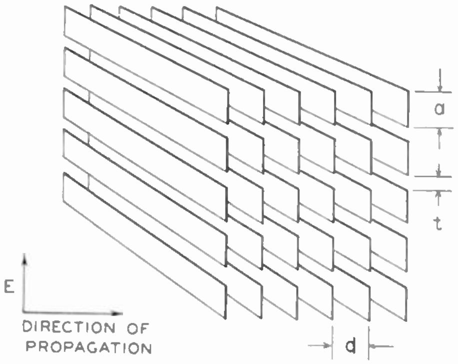

Fig. 1—Thin metallic strip artificial dielectric material.

PREPARATION AND MEASUREMENT OF TEST SAMPLES

Referring to Fig. 1, it will be noticed that if the dielectric were infinite in extent, thin metal plates might be passed through it at right angles to the voltage vector and midway between the metallic loading elements without affecting the field distribution within the dielectric. This means that a single metal-bounded sandwich section of the dielectric may be removed from the whole for measurement purposes; this done, the section removed may be bent into a doughnut shape, keeping the faces of the loading elements in the same planes as occupied in the whole. As the loaded section is now closed upon itself, all edge effects are eliminated and a tube of the material of length equal to the original depth is formed. This tube is now inserted in a short-circuited coaxial chamber for measurement,[8] as shown in Fig. 2 (following page). The spacing between the last metallic loading element and the short circuit is made $d/2$.

We now treat the metallic loaded section as if it were a sample of a homogeneous dielectric material. If a sample of homogeneous dielectric material is inserted

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* Decimal classification: R282.9×R310. Original manuscript received by the Institute, August 11, 1950; revised manuscript received, February 23, 1951.

The material contained in this paper has been submitted to the University of Minnesota in partial fulfillment of the requirements for the degree of Professional Engineer.

† Bell Telephone Laboratories, Inc., Holmdel, N. J.

[1] W. E. Kock, "Metallic delay lenses," *Bell Sys. Tech. Jour.*, vol 27, pp. 58–82; January, 1948.

[2] S. B. Cohn, "Analysis of metal-strip delay structures for microwave lenses," *Jour. Appl. Phys.*, vol. 20, pp. 257–262; March, 1949.

[3] L. Brillouin, "Wave guides for slow waves," *Jour. Appl. Phys.*, vol. 19, pp. 1028–1041; November, 1948.

[4] M. A. Brown, "The design of metallic delay dielectrics," *Jour. IEE* (London), vol. 97, part III, pp. 45–47; January, 1950.

[5] B. A. Lengyel, "A Michelson type interferometer for microwave measurements," PROC. I.R.E., vol. 37, pp. 1242–1244; November, 1949.

[6] For a general background on the various methods used in measuring the dielectric properties of ordinary homogeneous dielectric materials, the reader is referred to a recent book edited by C. G. Montgomery, "Technique of Microwave Measurements," Radiation Laboratory Series 11, Chap. 10 (by R. M. Redheffer), McGraw-Hill Book Co., Inc., New York, N. Y.; 1947.

[7] Since the writing of this paper an article has appeared by S. B. Cohn, "Electrolytic tank measurements for microwave metallic delay-lens media," *Jour. Appl. Phys.*, vol. 21, pp. 674–680; July, 1950. This paper shows how the low-frequency index of refraction of such a medium may be calculated from electrolytic tank measurements on individual loading elements.

[8] The slight "curvature effect" introduced in place of the "edge effect" is comparatively small and calculations indicate that for the cases considered the dielectric constant is decreased only a few per cent by this transformation.

at the short-circuited end of a line, the standing-wave pattern will shift by an amount dependent on the length and dielectric constant of the sample inserted. The dielectric constant of the material is calculated, as shown in Fig. 3, from the shift in the standing-wave pattern.

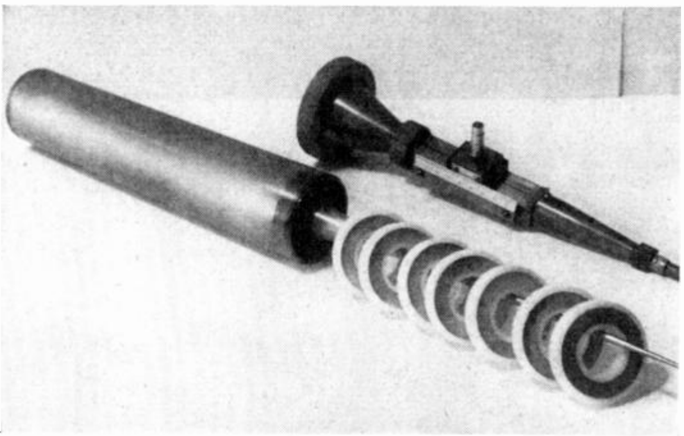

Fig. 2—Photograph of an elongated coaxial chamber with a few sections of the strip-type dielectric material.

The loss in the sample of dielectric material is calculated from the change in the magnitude of the standing wave. With the dielectric sample inserted, the standing-wave ratio will be maximum when the electrical length of the sample is a multiple of a half-wavelength, and minimum when the electrical length of the sample is an odd multiple of a quarter-wavelength. For low-loss materials, transmission-line theory gives

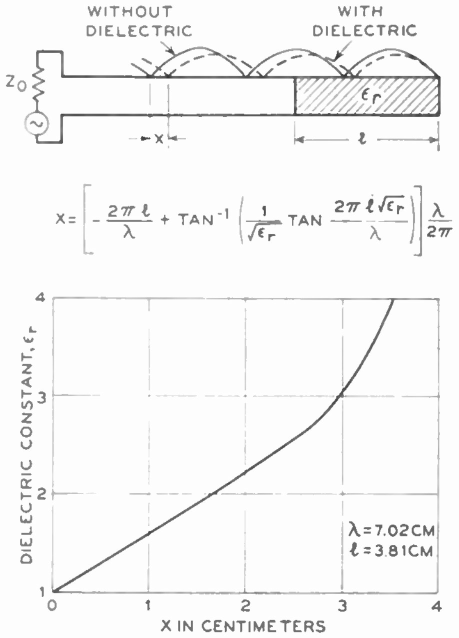

Fig. 3—Standing-wave relations used for determining the dielectric constant.

$$\text{for } l = m\frac{\lambda}{2} \quad \text{where } m \text{ is an integer}$$

$$\text{loss} = \frac{8.7}{l}\left[\frac{\sqrt{\epsilon_r}}{S_T} - \frac{\sqrt{\epsilon_r}}{S_0}\right] \text{db per meter,} \tag{1}$$

$$\text{for } l = (2m-1)\frac{\lambda}{4}$$

$$\text{loss} = \frac{8.7}{l}\left[\frac{1}{\sqrt{\epsilon_r}S_T} - \frac{\sqrt{\epsilon_r}}{S_0}\right] \text{db per meter,} \tag{2}$$

where S_T = voltage standing-wave ratio with sample inserted, S_0 = voltage standing-wave ratio with sample removed, l = length of sample in meters, and ϵ_r = the dielectric constant relative to air.

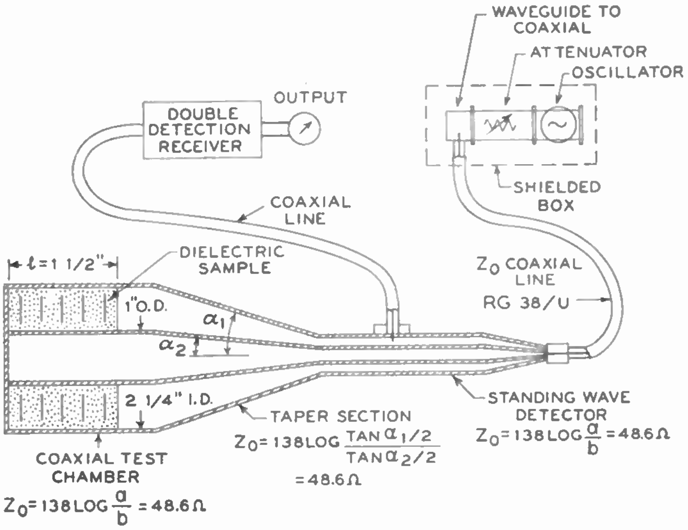

Fig. 4—Apparatus arrangement for measuring dielectric properties.

APPARATUS

Fig. 4 is a sketch of the apparatus used in making the measurements. The coaxial-test chamber used for dielectric constant measurements is shown. This chamber allows for a representative sample of the dielectric, and is satisfactory for measurements on materials such as those shown in Fig. 1, where $a+t$ is equal to $\frac{5}{8}$ inch. The enlarged section is made short to prevent the supporting of higher order modes. The tapered section is made smooth and accurately coaxial to prevent any tendency to generate unwanted modes. By taking these precautions no trouble due to multiple moding was experienced with this short test section. The remaining equipment shown on the sketch is composed of standard components.

THEORETICAL RELATIONS

The metal-loaded material shown in Fig. 1 can be represented by the strip-loaded transmission line shown in Fig. 5(a). The equivalent transmission line with lumped circuit elements is shown in Fig. 5(b). Using simple transmission-line equations, the impedance looking into the short-circuited line AB–GH (Fig. 5(a)) is

$$Z_{AB} = R_{AB} + j\omega L_{AB}$$

$$= R_{AB} + j120\pi \frac{d}{b} \tan\left(2\pi \frac{a/2}{\lambda}\right), \qquad (3)$$

where b is the assumed width of the line (in the magnetic plane at right angles to the paper). Hence, the loading inductance in the equivalent loaded line (Fig. 5(b)) is

$$L_{AB} = 20 \frac{d\lambda}{b} \tan\left(\pi \frac{a}{\lambda}\right) 10^{-8}. \qquad (4)$$

The total inductance per unit length of the equivalent line is the inductance of a tape line of spacing t plus the loading inductances,

$$L' = 40\pi \frac{t}{b} 10^{-8} + 2L_{AB}/d$$

$$= 40\pi \frac{t}{b}\left(\frac{\lambda}{\pi t} \tan \frac{\pi a}{\lambda} + 1\right) 10^{-8}. \qquad (5)$$

When d/t is small, the capacity per unit length is

$$C' = \frac{1}{36\pi\, 10^9} \frac{b}{t}.$$

In general

$$C' = \frac{1}{36\pi\, 10^9} \frac{b}{t} K. \qquad (6)$$

In a private communication, Schelkunoff gives the formula

$$K = \frac{\pi t}{2d \cosh^{-1} e^{\pi t/2d}} \qquad (7)$$

where $d \leqq t$.

The velocity in the equivalent line is

$$V' = \frac{1}{\sqrt{L'C'}} = \frac{3 \times 10^8}{\sqrt{K\left(\frac{\lambda}{\pi t} \tan \frac{\pi a}{\lambda} + 1\right)}}. \qquad (8)$$

The equivalent dielectric constant is $[3\times10^8/V']^2$; or

$$\epsilon_r = K\left(\frac{\lambda}{\pi t} \tan \frac{\pi a}{\lambda} + 1\right). \qquad (9)$$

The impedance of the equivalent line is

$$Z_0' = \sqrt{\frac{L'}{C'}} = 120\pi \frac{t}{b} \sqrt{\frac{1 + \frac{\lambda}{\pi t} \tan \frac{\pi a}{\lambda}}{K}}. \qquad (10)$$

By (9)

$$Z_0' = 120\pi \frac{t\sqrt{\epsilon_r}}{bK}. \qquad (11)$$

The attenuation of the equivalent line is largely caused by the series resistance R_{AB} Fig. 5(b)), which is the resistive component of the input impedance to line AB–GH (Fig. 5(a)). For $a < \lambda/4$,

$$R_{AB} \approx \frac{a}{2} R,$$

where R is the resistance per unit length of line AB–GH. For copper ($g = 5.8\times10^7$),

$$R_{AB} \approx \frac{a}{2} \frac{9.06 \times 10^{-3}}{b\sqrt{\lambda}}.$$

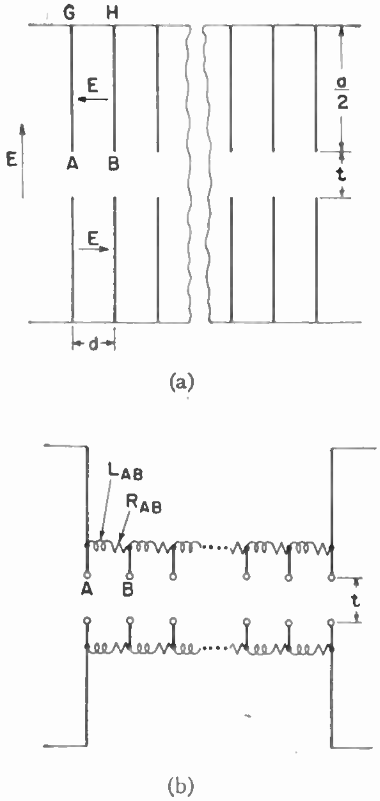

Fig. 5—Equivalent transmission line with lumped circuit elements. (a) Cross section of strip-loaded material of width b. (b) Equivalent transmission line of width b.

The approximate attenuation of the equivalent line is

$$T' = 8.7 \frac{2R_{AB}}{d} \frac{1}{2Z_0'}$$

$$= \frac{aK}{td\sqrt{\epsilon_r}\sqrt{\lambda}} 1.05\times10^{-4} \text{ db per meter (for copper)}. \qquad (12)$$

The units used in these equations are mks, and thus all dimensions are in meters.

RESULTS

Fig. 6 shows the coaxial measuring section together with several single-layer samples of loaded and unloaded dielectric materials in the foreground. On the extreme lower left is depicted the coaxial equivalent of a single section of a metallic strip dielectric of the type shown in Fig. 1. Low-loss styrofoam[9] is used as the spacer material.

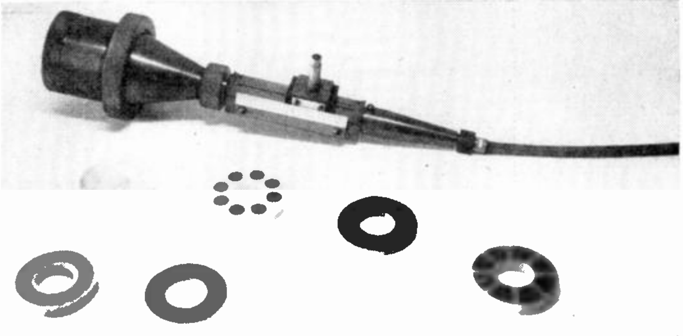

Fig. 6—Photograph showing coaxial measuring section together with several samples of dielectric materials.

Table I below lists the important dimensions, as well as the measured and calculated values for the dielectric constant ϵ_r of several different strip dielectrics. It is seen that the measured and calculated values of the dielectric constant agree in most cases to better than 10 per cent.

TABLE I

a (inch)	d (inch)	t (inch)	K (Equation (7))	λ (cm)	Dielectric Constant ϵ_r Calculated (Equation (9))	Measured (Equation (9))
$\frac{1}{2}$	$\frac{3}{32}$	$\frac{1}{8}$	0.755	3.18	8.38	
				7.00	4.13	4.32
0.365	$\frac{3}{16}$	0.26	0.766	3.18	2.39	2.10
				7.00	1.95	1.68
0.365	$\frac{3}{8}$	0.26	0.622	7.00	1.55	1.58

Loading arrangements other than the strip type have also been investigated. The solid curve of Fig. 7 shows the variation with wavelength of the dielectric constant ϵ_r of a dielectric material consisting of thin, isolated, $a=\frac{1}{2}$-inch square metal loading elements, spaced $\frac{1}{8}$ inch. The space between layers, $d=\frac{3}{8}$ inch, is again filled with low-loss styrofoam. The coaxial equivalent of a single section of this material is shown on the extreme right of Fig. 6. The dotted curve of Fig. 7 is calculated from (7) and (9) for a continuous strip dielectric with the same dimensions. The calculations for the strips gave, as expected, a slightly higher dielectric constant than the measured values for the closely spaced squares. According to (9), resonance will occur when $a=\lambda/2$. It is seen that this point is rapidly being approached at the left side of Fig. 7. To avoid anomalous dispersion regions, these particular materials should not be used for a wavelength below about 5 cm.

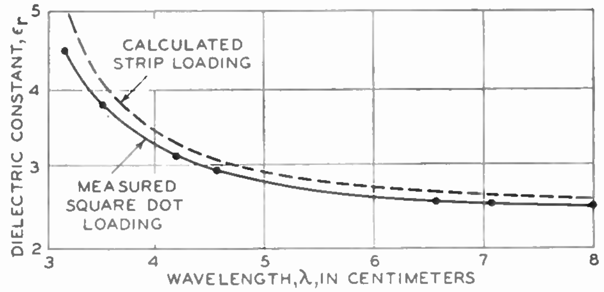

Fig. 7—Variation of the dielectric constant of two artificial dielectrics with wavelength. $a=1/2$ inch. $t=1/8$ inch. $d=3/8$ inch.

The transmission loss, as stated earlier, may be calculated from (1) or (2) if we know the magnitude of the standing-wave ratio in the chamber when the electrical length of the dielectric sample is either a multiple of a half wavelength or an odd multiple of a quarter wavelength. The standing-wave ratio for the air-filled chamber must also be known.

Fig. 8 shows a measured curve of the variation of the standing-wave ratio with sample length for a strip-type dielectric material made up in sections $d=\frac{3}{8}$ inch thick. This curve was taken at a frequency of 3,919 mc and the length was changed in steps equal to d; the circled points show the information so obtained. The exact maximum and minimum values on the curve, indicated by X's, were arrived at by slight frequency variations in the "peak" and "valley" regions. Only a single maximum or minimum need be known to solve either (1) or (2); however, when several cycles are taken, as in

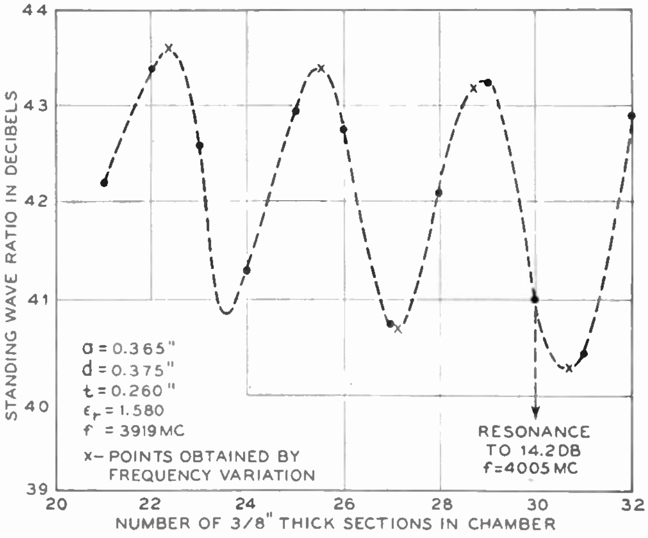

Fig. 8—Variation of the measured standing-wave ratio with the length of a dielectric sample.

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[9] Low-loss styrofoam has a measured dielectric constant of only 1.03±0.01 in the microwave range, and thus forms an attractive light-weight spacer material.

Fig. 8, the multiple solutions possible improve the accuracy of the loss measurement. By the use of the longer chamber, shown in Fig. 2, it was possible to allow for the length variation needed in obtaining such a curve.

When extremely low-loss materials are being measured, as is the case for most metal-loaded dielectrics, the observed standing-wave ratios must be corrected by taking into account the losses in the line between the probe of the standing-wave detector and the face of the dielectric sample. This correction was calculated and applied to the measurements of Fig. 8 before using (1) and (2). Furthermore, care must be taken to avoid making measurements in regions where the chamber itself may become resonant at a higher mode. One such resonance was found in taking the curve of Fig. 8 when 30 sections were in the chamber. Such resonances might be eliminated by placing the proper "killers" in the enlarged chamber section; but in most cases this will not be necessary if sufficient points are taken on a curve for a resonance to be recognized when it occurs.

The measurements shown on Fig. 8 were taken with the dielectric material pictured in Fig. 2, which was an aluminum-strip type where $a=0.365$ inch, $t=0.26$ inch, and $d=0.375$ inch. The measured transmission loss was only 0.0535 db per meter, which is equivalent to a loss tangent of 0.00012. The calculated loss, as given by (12), is 0.0358 db per meter.[10] This agreement between the measured and the calculated losses is quite satisfactory.[11] Equation (12) applies for copper-loading elements. If material of higher resistivity is used for loading, the transmission loss will increase by the square root of the resitivity ratio. Equation (12), therefore, may be used for estimating the effect of changes in the design of strip-loaded dielectric material. For example, the equation shows that as the density of the loading elements is increased the transmission loss will also increase.

CONCLUSIONS

The short-circuited coaxial-line method has been found useful for the measurement of the dielectric constant of metal-loaded artificial dielectric material in the microwave region. Samples must be prepared in a particular manner to permit measurements to be made in a coaxial cavity, but edge effects are eliminated in the process and only small-sized samples are required for the measurements. The loss factor of artificial dielectrics also may be measured by this method although considerable care is required to obtain reliable measurements on extremely low-loss materials. The measurements of dielectric constant and transmission loss are in good agreement with values calculated from approximate formulas based on simple transmission-line theory.

ACKNOWLEDGMENT

The author wishes to express his appreciation to members of the Holmdel Radio Laboratories, and especially to W. E. Legg, for assistance and co-operation during the course of this work.

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[10] This loss was calculated on the basis of the dc resistivity of aluminum, as given in handbooks. Actually, at microwaves, the loss will be a few percent higher because of surface roughness. See paper by A. C. Beck and R. W. Dawson, "Conductivity measurements at microwave frequencies," PROC. I.R.E., vol. 38, pp. 1181–1190; October, 1950.

[11] In the case of metal-loaded dielectrics, the so-called dielectric loss, for all practical purposes, is almost entirely in the form of power dissipated in the metallic loading elements.

A Precise Sweep-Frequency Method of Vector Impedance Measurement*

D. A. ALSBERG†, MEMBER, IRE

This paper is published with the approval of the IRE Professional Group on Instrumentation, and is secured through the co-operation of that Group.—*The Editor.*

Summary—**The impedance of a two-terminal network is defined completely by the insertion loss and phase shift it produces when inserted between known sending and receiving impedances.**

Recent advances in precise wide-band phase and transmission measuring circuits have permitted practical use of this principle. Reactive and resistive impedance components are read directly from a simple graphical chart in which frequency is not a parameter. The basic principle described promises attractive possibilities in many cases of impedance measurements where present methods are inadequate.

INTRODUCTION

IMPEDANCE is a fundamental property of any transmission facility; thus, from the earliest beginning of electrical transmission impedance has been a parameter of importance. As the number of links constituting a transmission facility increases, the properties of its component elements must be measured with increasing precision. Extreme examples are the transcontinental television transmission systems now nearing completion. A representative planned coaxial-cable system is made up of about 1,000 repeaters in tandem from New York to Los Angeles; a corresponding microwave radio-relay system consists of about 116 repeater

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† Bell Telephone Laboratories, Inc., Murray Hill, N. J.

stations in tandem. The most precise impedance requirements are assigned to terminations used to interconnect networks and other transmission elements and to components. Often even small impedance irregularities in their summed effect result in severe impairment of the transmission characteristics. In many cases it is inefficient to measure impedances at discrete frequencies, and sweep-frequency measurements must be made in order to detect significant irregularities. Often impedances must be measured in positions which do not permit short leads to the measuring circuit.

Impedance bridges do not lend themselves readily to sweep-frequency techniques, and must be located close to the impedance to be measured. Similarly, slotted lines and the three-voltmeter method do not adapt themselves to sweep methods.

Numerous devices, such as hybrids and directional couplers, have been developed which use variations of the reflected-energy principle. These adapt themselves to sweep-frequency techniques. However, they all require some form of mutual coupling, which restricts the bandwidth over which they are usable for sweep-frequency methods. The internal phase-shift changes and attendant calibration corrections of these devices are usually so severe that sweep-frequency measurements have only been used to measure the magnitude but not the phase, of an impedance.

Impedances which have to be measured with the greatest precision are those in the vicinity of transmission-line impedances, terminations, and connectors. Commercially available bridges do not possess adequate accuracy in many applications.

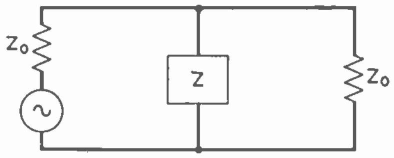

Fig. 1—Impedance shunt insertion.

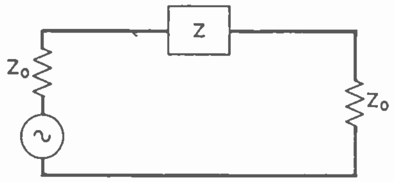

Fig. 2—Impedance series insertion.

GENERAL PRINCIPLE

Since the development of highly precise phase- and transmission-measurement methods, a principle of impedance measurement has become practical, which, heretofore, has had little attention despite its many inherent advantages. Let us assume, in the ideal case, a transmission line terminated on both the sending and receiving ends with its characteristic impedance. If we now insert an impedance Z in shunt (Fig. 1) or in series (Fig. 2) with this line, the insertion loss and phase shift thus produced are directly a measure of the impedance Z in terms of the characteristic impedance of the line. A measuring circuit using this principle consists essentially of an insertion loss and phase-shift measuring circuit (Fig. 3) incorporating a fixed reference transmission line and a measuring transmission line having some provision for inserting the two-terminal impedance Z into the line.

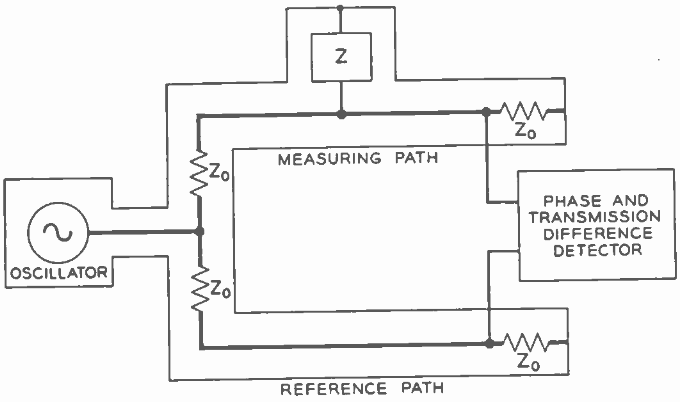

Fig. 3—Basic measuring circuit.

Some advantages of the insertion loss and phase principle are immediately evident.

1. The perturbation created in the transmission line by insertion of the impedance is directly the measure of the impedance. No mutual coupling or probes are required as an intervening link.
2. Impedance is measured as a transmission property, thus constant impedance corresponds to constant transmission, independent of frequency.

These two properties in combination permit precise sweep-frequency measurements over a much wider band of frequencies than are possible with any other known method.

3. As the transmission line is terminated in its characteristic impedance, no restriction is placed upon its length. Thus the junction point for insertion of the unknown impedance can be as remote from the measuring circuit as desired, without any impairment of accuracy.

Definitions and the exact equations for the relationship of insertion loss and phase and impedance or admittance are given in the Appendix.

In practice it would be cumbersome to solve these equations for every measurement. To avoid this, convenient charts have been prepared which are entered with the measured insertion loss and phase so as to read directly resistive and reactive impedance components.

The first chart (Fig. 4) covers all impedances containing positive resistances, and is drawn on a normalized impedance basis. The impedance grid is similar to the familiar Smith chart. Insertion loss and phase are represented by circles and radii centered on the point of zero impedance. The chart yields answers for either shunt impedance or series admittance to two significant figures. It is used for solving problems where this ac-

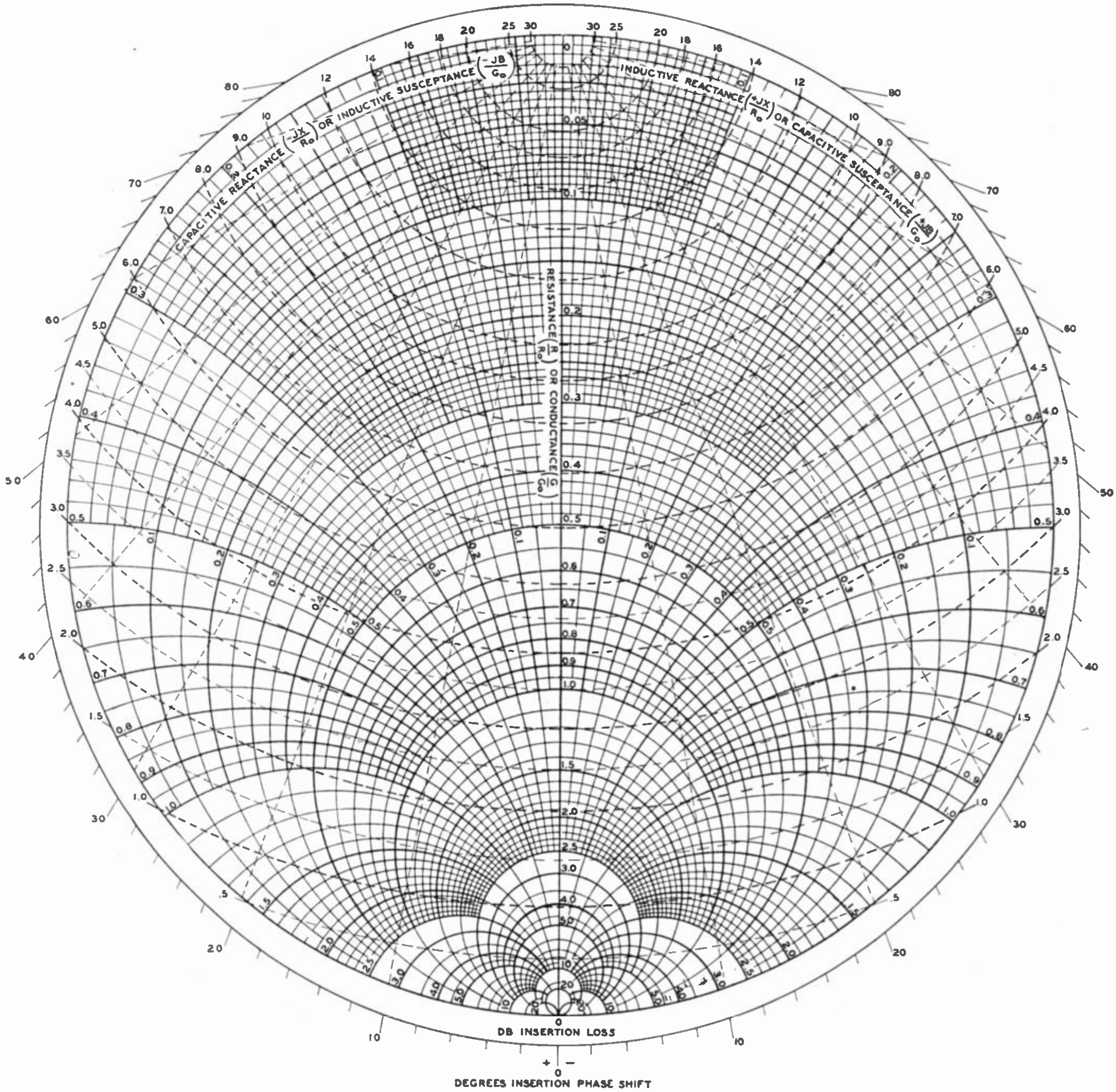

Fig. 4—Universal impedance and admittance chart. Insertion and phase shift produced by normalized shunt impedance (Z/R_0) or series admittance (Y/G_0) between terminations R_0 or G_0.

curacy is adequate, and it is used also for evaluating the range of loss and phase shift a measuring set must cover in order to measure a predetermined range of impedances.

The following chart (Fig. 5) was drawn specifically for 75- and 50-ohm circuits, to cover a range of impedance of ±20 per cent about the circuit impedance. It is in this region where the majority of the most precise impedance measurements have to be made. It may be read to an accuracy of about 0.1 per cent. Resistance and reactance are read directly in ohms. It should be noted that in order to cover an impedance variation of ±20 per cent, or 10 per cent reflection coefficient, about the circuit impedance, a loss range of only ±0.5 db and a phase range of ±3.5 degrees are required.

If other impedance ranges have to be read with higher precision than is possible with Fig. 4, existing published equalizer charts[1] may be used, or, when warranted, special charts may be readily constructed.

Limitations and Errors of Measurement

As may be seen from inspection of the impedance charts, accurate measurement of impedances relies on the ability to measure with high accuracy small losses and phase shifts. The alternative reflection loss and phase method relies on measuring larger losses and phase shifts with less accuracy.

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[1] "Equalizer Charts," Bell Telephone System Monograph B-1643.

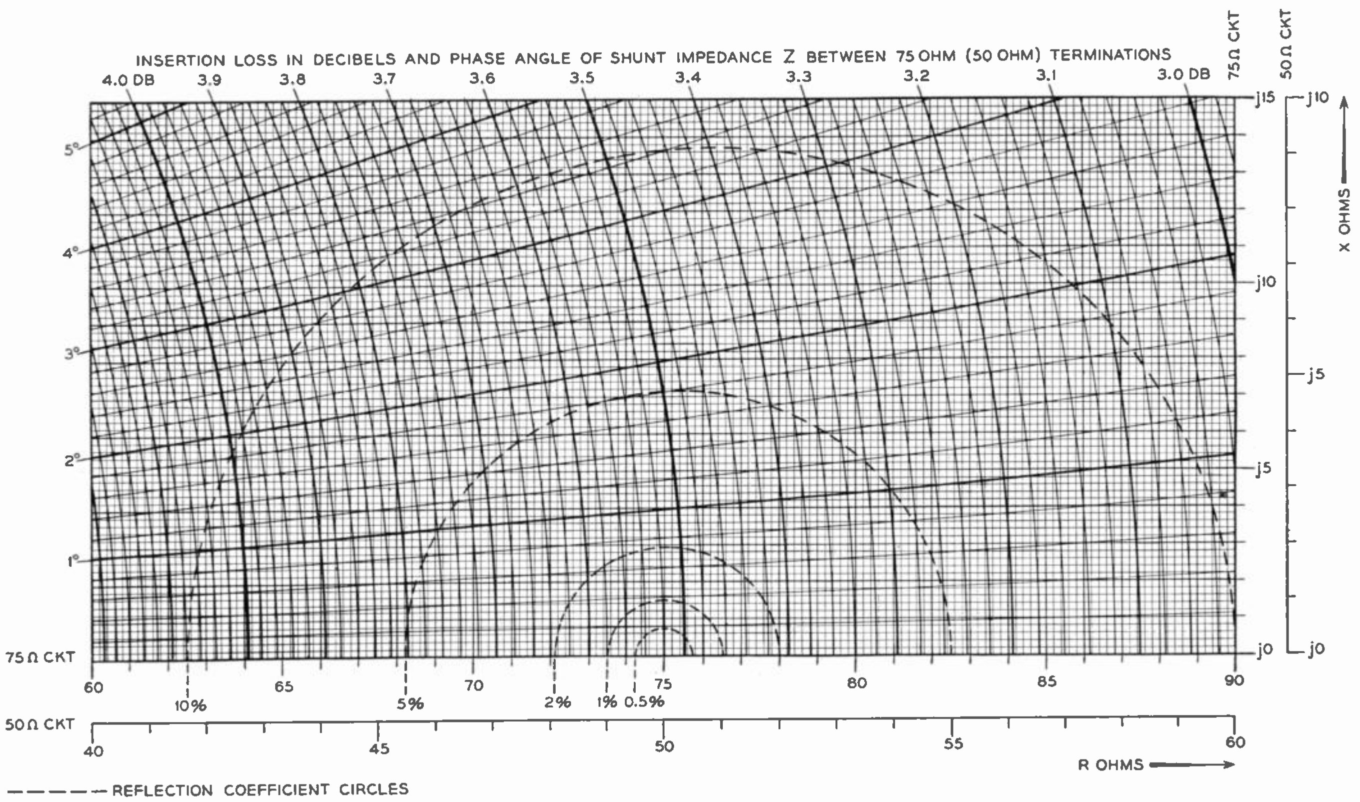

Fig. 5—Impedance chart for impedances in the vicinity of 50 and 75 ohms.

At high frequencies the effects of residual capacitances and inductances are often sufficiently large to make it necessary to use shields. Such shields can be grounded conveniently when shunt insertion is used. When using series insertion for measuring large impedances, the uncontrolled stray impedances to ground limit the accuracy of measurement, and yet many satisfactory results have thus been obtained in measurements on spurious resonances of piezoelectric crystals.

The effect of the fixture used to insert the unknown impedance into the phase- and loss-measuring circuit may be difficult to evaluate, but it may be eliminated by a simple procedure. In a coaxial circuit, a tee is inserted in the line connecting the test circuit generator to the detector. To establish the reference zero of the circuit, the open end of the tee is terminated with a known impedance standard and the test circuit is adjusted to read the nominal loss and phase shift associated with the impedance standard. For instance, we find from Fig. 5 that a 75.00-ohm impedance shunted across a 75.00-ohm circuit produces a 0.0-degree insertion phase shift and a 3.52-db loss. This setting is then as accurate as the impedance standard is known. Using coaxial-line techniques and thin film resistors, it is possible to design standards for frequencies up to 80 mc, whose reactance component is less than 0.1 per cent of the nominal impedance and whose resistance component can be determined by dc measurement.

As the coaxial cables used to connect the measuring circuit have a surge impedance which is a function of frequency, and which is subject to manufacturing tolerances, the impedance Z_0 is not constant over a wide frequency band and is dependent on the specific cable used. If the direct-reading chart (Fig. 5), which assumes a constant R_0 of 75 ohms or 50 ohms, is used, an error then results. To determine if this error is negligible, the formulas derived in the Appendix may be used.

The impedance measured is the one at the junction point of the unknown impedance with the transmission line. Often the impedance to be measured cannot be inserted physically at the junction point, and a coaxial line is used between the coaxial tee and the unknown impedance. If the surge impedance of this line and the unknown impedance are significantly different, the measured impedance value must be corrected for the effects of the line by use of conventional transmission-line theory, most conveniently by use of the Smith Chart.[2] To avoid this correction, it is advisable to make the connection between unknown impedance and the junction point as short as possible, and to lengthen instead the cables connecting the junction point to the generator and detector. The impedance of these connecting cables can be measured precisely, however; if it is found to be substantially different from the design impedance R_0, significant errors can be computed from (20) in the appendix. Thus, the unknown impedance can be measured as remotely from the insertion-phase and -loss measuring circuit as desired without impairment of accuracy.

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[2] P. H. Smith, "An improved transmission-line calculator," *Electronics*, vol. 12, pp. 29–30; January, 1939; also, *Electronics*, vol. 17 pp. 130–133; January, 1944.

APPLICATION

The method described was demonstrated first by using an improved version of a phase and transmission measuring circuit, previously described.[3] The improved version (Fig. 6) has the following characteristics:

Frequency range	0.05 to 20 mc (9 octaves)
Loss range	0 to 70 db
Accuracy (absolute, recording)	±0.02 db (up to 30-db loss) ±0.1 degree
(differential, recording)	±0.01 db ±0.05 degree
(differential, with special technique)	±0.002 db ±0.01 degree
Zero characteristic less than	±0.1 degree and ±0.02 db from 0.05 to 20 mc.

Circuit impedance 75.0 ohms.

The circuit is self-tuning and the test-frequency oscillator is motor driven at the rate of 1 mc for sweep measurement from 0.05 to 20 mc. The insertion phase shift and loss may be read directly on the indicating meter, or may be automatically recorded.

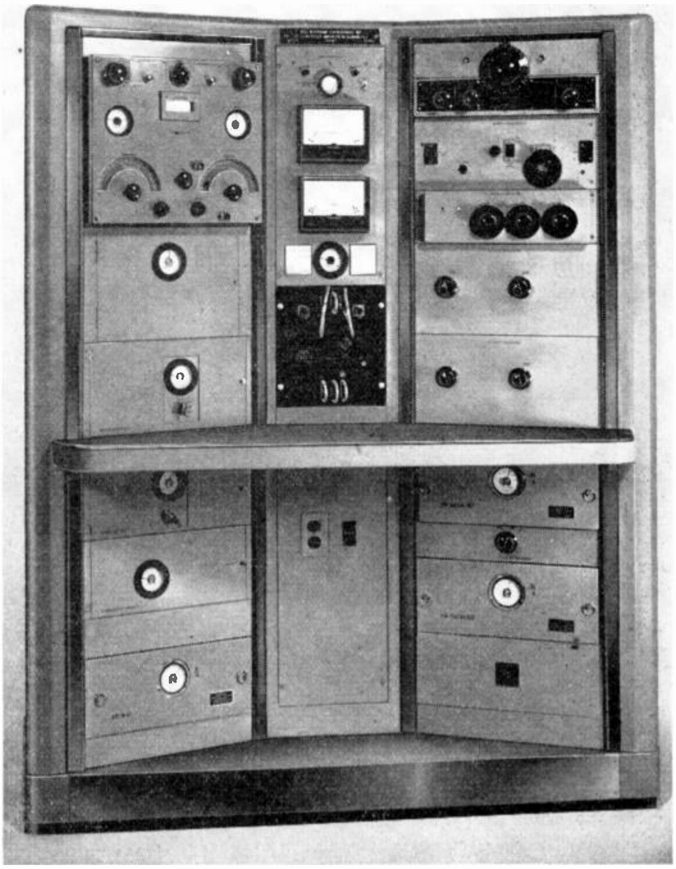

Fig. 6—50-kc to 20-mc phase, transmission and delay measuring set.

From Fig. 5 it is evident that, with a differential accuracy of 0.01 db and 0.05 degree, impedance accuracies of ±0.25 per cent are attained in the vicinity of 75 ohms. The 75-ohm coaxial tee (Fig. 7) is inserted into the test branch of the measuring circuit and the unknown impedance is inserted in the open arm of the tee. The motor drive of the test oscillator is started and the frequency is swept from 0.05 to 20 mc, or over any portion of the band. Deflections of the indicating meters reveal any impedance variations immediately.

Inspection of chart Fig. 4 shows that as the impedance increases, accuracy decreases, when shunt insertion is used. For instance, from Fig. 4 for an impedance of 750 ohms ($R/R_0=10$) a resolution of 0.01 db corresponds to a 20-ohm or 2.5-per cent uncertainty. However, high accuracy is attained for low impedances using shunt insertion. For example, from Fig. 4, an impedance of 1.5 ohms ($R/R_0=0.02$) corresponds to a loss of about 28 db; an accuracy of 0.02 db then yields an impedance uncertainty of about 0.003 ohms, or 0.2 per cent.

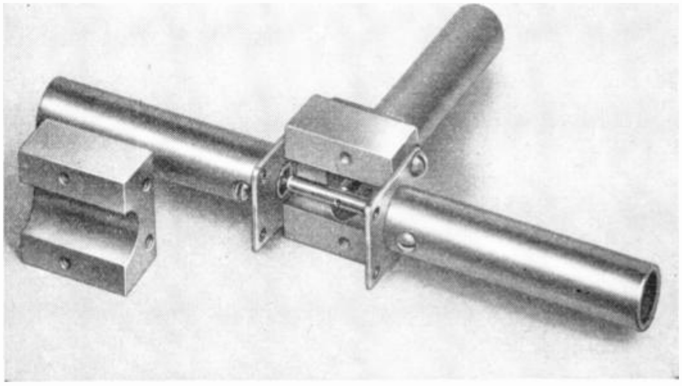

Fig. 7—75-ohm coaxial tee.

The differential accuracy limit of 0.01 db and 0.05 degree is set by the circuit stability. By increasing the indicating-meter sensitivity and averaging a number of consecutive measurements, it is possible to attain a statistical differential accuracy of 0.002 db and 0.01 degree, corresponding to impedance accuracies of ±0.05 per cent at 75 ohms.

The validity of the general impedance-measurement method was verified at frequencies up to 80 mc. The method has found practical use in the measurement of coaxial repeaters, quartz crystals, cable impedances, and precise terminations.

EXAMPLES

An example of a recording sweep measurement is shown in Fig. 8 on page 1398: 158 feet of RG 6/U cable were terminated with a resistance standard of 75.10 ohms. Both the transmission and phase records have two traces, one the reference zero trace, and one the measurement trace, the net measurement value being the difference between the two traces. As the deviation of the measured impedance from 75 ohms was small, phase and loss departure readings could be labeled directly in ohms without the need for conversion by use of graphical tables. The irregularity of the measurement trace shows clearly the frequency dependence and irregularity of the characteristic impedance of RG 6/U cable, a common fault of all flexible coaxial cables.

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[3] D. A. Alsberg and D. Leed, "A precise direct reading phase and transmission measuring system for video frequencies," *Bell Sys. Tech. Jour.*, vol. 28, pp. 221–238; April, 1949.

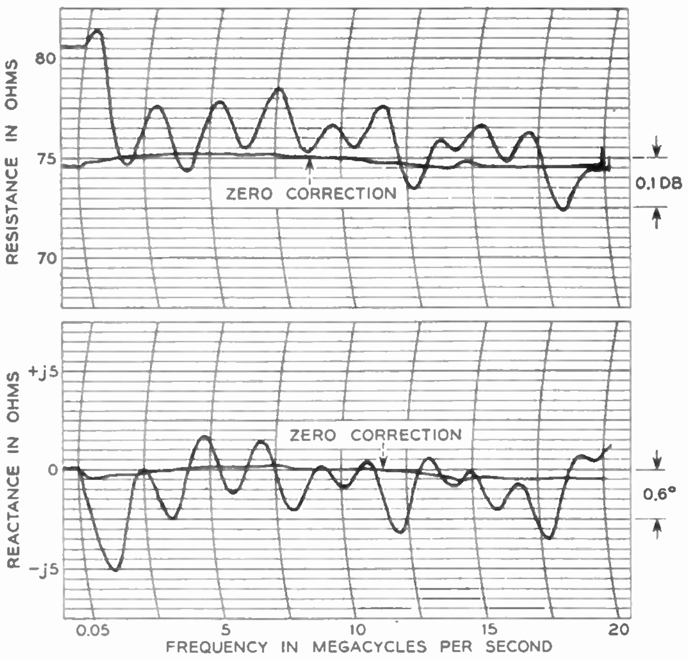

Fig. 8—Recording of input impedance of 158 ft of RG 6/U coaxial cable terminated in 75.10 ohms.

Another important application has been the measurement of quartz-crystal primary parameters. Fig. 9 shows the conventional equivalent circuit of a quartz crystal. The measurement procedure is as follows in a 75 ohm circuit:

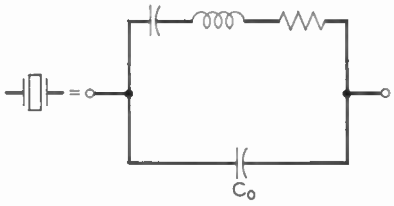

Fig. 9—Equivalent circuit for piezoelectric crystal.

The low-pass filter (Fig. 10) is terminated with a 75-ohm resistance standard. The crystal is inserted, as shown, and at a frequency remote from resonance the capacitor C_1 is adjusted until the filter is transparent to 75 ohms. Thus, the static crystal capacity C_0 is absorbed into the filter. After this initial adjustment the filter and crystal are inserted in the measuring circuit, as shown in Fig. 11. The resulting measurement then contains only the effective impedance of the series-resonance elements. As the static capacitance C_0 has been absorbed into the filter, zero phase shift occurs exactly at resonance. With this method the residual resistance, or Q, of high-frequency crystals, has been determined with higher accuracy than heretofore possible.

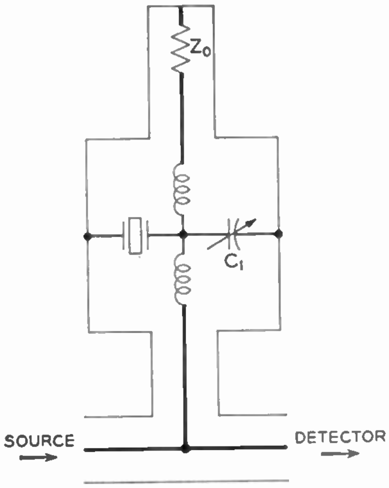

Fig. 10—Adjustment of low pass filter to absorb static crystal capacitance C_0.

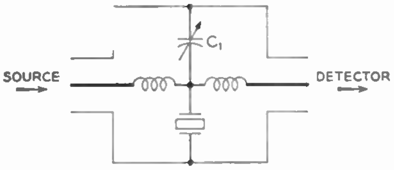

Fig. 11—Insertion of crystal and low pass filter in impedance measuring circuit.

A typical low-pass filter structure for crystal measurements is shwno in Fig. 12.

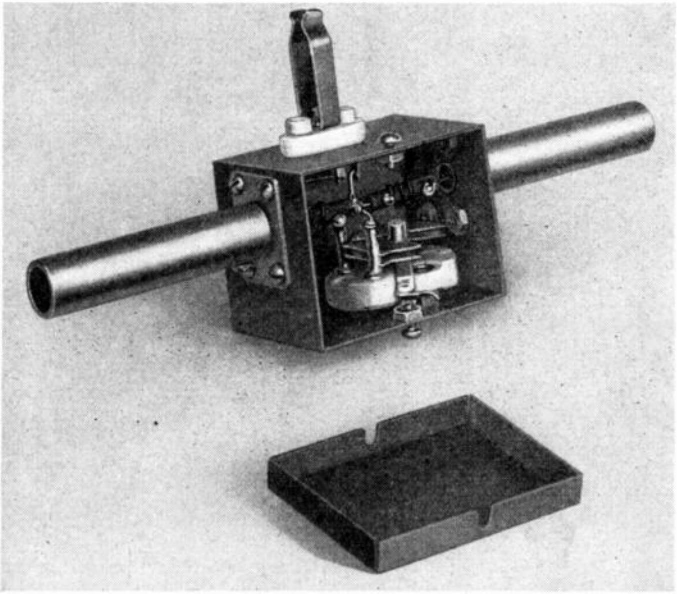

Fig. 12—Typical low-pass filter for piezoelectric crystal impedance measurements.

CONCLUSION

The examples used were cited as experimental verification of the possibilities of the basic method. It is the primary purpose of this paper to stimulate others to take advantage of the properties of the insertion phase and loss principle, which is a general principle of measurement not restricted to the coaxial measurement circuits. The fundamental properties are as follows:

1. The impedance is measured as a transmission property; thus, constant impedance corresponds to constant transmission, and frequency is not a parameter. This method may be adapted ideally to sweep-frequency measurement techniques.
2. The perturbation created in the transmission line by insertion of the impedance is the quantity directly of interest; in contrast to methods like slotted lines, where the disturbance created by probes is a source of error and limitation, no mutual impedance is required.
3. The measurement relies on the ability to measure precisely small phase shift and potential differences in contrast to methods using the reflected energy principle, such as hybrids, slotted lines, and directional couplers which require measuring large phase shifts and potential differences, though with less accuracy.
4. Measurement may be made at a remote distance from the measuring circuit.

Where an accurate impedance-measurement device for a limited impedance range is required, greatly simplified but very precise phase- and transmission-measuring circuits can be built. Where the general transmission measurement of a communications device requires more elaborate phase- and transmission-measuring facilities, the mere addition of a line-bridging device converts these facilities, at practically no added expense, to measure impedance as well.

In addition to enabling sweep measurements, the application of the insertion-phase and -loss principle has resulted in more precise vector-impedance measurements at high frequencies than have been obtained by other available methods. The method has not yet been fully exploited, and it offers the possibility of precision measurements at very high frequencies, now only attained in the best bridges at much lower frequencies.

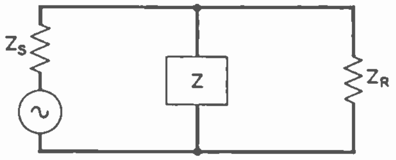

Fig. 13—Impedance shunt insertion definitions.

APPENDIX

Theory

The insertion loss α and phase shift β, produced by insertion of a two-terminal impedance between known sending and receiving impedances, is given by the following equations (see bibl. ref. 1):

For shunt impedance, (Fig. 13).

$$Z_0 = 2\frac{Z_r Z_s}{Z_r + Z_s}, \tag{1}$$

then

$$e^{\alpha+j\beta} = 1 + \frac{1}{2Z/Z_0} \tag{2}$$

if Z_0 is a pure resistance R_0.

$$e^{\alpha+j\beta} = 1 + \frac{1}{2Z/R_0} = 1 + \frac{1}{2(R/R_0 + jX/R_0)} \cdot \tag{3}$$

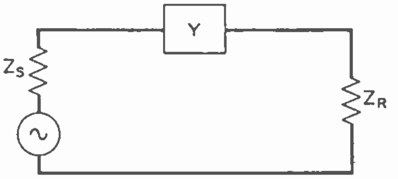

Fig. 14—Admittance series insertion definitions.

For series admittance (Fig. 14) let

$$Y_0 = \frac{2}{Z_s + Z_r}, \tag{4}$$

then

$$e^{\alpha+j\beta} = 1 + \frac{1}{2Y/Y_0} \tag{5}$$

if Y_0 is a pure conductance G_0.

$$e^{\alpha+j\beta} = 1 + \frac{1}{2Y/G_0} = 1 + \frac{1}{2(G/G_0 + jB/G_0)} \cdot \tag{6}$$

Both (3) and (6) are of the general form

$$e^{\alpha+j\beta} = 1 + \frac{1}{u + jv} \cdot \tag{7}$$

In (7) the loci of constant loss and phase shift are a family of orthogonal circles if v and u are plotted in rectangular co-ordinates.[1]

If we apply the complex transformation

$$u + jv = \frac{1 + \zeta}{1 - \zeta}, \tag{8}$$

where ζ is a complex number, to (7), it will result in a family of orthogonal circles representing u and v (Fig. 4). These are identical to the familiar Smith chart,[2] except for a scale factor of $\frac{1}{2}$ arising from the definition of Z_0 and Y_0. The outside rim of the chart $u=0$ is the unit circle in the ζ plane. The lines of constant insertion loss are now circles centered on $u=0$, $v=0$, the locus of which is defined by

$$\alpha = -20 \log \left| \frac{1+\zeta}{2} \right| \text{ decibels} \qquad (9)$$

Lines of constant inertion phase shift are the radii of the loss circles in the ζ plane.

Equations (3) and (6) could also be written in terms of shunt admittance and series impedance, resulting in a chart similar to Fig. 4. This subject will be discussed more extensively by the author in a forthcoming article on equalizer charts.

Error Computation

Error occurs when actual Z_0 is different from R_0 assumed in direct reading charts.

From (1) let us define

$$Z_0 = R_0 + \delta_z. \qquad (10)$$

The unknown impedance is defined as

$$Z = R_0 + \epsilon_z. \qquad (11)$$

The measurement procedure fixes the loss and phase readings of the measuring set to the values prescribed by the impedance chart for the design impedance R_0 when the impedance standard is inserted, even though Z_0 differs from R_0 by δ_z. This introduces an error Δ in loss and phase readings. From (2) we can write the following:

The nominal value of the table for zero reference is

$$e^{\alpha+j\beta} = 1 + \frac{1}{2R_0/R_0} = 1.5. \qquad (12)$$

The actual value measured for zero reference is

$$e^{\alpha'+j\beta'} = 1 + \frac{1}{2R_0/(R_0+\delta_z)} = 1.5 + \frac{\delta z}{2R_0}\cdot \qquad (13)$$

The nominal value of the table for Z is

$$e^{\alpha_1+j\beta_1} = 1 + \frac{1}{2Z/R_0}\cdot \qquad (14)$$

The actual value measured for Z is

$$e^{\alpha_1'+j\beta_1'} = 1 + \frac{1}{2Z/(R_0+\delta_z)}\cdot \qquad (15)$$

Therefore,

$$\Delta = -(\alpha+j\beta)+(\alpha'+j\beta')+(\alpha_1+j\beta_1)-(\alpha_1'+j\beta_1'). \qquad (16)$$

Substituting (12), (13), (14), and (15) in (16), we find as a complete expression for Δ

$$\Delta = lg\left[1 + \frac{\frac{\delta_z}{3R_0}\left(1+\frac{R_0}{2Z}\right) - \frac{\delta_z}{2Z}}{\left(1+\frac{R_0}{2Z}\right)+\frac{\delta_z}{2Z}}\right]. \qquad (17)$$

If δ_z is small, this yields

$$\Delta = lg\left(1 + \frac{\delta_z}{3R_0} - \frac{\delta_z}{2Z+R_0}\right)$$

$$= lg\left(1 + \frac{\delta_z}{3R_0} - \frac{\delta_z}{3R_0+2\epsilon_z}\right). \qquad (18)$$

If the deviation ϵ_z of Z from R_0 is small, (13) approximates to

$$\Delta = lg\left(1 - \frac{\delta_z\epsilon_z}{4.5R_0^2}\right), \qquad (19)$$

or as δ_z and ϵ_z are small,

$$\Delta = -\frac{\delta_z\epsilon_z}{4.5R_0^2} \text{ (nepers, radians).} \qquad (20)$$

For example, let

$$\delta_z = \delta_R + j\delta_X \text{ ohms} \qquad (21)$$

and

$$\epsilon_z = \epsilon_R + j\epsilon_X \text{ ohms.} \qquad (22)$$

If (16) and (17) are inserted into (15), assuming $R_0 = 75\omega$ and converting to db and degree,

$$\Delta \text{ loss} = 3.4 \times 10^{-4}\,(\delta_R\epsilon_R - \delta_X\epsilon_X) \text{ decibels} \qquad (23)$$

$$\Delta \text{ phase} = 2.26 \times 10^{-3}\,(\delta_X\epsilon_R + \delta_R\epsilon_X) \text{ degree.} \qquad (24)$$

In practice Δ can be neglected in most applications. For a given δ_z the error Δ decreases linearly as the unknown impedance Z approaches R_0.

When the error, as given by (20), becomes sufficiently large to warrant correction of the chart readings, at times it will be found more convenient to use the actual Z_0 and Y_0, as defined in (1) and (4), directly in reading the charts rather than to use the arbitrary chart-design impedance plus correction from (20).

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Speech-Reinforcement System Evaluation*

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This paper is published with the approval of the IRE Professional Group on Audio, and has been secured through the co-operation of that Group.—*The Editor.*

***Summary*—Speech-reinforcement systems in six large auditoriums were evaluated, using subjective rating tests, word-articulation tests, and, in two cases, a new test method. This method, called the "terminal-word test," makes possible the quantitative measurement of speech intelligibility for a sound system in actual use. A graphical method is presented for calculating the performance of a sound system in which account is taken of the frequency response of the system, the reverberation time of the room, the directivity index of the loudspeaker, and the room noise. Test results indicate that a flat frequency response in the range between 400 and 4,000 cps is required for good intelligibility. The graphical method indicates that little further increase in intelligibility would result from extending this range upward or downward. If the loudspeaker system is sufficiently directive in this frequency range and properly located in the room, room reverberation has little effect on speech intelligibility.**

I. INTRODUCTION

EARLY IN 1949, in connection with the Mid-Century Convocation at the Massachusetts Institute of Technology, the authors designed three sound systems for use in auditoriums, ranging in size from 237,000- to 5.5-million cubic feet. These auditoriums were highly reverberant when empty, and considerable difficulty in obtaining satisfactory speech intelligibility was anticipated.

After preliminary study, it was decided to restrict the frequency responses to the interval between 200 and 7,000 cps, which is the range known to be of principal importance to speech intelligibility.[1,2] Two purposes were served by this decision: excitation of the low-frequency room resonances was avoided, and the cost of the loudspeaker installations was reduced. Although it was expected that a loss of naturalness would be observed because of the restricted low-frequency response, we were surprised to learn that many people judged these systems superior to average installations. Our interest was aroused in performing a series of experiments to determine some of the psychological factors involved in sound-system design for speech. No effort has been made to extend the studies to music.

The experimental program was planned to yield answers to the following questions: (a) Should the bass response of the system be restricted to preserve intelligibility in reverberant space? (b) If bass response is restricted, to what extent is naturalness impaired? (c) Does the articulation index theory apply to sound systems in reverberant surroundings? (d) What is the best location for the loudspeakers? (e) Can a suitable test be devised for measuring the performance of a sound system under normal operating conditions?

II. DESCRIPTION OF SYSTEMS

Our studies eventually included a cathedral, a memorial auditorium, and a highly reverberant industrial building, in addition to the three auditoriums mentioned above. In every case the loudspeakers consisted of one or more multicellular high-frequency units selected and installed so that from any seat the listener could see the throat of one of the horn cells. In Walker Memorial (Morss Hall) at M.I.T. (volume 237,000 cubic feet), a horn two cells high and five cells wide was used. In the Rockwell Cage at M.I.T. (volume 1,100,000 cubic feet), two horns, each two cells high and five cells wide, and one horn two cells high and four cells wide, were used. In the Boston Garden (volume 5,500,000 cubic feet), two horns, each three cells high and four cells wide, and one horn three cells high and six cells

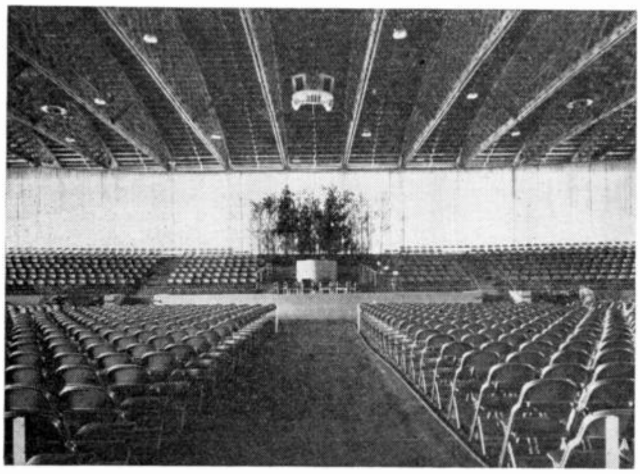

Fig. 1—View of Rockwell Cage showing loudspeaker installation. The speakers are mounted approximately 15 feet forward of and 20 feet higher than the microphone.

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* Decimal Classification: R391×534. Original manuscript received by the Institute, March 1, 1951; revised manuscript received, June 18, 1951.

† Massachusetts Institute of Technology, Cambridge, Mass.

[1] L. L. Beranek, "Design of speech communication systems," PROC. I.R.E., vol. 35, pp. 880–890; September, 1947.

[2] N. R. French and J. C. Steinberg, "Factors governing speech intelligibility," *Jour. Acous. Soc. Amer.*, vol 19, pp. 90–119; January, 1947.

wide, were used. In the cathedral (volume 1,400,000 cubic feet), the equivalent of a two-cell high, five-cell wide horn was used. In the memorial auditorium (volume 1,160,000 cubic feet), a two-cell high, five-cell wide horn was used. In the industrial building (volume 990,000 cubic feet), two horns, one two cells high and six cells wide, and the other two cells high and four cells wide, were used. In addition, in the Rockwell Cage and in the memorial auditorium, two direct-radiator low-frequency units were used during portions of the experiments. A general view of the Rockwell Cage is shown in Fig. 1.

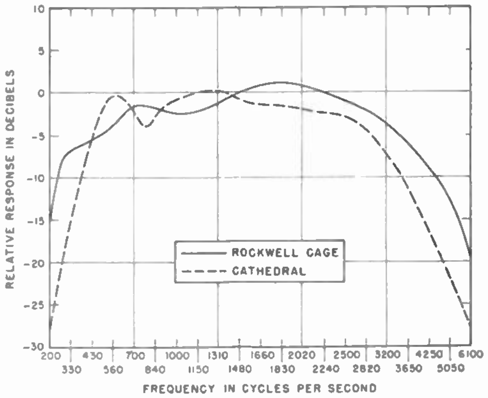

Fig. 2—Response of high-frequency channels of speech-reinforcement systems in Rockwell Cage and cathedral.

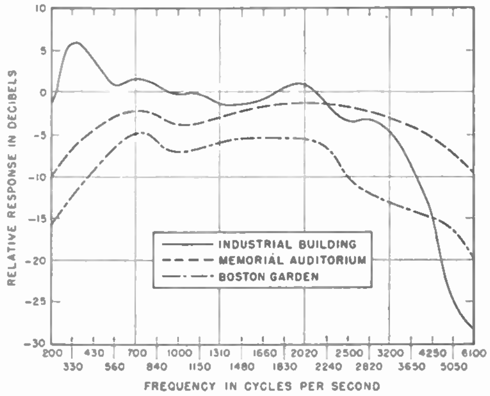

Fig. 3—Response of high-frequency channels of speech-reinforcement systems in industrial building, memorial auditorium, and Boston Garden.

In each auditorium, the sound pressure level produced by the sound system was measured as a function of frequency at several locations. In each test, the microphone was replaced by a warble-tone source having the same nominal internal impedance. The frequency response of the microphone was then added to that response curve to obtain an over-all response characteristic. Space averaging of the data was obtained by moving the microphone back and forth a few feet at each location. The resulting curves for the high-frequency channels alone, of five of the systems tested, are shown in Figs. 2 and 3. The abscissas of these figures show frequency in cycles per second plotted on a scale such that, in quiet, equal distances along the scale are of equal importance to speech intelligibility.[1,2] For example, the frequency range from 700 to 1,310 cps is as important to speech intelligibility as the range from 2,020 to 3,200 cps. Equipment by three different manufacturers was used. It should be noted that in nearly every case the system response falls off rapidly at high frequencies.

Distributions of relative sound levels over the main-floor seating area in four of the auditoriums are shown at two frequencies in Figs. 4 and 5. If 1-db contours were plotted on these figures, it would be seen that the sound levels varied over a range of as little as 5 db to as much as 12 db. Based on our observations, it is desirable to adjust the system for a maximum variation of about 6 db. However, this condition was achieved only in the memorial auditorium.

In two of the installations, considerable trouble from acoustic feedback was encountered when the systems were first tested. The feedback point was reduced 3 db in one case and seven in the other, by blocking off (with absorbent cotton) those horn cells which pointed directly at the walls on either side of the stage.

III. ACOUSTICAL CONDITIONS

The reverberation times as a function of frequency were measured in five of the auditoriums with no audience present (see Fig. 6). In the cathedral alone, data also were taken with an audience. The reverberation times were then calculated for the number of people present during the tests[3] (see Fig. 7). Detailed reverberation data were not taken in the industrial building. At 500 cps, stop-watch observations indicated a reverberation time of about 5 seconds.

Although the desired reverberation time for amplified speech is believed to be less than 1 second,[4] it was found, by means of the techniques described in this paper, that a high level of intelligibility can be achieved, even when the reverberation time at 500 cps is as high as 2 to 4 seconds.

IV. INTELLIGIBILITY TESTS

Two different methods were used for measuring the speech intelligibility. One of these was the familiar word-articulation test in which one or more talkers and a number of listeners participated. The test words were monosyllabic and were used in phonetically balanced groups of fifty. Each word was read in one of the following carrier sentences: "You will write ______"; "Please write down ______"; "This time put ______"; and "I will read ______."

The other method of testing intelligibility was devised especially for these studies. It was performed at a public gathering while the scheduled speaker was talk-

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[3] V. O. Knudsen and C. M. Harris, "Acoustical Designing in Architecture," John Wiley and Sons, Inc., New York, N. Y., chap. 8; 1950.

[4] R. H. Bolt and A. D. MacDonald, "Theory of speech masking by reverberation," *Jour. Acous. Soc. Amer.*, vol. 21, pp. 577–580; November, 1949.

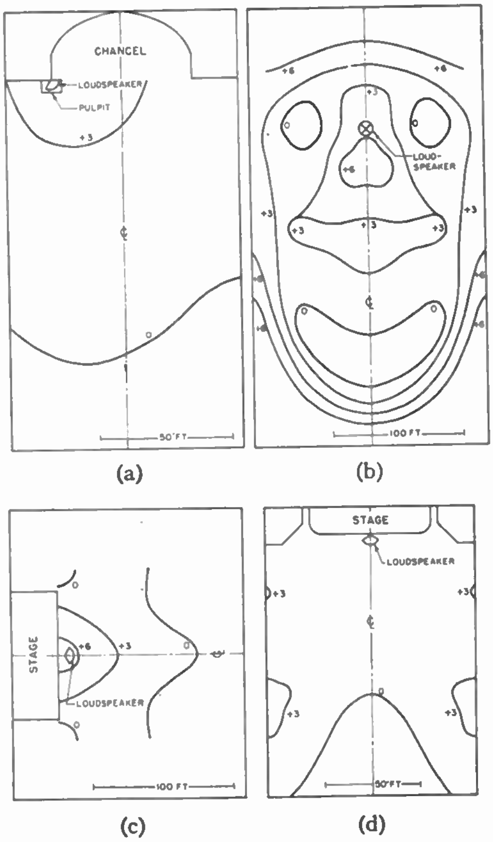


Fig. 4—Contours of constant sound pressure level at 500 cps for: (a) cathedral, (b) Boston Garden, (c) Rockwell Cage, and (d) memorial auditorium.

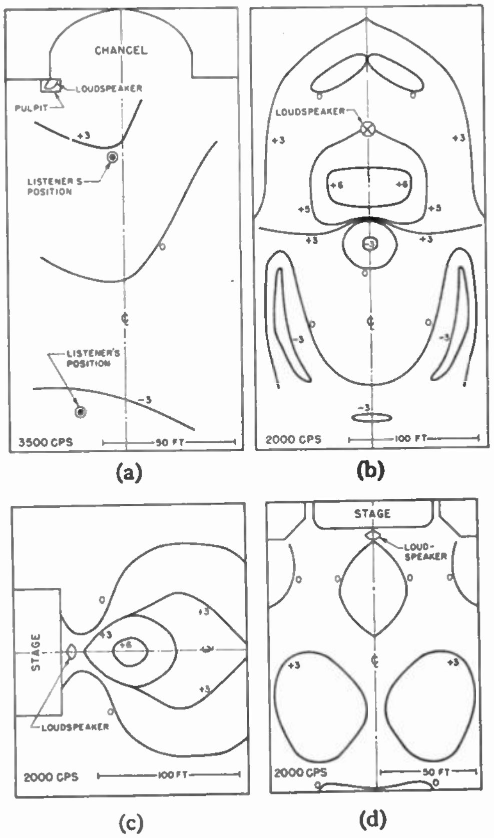
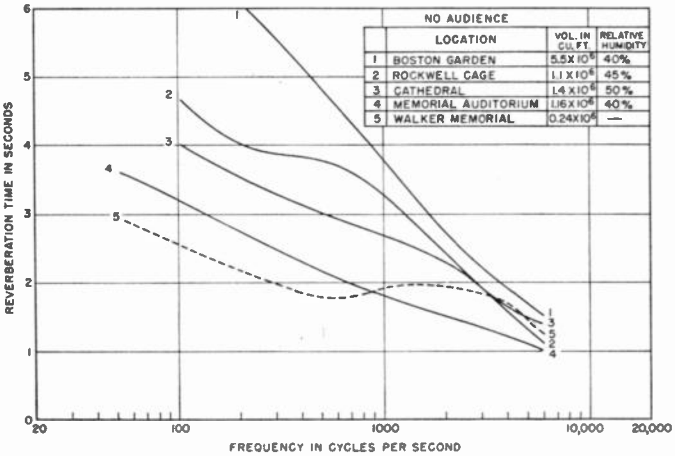
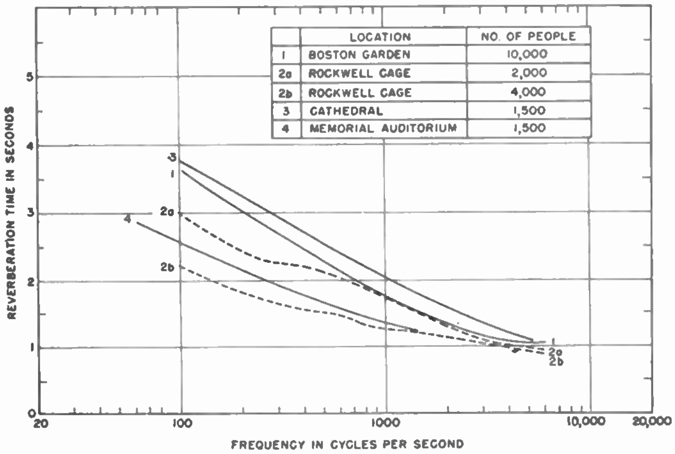


Fig. 5—Contours of constant sound pressure level at frequencies indicated for: (a) cathedral, (b) Boston Garden, (c) Rockwell Cage, and (d) memorial auditorium.



NO AUDIENCE			
	LOCATION	VOL. IN CU. FT.	RELATIVE HUMIDITY
1	BOSTON GARDEN	5.5X10^6	40%
2	ROCKWELL CAGE	1.1X10^6	45%
3	CATHEDRAL	1.4X10^6	50%
4	MEMORIAL AUDITORIUM	1.16X10^6	40%
5	WALKER MEMORIAL	0.24X10^6	—

Fig. 6—Measured reverberation times for empty auditoriums. The volumes and relative humidities are shown in the table.



	LOCATION	NO. OF PEOPLE
1	BOSTON GARDEN	10,000
2a	ROCKWELL CAGE	2,000
2b	ROCKWELL CAGE	4,000
3	CATHEDRAL	1,500
4	MEMORIAL AUDITORIUM	1,500

Fig. 7—Reverberation times calculated (except for the cathedral where the times were measured) assuming the occupancy given in the table.

ing. Certain members of the audience were asked to write down the terminal word of every sentence uttered by the speaker. Simultaneously, a recording of the speaker's voice was made with a disk or tape recorder connected to the microphone circuit. The word lists so obtained were graded with the aid of a master sheet prepared from the recording.

In the cathedral, word-articulation tests were performed on two occasions with no audience but with different observers. One group consisted of two talkers and five listeners, and the other of one talker and twenty listeners. The smaller group was composed of male college graduates. The larger group was composed of men and women whose educational backgrounds ranged from grammar school to college. For the smaller group, four locations were selected for the listeners: two on the main floor (see Fig. 5(a)) and two in the two galleries at the rear of the auditorium. Each listener sat in four different locations during the test, and was at each location once for each of the two talkers. In all, eight lists of fifty words each were read. For the larger crew, eight locations in the cathedral were selected, one for each of the eight word lists. The results of the tests, grouped so as to apply to four separated locations in the cathedral, are shown in Fig. 8. The average percentage of words correctly recorded by the small crew was 63, while that of the large crew was 41. However, the differences between the average score and the score at each of the four positions was very nearly the same for both crews. This illustrates an important consideration in articulation testing; namely, only comparisons among scores obtained by the same crew on the same day should be made because the composition of the crew, experience, and other factors, greatly influence the absolute value of the score.

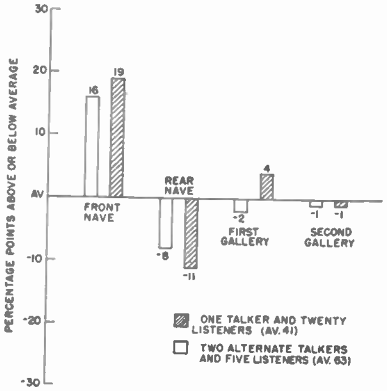


Fig. 8—Articulation scores obtained in the cathedral for each of two groups of observers plotted relative to the average score for the group.

In the front part of the nave, where the reverberation characteristics of the room are of least importance because of the high ratio of direct-to-reflected sound energy, the scores were 79 per cent for the small crew. Undoubtedly this score would have been higher if the sound system had had a greater frequency range.

From the frequency-response curve for the cathedral shown in Fig. 2, it can be seen that the response was uniform within ±3 db over only 65 per cent of the frequency range which is essential to good speech intelligibility. A restricted frequency range lowers the intelligibility of speech, and the score of 79 per cent observed is reasonable for the range used, as we shall see later. If the frequency response had been increased, the word intelligibility for both crews would have improved substantially, assuming low background noise level. A method for calculating the improvement is given in Section VI.

The lower scores at the rear of the nave and in the galleries are due, in part, to the reduced intensity of the sound (see Figs. 4 and 5) and, in part, to the reverberant conditions. The average sound levels were nearly the same in the galleries as at the rear of the nave. The poor acoustic conditions (reverberation and echoes) at the rear of the nave greatly lowered the scores. In this cathedral, the reverberation time was nearly independent of the number of people in the audience because of the high absorption of the seat cushions and the lower relative humidity when the cathedral was empty.

The results obtained from the terminal-word test described above are shown in the middle column of Table I. The word-articulation scores for comparable positions with the large group are also shown. The scores from the terminal-word test were expected to be higher than those from the word-articulation test because the listeners were aided by sentence context. It was anticipated that the scores would approach what is commonly called the "sentence intelligibility."

TABLE I

Comparison of Terminal-Word Scores Determined with Full Audience and Word-Articulation Scores Determined without Audience—Cathedral

Location	Terminal-Word Score	Word-Articulation Score
Front of nave	80	60
Center of nave	80	55
Rear of nave	69	35
Galleries	66	39

The data in Table I are plotted as crosses in Fig. 9 along with a curve of sentence intelligibility versus word articulation as determined by Egan from extensive experiments conducted at Harvard University.[8] These results show that the terminal-word test is approximately a measure of sentence intelligibility.

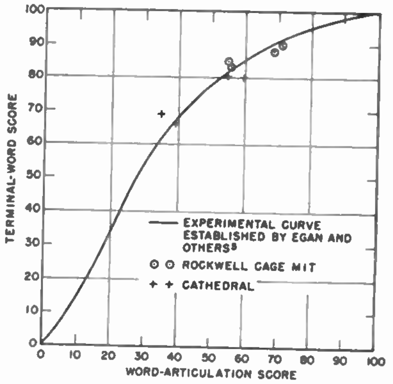


Fig. 9—Relation between terminal-word scores and word-articulation scores. An experimental curve of sentence intelligibility versus word-articulation scores is shown for comparison.

To determine the effect of restricting the bass response of a speech-amplifying system, tests were performed in the Rockwell Cage at M.I.T. using the loudspeaker arrangement of Fig. 1, and a cardioid microphone. From casual observation we decided that with the room empty, a flat frequency response down to very low frequencies produced undesirable reverbera-

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[8] J. P. Egan; see L. L. Beranek, "Acoustic Measurements," John Wiley and Sons, Inc., New York, N. Y., p. 628; 1949.

tion effects. Our first test was performed with only the moderate amount of bass shown by curve *B* (Fig. 10).

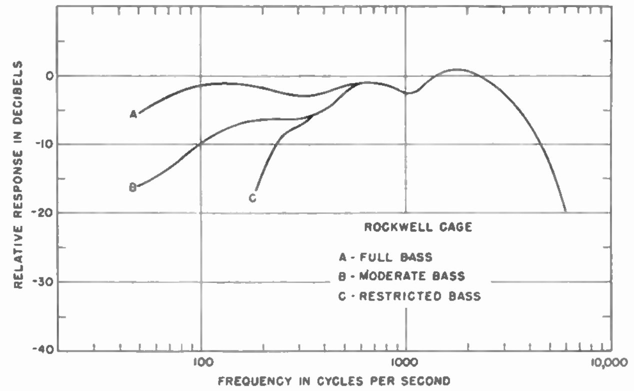

Fig. 10—Frequency-response characteristics of the Rockwell Cage sound-reinforcing system used during subjective tests.

Word articulation was measured with twenty-five listeners present for the two-system conditions shown by response curves *B* and *C*, Fig. 10. There were two talkers, each reading four hundred words, that is, two hundred words for each system. The twenty-five listeners were divided into four groups of about six persons each for the reading of a list of one hundred words. Then they were divided differently by a random-selection process to form another set of four groups for the reading of another list of one hundred words. By the time eight of the one hundred-word lists had been recorded, each person had been to each of four positions twice. For each of the one hundred-word lists, the two conditions of the system and the two voices had been employed. Results are shown in Table II.

TABLE II

Comparison of Sound System with Moderate and Restricted Bass (See Fig. 11)—Rockwell Cage

System	Condition	Word-Articulation Score
C	Restricted bass	62.8 per cent
B	Moderate bass	61.5 per cent

We see that the system performance with restricted bass is slightly superior to that with moderate bass. The difference is so small that either system is equally satisfactory. The low average scores of about 62 per cent were not surprising because of external traffic noise, the somewhat restricted high-frequency response of the system, the reverberant room conditions, and the variation of level in the room (see Figs. 4 and 5). There were only negligible differences among scores obtained with the two talkers.

In the second test, terminal-word articulation was measured with a full audience present, for the two system conditions shown by response curves *A* and *C* of Fig. 10. Thirty-nine observers recorded the last word of each sentence for 8 minutes without bass, then for 16 minutes with bass, and finally for 8 minutes without bass. Results are shown in Table III.

TABLE III

Comparison of Sound System with Full and Restricted Bass (Fig. 11)—Rockwell Cage

System	Condition	Terminal-Word Articulation Score
C	Restricted bass	86.6 per cent
A	Full bass	85.7 per cent

Our conclusion is the same as before, namely, that a slight advantage results from restricting the bass, but the difference between the performances of the two systems is negligibly small.

A further comparison of terminal-word intelligibility with sentence intelligibility is possible from this experiment. Data from four separated positions in the Rockwell Cage are shown in Table IV and are plotted as circles in Fig. 9. As was the case for the cathedral tests, the points fall near the Egan curve.

TABLE IV

Comparison of Terminal-Word Scores Determined with Full Audience, and Word-Articulation Scores Determined with Twenty-Five Listeners—Rockwell Cage

Location	Terminal-Word Score	Word-Articulation Score
Near center of room	90	71
Center edge	88	69
Rear corner	83	56
Center rear	85	55

Word-intelligibility tests were also performed in the industrial building. The results will be discussed in Section VI.

V. Preference, Naturalness, and Intelligibility Ratings

Manufacturers of sound equipment frequently state that it is true that only the frequencies from about 300 to 5,000 cps are needed for good speech intelligibility, but if naturalness is desired, the lower frequencies must also be reproduced. To test the validity of this statement, members of the audience were asked to make subjective judgments of the naturalness and intelligibility of speech in each auditorium studied. In the Boston Garden, where Winston Churchill and Harold E. Stassen spoke before near-capacity audiences, only enthusiastic appraisals of the system[6] were received. Favorable comments are still being received on the systems in the cathedral and in the memorial auditorium, despite the restricted bass, after two years of continuous use.

In general, qualitative comments are not wholly significant because the audience does not have specific

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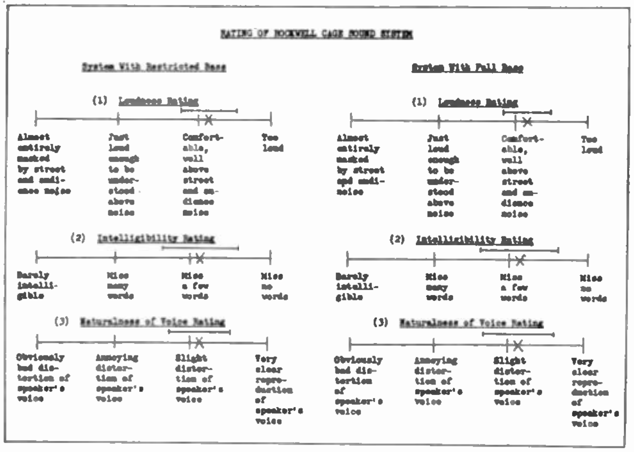
[6] This system was installed for the M.I.T. Mid-Century Convocation, and differs greatly from the one ordinarily used in the Boston Garden.

comparisons to make. To obtain more quantitative results, questionnaires were issued on each of three occasions in the Rockwell Cage at M.I.T. First, when three prominent Americans spoke; second, when a noted evangelist spoke; and third, when the word-articulation tests previously mentioned were being performed.

On the first occasion, with the bass condition of curve *C* of Fig. 10, twenty-seven observers judged the intelligibility, naturalness, and loudness on a rating scale of *Excellent*, *Good*, *Fair*, and *Poor*. The averages of their ratings were: Intelligibility—*Good to Excellent;* naturalness—*Good;* and loudness—*Good.* When the evangelist spoke, the system was operated part of the time with restricted bass, as shown by curve *C* of Fig. 10, and part of the time with full bass, as shown by curve *A* of Fig. 10. The three rating scales on which the 39 observers indicated their opinions, are reproduced in Fig. 11. The average results are indicated on each scale. These results show no discernible preference on the part of the listeners for one system over the other. On the third occasion, thirty observers, were asked to indicate a preference when the bass was switched in and out. Of these, thirty preferred the system with moderate bass (curve *B* of Fig. 10), and eighteen preferred it with no bass. On this occasion, the full bass condition was not compared.



RATING OF ROCKWELL CAGE SOUND SYSTEM

System With Restricted Bass

(1) Loudness Rating

Almost entirely masked by street and audience noise
Just loud enough to be understood above noise
Comfortable, well above street and audience noise
Too loud

(2) Intelligibility Rating

Barely intelligible
Miss many words
Miss a few words
Miss no words

(3) Naturalness of Voice Rating

Obviously bad distortion of speaker's voice
Annoying distortion of speaker's voice
Slight distortion of speaker's voice
Very clear reproduction of speaker's voice

System With Full Bass

(1) Loudness Rating

Almost entirely masked by street and audience noise
Just loud enough to be understood above noise
Comfortable, well above street and audience noise
Too loud

(2) Intelligibility Rating

Barely intelligible
Miss many words
Miss a few words
Miss no words

(3) Naturalness of Voice Rating

Obviously bad distortion of speaker's voice
Annoying distortion of speaker's voice
Slight distortion of speaker's voice
Very clear reproduction of speaker's voice

Fig. 11—Rating sheet used in Rockwell Cage test when a noted evangelist spoke. The crosses on each of the rating scales show the average of the ratings for 39 observers. The bars above the crosses are equal in length to twice the standard deviations.

VI. Calculation of Speech Intelligibility

A simple graphical method for calculating speech intelligibility was used to predict the articulation scores measured in the industrial building. By this method the articulation index area of Fig. 12 is combined with the over-all response curve of the sound system to produce the shaded area of Fig. 13,[1] bounded by the curves "amplified speech peaks" and "amplified speech minimums." This area was located vertically on the coordinate system by measuring the average sound pressure level in each of eight octave bands at several positions, while the talker was reading a word list. These eight octave band levels were then converted to spectrum levels (sound pressure level in a band 1 cps wide). Finally, 10 db were added to these spectrum levels to give a plot of the speech peaks as a function of frequency, because the average, as read on a vu meter, lies about 10 db below the speech peak level.[7] This plot, having only eight points, was used to locate vertically the more accurate plot of Fig. 13 by making it and the upper edge of the shaded region coincide as closely as possible.

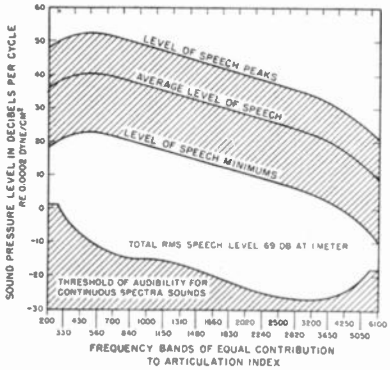


Fig. 12—Graph relating speech levels to frequency. The upper shaded area between the curves "level of speech peaks" and "level of speech minimums" is, by definition, the area corresponding to 100 per cent "articulation index." The speech levels were measured with a microphone placed 1 meter in front of a man speaking in a raised voice in an anechoic chamber.

The background noise in the room during the tests was also measured by the eight-band analyzer. After converting these sound levels to spectrum levels, they were plotted on the same graph (Fig. 13). The proportion of the shaded area lying above the background noise (or the threshold of hearing, whichever is greater) is known as the articulation index. For the example shown in Fig. 13, the articulation index is 0.51. This calculation was made for one particular position in the building, for one particular gain-control setting. The sound level varied by about 10 db over the floor area. Calculations were performed for three other positions, and the results are shown in Table V.

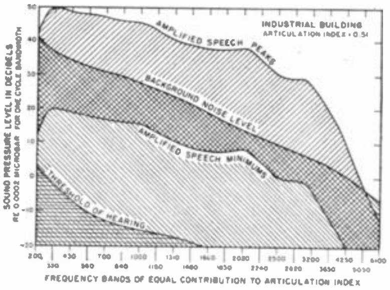


Fig. 13—Graph for determining an articulation index in the industrial building. The fraction of the articulation index area that lies above the "background noise level" is the articulation index, here equal to 0.51.

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[7] L. L. Beranek, "Acoustic Measurements," John Wiley and Sons, Inc., New York, N. Y., pp. 702–703; 1949.

TABLE V

Comparison of Calculated Articulation Indexes with Measured Articulation Scores—Industrial Building

Position	Calculated Articulation Index	Word-Articulation Score
A	83	90
B	71	76
C	62	65
D	51	59

The articulation index is approximately a measure of the percentage of the syllables correctly heard by a listener.[1] The word-articulation score is always higher; the exact relation depends upon the education of the listeners and their previous practice. The relations of Table V are typical of results obtained for college graduates without previous experience at this kind of test. A relation for college graduates with months of daily experience at this kind of test is shown in curve (a), Fig. 11 of the literature 1. Lower scores will be obtained for personnel whose reading habits are less well cultivated.

Bolt and MacDonald[4] have shown that the reverberation in a room, caused by the speech itself, is similar to background noise in reducing speech intelligibility. For each of the twenty bands of equal importance to speech intelligibility shown on the abscissas of Figs. 12 and 13, the number of decibels that the reverberant "speech" lies below the peaks of the direct speech is related to reverberation time, as shown in Table VI. For a multicellular horn with a mouth opening of 22 by 30 inches, the directivity index is approximately 8 db at 200 cps, 11 db at 300 cps, and 13 db for all frequencies above 500 cps. The reverberant "speech" level should be plotted as though it were background noise, and the higher of the two curves (room background noise or reverberant "speech" level) used as the background noise in determining the articulation index. It is seen that if the directivity index is high, speech intelligibility is affected little by reverberation. This agrees with the subjective impression of listeners. However, in a reverberant room, the background noise level will be higher than that in a nonreverberant room. Reduction of the reverberation time results in both a reduction of background noise and of the reverberant "speech," both of which increase the articulation index.

TABLE VI

Relation between Reverberation Time in an Auditorium and the Difference in Decibels between the Direct Speech Peak Levels and the Reverberant "Speech" Levels. The Parameter Is Directivity Index of the Loudspeaker[8]

Reverberation Time Seconds	Direct Speech Peak Level Minus Reverberant "Speech" Level—Decibels				
	Loudspeaker Directivity Index—Decibels				
	0	5	10	15	20
0.5	25	30	35	40	45
1	20	25	30	35	40
2	15	20	25	30	35
3	13	18	23	28	33
4	11	16	21	26	31

VII. Other Observations

An important reason for the satisfactory performance of these sound systems, even in rooms where the reverberation times at 500 cps are between two and four seconds, is the location of the loudspeakers. It was found that the best results were obtained when the loudspeaker was between 20 and 30 feet above the head of the talker, and the principal axis of the horn was pointed downward so as to keep as much of the sound off the side walls and ceiling as possible. Also, a more uniform distribution of sound over the seating area is obtained when the loudspeaker is high.

Experiments in Morss Hall at M.I.T., and at various other installations in Boston, showed that a single loudspeaker over the podium is superior to one or more loudspeakers on either side of the stage. With loudspeakers on opposite sides of the stage, distracting aural effects occur along and adjacent to the center line of the hall. With one loudspeaker, even though mounted high above the podium, the amplified speech appears to come from the talker's mouth because of the poor vertical directivity of the human ear.

The fact that so little difference in preference was found between systems with restricted bass and those with full bass, resulted partly from the satisfactory high-frequency response of all systems tested. If the high-frequency response is restricted, and if full bass is present, the quality of the reproduced speech is poor, especially in highly reverberant spaces. Situations such as this are common in the United States, for example, in large railway stations.

Conclusions

(1) The results reported herein are consistent with the prediction from articulation theory, that the overall frequency response of a speech-reinforcing system should extend at least from 400 to 4,000 cps, and lie within about ±3 db of the average value of the response through that region.

(2) The distribution of sound in an auditorium should be sufficiently uniform so that the requirements of the first conclusion are satisfied at all seat locations.

(3) Our results indicate that, contrary to general belief, the naturalness of the reinforced speech is not affected if frequencies below approximately 400 cps are attenuated. This conclusion is based on experiments where the response above 400 cps meets the requirements under the first conclusion above.

(4) The results show that the reinforcement of the low voice frequencies in highly reverberant auditoriums does not reduce the intelligibility of the amplified speech. Little, if any, harm results from the presence of

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[8] L. L. Beranek, "Acoustic Measurements," John Wiley and Sons, Inc., New York, N. Y., pp. 668–684; 1949.

the full bass range of frequencies, provided the response at the high end is good out to at least 4,000 cps.

(5) The preceding two conclusions might no longer be valid if the loudspeaker location were to differ from those used in these experiments and if the loudspeaker had poor directivity. The locations used here were selected after observations of many existing systems, which indicated that the loudspeakers should be hung high above the podium (20 to 30 feet) and be directed downward, so that the sound is absorbed by the people and upholstered seats, and so that the sidewall reflections are minimized. A single loudspeaker system so located yields more natural reproduction than a split loudspeaker system because (a) the sound appears to come from the talker and because (b) no regions of the seating exist at which the sounds from two loudspeakers overlap, with resulting disturbing psychological effects.

(6) The terminal-word test described herein for measuring the sentence intelligibility of a sound system under normal operating conditions yields results which are consistent with results from conventional word-articulation tests, which can rarely be carried out under normal operating conditions.

(7) Previously reported theories for the calculation of speech intelligibility predict word-articulation scores within acceptable tolerances.

Radiation Resistance of a Two-Wire Line*

JAMES E. STORER† AND RONOLD KING‡

Summary—**A general formula for the radiation resistance of a two-wire line is derived by means of a Poynting vector integration over a large sphere. The result is shown to be in agreement with that computed by other techniques. Formulas for the useful special cases of a lossless system and a nonresonant line are presented.**

I. Introduction

THE PROBLEM of the power radiated from a two-wire line is one of considerable interest. Although numerous papers[1-5] have appeared on the subject, none, to the authors' knowledge, has contained a general formula for the radiation resistance. It is the purpose of this paper to derive such a formula and compare it with previous work.

There are three techniques available for calculating the radiation resistance. One is the integration of the normal component of the Poynting vector over a large sphere surrounding the line. A second, usually called the emf method, is equivalent to the integration of the normal component of the Poynting vector over the surface of the wires. The third and last possibility is to obtain the radiation resistance directly from the integral equation for the distribution of current along the line. It will be shown that these methods all yield consistent results.

II. The Distribution of Current Along the Line

Before it is possible to obtain a formula for the radiation resistance, it is necessary to know the distribution of current along the line. This has been done in standard texts on the subject.[6] For purposes of reference, however, it is worth while to write down the integral equation for the current and its solution. Hallén's equation[7] for such a two-wire-line circuit, as indicated in Fig. 1, is

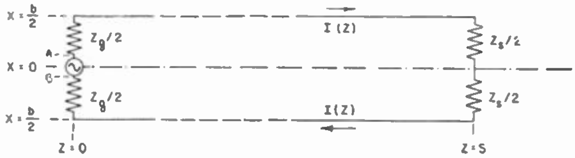

Fig. 1—Two-wire line circuit.

$$Z\delta(s) = z^i I(s) + \frac{j\zeta_0}{4\pi}\oint_{s'} I(s')\left[\beta_0{}^2\hat{s}\cdot\hat{s}' - (\hat{s}\cdot\nabla)(\hat{s}'\cdot\nabla')\right]\frac{e^{j\beta_0 R_{ss'}}}{\beta_0 R_{ss'}}\,ds', \qquad (1)$$

where

z^i = the internal impedance per unit length

Z = the input impedance across the generator terminals AB

$\delta(s)$ = the Dirac delta function defined by $\int_{-x}^{x}\delta(x)dx = 1,\ x>0$

$I(s)$ = the current in the line at a point s^{-z}

ζ_0 = the impedance of free space ($=120\pi$ ohms)

$\beta_0 = 2\pi/\lambda$ where λ is the free-space wavelength

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[4] C. W. Harrison, "On the pickup of balanced four-wire line," Proc. I.R.E., vol. 30, pp. 517–518; November, 1942.

[5] E. J. Sterba and C. B. Feldman, "Transmission lines," Proc. I.R.E., vol. 20, pp. 1163–1202; July, 1932.

[6] R. W. P. King, "Electromagnetic Engineering," vol. 1, chap. VI, McGraw-Hill Book Co., New York, N. Y.; 1945.

[7] J. Aharoni, "Antennae," chap. II, Oxford University Press, Oxford, England; 1946.

$\hat{s}$ = a unit vector tangent to the wires at a point s, and in the direction of the current

$R_{ss}' = \{[\text{distance between } s \text{ and } s']^2 + a^2\}^{1/2}$

a = the radius of the wires, satisfying the inequality $\beta_0^2a^2 \ll \beta_0^2b^2 \ll 1$

$\oint_s' ds'$ indicates an integration around the contour of the circuit.

It will be assumed, although not explicitly written, that all electromagnetic quantities have a time dependence of the form $e^{j\omega t}$.

Subject to the five conditions

1) $\beta_0^2b^2 \ll 1$,
2) $\beta_0^2a^2 \ll \beta_0^2b^2$,
3) $\alpha^2/\beta_0^2 \ll 1$, where α is the attenuation constant of the line,
4) The line is balanced, i.e., the currents in the two wires at the same position on the line are equal but opposite in direction, and
5) The current across the terminations is a constant, i.e., the terminations occupy regions of space that are small in comparison with the wavelength,

the solution of (1) can be written as a power series expansion in the parameter $\beta_0^2b^2$ as follows:

$$I(z) = I_0(z) + \beta_0^2b^2I_1(z) + \beta_0^4b^4I_2(z) + \cdots. \quad (2a)$$

King[8] has demonstrated that $I_0(z)$ can be written in the form

$$I_0(z) = I_0(0)\frac{\cosh[(\alpha + j\beta)(s - z) + \rho + j\Phi']}{\cosh[(\alpha + j\beta)s + \rho + j\Phi']}. \quad (2b)$$

The quantities ρ and ϕ' are defined in terms of the characteristic impedance Z_c, the generator impedance Z_g, and the terminal impedance Z_s as follows:

$$\rho + j\Phi' = \tanh^{-1} Z_s/Z_c.$$

The input impedance Z is found to be

$$Z \doteq Z_g + Z_c \tanh[(\alpha + j\beta)s + \rho + j\Phi'] + \text{terms of order } \beta_0^2b^2. \quad (3)$$

The $\beta_0^2b^2$-term in the current-distribution expansion is impossible to evaluate unless we assume a specific wire configuration for the terminations. Even in the simplest case of wire bridge terminations for the line, this $\beta_0^2b^2$ term appears to be so complicated that it would have to be evaluated numerically.

III. Calculation of the Radiation Resistance by the Integration of the Poynting Vector Over a Large Sphere

The integration of the normal component of the Poynting vector over the surface of a large sphere surrounding an antenna system is the usual technique to determine the power radiated from such a system.[9] Using the standard formula for the Poynting vector, and taking the limit as the spherical surface recedes to infinity, the following equation is obtained for the radiation resistance R:

Average power radiated

$$= \frac{1}{2}RI(0)I^*(0)$$

$$= \frac{1}{2}\frac{\zeta_0}{4^2\pi^2}\beta_0^2\int_0^{\pi}\int_0^{2\pi}\sin\theta d\theta d\phi\left[\vec{\eta}\cdot\vec{\eta}^* - (\hat{r}\cdot\vec{\eta})(\hat{r}\cdot\vec{\eta}^*)\right] \quad (4)$$

$$\vec{\eta} = \int_v \vec{J}(x', y', z')e^{-j\beta_0(\hat{r}\cdot\vec{r}')}dx'dy'dz'$$

where

θ, ϕ = spherical co-ordinates describing points on the spherical surface

$\hat{r}$ = a unit vector directed outward from the surface

$*$ indicates a complex conjugate

$\vec{r} = x'\hat{x} + y'\hat{y} + z'\hat{z}$

$J(x', y', z')$ = the current density at a point x', y', z'.

If the previously given current distribution (2) is inserted, the integral for $\vec{\eta}$ may be evaluated. It consists of three parts, a contribution from the line, and one from each termination. If only the first-order terms β_0b are kept, the integral for $\vec{\eta}$ becomes independent of the physical structure of the terminations (wire bridge, coil, and the like). The contributions from the terminations are equivalent to that of a single current filament across the terminations. Inserting this expression for $\vec{\eta}$ into (4), the radiation resistance may be evaluated. The details of this process are left to the Appendix. The final result is

$$R = \frac{\zeta_0}{4\pi}\beta_0^2b^2\frac{\cosh(\alpha s + 2\rho)}{|\cosh[(\alpha + j\beta)s + \rho + j\Phi')]|^2} \cdot \left[\cosh\alpha s - \frac{\sin 2\beta_0 s}{2\beta_0 s}\right] + \text{terms of order } \beta_0^3b^3. \quad (5)$$

It is apparent from the technique used to obtain (5) that the $\beta_0^3b^3$ terms would be difficult, if not impossible, to evaluate in a general way. It is worth while to see how formula (5) simplifies for certain special cases.

Case I: Line Approaching the Nonresonant State, i.e., $\rho > 3$.

$$R = \frac{\zeta_0}{2\pi}\beta_0^2b^2\left[\cosh\alpha s - \frac{\sin 2\beta_0 s}{2\beta_0 s}\right]e^{-\alpha s} \quad (\pm 0.5 \text{ per cent}).$$

Then for a long line, $B_0s > 50$, and good conductivity (α small)

$$R = \frac{\zeta_0}{2\pi}\beta_0^2b^2 = 60\beta_0^2b^2 \text{ ohms} \quad (\pm 0.5 \text{ per cent}). \quad (6a)$$

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[8] R. W. P. King, "Transmission-line theory and application," *Jour. Appl. Phys.*, vol. 14, pp. 577–600; November, 1943.

[9] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., New York, N. Y.; 1941.

Case II: Small Losses in Termination and Line, i.e., $(\alpha s+2\rho)<0.1$.

There

$$R = \frac{\zeta_0}{4\pi}\beta_0{}^2b^2\frac{1}{\cos^2(\beta_0 s+\Phi')}\left[1-\frac{\sin 2\beta_0 s}{2\beta_0 s}\right]$$

(± 0.5 per cent).

This resistance can be referred to the maximum current on the line, i.e., power radiated $=\frac{1}{2}R_{\max}|I_{\max}|^2$; then

$$R_{\max} = \frac{\zeta_0}{4\pi}\beta_0{}^2b^2\left[1-\frac{\sin 2\beta_0 s}{2\beta_0 s}\right] \qquad \text{(See Fig. 2).}$$

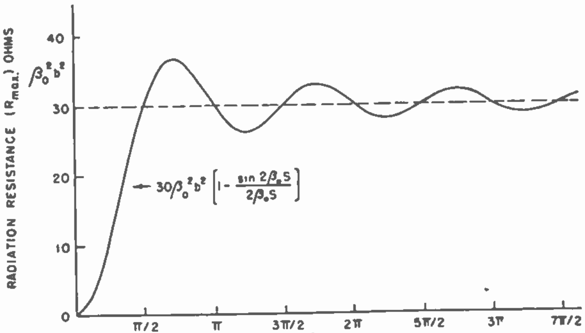

Fig. 2—Radiation resistance of a lossless line.

It is to be noted that this resistance is independent of the terminations and reasonably constant. In the case of wire bridge terminations ($\phi'\cong 0$) and a resonant line ($\beta_0 s=n\pi$), these formulas reduce to

$$R = R_{\max} = \frac{\zeta_0}{4\pi}\beta_0{}^2b^2 = 30\beta_0{}^2b^2 \text{ ohms.} \tag{6b}$$

The formulas for these last special cases are in agreement with those of previous authors. The difference between the above nonresonant-line expression and that of Sterba and Feldman[5] is attributable to the fact that radiation from the terminations has not been included in their work.

Integration of the normal component of the Poynting vector over a large sphere has the advantage over other methods of obtaining the power radiated from a two-wire line in that the radiation resistance may be evaluated without reference to the physical construction of the terminations. It is interesting to see, however, that other techniques lead to the same results in a more restricted fashion.

IV. EVALUATION OF THE RADIATION RESISTANCE BY OTHER TECHNIQUES

In the Poynting vector method of evaluating the radiation resistance, it is necessary to know the distribution of current. This being the case, it is apparent that the fundamental method of getting at the radiation resistance is to obtain it directly from the integral equation (1) defining the current distribution. This was attempted by King.[10]

The method used resembles the procedure familiar in the approximate solution of the integral equation for the current in an antenna. The principle involved is to substitute for the current under the sign of integration a zero^th^-order approximation of the current. In the case of the transmission line, this zero^th^-order current is the well-known solution of the conventional transmission-line equations which neglect radiation. For a low-loss resonant line, the distribution is very nearly sinusoidal. In the case of the antenna, the zero^th^-order current also is sinusoidal, but this is a very much poorer approximation of the actual current than is true for the transmission line. It might be supposed, therefore, that the evaluation of the radiation resistance of the transmission line should be correspondingly much more accurate than the corresponding calculation of the impedance of the antenna. But this is not necessarily true. If the radiation resistance of a resonant line were of the same order of magnitude as the radiation resistance of a resonant antenna, the more accurate zero^th^-order current available for the transmission line certainly should lead to a correspondingly more accurate radiation resistance. However, the radiation resistance of a resonant line is of the order of magnitude of 30 $\beta_0{}^2b^2$ ohms, and hence is a very small fraction of an ohm for $\beta_0 b\ll 1$. For an antenna, the radiation resistance at resonance is of the order of magnitude of 70 ohms. Clearly, in order to determine accurately the very small value of radiation resistance for a transmission line, a very much more accurate zero^th^-order distribution of current is necessary for the line than for the antenna. Actually, a distribution is required that is accurate to terms in $\beta_0{}^2b^2$. This can not be true of the conventional transmission-line distribution, since it is determined by ignoring radiation resistance which is known to be as large as 30 $\beta_0{}^2b^2$. Nevertheless, in the case of a resonant line an even number of half-wavelengths long, King arrived at essentially the correct result. But for other lengths, in particular, for a line an odd number of half-wavelengths long, an incorrect result was obtained as should, indeed, be expected.

This sensitivity to current distribution can be circumvented by the use of a variational principle. Then the eigen value so obtained, in this case Z, is correct to one order higher than the distribution function used to compute it. Hence, when the approximate current-distribution function (2b) is used, Z is correct to one higher order of $B_0{}^2b^2$ than (2b). Thus, Z so computed is correct to $B_0{}^2b^2$.

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[10] R. W. P. King, "Electromagnetic Theory," Chap. VI, Sec. 25, pp. 483–485, eq. 1–12, vol. I, McGraw-Hill Book Co., New York, N. Y.; 1945.

The original equation (1), can be written as follows:

$$Z\delta(s) = \frac{1}{I(0)} \oint_{s'} I(s')G(s, s')ds', \tag{7}$$

where

$$G(s, s') = z^i\delta(s - s') + \frac{j\zeta_0}{4\pi} \left[\beta_0{}^2\hat{s}\cdot\hat{s}' - (\hat{s}\cdot\nabla)(\hat{s}'\cdot\nabla')\right] \frac{e^{-j\beta_0 R_{ss'}}}{\beta_0 R_{ss'}} = G(s', s).$$

The variational expression for Z is

$$Z = \frac{1}{I^2(0)} \oint_s \oint_{s'} I(s)I(s')G(s, s')ds'ds.$$

This can be proved readily by performing the variation, i.e.,

$$\delta Z = \frac{1}{I^2(0)} \oint_s \oint_{s'} [I(s)\delta I(s') + I(s')\delta I(s)]G(s, s')ds'ds - 2\frac{1}{I^3(0)}\,\delta I(0) \oint_s \oint_{s'} I(s)I(s')G(s, s')dsds'.$$

Using the original equation (7), this becomes

$$\delta Z = \frac{1}{I(0)} \oint_{s'} \delta I(s')Z\delta(s')ds' + \frac{1}{I(0)} \oint_s \delta I(s)Z\delta(s)ds - 2\frac{\delta I(0)}{I(0)}Z = \frac{\delta I(0)}{I(0)}Z + \frac{\delta I(0)}{I(0)}Z - 2\frac{\delta I(0)}{I(0)}Z = 0.$$

A technique is accordingly made available for determining Z correct to the order $\beta_0{}^2b^2$. Assuming this, the next problem is to separate out the radiation resistance from the input impedance Z. When there are no losses in the line or terminations, this is simple, because then the real part of the input impedance must be the radiation resistance, since there are no other power losses in the system. In the case of no losses, $z^i = 0$ and $I(s)$ is real. Hence,

$$R = \text{real part}\,\frac{1}{I^2(0)} \oint_s \oint_{s'} I(s)I(s')G(s, s')dsds' = \frac{\zeta_0}{4\pi} \oint_s \oint_{s'} I(s)I(s')\left[\beta_0{}^2\hat{s}\cdot\hat{s}' - (\hat{s}\cdot\nabla)(\hat{s}'\cdot\nabla')\right] \frac{\sin \beta_0 R_{ss'}}{\beta_0 R_{ss'}}\, dsds'.$$

After a rather tedious integration, it can be shown[11] that this integral yields a result identical to that obtained previously (Case II). It must be emphasized that this method yields the correct result even if the lossless assumption is not made.

It is also possible to obtain the radiation resistance by integrating the normal component of the Poynting vector over the surface of the wires. This procedure is often referred to as the emf method. It can be shown[11] that this technique also yields results in agreement with those obtained in this paper.

V. RATIO OF OHMIC LOSSES TO RADIATION LOSSES

Antenna systems that consist essentially of a wire close to and parallel to a metal surface, are of practical importance. If the metal surface is sufficiently large compared with the distance $b/2$ of the parallel antenna from it, the conducting surface may be replaced by an image of the wire. The wire, plus its image, is equivalent to a two-wire line; ohmic losses and radiation are one-half those of a two-wire line. It is apparent that the efficiency of such a system is determined by the ratio of ohmic losses to the radiation resistance. Restricting the work to the case of lossless terminations and a near lossless line, the radiation resistance of the two-wire line is

$$R = \frac{\zeta_0}{4\pi}\beta_0{}^2b^2 \frac{1 - \dfrac{\sin 2\beta_0 s}{2\beta_0 s}}{\cos^2 (\beta_0 s + \Phi')};$$

the factor $1 - \sin 2\beta_0 s/2\beta_0 s$ is approximately equal to 1, provided $\beta_0 s$ has any appreciable value (see Fig. 2). As only the order of magnitude is of interest in this case, this factor can be set equal to 1. Therefore

$$R \cong \frac{\dfrac{\zeta_0}{4\pi}\beta_0{}^2b^2}{\cos^2 (\beta_0 s + \Phi')}.$$

As it has been assumed that there are no losses in the terminations, the total ohmic loss for the two-wire line is given by

$$\text{ohmic loss} = \frac{1}{2} R_{\text{ohmic}} |\,I(0)\,|^2 = \frac{1}{2} r_i \int_0^s |\,I(z)\,|^2 dz + \frac{1}{2} r_i \int_s^0 |\,I(z)\,|^2 dz,$$

where r_i is the resistance per unit length of the line wires. Therefore

$$R_{\text{ohmic}} = 2r_i \int_0^s \frac{|\,I(z)\,|^2}{|\,I(0)\,|^2}\, dz = 2r_i \int_0^s \frac{|\cosh [(\alpha + j\beta_0)(s - z) + j\Phi']\,|^2}{|\cosh [(\alpha + j\beta_0)s + j\Phi']\,|^2}\, dz \cong 2r_i \int_0^s \frac{\cos^2 [\beta_0(s - z) + \Phi']}{\cos^2 [\beta_0 s + \Phi']}\, dz = r_i s\, \frac{1 + \dfrac{\cos (\beta_0 s + \Phi') \sin \beta_0 s}{\beta_0 s}}{\cos^2 (\beta_0 s + \Phi')}.$$

For appreciable $\beta_0 s$, the factor

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[11] J. E. Storer, "Radiation Resistance of a Two Wire Line," Technical Report No. 69, Cruft Laboratory, Harvard University; March, 1949.

$$1 + \frac{\cos(\beta_0 s + \Phi') \sin \beta_0 s}{\beta_0 s}$$

is approximately equal to 1. Hence,

$$R_{\text{ohmic}}/R \cong \frac{s r_i}{\zeta_0/4\pi} \beta_0{}^2 b^2.$$

It is interesting to evaluate this for a particular case. Assuming the frequency is 300 mc and the wire is of copper and 1 mm thick, $r_i = 1.4$ ohms/meter, letting $s = 1$ meter. This becomes

$$R_{\text{ohmic}}/R = \frac{1 \times 1.4}{30} \cong \frac{1}{20}.$$

VI. CONCLUSIONS

It has been shown that various methods yield consistent results for the radiation resistance. A general formula was obtained for the radiation resistance, the two most useful special cases of this formula being

$$R = \zeta_0/2\pi(\beta_0{}^2 b^2)$$

for a long, near nonresonant line.

$$R_{\max} = \zeta_0/4\pi(\beta_0{}^2 b^2)\left[1 - \frac{\sin 2\beta_0 s}{2\beta_0 s}\right]$$

for a near lossless system.

APPENDIX

EVALUATION OF THE RADIATION RESISTANCE BY THE POYNTING VECTOR METHOD

Using the previously given current distribution (2) and remembering that the radius of the wires satisfies the inequality $\beta_0{}^2 a^2 <<< 1$, $\vec{\eta}$ can be expressed as follows:

$$\vec{\eta} = \hat{z} \int_0^s I(z) e^{-j\beta_0 \cos\theta z'} \left[e^{-j(\beta_0 b \sin\theta\cos\phi)/2} - e^{j(\beta_0 b \sin\theta\cos\phi)/2}\right] dz'$$

$$+ I(0) \int \hat{s}' e^{-j\beta_0 \hat{r}\cdot\vec{r}'} ds' + I(s) \int \hat{s}' e^{-j\beta_0 \hat{r}\cdot\vec{r}'} ds'.$$

Generator Termination Terminal

If terms only of the order $\beta_0 b$ are kept in the expression for $\vec{\eta}$, all three integrals simplify; the two involving the terminations become independent of the physical structure of the terminations (wire bridge, coil, etc.). Hence the contributions from the terminations to the radiation become equivalent to that of a single current filament across the terminations. Thus,

$$\vec{\eta} = \hat{z}[-j\beta_0 b \sin\theta\cos\phi] \int_0^s e^{-j\beta_0 z \cos\theta} I(z') dz'$$

$$+ I(0)\hat{x} \int_{-b/2}^{b/2} dx' + I(s)\hat{x} e^{-j\beta_0 s \cos\theta} \int_{-b/2}^{b/2} dx'$$

$$+ \text{terms of order } \beta_0{}^2 b^2.$$

It is now seen that terms of order $\beta_0{}^2 b^2$ in the current-distribution function can be dropped, since $\vec{\eta}$ is calculated correctly only to order $\beta_0 b$. The integrations necessary for $\vec{\eta}$ can now be readily performed, as the integrand is simply a product of an exponential and a trigonometric function. Hence

$$\vec{\eta} = bI(0)\left\{J_1(\cos\theta)\hat{x} + \frac{\cos\phi}{\sin\theta}\left[\cos\theta J_1(\cos\theta) - J_2(\cos\theta)\right]\hat{z}\right\},$$

where

$$J_1(\cos\theta) = 1 - \frac{\cosh(\rho + j\Phi')}{\cosh[(\alpha + j\beta)s + \rho + j\Phi']} e^{-j\beta_0 s \cos\theta}$$

$$J_2(\cos\theta) = \frac{\sinh[(\alpha + j\beta)s + \rho + j\Phi']}{\cosh[(\alpha + j\beta)s + \rho + j\Phi']} - \frac{\sinh(\rho + j\Phi')}{\cosh[(\alpha + j\beta)s + \rho + j\Phi']} e^{-j\beta_0 s \cos\theta}.$$

Now, $\hat{r}\cdot\hat{x} = \sin\theta\cos\phi$ and $\hat{r}\cdot\hat{z} = \cos\theta$. After substituting $\vec{\eta}$ into the formula for the radiation resistance, this becomes

$$R = \zeta_0 \frac{\beta_0{}^2 b^2}{4^2\pi^2} \int_0^\pi \int_0^{2\pi} \sin\ d\theta d\phi \left\{J_1 J_1{}^* + \frac{\cos^2\phi}{\sin^2\theta}[\cos\theta J_1 - J_2][\cos\theta J_1{}^* - J_2{}^*]\right.$$

$$- \left[J_1 \sin\theta\cos\phi + \frac{\cos\phi\cos\theta}{\sin\theta}(\cos\theta J_1 - J_2)\right]$$

$$\left.\cdot\left[J_1{}^* \sin\theta\cos\phi + \frac{\cos\phi\cos\theta}{\sin\theta}(\cos\theta J_1{}^* - J_2{}^*)\right]\right\}.$$

As J_1 and J_2 are not functions of ϕ, the ϕ part of this integration is readily performed, and the resulting terms simplify to

$$R = \zeta_0 \frac{\beta_0{}^2 b^2}{4^2\pi} \int_0^\pi \sin\theta d\theta [J_1(\cos\theta)J_1{}^*(\cos\theta) + J_2(\cos\theta)J_2{}^*(\cos\theta)]$$

$$= \zeta_0 \frac{\beta_0{}^2 b^2}{4^2\pi} \int_{-1}^{+1} [J_1(x)J_1{}^*(x) + J_2(x)J_2{}^*(x)] dx.$$

This last integration is just one of exponentials and is readily performed. The somewhat complex array of terms so obtained can be simplified by means of trigonometric identities to:

$$R = \frac{\zeta_0}{4\pi}\beta_0{}^2 b^2 \frac{\cosh(\alpha s + 2\rho)}{|\cosh[(\alpha + j\beta)s + \rho + j\Phi]|^2} \cdot \left[\cosh\alpha s - \frac{\sin 2\beta_0 s}{2\beta_0 s}\right].$$

An Electrostatic-Tube Storage System*

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Summary—**A storage system which will store binary (on or off) pulses has been constructed, and should prove useful in laboratory studies of certain communication problems. The system comprises two storage channels, each utilizing an MIT electrostatic storage tube and switching circuits which route incoming pulses into one channel while stored pulses are being recovered from the other. Pulses are stored in each tube in a square array of discrete spots of charge; each spot may assume one of two possible potentials, corresponding to the two possible states of a binary pulse. The order of occurrence of incoming pulses is preserved during storage, but the time relationship is not; the time relationship of pulses recovered from storage is determined by an independent pulse source under control of the user. Consequently, the system may be used to compress, expand, or delay a group of pulses. The capacity of each storage channel is, at present, 256 pulses. The system operates reliably at all frequencies up to 33 kc when storing incoming pulses and up to 70 kc when supplying stored pulses.**

ONE OF THE basic problems in communication is the obtaining of more efficient utilization of transmission channels, by reducing the necessary bandwidth, reducing the required average power, or improving the signal-to-noise ratio. This problem is most profitably attacked by applying information theory. The results obtained show that storage systems are necessary components of the information-processing equipment at both the sending and receiving ends of the transmission system; and furthermore, that each such storage system must be capable of absorbing alterations of the time scale because the instantaneous rate at which new information enters the system will, in general, be different from that at which stored information is recovered.[1] Other communication applications of storage systems are to obtain delays and to obtain uniform compression or expansion of the time scale of the input.

The storage system described in this paper was designed to serve as a general-purpose laboratory instrument, capable of providing a delay or uniform or arbitrary alterations of the time scale of the input. It stores binary pulses: pulses which have only the two states of on and off and which represent the binary digits of 1 and 0. The fact that the system stores only binary pulses does not limit its utility in communication because information can be represented arbitrarily closely by a sequence of binary numbers.

The characteristics of a storage system are primarily determined by the storage device which is used as the nucleus of the system. In the present system, the MIT electrostatic storage tube was selected as the storage device because it enabled the desired flexibility with respect to time-scale alterations to be obtained without the use of an excessively large quantity of equipment. The same flexibility characteristics, and a much greater operating speed, could have been obtained by using flip-flops (bistable multivibrators) instead of the storage tube. However, the size and cost of the equipment which would have been required more than offset the advantage of a higher operating speed. Other storage devices, such as magnetic drums and ultrasonic-line storage loops,[2] were considered unsuitable because these devices cannot be easily made to absorb time-scale alterations.

The storage system contains two storage channels, each consisting of an MIT tube and the circuits which are necessary to operate the tube. Incoming pulses are routed to one of the two storage channels while pulses which have previously been stored in the other channel are being recovered; these operations are reversed when the channel in which storage is taking place is completely filled.

A continuous flow of information can be maintained through the system since the two storage channels perform opposite functions at any given time; one receives incoming pulses while stored pulses are being recovered from the other. All circuits have been made insensitive to pulse-repetition-frequency effects. The order of stored pulses is preserved, but their time relationship is lost. The time relationship of pulses recovered from storage is determined by an external source. Consequently, neither the input pulses nor the output pulses need have any periodic relationship; a delay or an alteration of the time scale may be easily obtained.

The MIT electrostatic storage tube was developed for use in the Whirlwind computer, and has been described in the literature.[3] The tube comprises a target assembly and two electron guns, one of which produces and the other maintains the two stable target potentials used to represent the two states of binary pulses. The high-velocity gun emits a narrow beam which can be positioned to any point on the target and which is turned on only when it is desired to write or read a pulse. The holding-gun beam covers the entire target, and is turned on except when writing or reading is taking place.

To write a pulse, it is necessary to position the high-velocity beam, turn off the holding beam, apply an ap-

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* Decimal classification: R339. Original manuscript received by the Institute, August 11, 1950; revised manuscript received, February 16, 1951.

The work described in this paper is discussed in more detail in Technical Report No. 154, Research Laboratory of Electronics, MIT, based on a thesis, "Storage of Pulse Coded Information," submitted in partial fulfillment of the requirements for the degree of Master of Science at the Massachusetts Institute of Technology, September, 1949. This work has been supported in part by the Signal Corps, the Air Materiel Command and the ONR.

† Research Laboratory of Electronics, Massachusetts Institute of Technology, Cambridge, Mass.

[1] A. J. Lephakis, "Storage devices for communication," *Electronics*, vol. 23, pp. 69–73; December, 1950.

[2] Proceedings, "Symposium on Large-Scale Digital Calculating Machinery," Harvard University Press, Cambridge, Mass.; 1948.

[3] S. H. Dodd, H. Klemperer, and P. Youtz, "Electrostatic storage tube," *Elec. Eng.*, vol. 69, pp. 990–995; November, 1950.

propriate potential to the signal plate of the target assembly, energize the high-velocity beam for a short time, and finally turn on the holding beam. It is not necessary to erase before writing. A stored pulse is read by performing essentially the same operations, except that the high-velocity beam is intensity-modulated with a 10-mc radio-frequency voltage. The state of the stored pulse is determined by comparing in a phase-sensitive detector this voltage with the signal-plate rf output.

The design of the storage-system circuits was based on tentative storage-tube data which were available in the early part of 1949. Provisions were made to change easily the pertinent operating characteristics of the circuits so that future storage tubes might be accommodated. Although circuit flexibility has been obtained at the expense of more complicated equipment, it is believed that the increased complexity is justified by the fact that the equipment is not likely to become obsolete.

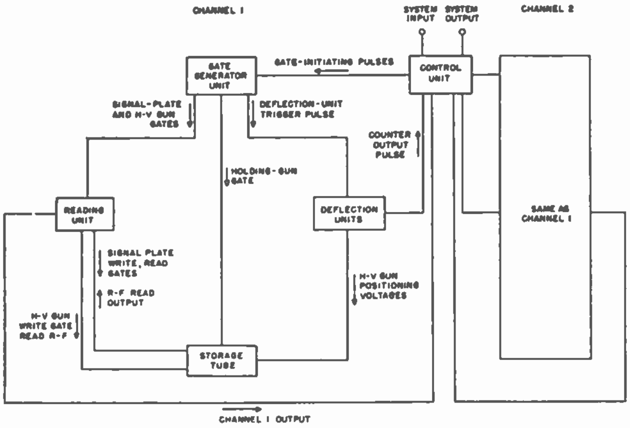

Fig. 1—Block diagram of the storage system.

A block diagram of the storage system is shown in Fig. 1. The circuits consist of four basic units. The deflection units generate the high-velocity-gun positioning voltages. The gate-generator unit generates the writing and reading gates which are required by the signal plate, the high-velocity-gun control grid, and the holding-gun control grid. Each gate is initiated by a trigger pulse. The gate-generator unit also provides the deflection-unit input pulses. The reading unit generates the pulsed oscillation which is applied to the high-velocity-gun control grid during reading, and detects the pulses which are read. The duration of the pulsed oscillation is determined by a gate from the gate-generator unit. The control unit provides the pulses which trigger the gate-generator units associated with the two storage tubes, and determines whether writing or reading is taking place in each storage tube. Switching of the writing and reading operations is caused by pulses which are produced by the deflection units whenever a storage tube is completely full or empty. When tube 1 is full, writing commences in tube 2 and reading commences in tube 1. Tube 1 must be emptied before tube 2 is filled. When tube 1 is empty, its circuits become quiescent and remain quiescent until tube 2 is filled, at which time writing commences in tube 1 and reading commences in tube 2. The circuits of tube 2 become quiescent as soon as this tube is empty, and when tube 1 is full, the cycle of operations is repeated.

It has been attempted to keep the current drain of the various circuits at a minimum, consistent with satisfactory operation, in order to simplify power-supply requirements. The low-impedance, high-current, circuits which are generally required to obtain extremely narrow pulses and extremely small rise and fall times of gates have been avoided. The widths of pulses generated by the equipment are approximately 0.2 μsec, and the rise and fall times of generated gates are approximately 0.2 μsec. Pulse transformers have been employed to invert the polarities of negative pulses and to obtain pulses at a low output-impedance level in order to prevent waste of current in inverter tubes and in cathode-follower output tubes. Pulse-repetition-frequency effects have been minimized by using clamping diodes in all capacitively coupled circuits.

The deflection units comprise two modified decoder circuits which supply 16 or 32 discrete voltage levels to the vertical and to the horizontal deflection plates of the high-velocity gun. Each such circuit consists of five pentode current sources, which supply weighted currents in the ratio 1:2:4:8:16, and are turned on and off by five flip-flops connected as a binary counter. The current-source outputs are added in a common load resistor, and the resulting voltage is applied to an amplifier which provides the balanced voltage necessary to drive the deflection plates. Each input trigger pulse shifts the output voltage to the next level. An established level is maintained constant until another input pulse is applied because direct coupling is used in the amplifier and between the current sources and the flip-flops. Thirty-two input pulses are required for a complete cycle of operation; at the end of each cycle, an output pulse is obtained from the counter. The output pulse of the horizontal counter is applied to the input of the vertical counter. In this manner, a square array of 32×32 spots, traced in sequence, is obtained on the storage-tube target. At the end of each frame an output pulse is obtained from the vertical counter and is applied to the control unit. A 16×16 array is obtained by by-passing the first stage of each counter.

The gate-generator unit supplies two sets of gates, one for reading and one for writing. All gates are generated by cathode-coupled monostable multivibrators. The gates are amplified and are applied through suitable circuits to the output lines. A typical output circuit is shown in Fig. 2. The gate is applied to the grid of a cathode follower, and the load is connected to the cathode of this tube. A discharge tube, which is normally not conducting, is connected across the load. Because of the load capacitance, the cathode potential of the cathode follower cannot change instantaneously. The positive-going edge of the gate, therefore, causes the grid-to-cathode potential of the cathode follower to increase by a considerable amount, and the resulting high cathode-follower current rapidly charges the capacitance. A large cathode resistor is used in order to allow most of the cathode-follower current to flow into the capacitance. The negative-going edge of the gate cuts off the cathode follower. A regenerative pulse amplifier, triggered by a pulse coincident with the negative-going edge of the gate, turns on the discharge tube which discharges the capacitance.

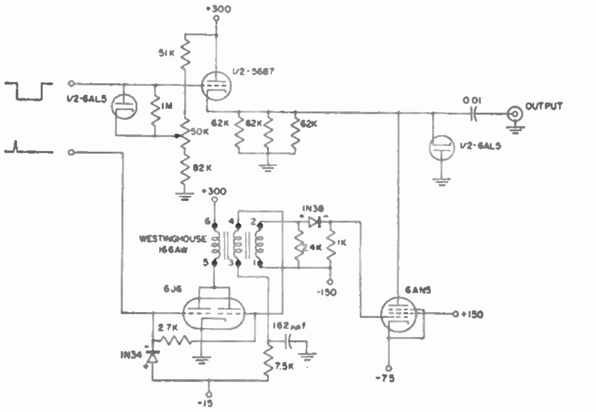

Fig. 2—Schematic diagram of the holding-gun-gate output circuit

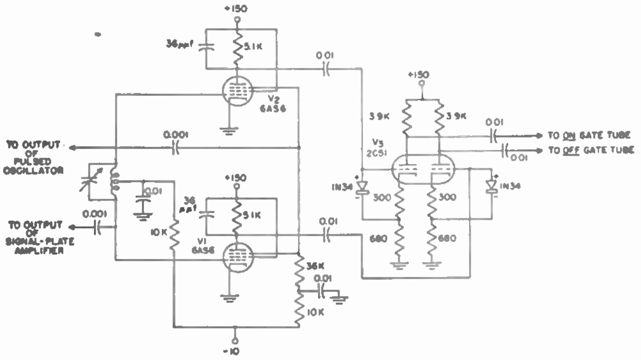

Fig. 3—The phase-sensitive detector used in the reading unit.

The reading unit consists of a 10-mc pulsed oscillator which is energized by the high-velocity-gun read gate, an amplifier to which the signal-plate rf output is connected, and a phase-sensitive detector. The schematic diagram of the latter is shown in Fig. 3. Each of the negative bias voltages which are applied to the suppressor grids and control grids of tubes V_1 and V_2 is sufficient to prevent plate-current flow; plate current will flow only if positive signals are simultaneously applied to both grids. The pulsed-oscillator output is applied to the suppressor grids of the tubes, and the signal-plate-amplifier output is applied to the control grids. The former signal has the same phase relationship at both tubes; the latter appears in phase opposition at the two control grids. Since the signal-plate output is either in phase or 180 degrees out of phase with the pulsed-oscillator signal, these signals coincide in one, and only one, of the two detector tubes. The output of the tube in which coincidence takes place is a negative gate. The positive gate obtained from the corresponding half of amplifier V_3 is applied to the suppressor grid of the *on* or *off* gate tube, which then passes a pulse supplied at the proper time by the control unit.

The major function of the control unit is to route the input pulses to the gate-generator units which operate the two storage tubes. Another function is to delay by appropriate amounts the pulses which initiate the signal-plate and the high-velocity-gun gates; at the start of a write or read cycle the holding beam is turned off first, then the signal plate is brought to the proper potential, and finally the high-velocity beam is turned on. Pulses are delayed by means of monostable multivibrators and pulse-forming tubes; the pulse applied to a multivibrator initiates a gate, and the pulse-forming tube generates a pulse at the trailing edge of this gate. The routing is accomplished by means of gate tubes which are opened and closed by flip-flops triggered by the deflection-unit-counter output pulses. Direct coupling is used between the gate tubes and the flip-flops.

The gate-generator units have been adjusted to allow a writing interval approximately 15 μsec and a reading interval of about 7 μsec. These units require 2 μsec to recover after a set of gates has terminated. The deflection units will provide a new spot location within 2 μsec after being triggered. On the basis of these figures, the maximum frequency limits for periodic operation of the storage system should be approximately 60 kc when writing and 110 kc when reading. It was found, however, that reliable operation did not occur under these conditions. Stored patterns degenerated, probably because of the low holding-gun duty cycle; the holding beam was on only 2/17 of the time during writing and 2/9 of the time during reading. The degeneration may have been aided by a slight defocussing of the high-velocity beam caused by stray magnetic fields.

Reliable periodic operation of the system, independent of frequency, was observed with holding-gun duty cycles of 1/2 or greater: writing frequencies of 33 kc or less, and reading frequencies of 70 kc or less. Tests involving nonperiodic operation have not been performed. However, data obtained from periodic operation indicate that the system will function properly if the minimum interval between adjacent pulses is at least 30 μsec during writing, and at least 15 μsec during reading.

ACKNOWLEDGMENT

The writer expresses his gratitude to J. B. Wiesner, H. E. Singleton, and S. H. Dodd and his colleagues of the Storage Tube Group of the MIT Servomechanisms Laboratory for their many valuable suggestions and advice, and to J. W. Forrester for extending the co-operation of the Servomechanisms Laboratory.

Determination of Aperture Parameters by Electrolytic-Tank Measurements*

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Summary—**In this paper it is shown how the electric and magnetic polarizabilities of an aperture may be determined accurately by electrolytic-analog measurements. Measured magnetic-polarizability data are given for rectangular-, rounded-slot-, cross-, rosette-, dumbbell-, and H-shaped apertures.**

INTRODUCTION

THE ELECTROMAGNETIC problem of an aperture of any shape in an infinitely thin conducting wall between any two regions has been solved in a general manner by Bethe.[1,2] In order for the solution to be valid, it is necessary that the aperture be small compared to a wavelength and compared to the distance to the nearest sharp bend of the wall or other perturbation. Although these limitations appear rather severe, Bethe's method has proven very useful in many applications.

It was shown by Bethe that the field in the vicinity of an aperture may be represented approximately by the original field E_0, H_0 at the location of the aperture before the aperture is cut in the wall, plus the fields of an electric and magnetic dipole located at the center of the aperture. The electric dipole is oriented perpendicular to the aperture and the magnetic dipole is in the plane of the aperture. The strengths of the dipoles are related to their respective original fields by constants of proportionality that are functions only of the shape and size of the aperture. These constants are known as the electric and magnetic polarizabilities P and M. The latter is a dyadic quantity, and hence the magnetic dipole moment is in the same direction as H_0 only if H_0 coincides with a principal axis of the aperture. The two principal axes of an aperture are orthogonal, and may be easily determined from the symmetry that usually exists for a practical aperture shape. The magnetic polarizabilities along the two principal axes are scalar constants that are in general unequal. Coupling formulas containing P and M are given for many important aperture configurations in the literature.[1,3–5]

Formulas for the polarizabilities are given by Bethe only for circular and elliptical apertures, and for long slits where the magnetic field is transverse to the slit. The lack of exact formulas for the polarizabilities of other aperture shapes has been a handicap to the practical application of Bethe's method. Because of the tremendous mathematical difficulty that would be involved in obtaining precise formulas for shapes other than those previously analyzed, a method of accurate measurement is necessary. Microwave measurement of the polarizabilities immediately suggests itself, but the probable error would be of the order of 10 per cent. In this article an electrolytic-tank method capable of an accuracy of about 1 per cent is described, and extensive graphical data for M are presented for many practical aperture shapes.

MEASUREMENT METHOD FOR THE ELECTRIC POLARIZABILITY

It will now be shown how the electric polarizability P of an aperture may be determined quantitatively from electrolytic-tank measurements. First, a formula will be derived that relates the electric polarizability of an aperture to the change in capacitance occurring when a magnetic-wall model of the aperture is inserted between a pair of parallel conducting plates. Then the electrolytic analog of this configuration will be presented, and the formula relating P to resistance measurements in an electrolytic cell will be given. CGS-Gaussian units are used unless otherwise indicated.

Assume a divided cell, as shown in Fig. 1(a). The cell consists of three equispaced horizontal electric walls and four vertical magnetic walls. The central electric wall, which is infinitely thin, contains an aperture of arbitrary shape whose electric polarizability is desired. With a voltage applied as shown, the field in region (2) is antisymmetrical to that in (1) about the central plane. Let the largest dimension l of the aperture be very small compared to the dimensions of the box. Then at distances $r \gg l$ from the aperture, the field in region (2) is the same as that existing if the aperture were not present plus the field of an electric dipole having the following dipole moment:

$$\Pi_e = \frac{P}{2\pi} E_0, \qquad (1)$$

where P is the electric polarizability and E_0 the exciting field. This relation is (25) of Bethe's Radiation Laboratory Report.[1] (Bethe inserted the $1/2\pi$ factor in order to rationalize the electric-polarizability formula for a

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† Sperry Gyroscope Company, Great Neck, N. Y.

[1] H. A. Bethe, "Lumped constants for small irises," *MIT Rad. Lab. Rep.* 43-22; March 24, 1943.

[2] H. A. Bethe, "Theory of diffraction by small holes," *Phys. Rev.*, vol. 66, p. 163; 1944.

[3] C. G. Montgomery, R. H. Dicke, and E. M. Purcell, "Principles of Microwave Circuits," Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y., vol. 8; 1948.

[4] M. Surdin, "Directional couplers in waveguides," *Jour. IEE*, vol. 93, pt. IIIA, p. 725; 1946.

[5] C. G. Montgomery, "Technique of Microwave Measurements," Radiation Laboratory Series, McGraw-Hill Book Co., New York, N. Y., vol. 11; 1947.

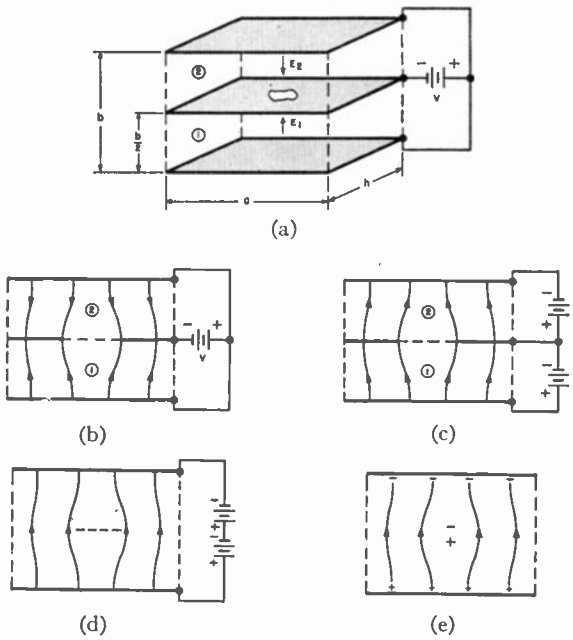
circular aperture.) The exciting field E_0 was defined verbally by Bethe on page 6 of his report, as follows:

$$E_0 = E_1 - E_2, \tag{2}$$

where E_1 and E_2 are the incident fields in regions (1) and (2). Since $E_2 = -E_1$, $E_0 = 2E_1$ and

$$\Pi_e = \frac{PE_1}{\pi}.$$

Similarly, the field in region (1) for $r \gg l$ is the original field plus that of a dipole oriented opposite to the previous one.



(a)

(b) (c)

(d) (e)

Fig. 1—Configuration for the electric-polarizability derivation.

The actual normal field in the plane of the aperture is zero in Fig. 1(a), and, therefore, one may close the aperture with a thin magnetic sheet without disturbing the field. This is shown in Fig. 1(b), where magnetic-wall surfaces are represented by dotted lines and electric-wall surfaces by solid lines. Because the two regions of Fig. 1(b) are isolated electrically, one may reverse the arrows on the field lines in region (2), as shown in Fig. 1(c), without otherwise affecting the field orientation. Now the electric field on the electric-wall portion of the partition is continuous across the partition, and hence the electric wall may be removed leaving only a thin magnetic-wall obstacle (Fig. 1(d)). The field in the entire cell of dimensions a, b, h is (for $r \gg l$) a superposition of the incident field E_1 plus the field of a single dipole replacing the obstacle of strength.

$$\Pi_e = -\frac{PE_1}{\pi}. \tag{3}$$

The minus sign takes account of the orientation of the dipole opposite to the impressed field.

Page and Adams give the following equation for the dielectric constant of a medium:[6]

$$\epsilon_r = 1 + 4\pi s_e, \tag{4}$$

where ϵ_r is the average dielectric constant relative to free space and s_e is the electric susceptibility of the medium, or the induced polarization per unit volume and unit impressed electric field. In the case of the cell of Fig. 1(e), the impressed field is E_1 and the volume is abh. The susceptibility, therefore, is

$$s_e = \frac{\Pi_e}{E_1 abh} = -\frac{P}{\pi abh} \tag{5}$$

and the dielectric constant is

$$\epsilon_r = 1 - \frac{4P}{abh}. \tag{6}$$

Let C_1 be the capacitance between the electric walls of the cell with the magnetic-wall obstacle inserted, and let C_2 be the capacitance with the obstacle removed. Then, since capacitance is proportional to dielectric constant,

$$\frac{C_1}{C_2} = \frac{\epsilon_r}{1} = 1 - \frac{4P}{abh}.$$

The electric polarizability is therefore given by the following expression:

$$P = \frac{abh}{4}\left(\frac{C_2 - C_1}{C_2}\right). \tag{7}$$

By means of this formula, the electric polarizability could be determined by capacitance measurements if the configurations were realizable. This is not the case, however, since not even a poor microwave approximation for a magnetic wall exists in nature. Examination of the analogy between conductance in the electrolytic tank and capacitance in free space shows, however, that the magnetic walls may be replaced by nonconductors in the tank, and hence the electrolytic-tank analog may be perfectly realized.[7]

In Fig. 2 (see page 1418) is shown the electrolytic cell which is the exact analog of the capacitance cell of Fig. 1(d). Two of the vertical boundaries are conductors (electric walls), the other two vertical boundaries, the bottom, and the surface of the liquid are nonconductors that simulate magnetic walls. The obstacle, which may be suspended by fine threads, is a thin, nonconducting model of the aperture under test. Let $G_1 = 1/R_1$ be the conductance that is analogous to C_1 and $G_2 = 1/R_2$ be the conductance analogous to C_2. Then in terms of the measurable resistances R_1 and R_2, P is given by

$$P = \frac{abh}{4}\left(\frac{R_1 - R_2}{R_1}\right). \tag{8}$$

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[6] L. Page and N. I. Adams, "Principles of Electricity," D. Van Nostrand and Co., Inc., New York, N. Y., p. 44; 1931.

[7] S. B. Cohn, "Electrolytic-tank measurements for microwave metallic delay-lens media," *Jour. Appl. Phys.*, vol. 21, pp. 674–680; July, 1950.

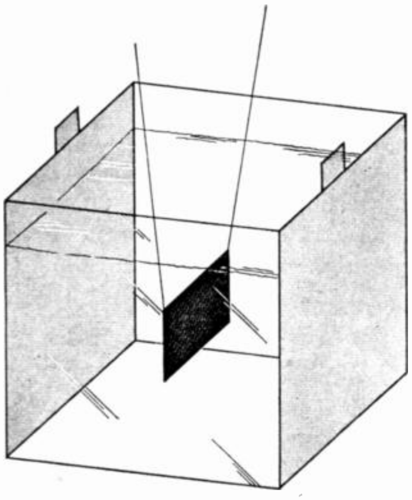
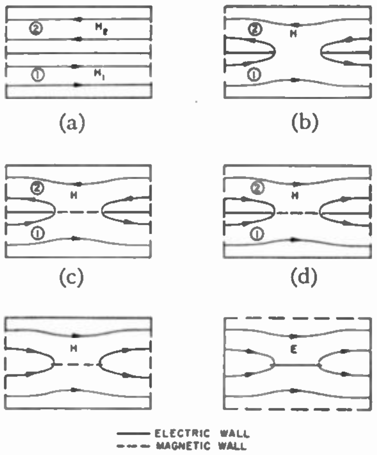


Fig. 2—Electrolytic cell containing a thin nonconducting obstacle.



(a) (b) (c) (d) (e) (f)

Fig. 3—Configuration for the magnetic-polarizability derivation.

MEASUREMENT METHOD FOR THE MAGNETIC POLARIZABILITY

The derivation of the relationship for the magnetic polarizability of an aperture is similar to that for the electric polarizability. As a final step, however, an application of Babinet's electromagnetic duality principle is necessary in order to obtain a configuration suitable for electrolytic measurement. The derivation is as follows:

Assume the divided rectangular cell shown in Fig. 3(a). In this case the cell consists of three equispaced horizontal electric walls, two vertical electric walls, and two vertical magnetic walls. Let equal but oppositely-directed uniform fields H_1 and H_2 exist in regions (1) and (2) of Fig. 3(a). Since the field lines are parallel to the electric walls and perpendicular to the magnetic walls, the boundary conditions in the box are satisfied. Now assume a small aperture in the infinitely thin central wall, with one of its principal axes oriented parallel to the original field. Then the field adjusts itself as shown in Fig. 3(b). As in the electric polarizability case, the field in region (2) far from the aperture is equal to the initial field H_2, plus the field of a magnetic dipole having the magnetic dipole moment

$$\Pi_m = \frac{M}{2\pi} H_0, \tag{9}$$

where the exciting field H_0 is given by

$$H_0 = H_1 - H_2 = 2H_1. \tag{10}$$

Because of the symmetry of the configuration, the aperture may be filled with an infinitely thin magnetic wall without disturbing the field (Fig. 3(c)). Since the two regions are now isolated, the field in region (2) may be reversed without affecting region (1) (Fig. 3(d)), and then the electric-wall portions of the central barrier may be eliminated without disturbing the field. These steps have produced the cell of Fig. 3(e) that contains a magnetic-wall model of the aperture. The average magnetic permeability of the cell containing the obstacle is given by

$$\mu_r = 1 + 4\pi s_m, \tag{11}$$

where μ_r is the average permeability relative to free space and s_m is the induced magnetic polarization per unit volume and unit impressed magnetic field. For the impressed field H_1 and cell volume abh, the permeability is

$$\mu_r = 1 + \frac{4M}{abh}. \tag{12}$$

Since, however, the cell has the field flowing in and out of its two magnetic wall boundaries, a direct electrolytic analog is not possible, and therefore the boundaries will be altered by application of Babinet's Principle. This principle states that if all the electric and magnetic walls of a nondissipative region are interchanged, and if the permeability and dielectric constant are interchanged, then E may be replaced by $-H$ and H may be replaced by E. These alterations transform Fig. 3(e) into 3(f), which has a direct electrolytic analog. The average dielectric constant of Fig. 3(f) is

$$\epsilon_r = 1 + \frac{4M}{abh}. \tag{13}$$

Let C_1 and C_2 be the respective capacitances between the electric walls of the cell with and without the obstacles present. Then

$$\frac{C_1}{C_2} = 1 + \frac{4M}{abh}. \tag{14}$$

The electrolytic analog of Fig. 3(f) is identical to that for the electric-polarizability measurement, except that a thin conducting model of the aperture parallel to the incident current flow is utilized instead of the nonconducting obstacle. In terms of the previously defined resistances, the magnetic polarizability is given by

$$M = \frac{abh}{4}\left(\frac{R_2 - R_1}{R_1}\right). \tag{15}$$

APPARATUS

A photograph of the electrolytic cell used for the magnetic-polarizability measurements is shown in Fig. 4. Two of the vertical walls inside the cell have a conducting surface of rhodium, while the other vertical walls and the bottom are nonconducting lucite. The internal dimensions of the cell are 6×6×6 inches. When in use, the cell was filled with a dilute aqueous solution of potassium chloride to a height of approximately $5\frac{1}{2}$ inches. The concentration was adjusted to give a cell resistance of about 7,000 to 10,000 ohms. The obstacles for the magnetic-polarizability measurements, which were supported in the cell by fine nylon threads, were rhodium-plated 0.0015-inch copper sheet.

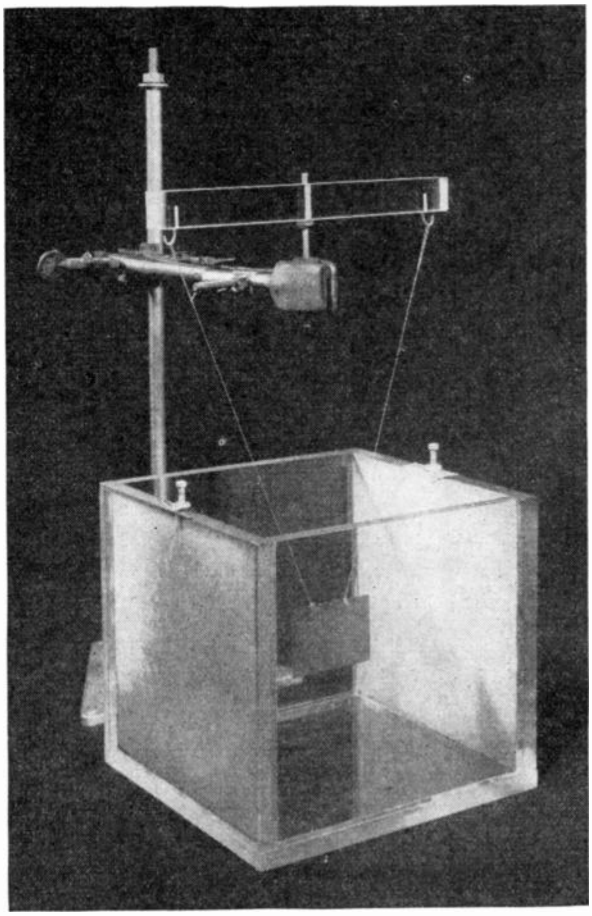


Fig. 4—Photograph of the cell and a suspended metallic obstacle.

The cell resistance was measured by a Wheatstone bridge utilizing a 1,000-cps generator and an oscilloscope detector. This equipment is discussed in more detail in a recent article on electrolytic-tank measurements for metallic delay lenses.[7] The bridge for the aperture measurements was made more sensitive, however, by the addition of 0.1-ohm steps to the variable arm. With this modification, a change of one part in 20,000 in the cell resistance was discernible by the detector.

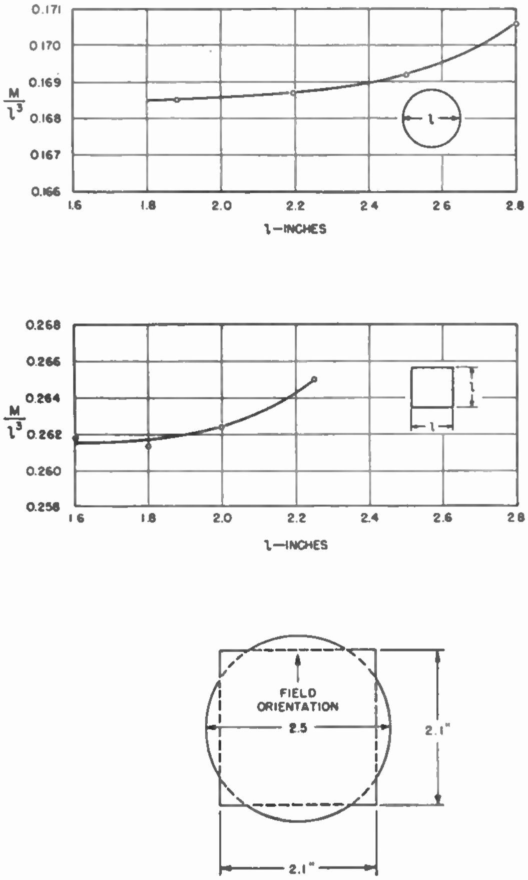


Fig. 5—Proximity effect and the criterion for maximum size.

MAJOR EXPERIMENTAL ERRORS

In the derivation of (15), a metallic obstacle small compared to the cell was assumed. Above a certain size, the accuracy of the equation would be affected by the proximity of the cell walls. It is desirable to use the largest obstacle for which proximity effects may be neglected, since the larger the obstacle the greater the difference between R_1 and R_2, and hence the greater the precision of measurement. In order to determine the maximum permissible size, circular and square obstacles of various sizes were tested. The ratio M/l^3, which should be constant in the ideal case, is plotted in Fig. 5. It is seen that the curves are almost flat for the smaller obstacles, and bend upward for the larger obstacles. A diameter of 2.5 inches has been arbitrarily chosen as the maximum allowable for the circle and a length of 2.1 inches for the square. In a similar manner, the maximum dimensions of all other shapes might be determined, but this is not a practical procedure since it requires the measurement of far too many obstacles. In order to reduce the number of measurements to one for each shape of obstacle, a criterion has been arbitrarily established that the proximity effect may be neglected for any obstacle that fits within the solid-line boundary shown in Fig. 5(c). This criterion is a reasonable one since the maximum allowable circle and square chosen from Figs. 5(a) and 5(b) fit the pattern.

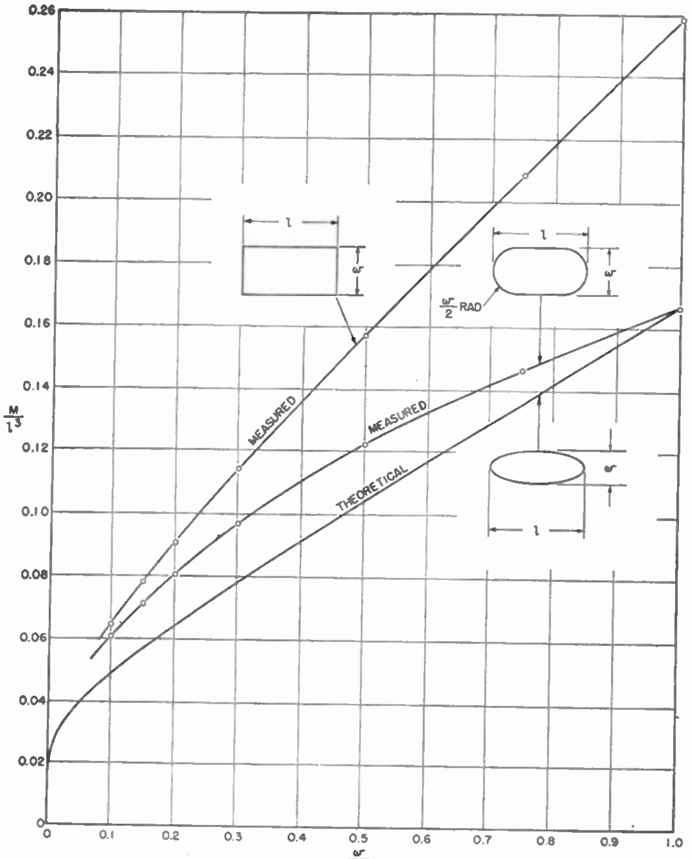


Fig. 6—Magnetic polarizabilities of rectangular, rounded-end, and elliptical slots, H parallel to l.

The theoretical M/l^3 ratio for a circle is exactly 1/6 or 0.16666 . . . . In Fig. 5(a) it is seen that for the smallest circle the ratio was 1.1 per cent higher than theoretical, while for the 2.5-inch circle the ratio was 1.57 per cent above theoretical. This error is due not only to the proximity effect but also to the finite thickness of the obstacle and to equipment errors. In order to compensate as well as possible for the residual error due to proximity, thickness, and the like, all later M/l^3 data have been multiplied by 0.987 since this factor reduces the measured values of Fig. 5(a) in the range of permissible circle diameters to approximately the theoretical value. Although it is not rigorously justifiable to use this factor for shapes other than circles, it is believed that more accurate results are achieved thereby than if no correction factor were used at all.

As a further check on the accuracy of the measurement method, three elliptical obstacles having different eccentricities were tested with the field parallel to the major axes. In each case, the measured value of M and the theoretical value computed from Bethe's exact formula checked to 1 per cent or better. The excellent agreement verified the method and the 0.987 factor.

THE MAGNETIC-POLARIZABILITY DATA

The data for rectangular apertures and for slots with rounded ends are plotted in Fig. 6 for the case of the incident field oriented parallel to the long dimension of the apertures. As a comparison the theoretical curve for the elliptical shape is included in the graph. Fig. 7 shows the measured data for dumbbell-, cross-, and rosette-shaped apertures. For the first, the incident magnetic field is assumed in the direction of l. For the others M/l^3 is independent of the orientation of the field. Note that all three shapes reduced to a circular aperture for $w/l=1.0$. Also note that the dumbbell reduces to a rounded slot for $w/l=0.1$.

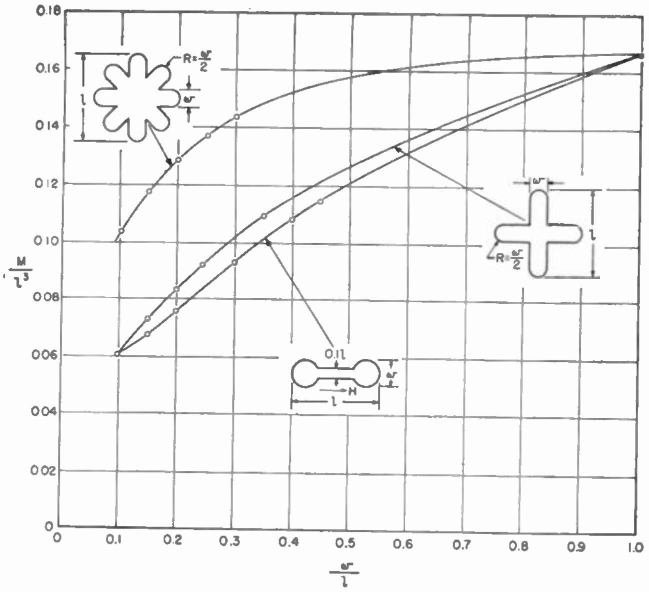


Fig. 7—Magnetic polarizabilities of rosette-, cross-, and dumbbell-shaped apertures.

For w/l between 0.1 and 1.0 the M/l^3 curves for the dumbbell and cross are very nearly the same as that for the rounded slot. This indicates that the magnetic polarizability is determined almost entirely by the shape and size of the extremities of the aperture along the magnetic-field direction and that the effect of the intermediate portions of the aperture on M is very small. This observation should prove of value for applying the available data to other related aperture shapes. For example, although the connecting bar of the dumbbell was $0.1l$ for Fig. 7, the curve may be used with very good accuracy for any bar width less than w.

The data for an H-shaped aperture are shown in Fig. 8 for the magnetic field in the l direction. Note that for w/l between 0.1 and 0.3, M/l^3 is within 15 per cent of the value for a rectangular slot having $w/l=0.5$. This is not surprising in view of the preceding discussion. For $w/l=0.5$, the H-shaped aperture reduces to a rounded slot, and therefore the value for the latter is plotted at this point in Fig. 8.

Values of M/l^3 taken from the original graphs are given in Table I, below. These values are believed to have an accuracy of the order of 1 per cent.

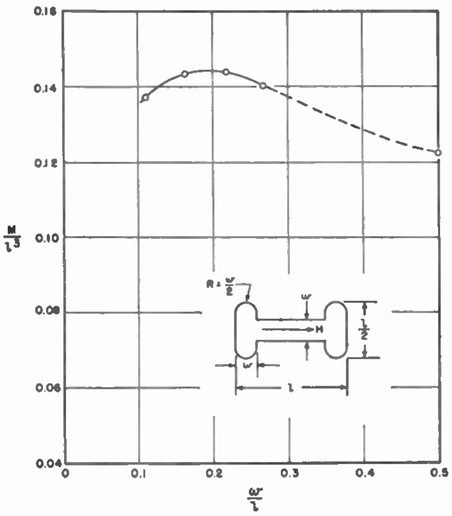


Fig. 8—Magnetic polarizability of an H-shaped aperture.

TABLE 1

Values of $M/2^3$

	$\frac{W}{2}=0.1$	0.15	0.2	0.25	0.3	0.35	0.5	0 .75	1.0
Rectangle	0.0645	0.0780	0.0906		0.1139		0.1575	0.2096	0.2590
Rounded slot	0.0610	0.0711	0.0801		0.0964		0.1222	0.1455	0.1667
Cross	0.0611	0.0728	0.0832	0.0930		0.1093			0.1667
Rosette	0.1028	0.1172	0.1282	0.1368	0.1434		0.1208		0.1667
Dumbbell*	0.0610	0.0675	0.0757	0.0848	0.0932	0.1012	0.1222		0.1667
H-shape	0.1358	0.1426	0.1442	0.1418					

* Width of bar = 0.12.

Combination Open-Cycle Closed-Cycle Systems*

J. R. MOORE†

Summary—**The ancient idea of a combination coarse and fine adjustment is shown to be applicable to the design of precision automatic control systems. In the particular class of systems discussed, the coarse adjustment is taken to be a separate element operated by the input, but outside the feedback loop. Its position outside the feedback loop qualifies the coarse controller as an open-cycle system, and makes it possible to introduce such elements without affecting the system's transient response adversely. In this way, interference equalization of dynamical distortion errors is possible without such critical dependence being placed on a knowledge of series elements of the system as is required for interference equalization by a controller in the feedback loop.**

Three broad types of open-cycle systems are discussed: series, parallel, and partially parallel. Each of these may be "algebraic," "differential," or a combination. The algebraic controllers are useful when the average value of the input signal is predictable—particularly where a repetitive duty cycle is encountered.

The advantages of adding completely parallel open-cycle elements for improving speed range and reducing over-all cost are shown. Idea is also applicable to nonlinear and multiplicoupled systems.

INTRODUCTION

IN THE DEVELOPMENT of methods for designing automatic control systems, the effort has been concentrated almost exclusively on systems with feedback, called "closed-cycle systems," as ideally illustrated in Fig. 1(a). The reason for this is that, by the use of feedback, the designer's ignorance of the exact nature of his system elements can be rendered unimportant, and a performance which, for the great majority of applications is satisfactory, can be obtained with a minimum of effort.

Although it is theoretically possible to make a com-

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pletely open-cycle control system (Fig. 1(b)) as accurate as a completely closed-cycle system by knowing almost everything about the load and unalterable elements, the computer required for such a system would usually be impossibly complicated. Furthermore, the effort would probably be foredoomed to failure because the actuators and unalterable elements found in practice almost invariably integrate their inputs, thereby causing errors to accumulate.

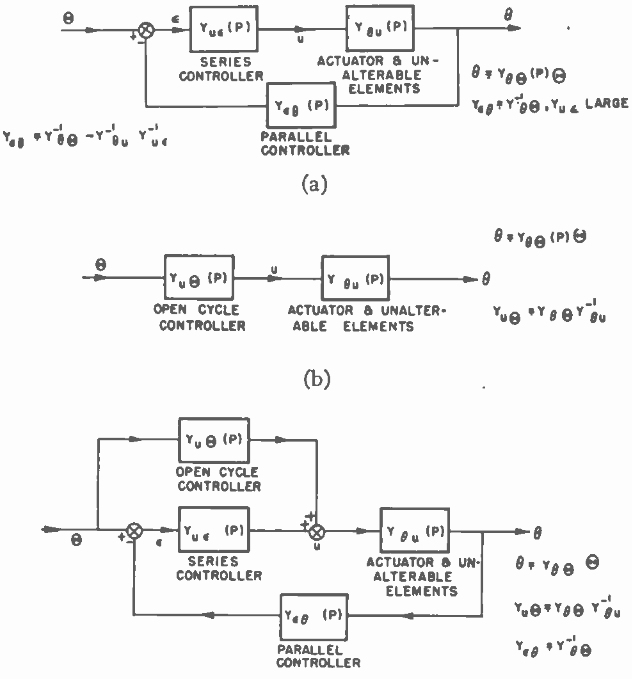

Fig. 1—(a) Idealized closed-cycle system. (b) Idealized open-cycle system. (c) Idealized combination (partially parallel) open-cycle closed-cycle system.

However, an open-cycle system may be combined with a closed-cycle system (as shown in Fig. 1(c)) to form an automatic control which is superior to either. Here the open-cycle system acts as the coarse controller providing the major portion of the output, while the closed-cycle system acts as the vernier.

The combination open-cycle closed-cycle system can be made to have the following advantages:

1. Small static errors.
2. Reduction of velocity, acceleration and higher derivative errors by factors of from two to more than 100 (compared with conventional closed-cycle systems).
3. Minimization of effects of disturbances not superposed on the desired input.
4. Maximum use of predictable input characteristics, particularly in applications with approximately periodic duty cycles.
5. Interference equalization without affecting the loop stability (contrary to the condition which exists with a parallel controller inside the feedback loop).
6. Possibility of making the closed-cycle system highly damped without reducing the velocity acceleration and higher derivative coefficients inordinately. (This overlaps advantages 2 and 3 above).
7. Possible reduction of required quality of large actuators and their controls without jeopardizing performance, thus leading to an over-all reduction in cost compared to completely closed-cycle systems of comparable performance.
8. Possibility of handling nonlinear and multipli-coupled systems with less apparatus than required for completely closed-cycle systems; furthermore, the form of this apparatus is often simpler and its perfection less critical than for corresponding parallel controllers.

The present paper attempts to illustrate these advantages and to document, formalize, and extend the basic idea of the combined coarse and fine adjustment (which many engineers and scientists have applied in some form or other to special problems of automatic control for several years[1]) via the use of open-cycle coarse control superposed on a closed-cycle vernier.

COMBINATION OPEN-CYCLE CLOSED-CYCLE SYSTEMS

Despite the shortcomings of many completely open-cycle control systems, open-cycle elements are often useful for coarse controls. As indicated previously, these can, in many cases, be combined profitably with closed-cycle systems as fine adjustments to produce combination systems which are better than either a completely closed- or completely open-cycle system by itself.

Two basic ideas are involved: (a) If the input (s) is approximately predictable as a function of time—particularly if it is approximately periodic (as in the duty cycles of many processes)—this information may be used to construct an "algebraic"[2] type of controller which almost produces the correct output without ever comparing it with the desired input function. Such a controller may involve cams and computers producing a coarse output directly (in which case they are truly "algebraic") or the cams and computers may control the input to an actuator, thereby producing a "quasi-algebraic" controller. And (b) If the input is unpredictable, but the dynamic characteristics of actuators, unalterable elements, and the load are predictable in form (by "dynamic characteristics" is meant the equations relating inputs and outputs) a controller can be built which has approximately the inverse characteristics and which produces the input to the actuators.

These ideas are not confined to linear systems or linear elements. Indeed some of their greatest utility occurs

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[1] One simple manifestation of the idea was mentioned by R. E. Graham as "feed forward" in his paper "Linear servo theory," *Bell Sys. Tech. Jour.*, vol. 25, pp. 616–651; October, 1946.

[2] An "algebraic" device is one whose output is (ideally) independent of the past history of its input.

where nonlinear elements are involved. They may, in general, take one or a combination of the three forms shown in Fig. 2.

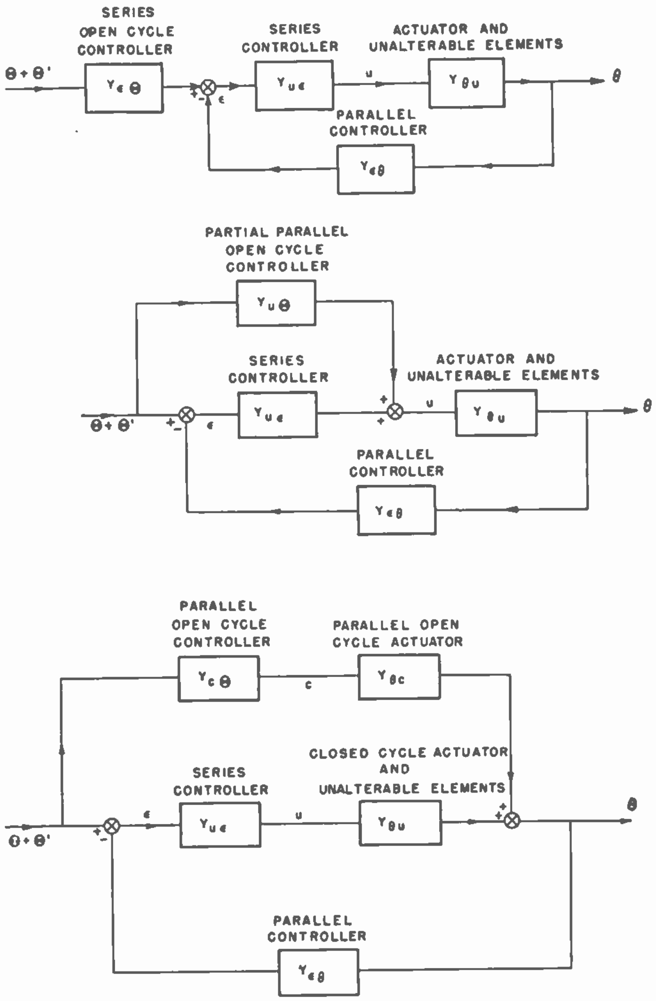

Fig. 2—Series showing partially parallel and completely parallel open-cycle elements.

THEORETICAL COMPARISON WITH CLOSED-CYCLE SYSTEM SYNTHESIS

It will be seen that, in any type of combination open-cycle closed-cycle system, the open-cycle portions operate directly on the inputs and, therefore, unless unstable in themselves, do not affect the stability of the system nor its response to other desired or undesired inputs. In this way, they differ from parallel equalizers or minor loop controllers. Consequently, open-cycle controllers make it possible to design the system loops for maximum external noise suppression and low-frequency accuracy, while in themselves providing the necessary higher frequency response.

The full significance of this statement will be brought out in the discussion (to follow) on partially parallel open-cycle controllers versus parallel closed-cycle controllers. However, a brief consideration of the general problem of synthesis of closed-cycle systems will serve to introduce the idea. Referring to Fig. 1(a), the idealized closed-cycle two-variable system is shown. Here the important elements are indicated by diagram blocks.

If the primary controller is series, it reduces errors by compensating for actuator inadequacies and providing sufficient error magnification over the desired operating "frequency" range to insure a satisfactorily small error. This may be considered control by "division" or "inversion," since the error is minimized by dividing it by the series controller amplification factor at each frequency, or by inverting undesirable characteristics of the unalterable elements. Here the parallel controller is used merely to convert the output into a form suitable for comparison with the input.

If the primary controller is parallel, it reduces errors by subtraction and so may be thought of as controlling by "interference." Here the series controller may still be used to minimize the error by inversion and to "linearize," or otherwise clean up, the actuator and unalterable element dynamics.

These ideas may be illustrated for the broad problems of controller synthesis by considering the error equation of an idealized linear closed-cycle system in the presence of errors in the controller functions themselves. Thus, referring to Fig. 1(a), the output is

$$\theta = \frac{(\Theta + \Theta')}{Y_{\epsilon\theta} + Y_{\theta u}^{-1} Y_{u\epsilon}^{-1}},$$

Here the transfer admittances Y_{ij} are assumed to be functions of the operator $P(=d/dt)$, and Θ' is the input noise. The equation is, of course, identical with the zero initial condition Laplace transform equation if P is replaced by the complex variable.

For purposes of the illustration, let us assume that the control system is a pure position servo. For this type of system, we wish to make $\theta = \Theta$. This will be written $\theta \mathrel{\overset{\cdot}{=}} \Theta$, where $\mathrel{\overset{\cdot}{=}}$ means "desired equal to."

As a result, the error is

$$\mathcal{E} = \theta - \Theta = \frac{[(1 - Y_{\epsilon\theta}) - Y_{\theta u}^{-1} Y_{u\epsilon}^{-1}]\Theta + \Theta'}{Y_{\epsilon\theta} + Y_{\theta u}^{-1} Y_{u\epsilon}^{-1}}. \quad (2)$$

We call the error in Θ, the "distortion error," and the error in Θ' the "input noise error."

If, now, the parallel controller is not to be used for interference equalization, we set

$$Y_{\epsilon\theta} = 1$$

giving

$$\mathcal{E} = \frac{-\Theta}{1 + Y_{\theta u} Y_{u\epsilon}} + \frac{Y_{\theta u} Y_{u\epsilon} \Theta'}{1 + Y_{\theta u} Y_{u\epsilon}}. \quad (3)$$

Obviously the only way in which $\partial\mathcal{E}/\partial\Theta$ can be made small is by making $Y_{u\epsilon}$ large. Evidently $\partial\mathcal{E}/\partial\Theta$ can theoretically be zero only if $Y_{u\epsilon}$ is infinite. This explains the term "synthesis by division." It is seen that if $Y_{u\epsilon}$ is large enough to make $\partial\mathcal{E}/\partial\Theta$ negligible, it is so large as to make $\partial\mathcal{E}/\partial\Theta'$ nearly unity.

Equation (2) shows that ideally $\partial\mathcal{E}/\partial\Theta$ may be made zero by use of a parallel controller if we make the coefficient of Θ zero. This requires that

$$Y_{\epsilon\theta} \doteq 1 - \frac{1}{Y_{\theta u}Y_{u\epsilon}}, \tag{4}$$

which gives

$$\frac{\partial\mathcal{E}}{\partial\Theta'} \doteq 1.$$

However, it is never possible to know any of the Y_{ij} exactly. Let us, therefore, indicate this ignorance as errors $\delta Y_{u\epsilon}$ and $\delta Y_{u\theta}$. This makes (3) more realistically

$$\frac{\partial\mathcal{E}}{\partial\Theta} = \frac{1 + \delta Y_{\epsilon\theta}Y_{\theta u}(Y_{u\epsilon} + \delta Y_{u\epsilon})}{1 + Y_{\theta u}(Y_{u\epsilon} + \delta Y_{u\epsilon})(1 + \delta Y_{\epsilon\theta})}, \tag{5}$$

while the error for the case of parallel equalization is (neglecting $\delta Y_{u\epsilon}$)

$$\mathcal{E} = \frac{-\delta Y_{\epsilon\theta}\Theta + \Theta'}{1 + \delta Y_{\epsilon\theta}}. \tag{6}$$

Evidently, with an unknown $\delta Y_{\epsilon\theta}$ the poles of the error function may have real parts not sufficiently negative to give a satisfactory degree of system stability. This is a limitation on error reduction in a closed-cycle system since it is not often possible to design for the average value of $Y_{\theta u}Y_{u\epsilon}$, but rather requires a complicated investigation of the worst combination of parameters and parameter errors. This precludes the possibility of using average values of system parameters.

By contrast, the partially parallel combination system of Fig. 1(c) relies on the open-cycle element for interference equalization. Here, however, the penalty for ignorance of $Y_{\theta u}$ is not nearly so great since $Y_{u\Theta}$ is outside the feedback loop. As will be shown later, the distortion error is, for a position servo,

$$\frac{\partial\mathcal{E}}{\partial\Theta} = \frac{Y_{\theta u}\delta Y_{u\Theta}}{1 + Y_{\theta u}Y_{u\epsilon}}. \tag{7}$$

Evidently, $\delta Y_{u\Theta}$ cannot affect the poles of $\partial\mathcal{E}/\partial\Theta$ except by adding poles of its own. Since these may always be made to have sufficiently negative real parts, errors in the open-cycle element (or ignorance of closed-cycle elements) cannot affect system stability adversely, and $\delta Y_{u\Theta}$ may be allowed to go as much negative as it can positive. This permits the design of the open cycle element for a suitably weighted average value of $Y_{\theta u}Y_{u\Theta}$, thereby permitting a splitting of uncertainty errors such as would seldom be feasible with a completely closed-cycle system.

BASIC BLOCK DIAGRAMS AND EQUATIONS OF LINEARIZED SYSTEMS

The ideas of the previous section may be understood by consideration of the basic block diagrams of three types of open-cycle closed-cycle systems. These are shown for a linearized single-output single-(nominally) input system in Figs. 3, 6, 7 and 8.

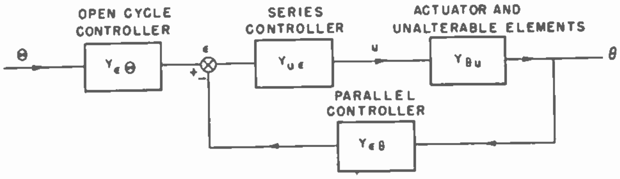

Fig. 3—The series open-cycle controller.

A. The Series Open-Cycle Controller

1. Compensation for Input Distortion Error

Considering first the system of Fig. 3, we note that the open cycle element $Y_{\epsilon\Theta}$ is placed in series with the input so that the system equation is

$$[1 + Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\theta}]\theta = Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\Theta}(\Theta + \Theta'). \tag{8}$$

Evidently this method is useful in the conventional sense if

$$\theta \doteq Y_{\theta\Theta}\Theta \doteq Y_{\epsilon\Theta}\Theta, \tag{9}$$

since this can be done without jeopardizing the stability or system response to disturbances entering elsewhere in the system. It is particularly desirable when the same result could not be conveniently accomplished by the more usual method of making

$$\Theta \doteq Y_{\epsilon\theta}\theta \tag{10}$$

because of the presence of an undesirable $Y_{\epsilon\theta}$ inside the feedback loop.

However, the series open-cycle controller may also be designed to minimize distortion error. Its greatest utility here comes when it is necessary to improve a system "as is" (without modifying it internally in any way, as would be required with a partially parallel controller), or when the inverse of the whole closed-loop system is simpler to mechanize with required accuracy than the inverse of some portion of the system containing the actuator and unalterable elements. As might be expected, the series open-cycle controller has a transfer function which is the inverse of the transfer function of the closed-loop part of the system multiplied by $Y_{\theta\Theta}$. Thus we start with the error equation

$$\begin{aligned}\mathcal{E} &= \theta - Y_{\theta\Theta}\Theta \\ &= \frac{[Y_{\epsilon\Theta} - Y_{\theta\Theta}(Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1})]\Theta + Y_{\epsilon\Theta}\Theta'}{Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1}}.\end{aligned} \tag{11}$$

To make $\partial\mathcal{E}/\partial\Theta$ zero, it is necessary that

$$Y_{\epsilon\Theta} \doteq (Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1})Y_{\theta\Theta}. \tag{12}$$

This makes

$$\mathcal{E} \doteq Y_{\theta\Theta}\Theta'. \tag{13}$$

Actually, as pointed out in the previous section, it is not physically possible to satisfy (12) exactly. Furthermore, it is desirable to compromise between elimination of distortion error and elimination of input noise error. Thus, we assume the more realistic form

$$Y_{\epsilon\Theta} = (Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1})Y_{\theta\Theta} + \delta Y_{\epsilon\Theta}, \qquad (14)$$

where the desired form of $\delta Y_{\epsilon\Theta}$ is controlled by the character of the input noise, the physical realizability of the controller, the nature of the closed cycle distortion error (distribution among various derivatives or frequencies), and the economics of the situation.

The resulting error expression is

$$\mathcal{E} = \frac{\delta Y_{\epsilon\Theta}}{Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1}}\Theta + \left(Y_{\theta\Theta} + \frac{\delta Y_{\epsilon\Theta}}{Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1}}\right)\Theta'. \qquad (15)$$

For physical realizability, $Y_{\epsilon\Theta}$ must never amplify high frequencies. This means that it must either be a constant or a rational fraction of powers of P whose denominator is at least of the same degree as the numerator. Thus

$$Y_{\epsilon\Theta} = \frac{K_{\epsilon\Theta}(1 + t_1P)(1 + t_2P)\cdots(1 + t_nP)}{(1 + T_1P)(1 + T_2P)\cdots(1 + T_{n+r}P)}, \qquad (16)$$

where r is a positive number.

The synthesis procedure will be illustrated for an actual application of the series open-cycle controller to improving the performance of servos on an early model of the Reeves Electronic Analogue Computer (REAC). The purpose of these servos is to convert a voltage into a shaft rotation. Usually this voltage comes from one part of the computer and the servos are used to rotate resolvers or potentiometers. These early computer servos, as designed, emphasized good static accuracy, but were subject to relatively large velocity errors. In particular, when the servos were adjusted for satisfactory damping, the positional error of the output was approximately 1.2 per cent of the numerical value of the rotational equivalent of the input velocity (based on the desired relation between input voltage and output rotations), whereas the positional error of the output was 0.05 per cent of the numerical value of the rotational equivalent of the input acceleration (both being measured in "per second" time units). These give velocity and acceleration error coefficients of 83 and 2,000, respectively.

For the application being described, the acceleration and higher order errors were satisfactory considering possible acceleration of the servo, but the velocity error had to be markedly reduced. Furthermore, it was desirable to work only on the input circuits (explaining the use of a series open-cycle controller instead of the partially parallel open-cycle controller to be described later). In terms of the error expression of (11), if we denote the input voltage E by

$$E = K_{E\Theta}\Theta,$$

we may treat the servo as a pure position servo with

$$\theta \doteqdot \Theta,$$

so that

$$\frac{\partial\mathcal{E}}{\partial\Theta} = \frac{1}{Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1}} - 1.$$

However, the error has already been stated in terms of $\dot{\Theta}$ and $\ddot{\Theta}$ as

$$\mathcal{E} = -\frac{\dot{\Theta}}{83} - \frac{\ddot{\Theta}}{2000} - \cdots. \qquad (17)$$

This means that

$$\frac{1}{Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1}} \approx 1 - \frac{P}{83} - \frac{P^2}{2000} - \cdots. \qquad (18)$$

From which it can be inferred that

$$Y_{\epsilon\theta} + Y_{\theta u}^{-1}Y_{u\epsilon}^{-1} \approx 1 + \frac{P}{83} + \frac{P^2}{1550} + \cdots. \qquad (19)$$

Since we desire only to reduce the P/83 term, it is satisfactory to strive for a

$$Y_{\epsilon\Theta} = 1 + \frac{P}{83} + \frac{\partial Y_{\epsilon\Theta}}{\partial(P^2)}P^2 + \cdots, \qquad (20)$$

where the terms beyond $\partial Y_{\Theta}/\partial P$ are not to be determined, it being required only that they be negligible compared to, or opposite and nearly the same size as corresponding terms in (19). This can be done with a simple rc lead network having a transfer function of the form

$$Y_{\epsilon\Theta} = \frac{1 + \tau P}{1 + K\tau P}$$
$$= 1 + \tau(1 - K)P - K(1 - K)\tau^2P^2 + \cdots. \qquad (21)$$

Taking a reasonable value of K to be 1/10, we have

$$0.9\tau = 0.012, \quad \tau = 0.0135,$$

which makes the series open-cycle controller error

$$\delta Y_{\epsilon\Theta} = -\left(\frac{1}{1550} + \frac{1}{61,000}\right)P^2 + \cdots$$
$$= -0.000661P^2 + \cdots. \qquad (22)$$

Thus we expect the total distortion error to be, from equations (15), (19), and (22),

$$\frac{\partial\mathcal{E}}{\partial\Theta} \approx \frac{-[0.000661P^2 + \cdots]}{1 + \frac{P}{83} + \frac{P^2}{1550} + \cdots}. \qquad (23)$$

The actual circuit used is shown schematically in Fig. 4. Here μ is so large that E_ϵ is negligible.

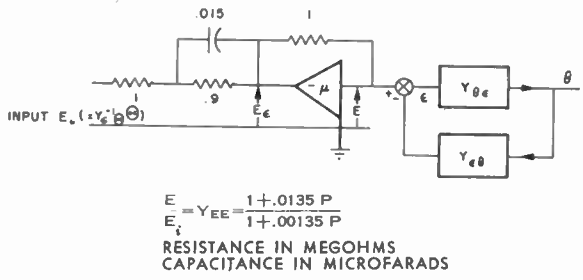

Fig. 4—REAC series open-cycle controller.

Fig. 5 below shows comparison photographs of input and error signals for the REAC servos with and without the series open-cycle controller for a step velocity, and sinusoids of three different frequencies. Evidently the velocity error has been reduced by a factor of about 15:1 for the amplitude of input signal used, whereas the sinusoidal error is improved by factors of 9.5:1, 8:1, and 4:1, respectively, for inputs of 0.5, 1, and 1.5 cps.

B. The Partial Parallel Open-Cycle Controller

Another type of open-cycle controller may be used in parallel with the series controller to produce an input directly to the actuator or actuator power amplifier. This is illustrated in Fig. 6.

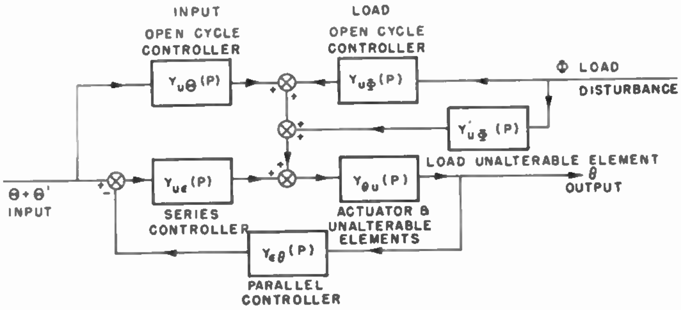

Fig. 6—Partially parallel open-cycle closed-cycle system with load compensation. The form shown here constitutes the generalized "feedforward" of R. E. Graham.

Here the linearized system equation is

$$(1 + Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\theta})\theta = Y_{\theta u}(Y_{u\Theta} + Y_{u\epsilon})(\Theta + \Theta') + Y_{\theta u}(Y_{u\Phi} + Y_{u\Phi}')\Phi \tag{24}$$

where Θ' is the input noise and Φ is the load disturbance. If

$$\theta = Y_{\theta\Theta}\Theta, \tag{25}$$

the error is

$$\mathcal{E} = \theta - Y_{\theta\Theta}\Theta \tag{26}$$

$$= \frac{[Y_{\theta u}Y_{u\epsilon}(1 - Y_{\epsilon\theta}Y_{\theta\Theta}) + Y_{\theta u}Y_{u\Theta} - Y_{\theta\Theta}]\Theta + Y_{\theta u}(Y_{u\Theta} + Y_{u\epsilon})\Theta' + Y_{\theta u}(Y_{u\Phi} + Y_{u\Phi}')\Phi}{1 + Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\theta}}. \tag{27}$$

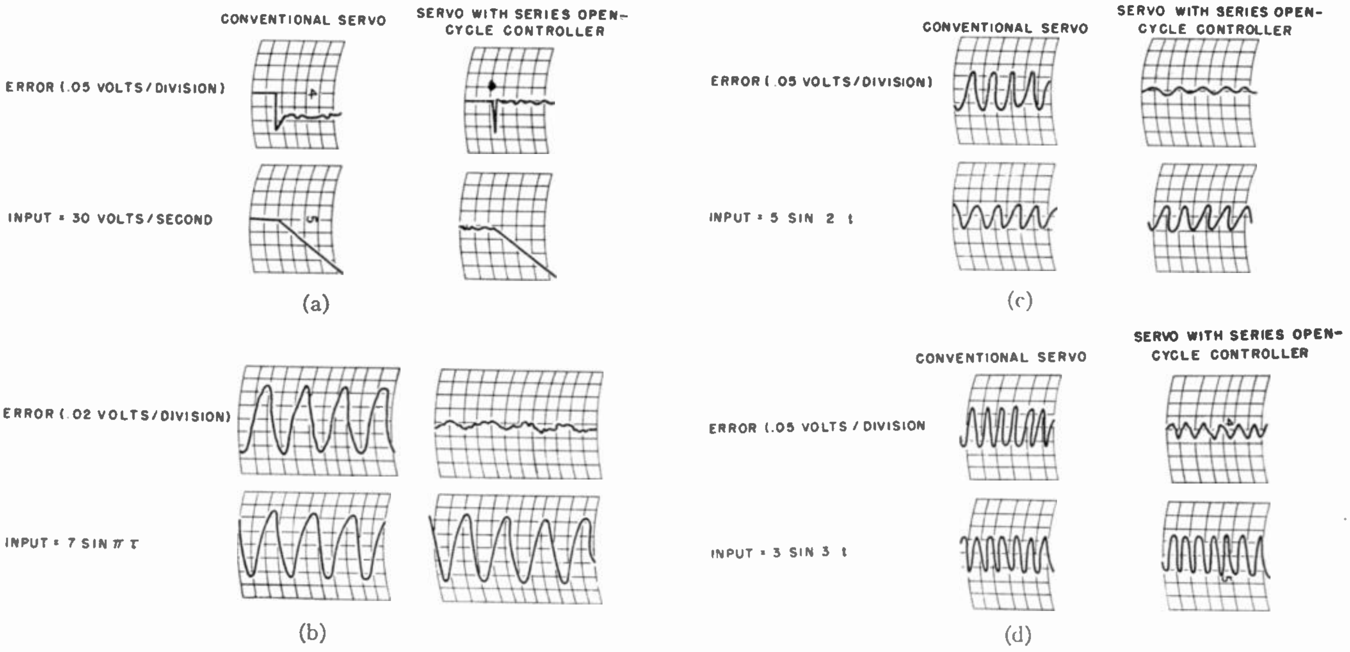

Fig. 5—The effect of open-cycle correction on a simulator servo. (a) Error, 0.05 volts per division; input = 30 volts per second. (b) Error, 0.02 volts per division; input = 7 sin $\pi\tau$. (c) Error, 0.05 volts per division; input = 5 sin 2 t. (d) Error, 0.05 volts per division; input = 3 sin $3t$.

Here it can be seen that if we choose

$$Y_{\epsilon\theta} \approxeq Y_{\theta\Theta}^{-1} \tag{28}$$

$$Y_{u\Theta} \approxeq Y_{\theta\Theta} Y_{\theta u}^{-1} \tag{29}$$

and

$$Y_{u\Phi} \approxeq - Y'_{u\Phi}, \tag{30}$$

(when the symbol $\approxeq$ means "equal over a satisfactory range"), the distortion and load errors can almost be "interfered out," just as with the series open-cycle system. The penalty for errors in $Y_{u\Theta}$ and $Y_{u\Phi}$, together with the input noise, Θ', is an expression of the form

$$\mathcal{E} = \frac{Y_{\theta u}\delta Y_{u\Theta}\Theta + [Y_{\theta\Theta} + Y_{\theta u}Y_{u\epsilon} + Y_{\theta u}\delta Y_{u\Theta}]\Theta' + Y_{\theta u}\delta Y_{u\Phi}\Phi}{1 + Y_{\theta u}Y_{u\epsilon}Y_{\theta\Theta}^{-1}}. \tag{31}$$

Again, since the open-cycle controller does not appear inside the feedback loop, uncertainties in it do not affect stability so that the closed loop may be designed for much higher attenuation than would otherwise have been possible, consistent with satisfactory response to high-frequency inputs.

However, addition of the open-cycle element makes the input noise error worse at high frequencies, illustrating the necessity for a compromise between distortion error and input noise error.

Finally, if Θ' and Φ are impulsive, the system should have good damping in all of its normal modes. In a completely closed-cycle system, good damping is obtained at the expense of gain and distortion error. The combination system permits good damping while, at the same time, keeping the distortion error low.

C. Completely Parallel Open-Cycle Controllers

The previous section considered the use of open-cycle controllers affecting the input to the system actuator or power amplifier. The method of the present section makes use of a separate actuator to feed the coarse correction into an adder where it is combined with the closed-cycle system output. Fig. 7 on page 1428 shows such an idealized servo with negligible load.

The equations of an idealized linearized system are

$$(1 + Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\theta})\theta = (Y_{\theta u}Y_{u\epsilon} + Y_{\theta c}Y_{c\Theta})(\Theta + \Theta') \tag{32}$$

and

$$\mathcal{E} = \frac{[Y_{\theta c}Y_{c\Theta} - Y_{\theta\Theta} + Y_{\theta u}Y_{u\epsilon}(1 - Y_{\theta\Theta}Y_{\epsilon\theta})]\Theta + (Y_{\theta c}Y_{c\Theta} + Y_{\theta u}Y_{u\epsilon})\Theta'}{1 + Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\theta}}. \tag{33}$$

As in the previous section, the best results are obtained by making

$$1 - Y_{\theta\Theta}Y_{\epsilon\theta} \approxeq 0 \tag{34}$$

and

$$Y_{c\Theta} \approxeq Y_{\theta\Theta}Y_{\theta c}^{-1}. \tag{35}$$

The practical or theoretical inability to satisfy (35) yields an error equation

$$\mathcal{E} = \frac{Y_{\theta c}\delta Y_{c\Theta}\Theta + (Y_{\theta\Theta} + Y_{\theta u}Y_{u\epsilon})\Theta'}{1 + Y_{\theta u}Y_{u\epsilon}Y_{\epsilon\epsilon}}. \tag{36}$$

The greatest advantages of a completely parallel open-cycle system accrue when the closed-cycle actuator and unalterable elements are of limited capacity, or when practical and economic factors (such as will be considered in more detail later) intervene.

PRACTICAL ADVANTAGES OF THE PARALLEL OPEN-CYCLE CLOSED-CYCLE SYSTEMS

The purely theoretical advantages of idealized completely parallel open-cycle systems are complemented by the practical advantages of the actual devices used. Thus, any actuator has a very definite ratio of maximum to minimum speed (set by friction, slot locking, or other factors) over which it will run smoothly. This range may be increased by the introduction of "jitter" but is, nevertheless, often less than desired for wide range applications. As a result, when the servo must operate smoothly at speeds which are a small fraction of its maximum, a single actuator may be unable to accomplish the task, whether inside or outside the feedback loop. For such a situation, it is often better to resort to the completely parallel open-cycle closed-cycle system of Fig. 7 in which the open-cycle actuator is relied upon to produce approximately the required output speed, and the closed-cycle system serves to add or subtract small corrections to the open-cycle output. In this way, the maximum speed (referred to the output) required of the closed-cycle actuator and, correspondingly, its maximum power, may be held to, say, less than 10 per cent of what would have been required had no open-cycle element been used. Thus, if the closed-cycle actuator has only the same speed range as its open-cycle counterpart, it can operate at more than ten times the gear reduction, thereby increasing the speed range of the system by a similar factor. In such operation, the closed-cycle element adjusts the output at all speeds of operation and continues to drive the load at very low speeds after the open-cycle actuator has slipped to a halt.

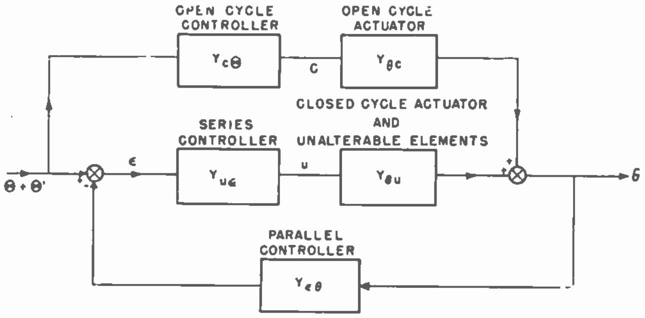

Fig. 7—Completely parallel combination system.

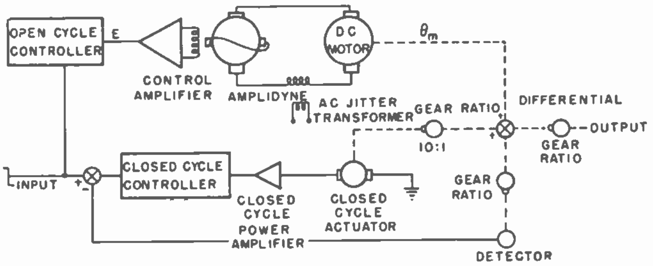

Fig. 8—Increased speed range with parallel open-cycle controller.

Such a system has other advantages besides the mere increase of operating speed range. A major one of these is economic. Thus, the cost of a closed-cycle system for fine control of a large power actuator is proportionally more than that for a small power actuator. Furthermore, for many operations in which a completely closed-cycle system could meet the specifications, the actuator and its controls would have to be of much higher quality than if it were to be part of a combination system. As a result, the closed-cycle vernier can often be added to a relatively cheap open-cycle actuator at much less cost than a single good actuator and closed-cycle system to control it.

As an example, consider the system of Fig. 8. Here the open-cycle controller is considered to be a large dc motor (shown here to be controlled by an amplidyne for purposes of illustration). The speed range of the motor is limited to, say, 50:1 even with jitter and an open-cycle controller.[3] Furthermore, the large motor can be relatively cheap for its power rating, and sluggish to respond, so that the speed and position may be instantaneously in error by as much as 10 per cent of the maximum. We now add a closed-cycle vernier such that ten revolutions of the closed-cycle actuator correspond to one revolution of the open-cycle actuator at the load. This actuator, however, is smaller and can have a much faster response than the open-cycle system actuator. Moreover, the speed range of the closed-cycle system can be made, say, 500:1. Since this 1:500 is a fraction of the 10 per cent error in the open-cycle system, we have succeeded in making the total operating speed range of the system 5,000:1. Furthermore, we have been able to use a high-power system much poorer in performance and therefore cheaper than would have been required to obtain even a 500:1 speed range with a completely closed-cycle system. As a result, the small closed-cycle system plus the large open-cycle system actually costs less than a large closed-cycle system having poorer performance.

Nonlinear Open-Cycle Controllers

A. Basic Theory

Although linearized approximations to actual transfer functions have been analyzed in examples of the previous sections, there is no such basic limitation on the combination open-cycle closed-cycle idea. Thus, if the actuator and unalterable elements have nonlinear characteristics, the partially parallel open-cycle controller must have the inverse characteristics. Referring to Fig. 7, suppose that, for

$$Y_{\theta\Theta}(P) = N_{\theta\Theta}(P)/D_{\theta\Theta}(P),$$
$$F_\theta(D_{\theta\Theta}\theta) = F_u(u) \tag{37}$$

where F_θ and F_u are any physically realizable functions. This requires that

$$F_\theta(N_{\theta\Theta}\Theta) \approx F_u(u) \tag{38}$$

be the equation of the partially parallel open-cycle controller, normally requiring a computer.

B. Example of a Hydraulic-Rate Servo

To illustrate, suppose that we wish to design a speed-controlled ("rate") servo. Here

$$\dot{\theta} \doteqdot K_{\theta\Theta}\Theta. \tag{39}$$

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[3] The reader will readily note that the open-cycle controller might be aided by a tachometer-accelerometer feedback around the large motor. This would correspond to an internal feedback loop in a closed-cycle system, and should be considered part of the open-cycle controller.

If the actuator is hydraulic and controlled by a valve, we consider the actual nonlinearities of the pressure drop across the valve. Referring to Fig. 9, a typical valve-controlled hydraulic ram is shown. Here we take the empirical formula for flow versus pressure

$$K_{\delta\dot{\theta}}\dot{\theta} = \delta(\Delta p)^n \qquad (40)$$

where

$$\frac{\Delta p}{2} = p_1 - p = p' - p_0.$$

The load force is proportional to

$$p - p' = p_1 - p_0 - \Delta p$$

so that, if L is the load force

$$p - p' = K_{p\ddot{\theta}}\ddot{\theta} + K_{pL}L = p_1 - p_0 - \Delta p. \qquad (41)$$

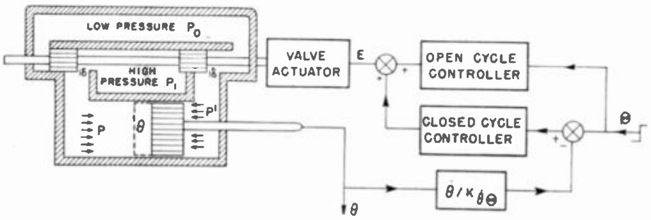

Fig. 9—Nonlinear open-cycle controller with hydraulic actuator.

The valve motion is related to the control voltage by the relation

$$(P^2 + 2\zeta_\delta\omega_\delta P + \omega_\delta^2)\delta = K_{\delta E}E. \qquad (42)$$

We now seek the relation between Θ and E required for the open-cycle controller to make

$$\dot{\theta} = K_{\dot{\theta}\Theta}\Theta. \qquad (43)$$

If we are successful in designing the open-cycle controller (and if (40) is sufficiently accurate), we may replace $\dot{\theta}$ by $K_{\dot{\theta}\Theta}\Theta$ in (40) and (41) to give

The reader will readily recognize that such a controller need not be restricted to open-cycle operation, but may also be used as a "linearizer" in series with the actuators and unalterable elements of a completely closed-cycle system for synthesis by "inversion." However, if used in the open-cycle controller, such a computer will have no effect on system stability.

Of course in any practical hydraulic system, the nonlinearity is masked by an output function feedback around the actuator-valve combination. This introduces problems of inner loop stability or sluggishness, but makes it possible to use a much more approximate open-cycle controller design.

C. Example of an Algebraic Cyclical Control

Where the desired output is approximately relatable to the time or an input variable by an algebraic (nondifferential) relationship, the open-cycle system may consist of a quasi-direct drive from input to output.

An example of such a condition is the flat card winding machine tension controller. Here, if a completely closed-cycle system (as shown in Fig. 10) is used, the tension actuator must be employed more to take up slack than to control tension. As a result, its ability to control tension is greatly impaired by the necessity for accelerating the mass of actuator and wire feed elements before the tension can even be affected. To illustrate this, we briefly analyze the problem of Fig. 10.

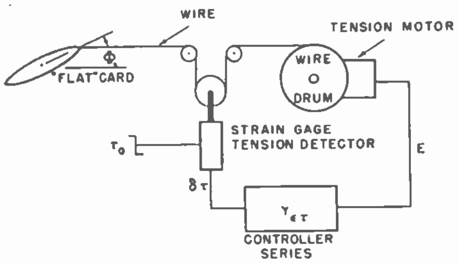

Fig. 10—Closed-cycle tension control system.

$$K_{\delta\theta}K_{\theta\Theta}(P^2 + 2\zeta_\delta\omega_\delta P + \omega_\delta^2)\left[\frac{\Theta}{(p_1 - p_0 - K_{P\ddot{\theta}}P\dot{\Theta} - K_{PL}L)^n}\right] = K_{\delta E}E. \qquad (44)$$

This shows what formula an idealized computer with Θ as an input and E as an output would take. Such a computer would, of course, be only approximate since perfect derivatives can never be mechanized. It would also be much more complicated than warranted for most applications, and is used here only as an example of a nonlinear open-cycle element design in all of which the dependent variable of the equations relating actuator output to input is replaced by the independent variable of the system.

Assuming that the friction and inertial torques of the rollers are negligible, the idealized tension equations are, for a separately excited dc or two-phase ac motor

$$\tau = -K_{\tau E}E + K_{\tau\dot{\theta}}P(1 + T_m P)\theta \qquad (45)$$

$$-E = Y_{E\tau}(\tau_0 - \tau) \qquad (46)$$

where τ_0 is the constant desired tension of the tension regulator.

It follows that

$$\tau = \frac{\tau_0 + K_{\tau\dot{\theta}}P(1 + T_mP)\theta}{1 + K_{\tau E}Y_{E\tau}}. \qquad (47)$$

The tension error is

$$\delta\tau = \tau - \tau_0 = \frac{K_{\tau\dot{\theta}}P(1 + T_mP)\theta - K_{\tau E}Y_{E\tau}\tau_0}{1 + K_{\tau E}Y_{E\tau}}. \qquad (48)$$

This shows $\delta\tau$ to depend on $\dot{\theta}$ and $\ddot{\theta}$. However, θ is determined by the position of the card given by the angle, ϕ. Thus, the tension regulation is evidently being greatly penalized by the full magnitudes of $\dot{\theta}$ and $\ddot{\theta}$ which may really be considered extraneous to the problem of tension control, since they depend on the size, shape and rotation speed of the card rather than on the tension.

Obviously, therefore, if we could minimize $\ddot{\theta}$ and modify E to eliminate the error chargeable to $\dot{\theta}$, the regulator could be made much more satisfactory. To do this, we add the open-cycle elements of Fig. 11. They have as their purpose the taking up of slack and the cyclical adjustment of winding speed so that $\dot{\theta}$ can be made nearly constant at the average rate of laying wire on the card.

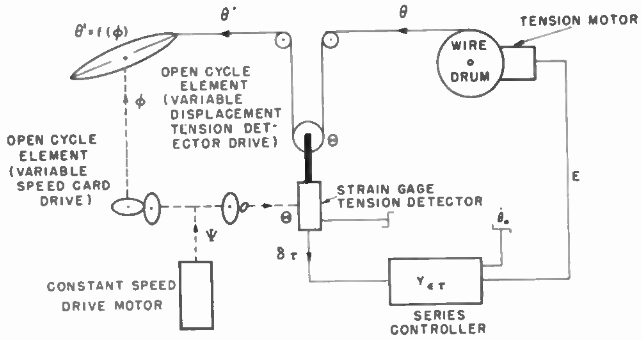

Fig. 11—Combination open-cycle closed-cycle tension regulator

In the particular manifestation of the system shown in Fig. 11, the open cycle elements are a cam drive on the tension detector and a variable speed drive on the card such that, if $\dot{\theta}_0$ is the average speed of the wire drum,

$$2\Theta(\psi) = f[\phi(\psi)] - \dot{\theta}_0t, \qquad (49)$$

where $\Theta(\psi)$ is the cam drive and $\phi(\psi)$ is the card winding variable speed cam gear drive employed, if at all, to make the design of the $\Theta(\psi)$ cam more feasible mechanically.[4] In addition, setting of $\dot{\theta}_0$ sets a voltage E_0 which, for constant velocity conditions, maintains the required current in the motor to maintain τ equal to τ_0 without requiring a strain gage error to do so.

The improvement to be expected from such a system will now be demonstrated by an idealized analysis. We begin by assuming that τ is the tension throughout the whole length of wire from card to drum at any instant. If θ is the angular displacement of the wire drum with slack take-up operating, Θ is the displacement of the slack take-up device, and θ' is the displacement which the drum would have without a slack take-up, we have

$$\theta = \theta' - 2\Theta. \qquad (50)$$

It follows, of course, that ideally we have

$$\dot{\theta} \underset{\cdot}{=} \dot{\theta}_0$$

so that

$$\ddot{\theta} \underset{\cdot}{=} 0.$$

If all of our idealized conditions can be met, there is no need for tension control other than calibration of the motor torque with supply current. However, in any practical case, imperfections will creep in all along the line and, indeed, it is the fact that a very considerable gain can be demonstrated despite these imperfections that makes the use of a combination open-cycle closed-cycle system desirable for this application.

Let the sum total imperfections be expressed as an equivalent error in $f[\phi(\psi)]$ so that substitution of (49) into (50) gives

$$\theta = \dot{\theta}_0t - \epsilon \qquad (51)$$

where

$$\epsilon = f[\phi(\psi)] - \theta'[\phi(\psi)]. \qquad (52)$$

We now add the last open-cycle element by making

$$E = \frac{K_{\tau\dot{\theta}}\dot{\theta}_0}{K_{\tau E}} + Y_{E\tau}(\tau_0 - \tau). \qquad (53)$$

Substituting (53) into (45) gives

$$\tau = \frac{\tau_0 + K_{\tau\dot{\theta}}P(1 + T_mP)\epsilon}{1 + K_{\tau E}Y_{E\tau}} \qquad (54)$$

so that

$$\delta\tau = \frac{K_{\tau\dot{\theta}}P(1 + T_mP)\epsilon - K_{\tau E}Y_{E\tau}\tau_0}{1 + K_{\tau E}Y_{E\tau}}. \qquad (55)$$

Comparing (55) with (48), it is observed that the error now involves ϵ where θ occurred before. Of course, $Y_{E\tau}$ can be selected to make the steady state error involving τ_0 negligible.

Evidently the tension error has been reduced by adding the open-cycle elements, an amount depending on the time variations of ϵ compared with the time variations of θ (in the system of Fig. 11). This permits an increase in winding speed which can greatly augment the output of the winding machine.

The reader will recognize that the specific example chosen does not represent the only means of accomplishing the desired objective of improved tension control. Thus there are immediately called to mind several improvements. In some of these, greater flexibility for use with cards of different shapes is attained by replacing

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[4] Without such a cam gear drive, the required variations in speed of the $\Theta(\psi)$ cam follower would be prohibitive.

the cams by computer-operated servos, providing adjustment of the functions ϕ and Θ to correspond to variations in card depth. Indeed, the best mechanical design is obtained by putting the tension adjustment at Θ and the slack take-up as a ψ-controlled variation in angular displacement of the wire drum. However, the example has been given, not to describe an improved tension control, but rather to illustrate advantages of the addition of completely parallel, algebraic open-cycle elements to an otherwise closed-cycle system.

The same idea may obviously be applied anywhere that an approximate duty cycle can be completely established for the output of an automatic control system, requiring, of course, that the significant rates of the difference between actual and approximate duty cycles are suitably less than the corresponding rates of the true desired output. It is also useful for partially parallel open-cycle systems. Here the actuator voltage is continually adjusted by an algebraic open-cycle element in accordance with the predictable portion of the duty cycle, and corrected for the unpredictable portion by the closed-cycle vernier.

MULTIPLE VARIABLE SYSTEMS

The previous sections have demonstrated the use of combination open-cycle closed-cycle systems in a variety of linear and nonlinear single-input single-output combinations. However, many of the more advanced problems of precision automatic control system design involve systems with several inputs and several outputs. Examples include multiple variable process controls, automatic pilots, blind landing equipment, and gimbal system servos (to name only a few).

As with the two variable systems, series, partially parallel, completely parallel open-cycle controllers, or combinations thereof, may be employed. Here, however, the interaction is often nonlinear and the requirement for computers in the open-cycle elements is increased.

Suppose that a physical system contains n inputs E_i and m outputs θ_j related by the m equations

$$F_j(E_1, E_2, \cdots E_n, \theta_1, \theta_2, \cdots \theta_m) = 0. \qquad (56)$$

Suppose, further, that it is desired to design a control system making use of the physical system by controlling its inputs or their effects in such a way that

$$\theta_j \doteq \theta_j(\Theta_1, \Theta_2 \cdots \Theta_n) \qquad (57)$$

where the Θ_j are now considered the inputs of the whole automatic control system.

In a completely closed-cycle system, the scheme would be to control m of the E_i by errors

$$\mathcal{E}_j = \Theta_j - (\theta_j')^{-1}[\theta_1, \theta_2, \cdots, \theta_m], \qquad (58)$$

and the major problem of controller design would consist of finding a suitable set of relations among the $\mathcal{E}_j$ and E_j. The $(n-m)E_i$, which do not enter into the control, constitute unwanted inputs (disturbances or "noises") and their effects must be minimized.

In a partially parallel combination open-cycle closed-cycle system, the problem is to relate Θ_j to the m desired unalterable element inputs E_j. This requires that a computer be designed such that

$$\overline{F}_j(E_1', E_2', \cdots, E_m', E_{m+1}, \cdots, E_n, \theta_1', \theta_2', \cdots, \theta_m') = 0 \qquad (59)$$

where the $\overline{F}_j$ represent practical approximations to the F_j. Here the θ_j' and $n-m$ of the E_i are considered controller inputs, and m of the E_i are considered controller outputs. The E_j' are added to the outputs E_k'' of a (by now greatly simplified) closed-cycle controller to establish the inputs E_j to the unalterable elements. Of course, it follows that

$$E_k'' = E_k''(\mathcal{E}_1, \mathcal{E}_2, \cdots \mathcal{E}_n). \qquad (60)$$

However, the E_k'' can each be made to depend primarily on one $\mathcal{E}_j$, since the open-cycle controller has insured that all $\mathcal{E}_j$ are much smaller than would be possible with a completely closed-cycle system.

Fig. 12 illustrates the system described by the foregoing equations. Here, however, we draw the diagram to represent one-way matrices defined by taking all similar quantities as a group. Thus letters without subscripts represent the matrices and

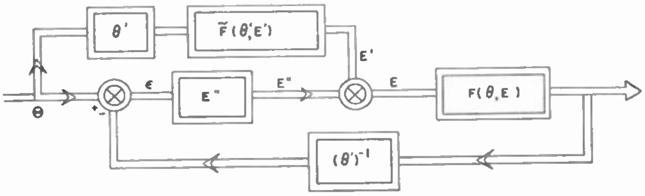

Fig. 12—Partially parallel open-cycle controllers in multiple variable systems.

$E \equiv E_1, E_2, \cdots, E_n$	(unalterable element input matrix)
$E' \equiv E_1', E_2', \cdots, E_m'$	(open-cycle controller output matrix)
$\theta \equiv \theta_1, \theta_2, \cdots, \theta_m$	(actual output matrix)
$\theta' \equiv \theta_1', \theta_2', \cdots, \theta_m'$	(desired output matrix)
$F \equiv F_1, F_2, \cdots, F_m$	(unalterable element relations)
$\overline{F} \equiv F_1', F_2', \cdots, F_m'$	(open-cycle controller)
$\Theta \equiv \Theta_1, \Theta_2, \cdots, \Theta_m$	(desired input matrix)
$E'' \equiv E_1'', E_2'', \cdots, E_m''$	(closed-cycle series controller)
$\mathcal{E} \equiv \mathcal{E}_1, \mathcal{E}_2, \cdots, \mathcal{E}_m$	(closed-cycle controller inputs).

Such partially parallel open-cycle controllers are particularly useful in precision autopilot designs and systems involving appreciable transport lag. Again, the job may often be performed better by a series open-cycle element or a combination of series and partially parallel elements.

CONCLUSIONS

Although there are many applications where the addition of open-cycle elements carries the design beyond the point of diminishing returns (or, in some cases of bad input noise, may actually make the system performance poorer), the requirements for better and better automatic control systems reveal many other situations in which the use of a combination open-cycle closed-cycle system can be expected to "save the day." Thus, it is believed that the concepts and synthesis techniques involving combination open-cycle closed-cycle systems, although certainly no panacea, represent a powerful addition to the repertory of the precision automatic control system designer.

NOTATION AND NOMENCLATURE

Θ = system "desired" or "true" input
Θ' = input disturbance or "noise"
θ = system output
$\mathcal{E}$ = system error [difference between actual noise free) input and that function of output over which control is desired]
$\mathcal{E}_n$ = noise error
Φ = load disturbance—flux
$N_{\alpha\beta}(P)D_{\alpha\beta}(P)$ = numerator and denominator polynomials in P
$Y_{\alpha\beta}(P)$ = operator applied to element input, β, to get element output, α, usually a rational fraction
E = actuator input or control voltage
T_i, τ_i = system parameters
τ = wire tension
$K_{\alpha\beta}$ = constant
K = per unit gain uncertainty or variation
$P = d/dt$ (also taken as the complex variable of Laplace transforms)
p = hydraulic pressure
c_i = reciprocal error coefficient of the ith derivative
ω = input circular frequency
$\doteqdot$ = desired equal to
$\cong$ = equal over a satisfactory range
y = motor field flux.

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CORRECTION

Abd El-Samie Mostafa, author of the paper, "Electron-Tube Performance with Large Applied Voltages," which appeared on pages 70–73 of the January, 1951, issue of the PROCEEDINGS OF THE I.R.E., has brought the following errors to the attention of the editors:

Page 71—Column 2, line 4
sin instead of rin

Equations (5), (7), and (8)
$-a_0E_c$ instead of $-a_0(E_c+B)$

Page 72—Column 2, line 1
$-E_c$ instead of E_c

Add to caption of Fig. 4
$E_b = 300$ volts, $E_c = 100$ volts, and $V_p = 130$ volts (R.M.S.)

The author's date of birth is April 27, 1913 instead of 1917.

Transient Response of a Narrow-Band Automatic Frequency-Control System*

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Summary—A method of analyzing the response of automatic frequency-control systems, operating in conjunction with a narrow band-pass filter, is presented in this paper. A time-lag, equivalent to the reciprocal of the bandwidth, is assigned to the filter. Circuit parameters are established for operation near the critically damped response condition. The equations of the system are derived with the aid of the Laplace transform. The method of residues is used to evaluate the transient response to a step-input frequency disturbance.

INTRODUCTION

A RECENT CIRCUIT PROBLEM required an automatic frequency-control system to operate in conjunction with a very narrow band-pass filter. The presence of such a filter introduces a time lag which complicates the behavior of the system, and makes its transient response difficult to predict. In a recent paper, van der Wyck[1] considers the dynamics of an afc system, and outlines conditions for a nonoscillatory response as derived from an unpublished work of de Cock Buning. However, the presentation is very brief, and no attempt is made to calculate the actual response time of the system.

This paper will present a method of analysis for narrow-band afc systems based on the Laplace transform. Conditions for a damped nonoscillatory transient response will be derived, and the response time of a system subjected to a step-input frequency disturbance will be calculated.

DESCRIPTION OF THE AFC SYSTEM

A block diagram of the afc system appears in Fig. 1. A portion of the output signal is amplified, limited, and applied to a frequency discriminator which is balanced for zero voltage output at the center frequency of the band-pass filter. A frequency variation in the output signal produces a discriminator output voltage which, acting through the integrating circuit and reactance tube, corrects the local oscillator frequency so that the output frequency is maintained within the band-pass region of the filter. The insertion of the integrating network permits adjustment of the transient response of the system.[2] The circuit parameters must be selected so that the steady-state output frequency deviation of the filter (resulting from an input frequency disturbance) will always remain in the band-pass region of the filter. This selection also must be consistent with any specified transient response.

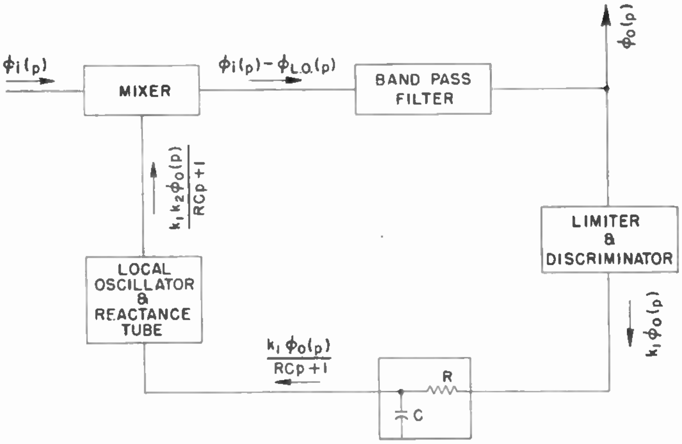

Fig. 1—Block diagram of the afc system. (The transforms of the variables are indicated as functions of P.)

ASSUMPTIONS AND DEFINITIONS

It will be assumed that at time $t=0$, an input frequency variation disturbs a previously attained steady-state condition in which the output frequency exactly coincided with the center frequency of the band-pass filter. The input disturbance will be designated as $\phi_i(t)$, and the resulting output frequency deviation as $\phi_0(t)$. The narrow band-pass filter will introduce a time delay τ, which is approximately equivalent to the reciprocal of its band-width at the one-half power points. It will also be assumed that the discriminator and reactance-tube responses are linearly proportional to frequency and voltage, respectively. From these assumptions the following definitions may be made:

$\phi_i(p)$ = Laplace transform of input frequency disturbance

$\phi_0(p)$ = Laplace transform of output frequency deviation

$\phi_{L.O.}(p)$ = Laplace transform of local oscillator frequency variation

$\phi_m(p)$ = Laplace transform of output of mixer

$e^{-\tau p}$ = Transfer function of band-pass filter[3]

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[1] van der Wyck, "Netherlands PTT single-sideband equipment," PROC. I.R.E., vol. 36, pp. 970–980; August, 1948.

[2] Other kinds of control circuits, such as integral plus proportional control, may sometimes be employed advantageously. For an excellent discussion of this aspect of the automatic control problem, see Hall, "Analysis and Synthesis of Linear Servomechanisms," Technology Press, MIT, Cambridge, Mass.; May, 1943.

[3] The transfer function of a network is defined as the ratio of the Laplace transform of the output signal to the Laplace transform of the input signal; the Laplace transform of $F(t)$ is defined as $f(p)=\int_0^\infty F(t)e^{-pt}dt$. Because of amplitude limiting, the afc system is insensitive to amplitude variations, and therefore it is not necessary to include the amplitude versus frequency characteristics of the filter in the transfer function for the filter.

$\frac{1}{RCp+1}$ = Transfer function of integrating network

k_1 = Sensitivity of discriminator in volts per cycle

k_2 = Sensitivity of reactance tube and local oscillator in cycles per volt.

$k = k_1 k_2$ = "Loop gain" of system.

SYSTEM EQUATIONS

By referring back to Fig. 1, in which the transforms of the frequency deviations are designated at various parts of the system, and recalling the definition of transfer function, the following algebraic equations may be formulated:

$$\begin{aligned}
\phi_m(p) &= \phi_i(p) - \phi_{L.O.}(p)\\
\phi_0(p) &= \phi_m(p)e^{-p\tau}\\
\phi_0(p) &= [\phi_i(p) - \phi_{L.O.}(p)]e^{-p\tau}\\
\phi_{L.O.}(p) &= \phi_0(p)\frac{k_1k_2}{RCp+1}.
\end{aligned}$$

Solving for $\phi_0(p)$ and writing K for k_1k_2,

$$\phi_0(p) = \phi_i(p)\left[\frac{e^{-p\tau}}{1+\dfrac{Ke^{-p\tau}}{RCp+1}}\right]. \qquad (1)$$

Equation (1) relates the output frequency deviation of the system to any input frequency disturbance. To obtain an equivalent equation in the real time domain, some functional form for $\phi_i(p)$ is specified, and $\phi_0(p)$ may then be transformed back into the time domain by means of the inverse transform.[4] Thus, if at $t=0$ a step-input frequency variation of M cps is assumed, $\phi_0(p)$ becomes M/p and $\phi_0(t)$ may be written as

$$\phi_0(t) = \underset{\beta_1\rightarrow\infty}{\frac{1}{2\pi j}}\int_{\alpha_1-j\beta_1}^{\alpha_1+j\beta_1}\frac{M}{p}\left(\frac{e^{-p\tau}}{1+\dfrac{Ke^{-p\tau}}{RCp+1}}\right)e^{pt}dp, \qquad (2)$$

where the path of integration includes all poles in the integrand.

ESTABLISHING THE TRANSIENT RESPONSE CONDITIONS

It is convenient to write (2) as

$$\phi_0(t) = \underset{\beta_1\rightarrow\infty}{\frac{1}{2\pi j}}\int_{\alpha_1-j\beta_1}^{\alpha_1+j\beta_1}\frac{M(RCp+1)e^{p(t-\tau)}}{p(RCp+1+Ke^{-p\tau})}dp. \qquad (3)$$

The form of the transient response may be predicted from a knowledge of the roots of the denominator in (3). To find these roots let

$$p(RCp+1+Ke^{-p\tau}) = 0.$$

One root equals zero. The other roots must satisfy the following:

$$RCp + 1 + Ke^{-p\tau} = 0. \qquad (4)$$

In general, p will be a complex number involving a real and an imaginary part. Positive real roots indicate an unstable response; negative real roots, a stable response degenerating as time increases. An imaginary root implies an oscillating response. If a nonoscillatory stable response is desired, it is necessary to establish circuit conditions which will either reduce the imaginary part to zero, or convert it to a real number. Let

$$p = \alpha + j\beta.$$

Substitute into (4):

$$RC\alpha + jRC\beta + 1 + Ke^{-\alpha\tau}(\cos\beta\tau - j\sin\beta\tau) = 0.$$

Separate real and imaginary terms, thus:

$$\cos\beta\tau = \frac{-(RC\alpha+1)}{Ke^{-\alpha\tau}} \qquad (5)$$

and

$$\sin\beta\tau = \frac{RC\beta}{Ke^{-\alpha\tau}}. \qquad (6)$$

Square, add (5) and (6), and solve for β, thus:

$$\beta = \sqrt{\frac{K^2e^{-2\alpha\tau} - (RC\alpha+1)^2}{(RC)^2}}. \qquad (7)$$

To eliminate the imaginary roots of (4), β itself must be imaginary. Therefore, to yield a nonoscillatory response, the system parameters must satisfy the following inequality:

$$Ke^{-\alpha\tau} \leqq (RC\alpha + 1). \qquad (8)$$

In principle, α can be determined by graphical solution of the above transcendental equations. However, such procedure is extremely tedious and time-consuming. Usually it is sufficient to solve the system for the critically damped condition for which β equals zero.[5] However, it may sometimes be desirable to operate the system with a very low-frequency, oscillating damped response, or sometimes with a slightly over-damped response. The former condition corresponds to a small positive number for β, the latter condition to a small imaginary number for β.

From (5) and (6),

$$\frac{\tan\beta\tau}{\beta\tau} = \frac{-RC}{(RC\alpha+1)\tau}. \qquad (9)$$

In practice, τ will usually be much less than unity, and $\tan\beta\tau/\beta\tau$ may be replaced approximately by unity, and (9) solved for α if β is limited to small values. Thus

$$\alpha = -\left(\frac{\tau+RC}{RC\tau}\right). \qquad (10)$$

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[4] The inverse transform $F(t)$, of $f(p)$ is defined as

$$F(t) = \frac{1}{2\pi i}\int_{\alpha_1-i\beta}^{\alpha_1+i\beta_1} f(p)e^{pt}dt.$$

Alternatively, a table of transform pairs which relates $f(p)$ with its corresponding $F(t)$, may sometimes be employed.

[5] The term "critically damped" is used here to indicate the transition from an aperiodic response to an oscillatory response.

Equation (10) indicates that near critical damping, the roots of (4) will be real and negative in character, and hence the system response to a step-input disturbance will be stable. From (8),

$$K\exp\frac{\tau + RC}{RC} \leqq \frac{\tau + RC}{\tau} - 1.$$

Usually RC will be much greater than τ so that the nonoscillatory transient response condition may finally be formulated as[6]

$$Ke \leqq \frac{RC}{\tau}\cdot$$

By setting $\beta =$ to zero in (7), the condition for the critically damped response may be established as

$$RC = \tau K\exp\frac{\tau + RC}{RC}\cdot \qquad (11)$$

CALCULATION OF $\phi_0(t)$

The steady-state frequency deviation resulting from a step-input frequency disturbance of M cps may be determined by evaluating the integral of (3) at the simple pole $p=0$. Let $\rho_1, \rho_2, \rho_3 \cdots \rho_m$ equal the roots of the quantity $(RCp+1+Ke^{-p\tau})$. Then, by the method of residues, $\phi_0(t)$ can be represented symbolically as

$$\phi_0(t)=\frac{M}{1+K}+M\sum_1^m \left.\frac{(RCp+1)e^{p(t-\tau)}}{\frac{d}{dp}(RCp+1+Ke^{-p\tau})}\right]_{p=\rho_1,\rho_2\cdots\rho_m} .$$

Imposing the restriction that the steady-state deviation must be less than half the bandwidth of the filter yields the following condition on K:

$$\frac{M}{1+K} \leqq \frac{BW}{2},$$

where $BW =$ the bandwidth at half-power points. Since it has been assumed that $BW = 1/\tau$, this inequality will be satisfied if

$$1 + K > 2\tau M. \qquad (12)$$

The complete time response, including both the steady-state and transient deviations, may be determined by evaluating the integral of (2) at all its poles. This evaluation is most readily performed by first expanding the integrand into a power series.[8] Thus

$$\phi_0(t) = \underset{\beta_1 \to \infty}{\frac{M}{2\pi i}} \int_{\alpha_1 - j\beta_1}^{\alpha_1 + j\beta_1} \left[\frac{e^{-p\tau}}{p} - \frac{K}{RC}\,\frac{e^{-2p\tau}}{p\left(p+\frac{1}{RC}\right)} \right.$$
$$+ \frac{K^2}{RC^2}\,\frac{e^{-3p\tau}}{p\left(p+\frac{1}{RC}\right)^2}$$
$$\left. - \frac{K^3}{RC^3}\,\frac{e^{-4p\tau}}{p\left(p+\frac{1}{RC}\right)^3} + \cdots \right] e^{pt}dp.$$

Integrating each term by the method of residues, and applying the time restrictions involved in the translation theorem of operational calculus,[9]

$$\begin{aligned}
\phi_0(t) = M &= 0 \quad \text{for} \quad t < \tau \\
&- MK\left[1 - e^{-(t-2\tau/RC)}\right] = 0 \quad \text{for} \quad t < 2\tau \\
&+ \frac{MK^2}{\overline{RC}^2}\left[\overline{RC}^2 - e^{-(t-3\tau/RC)}\left\{RC(t-3\tau) + \overline{RC}^2\right\}\right] = 0 \quad \text{for} \quad t < 3\tau \\
&- \frac{MK^3}{\overline{RC}^3}\left[\overline{RC}^3 - e^{-(t-4\tau/RC)}\left\{\frac{RC}{2}(t-4\tau)^2 + (t-4\tau)\overline{RC}^2 + \overline{RC}^3\right\}\right] = 0 \quad \text{for} \quad t < 4\tau. \qquad (13) \\
&+ \cdots .
\end{aligned}$$

Since for a stable response the roots cannot assume positive real values, the steady-state deviation may be determined by allowing t to approach infinity.[7] Thus

$$\underset{t\to\infty}{\phi_0(t)} = \frac{M}{1+K}\cdot$$

The number of terms which must be computed will depend on the rate of convergence of $\phi_0(t)$ to its final steady-state value. Calculations on practical afc systems (where τ is much less than one) indicate that usually only four or five terms will be required for an accurate description of the transient response.

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[6] This condition is in agreement with that given by van der Wyck in the paper previously cited in footnote reference 1.

[7] The summation term in the equation for $\phi_0(t)$ is valid only for simple roots. For higher order roots, other techniques must be used to evaluate the residues. However, since only real negative roots are involved, $\phi_0(t)$ will always equal $M/1+K$ as t approaches infinity.

[8] This kind of expansion for evaluating residues is suggested in the following paper: L. A. Pipes, "The analysis of retarded control systems," *Jour. Appl. Phys.*, vol. 18, pp. 617–623; July, 1948. This paper considers servomechanisms with a time lag in the feedback loop.

[9] Churchill, "Modern Operational Mathematics in Engineering," p. 21, McGraw-Hill Book Co., New York, N. Y.; 1944.

Time Response of a Typical System

As a practical illustration of the foregoing principles, consider an afc system having a filter bandwidth of only 10 cycles at the half-power points, and a bandwidth of 200 cycles at the 60 db points. Assume that proper adjustment of the limiter and discriminator circuits will allow the system to correct for a maximum step-input frequency disturbance of 100 cycles. Setting $\tau = 0.1$ second and $M = 100$ cps, K (as determined from (12)) must equal at least 19 if the steady-state deviation is not to exceed 5 cycles. Let $K = 20$. For the critically damped transient response, the time constant of the integrating circuit must satisfy (11). Graphical solution of this transcendental equation yields a value of 5.539 seconds for RC.

The complete response time of the system may now be calculated from the infinite series solution of $\phi_0(t)$. The solid line curve, designated as $K = 20$ in Fig. 2, represents the critically damped response; the dashed line, the step-input disturbance. It is interesting to note that a time lag of τ seconds occurs before the output frequency responds to the input disturbance, and that another time lag of τ seconds must occur before correction of the output frequency is experienced.

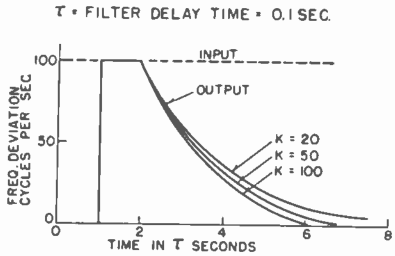

Fig. 2—The critically damped transient response to a step-input frequency disturbance for three values of "loop gain."

Conclusion

A method of calculating the transient response of a narrow-band afc system has been presented. Although the analysis was based on a step-input disturbance, any other kind of disturbance may be inserted into (2), and the resulting output-frequency deviation computed from a new series solution for $\phi_0(t)$.

Acknowledgments

Thanks are due P. Conley and C. E. Vogeley of the Westinghouse Research Laboratories, the former for helpful discussions during preparation of this paper, and the latter for proofreading the manuscript.

The Folded Fan as a Broad-Band Antenna*

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***Summary*—High-frequency model studies are reported which promise an antenna which operates over a 4-to-1 frequency range with a maximum power loss, due to mismatch, of 18.5 per cent. In the projected naval application, this loss is not unusual with currently used antenna systems. The antenna consists of an ordinary fan-type antenna folded back to ground, that is, electrically connected at the top to an identical fan grounded at its base.**

A means is described for matching the antenna to 52-ohm transmission line, although the folded structures center at about 160 ohms. The model described in detail is not necessarily of absolutely optimum proportions.

The maximum variation with azimuth of the radiation intensity of the antenna is less than 10 db. In the vertical plane, the antenna illuminates the horizon adequately, wasting no appreciable portion of the energy on the zenith.

Introduction

THE ANTENNA development described here was undertaken in an effort to solve partially the antenna problem aboard naval ships. In the past, the decks and superstructures of such vessels have become increasingly cluttered in a most confusing manner, with antennas of all types and sizes. In general, approximately one antenna is installed for each equipment: transmitter, receiver, or transceiver. As the most frequently used antennas in the communication band often have very poor impedance characteristics with several sharp resonances within the band, and deficient radiation patterns, often with severe nulls in the azimuth coverage, communication efficiency has not always been up to a practical minimum.

One mode of resolution of these difficulties proposes the wide use of multiplexing techniques. Each antenna would serve several units, either transmitters or receivers. We will concern ourselves with the transmitting problem. Multiplexing devices for the transmitting case generally require a broad-band antenna for a load.

It is further desired to use coaxial cable in place of the more bulky and bulkhead-weakening trunk, which leads from the transmitter to the antenna. The high standing-wave ratios and consequent power loss, or voltage breakdown and dielectric failure that would result with present antenna systems, prohibit the use of coaxial cable.

It was proposed that a broad-band type radiating structure to be fed by 52-ohm coaxial cable be developed for use in the 2- to 27-mc communication band.

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MODEL STUDIES

Impedance Measurements

A study of the impedance characteristics of various folded structures was carried out using ultra-high-frequency models. A scaling factor of 50 was used for much of the work, making the band of interest spread from 100 to 1,350 mc. Where other scaling factors were used, the data have been translated into terms of this frequency band.

The measurement equipment consisted of the following. A 20-foot-square simulated ground plane situated on the side of a building was used in conjunction with a Chipman measuring line.[1] The antenna models were excited by a small loop projecting above the surface of the ground plane and fed by a well-shielded transmitter. At the lower end of the frequency band, the General Radio model 757-A signal generator was used. At the high end, a local oscillator from an AN/SPR-2 receiver was used. The signal picked up by the sampling loop at the movable short circuit of the Chipman line was compared with the signal supplied from the Navy Type LAE-2 or LAF-3, and their associated direct-reading attenuators. The standing-wave ratio could be obtained directly in decibels from these attenuators, or the steepness of the current maxima could be obtained and translated to radius on the Smith chart. The AN/APR-1 receiver and its several radio-frequency heads were used in conjunction with an external microammeter. Except at the very lowest frequencies, where the receiver calibration was used, the frequencies were determined with the use of a series of transmission type cavity wavemeters.

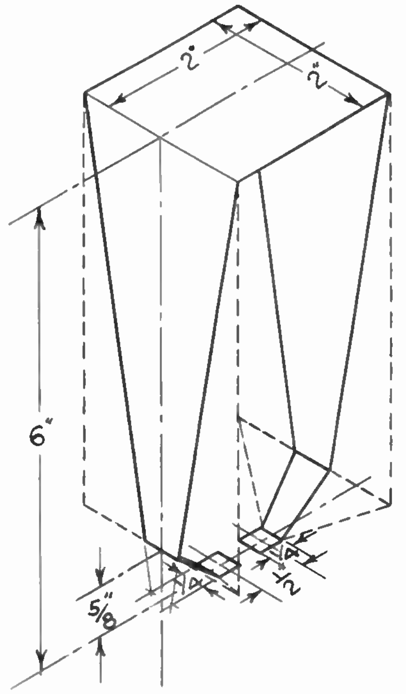

Fig. 1—Fat folded monopole.

The final design of antenna was not to be too large for installation on naval ships. A maximum height of 25 feet was selected as a nominal value to shoot for. This scaled to 6 inches. An early model investigated in these studies is illustrated in Fig. 1. One tab at the base was screwed to the inner conductor of the Chipman line; the other was soldered to the ground plane. It was desired to make the basic elements of the folded structure as broad-band as possible by themselves. The first model measured conformed to the dotted outline shown. Curve *B* in Fig. 3(a) represents the impedance characteristic of this model. The improvement noted in curve *A* was effected after trimming the model to conform to the solid lines in Fig. 1. Two effects are apparent here. First, trimming away the sides of the structure reduced the effective base capacitance, thus allowing the entire impedance characteristic to shift toward the inductive side of the chart. Second, the losses through radiation in the transmission-line mode were increased through a reduction in self-shielding, and the fatness in the monopole mode was reduced, bringing the *Q*'s of the two modes more nearly into balance.

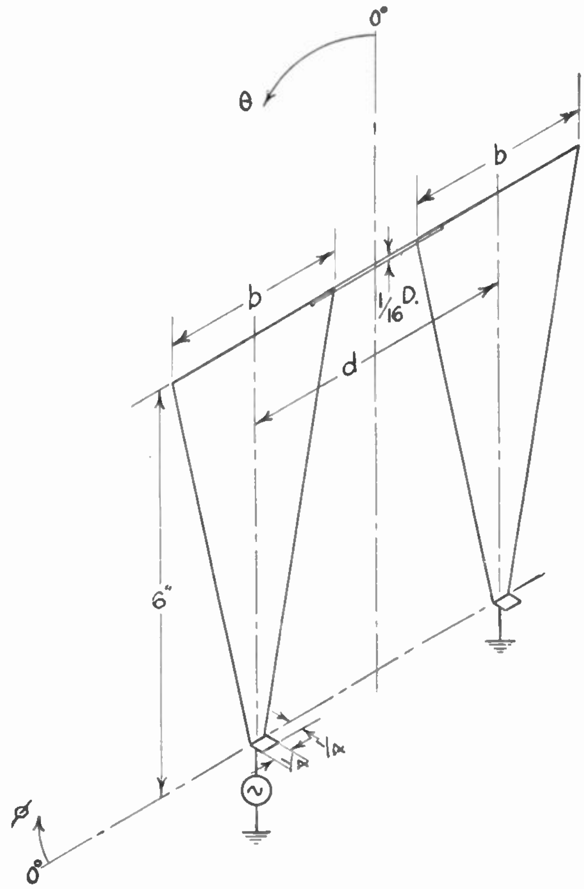

Fig. 2—Folded-fan antenna.

The effects of this trimming operation gave the clue to the folded-fan construction used in the remainder of the studies. In this device (see Fig. 2) the transmission

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[1] J. T. Bolljahn: "Chipman Line Analysis and Development," Technical Report 132, Antenna Laboratory, University of California, Berkeley, Calif.; October 31, 1947.

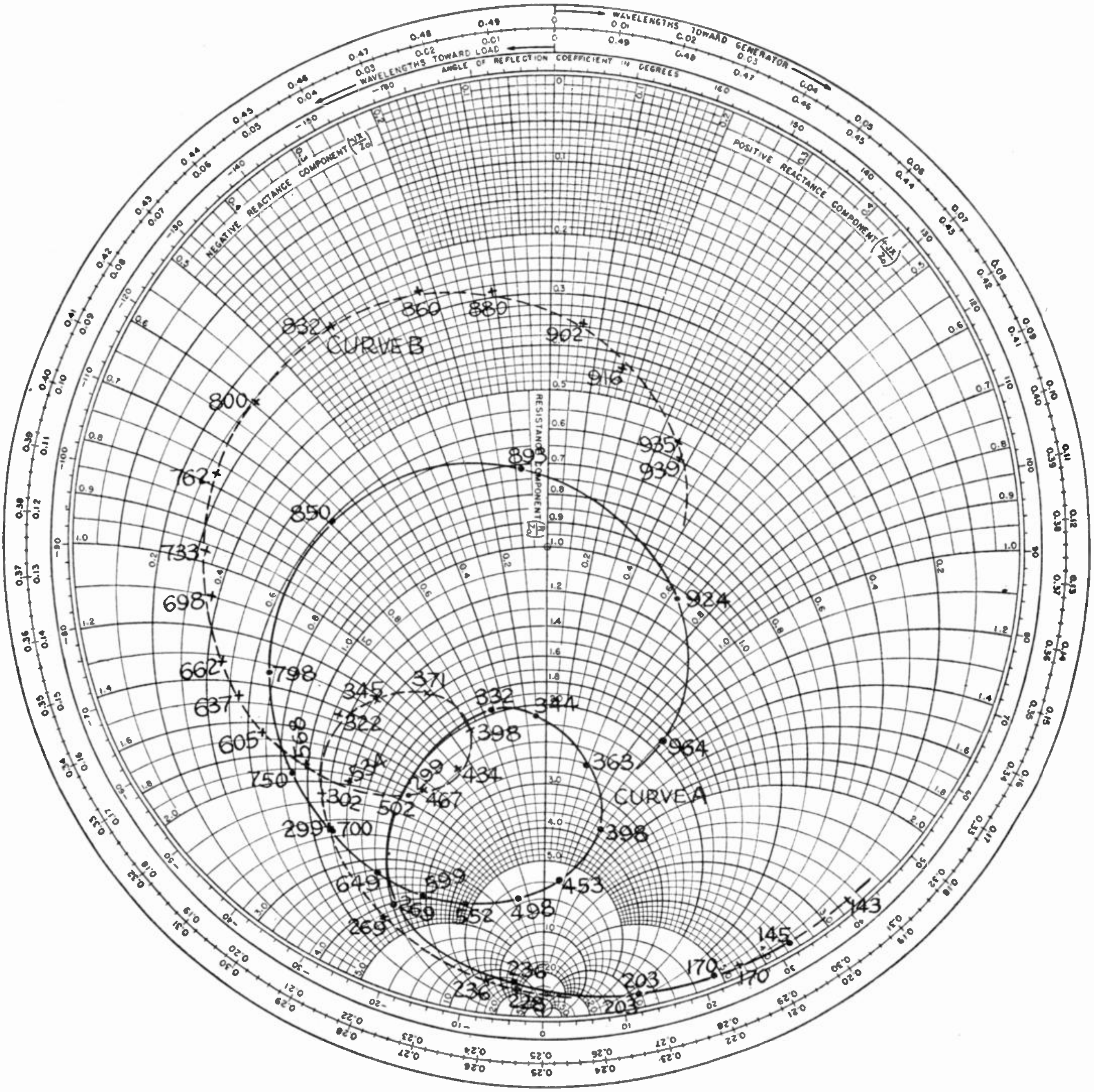

Fig. 3(a)—Smith chart plots of impedance referred to 52 ohms versus frequency in mc. (Chart from Philip H. Smith, "Transmission line calculator," *Electronics*, vol. 12, January, 1939; also vol. 17, p. 130, January, 1944.)

line has been opened up to the maximum extent possible, the entire structure lying in a plane. At the same time, the monopole is made fat by the use of the fan structure. A systematic survey of the impedance characteristics of this antenna for a wide array of different proportions was conducted. The dimensions *d* and *b* are identified in Fig. 2. Dimension *d* was varied from 2 to 6 inches, and *b*, from 1 to 5 inches. Some of the results are presented in Figs. 3(b) and 3(c). Fig. 3(b) shows the effect on the impedance locus of pulling two identical fans apart. Fig. 3(c) shows the effect of broadening the fans held at a constant distance on centers. In the latter, two isofrequency lines are drawn to show the manner in which the loci draw in toward a restricted region of the chart as the fans are enlarged.

Transmission-Line Matching

The impedance characteristic of Figs. 3(b) and 3(c) are compact but at too high an impedance level to be fed by 52-ohm cable. It was desired to match a representative model to 52 ohms by some convenient, practical method. The model of curve *C*, Fig. 3(c), was se-

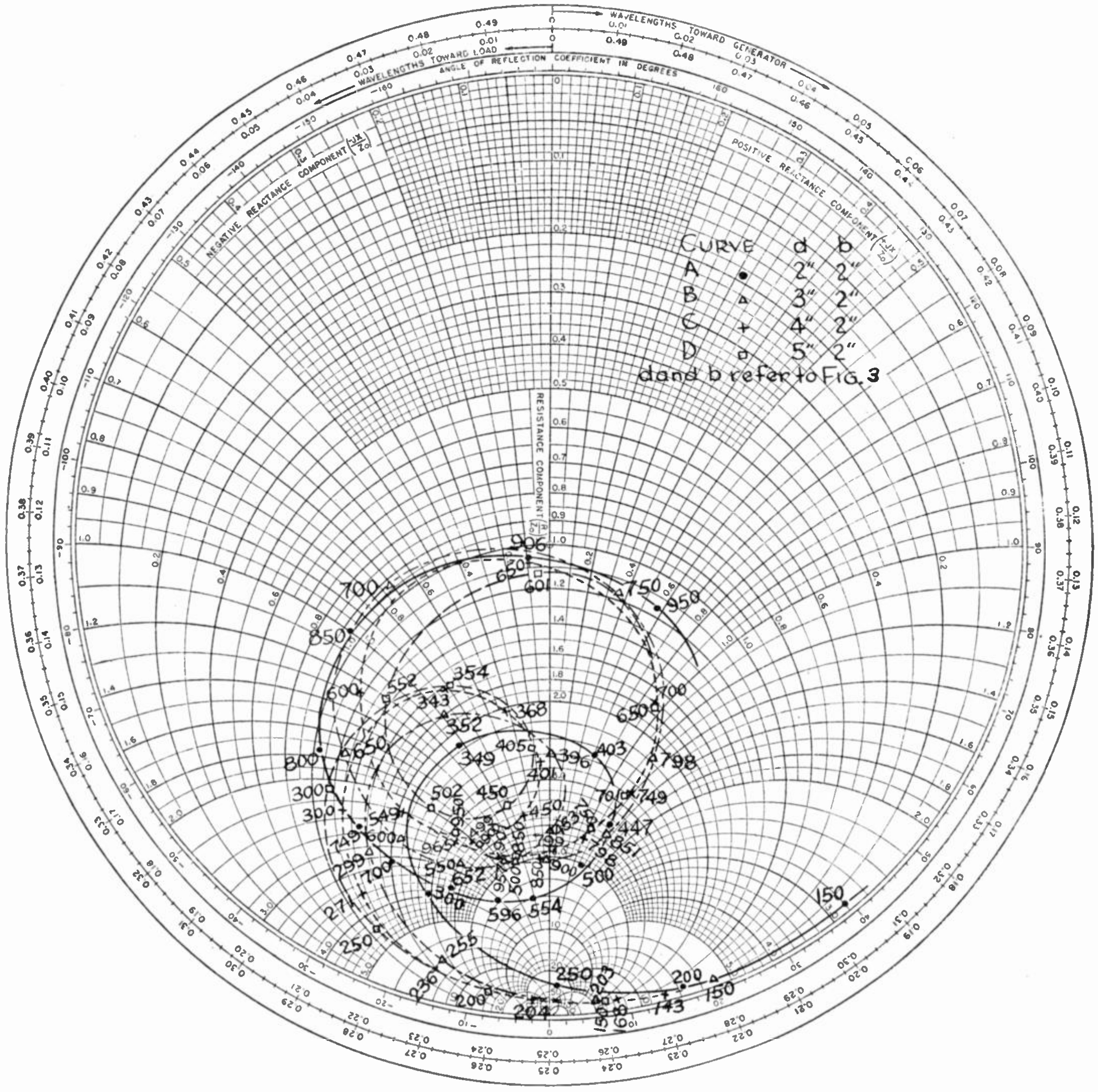

Fig. 3(b)—Smith chart plots of impedance referred to 52 ohms versus frequency in mc. B of Fig. 2 equal to 2 inches.

lected as having good balance of monopole and transmission-line Q's, and as having nominal dimensions. By a method already described[2] three sections of transmission line were selected to simulate an exponential taper. Values of characteristic impedance were selected which were available commercially, should a full-scale test be desired at a future time. The results of the analytical transformation were checked experimentally by means of an extention to the Chipman line.

The inner conductor was stepped to produce the 125-ohm and the 93-ohm sections. The line itself has an impedance of 67.5 ohms. The results were transformed analytically through a portion of the line itself and then referred to 52 ohms. The data in final form appear in Fig. 3(d). To obtain information in the higher end of the frequency range of interest, a smaller scaling factor was used (35), but the results are referred to the 6-inch height for uniformity of presentation.

It is noted that the VSWR of this antenna and matching system is within 2.5 from about 285 to about

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[2] R. L. Linton, Jr., "Design charts for transmission line matching systems," *Tele-Tech.*, vol. 9, p. 19; January, 1950.

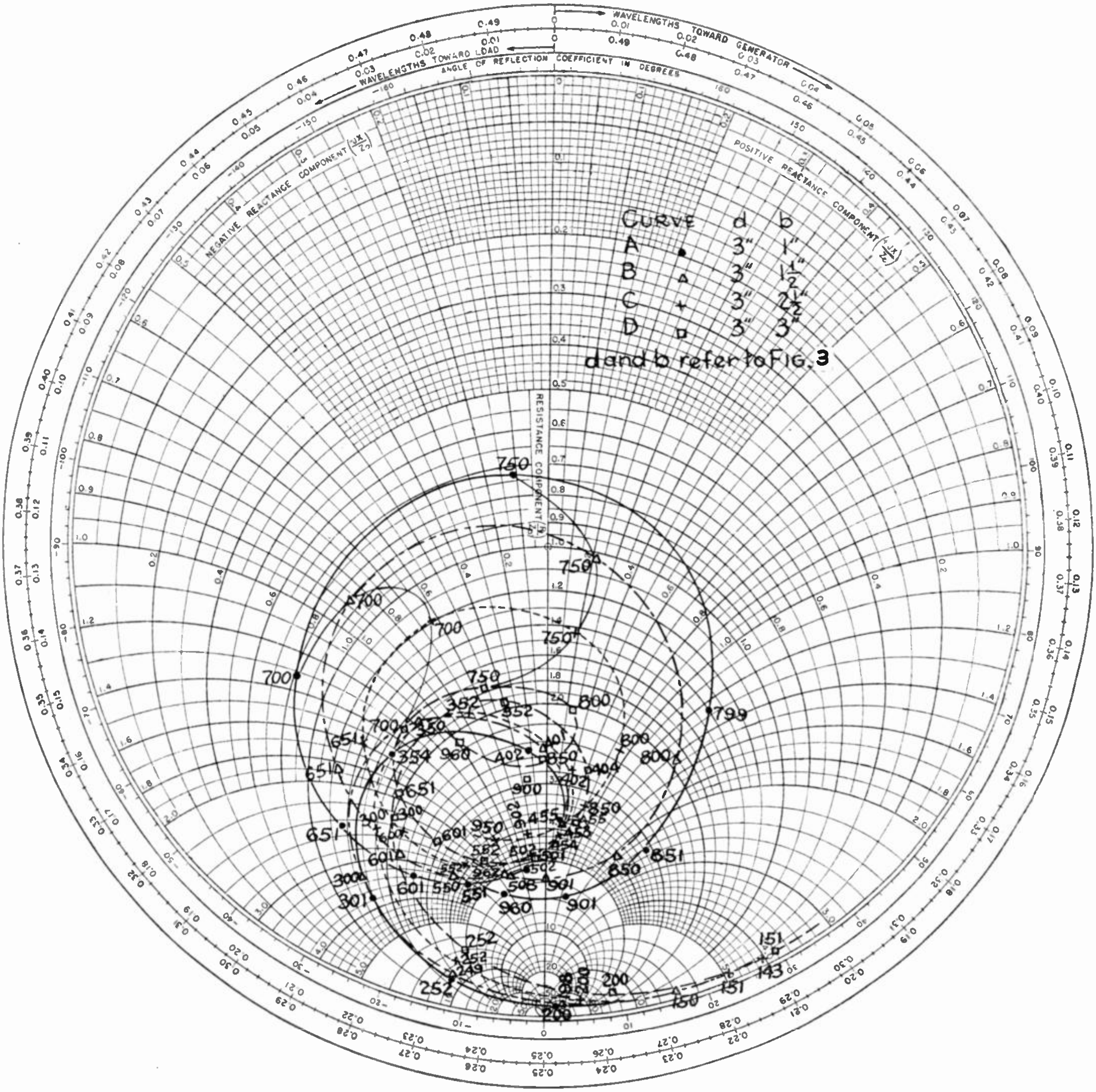


Fig. 3(c)—Smith chart plots of impedance referred to 52 ohms versus frequency mc. D of Fig. 2 equal to 3 inches.

1,160 mc, and within 4 from 265 to probably at least 1,350 mc.

Radiation Patterns

The radiation patterns of the model antenna in the horizontal and vertical planes were checked for acceptability for the projected application. For this purpose, a horizontal ground plane about 20 feet square with a turntable in the center was used. A lighthouse tube transmitter fed a one-inch stub antenna at the "focus" of a corner reflector which illuminated the ground plane. For the vertical pattern measurements, the model antenna and the corner reflector antenna were removed from the ground plane by about 10 feet. A Hoffman pattern recorder was used in conjunction with the AN/APR-1 receiver. The conventions for azimuth ϕ and zenith angle θ are indicated in Fig. 2. The patterns have been transcribed from the Hoffman pattern records and normalized to the outer edge of the charts.

Horizontal Patterns: The horizontal patterns are presented in Fig. 4. The ratio of maximum to minimum intensity throughout the useful band is well below 10 db.

As a rough check on the efficiency of the folded fan, the level of the pattern was compared, by direct substi-

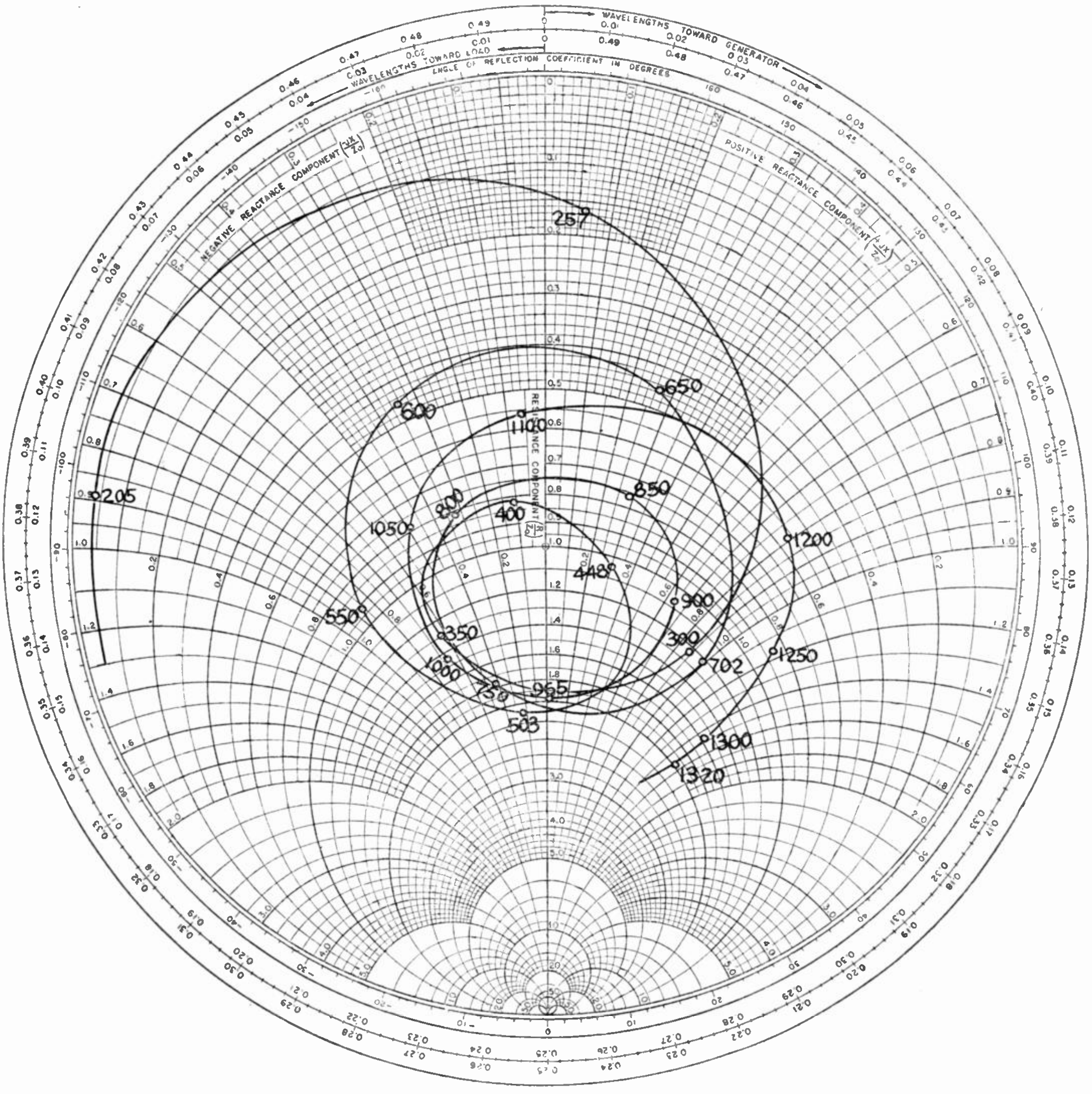


Fig. 3(d)—Smith chart plot of impedance referred to 52 ohms versus frequency in mc. Impedance transformed electrically through stepped matching transmission line. Characteristic impedance respectively 125 ohms, 93 ohms, and 67.5 ohms. Model 8 9/16 inches high. Frequencies for equivalent 6-inch antenna given.

tution, with that of a 1/16-inch diameter monopole of the same height at several frequencies. It is noted that the efficiencies of the two antennas are not radically different throughout the frequency range.

Vertical Patterns: For taking vertical radiation patterns, a balanced folded-fan dipole was constructed of the same proportions as those of the model previously selected. The balanced model was scaled to a factor of 74, but the frequencies given in Fig. 5 correspond to equivalent frequencies for the 6-inch unbalanced model. The model was mounted on a vertical phenolic rod and wooden tower. Patterns are given for three frequencies near the top of the band of interest in two vertical planes: the plane of the antenna and that perpendicular to it.

360-degree patterns are given so that the degree of symmetry displayed will indicate how well balanced the model was at the time the pattern was taken. The balance was adjusted by positioning the shorting bar of a balun incorporated at the feed of the balanced

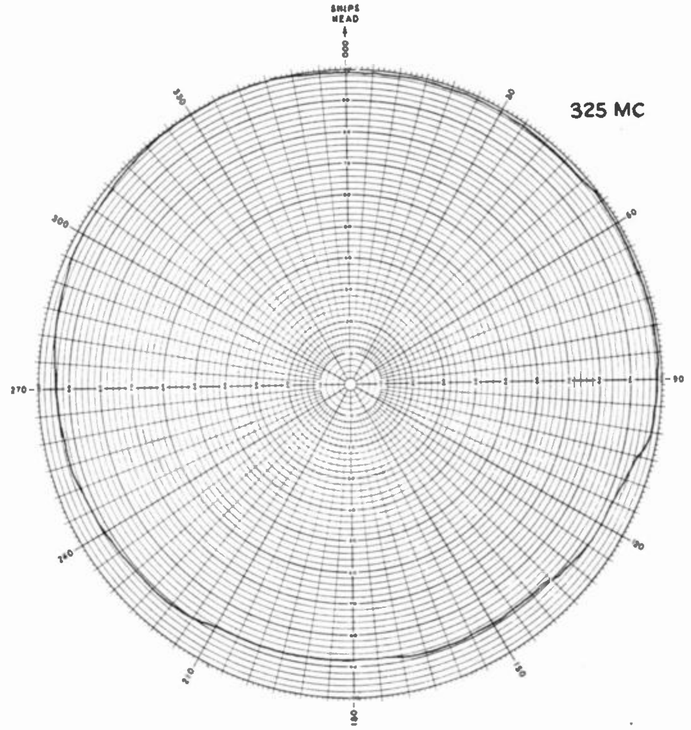


Fig. 4(a)—Horizontal radiation pattern of folded fan at 325 mc.

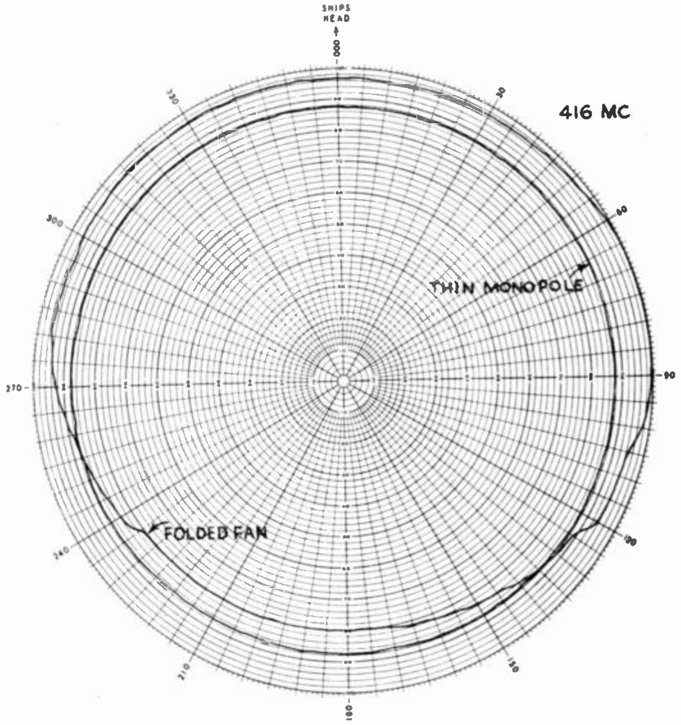


Fig. 4(b)—Horizonnal raniation pattern of folded fan and comparison monopole at 416 mc.

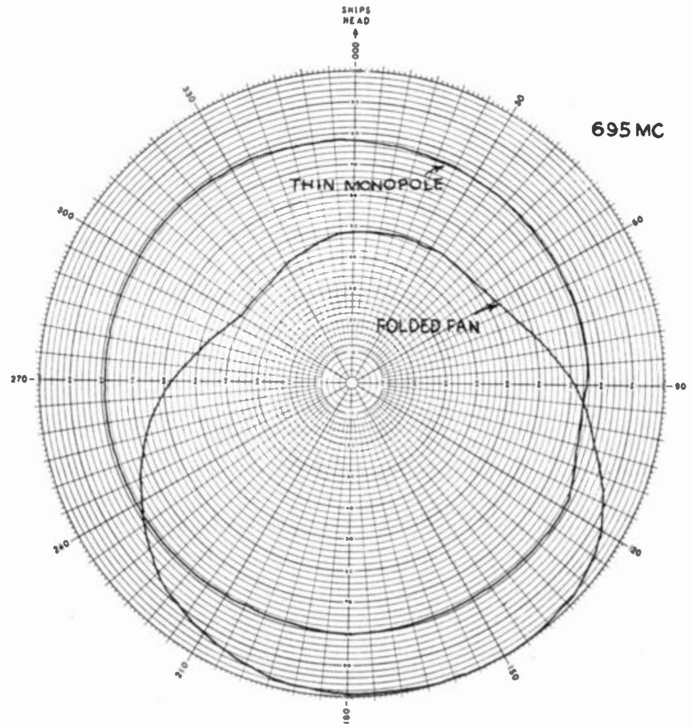


Fig. 4(c)—Horizontal radiation patterns of folded fan and comparison monopole at 695 mc.

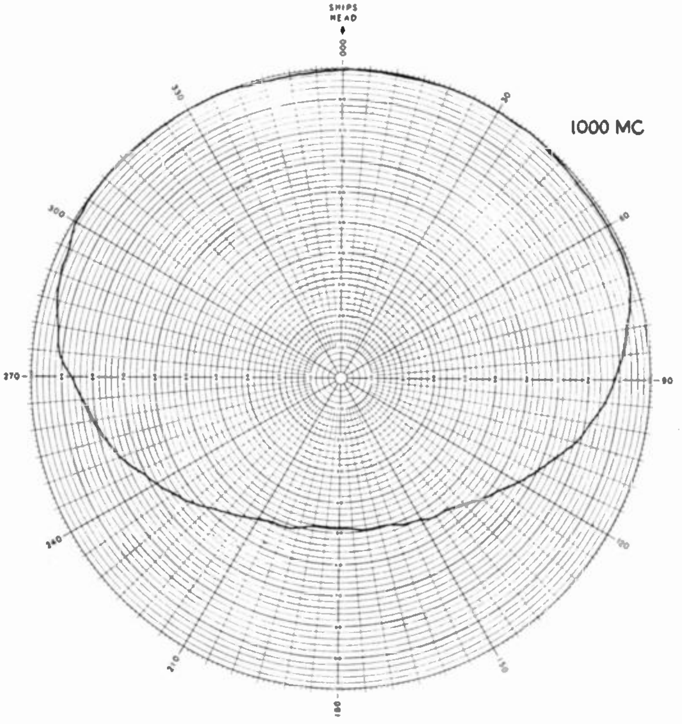


Fig. 4(d)—Horizonnal radiation pattern at folded fan at 1,000 mc.

model. In all cases, the horizon receives a substantial illumination, and no major portion of the energy is diverted toward the zenith. In the figures, the zenith angle is shifted slightly from that indicated by the axis of symmetry due to inadvertent error in placing the graph sheets in the recorder. Time was not available for taking patterns at lower simulated frequencies; a still higher scaling factor would be needed to avoid the effects of reflections from the ground plane.

CONCLUSIONS

The results given here indicate further investigations that might prove fruitful. The impedance and radiation characteristics described should be checked in a full-scale model. Further model studies should be conducted to determine the feasibility of substituting available ship structures for the grounded leg of the folded fan. Many naval vessels possess structures of sufficient height. Where structures are larger than 25 feet, the

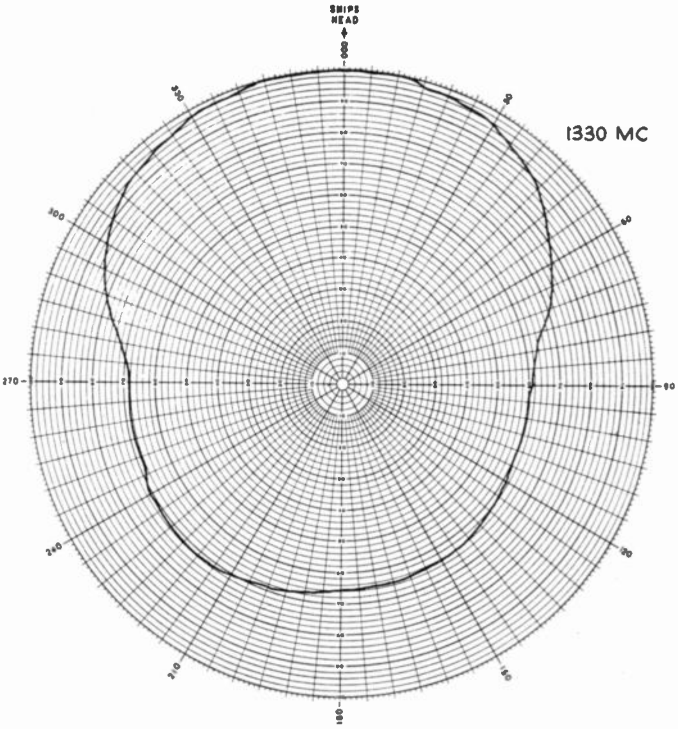


Fig. 4(e)—Horizontal radiation pattern of folded fan at 1,330 mc.

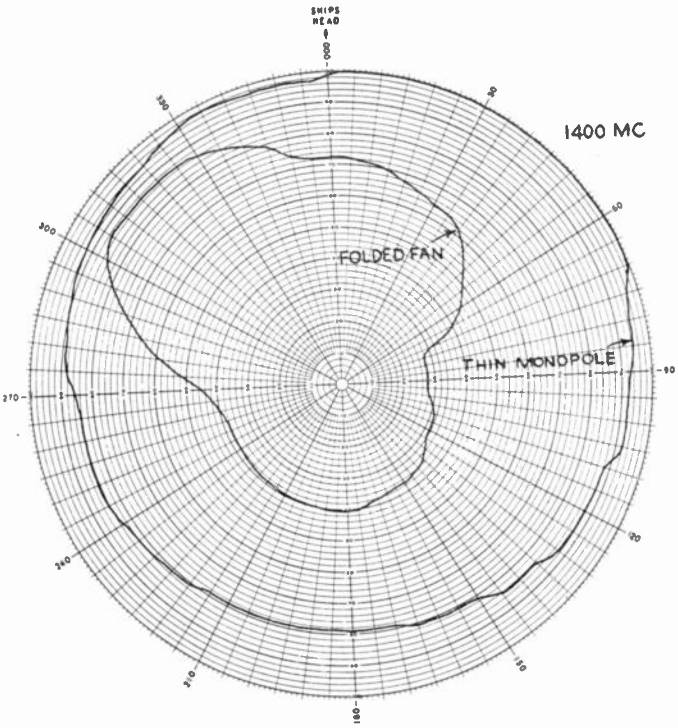


Fig. 4(f)—Horizontal radiation patterns of folded fan and comparison monopole at 1,400 mc.

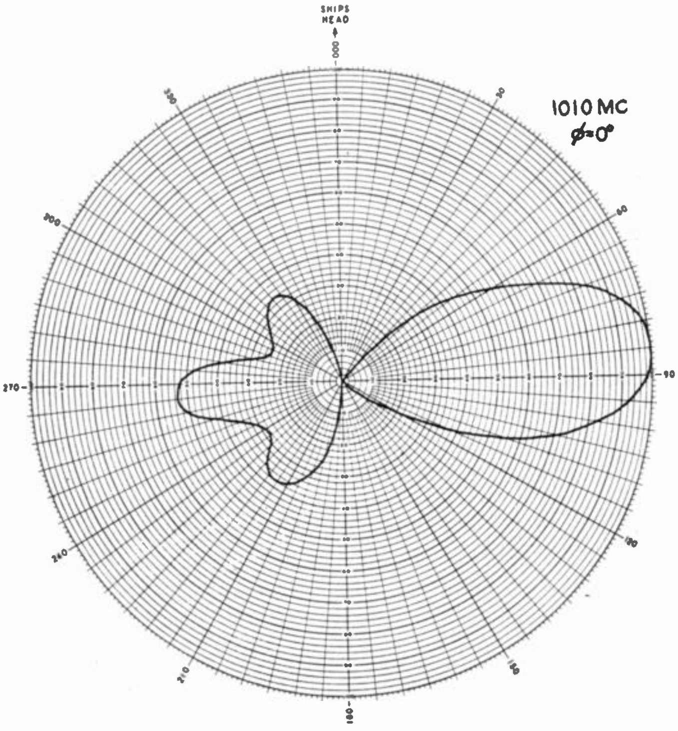


Fig. 5(a)—Vertical radiaton pattern of folded fan at equivalent frequency of 1,010 mc. Azimuth 0°.

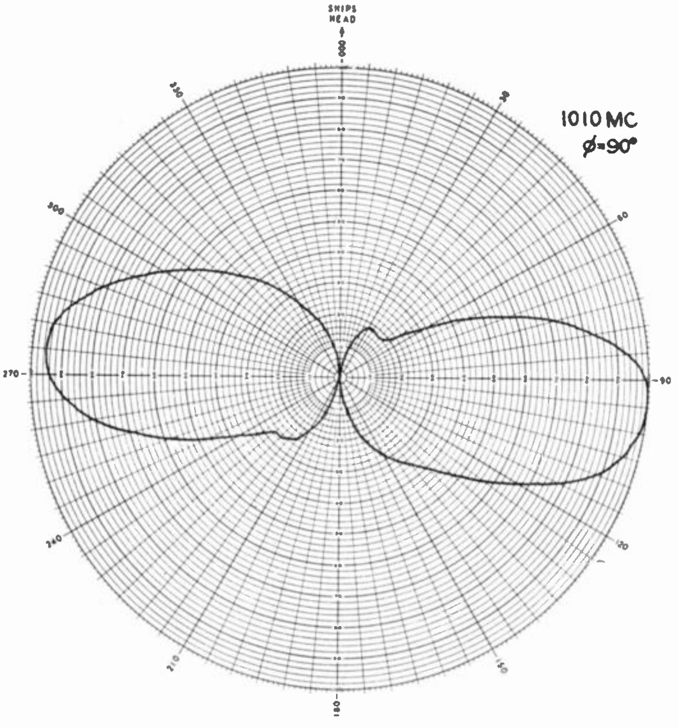


Fig. 5(b)—Vertical radiation pattern of folded fan at equivalent frequency of 1,010 mc. Azimuth 90°.

4-to-1 or 5-to-1 frequency range may be shifted to include the lower communication frequencies. For example, the stack of the *Iowa* class is three times as high as the full-scale antenna assumed in these studies. A folded fan of this height would have a 4-to-1 frequency range between 1.9 and 7.7 mc.

The present antenna seems to have possible application where a broad-band antenna is needed, either in the original projected application, or in other communication applications. At full-scale frequencies, the antenna would have a VSWR within 2.5 from 5.7 to 23.2 mc. Such a mismatch involves a power loss of only 18.5 per cent, which is not a serious loss compared to current naval practice in many instances. From 5.3 to 27 mc, the VSWR would be within 4, a power loss of 36 per cent. The matching cable needed to achieve these results

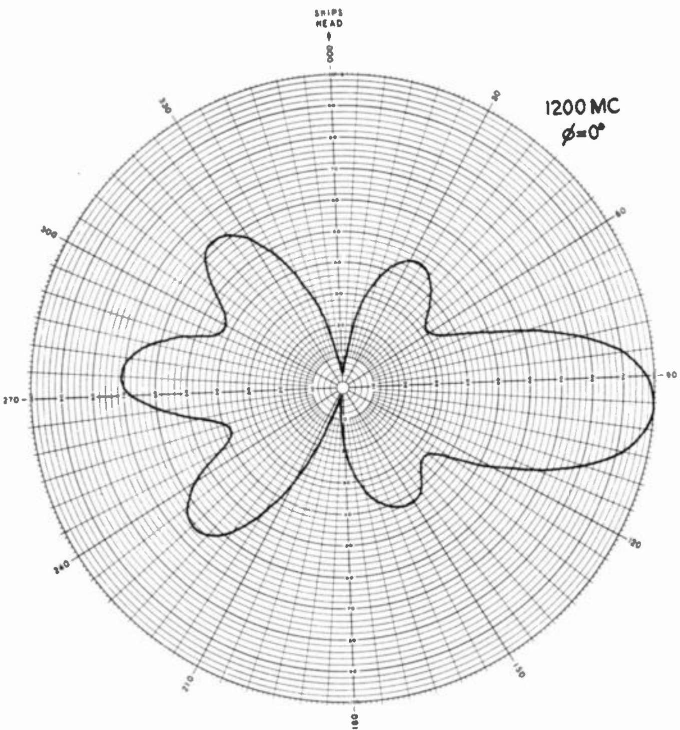


Fig. 5(c)—Vertical radiation pattern of folded fan at equivalent frequency of 1,200 mc. Azimuth 0°.

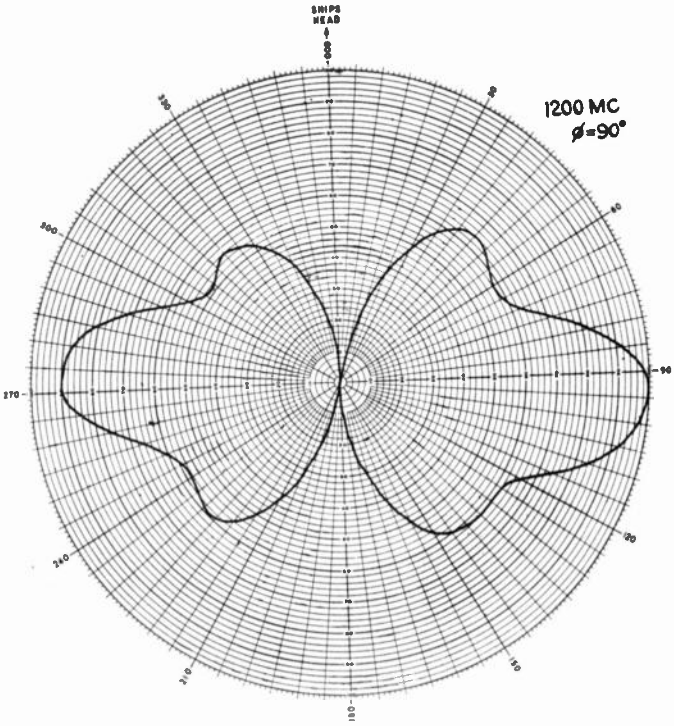


Fig. 5(d)—Vertical radiation pattern of folded fan at equivalent frequency of 1,200 mc. Azimuth 90°.

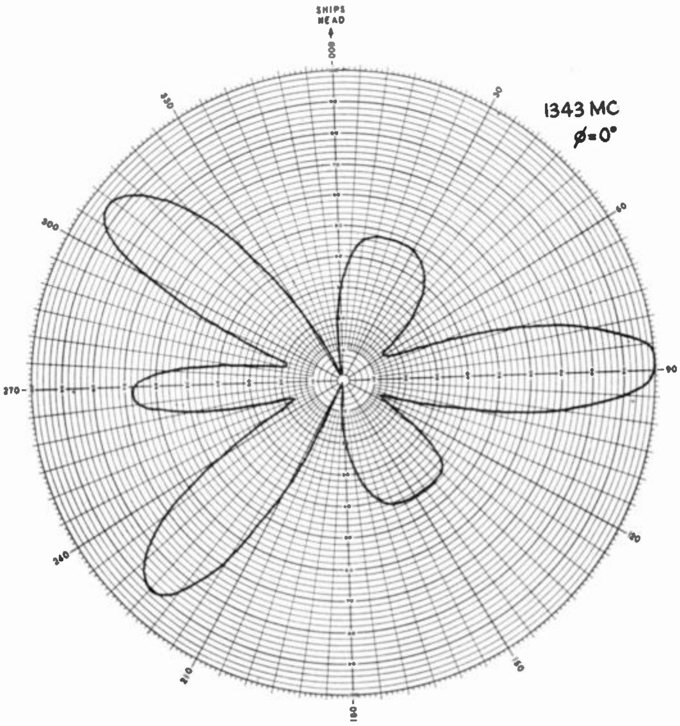


Fig. 5(e)—Vertical radiation pattern of folded fan at equivalent frequency of 1,343 mc. Azimuth 0°.

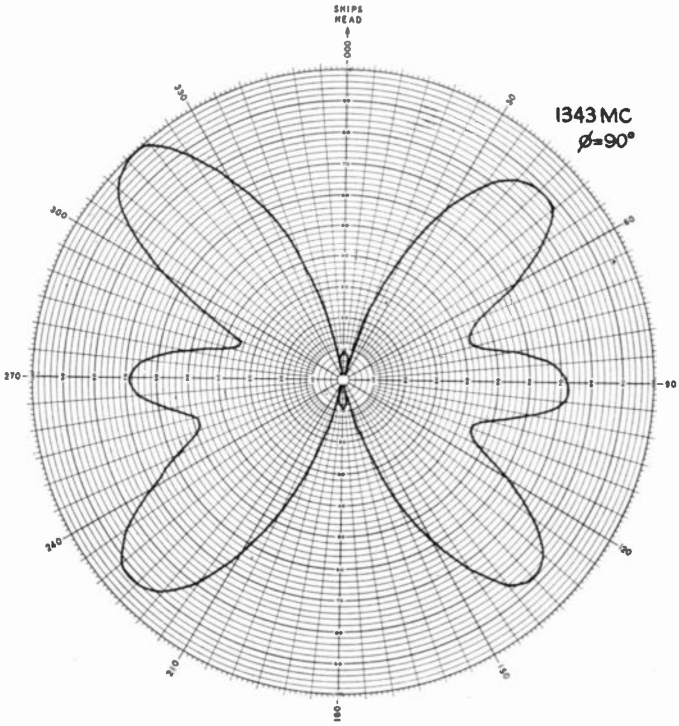


Fig. 5(f)—Vertical radiation pattern of folded fan at equivalent frequency of 1,343 mc. Azimuth 90°.

would be only 49.2 feet long, assuming air dielectric. The antenna would be 25 feet high, 23 feet in horizontal length, and of negligible width in the other dimension. The fans should be simulated with three or four wires. It should be pointed out that the particular proportions selected for detailed description are not necessarily optimum, and that even better results than those obtained may be possible with the folded-fan construction.

The balanced folded-fan dipole at ultra-high frequencies may have promise as a combined FM and television-receiving antenna.

ACKNOWLEDGMENT

The author acknowledges deep gratitude to Robert deLiban who directly supervised the work reported, and offered many helpful suggestions and much encouragement.

The Use of Complementary Slots in Aircraft Antenna Impedance Measurements*

J. T. BOLLJAHN†, ASSOCIATE, IRE, AND J. V. N. GRANGER†, SENIOR MEMBER, IRE

Summary—**This paper describes a method for eliminating the feed-cable effect in the measurement of aircraft wing-cap and tail-cap antenna impedances with the aid of models. The procedure employed allows greater accuracy in measurement than that obtainable with conventional techniques, but its application is restricted to use use on simplified models. The advantages and limitations of the method are discussed, and typical experimental results are presented.**

I. Introduction

MODELING techniques have been used successfully for a number of years in the measurement of antenna characteristics. These techniques are particularly well suited for measurements on antennas which are mounted on complex structures such as ships or aircraft, where adequate full-scale measurements are costly and difficult to accomplish.

Although the principles of modeling apply to both radiation patterns and impedance characteristics, their use in impedance measurements has been limited considerably by practical difficulties. In measuring the impedance of a model aircraft antenna in or below the frequency range where strong resonances may occur in the aircraft structure, for example, one must either use a model large enough to contain the measuring equipment, or make the measurements through a length of cable. The first alternative is impractical in most cases because of the model sizes required, while the second alternative leads to errors due both to the inaccuracy involved in transforming the measured impedance through a length of cable to determine the terminal impedance, and to the perturbation of the fields outside the model caused by the presence of the cable.

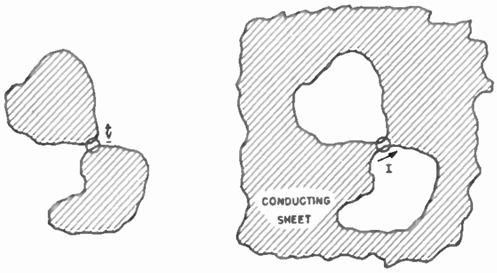
Measurements in this frequency range are of interest at present in connection with studies of wing-cap and tail-cap antennas. Since these antennas are formed by separating and insulating extremities of the aircraft from the main structure, it is seen that a systematic study of their impedance properties on full-scale aircraft would be completely impractical, and a consideration of new modeling techniques is in order.

This report describes a technique suitable for such work which employs a simplified aircraft model consisting of two or more strip conductors lying in a plane and arranged to simulate as closely as possible the shape of the aircraft. Impedance measurements are actually made on a slot which is complementary to this system of plane conductors and which is cut in a large conducting sheet. The impedance of a required configuration in the simplified aircraft model may be calculated from the corresponding measured value for the complementary case with the aid of Babinet's Principle.

Use of the slot analog rather than the model itself makes it possible to embed the feed cable in the ground plane, and thus minimize its effect on the external fields. In some cases the symmetry is such that an image plane may be used, and the feed-cable effect may be eliminated entirely by locating the measuring equipment behind the image plane. The accuracy of results obtained with this technique is limited, of course, by the approximations inherent in the model employed, but the experimental errors are considerably smaller than those which would occur in direct measurements on conventional models. The results of considerable experience lead to the conclusion that the net result of these two conflicting factors favors the slot technique for over-all accuracy.

II. Babinet's Principle

The basis for the slot analog is the electromagnetic Babinet Principle, and in particular the impedance result first pointed out by Booker.[1] This result relates the impedances seen by the two generators in the two problems pictured in Fig. 1. The configuration in the antenna problem consists of an arbitrary arrangement of plane conductors lying in a common plane and driven at one or more points by generators. In the slot problem, a corresponding arrangement of openings has been cut from an infinitely large, plane, perfectly conducting surface, and generators corresponding to those in the antenna problem are present as indicated. Designating the antenna and slot properties by the subscripts a and s, respectively, we may write the expression of interest as follows:



(a) (b)

Fig. 1—Complementary antennas.

$$Z_aZ_s = \frac{\eta^2}{4} \qquad (1)$$

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* Decimal classification: R221×R326.21. Original manuscript received by the Institute, May 16, 1950; revised manuscript received, February 24, 1951.

† Stanford Research Institute, Stanford, Calif.

[1] H. G. Booker, "Slot aerials and their relation to complementary wire aerials," *Jour. IEE*, part IIIA, vol. 93, pp. 620 ff.; March, 1946.

where

Z_a = antenna impedance, ohms
Z_s = slot impedance, ohms
$\eta=\mu_0/\epsilon_0=120\pi$ ohms.

The more general statement of Babinet's Principle from which this result is developed relates the electromagnetic fields in the two problems. Specifically, it states that the electric field in the antenna problem is of the same form as the magnetic field in the slot problem, and the magnetic field in the antenna problem is of the same form as the electric field in the slot problem. It is necessary to qualify this statement only to the extent that the field directions on one side of the plane containing the conductors must be reversed in one of the problems. The reason for this reversal may be seen by considering the field discontinuities involved. In either problem there must be discontinuities in the normal component of E and the tangential component of H at the conducting portion of the surface, corresponding to the charge and current distributions on the conductors. The reversal of fields on one side of the sheet in one of the problems is necessary to move the region in which field discontinuities occur from one set of conductors in the antenna problem to the complementary set of conductors in the slot problem.

III. Application to Impedance Measurements

As mentioned above, the chief advantage of using the slot analog in impedance measurements lies in the fact that the feed cable from the measuring equipment may be embedded in the conducting sheet, thus virtually eliminating its effect on the external fields. The inaccuracies associated with measuring through a long length of line are still present, however, since the operator and measuring equipment must be located sufficiently far from the slot that their presence does not affect the readings.

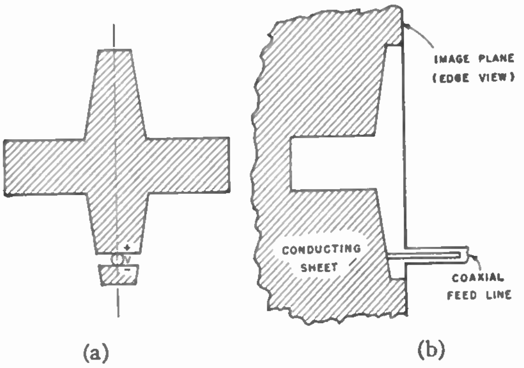

Fig. 2—Simulated wing-cap antenna.

In systems having image symmetry such as the one shown in Fig. 2(a), this difficulty may be avoided by making measurements on a half-slot and image plane as indicated in Fig. 2(b). In this case, a coaxial slotted line may be connected directly to the desired feed point. If the slotted line has a characteristic impedance Z_0, it is readily shown with the aid of (1) that the normalized admittance values of the half slot are the same as the corresponding impedance values for the antenna problem with the latter normalized to a characteristic impedance $Z_0'=\eta^2/8Z_0$.

It is of interest to note that if the antenna problem involves more than one feed point, and if the coupling between the various feedpoints is represented by constructing a suitable equivalent network, then the dual of the antenna network is an equivalent network for the slot problem, provided that the impedance elements in the two networks are normalized to Z_0' and Z_0, respectively.

IV. Simplified Aircraft Models

Since the fields in the slot problem are accurately related to those in the complementary plane-conductor problem by Babinet's Principle, it follows that the basic approximations involved in the equivalent slot technique are simply those involved in using a system of plane conductors to represent the aircraft structure.

The use of simplified models of planar form is based on the concept of equivalent cross sections in two-dimensional static problems. Consider two infinitely long conductors having different cross-sectional shapes. It is always possible, at least theoretically, to adjust the transverse dimensions of one of the conductors while retaining its cross-sectional shape, so that the two conductors have the same static capacitance per unit length. The two conductors adjusted in this fashion to have the same static capacitance per unit length are said to have equivalent cross sections. It is possible to show[2,3] that two linear antennas will have equivalent impedance characteristics if their cross sections are equivalent in this sense, and if their cross-sectional dimensions are small compared with the wavelength.

The application of equivalent cross sections to the construction of planar aircraft models involves the assumption that the approximations inherent in the equivalent cross-section concept remain valid for more complex structures, even when the cross-sectional dimensions of the elements are appreciable fractions of the wavelength. A discussion of these approximations and an experimental investigation of their validity in the case of aircraft structures are given in the literature.[4]

The use of the planar model is further supported by the agreement found in an experimental comparison of the electromagnetic resonance behavior of planar and conventional models of the same aircraft structure.[5] These results serve to resolve a difficulty which arises in the case of the tail-cap structure, namely that the driven

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[2] F. Bloch and M. Hammermesh, "Equivalent Radius of Thin Cylindrical Antennas," Rpt. 411-TM125, Radio Research Laboratory, Harvard University; June, 1944.

[3] C. Flammer, "Equivalent Radii of Thin Cylindrical Antennas with Arbitrary Cross Sections," Tech. Rpt. No. 4, Contract AF 19(122)78. Aircraft Radiation Systems Laboratory, Stanford Research Institute; February, 1950.

[4] J. V. N. Granger and T. Morita, "R-f current distributions on aircraft structures," Proc. I.R.E., to be published.

[5] A. S. Dunbar, "Electromagnetic Resonance Phenomena in Aircraft Structures," Tech. Rpt. No. 8. Contract AF 19(122)78. Aircraft Radiation Systems Laboratory, Stanford Research Institute; May, 1950.

vertical stabilizer does not fit into the planar models. Impedance measurements were made on the slot system complementary to the structure shown in Fig. 3, in which the vertical stabilizer is represented by a coplanar extension of the fuselage beyond the horizontal stabilizer. The resonances apparent in the impedance data shown in this figure correspond to those observed by Dunbar in back-scattering studies of a conventional model of the same airframe. Adoption of the complementary slot model for the tail-cap structure permits the use of image-plane measurements which are not possible when a conventional model is employed. The impedance data of Fig. 3 correspond in their important features to the limited data which are available from measurements on full-scale structures.

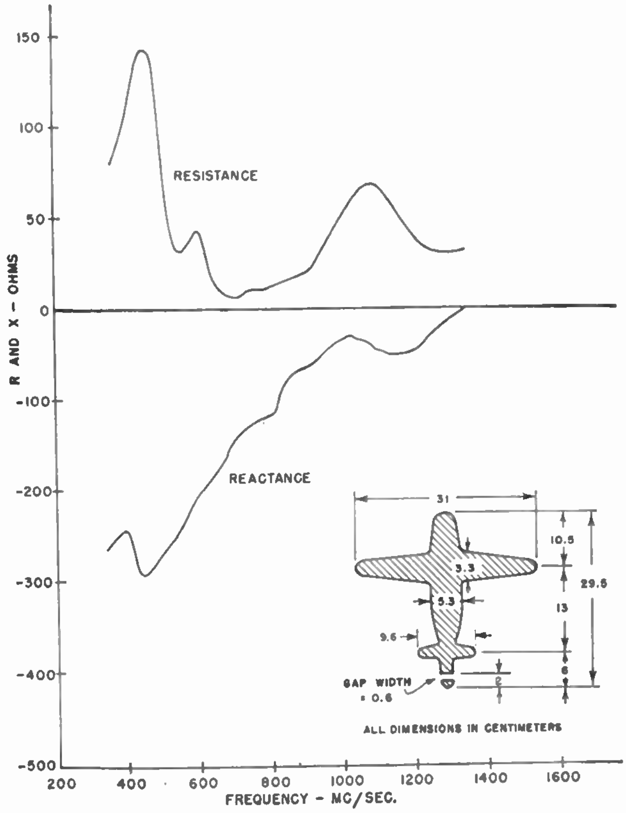

Fig. 3—Measured impedance characteristic of tail-cap antenna on simplified model

Measurements on simulated wing-cap antennas have been made with the slot arrangement shown in Fig. 2. The fore-and-aft symmetry of the slot representing the aircraft fuselage permits use of the image-plane technique which could not be used if a more accurate model were employed. Although this symmetry eliminates certain resonances associated with fuselage currents, measurements on such a model have yielded useful information on the effect of the length of the isolated section and the gap width on the impedance characteristics of the system.[6]

V. The Feed System

It is of interest to consider briefly the details of the feed region in image-plane slot measurements. The inner-conductor extension which spans the slot in the slot problem is seen to correspond to the gap between the main structure and the isolated section in the planar model. If this inner-conductor extension is made of a flat strip lying in the plane containing the slot, its width corresponds directly to the gap width in the planar model. For extensions made with circular sections, the equivalent strip width may be calculated by employing again the theory of equivalent cross sections. The interpretation of the effect of nonplanar feeders in terms of the equivalent gap width is of particular interest in connection with problems which do not have image symmetry and which do not, therefore, permit use of the image-plane technique. In such a problem it would be desirable to provide the feed at the center of the isolating gap by some scheme such as that shown in Fig. 4. The use of equivalent cross-section theory for interpreting the planar-model gap width in terms of the feeder diameter in the slot problem has been verified experimentally for small diameter feeders by comparison checks using image-plane measurements.

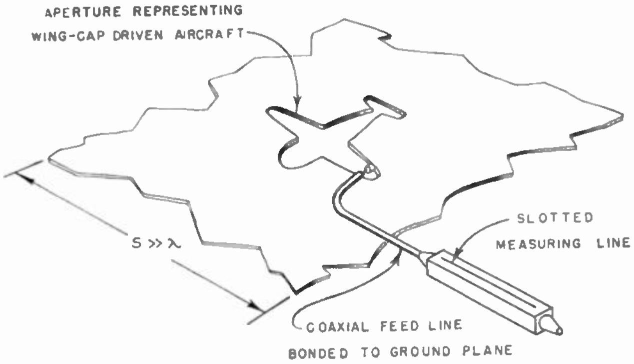

Fig 4—Full-slot analogue of wing-cap antenna.

A final consideration of importance in connection with the feed configuration concerns the base capacitance. In an actual wing-cap or tail-cap antenna the structure has a finite thickness at the gap, and fields will be set up across the gap in the region within the contours of the structure. In the planar model, there is no counterpart to this region, and hence, the effect of the internal fields is not accounted for. Impedance values measured with the planar model may be corrected for this effect by adding a base-shunting capacitance correction term. The required value of base capacitance may be estimated from the dimensions of gap region.

VI. Experimental Work

All measurements to date have been made in the frequency range 400–1,600 mc using the image-plane technique.

Initial measurements were made on a relatively narrow, linear, center-fed slot so that the equivalent dipole impedance thus determined could be compared with

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[6] J. V. N. Granger, "Wing-Cap and Tail-Cap Aircraft Antennas," Tech. Rpt. No. 6, Contract AF 19(122)78. Aircraft Radiation Systems Laboratory. Stanford Research Institute; March 1950.

available data on dipole impedances. The measured impedance values and the corresponding slot dimensions are shown in Fig. 5. Taking the equivalent diameter of the complementary strip dipole to be half the strip width, this dipole has a ratio of total length to equivalent diameter of 34.7. The resonant resistance is seen to have a reasonable value of about 60 ohms. The antiresonant resistance is 600 ohms, or about 11 per cent higher than the value of 540 ohms calculated from Schelkunoff's theory.[7] This check is considered to be adequate, in view of the general disagreement on the exact values of antiresonant resistance of such antennas.

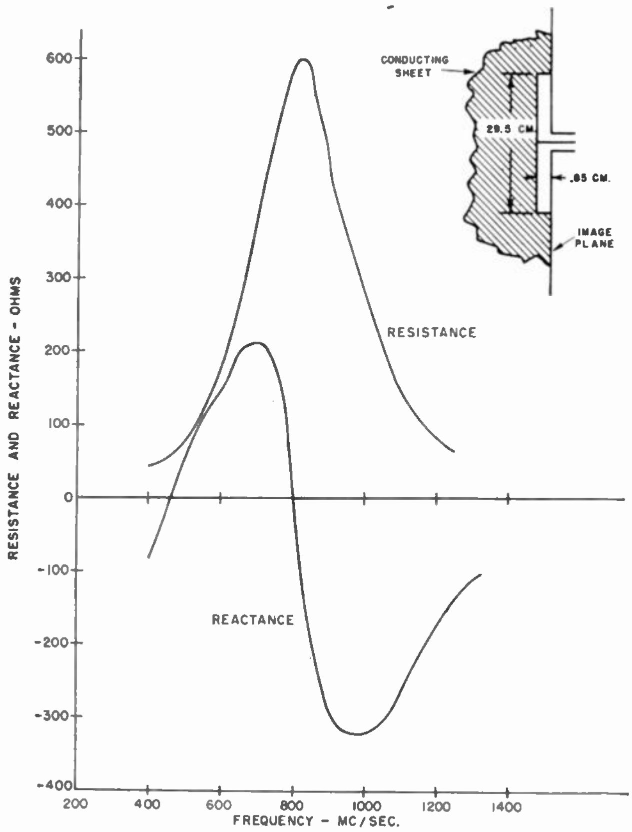

Fig. 5—Dipole impedance as determined from complementary slot measurements.

The complementary slot modelling technique has been employed in an extensive investigation of the impedance properties of wing-cap and tail-cap antennas. A typical measured impedance curve is shown in Fig. 3, in which the aircraft modelled was the B-29. An evaluation of the accuracy of the method is difficult because of the lack of adequate data on full-scale structures. Data obtained on the only comparable full-scale structure known to the writers are in reasonable agreement, when the effect of base capacitance is taken into account. These impedances are characterized by resistance values which vary between relatively narrow limits as compared with the corresponding variations for conventional wire antennas, and reactance curves which resemble that of a fixed capacitor in series with one or more low-Q, parallel resonant circuits.

ACKNOWLEDGMENTS

The writers wish to acknowledge the contributions made to this work by C. T. Tai in many discussions of the problem. The experimental results reported herein were measured by Mrs. Robert F. Reese.

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[7] S. A. Schelkunoff, "Theory of antennas of arbitrary size and shape," PROC. I.R.E., vol. 29, pp. 493–521; September, 1941.

Alignment and Adjustment of Synchronously Tuned Multiple-Resonant-Circuit Filters*

MILTON DISHAL†, SENIOR MEMBER, IRE

Summary—**A simple method of "tuning up" a multiple-resonant-circuit filter quickly and exactly is demonstrated. The method may be summarized as follows: Very loosely couple a detector to the first resonator of the filter; then, proceeding in consecutive order, tune all odd-numbered resonators for maximum detector output, and all even-numbered resonators for minimum detector output (always making sure that the resonator immediately following the one to be resonated is completely detuned).**

Also considered is the correct adjustment of the two other types of constants in a filter. Filter constants can always be reduced to only three fundamental types: f_0, $d_r(1/Q_r)$, and $K_{r(r+1)}$. This is true whether a lumped-element 100-kc filter or a distributed-element 5,000-mc unit is being considered. d_r is adjusted by considering the rth resonator as a single-tuned circuit (all other resonators completely detuned) and setting the bandwidth between the 3-db-down-points to the required value. $K_{r(r+1)}$ is adjusted by considering the rth and $(r+1)$th adjacent resonators as a double-tuned circuit (all other resonators completely detuned) and setting the bandwidth between the resulting response peaks to the required value.

Finally, all the required values for K and Q are given for an n-resonant-circuit filter that will produce the response $(V_p/V)^2=1+(\Delta f/\Delta f_{3db})^{2n}$.

I. INTRODUCTION

THIS PAPER attempts to answer two questions: "Exactly how can one 'tune up' a synchronously tuned multiple-resonant-circuit filter and be sure the tuning is correct?" and "Exactly how can one make

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* Decimal classification: R143.2×R386. Original manuscript received by the Institute, May 1, 1950; revised manuscript received, January 12, 1951. This paper was written in connection with work done on a contract for the Department of Air Force—AMC, Wright-Paterson Air Force Base.

† Federal Telecommunication Laboratories, Inc., Nutley, N. J.

sure that the mechanical design is actually supplying the required circuit constants?"

It should be noted that, for brevity, the paper will refer only to band-pass filters; the reader should realize that the discussion also applies similarly to the alignment and adjustment of low-pass, high-pass, and band-rejection filters when analogous frequency-variables and circuit constants are used.

The physical embodiment of a constant-K or equivalent type of filter, i.e., a filter having n complex frequency roots and no finite frequencies of infinite attenuation (all zeros at infinity), must exactly supply the numerical values of three kinds of quantities:[1] (1) resonant frequency f_0, (2) coefficients of coupling between adjacent resonators $K_{r(r+1)}$, and (3) resonator decrements $d_r(Q_r=1/d_r)$. It may be helpful to note that the above is true whether the elements of the unit are lumped, quasi-lumped, or distributed, so long as the percentage bandwidth is less than approximately 10 per cent for the latter two cases, as is most always true.

It should be realized that in the literature concerning filters a number of seemingly different types of constants have been used to describe the same or exactly equivalent networks. For example, classical filter theory which gives only approximate design data, usually produces the required values for L, C, M, and R; late in the 1930's, a number of papers described circuits by the so-called "ladder-network coefficients" for each arm; and at present many papers speak of the doubly or singly loaded Q of each resonator. In every case, the many different types of constants are all equivalent, but it has been the experience of the author that the constants f_0, $K_{r(r+1)}$, and Q_r are the "best" to use, particularly when dealing with *dissipative* filters.

For practical reasons usually involving mechanical tolerances, most selective-circuit designs incorporate a trimming adjustment for setting the resonant frequency of each resonator. After the filter is mechanically finished, the unit is aligned, i.e., all resonant frequencies are somehow correctly adjusted. Section III describes a method of alignment for multiple-resonant-circuit filters that is precise, requires no sweep-frequency generator, and can be performed by unskilled personnel.

The coefficient of coupling between adjacent resonators is usually not made variable as this adjustment requires a person "skilled in the art"; each K is carefully set by the designer as part of the mechanical design, which must be sufficiently stable to maintain it at the required value. Section IV describes an easy method for experimentally adjusting each coefficient of coupling to the exact desired value.

For the sake of completeness, a few comments are made about measuring Q, the third of these constants, in Section V.

Section VI presents some useful design data on what values of K and Q are required in a multiple-resonator-circuit filter.

II. SYMBOLS

n = total number of resonators used in a filter.

r = resonator number in filter chain. The resonator at the input end is numbered 1.

f_0 = resonant frequency of each resonator; this must include all coupling reactances. See Section IIIC.

$K_{r(r+1)}$ = coefficient of coupling between the rth and $(r+1)$th adjacent resonators. This may be defined fundamentally as the fractional bandwidth between the resulting response peaks that exist when each *pair* of adjacent resonators is considered separately (and the resonator Q's are infinite).

d_r = decrement of the resonator. This may be defined fundamentally as the fractional bandwidth between the 3-db-down points when each resonator is considered separately.

Δf_{3db} = total bandwidth between 3-db-down response points.

$V_{1,r}$ = voltage across *first* resonator when all following resonators up to the rth resonator have been correctly tuned.

$p_{AB}=K_{AB}(Q_AQ_B)^{1/2}$ = fraction of "critical coupling" in a double-tuned circuit made up of resonators A and B.

$(\Delta f_p)_A$ = total bandwidth between response peaks in resonator A, to which a generator is coupled, when A is coupled *only* to an adjacent resonator B.

$(\Delta f_p)_B$ = total bandwidth between the response peaks in the above-mentioned resonator B.

$F_p=(\Delta f_p/f_0)$ = fractional bandwidth between response peaks.

$t=Q_A/Q_B$.

Δf_β = total bandwidth between response points that are V_p/V_β down from the peak response.

V_p = voltage output at peak of response curve.

V_β = voltage output at point of response curve where the bandwidth is Δf_β.

III. ALIGNMENT OF MULTIPLE-RESONATOR FILTERS

A. General Principle

This paper will refer mainly to small-percentage-bandwidth node networks. The reader should realize that in accordance with the principle of duality and with the following substitutions of words: mesh for node, current for voltage, voltage for current, open circuit for short circuit, and the like, the alignment procedure applies similarly to mesh networks. The procedure applies also to the large-percentage-bandwidth constant-K configuration discussed in Section IIIC.

The fundamental principle proposed in this section is that alignment can best be done by *completely assembling the filter and then concentrating on the amplitude phenomena occurring in the first resonator of the filter chain at*

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[1] M. Dishal, "Design of dissipative band-pass filters producing desired exact amplitude-frequency characteristics," PROC. I.R.E., vol. 37, pp. 1050–1069; September, 1949. Also *Elec. Commun.*, vol. 27, pp. 56–81; March, 1950.

the desired resonant frequency. In Section IIIC, it is shown that if all the resonators are first completely detuned and if they are resonated in numerical order, calling the input resonator 1, then all odd-numbered resonators place an open circuit (high resistance) and all even-numbered resonators place a short circuit (low resistance) across the input terminals when correctly tuned.

B. Alignment Procedure

The alignment procedure will be described using the quadruple-tuned node-type band-pass filter shown in Fig. 1 as an example. Fig. 2 shows a practical physical embodiment of the circuit of Fig. 1.

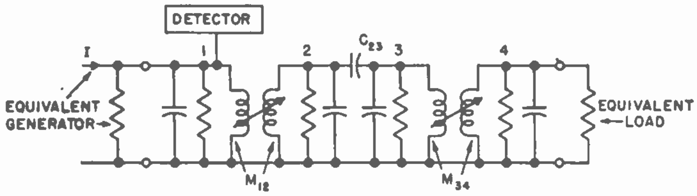

Fig. 1—Quadruple-tuned filter used to demonstrate alignment procedure of Section III. It should be realized that the alignment procedure applies to all the different types of synchronously tuned constant-K and coupled-resonant-circuit filters.

The procedure is applicable to all coupled-resonant-circuit filters, whether they be low-frequency constant-K configurations, medium-frequency coupled circuits, microwave quarter-wave-coupled waveguide filters, or the like.

1. Connect the generator to the first resonator of the filter and the load to the last resonator of the filter in exactly the same manner as they will be connected in actual use.
2. Couple a nonresonant detector directly and very loosely[2] to either the electric (voltage) or magnetic (current) field of the *first* resonator of the filter chain.
3. Completely detune[3] all resonators.
4. Set the generator frequency to the desired midfrequency of the filter.
5. Tune resonator 1 for ***maximum output*** indication on the detector. Lock the tuning adjustment.
6. Tune resonator 2 for ***minimum output*** indication on detector. Lock the tuning adjustment.
7. Tune resonator 3 for maximum output and lock the tuning adjustment.
8. Tune resonator 4 for minimum output and lock the tuning adjustment. The alignment of the filter shown in Fig. 1 is now complete.

If it is impracticable completely to detune all the resonators in a node network, a *single* device may be used to short-circuit ***the resonator immediately following the one being tuned*** since this will remove the effect of all the following resonators. It is important to make sure that this short circuit is fully effective at the measurement frequency.

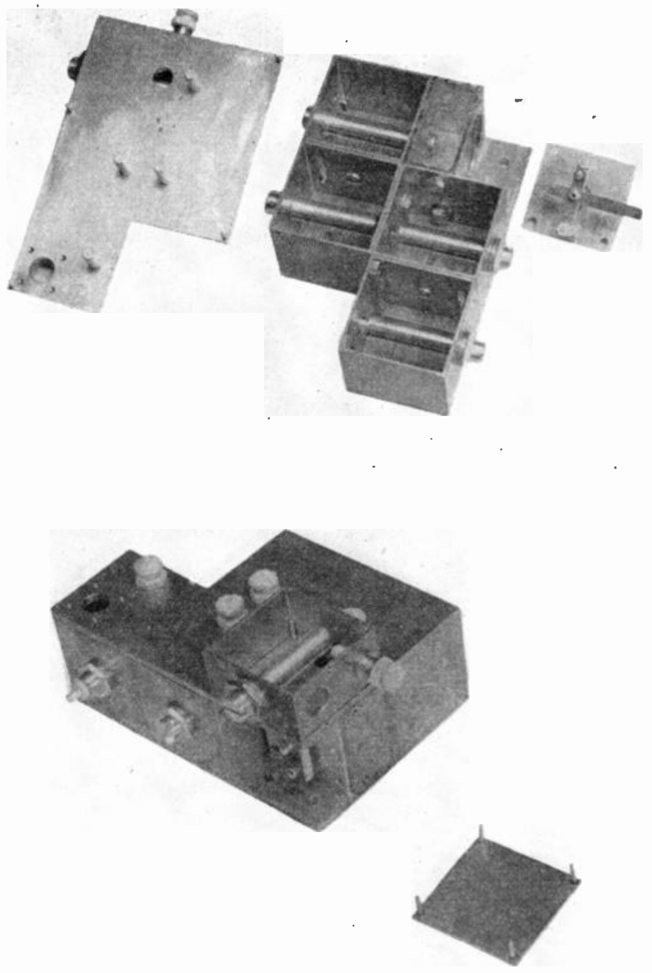

Fig. 2—A quadruple-tuned μhf filter embodying the circuit of Fig. 1. The midfrequency is 1,400 mc and the 3-db bandwidth is 6 mc. Note the smaller inductive coupling slot between resonators 1 and 2, the capacitive coupling hole between resonators 2 and 3, and the larger inductive coupling slot between resonators 3 and 4. The small plate with the "cross" mounted on it is the crystal mixer unit; when it is mounted in the last small cavity, the crystal is correctly coupled (capacitively) to both the fourth resonator and to the local-oscillator resonator.

Fig. 3 clearly demonstrates the amplitude-frequency phenomena that occur in each step of the alignment procedure. A sweep-frequency generator was used, and attention is called to the resonant-frequency marker. It should be clearly realized that since the alignment adjustments depend exclusively on the amplitude of the response at the resonant frequency f_0, a sweep-frequency generator is *not* required and all adjustments can be made with a single-frequency input f_0.

These oscillograms were obtained in aligning the quadruple-tuned filter shown in Fig. 1. It was designed to produce a Chebishev transfer shape[1] having a 1/2-db peak-to-valley ratio when loaded on one side only; i.e., it was fed by a high-impedance generator.

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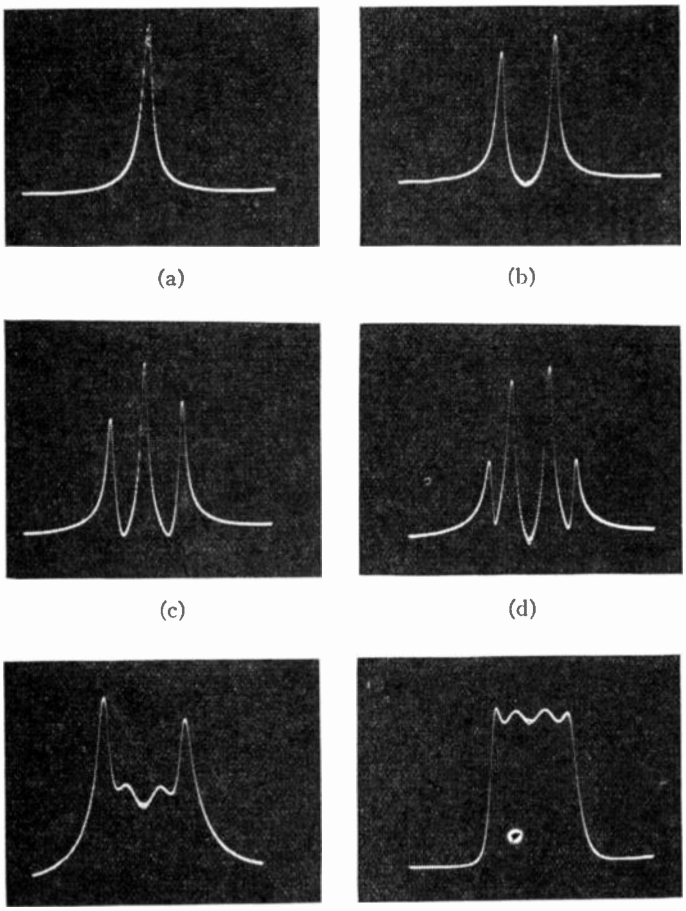
[2] A nonresonant detector (or generator) may be said to be "very loosely" coupled when it lowers the unloaded Q of the resonator by less than 5 per cent (say).

[3] A resonator is sufficiently detuned when its resonant frequency is at least 10 pass-band-widths away from the pass-band midfrequency.

Efficient filters with low internal losses, i.e., those using resonators having unloaded Q's very much greater than the fractional midfrequency $(f_0/\Delta f_{3db})$, produce deep and broad minimums when the even-numbered resonators are properly tuned, as may be seen in Fig. 3(b). Therefore, it is important to use a large-amplitude signal input and high detector gain so that the middle of the minimum can be tuned accurately to the midfrequency. If the maximum generator input and detector gain still produce a broad null, the tuning adjustments should be set midway between two points of equal output.


(a) (b)

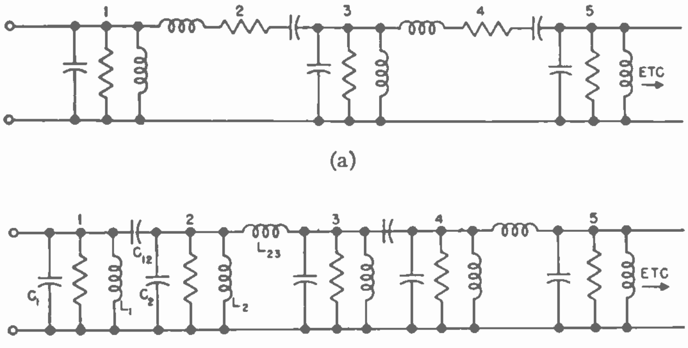
(c) (d)

(e) (f)

Fig. 3—Oscillograms of the amplitude-frequency phenomenon occurring in resonator 1 as the alignment steps of Section III are performed. (a) Resonator 2 is short-circuited or detuned and the input resonator 1 (odd-numbered) is adjusted for maximum marker signal of f_0. (b) Resonator 3 is detuned and the second resonator (even-numbered) is tuned for minimum amplitude of f_0. Oscillograms (c) and (d) show the continuation of the procedure of tuning odd-numbered resonators to produce maximum and even-numbered resonators to produce minimum values of f_0 response in resonator 1. It can be seen that as the rth resonator is tuned, there will be r peaks and $r-1$ valleys produced in resonator 1; this is a simple restatement of Foster's reactance theorem. Oscillogram (e) shows the voltage across resonator 1 as, with correct applied loading, the last resonator (even-numbered) of Fig. 1 is tuned for minimum output at f_0. Oscillogram (f) shows the resulting Chebishev transfer response shape (no tuning adjustments were retouched).

C. Simple Theory of Alignment Procedure

Perhaps the simplest way of showing that the alignment procedure is correct is to consider the large-percentage-bandwidth constant-K filter chain shown in Fig. 4(a), to which all small-percentage-bandwidth coupled-resonant-circuit filters are exactly equivalent no matter what type of coupling is used between adjacent resonant circuits; and then to consider as a further example the small-percentage-bandwidth node circuit shown in Fig. 4(b).



(a)

(b)

Fig. 4—(a) and (b) are, respectively, large- and small-percentage-bandwidth networks referred to in the explanations given in Section IIIC. It should be clearly realized that Fig. 3(b) shows alternate C and L couplings purely as an example; the explanation is true no matter what type of reactive coupling is used.

The reasoning applicable to the constant-K configuration of Fig. 4(a) requires previous knowledge of two simple facts.

First, complete detuning of all the resonators means that all the *series resonators* are effectively *open-circuited* and all the shunt resonators are effectively *short-circuited.*

Second, when correctly aligned, the resonant frequency f_0 of each separate resonator is identical.

Thus, when resonator 1 is tuned to f_0 (with resonator 2 open-circuited), maximum voltage will appear across the high parallel-resonant resistance of resonator 1. When resonator 2 is tuned to f_0 (with resonator 3 short-circuited), the low series-resonant resistance of 2 will shunt the terminals of 1, thus dropping the voltage across resonator 1 to a minimum. On tuning 3 to f_0 (with resonator 4 open-circuited), the high parallel-resonant resistance of 3 will remove the low series-resonant resistance of 2 from across the terminals of 1 so that the voltage across 1 will again rise to a maximum. Thus, starting at the front end of the filter, all odd-numbered resonators will produce a maximum voltage and all even-numbered resonators will produce a minimum voltage across resonator 1.

The reasoning applicable to the small-percentage-bandwidth node network of Fig. 4(b) requires previous knowledge of three simple facts.

First, complete detuning of a resonator means that the node involved is effectively short-circuited.

Second, when correctly aligned, the resonant frequency of each node is identical and the elements that resonate a node consist of every susceptance that touches the node, e.g., node 2 of Fig. 4(b) is resonated by adjusting C_2 (or L_2) to resonate with parallel combination of C_{12}, C_2, L_2, and L_{23}.

Third, if a group of reactances parallel resonate together, then any one of the reactances also series resonates with all the others, e.g., C_{12} series resonates with the parallel resultant of C_2, L_2, and L_{23}.

Thus, where node 1 is tuned to f_0 (with node 2 short-circuited), the high parallel-resonant resistance of C_1, L_1, and C_{12} will produce a voltage maximum at f_0. When node 2 is tuned to f_0 (with node 3 short-circuited), C_{12} will series resonate with the parallel resultant of C_2, L_2, and L_{23}, thus placing a short circuit across node 1 and producing a voltage minimum across node 1. The process repeats as alignment proceeds, producing maximums for alignment of odd-numbered and minimums for even-numbered resonators.

It will be shown in Section IVD that if we know the Q's of each resonator being used, then the ratios of maxima and minima occurring in resonator 1 can be used to set or check all the $(n-1)$ coefficients of coupling in a filter.

IV. Exact Adjustment of Coupling Between Resonators

A. General Principle

Before this section can be applied to the mechanical design and adjustment of a filter, it is, of course, necessary to determine by some synthesis procedure just what adjacent-resonant-circuit coefficients of coupling the mechanical embodiment must produce. As mentioned in the Introduction, no matter what type of constants are used to describe the synthesized network, they can always be transformed into f_0, $K_{r(r+1)}$, and Q_r.

The fundamental procedure being proposed in this section is to consider every pair of adjacent resonators as a double-tuned, i.e., two-pole, circuit (with all the other resonators completely detuned), and so be able to use the exactly known relation between the circuit constants and the resulting amplitude-frequency characteristic of a double-tuned circuit.

In a double-tuned circuit with Q_A and Q_B equal to infinity, the fractional bandwidth $(\Delta f_p/f_0)_A$ between primary response peaks is exactly equal to the coefficient of coupling between resonators A and B. (*In fact, this may be used as the basic definition of the constant that is commonly called the coefficient of coupling.*) When the resonators do not have infinite Q, the above equality is not exactly true, and Figs. 5 and 6 together with the described procedures supply two ways of finding the exact coefficient of coupling between adjacent resonators.

B. Adjustment Procedure

The procedure for measuring adjacent-resonator coupling, which is applicable to all coupled-resonant-circuit filters whether they be called low-frequency constant-K configurations, medium-frequency coupled circuits, or microwave quarter-wave coupled-waveguide filters, is as follows:

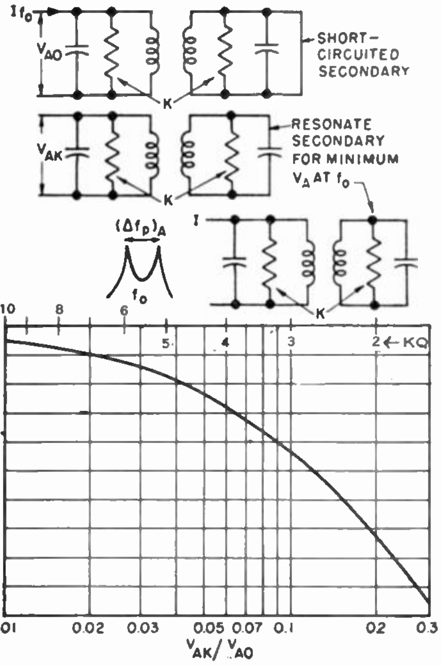


Fig. 5—Method of obtaining exact coefficient of coupling K between two resonators by measurements on only the primary circuit.

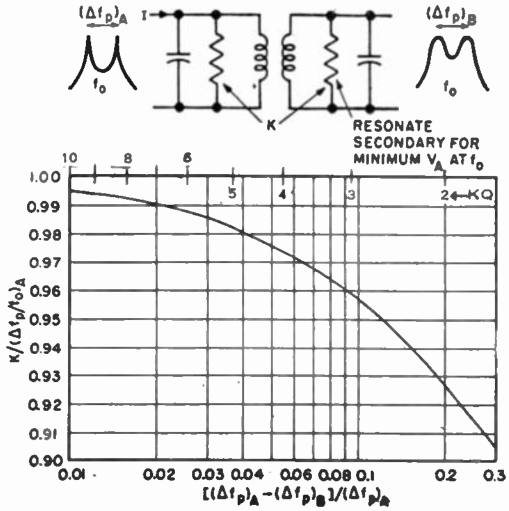


Fig. 6—Method of obtaining exact coefficient of coupling K between the two resonators by frequency measurements only.

1. Designate as A and B the two adjacent resonators between which the coefficient of coupling is to be adjusted.
2. Couple a nonresonant signal generator directly and very loosely to either the electric (voltage) or magnetic (current) field of resonator A.
3. Couple a nonresonant detector directly and very loosely to either the electric (voltage) or magnetic (current) field of resonator A.
4. Completely detune all the resonators in the filter chain.
5. Tune resonator A for maximum output from the detector. Record the signal-generator input and detector output.
6. Tune resonator B for minimum output from the detector (as in alignment procedure, Section III). Increase the signal-generator input to produce the same output obtained in step 5.

7. The ratio of the signal-generator input in step 5 to that in step 6 is the abscissa of the graph of Fig. 5. From the ordinate of this graph, read the ratio of the coefficient of coupling K between resonators A and B, to the percentage bandwidth $(\Delta f_p/f_0)_A$ between the response peaks that are now present across resonator A.

8. Carefully measure the bandwidth $(\Delta f_p)_A$ between the response peaks of resonator A.

9. The exact coefficient of coupling is equal to the fractional bandwidth between these peaks times the ordinate obtained in step 7.

If it is not convenient to measure the amplitude ratio described in step 7, the following procedure involving frequency measurements only can be used. Omit the amplitude measurements from the above procedure and after step 6 carefully measure, by means of a nonresonant detector loosely coupled to resonator B, the bandwidth $(\Delta f_p)_B$ between the response peaks appearing on the secondary side.

The fractional difference in peak bandwidth $[(\Delta f_p)_A-(\Delta f_p)_B]/(\Delta f_p)_A$ is the abscissa of the graph of Fig. 6. The exact coefficient of coupling is equal to the fractional bandwidth $(\Delta f_p/f_0)_A$ between *primary* peaks times the corresponding ordinate given in Fig. 6.

Examination of the ordinate values of Figs. 5 and 6 shows that even with a $(V_{A,A}/V_{A,B})$ ratio as small as 12 db, or a $[(\Delta f_p)_A-(\Delta f_p)_B]/(\Delta f_p)_A$ ratio as large as 25 per cent, the existing coefficient of coupling is only 8.5 per cent less than the percentage bandwidth between the primary peaks; therefore, in many cases it may be permissible simply to measure the bandwidth $(\Delta f_p)_A$ between primary response peaks and make the approximation that the coefficient of coupling is equal to about 0.96 times the measured fractional bandwidth.

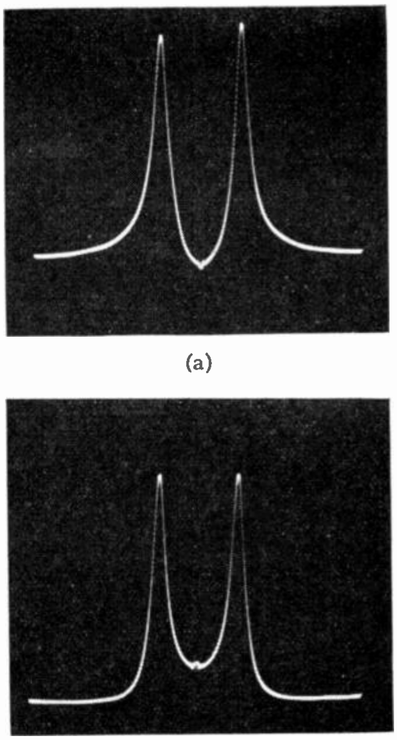
Figs. 7(a) and 7(b) show the frequency-amplitude relations on the primary and secondary sides when the above procedure is used.

C. Quantitative Theory of Measuring Coefficient of Coupling

All the amplitude-frequency characteristics of a double-tuned circuit are most conveniently related to each other by means of two variables, the Q ratio of primary to secondary Q_A/Q_B and the $K_{AB}(Q_AQ_B)^{1/2}$ product. For the preparation of Figs. 5 and 6, we need the relations between the above variables and four quantities: (1) drop in primary voltage when the secondary is correctly resonated $V_{A,B}/V_{A,A}$; (2) bandwidth between response peaks in the primary side $(\Delta f_p)_A$; (3) bandwidth between peaks on the secondary side $(\Delta f_p)_B$; and (4) relation of bandwidth between primary peaks to the coefficient of coupling $K_{AB}/(\Delta f_p/f_0)_A$.

Straightforward analysis of a correctly resonated double-tuned circuit results in the following equations for these quantities in which the Q ratio will be denoted by $t=Q_A/Q_B$ and the $K_{AB}(Q_AQ_B)^{1/2}$ product by p.

$$\left(\frac{V_{A,A}}{V_{A,B}}\right) = 1 + p^2 \tag{1}$$



(a)

(b)

Fig. 7—Response peaks of a properly resonated pair of adjacent resonators with all other resonators completely detuned. By measuring the frequency bandwidth between the peaks occurring in the input resonator (a), the exact coefficient of coupling may be obtained. (See Fig. 4.) The corresponding peak bandwidth for the second resonator (b) differs slightly from (a) because the resonators have finite values of Q. (See Fig. 5.)

$$\left(\frac{\Delta f_p}{f_0}\right)_A = K\left\{\left[1 + 2(1+t)\frac{1}{p^2}\right]^{1/2} - t\frac{1}{p^2}\right\}^{1/2} \tag{2}$$

$$\left(\frac{\Delta f_p}{f_0}\right)_B = K\left[1 - 1/2\left(t + \frac{1}{t}\right)\frac{1}{p^2}\right]^{1/2}. \tag{3}$$

The fourth relation mentioned above is, of course, obtained from (2).

Equations (1), (2), and (3) are used straightforwardly to obtain the graphs of Figs. 5 and 6 for the case where the Q ratio is unity (most filters are built using resonators having the same unloaded Q).

D. Checking Coefficients of Coupling

During construction, or after a filter has been constructed, it is often desirable to be able to set or to double check each coefficient of coupling without going through the procedure of converting each pair of adjacent resonators into a double-tuned circuit as is required by Section IVB.

As shown below this can be accomplished by measurements made entirely at the input resonator. There are, in practice, two cases which have to be considered: In the first case (usually the large-percentage-bandwidth filter) the unloaded Q's of the resonators being used are very much greater than the fractional midfrequency $(f_0/\Delta f_{3db})$ being used; i.e., the unloaded individual Q's are essentially infinite. In the second case (usually the small-percentage-bandwidth filter), the unloaded Q's of the resonators are only 4 or 5 (say) times $(f_0/\Delta f_{3db})$.

For the first case above, the K's can be set, or measured, in consecutive order, by measuring the bandwidth between the various response peaks appearing in resonator 1, as each of the following resonators is resonated in consecutive order (see Fig. 3).

It will be remembered from Fig. 3 that there will be r response peaks occurring in the input resonator when the rth resonator is correctly tuned.

To calculate the peak bandwidths that should be measured when the rth resonator is tuned, straightforward analysis shows that we must solve the polynomial (4) for its roots

$$F_p{}^r - (\sum K^2)F_p{}^{(r-2)} + (\sum K^2K^2)F_p{}^{(r-4)} - (\sum K^2K^2K^2)F_p{}^{(r-6)} \cdots = 0. \quad (4)$$

The polynomial stops at the first- or zero-power term; i.e., no negative exponents are considered.

The coefficient of the $F_p{}^{(r-2)}$ term is the sum of all the products of the coefficients of coupling squared taken one at a time; the coefficient of the $F_p{}^{(r-4)}$ term is the sum of all the products of K^2 taken two at a time, *but in any pair a subscript number must not appear more than once*; the coefficient of the $F_p{}^{(r-6)}$ term is the sum of all the K^2 products taken three at a time, *but in any triplet a subscript number must not appear more than once*; and so forth.

As an example, as the first to fifth resonators are tuned consecutively, the following 5 polynomials must be solved consecutively to calculate the fractional bandwidth $(\Delta f_p/f_0)$ that should occur between response peaks.

$$F_p = 0 \quad (4a)$$

$$F_p{}^2 - K_{12}{}^2 = 0 \quad (4b)$$

$$F_p{}^3 - (K_{12}{}^2 + K_{23}{}^2)F_p = 0 \quad (4c)$$

$$F_p{}^4 - (K_{12}{}^2 + K_{23}{}^2 + K_{34}{}^2)F_p{}^2 + (K_{12}K_{34}) = 0 \quad (4d)$$

$$F_p{}^5 - (K_{12}{}^2 + K_{23}{}^2 + K_{34}{}^2 + K_{45}{}^2)F_p{}^3 + (K_{12}{}^2K_{34}{}^2 + K_{12}{}^2K_{45}{}^2 + K_{23}{}^2K_{45}{}^2)F_p = 0. \quad (4e)$$

These first five polynomials require simple linear- and quadratic-equation solutions, and, because the coefficients are known numerical values, the Graeffe root-squaring process can be used to solve accurately any of the polynomials.

For the second case described above, the coefficients of coupling should be set or measured in consecutive order as follows: Accurately measure the Q of each resonator in the filter, then proceed step by step through the alignment procedure of Section III*B*, accurately measuring (and recording) the magnitudes of the alternate maxima and minima produced. Straightforward analysis shows that the ratio of the detector output obtained when resonator 1 is alone resonated to that obtained when resonator r is resonated is given by the continued fraction of (5).

$$\left(\frac{V_{1,1}}{V_{1,r}}\right) = \left[1 + \cfrac{p_{12}{}^2}{1 + \cfrac{p_{23}{}^2}{1 + \cdots \cfrac{}{1 + p^2{}_{(r-1r)}}}}\right]. \quad (5)$$

It is important to realize that Q_1 is the Q of resonator number 1 with both the generator and detector coupled to it.

Since we know the desired value for each K and have measured each Q, we know the value of each $P_{12}{}^2 = K_{12}{}^2Q_1Q_2$, and so on in (5); and the measured value of the voltage ratios $(V_{1,1}/V_{1,r})$ should, of course, equal those calculated from (5).

V. ADJUSTMENT OF RESONATOR DECREMENT (1/Q)

It has been the author's experience that any method of measuring Q that removes the resonator from the exact position it occupies in the filter chain is potentially inaccurate.

It has also been noted that measurements involving an amplitude-modulated oscillator can lead to erroneous results particularly in the uhf and microwave regions because of spurious frequency modulation. If an amplitude-modulated carrier is being used, an obvious check for appreciable spurious frequency modulation is to use an oscilloscope to examine the envelope of the wave form being measured and a narrow-band receiver to examine the frequency content of the supposedly purely amplitude-modulated carrier.

The most trustworthy method of making accurate unloaded or loaded Q measurements on a resonator that is part of a filter chain seems to be as follows:

1. Completely assemble the filter.
2. Completely detune all resonators except the one to be measured. Obviously, complete detuning of the resonator on each side of the one being measured should be satisfactory.
3. A nonresonant signal generator is coupled directly and very loosely to either the electric or magnetic field of the resonator.
4. A nonresonant detector is coupled very loosely and preferably to the field opposite to that being used for the generator; i.e., make sure that there is negligible direct coupling between generator and detector.
5. Using an unmodulated wave or an amplitude-modulated wave checked for negligible frequency modulation from the signal generator, measure the frequency difference Δf_β between the points that are V_p/V_β down from the peak response; the resonator Q is given by

$$Q = (f_0/\Delta f_\beta)[(V_p/V_\beta)^2 - 1]^{1/2}. \quad (6)$$

Obviously, when high Q's are to be measured, the apparatus must be capable of measuring very-small-percentage bandwidths. This may be accomplished by "beating down" the measurement frequency with a very stable local oscillator and a mixer and by making the measurements at the resulting difference frequency. By this means, the accuracy of the measuring apparatus is increased by the ratio of the original to the difference frequencies. The "cost" of this simplification is, of course, the necessity of using a very stable local oscillator.

VI. K'S AND Q'S TO PRODUCE RESPONSE SHAPE $V_p/V = [1 + (\Delta f/\Delta f_{3db})^{2n}]^{1/2}$.

A straightforward synthesis procedure[1,4,5] shows that for the transfer response shape given just above, and if infinite Q resonators are used, i.e., in practice, resonators whose unloaded Q's are greater than $10/[\sin (90°/n)]$ times the fractional midfrequency $(f_0/\Delta f_{3db})$, it is possible to write very concisely the exact values of K and Q required for any number n of resonators for two important practical cases. The rate of cutoff obtained with the transfer response shape given in the title of this section is exactly $6n$ db per octave and the size of the octave is the bandwidth between the midfrequency and any frequency past the 3-db-down frequency.

For the case where *one end of the network can have only a pure reactance placed across it* (by either a reactive generator, e.g., the plate of a pentode tube, or by a reactive load), the required Q and K values are given exactly by (7(a)) and (7(b)).

$$\frac{Q_1}{(f_0/\Delta f_{3db})} = \sin \frac{90°}{n}, \quad Q_{2\to n} = \infty \tag{7a}$$

$$\frac{K_{r(r+1)}}{(\Delta f_{3db}/f_0)} = \frac{\cos (r\, 90°/n)}{\{[\sin (2r-1)(90°/n)][\sin (2r+1)(90°/n)]\}^{1/2}} \tag{7b}$$

where r is made equal to 1, 2, 3, and so forth, up to $(n-1)$; and n is the total number of resonators used. For the above design equations, *the end resonator that is loaded is called resonator* 1.

For the case where *both ends of the network must have resistances placed across them* (e.g., a 50-ohm generator and 50-ohm load are being used), then the required Q and K values are given exactly by (8(a)) and (8(b)).

$$\frac{Q_1(=Q_n)}{(f_0/\Delta f_{3db})} = 2 \sin \frac{90°}{n}, \quad Q_{2\to(n-1)} = \infty \tag{8a}$$

$$\frac{K_{r(r+1)}}{(\Delta f_{3db}/f_0)} = \frac{0.5}{\{[\sin (2r-1)(90°/n)][\sin (2r+1)(90°/n)]\}^{1/2}}, \tag{8b}$$

where r is set equal to 1, 2, 3, and so forth, up to $(n-1)$; and n is the total number of resonators being used. Since the resulting circuit is symmetrical, it makes no difference which end resonator is called resonator 1.

For the unfortunately practical case where the unloaded Q of the resonators being used is not infinite, it does not seem to be possible to obtain elegantly simple design equations like (7) and (8). Figs. 8 and 9 give the K and Q values required to produce exactly the transfer response shape $(V_p/V)^2 = 1 + (\Delta f/\Delta f_{3db})^{2n}$ for triple- and quadruple-resonator filters, respectively, for the reactive-generator case. The abscissa of these graphs is the ratio of the unloaded $Q(Q_0)$ of the resonators being used to the fractional midfrequency $(f_0/\Delta f_{3db})$.

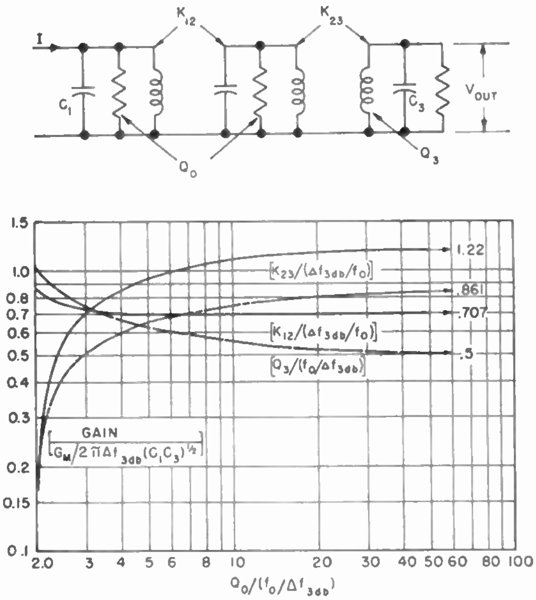


Fig. 8—Exact design for a finite-Q triple-tuned node circuit producing a Butterworth response shape when driven by an infinite-resistance generator.

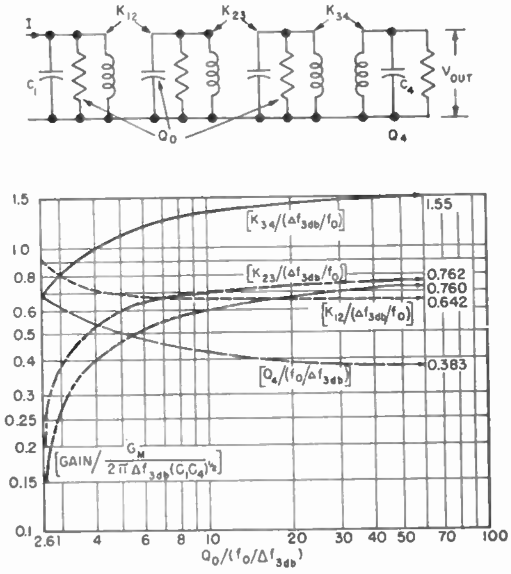


Fig. 9—Exact design for a finite-Q quadruple-tuned node circuit producing a Butterworth response when driven by an infinite-resistance generator.

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[4] E. L. Norton, U. S. Patent No. 1,788,538; January, 1931.
[5] W. R. Bennett, U. S. Patent No. 1,849,656; March, 1932.

The Multisection *RC* Filter Network Problem*

L. STORCH†

DISCUSSION

IN ORDER to analyze the network of Fig. 1, the author of a recent paper[1] finds it necessary to postulate that equations (1) to (4) represent the desired voltage transfer ratio $(E_0/E_i)_n$ and input admittance G_n, which assumptions he then validates by an inductive proof:

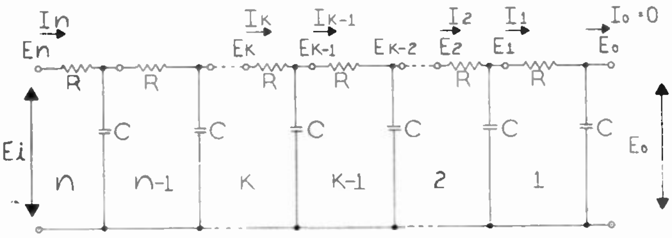

Fig. 1—(slightly modified). The n-section RC filter network.

$$\left(\frac{E_0}{E_i}\right)_n = \frac{1}{1 + a_1Tp + a_2T^2p^2 + \cdots + a_{n-2}T^{n-2}p^{n-2} + a_{n-1}T^{n-1}p^{n-1} + T^np^n} \tag{1}$$

$$G_n = \frac{Cp[n + b_1T_p + b_2T^2p^2 + \cdots + b_{n-2}T^{n-2}p^{n-2} + T^{n-1}p^{n-1}]}{1 + a_1T_p + a_2T^2p^2 + \cdots + a_{n-2}T^{n-2}p^{n-2} + a_{n-1}T^{n-1}p^{n-1} + T^np^n}, \tag{2}$$

where the coefficients of $(pT)^m$ in the case of n sections are:

$$a_{m,n} = \frac{(n+m)!}{(n-m)!(2m)!} \tag{3}$$

$$b_{m,n} = \frac{(n+m)!}{(n-m-1)!(2m+1)!}. \tag{4}$$

The ability to anticipate exactly the correct solution, which consists of four intricate equations, would seem to demand extraordinary intuition or prescience. It is the purpose of this note to show how the network characteristics can be obtained straightforwardly from the basic circuit equations.

This approach should be more meaningful to the engineer who is interested in the method of solving the actual problem, rather than in a formal proof of a set of elaborate postulates. In addition, it supplies him with a mode of attack which is valuable also when facing other circuit problems of the recurrent network type.

For any section k in Fig. 1 $(2 \leqq k \leqq n)$, by the junction law of currents:

$$I_k = \frac{E_k - E_{k-1}}{R} = \frac{E_{k-1} - E_{k-2}}{R} + pCE_{k-1}.$$

Therefore $(T = RC)$:

$$E_k = (2 + pT)E_{k-1} - E_{k-2} \tag{5}$$

with

$$E_0 = E_0 \quad \text{and} \quad E_1 = (1 + pT)E_0.$$

The single recursion process for E_k, carried on until $k = n$, is sufficient to produce the complete solution:

$$\left(\frac{E_0}{E_i}\right)_n = \frac{E_0}{E_n} \tag{6}$$

$$G_n = \frac{I_n}{E_n} = \frac{pC}{pT} \cdot \frac{E_n - E_{n-1}}{E_n}. \tag{7}$$

Furthermore, explicit expressions for the coefficients $a_{m,k}$ of the polynomial

$$\frac{E_k}{E_0} = \sum_{m=0}^{k} a_{m,k}(pT)^m, \tag{8}$$

which is generated by the recursion process, can also be derived from the fundamental equation (5).

Substituting (8) in (5) and collecting terms in $(pT)^m$:

$$a_{m,k} = 2a_{m,k-1} + a_{m-1,k-1} - a_{m,k-2} \tag{9}$$

with

$$a_{0,0} = 1, \qquad a_{0,1} = a_{1,1} = 1,$$

and by (8)

$$a_{m,k} = 0 \quad \text{when} \quad m > k, \text{ or } m < 0 \; .$$

Let us write (9) in the more symmetrical form:

$$\Delta_{m,k} = (a_{m,k} - a_{m,k-1}) = (a_{m,k-1} - a_{m,k-2}) + a_{m-1,k-1}, \tag{10}$$

and tabulate the first few terms[2] by forward differencing.

In view of the obvious nexus between the columns of Table I and sequences of binominal coefficients, these are listed in Table II. A comparison indicates that

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* Decimal classification: R143.2. Original manuscript received by the Institute, April 19, 1950; revised manuscript received, December 13, 1950.

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[1] E. W. Tschudi, "Admittance and transfer function for an n-mesh RC filter network," PROC. I.R.E., vol. 38, pp. 309–310; March, 1950.

[2] This type of derivation is chosen since it is more graphic and also requires less space.

TABLE I. $a_{m,k}$ AND $\Delta_{m,k}$

k	$a_{0,k}$	$\Delta_{1,k}$	$a_{1,k}$	$\Delta_{2,k}$	$a_{2,k}$	$\Delta_{3,k}$	$a_{3,k}$
0	1	0	0	0	0	0	0
1	1	1	1	0	0	0	0
2	1	2	3	1	1	0	0
3	1	3	6	4	5	1	1
4	1	4	10	10	15	6	7
5	1	5	15	20	35	21	28
6	1	6	21	35	70	56	84
7	1	7	28	56	126	126	210
8	1	8	36	84	210	252	462

TABLE II. BINOMIAL COEFFICIENTS

k	$\binom{k}{0}$	$\binom{k}{1}$	$\binom{k}{2}$	$\binom{k}{3}$	$\binom{k}{4}$	$\binom{k}{5}$	$\binom{k}{6}$
0	1	0	0	0	0	0	0
1	1	1	0	0	0	0	0
2	1	2	1	0	0	0	0
3	1	3	3	1	0	0	0
4	1	4	6	4	1	0	0
5	1	5	10	10	5	1	0
6	1	6	15	20	15	6	1
7	1	7	21	35	35	21	7
8	1	8	28	56	70	56	28

$$a_{0,k} = \binom{k}{0}, \qquad a_{1,k} = \binom{k+1}{2},$$

$$a_{2,k} = \binom{k+2}{4}, \qquad a_{3,k} = \binom{k+3}{6}.$$

Apparently, the general solution is

$$a_{m,k} = \binom{k+m}{2m},$$

which can be verified by substitution in (9).

Factoring $pC \cdot E_0/E_n$ in (7):

$$\frac{E_n - E_{n-1}}{(pT)E_0} = \sum_{m=0}^{n-1} b_{m,n}(pT)^m.$$

Consequently:

$$b_{m,n} = a_{m+1,n} - a_{m+1,n-1}$$
$$= \binom{n+m+1}{2m+2} - \binom{n+m}{2m+2} = \binom{n+m}{2m+1}.$$

The explicit solution for the coefficients is, therefore:

$$a_{m,n} = \binom{n+m}{2m} = \frac{(n+m)!}{(n-m)!(2m)!} \tag{11}$$

$$b_{m,n} = \binom{n+m}{2m+1} = \frac{(n+m)!}{(n-m-1)!(2m+1)!}. \tag{12}$$

This completes the solution of the problem.

A quite different method of analysis, which may be of considerable interest, will now be outlined briefly.

A resistor $R/2$ connected in series with the upper output terminal in Fig. 1 does not alter the desired open-circuit characteristics. But the structure may now be considered as a chain of n symmetrical T sections, with a resistor $R/2$ in each series arm and a capacitor C in each shunt arm, which is fed through a source impedance $R/2$.

In terms of the image parameters θ and Z_0 of a single T section,[3]

$$\cosh \theta = 1 + \frac{pT}{2}, \quad Z_0 = \frac{\sinh \theta}{pC} = \frac{R}{2} \cdot \frac{\sinh \theta}{\cosh \theta - 1}, \tag{13}$$

the properties of the complete network are described in matrix form by

$$\begin{pmatrix} E_n \\ I_n \end{pmatrix} = \begin{bmatrix} 1, & \dfrac{R}{2} \\ 0, & 1 \end{bmatrix} \begin{bmatrix} \cosh \theta, & Z_0 \sinh \theta \\ \dfrac{\sinh \theta}{Z_0}, & \cosh \theta \end{bmatrix}^n \begin{pmatrix} E_0 \\ I_0 \end{pmatrix}.$$

Substituting[4]

$$\begin{bmatrix} \cosh \theta, & Z_0 \sinh \theta \\ \dfrac{\sinh \theta}{Z_0}, & \cosh \theta \end{bmatrix}^n = \begin{bmatrix} \cosh n\theta, & Z_0 \sinh n\theta \\ \dfrac{\sinh n\theta}{Z_0}, & \cosh n\theta \end{bmatrix}$$

and performing the multiplication after setting $I_0 = 0$,

$$\frac{E_n}{E_0} = \cosh n\theta + \frac{R}{2Z_0} \sinh n\theta$$

$$\frac{I_n}{E_0} = \frac{\sinh n\theta}{Z_0}.$$

After substituting for Z_0 from (13) and simplifying:

$$\frac{E_n}{E_0} = \frac{\sinh (n+1)\theta}{\sinh \theta} - \frac{\sinh n\theta}{\sinh \theta} \tag{14}$$

$$G_n = \frac{I_n}{E_n} = \frac{pC \sinh n\theta}{\sinh (n+1)\theta - \sinh n\theta} \tag{15}$$

where

$$\theta = \cosh^{-1}\left(1 + \frac{pT}{2}\right).$$

This represents an alternate and compact solution[5] of the problem in terms of trigonometric functions of the iterative transfer constant of the RC section, which is also the image transfer constant of the full T section.

The same results can be obtained without the artifice of inserting a resistor $R/2$ in series with the output terminal. The transmission matrix of the RC section is expressed as a similarity transformation of the diagonal matrix containing its latent roots, in which form it is

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[3] F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Co., Inc., New York, N. Y., p. 194; 1943. Equations (75a) and (76); or directly from open circuit conditions, which determine the general circuit parameters A and $C(A = \cosh \theta, C = \sinh \theta/Z_0)$ in terms of R and C.

[4] L. A. Pipes, "Applied Mathematics for Engineers and Physicists," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 264–265; 1946.

[5] Reduces to the polynomial form, if desired, since

$$\frac{\sinh n\theta}{\sinh \theta} = \sum_{n} \binom{n+m}{2m+1} (pT)^m$$

by expanding in terms of

$$2 \sinh \frac{\theta}{2} = \sqrt{pT}$$

according to #3.173 (modified), Smithsonian Mathematical Formulae, Smithsonian Institution, p. 67; 1939.

raised to the n'th power.[6] This process, however, involves more advanced concepts of the matrix calculus. Also, the solution in terms of hyperbolic functions can be obtained from the finite difference equation (5) by conventional methods, but requires more algebraic manipulation in that case.

CONCLUSIONS

It may be concluded that methods for the solution of the given problem are available which take nothing for granted but Ohm's and Kirchhoff's laws, and yet arrive at the goal with no more, or even less, algebraic manipulation than the "postulatory" method in the paper under discussion. Two of the possible methods have been presented in this note.

The finite difference method leads to the solution almost immediately. Its fundamental recursion formula (5) follows directly from the basic relation $\sum i=0$ at a network junction, which can be formulated by inspection. A few more steps lead to explicit expressions for the coefficients $a_{m,n}$ and $b_{m,n}$, although, in general, solution by actual recursion according to (5) would be adequate. The second method, using image parameters suitable for cascade connection of symmetrical structures, may be even more attractive to the communications engineer.

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[6] The first method is closely related, in mathematical terms, to a similarity transformation which equalizes the elements in the principal diagonal.

The Measurement of Antenna Impedence Using a Receiving Antenna*

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AND D. G. WILSON§, MEMBER, IRE

Summary—**Energy from a remote transmitter excites a receiving antenna that is erected vertically over a large conducting plane and base-loaded by a vertical slotted coaxial cavity of variable length. From measurements of the location and half-power width of resonance curves, the combined phase and damping factors for the two ends of the cavity are determined. By measuring these factors for the lower end of the cavity separately, those of the upper end are determined and used to calculate the impedance of the antenna. In effect, the receiving antenna is the generator driving the coaxial line, and it is the impedance of this generator that is measured. Curves of the measured impedance as a function of the electrical length of the antenna are given. Excellent agreement is obtained between impedances measured in this manner for the receiving antenna and corresponding impedances of the same antenna when base-driven through the slotted section. Both sets of measurements are in good agreement with the King-Middleton second-order theory.**

IS THE IMPEDANCE of an antenna that is loaded and used for reception equal to the impedance of the same antenna when used for transmission with the load replaced by a generator? The answer is simple. Yes, the impedances necessarily are equal if they are *defined to be the same*, and this is both useful and conventional. This is done in terms of an "equivalent" series circuit in which the receiving antenna is replaced by a concentrated emf V and a lumped internal impedance Z_0 in series with the lumped load Z_L. This circuit is rigorously equivalent to the antenna with load for the current I_0 entering and leaving the load, as is readily established using Thévenin's Theorem.[1] In the symmetrical, center-loaded antenna in Fig. 1(a) the open circuit voltage across AB as in Fig. 1(b), due to the action of the electric field, is $V=V_{AB}$ (open). Thévenin's Theorem states that I_0 in Fig. 1(a) is the same as I_0 in Fig. 1(c) if $V=V_{AB}$ (open) and Z_0 is the impedance looking into the terminals AB in Fig. 1(b) with the electric field E equal to zero. Hence,

$$I_0 = V/(Z_0 + Z_L). \tag{1}$$

Note that by definition Z_0 is the impedance of the antenna *as if driven by a potential difference across its terminals*. Hence, it is identically the transmitting impedance.

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[1] Cruft Electronics Staff, "Electronic Circuits and Tubes," McGraw-Hill Book Co., New York, N. Y., p. 110; 1947.

V in (1) is that concentrated voltage which would maintain the same current I_0 in the load Z_L when connected in series with the lumped impedance Z_0 of the antenna as exists by action of the electric field along the loaded antenna. $V = V_{AB}$ (open) is defined in the literature[2,3] in terms of the electric field E and the dimensions and orientation of the antenna. Note that the simple circuit in Fig. 1(c) is equivalent to the actual antenna *only* in the determination of I_0. If the voltage V is connected in series with Z_L and the antenna, I_0 is the same as in the receiving antenna if V is properly defined, but I_z elsewhere on the antenna is not.

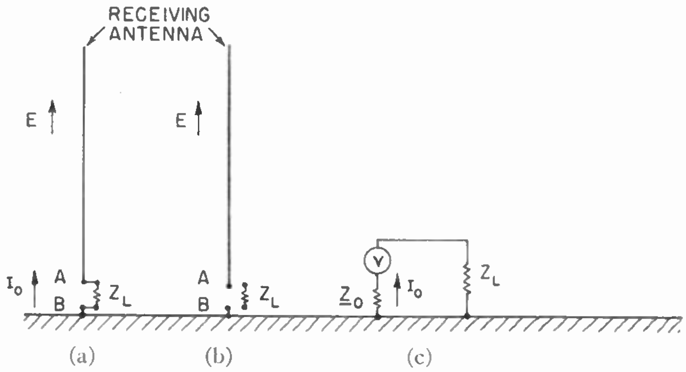

Fig. 1—Equivalent circuit for I_0 in receiving antenna.

The impedance of the receiving antenna, defined as Z_0 in (1), may be measured with the apparatus shown in Fig. 2. Z_L is the input impedance of a vertical section of slotted line terminated in a sensitive detector and a piston, and the antenna is the vertical extension of the inner conductor above a large ground screen. V is the equivalent concentrated generator driving the line, and Z_0 is the internal impedance of the generator, i.e., the impedance of the antenna. Thus, the experimental problem is to measure the internal impedance Z_0 of the generator driving the line. This can *not* be done using either the conventional standing-wave-ratio method[4] or the distribution-curve method[4] since these involve only the load impedance. However, Z_0 is readily determined using the resonance-curve method.[4,5] It is necessary merely to determine the position and half-power width of the resonance curve obtained by moving the piston terminating the line for each length h of the antenna.

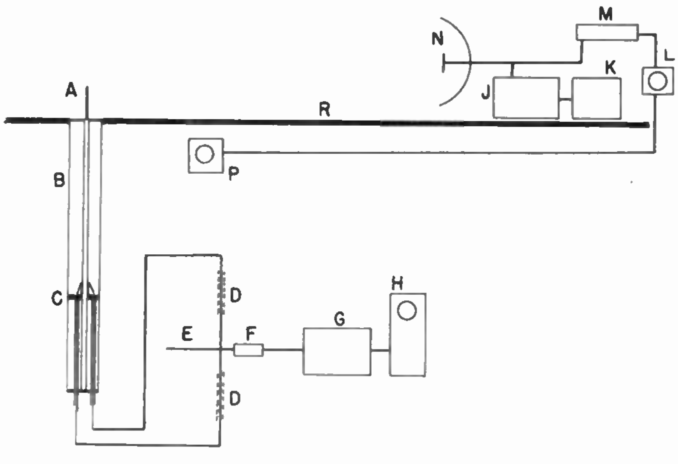

Fig. 2—Schematic diagram of complete equipment.

A = receiving antenna
B = measuring line
C = movable piston
D = double line stretcher
E = shunt stub
F = bolometer mount
G = tuned bolometer amplifier
H = Ballantine voltmeter
J = transmitter
K = modulator
L = wavemeter crystal current
M = wavemeter
N = transmitting antenna
P = remote crystal current
R = ground plane.

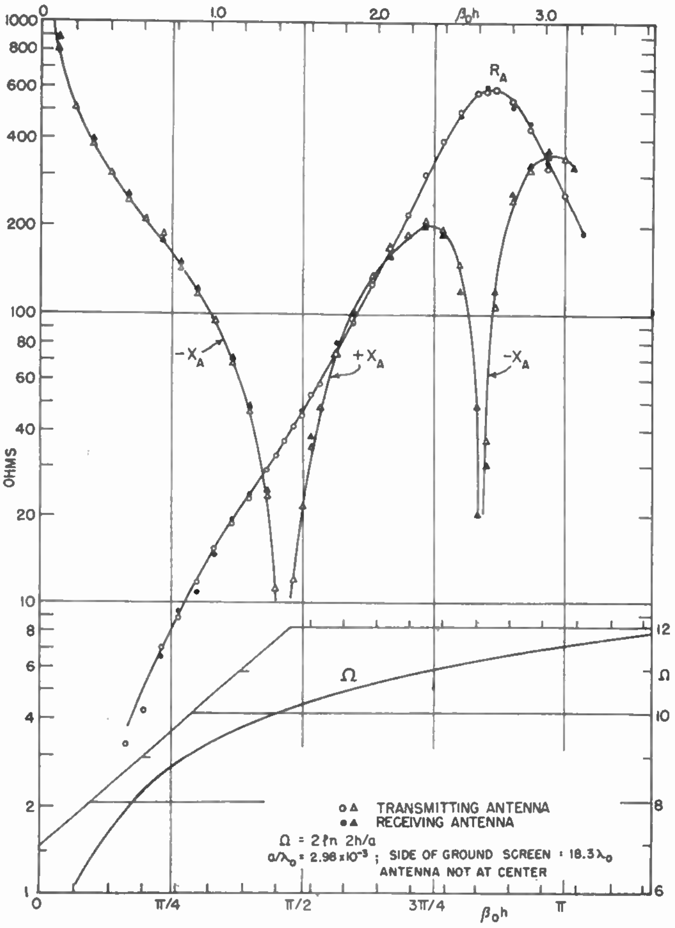

Fig. 3—Impedance of antenna by reception and transmission methods.

The resonance-curve method is expressed concisely using the terminal functions ρ_o and Φ_o of Z_0 at $z=0$, and ρ_s and Φ_s of Z_s at the end $z=s$ of the coaxial line with characteristic impedance $Z_c \doteq R_c$. These functions are defined by

$$Z_0 = Z_c \coth(\rho_o + j\Phi_o); \quad Z_s = Z_c \coth(\rho_s + j\Phi_s). \quad (2)$$

The over-all phase function $(\beta s + \Phi_o + \Phi_s)$ is determined directly from the resonant length of the line; the over-all attenuation function $(\alpha s + \rho_o + \rho_s)$ from the half-power width, using

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[2] R. King and C. W. Harrison, Jr., "The receiving antenna," Proc. I.R.E., vol. 32, pp. 18–49; January, 1944.

[3] R. W. P. King, H. R. Mimno, and A. H. Wing, "Transmission Lines, Antennas, and Wave Guides," McGraw-Hill Book Co., New York, N. Y., p. 164; 1945.

[4] D. D. King, "Impedance measurements on transmission lines," Proc. I.R.E., vol. 35, pp. 507–514; May, 1947.

[5] R. King, "Transmission-line theory and its application," Jour Appl. Phys., vol. 14, pp. 577–600; November, 1943.

$$B = \beta s_n + \Phi_0 + \Phi_s = n\pi; \quad n = 1, 2, \cdots, \tag{3}$$

$$A = \alpha s_n + \rho_o + \rho_s = \beta W/2, \tag{4}$$

where α and $\beta = 2\pi/\lambda$ are, respectively, the attenuation and phase constants of the line, W is the full width of the resonance curve, and s_n is the length of the line at resonance. By predetermining Φ_s and ρ_s for the piston with its two detecting loops, measuring s directly, obtaining β from the measured wavelength, and computing the small quantity α from the dimensions and material of the line, ρ_o and Φ_o may be evaluated from (3) and (4).

$Z_0 = R_0 + jX_0$ is determined from curves of ρ_o and Φ_o, using

$$Z_0 = R_0 + jX_0 = \frac{R_c[\sinh 2\rho_o - j \sin 2\Phi_o]}{\cosh 2\rho_o - \cos 2\Phi_o}. \tag{5}$$

Since ρ_o and Φ_o are much more slowly varying than R_o and X_o, it is more accurate to draw smooth curves of ρ_o and Φ_o through the experimental points and use these to determine R_o and X_o than to substitute the experimental values of ρ_o and Φ_o directly in (5). Experimental curves of R_o and X_o are in Fig. 3. In Figs. 4 and 5 experimental points are compared with theoretical curves of the King-Middleton second-order theory.[6]

The vertical slotted line used in the measurements is that described by D. D. King.[7] It is provided with a long taper so that the gap may be kept sufficiently small to make terminal effects negligible. Styrofoam was used for the single insulator at the upper end of the line. The ground screen was a 36-foot square of sheet aluminum. The operating frequency was 500 mc.

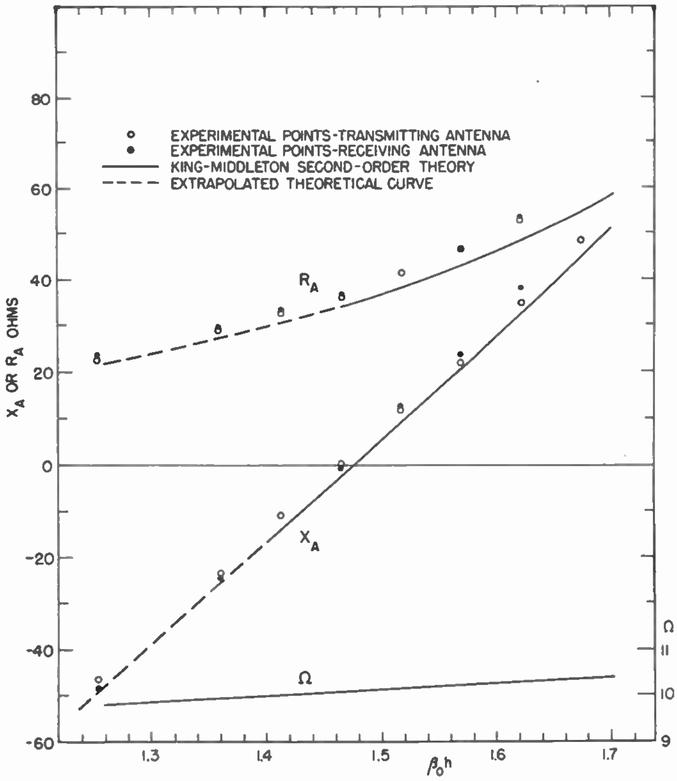

Fig. 4—Theoretical impedance near resonance with experimental points from Fig. 3.

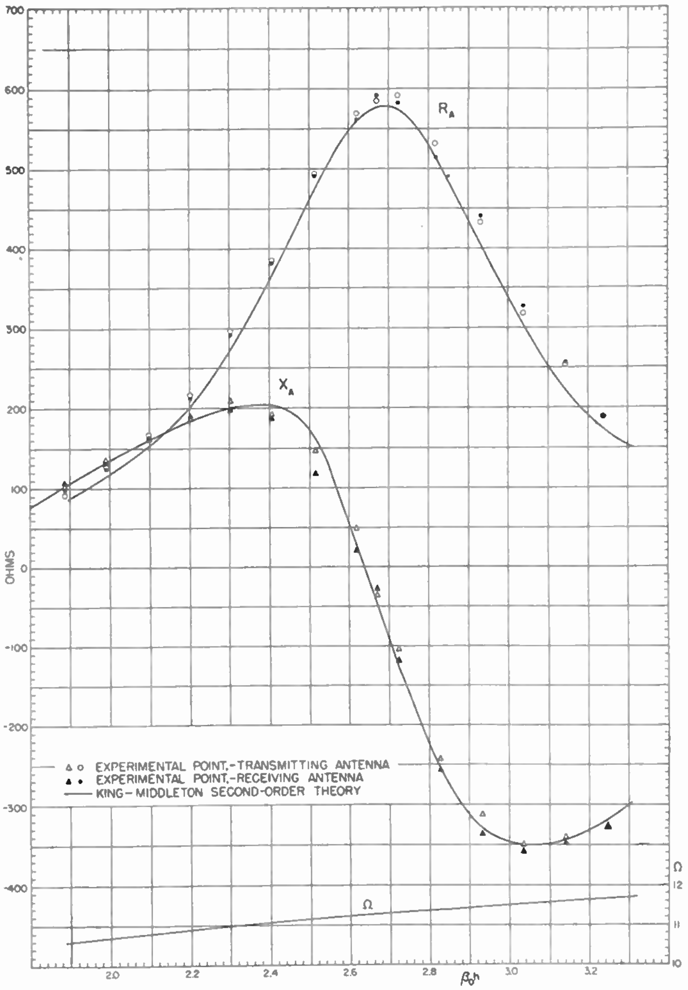

Fig. 5—Theoretical impedance near antiresonance with experimental points from Fig. 3.

The method used to determine ρ_s was to replace the receiving antenna by a metal cap, excite the line by a loosely coupled generator, and measure ρ_s with an auxiliary probe when $\rho_o = 0$. Φ_s, for the piston at the lower end of the line was determined by locating the minimum in the distribution curve using an auxiliary probe.

Although, in general, the measurement of the impedance of an antenna is more convenient when it is driven than when used for reception, since the standing-wave-ratio and distribution-curve methods are available, the receiving-antenna method has the advantage of requiring no transmitter at the location of the measurement. Thus, for example, only light test equipment is needed for measuring the impedance of an antenna in an aircraft or vehicle in motion, using a slotted line with predetermined ρ_s and Φ_s and a signal from a more or less distant high-powered transmitter.

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[6] R. King and D. Middleton, "The cylindrical antenna, theory and experiment," *Quart. Appl. Math.*, vol. 3, pp. 302–335; January, 1946; also *Jour. Appl. Phys.*, vol. 17, pp. 273–284; April, 1946.

[7] D. D. King, "The measured impedance of cylindrical dipoles," *Jour. Appl. Phys.*, vol. 17, pp. 844–852; October, 1946.

Contributors to Proceedings of the I.R.E.

D. A. Alsberg

D. A. Alsberg (A'46–M'48) was born in Kassel, Germany, on June 5, 1917. He obtained his undergraduate training at the Technische Hochschule in Stuttgart, Germany, and completed his work there in 1938. After coming to the United States, he was engaged in graduate study at the Case School of Applied Science (now Case Institute of Technology) from 1939 to 1940. Following three years as development engineer with several companies in Ohio, he entered the United States Army and served at the Aberdeen Proving Ground, in Maryland and in Europe.

In 1945, Mr. Alsberg joined Bell Telephone Laboratories, Inc., where he has been concerned with phase, transmission, and related measurements problems in connection with coaxial-cable carrier and microwave radio-relay systems development.

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Leo L. Beranek

Leo L. Beranek (S'36–A'41–SM'45) was born in Solon, Iowa, on September 15, 1914. He received the B.A. degree, "with distinction," from Cornell College, Iowa, in 1936, the M.S. and D.Sc. degrees from Harvard University in 1937 and 1940, respectively, and an honorary D.Sc. degree from Cornell College in 1946.

Preceding his appointment to the Massachusetts Institute of Technology staff, as a physics research associate, in February, 1946, Dr. Beranek was associated professionally with Harvard University. From 1937 to 1946, he served successively as a research assistant, instructor, director of sound-control research, and director of Electroacoustic and Systems Research Laboratories. In February, 1947, Dr. Beranek was appointed associate professor of communications engineering in the Department of Electrical Engineering at M.I.T., and technical director of the Acoustics Laboratory.

Dr. Beranek is the author of two books and numerous papers and articles. In 1944, the Acoustical Society of America awarded him their biennial award for outstanding contributions to acoustics. In recognition of "outstanding services to his country," Dr. Beranek received the President's Certificate of Merit, in October, 1948.

Dr. Beranek is a fellow of the American Physical Society, a fellow and associate editor of the Acoustical Society of America, a member of the American Association for the Advancement of Science, and a member of Sigma Xi and Eta Kappa Nu. He has also served as national chairman of the Professional Group on Audio of the IRE, as vice-president of the Acoustical Society of America, and as chairman of the Acoustics Section Z-24 of the American Standards Association. He is now chairman of the Acoustics Panel of the Research and Development Board of the Department of Defense.

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J. T. Bolljahn

J. T. Bolljahn (A'43) was born in Oakland, Calif., in 1918. He received the B.S. and Ph.D. degrees from the University of California in 1941 and 1950, respectively. From August, 1941 until January, 1946, he was employed by the Naval Research Laboratory in Washington, D. C. His work in this position was concerned with the development of aircraft and shipboard antennas. From February, 1946, until September, 1949, he was a member of the staff of the University of California Antenna Laboratory.

Dr. Bolljahn joined the staff of the Stanford Research Institute as a senior research engineer in September, 1949. He has recently received a part-time appointment as acting associate professor in the electrical engineering department of Stanford University for the Summer Quarter, 1951. He is a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

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Seymour B. Cohn

Seymour B. Cohn (S'41–A'44–M'46) was born at Stamford, Conn., on October 21, 1920. He received the B.E. degree in electrical engineering from Yale University in 1942, also the M.S. degree in communication engineering in 1946, and the Ph.D. degree in engineering sciences and applied physics in February, 1948, from Harvard University. From 1942 through 1945, Dr. Cohn was employed as a special research associate by the Radio Research Laboratory of Harvard University. During part of this time, he represented that Laboratory as a Technical Observer with the United States Army Air Force in the Mediterranean Theater of Operations.

Since March, 1948, Dr. Cohn has been employed by the Sperry Gyroscope Company as a project engineer in the microwave department. He is a member of Tau Beta Pi and an associate member of Sigma Xi. He is now serving on the Papers Review Committee of the IRE.

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Ramsay P. Decker

Ramsay P. Decker was born on March 26, 1926, in Chicago, Illinois. He served in the United States Army in Italy from January, 1945, to June, 1946. In December, 1948, he received the B.S. degree in electrical engineering from the Technological Institute of Northwestern University.

After graduation, Mr. Decker joined the Research Division of the Collins Radio Company, where he has been engaged in work on high-power vacuum-tube control circuitry, propagation-data analysis, and uhf airborne transmitter development.

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Milton Dishal

Milton Dishal (A'41–SM'46) was born on March 20, 1918, in Philadelphia, Pa. He received the B.S. degree from Temple University in 1939, and the M.A. degree in physics in 1941. From 1939 to 1941, Mr. Dishal was a teaching fellow in physics at Temple University.

In 1941 he entered the employ of Federal Telecommunications Laboratories, where he is now a senior project engineer, engaged in the development of radio receivers having special characteristics.

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W. G. Findley

Warren G. Findley was born in New York, New York, on September 23, 1906. He received the A.B. degree, with high honors in mathematics, from Princeton University in 1927. In 1929 and 1933, respectively, he received the degrees of M.A. and Ph.D. in educational psychology and statistics from Columbia University.

From 1927 through 1938, Dr. Findley served as College Personnel

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Officer at Cooper Union Institute of Technology, during which period he also developed aptitude tests for the College Entrance Examination Board. From 1938 through 1946, he was Assistant Director of Examinations and Testing in the New York State Education Department. He interrupted this work for 3 months in 1944 to supervise the construction of tests for the U. S. Armed Forces Institute at the University of Minnesota. In the capacity of Chief of the Evaluation Branch in the Educational Advisory Staff, he guided the development of tests at the Air University, Maxwell Air Force Base, Alabama, from 1946 through 1948. Since 1948 he has been Director of Test Development for the Educational Testing Service in Princeton, N. J., where a number of the tests and testing programs mentioned in his article, "Using Tests to Select Engineers," are built.

Dr. Findley is a fellow of the American Psychological Association and the American Association for the Advancement of Science, and a member of the American Educational Research Association, the Psychometric Society, and Phi Beta Kappa. He has contributed articles on testing to a number of professional journals and yearbooks.

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Irvin H. Gerks (A'32–M'41–SM'43) was born in New London, Wis., in 1905. He received the B.S. degree from the University of Wisconsin in 1927, and the M.S. degree from the Georgia School of Technology in 1932, both in electrical engineering.

Irvin H. Gerks

From 1927 to 1929, Mr. Gerks was employed by Bell Telephone Laboratories, where he developed and tested various special equipment for telephone centrals, such as tone signaling and centralized time announcing. Following this period, he became an instructor in and, later, an assistant professor of, communication and electronic engineering at the Georgia School of Technology, where he remained until 1940.

During the war years, Mr. Gerks was an officer in the Army Signal Corps. He was stationed first at Wright Field and later transferred to the Pacific Theater. At Wright Field he was in charge of the Communication and Navigation Division of the Aircraft Radio Laboratory.

At present Mr. Gerks is employed by the Collins Radio Company, where he is engaged in radio wave-propagation studies.

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John Van Nuys Granger (S'42–A'45–M'46) was born in Iowa, in 1918. He received the A.B. degree from Cornell College in 1941, and the M.S. and Ph.D. degrees from Harvard University in 1942 and 1948, respectively. During a part of 1942, he was a member of the staff of the Pre-Radar School at Cruft Laboratory, Harvard University, Cambridge, Mass. In November, 1942, he joined the Radio Research Laboratory of Division 15, OSRD, where he remained until 1945. During that interval he served with the American British Laboratory in Great Malvern, Worcestershire, England, and as a technical observer with the U. S. Air Forces in France and the Low Countries.

J. V. N. Granger

In 1945 he joined the staff of the Central Communications Research Laboratories of Division 13, OSRD, at Harvard, leaving in 1946 to resume his studies. In 1947 he became group leader of the Antenna Group at the Cruft Laboratory. In May, 1949, he was named supervisor of the Aircraft Radiation Systems Laboratory, Stanford Research Institute, Stanford, Calif. He has recently received a part-time appointment as acting associate professor in the electrical engineering department of Stanford University for the Summer Quarter, 1951.

Dr. Granger is a member of the Panel on Antennas and Propagation of the Research and Development Board, the American Physical Society, Sigma Xi, and also of Commission 6 of the U.S.A. National Committee of the URSI.

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Elmer O. Hartig (S'47) was born in Evansville, Ind., on January 28, 1923. He received the B.S. degree in electrical engineering from the University of New Hampshire in 1946. He attended Harvard University in the Department of Engineering Sciences and Applied Physics, where he received the S.M. degree in 1947, and the Ph.D. in 1950.

Elmer O. Hartig

From 1944 to 1946, while on active duty in the U. S. Army, Dr. Hartig was associated with the Manhattan Project at Columbia University and Los Alamos, N. M. From 1948 to 1950, while at Harvard's Cruft Laboratory, he served as a research assistant, doing research in antennas and pulsed circuits. Dr. Hartig has been associated with the aerophysics department of the Goodyear Aircraft Corporation since July, 1950, where he is heading the microwave and antenna group. He is a member of Sigma Xi.

John Alexander Kessler was born in Buffalo, N. Y., on December 19, 1920. He received the A.B. degree (cum laude) from Harvard College in 1942, in the Department of Music, and the S.M. degree from Harvard University in 1947, in the Department of Engineering Sciences and Applied Physics (acoustics and communications engineering).

J. A. Kessler

Mr. Kessler has been affiliated with the research staff at the Massachusetts Institute of Technology, and with the Electro-acoustic Laboratory at Harvard University. Mr. Kessler is now a staff member of the Division of Industrial Co-operation, and liaison officer of the Acoustics Laboratory at M.I.T.

Mr. Kessler is a member of the Acoustical Society of America, serving on the Music Committee and subcommittee Z 24-X-5 (Measurements of Acoustic Properties of Materials), and is a member of the Society of Sigma Xi.

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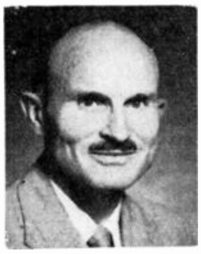
Ronold King (A'30–SM'43) was born on September 19, 1905, at Williamstown, Mass. He received the B.A. degree in 1927, and the M.S. degree in 1929 from the University of Rochester, and the Ph.D. degree from the University of Wisconsin in 1932. He was an American-German exchange student at Munich from 1928 to 1929, a White Fellow in physics at Cornell University from 1929 to 1930, and a Fellow in electrical engineering at the University of Wisconsin from 1930 to 1932. He continued at Wisconsin as a research assistant from 1932 to 1934. From 1934 to 1936, he was an instructor in physics at Lafayette College, serving as an assistant professor in 1937.



Ronold King

During 1937 and 1938, Dr. King was a Guggenheim Fellow at Berlin. In 1938 he became instructor in physics and communication engineering at Harvard University, advancing to assistant professor in 1939 and to associate professor in 1942. He was appointed Gordon McKay professor of applied physics at Harvard University in 1946.

Dr. King is a Fellow in the American Physical Society, the American Association for the Advancement of Science, and the American Academy of Arts and Sciences. He is a member of Phi Beta Kappa and also of Sigma Xi.

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A. J. Lephakis

A. J. Lephakis (A'44–M'50) was born in Glen Cove, L. I., N. Y., on May 14, 1921. Before the war, he attended the Massachusetts Institute of Technology. From 1943 to 1946, he served with the United States Army Air Force. In 1946, he returned to M.I.T., and received the S.B. and S.M. degrees in electrical engineering in 1948 and 1949, respectively. He remained at M.I.T. as a research assistant, and has been associated with the communication group at the Research Laboratory of Electronics since that time.

Mr. Lephakis is a member of Sigma Xi and an associate of the American Institute of Electrical Engineers.

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R. L. Linton, Jr.

R. L. Linton, Jr. (S'41–A'44–M'46) was born on February 10, 1921, in Clemenceau, Ariz. He received the B.S. degree in electrical engineering at the University of California in May, 1942. He was employed by the Navy Department from then until April of 1944. For the Navy Department he was engaged in ultra-high-frequency antenna development at the Naval Research Laboratory, and in microwave airborne-radar and test-equipment design at the Bureau of Ships. Mr. Linton then became connected with the University of California at Berkeley. Here he worked at the Radiation Laboratory until June, 1945, when, at the formation of the Antenna Laboratory, he joined that group to be occupied with instrument and antenna development. Concurrently with his activity at the Antenna Laboratory, Mr. Linton was also a part-time lecturer for the Electrical Engineering Division from the fall of 1946 till mid-1949.

Mr. Linton is now an electronics engineer in the Electronics and Guidance Section of Consolidated-Vultee Aircraft Corporation at San Diego, Calif. He is a member of the American Association for the Advancement of Science and of the Tau Beta Pi Association, and an associate of the Society of Sigma Xi.

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J. F. Marshall

John F. Marshall was born on December 21, 1918, in New York City. He received the A.B. degree from Swarthmore College in 1941, and the Ph.D. degree from the University of Notre Dame in 1949.

Dr. Marshall joined the staff of the Bartol Research Foundation as a physicist in 1942. As a research assistant from 1946 to 1949, and an instructor from 1949 to 1950 at the University of Notre Dame, he specialized in theoretical nuclear physics.

At present Dr. Marshall is engaged in the study of the theory of secondary electron emission at the Bartol Research Foundation.

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Theodore Miller

Theodore Miller (S'40–A'42–M'48) was born in Austria, on February 3, 1917. He received the B.S. degree in electrical engineering from the Carnegie Institute of Technology in 1946 and the M.S. degree in mathematics from the University of Pittsburgh in 1948. From 1941 to 1944 he was employed by the C. L. Hofmann Corporation of Pittsburgh as an electronics engineer, in which capacity he did development work on hearing-aid devices. Since 1944 he has been associated with the Westinghouse Research Laboratories, where he has been concerned with microwave research, radar and sonar engineering, and, at present, color-television research. He is the author of several technical papers on microwave apparatus and low-frequency instruments.

Mr. Miller is a member of the American Physical Society.

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J. R. Moore

J. R. Moore was born on July 5, 1916, in Saint Louis, Mo. After receiving the B.S. degree in mechanical engineering from Washington University in Saint Louis, in 1937, Mr. Moore joined the General Electric Company, completing the three-year course in advanced engineering in 1940. Continuing in the employ of General Electric until 1946, he became assistant head of the engineering mechanics group of the aeronautics and marine engineering division, and head of the theoretical section, project Hermes, respectively.

Mr. Moore was responsible for the development of wartime airborne and naval-gun fire-control equipment and of automatic production machinery, and worked on simulation problems for guided missiles. He was associate professor of mechanics, and director of the Dynamical Control Laboratory at Washington University from 1946 until February, 1948, when he moved to California to join the North American Aviation, Inc., Aerophysics Laboratory staff, where he is now assistant chief of guidance.

Mr. Moore is a member of two RDB groups. Most of his work has been in the fields of computing mechanisms, three-dimensional dynamics, and automatic control. Since the fall of 1948, he has also been a visiting associate professor of engineering at the University of California at Los Angeles, Calif., where he teaches a graduate course in advanced servomechanism theory.

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T. Morita

Tetsu Morita (S'44–A'49) was born in Seattle, Wash., on February 5, 1923. He attended the University of Washington from 1940 to 1942, and received the B.Sc. degree in electrical engineering from the University of Nebraska in 1944. He received the M.S. and Ph.D. degrees from Harvard University in 1945 and 1949, respectively.

During 1944 Dr. Morita assisted in the Army Specialized Training Program at the University of Nebraska. From 1946 to 1947 he was a teaching fellow at Harvard University, where he was a research assistant during the period from 1947 to 1949. Dr. Morita has been a research fellow in electronics at Harvard since 1949, and he is, at present, group leader of the Antenna Group at Cruft Laboratory.

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M. A. Pomerantz

Martin A. Pomerantz was born on December 17, 1916, in New York City. He received the A.B. degree from Syracuse University in 1937, the M.S. degree from the University of Pennsylvania in 1938, and the Ph.D. degree from Temple University in 1951.

Dr. Pomerantz joined the staff of the Bartol Research Foundation as research assistant in 1938. In 1941, he was promoted to Research Fellow. At present he is engaged, as a physicist, in the field of cosmic radiation, as well as that of secondary electron emission.

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As a leader of a number of National Geographic Society cosmic-ray expeditions, including several to Hudson Bay, Dr. Pomerantz has conducted cosmic-ray experiments at very high altitudes with instruments carried aloft by free balloons.

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William H. Radford (A'41–SM'45) was born in Philadelphia, Pennsylvania, on May 20, 1909. He received the B.S. degree in electrical engineering from the Drexel Institute of Technology in 1931, and the M.S. degree from the Massachusetts Institute of Technology in 1932. He has been a member of the staff of the department of electrical engineering at M.I.T. since 1932, and is now Professor of Electrical Communications there.

W. H. Radford

Professor Radford has been active as a consultant to government and industry on radio-communication facilities and specialized electronic applications since 1935. He served as a section member and consultant to the National Defense Research Committee, from 1940 to 1943. In 1941, he assisted in establishing the M.I.T. Radar School, which later became the principal center for training electronics specialist radar officers for the United States Navy. He was closely associated with the radar school throughout its period of operation, and became associate director in 1944. He is now devoting a large portion of his time to supervision of government-sponsored research programs in radio communications.

Professor Radford is a registered professional engineer in the Commonwealth of Massachusetts. He is a member of the American Institute of Electrical Engineers, the American Society of Engineering Education, the American Association for the Advancement of Science, Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

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William M. Sharpless (A'28–M'38–SM'43) was born on September 4, 1904, in Minneapolis, Minn. He received the B.S. degree in electrical engineering in 1928 and the E.E. degree in 1951, both from the University of Minnesota. As a member of the technical staff of the Bell Telephone Laboratories since 1928, he has been engaged in radio research projects.

W. M. Sharpless

Mr. Sharpless' most recent work has had to do with the studies of the angle of arrival of microwaves, and the design of artificial dielectrics and microwave antennas.

Mr. Sharpless is a member of the American Physical Society, and is a licensed professional engineer in the State of New Jersey.

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Leo Storch was born on March 3, 1921, in Vienna, Austria. In January, 1944, he received the B.E.E. degree, cum laude, from the School of Technology, College of the City of New York. He was awarded the M.S. degree by the Graduate School of Stevens Institute of Technology in June, 1947.

Leo Storch

From 1944 to 1947, Mr. Storch worked as a development engineer with the Western Electric Company, Kearny, N. J. He was engaged in the circuit design for a wide variety of testing and measurement equipment. Subsequently, he transferred to RCA Victor, Camden, N. J., as advanced development engineer in the aviation radio department, where he was primarily concerned with the development of antenna impedance-matching networks and automatic tuning systems for wide-band aircraft communications transmitters.

Mr. Storch recently joined the Research and Development Laboratories of the Hughes Aircraft Company, Culver City, Calif., as research physicist in the communications group.

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James E. Storer was born in Buffalo, N. Y., on October 26, 1927. He received the A.B. degree in physics in June, 1947, from Cornell University, followed by the M.S. and Ph.D. degrees in in 1948 and 1951, respectively, both from the Department of Engineering Sciences and Applied Physics at Harvard University. While at Harvard, Mr. Storer held an Atomic Energy Commission Fellowship. He is a member of the Sigma Xi Society.

James E. Storer

Mr. Storer is at present engaged in research on thin wire antennas at the Electronics Research Laboratory of Harvard University.

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Jerome Bert Wiesner (S'36–A'40–SM'48) was born on May 30, 1915, in Detroit, Michigan. He received his B.Sc. and M.S. degrees from the University of Michigan in 1937 and 1940, respectively. In 1950 he was granted his Doctor of Philosophy degree from the University of Michigan.

J. B. Wiesner

Dr. Wiesner was chief engineer of the Acoustical and Record Laboratory of the Library of Congress from 1940 to 1942, at which time he became a member of the staff at the Radiation Laboratory at the Massachusetts Institute of Technology. In 1945, he went to the Los Alamos Laboratory in New Mexico as a member of the staff, returning to M.I.T. in 1946. He has been assistant director of the Research Laboratory of Electronics at M.I.T. for the past three years, and is now its associate director. In addition to his work in the laboratory, he is professor of electrical engineering at M.I.T.

Dr. Wiesner is a member of the Acoustical Society of America, the Federation of American Scientists, and the American Association for the Advancement of Science. In addition, he is a member of Eta Kappa Nu, Sigma Xi, and Phi Kappa Phi.

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Donald G. Wilson (S'38–A'40–M'48) was born in Bridgeport, Conn., on September 20, 1917. He received the B.E.E. degree from Rensselaer Polytechnic Institute in 1938, and the S.M. degree in communication engineering from Harvard University in 1939. After a year in industry, he taught for two years at Rensselaer.

Donald G. Wilson

From 1941 to 1945, Mr. Wilson was engaged in microwave propagation research at the Radiation Laboratory of the Massachusetts Institute of Technology. In 1945 he returned to Harvard University for graduate study, and received the Ph.D degree in 1948.

He joined the department of electrical engineering at the University of Kansas in 1947 as an associate professor. In 1948 he was appointed chairman of the department of electrical engineering, and was advanced to the rank of professor in 1950.

Dr. Wilson is a member of the American Institute of Electrical Engineers, Sigma Xi, and Tau Beta Pi.

Institute News and Radio Notes

Calendar of
COMING EVENTS

Third Annual Technical Conference, Kansas City Section, Hotel President, Kansas City, Mo., November 16, 17

First JETEC General Conference, Absecon, N. J., November 29–December 1

IRE Nuclear Symposium, Brookhaven National Laboratory, Upton, L. I., N. Y., December 3, 4

Joint IRE/AIEE Computer Conference, Benjamin Franklin Hotel, Philadelphia, Pa., December 10–12

Symposium on Williams Electrostatic Storage, National Bureau of Standards, Washington, D. C., December 13–14

IAS-ION-IRE-RTCA Conference on Air Traffic Control, Astor Hotel, New York, N. Y., January 30

1952 IRE National Convention, Waldorf-Astoria Hotel and Grand Central Palace, New York, N. Y., March 3–6

IRE National Conference on Airborne Electronics, Hotel Biltmore, Dayton, Ohio, May 7–9

4th Southwestern IRE Conference and Radio Engineering Show, Rice Hotel, Houston, Tex., May 16–17

TECHNICAL COMMITTEE NOTES

The August meeting of the **Standards Committee** was preceded by a meeting of the Administrative Committee. A. G. Jensen, presided as Chairman at both meetings. W. H. Pease was named as Chairman of the new **IRE Technical Committee on Servo-Systems**, recently organized, and W. D. Goodale, Jr., was appointed Chairman of the **IRE Technical Committee on Electroacoustics.** C. H. Page opened the question as to what the Committee's opinion was on the use of formulas in definitions. It was pointed out that some terms are inherently mathematical and could be defined best by giving a formula, although the tendency in the past has been to avoid this use. The Committee decided that it is better to use formulas, rather than using long, wordy, and often unclear, definitions. The Standards Manual will be revised shortly and this concept will be reflected in the Manual. Other additions to the Manual will be welcomed by Messrs. Weber, Baldwin, and Jensen.

The **Standards Committee** which convened on September 13, under the Chairmanship of A. G. Jensen, recommended for publication a paper on, "Fundamental Considerations Regarding the Use of Relative Magnitudes," by J. W. Horton. Dr. Horton is the Chief Consultant of the U. S. Navy Underwater Sound Laboratory, New London, Conn. His paper will be considered for an early publication in the PROCEEDINGS, by the editorial department.

The Committee on Antennas and Waveguides held a meeting on September 11, under the Chairmanship of Gardner Fox. This Committee is still engaged in the preparation of definitions which will be submitted to the Standards Committee for approval.

Consideration is being given to the necessity for broadening the scope of the **Mobile Communications Committee** and to the possibility of changing the name of the Committee to include land-mobile, air-mobile, and sea-mobile communications.

A Task Group of the **Receivers Committee** has prepared a paper on Methods of Calibration of Radiation.

Work is underway towards the preparation of the Report of the **Annual Review Committee.** Ralph Batcher, Chairman of the Annual Review Committee has requested that each Technical Committee Chairman appoint a member to the Annual Review Committee. The deadline for the receipt of material for the Annual Review Report is November 15.

A meeting of the **Measurements and Instrumentation Committee** was held on Sept. 14, F. J. Gaffney, Chairman. The Chairman announced that it would not be necessary to reactivate the **Subcommittee on Basic Techniques of Instrumentation** as the work of this Subcommittee has been taken over by the **Subcommittee on Basic Standards and Calibration Methods (25.1).** J. L. Dalke, Chairman of **Subcommittee 25.2, Dielectric Measurements,** reported that his Subcommittee is endeavoring to extend the frequency range presently defined and will redefine a series of terms on the low-frequency range. Definitions are needed that will extend over the entire frequency range, realizing of course that different methods of measurement will be required for the various frequency ranges. This Subcommittee 25.2 is also planning to review standards of other organizations and, if possible, incorporate them into IRE Standards. A report of the work in progress in **Subcommittee 25.3, Magnetic Measurements,** was presented by F. J. Gaffney in the absence of the Chairman, C. D. Owens.

Subcommittee 25.4, Audio-Frequency Measurements, under the Chairmanship of Dr. Peterson is collecting material for use in their standardizing work. **Subcommittee 25.11, Statistical Quality Control,** under the Chairmanship of Mr. Steen is endeavoring to prepare a bibliography of articles and information which will be helpful to the field of Quality Control. Another project of this Subcommittee is to make a survey of the field of other committees, subcommittees, and organizations toward the end of avoiding duplication of effort.

A. V. Loughren, Vice-President in Charge of Research, Hazeltine Electronics Corporation, has been appointed to replace Haraden Pratt on the **Joint Technical Advisory Committee.**

The IRE-IAS-ION-RTCA Symposium on Electronics in Aviation will be held on January 30, 1952 in New York City. Details will be announced as they are formulated.

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FEEDBACK CONTROL CONFERENCE PLANNED

A special two-day conference on major developments in feedback control systems will be held December 6–7, at the Chalfonte-Haddon Hall Hotel, Atlantic City, N. J., it has been announced. The meeting will be sponsored by the American Institute of Electrical Engineers Committee on Feedback Control Systems.

Not only will developments of the past few years be reviewed during the conference but the program provides for covering many new, significant advances in design and components, according to Jerome Zauderer, chairman of the information committee.

Of special interest at the Atlantic City meeting will be a session on biological servomechanisms. Topics to be discussed also will include the role of statistics, discontinuous control systems, control systems operating from discrete data inputs, and a sampled data control system technique.

Scheduled to appear on the program are scientists and engineers from Massachusetts Institute of Technology, Columbia University, Minnesota, University of Connecticut, and representatives of several large industrial organizations.

The program and other information can be obtained by writing to Mr. Jerome Zauderer, American Measuring Instruments Corporation, 21–25 44th Avenue, Long Island City, N. Y.

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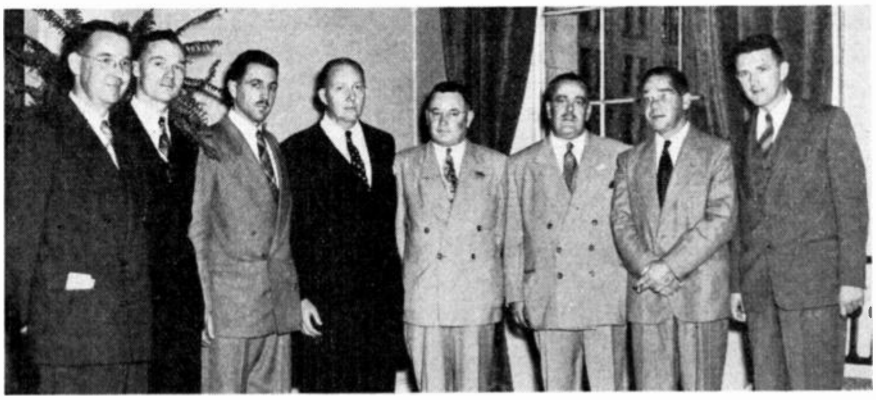
HOUSTON SECTION TO SPONSOR SOUTHWESTERN IRE CONFERENCE

The Fourth Southwestern IRE Conference and Radio Engineering Show will be held in the Rice Hotel, at Houston, Texas, May 16 and 17, 1952. The Houston Section will act as host and sponsor.

Provisions have been made for the holding of all meetings, technical sessions, and equipment exhibits on one floor. Space will be available also in case of simultaneous technical sessions. Equipment exhibitors desiring to demonstrate sound producing apparatus will have isolated rooms on the same floor at their disposal.

The Houston Section's publication *Scope* plans two special Conference issues for the months of March and May, to be given region distribution.

President Coggeshall Visits Vancouver Section



A most cordial reception was given President I. S. Coggeshall by the Vancouver Section of the IRE, during a recent trip to the West Coast. Section officers meeting with President Coggeshall are (left to right): Lorne Kersey, Student Branch Co-ordinator; J. S. Gray, Membership Committee; A. H. Gregory, Vice-Chairman; President I. S. Coggeshall; G. C. Chandler, Chairman; D. D. Carpenter, Secy-Treas.; B. R. Tupper, Junior Past Chairman; and Miles Green, Asst. Secy-Treas.

Professional Group Notes

The **IRE Professional Group on Antennas and Propagation** held a joint meeting with the U. S. National Committee, URSI, on October 8, 9, and 10, at Cornell University, Ithaca, N. Y. C. R. Burrows of the School of Electrical Engineering, Cornell, and A. H. Waynick of Pennsylvania State College took charge of the program which included subjects on Radio Standards, Measurements, Tropospheric Propagation, Astronomy, and Ionospheric Propagation.

The **IRE Professional Group on Broadcast and Television Receivers** held an Administrative Committee meeting during the Radio Fall Meeting at the King Edward Hotel in Toronto, Ontario, October 29, 30, and 31.

The petition for the formation of an **IRE Professional Group on Electronic Computers** was approved by the Committee on Professional Groups at the last meeting. This Group is being promoted on the West coast by H. T. Larson of Hughes Aircraft Company, Culver City, Calif., and on the East coast by M. M. Astrahan of International Business Machines, Poughkeepsie, N. Y. Through the combined efforts of both promoters, the Group has a nationwide representation on its Administrative Committee.

Nathan Marchand, Chairman of the **IRE Professional Group on Information Theory,** has announced the appointment of D. L. Trautman of the University of California, as a member of the Administrative Committee of the National Professional Group. Professor Trautman is Chairman of the Los Angeles Group Chapter of the Information Theory Professional Group.

Eugene Mittelmann, Chairman of the **IRE Professional Group on Industrial Electronics** has reported that the recent membership drive conducted in the Sections of Akron, Chicago, Cleveland, Los Angeles, Milwaukee, New York, Pittsburgh, and Schenectady, has increased the membership of the Group to 848, as of September 15, 1951.

The petition for the formation of an **IRE Professional Group on Electron Devices** was approved by the Committee on Professional Groups at its last meeting. The Administrative Committee of this Group is being formulated; G. D. O'Neill, of Sylvania Electric Products Company, Bayside, L. I., is acting Chairman.

An IRE Symposium at Brookhaven National Laboratory, December 3 and 4, is being sponsored by the **IRE Professional Group on Nuclear Science** in co-operation with the Atomic Energy Commission. The afternoon and evening of December 3, will be composed of a program of papers on Nuclear Developments, with a social hour at 5:30, followed by dinner at 7:00, and resuming at 8:15 with papers on Atomic Energy, plus a panel discussion. The second session on the morning of December 4, consists of papers on Scintillation Counters, Functions of Research Reactors, Nuclear Reactors, and Fission Products in Industry.

A new **IRE Professional Group on Microwave Electronics** is being formulated. Those interested in becoming members of this Group are requested to contact Ben Warriner, IV, 151 Jefferson Avenue, Thornwood, N. Y. The scope of this Group is to encompass microwave theory, circuitry and technique, microwave measurements, tubes.

A national meeting on Land Mobile Communications was held at the Sheraton Hotel, Chicago, Ill., October 25 and 26, sponsored by the **IRE Professional Group on Vehicular Communications.** The welcoming address was given R. V. Dondanville, Chairman of the Chicago Group Chapter. Austin Bailey, Chairman of the National Group, and F. T. Rudelman, Vice-Chairman, acted as Moderators for the two-day session. Guest speakers at the dinner on October 25, and luncheon, October 26, were E. M. Webster of the Federal Communication Commission and J. D. O'Connell, Office of the Chief Signal Office, U. S. Army, respectively. Interesting papers on different phases of mobile communication were presented by: G. C. Terrell, L. P. Morris, M. E. Bond, J. F. Byrne, R. H. McRoberts, A. A. Macdonald, H. H. Davids, M. R. Friedberg, C. P. Williams, D. W. Bodle, J. S. Brown, H. K. Lawson.

The **IRE Professional Group on Audio** was sponsor to the Audio Session at the National Electronics Conference held last month in Chicago. The Audio Session of the Radio Fall Meeting in Toronto, Canada was also sponsored by the Audio Group, under the Chairmanship of Frank Slaymaker of the Stromberg-Carlson Company. Plans are being formulated to transmit the papers presented at these Audio Sessions to all members of the IRE/PGA who have paid their assessments. A meeting of the Administrative Committee of this Group was held October 22, 1951, in Chicago. The September, 1951, issue of the NEWSLETTER of the IRE/PGA contains a technical editorial by H. F. Olson, entitled "Selecting a Loudspeaker." An announcement has been made in the NEWSLETTER of the arrangements made by the Group with Shure Brothers, Incorporated, to send a Shure Reactance Slide Rule to all members of the Group who send in the coupon contained in the NEWSLETTER. A similar arrangement has been made with the Ohmite Manufacturing Company by which they will send an Ohmite Ohm's Law Calculator. Previous plans had been made with the Mark Simpson Manufacturing Company, to send a Masco Sound Surveyor Slide Rule to all members of the Group on the same no-charge basis.

IRE-AIEE Computer Conference Program Announced

The Joint IRE-AIEE Computer Conference to be held December 10, 11, and 12, in Philadelphia, Pa., will present papers from some 12 computing groups or manufacturing organizations which have succeeded in obtaining useful results from present large-scale digital computers. The following program lists the machines which will be discussed.

Monday, December 10, Benjamin Franklin Hotel:

1101, Engineering Research Associates; TESTRAC, Burroughs Adding Machine Corporation; UNIVAC, Remington Rand, Eckert Mauchly Division.

Tuesday, December 11, Edison Building:

Card-Programmed Electronic Calculator, IBM Corporation; Institute for Advanced Study Machine, Institute for Advanced Study, Princeton, N. J.; ORDVAC, University of Illinois; SWAC, Institute for Numerical Analysis, Los Angeles, Calif.; Harvard MARK III, Dahlgren Proving Grounds; University of Manchester Electronic Computer, Ferranti Limited, England.

Wednesday, December 12, Benjamin Franklin Hotel:

Whirlwind I, Massachusetts Institute of Technology; EDSAC, Cambridge University, Eng.; SEAC, Bureau of Standards, Washington, D. C.

A Keynote address by W. H. MacWilliams, Jr., of the Bell Telephone Laboratories will precede the formal presentation of papers. J. W. Forester will summarize the Conference on Wednesday afternoon. Inspection trips have been arranged for visiting the Bureau of Census UNIVAC at Remington-Rand, Eckert-Mauchly Division, Burroughs Research Division, Moore School of Electrical Engineering, University of Pennsylvania, and Technitrol Engineering Company. Reservations for trips will be made at the Conference due to the space limitations of these groups.

Notice

IRE MEMBERS

Rates of the Proc. of the IEE

There has been a slight increase due to publication costs in the reduced rates given to those members of the IRE who subscribe to the *Proceedings of the Institution of Electrical Engineers*. These rates, effective January 1, 1952, are as follows:

	Sterling
Part I (General)	12s. 6d.
Part II (Power Engineering)	17s. 6d.
Part III (Radio and Tele-communication Engineering)	17s. 6d.
Part IV (Collected Monographs)	7s. 6d.
All Four Parts Together	£2.10s. 0d.

STUDENT AWARDS ANNOUNCED

The IRE Board of Directors established a plan last year whereby students in the different Student Branches may be given an award by the local Section. The Annual IRE Student Branch Awards for 1951, as well as the name of the Student Branch in which the student winner was enrolled, are listed as follows:

Student Branch Award Winner	*Student Branch*
R. D. Wengenroth	Rensselaer Polytechnic Inst.
G. M. Badoyannis	Rutgers Univ.
C. C. Townsend	Princeton Univ.
J. E. Shea	Univ. of Connecticut
T. G. Lynch	Univ. of British Columbia
R. D. Gloor	Univ. of Louisville
E. J. Breiding	Univ. of Kentucky
S. C. Bogan	Yale Univ.
C. G. Blanyer	Univ. of Illinois
J. R. Wood	Utah State Agricultural College
D. J. Groszewski	Univ. of Dayton
R. H. Wilcox	Lafayette College
H. D. Ruhl, Jr.	Michigan State College
R. O. Rheaume	Univ. of Detroit
F. H. Tendick, Jr.	Univ. of Michigan
H. J. Hummel	Illinois Inst. of Technology
R. A. Huwe	Univ. of Minnesota
J. R. Hall	Seattle Univ.
O. W. Fix	Univ. of Washington
D. O. Martin	Southern Methodist Univ.
V. C. Hathaway	Northwestern Univ.
E. C. Pauly	Univ. of Florida
D. M. Culler	Carnegie Inst. of Technology
H. R. Stillwell	Univ. of Pittsburgh
W. J. Huhtala	Michigan College of Mining and Technology
R. C. Ritchart	Univ. of Wisconsin
A. F. Petrie	Marquette Univ.
W. F. Kyle	California State Polytechnic College
F. R. Goodman	California Inst. of Technology
J. E. Niebuhr	Univ. of Southern California
P. H. McBride	Univ. of Miami
K. O. Timothy	Univ. of Utah

UHF SYMPOSIUM HELD

The IRE Professional Group on Broadcast Transmission Systems sponsored an UHF Symposium, the first of its kind, which was held at the Franklin Institute, Philadelphia, Pa., Sept. 17. Over 150 persons from 20 states heard eight leading specialists present outstanding papers on this all-important phase of television. The talks, well illustrated and supported by exhibits, were presented in the following order.

"Some Experiments with 850-Mc Television Transmissions in the Bridgeport, Connecticut, Area," G. H. Brown, RCA Laboratories Division; "DuMont 700-Mc UHF Installation," William Sayer, Jr., and Elliot Mehrbach, Allen B. DuMont Laboratories, Inc.; "Impedance and Frequency Measurements at UHF," R. A. Soderman and F. D. Lewis, General Radio Company; "Side Fire Helix UHF Television Transmitting Antenna," L. O. Krause, General Electric Company; "A Fundamental Approach to UHF Television Receivers," W. B. Whalley, Sylvania Physics Laboratories; "Progress Report on the RCA-NBC UHF-Project at Bridgeport, Connecticut," Raymond Guy, National Broadcasting Company; "Transmission Line Problems in the UHF Television Band," J. M. DeBell, Jr., Allen B. DuMont Laboratories; "An Electronic Radio Field Strength Analyzer for Use in Television Station Field Surveys," F. W. Smith, National Broadcasting Company.

Lewis Winner of *TeleVision Engineering* and Chairman of the IRE Professional Group on Broadcast Transmission Systems, opened the symposium with introductory remarks, and an "UHF Information Please Roundtable," completed the day.

IEE TV CONVENTION SLATED

The committee of the radio section, acting on behalf of the Council of the Institution of Electrical Engineers, are arranging a convention to be known as "The British Contribution to Television," to be held in London, April 28 to May 3, 1952, and they cordially invite IRE members to attend.

The convention will be organized in nine sessions covering the complete field of television. Each session will be devoted to the presentation and discussion of technical papers, and will be supported by demonstrations where applicable. Visits of inspection to organizations concerned with every aspect of television will be included in the program.

The sessions into which the convention will be divided are as follows: Program origination; Point-to-point transmission; Broadcasting stations; Propagation; Receiving equipment (2 sessions); Nonbroadcasting applications; and System aspects. An historical paper and a broad survey paper to act as an introduction to the whole convention will also be presented.

It is expected that proofs of all papers will be available in advance. A special issue of *The Proceedings of the Institution of Electrical Engineers* will be published containing all proceedings of the convention.

Full particulars, and a form of application for registration of those members wishing to attend the convention, will be issued shortly to all the regular recipients of Part III (Radio and Communication Engineering) of *The Proceedings of the Institution of Electrical Engineers*. Members who do not subscribe to this publication and who wish to take advantage of this invitation should notify the Executive Secretary of the Institute of Radio Engineers, 1 E. 79 St., New York 21, N. Y., so arrangements can be made for the supply of registration forms.

UHF SYMPOSIUM SPEAKERS

Speakers at the UHF Symposium of the IRE Professional Group on Broadcast Transmission Systems, held September 17, at the Franklin Institute in Philadelphia, are (left to right): front row; F. W. Smith, NBC; L. O. Krause, General Electric; R. A. Soderman and F. D. Lewis, General Radio; W. B. Whalley, Sylvania Electric; G. H. Brown, RCA Laboratories Division, RCA; William Sayer, Jr. and Elliot Mehrbach, DuMont; J. M. DeBell, Jr., DuMont; and Raymond Guy, NBC. In the rear appear the two Moderators for the session (left to right): D. D. Israel and S. L. Bailey.

IRE People

Bruce Williams (S'47–A'50) has been appointed as sales engineer for the John A. Green Company, Dallas, Tex. In his position he will call on industrial accounts, jobbers, research laboratories, and manufacturers in the states of Texas, Oklahoma, Louisiana, Arkansas, and New Mexico.


Bruce Williams

Mr. Williams was born on October 20, 1919, in Maryland, and attended the University of Cincinnati and the Oklahoma A and M College, Stillwater. He received his B.S. and M.S. degrees in electrical engineering at Oklahoma A and M College, in 1947 and 1949, respectively. He was also an instructor in electrical engineering from 1947–1948. Mr. Williams was a research engineer at the Field Research Laboratory of the Magnolia Petroleum Company, Dallas, until he joined the staff of the John A. Green Company.

Mr. Williams recently was Exhibits Chairman for the 1951 Southwestern IRE Conference which was held in Dallas. He is a member of Delta Tau Delta fraternity, and a member of the AIEE and the ARRL.

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O. L. Angevine, Jr. (S'36–A'37–SM'44), formerly chief engineer of the sound equipment division, Stromberg-Carlson Company, recently accepted an appointment as chief engineer of the Caledonia Electronics and Transformer Corporation, it has been learned.

Born in Rochester, N. Y., in 1914, Mr. Angevine received the B.S. degree in electrical engineering in 1936 from the Massachusetts Institute of Technology. Upon graduation he joined the staff of the Stromberg-Carlson Company as an engineer in the telephone laboratory. He was appointed staff engineer for the vice-president in charge of engineering in 1941, and chief engineer of the sound equipment division in 1946.

A former Chairman of the IRE Professional Group on Audio and of the Rochester Section, Mr. Angevine is active on the Video and Audio Techniques Committee. He belongs to the American Institute of Electrical Engineers, to the Acoustic Society of America, and is currently serving as chairman of the sound equipment section, engineering department, of the Radio-Television Manufacturers Association.

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Marvin Hobbs (A'35–M'41–SM'43) has been named adviser to the Chairman of the Munitions Board. In this capacity he will co-ordinate all phases of the Defense Department's planning to meet the requirements for military electronics production. He will assist the Vice-Chairman for Production and Requirements, and the Military Director for Production.

Mr. Hobbs was born in Kyana, Ind., in 1912. He received the degree of B.S.E.E. at Tri-State College, Ind., in 1930. His engineering and production experience in the radio electronics and television industry extends over a period of 20 years. During World War II, he was associated with the War Production Board and with the Army Air Forces in the Pacific area. After the war he worked as a consulting engineer in Chicago for a number of radio and television manufacturers, including RCA and the Scott Laboratories. In 1950 he was appointed Deputy Executive Director of the Electronics Division of the Munitions Board in the Department of Defense, and has worked in this capacity until now.

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Quincy A. Brackett (M'41–SM'43), an assistant to Dr. Lee DeForest, inventor of the radio tube, died recently at the St. Andrew's Hospital, Boothbay, Me., after a long illness. His age was 66.

Mr. Brackett was graduated from Harvard University in 1907 and worked in the New York City laboratories of the Western Electric Company with Dr. DeForest for three years. After being associated with radio station KDKA, he was employed by the Westinghouse Electric Corporation during World War I, supervising production of radio equipment for the Army and Navy.

From 1921 to 1935, he worked with Westinghouse's East Springfield, Mass. plant, where radio transmitters, receivers, and other equipment were produced. In 1935, he helped found radio station WSPR in Springfield, and was president of that station from then until last spring, when he went into semi-retirement as vice president.

E. U. Condon (M'42–SM'43), Director of the National Bureau of Standards and noted nuclear physicist, has been appointed as director of research and development of the Corning Glass Works, Corning, N. Y. Dr. Condon has resigned as Director of the National Bureau of Standards, effective September 30, 1951.

For a photograph and biography of E. U. Condon, see page 707 of the June, 1951, issue of Proceedings of the I.R.E.

Norman L. Winter (A'47–M'47), chief sales engineer for Sperry Gyroscope Company, Great Neck, L. I., N. Y., has been appointed Chairman of the Navigation Technical Group of the Research and Development Board, Department of Defense, it was announced. This group was recently established to advise the Board on the integration and consolidation of air, land, and sea navigation research and development projects.


N. L. Winter

Mr. Winter has been associated with the Board since its establishment. He was Executive Director of the Committee on Electronics from 1946 until 1949, when he joined the staff of Sperry. Since that time he served the Board in a part-time capacity as a consultant on electronic, aircraft, and navigation problems.

In 1942 Mr. Winter was called to active duty with the Army in the Office of the Chief Signal Officer, in command of the Electronics Branch of the Engineering and Technical Service. For his services in this and other capacities during World War II, he was awarded the Legion of Merit.

From 1929 to 1941 he was employed by the General Electric Company, successively, as meter design engineer, motor design engineer, and motor application engineer.

Mr. Winter was graduated from Purdue University with the B.S. degree in electrical engineering, and has done work at the Massachusetts Institute of Technology, the University of Indiana, and Harvard.

He is a member of the American Institute of Electrical Engineers, the Indiana Engineering Society, the Air Force Association, and the Army Signal Association.

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Leo G. Sands (A'44–M'45–SM'50), formerly staff assistant to the general sales manager at Bendix Radio, was named director of public relations and advertising at that organization, it has been learned. A specialist in the design of control amplifier circuits, Mr. Sands has been actively concerned with the entire field of radio and electronic circuits. His career has included the following positions: Chief Inspector of Airborne Radar Eequipment for the U. S. Army at the Sacramento Air Depot; service manager for various radio stores on the west coast, installations engineer of the Remler Company; and design engineer of the Coast Radio Company, San Jose, Calif.

Cyrus D. Backus (A'19–M'26–SM'43), former examiner in the radio division of the Patent Office, Washington, D. C., and an authority in the communications field, died recently at his home in Silver Spring, Md.

A native of Groton, N. Y., Mr. Backus was graduated in law and philosophy from Cornell University in 1896. In 1903 he received the degree of M.E.E. from George Washington University and later was named head of its electrical communication department. Mr. Backus was chief of Division 51, the principal radio division of the Patent Office, from the early years when that division was formed. He was associated with the Patent Office for 40 years, retiring in 1943, and for four years was patent law consultant of the International Telephone and Telegraph Company in New York, N. Y. He then returned to private law practice. A member of the District Bar in Washington, D. C., he was admitted to practice before the Court of Customs and Patent Appeals and the Supreme Court.

Mr. Backus was a Fellow of the American Association for the Advancement of Science and belonged to the American Institute of Electrical Engineers.

J. W. Hines (S'46–A'48) has been appointed as the new sales engineer of Magnecord Incorporated, Chicago, Ill. Mr. Hines will do liaison work between the engineering and sales department of Magnecord Incorporated, and will also handle technical service and sales problems.

J. W. Hines

Mr. Hines was born in Wilkinsburg, Pa., on September 26, 1923, and received his degree of B.S.E.E. at the Carnegie Institute of Technology in 1947. He has been active in the electronics field for the past 8 years, and served as a newsreel cameraman for the Signal Corps in Europe, during World War II.

Previous to his staff appointment with Magnecord, Mr. Hines was chief engineer for radio station WBVP, Beaver Falls, Pa.

Mr. Hines is a member of the Chicago section of the IRE.

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William J. Warren (SM'46) has joined the staff of Shell Development Company in Emeryville, California, and has been assigned to the Associate Directors' staff.

Mr. Warren was born in 1910 at Eureka, Calif. He received the B.S. degree in electrical engineering in 1931, from Santa Clara University and the Ph.D. degree in 1936, from the University of Illinois, where he taught from 1934 to 1937 and from 1938 to 1941. He was employed from 1937 to 1938 by the General Electric Company at Schenectady, N.Y., as a test engineer.

Before joining the Shell Development Company staff, Mr. Warren was affiliated with the University of Santa Clara where he was Professor of electrical engineering and part-time consultant for several industrial firms, since 1941.

Mr. Warren is a member of Sigma Xi, Tau Beta Pi, Eta Kappa Nu, Phi Kappa Phi, and a member of the AIEE.

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Loyd C. Sigmon (A'29–SM'46) has been awarded an honorary degree of electrical engineer by the Milwaukee School of Engineering during their June, 1951 commencement exercises, at which he was their featured speaker. Mr. Sigmon, who is chief engineer, vice president, and assistant general manager of radio station KMPC, Hollywood, Calif. advised the graduates that they have a knowledge which is needed now as in no other time in American history.

In prior years Mr. Sigmon attended the school of electrical engineering, Milwaukee, Wis., and from 1935 to 1940, was chief engineer for radio station KCMO, Kansas City, Mo. He then held the position of director of engineering for KMPC, Los Angeles, until his entry into the Armed Services, during World War II, where he served with the Army Signal Corps, with the rank of Lieutenant Colonel, as radio officer in the Signal Corps Communications Division of the European Theatre of Operations.

Mr. Sigmon was born on May 5, 1909, in Stigler, Okla. He is the holder of the Legion of Merit and the Order of the British Empire. He is an honorary member of the French Signal Corps.

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H. I. Romnes (SM'46), formerly general manager of the long lines department of American Telephone and Telegraph Company, has recently been appointed director of operations.

Mr. Romnes was born in Wisconsin, in 1907, and received the B.S. degree in electrical engineering from the University of Wisconsin in 1928. His professional experience includes service with the Wisconsin Telephone Company, and membership on the technical staff of the Bell Telephone Laboratories, Inc. In January, 1945, he joined the American Telephone and Telegraph Company, in charge of the toll transmission troup in the operation and engineering department, where he was responsible for long-distance transmission facilities of all types.

In 1950 Mr. Romnes became associated with the Illinois Bell Telephone Company, Chicago, Ill., as chief engineer, but returned to AT&T in December of that year.

Ernest R. Cram, a charter member of the IRE and pioneer in wireless telegraphy, died recently in the Long Island College Hospital, Brooklyn, N. Y. He was 70 years of age.

A native of Boston, Mass., Mr. Cram received his education at Harvard University and was first associated with the Stone Wireless Telegraph Company of Boston, for which he took out several patents on tuning circuits. Later, he joined the U. S. Signal Corps, Washington, D. C., as civilian engineer, and after 25 years of service in this line became associated with the Radio Corporation of America, with respect to patent and legal matters.

In the Signal Corps, his chief duties were the development of mica transmitting condensers, which led to their later commercial manufacture by the Dubilier Condenser Company, and as assistant to (then) Major George Owen Squier, U. S. Army Signal Corps, in experimenting with wired wireless, forerunner of the coaxial cable. He also assisted, for a time, John Hays Hammond, Jr., in the development of radio-controlled torpedoes.

Mr. Cram was founder of the Society of Wireless Telegraph Engineers, which later became part of the IRE.

J. W. Nelson, Jr. (A'46–SM'47) has been appointed as the manager of the newly formed application engineering section within the General Electric Government Sales Department, Syracuse, N. Y. His primary association with the new section will be to work closely in the field with all branches of the armed services, assisting them in the use of present electronic equipment, and locating possible application for which new electronic devices might be developed.

J. W. Nelson, Jr.

A native of Berkeley, Calif., Mr. Nelson received the degree of B.S. in electrical engineering from the University of California in 1941. During 1941–1942, he served as a research associate in the radiation laboratory at Massachusetts Institute of Technology, Cambridge, Mass., engaged in microwave research. Mr. Nelson served the next four years with the Air Force in World War II, as a development engineering officer, and later joined General Electric as a development engineer in their Government Division at Syracuse. He became sales engineer in 1947, and in 1949 he was named sales manager for the Air Force equipment section of General Electric. This was the position Mr. Nelson held prior to this announcement.

He is a member of Sigma Xi and RESA Tau Beta Pi, and Eta Kappa Nu.

Books

Propagation of Short Waves Edited by Donald E. Kerr

Published (1951) by McGraw-Hill Book Co., 330 W. 42 St., New York 18, N. Y. 692 pages+22-page index+14-page appendix+xvii pages. 299 figures. 6×9½. $10.00.

The next-to-last volume to be issued as part of the Radiation Laboratory Series will be a most welcome addition to the libraries of those working in the field of tropospheric radio-wave propagation. One cannot help but be immediately impressed by the fact that here, at hand, is a single volume containing an excellent summary of hundreds of wartime reports on vhf and uhf transmission. In spite of the many problems involved in the selection and review of these reports, the finished text is well constructed and concise, without too much loss of pertinent detail.

Naturally, it was impossible to incorporate all of the various phases of uhf propagation phenomena into this one volume. Very little material appears relating to diffraction by obstacles or forecasting techniques; however, in both instances, the subjects were either not studied thoroughly by Group 42, or overlapping material has already appeared in print. This is not mentioned as detracting from the usefulness of the volume, but only to illustrate its limitations.

The coverage of the book may best be summarized in a simple listing of the chapters. These are: "Elements of the Problem," "Theory of Propagation in a Horizontally Stratified Atmosphere," "Meteorology of the Refraction Problem," "Experimental Studies of Refraction," "Reflections from the Earth's Surface," "Radar Targets and Echoes," "Meteorological Echoes," and "Atmospheric Attenuation." The volume also contains an Appendix dealing with two additional views on the mathematical theory of scattering. These are the Lorentz Reciprocity Theory and Coherent vs. Incoherent Scattering.

This reviewer was particularly interested in the sections of the book which developed the radio-meteorology relationship. This is a subject that has sorely needed a comprehensive background, understandable to the radio engineer. Chapter 3 accomplishes this purpose and should find considerable use as a standard reference in this field. The qualitative aspects of the numerous experiments performed in order to correlate radio-meteorological pattern are extremely interesting. The treatment of this material here is much more descriptive and illustrative than it is in any comparable volume published to date. Accent is primarily placed upon modified-index distribution (M-profile) which renders a good agreement over medium-length paths. The higher fields over longer pathlengths may probably be accounted for by theories developed after the dissolution of Group 42.

A fair section of this book is devoted to the problems of scattering and target echoes. The theories and mathematical formulas relative to these problems are very well developed, and the authors are the first to admit that, at present, the cases of most practical interest are unfortunately beyond the scope of existing methods. Good discussions are noted in this chapter on clutter echoes, isolated targets, and the sea echo.

The authors have successfully avoided the duplications evident in British and American reports on the subject. The gaps in the coverage of the material which have been mentioned above do not detract seriously from the very apparent usefulness of the book. The only objection which can be raised is the inaccessibility of the many reports used as source material for which the authors can hardly be blamed.

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Review of Current Research and Directory of Member Institutions, 1951

Published (1951) by the Engineering College Research Council of the American Society for Engineering Education, Room 7-204, 77 Mass. Ave., Cambridge 39, Mass. 215 pages+28-page index+x pages. 6×9. $2.25.

For the benefit of readers unfamiliar with the previous (1949) edition of this volume, it is a valuable listing of educational institutions active in pure and applied research, with an abstract of their current activities. The present edition has 35 per cent more pages than its predecessor.

About 80 colleges comprising the active members of the Engineering College Research Council are listed. Several pages are devoted to each one with information on its research officers, policies, personnel, expenditures, and projects now active.

Typical of some twelve particularly active research schools is an annual expenditure of over a million dollars to support a research staff of more than a hundred professors, consultants, other scientists and advanced students engaged on these projects.

The projects themselves cover the entire range of scientific subjects with emphasis on applied science and engineering. Such emphasis is to be expected as most of the work is sponsored by government agencies for military objectives. It is presumptive, however, that one of the principal aims is the subsidizing of education for increasing the technological potential of the nation; one might even go so far as to say that the many useful results may be regarded as by-products of a system geared to the development of trained and inspired scientists and engineers.

Well indexed by subjects, the volume is a convenient guide to those colleges that may be specializing in any particular field. It is a concise record of the current activities in one of the most ambitious educational experiments in American history.

Harold A. Wheeler
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Servomechanisms and Regulating System Design, Vol. I by Harold Chestnut and Robert W. Mayer

Published (1951) by John Wiley & Sons, Inc., 440 Fourth Ave., New York 16, N.Y. 497 pages+7-page index +6-page bibliography + xiii pages. 304 figures. 5¾×9½. $7.75.

This is a book of good quality which does not particularly expand the field of its title but which does add a well-rounded work to those already on the market. It is intended for the "training of design and application engineers" and the particular volume under review is "adapted to the needs of engineers and engineering students who have not had previous training or experience in the field of closed-loop control systems."

To put down some thoughts which, although not unimportant, can be disposed of by a brief listing: (1) The present volume is the first of two; since the second has not been published, it is necessary for Volume I to stand on its own feet here. (2) The book has a profusion of excellent illustrations and charts. (3) Worked-out examples illustrating text material are good. (4) Typography and general appearance are good and typographical errors are infrequent. (5) There is a good bibliography of more than 100 items, almost all in English. (6) Problems for students to solve, covering some 50 pages at the end of the text, are good. (7) The book, in the General Electric series, is meant not only for electrical engineers but apparently also for other engineers who take General Electric Company courses. As a result there is a considerable amount of elementary material in the first third of the book.

Omitting consideration of elementary matter, Chestnut and Mayer's book covers material which might be described as customary, with the exception that considerably more emphasis is placed on attenuation and phase characteristics than usual. Chapter subjects (not titles) include stability, transfer functions, system types, complex-plane plots, attenuation and phase characteristics and applications, integral-network relations, multiple-loop systems, and transient performance.

On summary glance the text would seem to be excellent. It does not, on full reading, live up to original expectations. There is no one major reason for this; instead the result arises from many small causes; the book is over-written and could be appreciably reduced while supplying the same technical information; a more than occasional lack of balance exists; empirical information is introduced without even qualitative justification or explanation; sometimes major jumps depend on slight or nonexistent bases; a large number of curves in the text do not have the co-ordinates labeled; the design objective of the text tends to be slightly emphasized, with the result that discussions of which charts will produce an answer the soonest seem to be over-emphasized; and so on, and so on. Altogether it comes down to the fact that so many disturbing simple items can be found so frequently throughout the text, that the book falls short of being excellent.

A notation developed by a subcommittee of the AIEE Committee on Feedback Control Systems is used. This notation, which does not have the blessing of the AIEE Standards Committee nor the IRE's, nor the ASA's, is not only used here but, in addition, will appear in another servomechanisms text to be published shortly. Thus there will be a popularization of a notation which has been developed on the basic premise that the field of feedback control is one to itself and that correlation with other fields of electrical engineering is not necessary. An example of the use of the notation in the book under review will show the difficulty which a student would have in passing from one class to another, or a listener in passing from one session to another, if the Chestnut and Mayer notation were used in one. On one figure (page 282) appear $E=I$ and $O=C$, in which no symbol refers to voltage, current, zero, or capacitance. R (for reference input) and R (resistance) appear on one diagram, and E (error) and E (voltage) likewise. Many common symbols (B, C, E, H, M, Q, R, V, Z) have meanings completely different from the usual and standardized ones. This constitutes a considerable problem for the IRE Symbols Committee and the newly created IRE Technical Committee on Servomechanisms.

From the point of view of the student who wants to learn the theoretical and fundamental background of the servomechanism field and who can pick up his design and empirical information later, it may not be fully desirable that new books cover to a large extent the field already covered by predecessors, even though a reasonable choice is valuable. In January, 1946, Professor Guillemin raised the question of why servomechanisms were not designed from the desired over-all transfer characteristic. Recent work in the theses of Clanton, Truxal, and Aaron, as well as some unpublished work, show the potentialities of this approach. It is true that quite substantial problems remain to be overcome, but none the less there is a good possibility that the next major advance in a textbook in this field will be achieved by the one which starts from an attempt to synthesize an over-all characteristic and considers the details in the present book as cases evolving from that general problem.

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An Introduction to Electron Optics by L. Jacob

Published (1951) by John Wiley & Sons Inc., 440 Fourth Ave., New York 16, N. Y. 137 pages+4-page index+1-page references+6-page bibliography. 48 figures. 4×6½. $2.00.

The author is affiliated with the Universities of Manchester and Liverpool, England.

This little book covers the entire field of electron optics from fundamental principles to lens aberrations, phase focusing, beam properties, and beam deflection. Field mapping and ray tracing in electrostatic fields, as well as the properties and aberrations of electrostatic lenses are included in about half of the book; however, magnetic lens fields are considered in somewhat less detail than the former.

The title, "Introduction to Electron Optics," is a little misleading. It constitutes rather, a summary of a very small considerable fraction of existing electron-optical literature. The beginner would have difficulty in discriminating between the more important and the less important techniques and findings of electron optics as they are reported here. Some of the general laws of electron optics are not brought out clearly and, too frequently, the logical basis of mathematical developments is not indicated, and symbols are inadequately defined.

Nevertheless, the worker in the field of electron optics will appreciate the many references to papers which may, heretofore, have escaped his notice. As a "refresher," particularly with reference to recent British work in the field of electron optics, the book should all the more so be useful, since the remarkable amount of material which has been compressed into this small volume appears to be relatively free of errors.

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Television and FM Antenna Guide by Edward M. Noll and Matthew Mandl

Published (1951) by the Macmillan Co., 60 Fifth Ave., New York 11, N. Y. 308 pages+3-page index +ix pages. 225 figures. $5.50.

With the advent of television, rooftop antennas have reappeared. They have sprouted in a way that is reminiscent of the appearance of outside antennas 30 years ago in the early days of broadcasting. The TV antenna, however, is a different species because television waves are short. This is both an advantage, because directional types are feasible, and a handicap, because the effective apertures are small. For example, a 1-megacycle broadcast antenna 50 feet long is but a small fraction of a wavelength, yet it may have an effective aperture of 50,000 square feet over which it can extract energy from a passing radio wave. On the other hand, a half-wavelength TV antenna operating at 100 megacycles has a maximum effective aperture of less than 15 square feet over which it can collect energy from a television wave. This handicap of small aperture may be compensated in part by the use of directional-antenna arrays, which is why we find a great variety of stacked, colinear, Yagi, and corner-reflector types where signals are weak. Some of these have been adapted from earlier amateur vhf practice, but many have been designed especially for the wide band requirements of television. Considering the many types in use, it is rather surprising that long-wire and rhombic types have not found wider application in rural areas where the necessary acreage is available.

Television-receiving antennas form an extensive subject. It is natural, therefore, that an entire book has been devoted to this topic. This book, "Television and FM Antenna Guide," by Edward M. Noll and Matthew Mandl, should be of assistance to all interested in TV and FM receiving antennas, and in particular to technicians concerned with TV-antenna installation. The treatment is elementary, easily followed, and intensely practical. It is essentially nonmathematical.

The first 26 pages give an introductory picture of wave propagation at very high frequency. Included are FCC charts for the calculation of signal range. The next 30 pages cover some pertinent properties of transmission lines. This is followed by 56 pages on important practical considerations regarding antennas with a discussion of feeds and patterns of a number of simple antennas and directive systems.

The remainder of the book (187 pages) is in the form of a guide arranged for easy reference, which takes up first such topics as antenna-site selection, installation tools and procedures, transmission-line installation methods, input systems, booster amplifiers, and so forth. This is followed by a series of short, well-illustrated descriptions of numerous commercial TV antenna types ranging from simple dipoles to multielement stacked arrays. Many patterns that apparently are power plots are presented. There are also sections on long wire types, indoor type, antenna rotators, and multiple-outlet systems for apartments.

This book should be of great interest and value to all concerned with TV receiving antennas.

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New Publications

TV Factbook No. 13, published July, 1951, by Television Digest, 1519 Connecticut Ave., Washington, D. C. Martin Codel, editor and publisher. This new, 96-page, $5 edition is the 13th in a semiannual series and includes such reference material as:

Personnel and facilities data, with digests of rate cards for all of the 107 television stations and four networks now operating, together with listings of actual and proposed TV stations in Latin America and Canada; complete tabulation of the now frozen 416 applications for new stations pending before FCC; official tables of present vhf and proposed new vhf- and uhf-channel allocations by states, cities, and channels indicating educational assignments, proposed station shifts, and so forth.

New features of this edition are a population-dwelling-sales analysis of the 162 most important markets of the United States; directories of engineers, attorneys, and related services specializing in TV; a 34×22″ wall map in color, showing present television areas and actual and projected coaxial-microwave network routes.

Brought up to date are directories of the 92 TV-receiver manufacturers in the United States, 13 in Canada, 38 picture-tube and 12 receiving-tube makers, 27 concerns manufacturing TV transmitting and associated equipment, 465 firms providing films and other programs to TV stations, lists of station sales representatives, labor unions, trade groups, research firms, and so on. In addition there are tabulations of television-radio receiver production by months since 1946 and the latest count of TV sets in use by areas.

Books

High-Frequency Measurements, Revised Second Edition by August Hund

Published (1951) by McGraw-Hill Book Co., 330 W. 42 St., New York 18, N.Y. 631 pages + 35-page index + 7-page appendix + xi pages. 417 figures. 6×9¼. $10.00.

This second edition of a pioneer book in high-frequency measurements has been revised extensively in order to include the developments of the last 19 years. "High frequency" now is interpreted to include all frequencies above 20 kilocycles per second up to super-high frequencies (microwaves). To increase the usefulness of the book, standard measurements and audio frequencies are also given. The chapter on Line and Antenna Determinations has been completely rewritten, the chapter on Modulation Measurement has been greatly expanded, the use of radio-frequency and very-high-frequency bridges has been added, and there is a discussion of signal-to-noise measurements. Because of space limitation, specific microwave techniques have not been included.

The book now covers in its authoritative and comprehensive manner introductory chapters on Fundamental Relations and Circuit Properties, High-Frequency Sources and useful laboratory apparatus, such as oscillographs, bridges, and attenuation networks. It then takes up the measurement of small high-frequency currents, and in individual chapters, the Measurement of Voltage, Frequency, Capacitance, Self-Inductance, Mutual Inductance, Effective Resistance, High-Frequency Power and Losses, Logarithmic Decrement, and related quantities. In each instance, basic relations are given, typical values mentioned, and possible errors and necessary precautions discussed. A brief chapter on Ferromagnetic Measurements follows; Tube Measurements are discussed more fully, with attention to transit-time effects. The chapter on Modulation Measurements is rather comprehensive with considerable stress on frequency and phase modulation. The chapter on Measurements on Lines and Aerial Systems briefly reviews transmission-line relations and discusses measurements on parallel-wire and coaxial lines; it does not treat measurements on waveguide systems.

Determinations on Wave Propagation include field-strength measurements and polarization effects in skywave and ionosphere propagation. A final chapter on miscellaneous matters includes measurements on Quartz Crystals, Standard Field Calibration, Signal-to-Noise Measurements, and a brief treatment of latest developments in bridge measurements.

Throughout the book there are frequent references to the companion volume, "Short-Wave Radiation Phenomena," published by the McGraw-Hill Book Co., in 1951, and to a previous book, "Phenomena in High-Frequency Systems," also published by McGraw-Hill, in 1936. Many of the illustrations have been rather drastically reduced in size from the original, probably in order to save space; in several places this will make the reading difficult.

A great improvement over the first edition is the consistent use of the now internationally adopted mks system. The author has continued the use of mathematical nomenclature and the procedures recommended by the Standards Committee of the IRE. With this timely revision, the book will retain its leadership in the frequency range of greatest interest to practical radio engineers.

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Transient Analysis in Electrical Engineering by Sylvan Fich

Published (1951) by Prentice-Hall, Inc., 70 Fifth Ave., New York, N. Y. 291 pages+6-page index+8-page appendix+ix pages. 92 figures. 5½×8½. $7.35.

This book is a new addition to the Prentice-Hall Electrical Engineering Series, edited by Professor W. L. Everitt. It was written with the aim of extending transient analysis in electrical engineering to include modern operational methods without eliminating the presentation of classical theory. The only prerequisites for understanding the subject matter are the conventional courses in calculus, college physics, and steady-state circuit theory. No previous knowledge of differential equations is assumed.

The first five chapters are devoted to the solution of linear differential equations and their application to the classical solution of electrical and mechanical transients. The concept of a complex frequency is introduced by the definition of a complex angle in the classical analysis of an oscillating circuit. The Laplace transform is developed in Chapter 6 and applied to networks in the following chapter. Methods for solving higher-degree algebraic equations are presented in Chapter 8 and applied to circuit problems in Chapter 9. The Fourier series, integral, and transforms are developed in Chapter 10. This is followed by an introduction to complex-variable theory and its application to the calculation of inverse Laplace transforms. Chapter 13 deals with distributed parameters. A resume of electrical analogues of engineering systems is given in the last chapter.

This book should prove to be an excellent text for an undergraduate course. As suggested by the author, the first nine chapters might be used for this purpose. The reviewer is less enthusiastic about the suggested use of the last nine chapters as the basis for a first graduate course. Although admirably presented, the ground covered in this section is so great that the treatment is necessarily abbreviated. To illustrate, it may be questioned whether the two-page derivation of Heaviside's expansion theorem or the 13-page chapter on systems having distributed parameters will meet the needs of the graduate students. And, although it is a matter of personal opinion, the reviewer feels that offering the concepts of the complex variable at a point so late in the book has weakened the presentation. But on the whole the book is very well written and the author is to be commended particularly for his valuable treatment of the physical basis of the mathematical results, as well as for the style, clear and readable throughout, and for his ability to lead the reader to an understanding of the subject.

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Radio Communication at Ultra High Frequency by J. Thomson

Published (1950) by John Wiley & Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 200 pages+3-page index+xix pages. 85 figures. 5×8¾. $4.50.

According to the preface, the aim of the author in writing this book was "to provide an account of modern developments in telecommunications employing radio waves of lengths ranging from a few meters to a few millimeters." It is addressed to students, and practicing communications engineers, as well as to workers in other electronic fields who might profit from it. The title of the book, would lead the prospective reader to expect a fairly complete discussion of the system considerations involved in uhf communications.

The author has not completely succeeded in accomplishing his own aim or in fulfilling the expectation mentioned above. The discussion of system considerations is restricted to a small portion of the final chapter of the book, while the remainder of the book is devoted to a discussion of uhf apparatus.

In several places the author relegated various developments to the indefinite future which actually were in commercial use at the time the book was written. One example of such questionable statements is in the final paragraph where he states: "Up to the moment of writing, the life of a U.H.F. valve is not satisfactory for unattended operation." In view of the General Electric Company's New York-Schenectady microwave relay system and the Bell System's New York-Boston and New York-Chicago systems, all of which were in commercial operation several years before the copyright date, this statement is difficult to understand.

In addition, there are a few minor errors such as the assigning of units to the dimensionless relative permeability on page 1, the use of $\overline{e_1}$ for $\sqrt{\overline{e_1^2}}$ throughout Chapter 4, and a few obvious typographical inaccuracies which could have been eliminated by careful proofreading.

This reviewer found the discussion of velocity-modulation devices and the chapter on receiver-input circuits interesting and well presented.

In spite of the shortcomings mentioned above, the author has produced a very readable book which should provide a valuable introduction to the field for students and which should prove to be, in part, a useful reference for the practicing engineer.

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The Annual Index to these Abstracts and References, covering those published in the PROC. I.R.E. from February, 1950, through January, 1951, may be obtained for 2s.8d. postage included from the *Wireless Engineer*, Dorset House, Stamford St., London S.E., England. This index includes a list of the journals abstracted together with the addresses of their publishers.

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ACOUSTICS AND AUDIO FREQUENCIES

016:534 2309

References to Contemporary Papers on Acoustics—A. Taber Jones and R. T. Beyer. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 377–385; May, 1951.) Continuation of 1812 of September.

534+621.395.61/.62](083.71) 2310

Standards on Electroacoustics: Definitions of Terms, 1951—(PROC. I.R.E., vol. 39, pp. 509–532; May, 1951.) Reprints of this Standard, 51 IRE 6 S1, may be purchased while available from the Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., at $1.00 per copy.

534.213–14 2311

Random Noise in an Attenuating Fluid Medium—R. E. Roberson. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 353–358; May, 1951.) "It is assumed that acoustic background noise is caused by a distribution of random "white" noise sources whose physical mechanism is unspecified. A law expressing the amplitude-distance attenuation characteristic of the medium is also assumed. Several distributions of noise sources are considered: uniform volume distributions, uniform surface dipole distributions, and two mixed cases. The noise dropoff, with frequency at a point below the surface, is found for each case. For an infinite volume of noise sources, this dropoff is 6 db/octave at all frequencies. It is shown how this simple model can be generalized to other attenuation laws and other spatial and amplitude distributions of noise sources."

534.232 2312

The Emission of Sound by a Piston—D. N. Chetaev. (*Compt. Rend. Acad. Sci.* (URSS), vol. 76, pp. 813–816; February 21, 1951. In Russian.) Integral (1) determining the amplitude of the velocity potential of steady-state sound oscillations is considered, and the sound field at an arbitrary point P, determined. A formula (5) determining the radiation resistance is also derived.

534.232:538.652:621.314.212 2313

Electroacoustic Transformation by Means of Magnetostriction, with Special Reference to Radiation from Transformers—H. H. Rust. (*Z. angew. Phys.*, vol. 2, pp. 487–491; December, 1950.) Magnetostriction curves, obtained with Ni, Fe, and Fe/Si oscillators, and taking account of magnetostrictive hysteresis, indicate complex oscillations rich in harmonics. These results apply directly to transformers, which in normal operation must incidentally radiate such magnetostrictively-excited mechanical waves with adverse effects on the properties of any oil used for insulation. Suggestions are made for eliminating these harmful effects.

534.24+534.373 2314

Scattering and Absorption by an Acoustic Strip—A. Levitas and M. Lax. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 316–322; May, 1951.) An analytical method is described for determining the scattering and absorption of sound for a nonuniform-boundary case represented by the application to an infinite wall of a finite-width strip of soundproofing material.

534.24 2315

Focusing of Sound Waves by a Parabolic Reflector—L. D. Rozenberg. (*Zh. Tekh. Fiz.*, vol. 20, pp. 385–396; April, 1950.) A mathematical investigation is presented of the sound field near the focus. Formulas are derived for determining the pressure at the focus, and the effects of increasing the aperture of the reflector on its focusing properties, while keeping the focal length constant, are examined. Methods are indicated for choosing the optimum focal length for a given aperture of the reflector, and for determining the radius of the diffraction circle at the focus. Cases when the source of sound is at a finite distance from the reflector, and not on the axis, are also considered. Some experimental results are included.

534.24 2316

On the Nonspecular Reflection of Sound from Planes with Absorbent Bosses—V. Twersky. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 336–338; May, 1951.) The analysis developed in 9 of February for nonabsorbent surfaces is extended to the case of absorbent surfaces. The results for the two cases are compared. The effect of the finite impedance may be either to decrease or increase the radiation reflected at the specular angle.

534.24 2317

Sound Scattering of a Plane Wave from a Nonabsorbing Sphere—R. W. Hart. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 323–329; May, 1951.) An analytical treatment is developed. Consideration is restricted to the case where the acoustic properties of the sphere are not very different from those of the surrounding medium. See also 2139 of October (Hart and Montroll).

534.24 2318

On the Reflection of a Spherical Sound Wave from an Infinite Plane—U. Ingard. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 329–335; May, 1951.) Boundary conditions are given in terms of a normal impedance independent of the angle of incidence. The integral for the reflected wave is expressed in a form such that the wave can be considered as originating from an image source having a certain amplitude and phase. Graphs for determining these values are given in terms of a "numerical distance" which depends on the normal impedance and the position of the field point.

534.321.9:537.228.1 2319

Design of Variable Resonant Frequency Crystal Transducers—W. L. Hall and W. J. Fry. (*Rev. Sci. Instr.*, vol. 22, pp. 155–161; March, 1951.) A description of a system employing liquid mercury as a backing of continuously variable dimensions. The important aspects, viz., tight coupling of the crystal and mercury backing, and decoupling of the crystal and mercury from the supporting structure, are considered in detail. Construction procedure on a unit to cover the frequency range 40 to 80 kc is indicated. Experimental results on the magnitude of the electrical input impedance as a function of frequency and mercury-column length are given. The unit is compared with transducers having fixed resonance frequency.

534.414 2320

The Wavelength of a Spherical Resonator with a Circular Aperture—H. Levine. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 307–311; May, 1951.) An expression for the wavelength is derived in the form of an expansion exact as far as terms in $(a/R)^2$, where a is the aperture radius and R the sphere radius. A procedure for determining the wavelength approximately for a resonator of arbitrary shape is also described.

534.414 2321

Variation of the Resistance in the Resonator Neck with Intensity of Incident Sound—R. K. Vepa. (*Science and Culture* (Calcutta), vol. 16, pp. 482–483; April, 1951.) Measurements were made on a resonator formed by a plate 0.65 cm thick with a 1.25-cm circular orifice, appropriately spaced from a rigid backing. Results are compared with earlier measurements using a thinner plate and smaller orifice.

534.771 **2322**

Upper Limit of Frequency for Human Hearing—J. H. Combridge, J. O. Ackroyd and R. J. Pumphrey. (*Nature* (London), vol. 167, pp. 438–439; March 17, 1951.) Comment on 2959 of 1950 (Pumphrey) and author's reply.

534.78 **2323**

Effect of Delay Distortion upon the Intelligibility and Quality of Speech—J. L. Flanagan. (*Jour. Acoust. Soc. Amer.*, vol. 23, pp. 303–307: May, 1951.) Speech articulation tests were made on an all-pass system capable of advancing or delaying one frequency band relative to the rest of the spectrum. Measurements were made at signal/noise ratios of 30 db and 0 db. The results indicate that maximum impairment of intelligibility occurs when the delays or advances are of the order of ¼ second, and when the band delayed or advanced is near the center of the speech spectrum.

534.79 **2324**

Calculation and Measurement of the Loudness of Sounds—L. L. Beranek, J. L. Marshall, A. L. Cudworth, and A. P. G. Peterson. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 261–269; May, 1951.) An equivalent-tone method is described, in which the spectrum of the sound is divided into frequency bands which are treated as pure tones in calculating their loudness. Calculations for bands of white noise and for complex tones are compared with subjectively obtained data. The agreement is good.

534.833.4 **2325**

Absorption of Sound by Resonant Panels—G. G. Sacerdote and A. Gigli. (*Jour. Acous. Soc. Amer.*, vol. 23, pp. 349–352; May, 1951.) The experimental determination of resonance frequency and absorption of resonators, formed by plywood plates with uniformly spaced circular holes placed at various distances from the wall, is described. Measurements were made at normal incidence and with diffuse sound in a reverberant room. Moderate agreement between theoretical and experimental results is noted.

621.395.61/.62 **2326**

The Piezoelectric Sound Detector and its Electrical and Acoustic Equivalent Circuits—F. A. Fischer. (*Arch. elekt. Übertragung*, vol. 4, pp. 435–436; October, 1950.) An equivalent electrical circuit is derived which is applicable for operation of the piezoelectric transducer as generator, or as detector of sound. The paper is complementary to that noted in 2966 of 1950.

621.395.61/.62 **2327**

Post-War Developments in Electroacoustics [by Telefunken]—F. Bergtold. (*Telefunken Ztg.*, vol. 23, pp. 106–110; September, 1950.) Apparatus described includes moving-coil microphone, pickup, sound-distribution system, loudspeaker and arrays, cinema installation, and house-communication system.

621.395.623.7 **2328**

Loudspeaker Damping—A. Preisman. (*Audio Eng.*, vol. 35, pp. 22–23, 38 and 24, 45; March and April, 1951.) Loudspeaker characteristics are discussed theoretically, and an experimental method is described for determining the constants of the unit. The Q is determined from the shape of the impedance/frequency curve, and the source resistance for critical damping is calculated from the voice coil and motional impedance at resonance. An alternative method, based on consideration from a mechanical viewpoint, is also presented.

621.395.625.2 **2329**

Gramophone Turntable Speeds—G. F. Dutton. (*Wireless World*, vol. 57, pp. 227–231; June, 1951.) Suitable speeds for microgroove recordings are considered in relation to public demand, record materials, groove spacing, needle size, and distortion with different tangential velocities. Results are presented graphically, and a summary gives optimum speeds for different record diameters.

621.395.625.3:538.221 **2330**

Mixed Ferrites for Recording Heads—Herr. (*See* 2445.)

621.396.645.37.029.4 **2331**

Independent Control of Selectivity and Bandwidth—Villard. (*See* 2395.)

ANTENNAS AND TRANSMISSION LINES

621.39.09 **2332**

A New Solution of the Fundamental Problem of the Propagation of Electromagnetic Processes in a Multi-Wire System—N. A. Brazma. (*Compt. Rend. Acad. Sci.* (URSS), vol. 76, pp. 41–44; January 1, 1951. In Russian.)

621.392.09 **2333**

Surface-Wave Transmission Line—R. H. Nelson. (*Wireless Eng.*, vol. 28, p. 162; May, 1951.) Comment on 1300 of July (Barlow) and 563 of March (Rust).

621.392.22 **2334**

The Behaviour of Electromagnetic Waves in Highly Nonuniform Lines—H. Meinke. (*Z. angew. Phys.*, vol. 2, pp. 473–478; December, 1950.) The significant characteristic of the wave field in the region of a nonuniformity is the appearance of a wedge-shaped intrusion, resulting from longitudinal field components, in the field pattern in the neighborhood of the point of zero transverse electric field strength. This intrusion is calculated for the case of a field with constant curvature, and examples are given of the effects due to irregularities of arbitrary form.

621.392.26†:538.561 **2335**

Theory of the Excitation of Oscillations in a Waveguide by Means of a Linear Aerial—A. I. Akhiezer and G. Ya. Lyubarski. (*Zh. Tekh. Fiz.*, vol. 20, pp. 1049–1064; September, 1950.) One of the main problems in the theory of antennas is the determination of the current distribution in an antenna to which given electromotive forces are applied. It has been shown by Leontovich and Levin (2618 of 1945) that in the case of an antenna in unlimited space, the problem is reduced to the solution of a linear integro-differential equation. A study is here presented of the current distribution in a linear antenna mounted along the axis of a cylindrical waveguide. In this case it is necessary to solve an equation of the same type as for an antenna in unlimited space. No effective methods of solving this equation for an antenna of arbitrary dimensions are known. The discussion is therefore limited to the case of a sufficiently long and thin antenna and, using a method proposed by Leontovich and Levin (2618 of 1945), an approximate solution of the equation is found by expanding the current in a series of powers of the inverse logarithm of the ratio of the length to the radius of the antenna.

The following two cases are considered separately: (a) when the wavelength differs considerably from the critical wavelength of the waveguide; (b) when this difference is not great. The current distribution in a tuned antenna differs very much from that in an antenna in unlimited space. Simple formulas for determining the amplitudes of the waves excited in the waveguide are also derived.

621.392.26†:621.39.09 **2336**

A Study of Asymmetrical Electromagnetic Waves from the Open End of a Circular Waveguide—L. A. Vainshtein. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 485–488; September 21, 1950. In Russian.) The methods used in 1283 of 1949 are not applicable to the case of asymmetrical electromagnetic waves, since in this case, owing to diffraction, two longitudinal components (1) and (2) of the electric vector appear at the open end, and the problem therefore cannot be reduced to a single integral equation. By using a generalized method similar to that presented in an earlier paper on sound radiation (*Compt. Rend. Acad. Sci.* (URSS), vol. 58, p. 1957; 1947) an exact solution can nevertheless be found. A system of equations (7) to (10) is derived, and methods are indicated for solving it.

621.392.26†:621.39.09 **2337**

The Diffraction of Waves at the Open End of a Circular Waveguide with Diameter Greater than the Wavelength—L. A. Vainshtein. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 909–912; October 11, 1950. In Russian.) The physical meaning of the formulas (see 1283 of 1949 and 2336 above) determining the radiation field under the above conditions, is discussed. Starting with the case of symmetrical waves, formula (7) determining the radiation field in the back half-space ($0<\theta<\lambda/2$) is considered, θ being the angle between the Z axis and the radius vector of the field point. With increase of distance from the edge of the waveguide, the primary cylindrical waves gradually become spherical. From a corresponding formula (12) for the front half-space, ($\lambda/2<\theta<\lambda$), it follows that waves from different sections of the edge interfere with one another and produce a complex spherical wave. Similar results are also obtained in the case of asymmetrical waves, but the process of development is more complicated.

621.392.43 **2338**

The Exponential Line at Cut-off Wavelength and in the Stop Range—A. Ruhrmann. (*Arch. elekt. Übertragung*, vol. 4, pp. 401–412; October, 1950.) The theory of current and voltage distribution at cutoff wavelength is discussed. The definition of characteristic impedance used in the pass range may be retained, but the transmission equations assume an indefinite form and require transformation. The cases of termination by characteristic impedance, short-circuiting, and open-circuiting, are considered separately; understanding of the mode of operation with complex termination is facilitated by reference to a circle diagram. On the basis of quadripole theory, it is possible to define a characteristic impedance within the stop range, although no wave propagation is to be inferred from the equations or circle diagrams. Attenuation factor and transmission factor are defined, the concept of a wave propagation process analogous to that in the pass range being made possible by introducing complex parameters. See also 1583 of 1950.

621.392.5:681.142 **2339**

Magnetic Delay-Line Storage—An Wang. (PROC. I.R.E., vol. 39, pp. 401–407; April, 1951.) A number of magnetic cores are connected together to form a static magnetic delay line in which a series of binary digits can be stored and read out. The operation of this type of line is briefly described and carefully analyzed, and the optimum operating conditions are derived. The effects of eddy-current loss and leakage inductance are considered, and criteria for stability of the system are discussed.

621.396.67 **2340**

Wrotham Aerial System: Part 1—New Design of Slot-Radiator for V.H.F. Broadcasting—C. Gillam. (*Wireless World*, vol. 57, pp. 210–214; June, 1951.) An omnidirectional horizontal-polarization radiator with a gain of 9 db, is obtained with 32 folded slots arranged in 8 tiers of 4, spaced uniformly around a vertical cylinder. The slots are cophased and fed by a branched transmission line with impedance-matching transformers, and are designed to handle simultaneously either three 25-kw FM transmissions or one 25-kw FM and one 18-kw AM transmission in the frequency band 87.5 to 95 mc.

621.396.67 **2341**

The Aerial Installations for the [German] Post-1945 High Power Transmitters—W. Berndt. (*Telefunken Ztg.*, vol. 23, pp. 39–52;

September, 1950.) Descriptions are given of the antenna installations for the medium- and long-wave broadcast and telegraphy transmitters described in 2565 below. A certain amount of improvisation was necessary. A feature common to all the installations is the spatial separation of transmitter and antenna system, the two being connected by hf cable.

621.396.67 2342

A Helix Theorem—J. D. Kraus. (Proc. I.R.E., vol. 39, p. 563; May, 1951.) For helical antennas of at least a few turns, with pitch angles of 10° to 15°, it is postulated that "when the circumference of an axial or end-fire helix is about one wavelength, (T_1 transmission mode dominant), there is a band of frequencies over which the phase velocity of wave propagation on the helix tends toward a value that makes the directivity a maximum."

621.396.67 2343

Radiation Properties of Spherical Antennas as a Function of the Location of the Driving Force—P. R. Karr. (*Bur. Stand. Jour. Res.*, vol. 46, pp. 422–436; May, 1951.) A theoretical analysis is made of the radiation from a conducting sphere fed at a narrow nonequatorial zone. Variations of radiation pattern, current distribution, and input admittance with the latitude of the feed zone, are studied. As long as the radius a of the sphere does not exceed $\lambda/2\pi$, the radiation conductance varies approximately as $\sin^4 \theta_0$, where θ_0 is the colatitude of the feed zone. For $a>\lambda/2\pi$, the maximum value of radiation conductance may occur at values of θ_0 other than 90°.

621.396.67 2344

Slot Radiators—N. A. Begovich. (Proc. I.R.E., vol. 39, p. 508; May, 1951.) Correction to paper abstracted in 2711 of 1950.

621.396.67 2345

Biconical Electromagnetic Horns—W. L. Barrow, L. J. Chu, and J. J. Jansen. (Proc. I.R.E., vol. 39, pp. 434–435; April, 1951.) Corrections to paper noted in 1404 of 1940.

621.396.67:538.566 2346

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 1—Transmission between Elliptically Polarized Antennas—V. H. Rumsey. (Proc. I.R.E., vol. 39, pp. 535–540; May, 1951.) A method of analysis is discussed which makes use of the impedance concept of transmission-line theory. It is shown that P, the ratio of two orthogonal tangential components of the electric field, is related to q, the ratio between the left- and right-handed circularly polarized components corresponding to the orthogonal components, in the same way as impedance is related to reflection coefficient. Representation of polarization on a transmission-line impedance chart is described, and solutions of various polarization problems in terms of impedance analogies are discussed.

621.396.67:538.566 2347

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 2—Geometrical Representation of the Polarization of a Plane Electromagnetic Wave—Deschamps. (*See* 2418.)

621.396.67:538.566 2348

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 3—Elliptically Polarized Waves and Antennas—Kales. (*See* 2419.)

621.396.67:53°.566 2349

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 4—Measurements on Elliptically Polarized Antennas—J. I. Bohnert. (Proc. I.R.E., vol. 39, pp. 549–552; May, 1951.) Two methods for measuring polarization characteristics are outlined. One uses a rotating linear-polarization antenna, the other uses two circular-polarization antennas.

621.396.671.012 2350

Pattern Calculator for A.M.—G. R. Mather. (*Electronics*, vol. 24, pp. 100–101; April, 1951.) A graphical method for calculation of the radiation pattern of two-tower directive arrays.

621.396.677 2351

The Electric and Magnetic Constants of Metallic Delay Media containing Obstacles of Arbitrary Shape and Thickness—S. B. Cohn. (*Jour. Appl. Phys.*, vol. 22, pp. 628–634; May, 1951.) Methods of deriving the dielectric constant and permeability of a metallic-obstacle medium are given. The equivalent shunt capacitive susceptance and series inductive reactance of the individual obstacles are also determined. By means of a correspondence established between an infinitely thin conducting obstacle and an aperture in an infinitely thin conducting wall, formulas are derived for different shapes of obstacle. The effect on the magnetic field of obstacles of moderate thickness is evaluated. For obstacles of arbitrary shape and thickness, the constants can be determined by the electrolyte-tank method; this is demonstrated for a particular example.

621.396.677 2352

Directional Aerials for Radio Stations—K. O. Schmidt. (*Fernmeldetech. Z.*, vol. 4, pp. 49–56; February, 1951.) Review and discussion of different types of radiators for use at wavelengths between 3 cm and 3 m.

621.396.677:621.396.11 2353

A Wide Band Aerial System for Circularly Polarized Waves, Suitable for Inospheric Research—G. J. Phillips. (Proc. I.R.E., vol. 98, pp. 237–239; Part III, May, 1951.) The system described can be used to select, without readjustment, one or other of two waves circularly polarized in opposite senses and incident vertically, within the frequency range 2 to 6 mc. Two mutually perpendicular horizontal dipoles are associated respectively with two phase-shifting LC lattice networks. In the antennas, emfs initially 90° out of phase, may be added in phase, thus giving selection of a circularly polarized component. A discrimination ratio of at least 12:1 between the components has been obtained.

621.396.677.001.4 2354

Reflecting Surface to Simulate an Infinite Conducting Plane—S. J. Raff. (*Jour. Appl. Phys.*, vol. 22, pp. 610–613; May, 1951.) A finite reflecting surface, which simulates an infinite plane, is required for calibrating measurements of reflections back to a microwave transmitting antenna. The Fresnel-zone method of physical optics is used for the design calculations. Variational calculus is used to determine the optimum reflector shape for a given antenna pattern, reflector size, and range of antenna-to-reflector distance. Theoretical values are compared with results obtained on an example constructed for use at 25λ from a dipole antenna.

621.392 2355

Transmission Lines and Networks [Book Review]—W. C. Johnson. Publishers: McGraw-Hill Book Co., New York, N. Y., 1950, 361 pp., $5.00. (*Electronics*, vol. 24, pp. 278–280; April, 1951.) "Although basically a textbook for undergraduate students, the material covered should be of considerable interest to practicing engineers in both the power and communication fields."

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.6:621.317.726 2356

Calculation of CR Elements for the case of Varying Voltage and/or Nonlinear Resistances—H. Elger. (*Arch. elekt. Übertragung*, vol. 4, pp. 413–426; October, 1950.) Methods are presented for calculating the buildup conditions in CR elements when either steady or time-varying voltage is applied. Consideration is given to potential-divider circuits. The effect of nonlinear resistances is investigated, as, e.g., in the demodulation of rectified modulated voltage with a square-law rectifier. Methods based on the theory are discussed for varying time constants within wide limits, and an example shows how the readings of a diode peak-voltage meter, operating at very high voltages, may be corrected for errors due to small discharges through the insulation between pulses.

621.3.015.7:621.387.4† 2357

Single-Channel Analyzer—J. E. Francis, Jr., P. R. Bell, and J. C. Gundlach. (*Rev. Sci. Instr.*, vol. 22, pp. 133–137; March, 1951.) An analyzer for proportional and scintillation counters counts the number of pulses the heights of which lie between E and $E+\Delta E$; ΔE is constant to within ± 1.2 per cent for $E=0-90v$.

621.3.016.352 2358

Relation of Nyquist Diagram to Pole-Zero Plots—H. F. Spirer. (Proc. I.R.E., vol. 39, p. 562; May, 1951.) Comment on 1086 of June (Harman).

621.314.212:534.232:538.652 2359

Electroacoustic Transformation by means of Magnetostriction, with Special Reference to Radiation from Transformers—Rust. (*See* 2313.)

621.314.3† 2360

High-Gain Magnetic Amplifier—R. Feinberg. (*Wireless Eng.*, vol. 28, pp. 151–155; May, 1951.) Self-excitation is an effective method of obtaining feedback in a transductor. High values of current amplification can be obtained by making the number of turns of the self-excitation winding sufficiently large in relation to the number of turns of the load winding. Turns relations for stable operation are discussed. When operating unstably, the system may be used as an on-off trigger relay.

621.314.3†:621.318.42 2361

Design of [Magnetic-] Amplifier Inductors with Series-Connected Ohmic Loads—E. Helmes. (*Arch. elekt. Übertragung*, vol. 5, pp. 39–46; January, 1951.) A formula is derived expressing the effective permeability and self-inductance and the transfer impedance between the coil and the load in terms of the core cross section and the number of turns. Families of curves are plotted from measurements on (a) a two-element inductor with transformer-sheet core, (b) a three-element inductor with mumetal core. These curves can be applied to other core materials by changing the scale.

621.314.3†:[621.396.615.17+621.396.619.2 2362

The Use of Saturable Reactors as Discharge Devices for Pulse Generators—W. S. Melville. (*Proc. IEE*, vol. 98, pp. 185–204; Discussion, pp. 204–207; Part III, May, 1951.) The development of materials used for saturable reactors is outlined; a magnetic material with rectangular hysteresis loop is required. Magnetic discharge devices can often replace electronic discharge devices. The merits of the two types are compared. The design and operation of saturable-reactor circuits for radar pulse-modulation and ignitron ignition are described.

621.315.592†+621.314.632 2363

The Characteristics and some Applications of Varistors—F. R. Stansel. (Proc. I.R.E., vol. 39, pp. 342–358; April, 1951.) A tutorial paper reviewing the properties of semiconductor nonlinear circuit elements, with particular reference to those available commercially. The principles and limitations governing the use of these elements are summarized. The varying importance of the different parameters with different types of application is illustrated by short discussions of the design of voltage limiters, power rectifiers, hf modulators, and compandors.

621.316.726:681.142 2364

Automatic Frequency Control—J. M. M. Pinkerton. (*Electronic Eng.*, vol. 23, pp. 142–143; April, 1951.) Description of a method of controlling the clock pulse frequency in the storage system of a digital computer. The phase of a pulse, which has traveled down an ultrasonic delay line, is compared with that of a later pulse of the same series which has not been delayed. The phase difference is used to derive a voltage for control of the frequency of the master oscillator producing the clock pulses, by means of a reactance tube.

621.316.86 2365

Pyrolytic Film Resistors: Carbon and Borocarbon—R. O. Grisdale, A. C. Pfister, and W. van Roosbroeck. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 271–314; April, 1951.) A description of the production and structure of thin carbon films, deposited on ceramics or fused silica by the pyrolysis of hydrocarbon vapors, which are capable of providing resistors of high stability with resistances from a few ohms to tens of megohms. The incorporation of boron in the film results in a smaller temperature coefficient than that of many wire-wound resistors, and the negligible skin effect permits advantageous use of these film resistors at high frequencies. Resistance values up to $10^9\Omega$ have been obtained in the borocarbon type. See also 583 of April.

621.317.353.2.012.3 2366

Mixer Harmonic Chart—T. T. Brown. (*Electronics*, vol. 24, pp. 132, 134; April, 1951.) The chart facilitates identification of spurious frequencies resulting from beating of various harmonics of two inputs, the frequency of one being variable.

621.318.572 2367

A Three-State Flip-Flop—A. D. Booth and J. Ringrose. (*Electronic Eng.*, vol. 23, p. 133; April, 1951.) With Type-6J6 and Type-6SN7 tubes, particular values of cathode resistor were found to give three stable states in flip-flop units. An explanation is given of the circuit operation.

621.319.53:621.396.9 2368

High-Voltage Pulse Modulators for Radar Pulse Transmitters—Tigler. (*See* 2431.)

621.385.3:546.289 2369

Duality as a Guide in Transistor Circuit Design—R. L. Wallace, Jr., and G. Raisbeck. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 381–417; April, 1951.) The properties of a transistor are compared with those of a vacuum-tube triode, and the relation between them is found to be such that, by interchanging current and voltage, a known vacuum-tube circuit can be transformed into one suitable for use with transistors. The necessary changes in the circuit elements are considered, and practical examples of these networks (or duals) are given. Circuits which permit the simultaneous use of vacuum tubes and transistors, such as the Doherty amplifier, are also discussed.

621.392 2370

The Potential Analogue Method of Network Synthesis—S. Darlington. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 315–365; April, 1951.) The method developed is based on the analogy between the gain and phase of linear networks and the two-dimensional potential and stream functions produced by charges corresponding to the network singularities.

621.392 2371

The Synthesis of RC Networks to Have Prescribed Transfer Functions—H. J. Orchard. (PROC. I.R.E., vol. 39, pp. 428–432; April, 1951.) A more general method than that of Guillemin (2462 of 1949) is described. The resulting network is in the form of a lattice, and is capable of providing any transfer function physically realizable by an RC network. The design procedure is simple. A numerical example is included.

621.392.4 2372

An Application of Equaliser Curves to the Design of Two-Terminal Networks—P. W. Seymour. (*P.O. Elect. Eng. Jour.*, vol. 44, Part I, pp. 31–32; April, 1951.) Description of a method for deriving the circuit constants of a 2-terminal network from available design data for a 4-terminal constant-impedance equalizer having an insertion-loss/frequency characteristic similar to the impedance/frequency characteristic required for the 2-terminal network.

621.392.5 2373

Fluctuations in a Linear System with Periodically Varying Parameters—S. I. Borovitski. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 233–236; September 11, 1950. In Russian.) A system is considered with parameters varying in accordance with the equation (1). Statistically, the behavior of such a system can be represented by the Einstein-Fokker equation (2). With the aid of the main solution (3) of this equation, the case of a statistically stable system is investigated. The spectrum of a perturbation consists of discrete lines on a continuous background. The discussion is illustrated by experimental curves obtained in analyzing the output of a super-regenerative receiver in the absence of a signal.

621.392.5 2374

Response Characteristics of Resistance-Reactance Ladder Networks—R. R. Kenyon. (PROC. I.R.E., vol. 39, pp. 557–559; May, 1951.) Generalized expressions for the transfer functions for resistance-reactance ladder networks are given and discussed in detail. The output voltage as a function of time is derived for the cases of unit-impulse and unit-step input voltage. Two methods are discussed for determining the output voltage for the case of an arbitrary input.

621.392.5 2375

Passive Pulse-Sharpening Circuits—L. Reiffel. (*Rev. Sci. Instr.*, vol. 22, pp. 214–216; March, 1951.) A circuit is described using a Ge damping diode in conjunction with a resonant circuit to shorten pulses obtained from a GM counter or other pulse detector.

621.392.52 2376

Relations between Signals and Spectra—K. Fränz. (*Arch. elekt. Übertragung*, vol. 5, pp. 10–14; January, 1951.) Theoretical proof of certain laws of filter theory. The major part of the energy of any signal of given duration is confined to a narrow spectrum whose limits are independent of the waveform. The connection between filter bandwidth and signal buildup is derived. The pass-band response curve of a filter with monotonic buildup must be roughly bell-shaped.

621.392.52:621.396.611.21 2377

Lattice-Type Crystal Filter—R. Lowrie. (*Electronics*, vol. 24, pp. 129–131; April, 1951.) Description of a 2-section filter incorporating eight crystals, with a pass band of width 3.9 kc centered at 2 mc, and a bandwidth of 12 kc at 60 db attenuation.

621.394/.396].6 2378

A New Colour-Coded Wiring System—N. G. Partridge. (*Electronic Eng.*, vol. 23, pp. 138–139; April, 1951.) The method is based on the consecutive numbering of all the items involved, whether wires, cableforms, chassis, or racks, using the international color-code system to enable the allotted number to be carried by each item in the form of colored bands or labels.

621.394/.396].6:621.392.012 2379

Correlation of Circuit Diagram and Wiring Development of Electronic Systems—A. W. Keen. (*Electronic Eng.*, vol. 23, pp. 144–145; April, 1951.)

621.396.6:061.4 2380

Trends in Components—(*Wireless World*, vol. 57, pp. 185–188; May, 1951.) A survey of the components and accessories shown at the annual private exhibition organized by the Radio and Electronic Component Manufacturers' Federation in London, April, 1951.

621.396.611+621.317.7].029.63 2381

Circuits and Measurement Apparatus for the 30-cm Band—Safa. (*See* 2479.)

621.396.611.21 2382

Some Notes on Overtone Crystals and Maintaining Oscillators operating in the Frequency Range of 33-55 Mc/s—J. B. Supper. (*Proc. IEE*, vol. 98, pp. 240–247; Part III, May, 1951.) The terms "minimum impedance" and "inductive impedance" are proposed for identifying two general forms of maintaining circuit investigated, and the concept of "impedance diameter" as a measure of crystal goodness is introduced. Measurements on different types of crystal and their behavior in the inductive-impedance oscillator, are recorded and discussed. The effect of plating area on crystals is considered, and attention is drawn to the improvement obtained by reducing the plating diameter below the present standard value.

621.396.611.21 2383

Amplitude of Vibration in Piezoelectric Crystals—E. A. Gerber. (*Electronics*, vol. 24, pp. 142, 216; April, 1951.) The amplitude of vibration of a crystal has some influence on the crystal parameters. Simple expressions are derived relating amplitude to the rf current through the crystal and to the voltage across it, and comparison is made with the results of more general theories of a vibrating piezoelectric plate. Only thickness modes of vibration are considered. Experiments verifying the formulas are described.

621.396.611.3 2384

The Calculation of the Input Impedance of Coupled Oscillatory Circuits—F. A. Fischer and U. John. (*Arch. elekt. Übertragung*, vol. 5, pp. 33–38; January, 1951.) Description of a method based on the representation of input impedance as a function of the difference between the damping and the resonance frequencies of the two circuits and the coupling factor. Curves and diagrams are plotted for a typical case. The basic formula is extended to the case of *n* coupled circuits.

621.396.611.4 2385

Application of the Method of Curvilinear Coordinates to the Calculation of a Π-Type Cavity Resonator—V. L. Patrushev. (*Zh. Tekh. Fiz.*, vol. 20, pp. 727–734; June, 1950.) Krasnushkin has studied the propagation of waves in waveguides of rectangular cross section by using the method of normal waves. In a previous paper (*Bull. Acad. Sci.* (URSS), *Sér. phys.*, vol. 12, p. 684; 1948) the author applied this method to the investigation of the em fields and natural frequencies of cavity resonators having rotational symmetry. Under certain conditions, a Π-type cavity resonator can be regarded approximately as a coaxial line with capacitance loading, and therefore belonging to this group of resonators. In the present paper it is shown that by introducing curvilinear co-ordinates, a rigorous solution of the problem is possible in principle.

621.396.611.4 2386

Design of the Π-Type Cavity Resonator—V. L. Patrushev and O. V. Romanova. (*Zh. Tekh. Fiz.*, vol. 20, pp. 798–801; July, 1950.) A formula is derived for determining the length of the plunger used for tuning the resonator. The discussion is illustrated by two numerical examples which have been verified experimentally.

621.396.615.17 2387

Theory of the Symmetrical Multivibrator—N. A. Zheleztsov. (*Zh. Tekh. Fiz.*, vol. 20, pp. 788–797; July, 1950.) The tube characteristic is assumed to consist of a number of linear sections. Analysis of the movement of the operating point along these sections makes it possible to determine the buildup of the discontinuous oscillations, and to prove the singularity and stability of the discontinuous periodic solution.

621.396.645 2388

Application of the Properties of Newtonian Potentials to the Design of Frequency-Modulation Amplifiers—P. Belgodère and A. Fromageot. (*Onde élect.*, vol. 31, pp. 18–32; January, 1951.) Use of a constant-gain amplifier to provide constant group-transmission time, (856 of 1950) requires that the width of the pass band be unnecessarily large. From a theoretical treatment, an alternative method of design is derived. This has yet to be checked experimentally, in particular as regards tolerances on tuning frequencies and circuit parameters.

621.396.645:535.247.4 2389

A Balance Indicator with High Input Impedance using a Cathode Follower—Dighton. (*See* 2513.)

621.396.645:621.317.083.4 2390

Sensitive Null Detector—M. G. Scroggie. (*Wireless World*, vol. 57, pp. 175–178; May, 1951.) Description of a selective af bridge amplifier for use at frequencies between 50 and 1,500 cps, with a "magic eye," milliammeter, or telephones as the output indicator. An agc circuit permits a range of signal input of 10 μv to 10v, and the time constant is such that a transient indication is given of a change in input at any signal level.

621.396.645:621.317.6 2391

The Determination of Amplifier Sensitivity with the Aid of the Noise Diode—Squires. (*See* 2478.)

621.396.645.012.8 2392

Network Representation of Input and Output Admittances of Amplifiers—F. W. Smith. (Proc. I.R.E., vol. 39, p. 439; April, 1951.) Comment on 2194 of 1949 (Vallese).

621.396.645.211 2393

Maximum Output from a Resistance-Coupled Triode Voltage Amplifier—J. M. Diamond. (Proc. I.R.E., vol. 39, pp. 433–434; April, 1951.) Simple formulas are derived for optimum load resistance with respect to output voltage, and for maximum voltage swing obtainable.

621.396.645.36.029.4 2394

Bridge-Compensated Differential Amplifiers—J. Labus. (*Arch. elekt. Übertragung*, vol. 4, pp. 437–440; October, 1950.) Sensitive push-pull amplifiers used for special purposes (e.g., electrocardiography), are liable to interference from ac fields at the input terminals. Several previously proposed circuits for eliminating this interference are briefly reviewed, and a method using a resistance-bridge network connected across the input, is described. This suppresses the interfering voltages before they reach the grids of the first-stage tubes, and hence prevents the production of harmonics.

621.396.645.37.029.4 2395

Independent Control of Selectivity and Bandwidth—O. G. Villard, Jr. (*Electronics*, vol. 24, pp. 121–123; April, 1951.) The feedback circuit in an RC af amplifier is designed to have constant amplitude/frequency but variable phase/frequency characteristics. The feedback is positive at the resonance frequency, negative at frequencies far from it. The complete circuit is shown for an amplifier with a constant percentage bandwidth/frequency variation, and a choice of three bandwidths at any desired selectivity.

621.396.822 2396

Thermal Fluctuation of Charge in Linear Circuits—E. A. N. Whitehead. (*Elliott Jour.*, vol. 1, pp. 32–34; March, 1951.) Derives the usual expressions for the noise power developed across an impedance; the noise generators are considered as being in parallel with the various circuit components.

621.392 2397

Transmission Lines and Networks [Book Review]—Johnson. (*See* 2355.)

621.392.025 2398

Alternating Current Circuits [Book Review]—R. M. Kerchner and G. F. Corcoran. Publishers: J. Wiley & Sons, New York, N. Y., 3rd edn. 1950, 586 pp., $5.50. (Proc. I.R.E., vol. 39, p. 448; April, 1951.) ". . . One of the very best summaries of the elementary background of ac circuit analysis . . . ."

GENERAL PHYSICS

534.01+538.56 2399

Theory of Waves and Oscillations in Nonhomogenous Discrete Structures—P. E. Krasnushkin. (*Zh. Tekh. Fiz.*, vol. 20, pp. 1065–1083; September, 1950.) A general discussion applicable to various discrete oscillating systems, such as molecular chains of certain organic compounds, nonuniform waveguides and strings, the ionosphere, etc. The following two types of oscillations are met with in such systems: (a) "collective" oscillations of sinusoidal form spread more or less uniformly over all elements of the system, and (b) "local" oscillations of exponential form in parts of the space occupied by the system. In order to investigate the nature of these two types of oscillations and the conditions of their appearance, consideration is given to the general case of a chain structure consisting of cells, each of which represents an oscillating system with one degree of freedom. The local oscillations approximate in frequency and shape the natural oscillations of isolated cells, and the collective oscillations appear as a result of the interaction between the local oscillations, depending on the degree of resonance coupling, a conception introduced by Mandelstam. The loosening of the coupling may result in the appearance of collective oscillations only within parts of the system, and their penetration into other parts in the form of exponential "tails." Since the collective oscillations are essentially standing waves, the regions limiting them are called wave barriers. Three different types of these barriers are specified, and their effect on the operation of the system is discussed.

535.12 2400

Wave Propagation in an Anisotropic Medium and the Corresponding Principal Directions—M. Pastori. (*Nuovo Cim.*, vol. 6, pp. 187–193; May 7, 1949.) At any point within the medium, at least three principal directions exist such that the sum of the squares of the velocities of the three corresponding wavefronts is constant.

535.215:538.221 2401

The Surface Photoelectric Effect in Ferromagnetics—S. V. Vonsovski and A. V. Sokolov. (*Compt. Rend. Acad. Sci.* (URSS), vol. 76, pp. 197–200; January 11, 1951. In Russian.) The anomalies in the photoelectric effect in ferromagnetics, observed by Cardwell (*Phys. Rev.*, vol. 76, p. 125; 1949) are discussed from the standpoint of the interaction between the outer *s*- and inner *d*-electrons, a concept developed by Vonsovski (2074 of 1947). The photoelectric current and the effective work function in ferromagnetics depend on the value of their spontaneous magnetization.

537.529 2402

A Review of Spark Discharge Phenomena—F. M. Bruce. (*Jour. Brit. IRE*, vol. 11, pp. 121–135; April, 1951.) The Townsend and streamer theories are discussed; the latter requires further evidence for its full substantiation, while the range of application of the former is likely to be greatly increased. Results of investigations in progress will be of importance, not only in the field of measurement over a wide variety of waveforms, but also in the extended use of gaseous insulation. Consideration is given to the uniform-field gap for standardizing methods of measurement. HF breakdown is not discussed here; this was dealt with in a paper noted in 613 of April.

537.533.8 2403

Some Peculiarities of the Secondary-Electron Emission from Thin Films of Calcium Chloride—V. N. Favorin. (*Zh. Tekh. Fiz.*, vol. 20, pp. 916–922; August, 1950.)

537.562:537.311 2404

Convergence of the Chapman-Enskog Method for a Completely Ionized Gas—R. Landshoff. (*Phys. Rev.*, vol. 82, p. 442; May 1, 1951.) A note relevant to 335 of March (Cohen, Spitzer, and Routly).

537.71 2405

Generalized Electrical Formulas—V. P. Hessler and D. D. Robb. (*Elec. Eng.*, vol. 70, pp. 332–336; April, 1951.) "A set of generalized electrical formulas is developed to which units of any absolute system may be applied. The generalization is accomplished with the aid of two additional constants, *n* and *u*. The general formulas may be reduced to the usual rationalized or unrationalized forms, or to the Gaussian or Heaviside forms, by the substitution of tabulated numerical values of the constants *n* and *u* in the general form."

538.221 2406

Single-Domain Structure in Ferromagnetics, and the Magnetic Properties of Finely Dispersed Substances—E. Kondorski. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 213–216; September 11, 1950. In Russian.)

538.221 2407

The Dependence of Magnetization Curves on Temperature and the Hysteresis Loop of High-Coercivity Alloys—Ya. S. Shur and N. A. Baranova. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 225–228; September 11, 1950. In Russian.)

538.24 2408

The Effect of Directed Stresses on the Shape of the Magnetization Curve in Strong Fields—L. V. Kirenski and L. I. Slobodskoi. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 457–459; September 21, 1950. In Russian.) A formula (1) expressing the intensity of magnetization is quoted, and relations between the various constants are discussed, particularly for the case when the elastic stresses in the sample are directed along the magnetizing field.

538.249 2409

On Certain Laws Governing Magnetic Viscosity—R. V. Telesnin. (*Compt. Rend. Acad. Sci.* (URSS), vol. 75, pp. 659–660; December 11, 1950. In Russian.)

538.249 2410

The Dependence of Magnetic Viscosity on Temperature—R. V. Telesnin and E. F. Kuritsyna. (*Compt. Rend. Acad. Sci.* (URSS), vol. 75, pp. 797–798; December 21, 1950. In Russian.)

538.56 2411

Applications of the Radiation from Fast Electron Beams—H. Motz. (*Jour. Appl. Phys.*, vol. 22, pp. 527–535; May, 1951.) The radiation from beams of fast electrons, passing through a succession of transverse electric or magnetic fields of alternating polarity, is examined. The frequencies and the angular distribution of radiated energy are calculated; the coherence of the radiation is discussed. Several applications appear possible, one of which is the production of millimeter waves of considerable power. Another is the monitoring of the speed of electrons having energies up to 10° ev.

538.56:535.42 **2412**

Diffraction of Centimetre Electromagnetic Waves by Metal Disks—H. Severin. (*Z. angew. Phys.*, vol. 2, pp. 499–505; December, 1950.) The diffraction phenomena observed along the axis of, and close up to, a conducting disk, for normal incidence of a plane wave, are compared with three approximate theoretical solutions. A wavelength of 10 cm and disks of thickness 2 mm and radius 0.5λ, 1λ, 1.5λ, and 2λ, were used. Best agreement with measurements is provided by a theory which assumes the disk to be covered with a layer of magnetic dipoles.

538.56:535.42 **2413**

On the Diffraction of a Plane Electromagnetic Wave by a Paraboloid of Revolution—C. W. Horton and F. C. Karal, Jr. (*Jour. Appl. Phys.*, vol. 22, pp. 575–581; May, 1951.) A theoretical investigation of the diffraction by the convex surface of the paraboloid. Expressions are derived for the components of the incident, scattered and refracted waves. The case of a perfectly conducting paraboloid and a wave front perpendicular to the axis of rotation, is considered. The variation of amplitude of the scattered wave with distance along axis of rotation is compared with the corresponding curve for a sphere of radius equal to the radius of curvature of the paraboloid at its nose.

538.56:535.42 **2414**

On the Diffraction of Electromagnetic Waves by Two Conducting Parallel Half-Planes—M. G. Cheney, Jr., and R. B. Watson. (*Jour. Appl. Phys.*, vol. 22, pp. 675–679; May, 1951.) The diffraction produced by the edges of two half-planes, arranged one behind the other relative to the signal source, is investigated experimentally and theoretically. Agreement between the two sets of results is not very close.

538.561 **2415**

The Problem of the Excitation of Electromagnetic Oscillations—B. Ya. Moyzhes. (*Zh. Tekh. Fiz.*, vol. 20, pp. 698–715; June, 1950.) A general method has been recently proposed by G. A. Grinberg ("Selected Problems of the Mathematical Theory of Electric and Magnetic Phenomena", published by the Academy of Sciences of U.S.S.R., 1948) for solving a large group of problems in connection with the excitation of waveguides and other systems by a given distribution of currents, or by slots for which the tangential component of the electric field is known. In this method one or, more generally, two independent equations, are derived from Maxwell's equation. Each of these equations includes a scalar function (field component or a component of auxiliary function potentials) for which the boundary conditions have to be established separately. In the present paper this method is discussed in detail and applied to the cases of a cylindrical waveguide and a sectoral horn.

538.566 **2416**

Synthesis and Analysis of Elliptic Polarization Loci in Terms of Space-Quadrature Sinusoidal Components—M. G. Morgan and W. R. Evans, Jr. (Proc. I.R.E., vol. 39, pp. 552–556; May, 1951.) A mathematical analysis of elliptically polarized waves, by means of which the elliptic locus, resulting from three mutually orthogonal component field vectors, may be specified in terms of those vectors, or the vectors specified in terms of the locus. The simpler case of two-component vectors is considered first.

538.566:535.43 **2417**

Electromagnetic Scattering from Spheres with Sizes Comparable to the Wavelength—A. L. Aden. (*Jour. Appl. Phys.*, vol. 22, pp. 601–605; May, 1951.) The general formula for the back-scattering cross section of a sphere is difficult to evaluate for a complex refractive index, owing to the lack of the necessary tables of Bessel functions. The evaluation may be carried out by transforming the formula by means of logarithmic derivative functions. Back-scattering cross sections were measured for water spheres with sizes comparable to the wavelength (16.23 cm) using a standing-wave method. The water was contained in a thin hemispherical shell of dielectric, mounted on an aluminium disk. Very good agreement with theory was obtained.

538.566:621.396.67 **2418**

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas: Part 2—Geometrical Representation of the Polarization of a Plane Electromagnetic Wave—G. A. Deschamps. (Proc. I.R.E., vol. 39, pp. 540–544; May, 1951.) The polarization and amplitude of an elliptically polarized plane wave may be specified by three quantities which define the ellipse traced out by the field vector, and which may be represented, according to a method used by Poincaré in optics, by the co-ordinates of a point on a sphere. Methods of solving problems using this concept are discussed.

538.566:621.396.67 **2419**

Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas. Part 3—Elliptically Polarized Waves and Antennas—M. L. Kales. (Proc. I.R.E., vol. 39, pp. 544–549; May, 1951.) A complex vector algebra is presented by means of which an elliptically polarized wave may be completely specified. The field at a point may be resolved into two space components, in general not in the same direction, having time variations in phase quadrature. Thus if the two space vectors are U_r and U_i, the complex vector given by U_r+jU_i completely defines the field at the point. The algebraic properties of such vectors are discussed, and resolution of the field into orthogonal elliptically polarized components, and the concept of phase, are studied. Relations useful in measurements and antenna problems are given.

537.311.33 **2420**

Semi-Conductors. [Book Review]—D. A. Wright. Publishers: Methuen and Co., London, Eng., 130 pp., 7s. 6d. (*Wireless Eng.*, vol. 28, p. 164; May, 1951.) "Specially concerned with the theory of electron flow in semi-conductors, and across the boundary between them and either a metal or a vacuum. . . . This monograph will undoubtedly be of great use to students of the electron physics of solids."

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523+551+621.396.93 **2421**

Recent Work of the Radiophysics Division C.S.I.R.O.—E. G. Bowen. (Proc. I.R.E. (Australia), vol. 12, pp. 99–108; April, 1951.) Developments in radio techniques for meteorology, astronomy, and navigation, are described. A digital computer has been designed and built by the Division to deal with its internal computing requirements.

523.72:621.396.822 **2422**

Radio Helioscopy—M. Waldmeier. (*Naturwiss.*, vol. 38, pp. 1–4; January, 1951.) A survey based on papers presented at the General Assembly of the International Scientific Radio Union. Results of solar-radiation measurements within the wavelength range 1 cm to 10 m, are discussed, and the adequacy of theories so far advanced, is examined. Most recent observations tend to support theories according to which the radiation disturbances are caused by coronal plasma oscillations excited by corpuscular rays or protuberances in motion.

523.746:538.12 **2423**

The Propagation of the Electromagnetic Field of Sunspots in the Sun's Atmosphere—P. E. Kolpakov and Ya. P. Terletski. (*Compt. Rend. Acad. Sci.* (URSS), vol. 76, pp. 185–188; January 11, 1951. In Russian.) Because of its electrical conductivity, the highly ionized atmosphere of the sun might be expected to act as a screen for the electromagnetic field of the sunspots. That indications of this field are nevertheless observed in the middle and upper layers of the sun's atmosphere, is due to motion of the latter. A mathematical analysis of the propagation process, leading to the derivation of two differential equations (bottom of p. 187) is presented.

537.591:[523.854:621.396.822 **2424**

Cosmic Rays as a Source of Galactic R.F. Radiation—V. L. Ginzburg. (*Compt. Rend. Acad. Sci.* (URSS), vol. 76, pp. 377–380; January 21, 1951. In Russian.) Galactic rf radiation cannot be explained by the thermal radiation of interstellar electrons. It is suggested that it may be produced by the radiation from the electronic component of cosmic rays traveling at relativistic velocities in the magnetic fields near and between the stars. On this assumption, and knowing the intensity of the magnetic field, it is possible to determine the concentration of the corresponding cosmic particles. Results of a detailed analysis are given separately for the cases of general galactic emission, and emission from discrete sources. The possibility of cosmic-ray radiation in the magnetic field of the earth is also discussed. The results obtained are not conclusive.

550.38 **2425**

Geomagnetic Indices for the Period from 22nd Dec. 1950 to 31st March 1951—(*Z. Met.*, vol. 5, p. 123; April, 1951.) Observations made at Niemegk are presented in chart form.

551.510.535 **2426**

Sporadic E Movements on 21 June 1949—N. C. Gerson. (*Tellus*, vol. 3, pp. 56–59; February, 1951.) An analysis of the reports from about 300 American radio amateurs, of radio contacts made by reflection from sporadic-*E* regions. Between midnight and 0400 hours, two clouds of E_s ionization were reported which drifted westward across the United States with average velocities of about 250 km/hr.

551.510.535 **2427**

Magneto-Ionic Triple Splitting of Ionospheric Waves—O. E. H. Rydbeck. (*Onde élect.*, vol. 31, pp. 70–81 and 153–156; February and March, 1951.) French version of paper noted in 1147 of June.

551.510.535:621.396.11 **2428**

The Effect of the Lorentz Polarization Term on the Vertical Incidence Absorption in a Deviating Ionosphere Layer—J. M. Kelso. (Proc. I.R.E., vol. 39, pp. 412–419; April, 1951.) Using the two parabola approximations, and neglecting the effect of the earth's magnetic field, the influence of the Lorentz term on the apparent scale height and total absorption of a Chapman layer is calculated for a wave reflected in the layer. The absorption is increased by 4 per cent or more when the Lorentz term is included. See also 638 of 1950.

551.594.25 **2429**

The Origin of the Electric Charge on Rain—J. A. Chalmers. (*Quart. Jour. R. Met. Soc.*, vol. 77, pp. 249–259; April, 1951.) Previously measured values of charge on rain during periods when point discharge occurs, can be accounted for quantitatively on reasonable assumptions in terms of the process of ion capture. The results show that separation of charge must operate at levels down to about 800 m, and the consequences of this are discussed in relation to theories of the process of separation of charge.

621.317.79:621.396.822 **2430**

High-Sensitivity High-Frequency Noise-Measurement Apparatus Calibrated Absolutely in KT₀ Units—Röschlau. (*See* 2487.)

LOCATION AND AIDS TO NAVIGATION

621.396.9:621.319.53 **2431**

High-Voltage Pulse Modulators for Radar Pulse Transmitters—H. Tigler. (*Arch. elekt. Übertragung*, vol. 5, pp. 47–51 and 91–98; January and February, 1951.) Descriptive review of 9 different circuits for pulse generation using vacuum tubes, thyratrons, and spark discharge systems. Control of the spark gap, and voltage multiplication by means of Marx circuits, are discussed. This circuit and the spark discharge are especially suitable for short pulses at high power and high voltage.

621.396.93+523+551 **2432**

Recent Work of the Radiophysics Division C.S.I.R.O.—Bowen. (*See* 2421.)

621.396.933 **2433**

A Source of Error in Radio Phase Measuring Systems—R. Bateman, E. F. Florman, and A. Tait. (PROC. I.R.E., vol. 39, pp. 436–438; April. 1951.) Discussion on 2515 of 1950.

621.396.933 **2434**

A General Survey of Electronics in Air Transport—C. H. Jackson. (*Jour. Brit. IRE*, vol. 11, pp. 139–155; Discussion, pp. 156–159; April. 1951.) Communications, navigation aids, and aids to approach and landing, are reviewed and related to standards of safety. Details of technique and function are not considered.

621.396.933 **2435**

Radio on the Airways—(*Wireless World*, vol 57, pp. 199–202; May, 1951.) A general description of mf omnidirectional beacons and radio ranges, and also vhf marker beacons, as used on the main air routes in Great Britain. The beacons and radio ranges, operating in the 200 to 400-kc band, provide airway entrance markers and 4-direction course indication, respectively. Position information is given by the vertically radiating vhf marker system.

621.396.9 **2436**

Radar Systems and Components [Book Review]—Bell Telephone Laboratories. Publishers: B. Van Nostrand Co., New York, N. Y., and Macmillan and Co., London, Eng., $7.50 or 56s. (*Engineering* (London), vol. 171, p. 392; April 6, 1951.) A collection of papers covering the magnetron, the klystron, the resonant cavity, radar antennas, etc. Valuable for the specialist and for the general student.

MATERIALS AND SUBSIDIARY TECHNIQUES

338.987.4:621.396/.397].6 **2437**

Conservation of Critical Materials—W. W. MacDonald. (*Electronics*, vol. 24, pp. 84–87; April, 1951.) Discussion of design modifications to economize in the use of scarce metals while maintaining receiver performance. Television receivers are particularly considered.

537.311.33 **2438**

Electrical Properties of Grey Tin—G. Busch, J. Wieland, and H. Zoller. (*Helv. Phys. Acta*, vol. 24, pp. 49–62; February 15, 1951. In German.) Grey tin of high purity was prepared by prolonged cooling of spectroscopically pure metallic tin, and numerous alloys were made by adding small amounts of Al. Conductivity was determined by measuring the *Q* factor of a coil with a core of grey tin powder, at frequencies up to 30 mc; at 0° C the value found was $5\times10^3\Omega^{-1}$ cm^{-1}. Hall effect and variation of resistivity with applied magnetic field were measured by conventional dc methods. Remarkably large variations of resistivity were observed. The experiments show that grey tin is a semiconductor of high electrical conductivity, with properties very similar to those of Si and Ge.

537.311.33 **2439**

The Effect of Pressure on the Electrical Resistance of certain Semi-Conductors—P. W. Bridgman. (*Proc. Amer. Acad. Arts and Sci.*, vol. 79, pp. 127–148; April, 1951.) Measurements made on Ge, Si, and several oxides, are described. The resistances of all the oxides decrease with rising temperature up to 200°C, but there is no common type of variation with change of pressure up to 50,000 kg/cm². The Ge and Si were investigated under hydrostatic conditions to 30,000 kg/cm² at room temperature only. Differences of behavior as between *n*- and *p*-types are indicated and discussed.

537.311.33 **2440**

The Diffusion of the Current Carriers in Semiconductors with Mixed Conductivity—V. A. Lashkarev. (*Compt. Rend. Acad. Sci.* (URSS), vol. 73, pp. 929–932; August 11, 1950. In Russian.) It is shown that "bi-polar" diffusion is due to a thermodynamically unbalanced state. The conditions are derived under which such a state is established, all excitation except thermal being excluded.

538.221 **2441**

Interaction between the *d*-Shells in the Transition Metals: Part 2—Ferromagnetic Compounds of Manganese with Perovskite Structure—C. Zener. (*Phys. Rev.*, vol. 82, pp. 403–405; May 1, 1951.) A discussion of the correlation between conductivity and ferromagnetism found by van Santen and Jonker (656 of April).

538.221 **2442**

Ferromagnetism in the Manganese-Indium System—W. V. Goeddel and D. M. Yost. (*Phys. Rev.*, vol. 82, p. 555; May 15, 1951.) About 25 alloys were prepared, with Mn contents ranging from 3 to 91 per cent by weight, in steps of about 4 per cent. Over half (3 to 50 per cent Mn) were found to show ferromagnetism believed to be due to a single phase (Mn_2In). Alloys containing up to 49 per cent Mn appear to be composed of In+Mn_2In, no eutectic being formed.

538.221 **2443**

An Investigation of the Magnetic Properties of Alloys of Manganese with Nickel and Cobalt—F. Gal'perin. (*Compt. Rend. Acad. Sci.* (URSS), vol. 75, pp. 515–518; December 1, 1950. In Russian.)

538.221 **2444**

An Investigation of the Magnetic Properties of Well Ordered Alloys—F. Gal'perin. (*Compt. Rend. Acad. Sci.* (URSS), vol. 75, pp. 647–650; December 11, 1950. In Russian.)

538.221:621.395.625.3 **2445**

Mixed Ferrites for Recording Heads—R. Herr. (*Electronics*, vol. 24, pp. 124–125; April, 1951.) Short discussion of the advantages of using ferrite materials instead of laminations in magnetic recording heads.

538.249 **2446**

Variation with Frequency of the Magnetic After-Effect in Powder Cores—R. Feldtkeller and H. Hettich. (*Z. angew. Phys.*, vol. 2, pp. 494–499; December, 1950.)

546.431.22 **2447**

Jumps in the Conductivity of Barium Titanate—N. A. Tolstoi. (*Zh. Tekh. Fiz.*, vol. 20, pp. 970–974; August, 1950.)

546.431.82 **2448**

X-Ray Investigations of the Ferroelectricity of Barium Titanate—W. Känzig. (*Helv. Phys. Acta*, vol. 24, pp. 175–216; April 10, 1951. In German.)

546.431.82 **2449**

The Nature of Electromechanical Oscillations in $BaTiO_3$ Ceramics—N. A. Roi. (*Compt. Rend. Acad. Sci.* (URSS), vol. 73, pp. 937–940; August 11, 1950. In Russian.) Discussion, with experimental curves, is presented for the following cases: (1) alternating field applied to nonpolarized sample (quasi-electrostriction); (2) weak alternating field applied to weakly polarized sample (linearized quasi-electrostriction); (3) weak alternating field applied to strongly polarized sample (linear piezoelectric effect).

546.431.82:548.55 **2450**

Elastic and Electromechanical Coupling Coefficients of Single-Crystal Barium Titanate—W. L. Bond, W. P. Mason, and H. J. McSkimin. (*Phys. Rev.*, vol. 82, pp. 442–443; May 1, 1951.) A report of measurements made on large multidomain single crystals.

548.0:537 **2451**

Ferroelectricity—B. T. Matthias. (*Science*, vol. 113, pp. 591–596; May 25, 1951.) A general discussion of known ferroelectric materials, and an examination of explanatory theories that have been advanced.

549.514.51 **2452**

Zero-Temperature-Coefficient Quartz Crystals for Very High Temperatures—W. P. Mason. (*Bell Sys. Tech. Jour.*, vol. 30, pp. 366–380; April, 1951.) Crystals with zero temperature coefficient of frequency were obtained by making measurements of a series of rotated *Y*-cuts in the thickness shear mode, and a series of rotated *X*-cuts in the longitudinal length mode, and hence determining the orientation for AT-, BT-, CT- and DT-type crystals with low temperature coefficients, passing through zero value at a prescribed temperature. Calculations are given for crystals operating at 200°C. An AT-type crystal was investigated experimentally, and the calculated results agreed reasonably well with measured values. The maximum temperature at which the temperature coefficient of an AT-type crystal can have zero value is 190°C.

620.193.21:679.5 **2453**

Outdoor Weather Aging of Plastics under Various Climatological Conditions—S. E. Yustein, R. R. Winans, and H. J. Stark, (*ASTM Bull.*, no. 173, pp. 31–43. Discussion, p. 43; April, 1951.) A report on electrical and mechanical tests carried out on five types of clear transparent sheet plastics, six types of laminated material, and five types of moulded terminal bars, after prolonged exposure to tropical, dry desert, temperate, subarctic and arctic conditions.

621.3.015.5:621.315.61 **2454**

Electrical Breakdown over Insulators in High Vacuum—P. H. Gleichauf. (*Jour. Appl. Phys.*, vol. 22, pp. 535–541; May, 1951.) Experimental investigations in the pressure range $5\times10^{-3}-10^{-7}$ mm Hg are described.

621.315.61:539.23 **2455**

The Electric Tunnel Effect across Thin Insulator Films in Contacts—R. Holm. (*Jour. Appl. Phys.*, vol. 22, pp. 569–574; May, 1951.) Previous calculations apply to either very weak or very strong electric fields. The important practical case of intermediate-strength fields is here considered. The image force is neglected, but it is shown how this can be allowed for, approximately. Calculated values of tunnel resistivity are plotted against applied voltage for metallic and for semiconducting contact members. Some inadequacies of the theory are discussed.

621.315.61.011.5 **2456**

"Heat Developed" and "Powder" Lichtenberg Figures and the Ionization of Dielectric Surfaces Produced by Electrical Impulses.—A. M. Thomas. (*Brit. Jour. Appl. Phys.*, vol. 2, pp. 98–109; April, 1951.) Some experiments on both types of figures are reported, and their characteristics outlined. An explanation is suggested of the mode of formation of "heat developed" figures which are associated with the state of the surface of certain kinds of solid dielectrics. "Powder" figures are used to investigate the effect of repeated impulses of alternating polarity, and they show that the effect of a discharge of given polarity is not cancelled by a succeeding discharge of opposite polarity. The phenomena are discussed in relation to theories of surface breakdown and spark discharge.

621.315.612 2457

Contribution to the Study of Physico-Chemical Phenomena in the Ceramics Industry—R. Lecuir. (*Ann. Radioélect.*, vol. 6, pp. 20–50; January, 1951.) A discussion of the forming and sintering of oxides not possessing the plasticity characteristic of clays, which therefore require organic additions to the mix varying according to the forming technique used. The effect of the state of aggregation of the initial powder and of the application of high pressures on the compactness of the product, are examined together with other factors affecting the amount of shrinkage on sintering and the sintering temperature required.

621.315.612.011.5 2458

Dielectric Losses in Ceramic Dielectrics and in Barium Titanate at High Frequencies—A. L. Khodakov. (*Zh. Tekh. Fiz.*, vol. 20, pp. 529–533; May, 1950.) Measurements were made of tan δ at frequencies from 10 to 200 mc and at temperatures from 15° to 180°C. In the case of $BaTiO_3$, tan δ decreases with temperature but remains practically constant within the frequency range specified.

621.318.2 2459

The Determination of the Optimum Parameters of Magnetic Systems with Permanent Magnets—A. Ya. Sochnev. (*Compt. Rend. Acad. Sci.* (URSS), vol. 76, pp. 65–68; January 1, 1951. In Russian.)

621.775 2460

Special Nickel-Iron Alloys Prepared by Powder Metallurgy—Thien-Chi N'Guyen and B. Michel. (*Ann. Radioelect.*, vol. 6, pp. 3–19; January, 1951.) The first of a series of articles on the manufacture of magnetic materials, particularly Ni-Fe alloys. A brief general discussion of ferromagnetism is presented. The superiority of powder-metallurgy techniques for producing these alloys is shown by a study of the surface texture and crystal structure of a sintered 50 per cent Ni alloy.

661.1.037.5 2461

The Physical Aspect of Glass-Metal Sealability in the Electronic Tube Industry—G. Trébuchon and J. Kieffer. (*Glass Ind.*, vol. 32, pp. 165–174, 202, 240–247, 255 and 290–295; April to June, 1951.) English translations of paper noted in 2253 of 1950, 135 of February, and 929 of May.

666.2:549.623.5 2462

Glass/Mica Vacuum-tight Seals—J. Labeyrie and P. Léger. (*Le Vide*, vol. 6, pp. 936–940; January, 1951.) See 1437 of 1950 (Labeyrie).

537.311.33 2463

Electrons and Holes in Semi-Conductors [Book Review]—W. Shockley. Publishers: D. Van Nostrand Co., New York, N. Y., 1950, 543 pp., $9.75. (Proc. I.R.E., vol. 39, pp. 449–450; April, 1951.) "This excellent book might almost be said to consist of a set of three monographs of increasingly rigorous treatment. Part I, 'Introduction to Transistor Electronics', is entirely descriptive . . . Part II . . . is 'Descriptive Theory of Semiconductors'. . . . In Part III, 'Quantum Mechanical Foundations', many of the concepts presented in earlier portions of the book are subjected to rigorous examination."

549.514.51:621.396.611.21 2464

Quartz Vibrators and their Applications [Book Review]—P. Vigoureux and C. F. Booth. Publishers: H. M. Stationery Office, London, Eng., 30/-. (*Jour. Brit. IRE*, vol. 11, p. viii; April, 1951.) The properties of quartz, manufacture of crystals, and applications in telecommunications, etc., are considered.

MATHEMATICS

512.831 2465

The Principle of Minimized Iterations in the Solution of the Matrix Eigenvalue Problem—W. E. Arnoldi. (*Quart. Appl. Math.*, vol. 9, pp. 17–29; April, 1951.)

517.534:538.566 2466

On the Method of Saddle Points—B. L. van der Waerden. (*Appl. Sci. Res.*, vol. B2, no. 1, pp. 33–45; 1951.) In attempting to solve the problem of radio propagation over a plane earth, earlier investigators encountered a difficulty in evaluating the integral $\int_c we^{-\lambda u}dv$ by the method of steepest descents, because of the proximity of a pole to a saddle point. The problem is here transformed to one of integration in the u-plane. A solution is obtained in which the part of the integral corresponding to the pole can be readily separated out.

681.142 2467

Mechanized Reasoning. Logical Computors and their Design—D. M. McCallum and J. B. Smith. (*Electronic Eng.*, vol. 23, pp. 126–133; April, 1951.)

681.142 2468

Visual Presentation of Binary Numbers—E. H. Lenaerts. (*Electronic Eng.*, vol. 23, pp. 140–141; April, 1951.) By using a raster timebase, the pulses representing a number can be displayed as vertical deflections of the trace, or by a pattern of bright dots on a background of fainter dots representing the zeros. In a better method described, the pulse train is superimposed upon the frame-deflection timebase, while the brightness of the spot is modulated by clock pulses timed to occur in all positions where a spot is possible. The pulses thus appear as short vertical lines with dots interposed which represent the zeros.

681.142:517.9 2469

Solution of a System of Linear Equations with a Slightly Unsymmetrical Matrix by Using a Network Analyzer—H. L. Knudsen. (*Trans. Dan. Acad. Tech. Sci.*, no. 2, 16 pp.; 1950. In English.) The method is based on iteration, only a few steps being necessary when the asymmetry is only slight. No equipment other than the network analyzer is required.

681.142:621.316.726 2470

Automatic Frequency Control—Pinkerton. (*See* 2364.)

MEASUREMENTS AND TEST GEAR

531.765:529.786 2471

Comparing Outputs from Precision Time Standards—J. M. Shaull and C. M. Kortman. (*Electronics*, vol. 24, pp. 102–107; April, 1951.) Description of equipment developed at the National Bureau of Standards for monitoring the time-keeping of a group of standard quartz clocks. The chronograph records time differences of two clocks to within 1 ms, using spark-generating equipment whose rate is controlled by one clock, while the drum speed is governed by the other. A motor-driven switch connects each of several clocks in turn to the spark generator every 15 minutes, thus providing intercomparison data. The chronoscope uses a 3 inch cr tube, one clock and frequency divider being applied to produce a circular sweep with small fixed marker dots at 0.1-ms intervals, while a pulse from the circuit of the second clock produces a larger bright spot on the sweep. The chronograph and chronoscope are locked in time phase, so that observation of the position of the bright spot enables the time difference between the two clocks to be determined to within 20 μs. See also *Tech. Bull. Nat. Bur. Stand.*, vol. 35, pp. 14–16; January, 1951.

535.322.4:546.217 2472

A Phase-Shift Refractometer—C. W. Tolbert and A. W. Straiton. (*Rev. Sci. Instr.*, vol. 22, pp. 162–165; March, 1951.) A description of apparatus for measuring small changes in the dielectric constant or refractive index of air, by determining the phase change of a 9.375-kmc wave over a 3 foot path. The test path may be confined to a waveguide through which air is drawn, or it may be the space between two antenna systems. The waveguide system can be used to measure rapid changes in the refractive index with an error <1 part in 10^6; the antenna method has larger errors because of flexibility of supports and external reflections.

621.317.083.4:621.396.645 2473

Sensitive Null Detector—Scroggie. (*See* 2390.)

621.317.335.3 †+621.317.374].029.64 2474

Measurement of Dielectric Constant and Losses of Solid Dielectrics by means of Waveguides—G. D. Burdun. (*Zh. Tekh. Fiz.*, vol. 20, pp. 813–821; July, 1950.) Waves of the H_{01} mode are excited in a rectangular waveguide, the end of which is closed by a piece of the dielectric under investigation. The distribution of the dielectric field intensity inside the waveguide is measured by means of a probe and indicator, and from these measurements the properties of the dielectric are determined. The theory of the method is discussed, and the results of measurements with various dielectrics on wavelengths between 1.6 and 3.2 cm are presented. The accuracy of the method is to within about 1 or 2 per cent.

621.317.35:621.397.5 2475

Notes on TV Waveform Monitor Frequency Response—W. L. Hurford. (Proc. I.R.E., vol. 39, pp. 562–563; May, 1951.) A comparison of monitors having (a) very wide response band, (b) the response specified in the IRE Standards on Television (see 2035 of 1950), and (c) a sharp cutoff at twice the bandwidth of the IRE curve. From the response of these devices to clean, sharp pulses, and to pulses with spikes, it is concluded that "the IRE response is an excellent choice."

621.317.39:[621.318.4+621.319.4 +621.396.611.1 2476

Devices for the Measurement of the Temperature Coefficients of Coils, Capacitors and Oscillatory Circuits—C. Schreck. (*Fernmeldetech. Z.*, vol. 4, pp. 30–36; January, 1951.) Review of the development of methods and apparatus necessitated by the continual demand for higher frequency constancy. "Static" temperature coefficients (i.e., those related to external heating effects) have received more attention than "dynamic" coefficients (related to internal heating effects).

621.317.4:621.317.755 2477

The Electron-Beam Ferroscope—P. E. Klein. (*Arch. tech. Messen.*, no. 181, pp. T34–T-24; February, 1951.) Description and circuit details of a cro unit for displaying the magnetization curve of high-permeability iron-alloy samples. The voltage drop across a resistor in the test-circuit primary is applied to the X-plate amplifier; the output from the secondary is fed to the Y-plate amplifier, either direct or through an integrating circuit. By means of a 3-position switch the waveform of the primary current, secondary voltage, or B-H characteristic of the magnetic circuit, can be displayed. Typical traces are shown.

621.317.6:621.396.645 2478

The Determination of Amplifier Sensitivity with the Aid of the Noise Diode—W. K. Squires. (*Sylvania Technologist*, vol. 4, pp. 35–

37; April, 1951.) A measure of amplifier performance designated "sensitivity factor" is introduced. It is expressed quantitatively as the ratio of standard noise output to actual noise output at maximum gain, and its use enables the gain and noise factor to be correlated. A method of measuring the sensitivity factor is described.

621.317.7+621.396.611].029.63 2479
Circuits and Measurement Apparatus for the 30-cm Band—E. Safa. (*Onde élect.*, vol. 31, pp. 33–43; January, 1951.) Illustrated review outlining the characteristics of tubes and associated coupling circuits, and giving details of generators, wavemeter, wattmeter, and curve tracer, designed for the uhf band.

621.317.715:621.396.611.33/.34 2480
Coupling of A.C. Galvanometer to A.C. Amplifier—C. T. J. Alkemade and P. M. Endt. (*Appl. Sci. Res.*, vol. B2, no. 1, pp. 46–52; 1951.) A galvanometer with a 50-cps magnetic field is considered. Untuned transformer coupling provides higher gain than capacitor coupling, but magnetic relaxation phenomena in the transformer cause slow changes in sensitivity, which make capacitor coupling preferable.

621.317.725 2481
An Instantaneous Peak Voltmeter—M. W. Tobin, H. Grundfest, and R. L. Schoenfeld. (*Rev. Sci. Instr.*, vol. 22, pp. 189–190; March, 1951.) The operation of the diode capacitor peak-voltmeter circuit is discussed, and a circuit is described which provides a measurement of pulse amplitude unaffected by the amplitude of the previous pulse. This is used for investigating pulses of duration 0.1–20 ms at frequencies as low as 0.2 cps.

621.317.725 2482
Audio-Frequency Valve Voltmeter—S. Kelly. (*Wireless World*, vol. 57, pp. 215–218; June, 1951.) Details are given of a self-calibrating, portable instrument designed for a voltage range of 1 mv to 10 v in four decades, with an input impedance >10 MΩ across 10 pF and an output impedance <500Ω.

621.317.726:621.3.011.6 2483
Calculation of CR Elements for the case of Varying Voltage and/or Nonlinear Resistances—Elger. (*See* 2356.)

621.317.76:621.396.615 2484
An Instrument for Recording the Frequency Drift of an Oscillator—W. W. Boelens. (*Philips Tech. Rev.*, vol. 12, pp. 193–199; January, 1951.) The meter was designed for measuring the frequency variation of the local oscillator of FM receivers in the 88 to 108-mc band. The reference frequencies, of which there are ten in the band 80.4 to 118.5 mc, are obtained from a 4.232-mc crystal oscillator by a multiplication and mixing process. The frequency drift is indicated by the variation of a direct current, which can be read on a dc meter or applied to a recording instrument.

621.317.79:621.3.018.78†:621.396.61 2485
Measurement of Distortion in Broadcast Transmitters—Müller. (*See* 2566.)

621.317.79:621.396.67 2486
A Phase Front Plotter for Testing Microwave Aerials—C. A. Cochrane. (*Elliott Jour.*, vol. 1, pp. 29–30; March, 1951.) A search antenna, servo-controlled via an rf phase discriminator, is used to find lines of constant phase, accurate to within about $\pi/16$ near the antenna. The search antenna is an open-circuited circular waveguide, suitable for wavelengths near 3.2 cm.

621.317.79:621.396.822 2487
High-Sensitivity High-Frequency Noise-Measurement Apparatus Calibrated Absolutely in kT$_0$ Units—H. Röschlau. (*Arch. elekt. Übertragung*, vol. 4, pp. 427–434; October, 1950.) Requirements for a receiver to have high sensitivity, appropriate for the investigation of cosmic noise sources on a wavelength of 1.5 m, are examined. The choice of input tube and input circuit are discussed in detail, a cavity-resonator tank circuit being used on account of its high resonance resistance, together with a pentode with regeneratively-coupled screen grid. Three alternative methods of performing the absolute calibration are described; a method using a noise diode Type SA102 was most exact and gave a value better than 0.05 kT$_0$, for the sensitivity. Details are given of the method used for coupling the receiver to the antenna array.

621.317.799†:621.385.012 2488
Tube Characteristic Tracer using Pulse Techniques—H. M. Wagner. (*Electronics*, vol. 24, pp. 110–114; April, 1951.) A full description of an instrument designed primarily to obtain characteristic curves in the positive-grid region for small tubes used at high pulse power levels. The curves are displayed on a cro screen and can be recorded photographically. See also 2576 of 1950 (Leferson) and 692 of April (Graffunder and Schultes).

621.396.615.015.7.001.4† 2489
Radar Test Generator—K. S. Stull. (*Electronics*, vol. 24, pp. 93–95; April, 1951.) Circuit details and description of equipment providing triggered or free-running pulses of duration 0.25, 0.5, or 1.0 μs, and also cw signals in the range 47–76 mc, for testing wide-band circuits. The output voltage is variable from 0.1 μv to 0.1 v.

537.7 2490
Electrical Measurements and the Calculation of the Errors Involved: Part 1 [Book Review]—D. Karo. Publishers: Macdonald & Co., London, Eng., 1950, 191 pp., 18s. (*Nature*, (London), vol. 167, p. 745; May 12, 1951.) ". . . an extremely useful book, particularly for finals students, research workers, and engineers."

621.317.755 2491
Encyclopedia on Cathode-Ray Oscilloscopes and their Uses [Book Review]—J. F. Rider and S. D. Uslan. Publishers: J. F. Rider, New York, N. Y., 1950, 992 pp., $9.00. (*Electronics*, vol. 24, pp. 282, 284; April, 1951.) "Although the authors explain in the foreword to this book that some readers having special interest may find that it has limited coverage, the book quite adequately backs up its title for the average reader. . . . A novel aid . . . is a collection of synthesized waveform patterns, a total of 1580 extending over 79 pages. These are provided for those readers who do not have an harmonic wave analyzer available."

621.396.615.17 2492
Time Bases (Scanning Generators) [Book Review]—O. S. Puckle. Publishers: Chapman and Hall, London, Eng., 2nd edn, 387 pp., 30s. (*Electrician*, vol. 146, p. 1131; April 6, 1951.) This edition includes a new chapter on Miller capacitance timebases, as well as many other modifications and additions to the original 1943 edition.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

620.179.16 2493
Nondestructive Testing of Large Forgings by an Ultrasonic Method—W. Felix. (*Schweiz. Arch. angew. wiss. Tech.*, vol. 17, pp. 107–113; April, 1951.) An account of experience gained over a period of several years in the operation of actual tests using pulse-type equipment.

621.316.7.076.7+621-526 2494
Control Systems and their Application to Industry—W. R. Blunden. (*Jour. Inst. Eng.* (Australia), vol. 23, pp. 89–94; April/May, 1951.) A general survey, in which the different types of control system are classified both according to operation and application, and some of the more important types are described briefly.

621.317.083.7:551.508.1 2495
Cosmic-Ray Radiosonde and Telemetering System—M. A. Pomerantz. (*Electronics*, vol. 24, pp. 88–92; April, 1951.) Details of equipment comprising four GM counters which trigger a multivibrator controlling the keying of an uhf transmitter. Altitude and temperature are indicated by modulation intervals and modulation frequency.

621.365.5 2496
High-Frequency Generators and their Applications—W. Burkhardtsmaier. (*Telefunken Ztg.*, vol. 23, pp. 73–82; September, 1950.) Commercially available generators, with outputs ranging from 1.5 to 70 kw, are described and illustrated. Industrial heating applications are considered. A frequency of 400 kc is used for inductive heating, and 20 mc for dielectric heating.

621.365.54†:621.793 2497
Metal Evaporator uses High-Frequency Heating—R. G. Picard and J. E. Joy. (*Electronics*, vol. 24, pp. 126–128; April, 1951.) Difficulties due to nonwetting of filament, and its reaction with the material to be evaporated, are avoided by the use of a water-cooled hf heating coil. The method enables metals, that would react with the usual tungsten or tantalum filament, to be evaporated.

621.383:551.576 2498
Electronic Theory for Design of a Radar System using Light Waves (in particular for a Cloud Height Indicator)—A. Baude. (*Onde élect.*, vol. 31, pp. 44–48 and 90–101; January and February, 1951.) The principle and theory of distance measurement by pulsed light waves is discussed. An illustrated description is given of two sets of apparatus developed for measurement of cloud height and estimation of layer thickness. At 1,500m, error may be about 10m. Clear echoes from above 10km have been obtained despite intervening cloud layers.

621.384.611.2† 2499
The 300 MeV Synchrotron at the Massachusetts Institute of Technology—(*Engineer* (London), vol. 191, pp. 440–442; April 6, 1951.) A general account.

621.385.833 2500
Theory of the Three-Electrode Electrostatic Lens—É. Regenstreif. (*Ann. Radioélect.*, vol. 6, pp. 51–83 and 114–155; January and April, 1951.) An expanded account of work noted in 1213, 1743 and 2314 of 1950, 1455 of July, and 1742 of August.

621.385.833 2501
The Visualization of Atomic Distances by means of the Electron Microscope—L. Wegmann. (*Helv. Phys. Acta*, vol. 24, pp. 63–71; February 15, 1951. In German.) By stopping out certain concentric lens zones, it is theoretically possible to effect an improvement of the resolving power of uncorrected electron lenses sufficient to render visible the atomic lattices of crystals. The practical difficulties are explained briefly.

621.385.833 2502
Scattering Phenomena in Electron Microscope Image Formation—C. E. Hall. (*Jour. Appl. Phys.*, vol. 22, pp. 655–662; May, 1951.)

621.385.833 2503
Electronoptical Theory of the Deflection of an Extended Electronoptical Image by means of Crossed Electrical Deflection Systems—J. Himpan. (*Ann. Phys.*, (Lpz.), vol. 8, pp. 405–422; February 15, 1951.) The discussion presented is valid for mutually perpendicular deflection systems with plates of any type and arrangement, and it takes third-order aberrations into account. The conditions under

which deflection in two dimensions can be performed without introducing distortion are established. For deflections greater than those permitted in the ideal case, six different types of defect are recognized, viz., three types of over-all image distortion, and three of image-point deformation. Simple formulas are derived expressing these defects. Their magnitudes, in particular cases, are calculated by use of constants, which can be determined from a few measurements.

621.385.833 **2504**

Calculation of the Optical Constants of Powerful Magnetic Electron Lenses—W. Glaser. (*Ann. Phys.*, (Lpz.), vol. 8, p. 423; February 15, 1951.) Corrections to paper noted in 703 of April.

621.385.833:537.533.72 **2505**

The Significance of the Concepts 'Focus' and 'Focal Length' in Electron Optics and Strong Electron Lenses with Newtonian Image-Formation Equation—W. Glaser and O. Bergmann. (*Z. angew. Math. Phys.*, vol. 1, pp. 363–379; November 15, 1950.) The functions determining the relation between the position of the object and that of the image, for linear magnification, are more complex in the case of electron optics than in the case of light. These functions are approached, in the neighborhood of two conjugate points, by the Newtonian osculating equation for the image function, including terms up to the fourth order. Each pair of conjugate points thus possesses corresponding foci, principal points, and focal lengths resulting from the Newtonian equation. If the osculating cardinal elements are independent of the pair of conjugate points chosen, they characterize by themselves the formation of the image, and are identical to the magnitudes defined in the usual way for light. Such fields, for which the image-formation equation of ordinary optics is strictly valid, are termed "fields with Newtonian representation."

A study is made of these strong fields, and examples of them are given which approximate the fields actually existing in electron lenses. In order to keep a physical significance for focal length, whatever the magnification, as close an approximation as possible to the empirical field must be obtained by one of the Newtonian type. A method for doing this is indicated, and experimental methods of determing the focal points and focal lengths of such approximate fields are examined.

621.385.833:621.311.1 **2506**

Modern Low-Power High-Voltage Generators—J. Vastel. (*Ann. Radioélect.*, vol. 6, pp. 84–94; January, 1951.) Various forms of generator suitable for supplying voltages in the range 30–80 kv, constant to within about 1 part in 100,000, are discussed, and the causes of ripple in the output are analyzed. Particular equipments described for supplying electron microscopes, etc., use low- and high-frequency oscillators (600 cps–46 kc), in association with voltage multipliers, giving output currents of about 100 μa and voltages of 60–80 kv.

621.386 **2507**

The Intensification of X-Ray Fluorescent Images—W. S. Lusby. (*Elec. Eng.*, vol. 70, pp. 292–296; April, 1951.) Paper given at AIEE Winter General Meeting, New York, January, 1951. The intensifier has a Cs-Sb photocathode arranged close to the ZnS input screen, an accelerating voltage of 30 kv causing the emitted photoelectrons to impinge on an Al-backed Zn-CdS output screen of reduced size at the far end of the 17-inch tube. Requirements for medical and industrial applications are discussed. Experimental installations, giving a brightness amplification of slightly over 100 times, have been put into operation.

621.386.1:621.385 **2508**

Radiographic Examination of Electronic Valves—H. B. van Wijlen. (*Philips Tech. Rev.*, vol. 12, pp. 207–209; January, 1951.) The examination of the electrode structure of tubes by means of X rays is described. A resolution of 6 μ is obtained with an image of approximately the same size as the object.

621.387.4† **2509**

A Secondary-Electron Photon Counter—S. F. Rodionov and A. L. Osherovich. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 461–463; September 21, 1950. In Russian.) A description is given of a photomultiplier device for counting "visible" photons (λ=3600–6500 Å). A Sb-Cs photocathode developed by Kubetski is used, and light fluxes of the order of $10^{-14}-10^{-15}$ lumens can be measured.

621.387.4† **2510**

The Discharge Mechanism for Oversize Pulses in Counters with Vapour Filling—H. Neuert. (*Ann. Phys.* (Lpz.), vol. 8, pp. 341–349; February 15, 1951.)

621.387.462† **2511**

Silver Bromide Crystal Counters—K. A. Yamakawa. (*Phys. Rev.*, vol. 82, pp. 522–526; May 15, 1951.)

621.387.464† **2512**

The Scintillation Counter—W. Hanle. (*Naturwiss.*, vol. 38, pp. 176–185; April, 1951.) A survey of the development and applications of photomultiplier-type counters, with a list of 156 references.

621.396.645:535.247.4 **2513**

A Balance Indicator with High Input Impedance using a Cathode Follower—D. T. R. Dighton. (*Jour. Sci. Instr.*, vol. 28, pp. 101–102; April, 1951.) "The use of the cathode follower circuit with high value grid resistances is discussed, and it is shown that a high-slope pentode can give an impedance conversion of 10^7. A cathode follower circuit, suitable for a null-balance indicator for photometric work, is described. The input impedance is 500 MΩ, the grid current about 10^{-10}a, and the detection limit 1 to 2 mv change of grid potential. A simple method of compensating for slow variations of heater voltage is employed."

621.38 **2514**

Survey of Modern Electronics [Book Review]—P. G. Andres. Publishers: J. Wiley and Sons, New York, N. Y., 1950, 522 pp., $5.75. (*Jour. Appl. Phys.*, vol. 22, pp. 685–686; May, 1951.) "Written as a text for a short survey course for students in electrical engineering . . . ."

621.365.54† **2515**

Induction Heating [Book Review]—N. R. Stansel. Publishers: McGraw-Hill Publishing Co., London, Eng., 1949, 212 pp., 34s. (*Nature* (London), vol. 167, p. 700; May 5, 1951.) Of the nature of a handbook giving formulas and data relating to the electrical and thermal quantities involved.

PROPAGATION OF WAVES

538.566 **2516**

Is there a Zenneck Wave in the Field of a Radiator?—H. Ott. (*Arch. elekt. Übertragung*, vol. 5, pp. 15–24; January, 1951.) Application of the modified saddle-point integration method (1024 of 1946) confirms the existence of a surface wave, but with coefficient half that of Sommerfeld's residuum, and only in the region of the boundary (earth) surface and for large values of refractive index. It is a component of a more general "surface effect" of fundamental significance. The validity of other theoretical solutions is discussed.

538.566 **2517**

The Nonexistence of the Surface Wave in the Radiation from a Dipole over a Plane Earth—T. Kahan and G. Eckart. (*Arch. elekt. Übertragung*, vol. 5, pp. 25–32; January, 1951.) Discussion of Sommerfeld's treatment of the problem, drawing attention to the mathematical error involved (2892 of 1949). Objections to Ott's theory (2516 above) are pointed out.

538.566 **2518**

The Radiation Principle—A. G. Sveshnikov. (*Compt. Rend. Acad. Sci.* (URSS), vol. 73, pp. 917–920; August 11, 1950. In Russian.) The radiation principle (equation 2), introduced by Sommerfeld to ensure the unique solution of the wave equation (1), varies in accordance with the region to which the latter is applied. Accordingly, either the principle of limiting absorption proposed by Ignatovski in 1905 or the principle of limiting amplitude proposed by Tikhonov and Samarski (*Zh. Eksp. Teor. Fiz.*, vol. 18, no. 2, p. 243; 1948) should be applied.

538.566:517.534 **2519**

On the Method of Saddle Points—van der Waerden. (*See* 2466.)

621.396.11 **2520**

Comparison of Ionospheric Radio Transmission Forecasts with Practical Results—A. F. Wilkins and C. M. Minnis. (*Proc. IEE*, vol. 98, pp. 209–220; May, 1951.) "The production of muf forecasts for oblique transmission involves numerous operations on basic information obtained at vertical incidence. At each stage, errors are introduced whose cumulative effect determines the difference between predicted and observed circuit performance. The sources of the errors are examined and tentative values assigned to them with special reference to F_2 region. The computed value of the total error is compared with results obtained on commercial and Service circuits, and with observations made by other means. It is concluded that, although on the average, agreement is good, discrepancies remain which need further examination after the elimination of known sources of error. In a few cases, comparisons of predicted and actual times of fades due to ionospheric absorption have been made. Although the agreement between these times is reasonably good, it is believed that predictions of the actual field strength may be in error by large amounts."

621.396.11 **2521**

Evaluation of Ionosphere Observations—W. Becker. (*Arch. elekt. Übertragung*, vol. 4, pp. 391–400; October, 1950.) Development of an earlier theoretical paper (2844 of 1944). The magnitudes of inaccuracies due to using ray theory instead of rigorous wave theory are investigated, neglecting the earth's curvature and magnetic field, and assuming a vertical distribution of ionization decreasing gradually to zero both above and below a layer of maximum density. Application of the calculated results to the evaluation of fixed-frequency and swept-frequency records shows that the ray theory is reliable down to low values of relative thickness of ionosphere layers for all frequencies except those within a small range around the critical frequency and, for very oblique incidence, those below 0.2 times the critical frequency.

621.396.11.029.45 **2522**

The Ionospheric Propagation of Low- and Very-Low-Frequency Radio Waves over Distances less than 1,000 km—R. N. Bracewell, K. G. Budden, J. A. Ratcliffe, T. W. Straker, and K. Weekes. (*Proc. IEE*, vol. 98, pp. 221–236; May, 1951.) Results are summarized of experimental work performed at the Cavendish Laboratory over a period of years. Waves of frequency 16 to 30 kc are reflected as if from a sharp horizontal boundary at a height of 72±3 km (with the sun overhead), and waves of frequency 30 to 150 kc at about 75 km at oblique incidence, and perhaps 10 km higher at vertical incidence. The polarization is approximately circular at steep incidence, and linear on 16 kc at oblique incidence (65°). Absorption increases rapidly with frequency; differences are observed in behavior around sunrise at steep and oblique incidence; sudden ionospheric disturbances are associated with decreases in the apparent height of reflection. Present theories of reflection of very long waves are outlined.

621.396.11.029.64:621.396.621.087.4 2523

A Receiver for Measuring Angle-of-Arrival in a Complex Wave—F. E. Brooks, Jr. (Proc. I.R.E., vol. 39, pp. 407–411; April, 1951.) Descriptions of the design, construction, and calibration of a field-strength and wave-direction recorder developed at the University of Texas for operation on a wavelength of 3.2 cm. Two plane wave fronts can be measured simultaneously. The absolute amplitude of the dominant wave can be determined to within ±1 db over a range of 70 db, the relative amplitude of the weaker to within ±0.5 db. The application of phase-interferometry technique enables the angles of incidence to be measured to within ±0.01°.

621.396.812.029.62/.63 2524

Investigations of the Influence of the Troposphere on the Propagation of Ultra-Short Waves—R. Schachenmeier. (*Arch. elekt. Übertragung*, vol. 5, pp. 1–9; January, 1951.) Description of a method developed in 1941 to enable reliable calculations of field strength to be made from a knowledge of tropospheric conditions. The theory is based on the combined effects of atmospheric refraction of the ray, and the diffraction effect due to the earth's curvature. Values calculated from meteorological observations are in satisfactory agreement with measured field strengths at meter and decimeter wavelengths, for different land/sea paths. The cases of propagation beyond and within the optical range are treated separately.

621.396.812.029.64 2525

Attenuation of Radio Signals caused by Scattering—A. H. LaGrone, W. H. Benson, Jr., and A. W. Straiton. (*Jour. Appl. Phys.*, vol. 22, pp. 672–674; May, 1951.) An equation is developed for determining the total energy scattered during passage of a beam through unit volume of a scattering medium. Curves are plotted from which the attenuation due to scattering can be found.

621.396.812.3:551.510.535 2526

Multiple Reflections and Undulations in the F_2-Region of the Ionosphere—S. S. Banerjee and R. R. Mehrotra. (*Science and Culture* (Calcutta), vol. 16, pp. 72–73; August, 1950.) Anomalous variations of the amplitude of multiple reflections are observed even when transmitter and receiver are close together. In some cases, at frequencies between 6 and 11 mc (low compared with f_cF_2 at the time), the amplitude of any higher-order echo was greater than that of any lower-order echo. The effect may be caused by undulations in the lower structure of the F_2 layer as suggested by Ratcliffe (193 of 1949).

621.396.812.3:551.510.535 2527

Anomalous Behaviour of Multiply Reflected Echoes from the Ionosphere—S. N. Mitra. (*Science and Culture* (Calcutta), vol. 16, pp. 425–426; March, 1951.) An alternative explanation of the phenomenon noted by Banerjee and Mehrotra (2526 above). It is suggested that on normal quiet days the effect may be due to interference bewteen the two magneto-ionic components of the downcoming wave. Experiments have shown that the use of polarized waves removed the anomaly, the first reflection becoming always the strongest.

621.397.8.08 2528

U.H.F. TV Propagation Measurements—Cook and Artman. (*See* 2563.)

RECEPTION

621.396.621 2529

The Development of Commercial Receivers [by Telefunken]—H. Hart, G. Schaffstein, and G. Vogt. (*Telefunken Ztg.*, vol. 23, pp. 83–92; September, 1950.) Descriptions are given of the first four communications receivers developed by Telefunken after the war, for press and government use, viz., the EPK/1, for telegraphy and telephony; the E11-1/48 and EPH/L/2 "Hell" system teleprinter receivers, and the "Ball E1" relay and monitor receiver.

621.396.621 2530

The Development of the Telefunken Broadcast Receiver since 1945—W. F. Ewald. (*Telefunken Ztg.*, vol. 23, pp. 97–105; September, 1950.) Account of the difficulties surmounted in restoring production in the absence of nearly all normal facilities, and descriptions, with constructional and circuit details, of a range of portable and table models marketed up to 1950.

621.396.621.087.4:621.396.11.029.64 2531

A Receiver for Measuring Angle-of-Arrival in a Complex Wave—(*See 2523.*)

621.396.822 2532

Receiver Noise—W. Kleen. (*Fernmeldetech. Z.*, vol. 4, pp. 19–25, 56–63 and 182; January, February, and April, 1951.) Review of present-day theory and technique for determining optimum noise factor. Relations connecting antenna and circuit noise with transmission range are derived, and the physical basis of tube noise is described. Cosmic noise is discussed. A quantitative study is made of input-circuit noise for disk-seal tubes. Noise figures for the klystron, traveling-wave tube, etc., are discussed. For wavelengths below about 2m, the inherent noise of the receiver determines the signal/noise ratio. Over 50 bibliographical references are given.

STATIONS AND COMMUNICATION SYSTEMS

621.39.001.11 2533

Symposium on Information Theory [London, 1950]—F. A. Fischer. (*Fernmeldetech. Z.*, vol. 4, pp. 79–85; February, 1951.) German report and comment. See also 984 of May (Jackson).

621.39.001.11 2534

Geometrical Interpretation in Hilbert Space of the Properties of Periodic or Pulsed Systems—R. Vallée. (*Ann. Télécommun.*, vol. 6, pp. 61–66; March, 1951.) The simpler properties of Hilbert space are summarized; a method is given for representing voltage and current as vectors in this space. Ohm's law is interpreted as a matrix operation: $\vec{V}=(Z)\vec{I}$, and Joule's law as a scalar product: $W=\vec{V}.\vec{I}$. Power factor is taken as the real part of $\vec{V}.\vec{I}/|\vec{V}|.|\vec{I}|$. A geometrical interpretation of the quantity of information contained in a periodic signal is deduced with a generalization of the above concepts for the case of a closed linear network. Similar relations are deduced for pulsed systems, the analogy resting on a time-frequency relation.

621.39.001.11:[621.317.083.7+621.398 2535

Application of Shannon's Communication Theory to Telecontrol, Telemetry, or Measurement—J. Loeb. (*Ann. Télécommun.*, vol. 6, pp. 67–76; March, 1951.) That part of Shannon's theory relating to sources capable of producing a finite number of different symbols is summarized; the properties of transmission channels in the absence and in the presence of noise, and the necessity for matching source to channel by suitable coding, are discussed. Loss of information due to noise is evaluated in the light of the concept of conditional entropy. The theory of a transmission channel in the presence of noise is treated as a generalization of which the Baudot-Verdan system of radiotelegraphic communication is a particular anticipatory application.

621.39.001.11:621.396.619.14 2536

Correlation Functions and Spectra of Phase- and Delay-Modulated Signals—L. A. Zadeh. (Proc. I.R.E., vol. 39, pp. 425–428; April, 1951.) "A delay-modulated signal may be regarded as the response of a delay modulator to the carrier of the signal. By using this point of view, a general expression for the correlation function of a delay-modulated signal is obtained. This expression is given in an operational form in which the operand is the correlation function of the carrier, and the operator is the correlation function of the delay modulator. The general result is applied to the determination of the correlation function of a delay-modulated signal having a periodic carrier, and, more particularly, to the determination of the correlation function of a phase-modulated signal."

621.395.44 2537

A Twelve-Channel Carrier Telephone System for Use on Open Wire Lines—T. B. D. Terroni. (*Strowger Jour.*, vol. 7, pp. 180–193; April, 1951.) A development of the cable system noted in 2891 of 1950 is described.

621.396.5 2538

U.S.W. Radiotelephony Technique—W. Runge. (*Telefunken Ztg*, vol. 23, pp. 67–72; September, 1950.) Meter-wave apparatus developed by Telefunken since the end of the war includes: (a) an FM mobile telephone installation especially for the police; (b) an easily portable FM transmitter for reporting and emergency purposes; (c) a fixed transmitter-receiver link of very high quality for connecting studio to broadcast transmitter. Descriptions are given of (a) and (b), (c) being reserved for a later paper.

621.396.5:621.395.722 2539

The New London Radio-Telephony Terminal—C. W. Sowton and D. B. Balchin. (*P.O. Elec. Eng. Jour.*, vol. 44, pp. 25–30; April, 1951.) The technical problems encountered in the construction of the new 48/80-circuit international radio telephone terminal in London are discussed, and the special control and supervision equipment installed is described with an outline of circuit operation. The use of automatic control circuits and standardized equipment results in considerable economies in staff and apparatus.

621.396.61/.62:623.6 2540

Progress in Military (Land Forces) Radiocommunications—Morand. (*Onde élect.*, vol. 31, pp. 3–17; January, 1951.) Description of equipment of American pattern selected for production in France from 1945 onward. Concise details and illustrations are given of (a) three short-range telephony intercommunication sets, two using FM, the third PHM for use in vehicles; (b) two medium-range portable W/T-R/T sets using AM; (c) long-range equipment comprising a mobile station with trailer power unit; power is 250 w for R/T, 400 w for W/T, frequency range 2 to 18 mc; a 4-channel multiplex mobile station with transmitter power 50 w, designed to replace land-line systems. The trend of technical development in miniaturization, tropicalization, etc., is outlined.

621.396.65 2541

Short-Range Communication by V.H.F. Radio—(*GEC Telecommun.*, vol. 1, no. 2, pp. 61–79; 1946.) A summary of the factors governing the choice of systems for point-to-point and mobile services. Advantages of the vhf band mentioned are the reduced noise level, the convenient size of antenna systems, and the restriction of transmissions to the service area. Simplex, duplex, and relay systems using both AM and FM, are discussed, and a complete range of equipment for both fixed and mobile stations is described.

621.396.712 2542

150-kW Medium-Wave Broadcast Transmitter at Daventry—(*Engineering* (London), vol. 171, pp. 506–507; April 27, 1951.) A description of the new British Third Program transmitter and antenna system. The transmitter, which uses air-cooled tubes, consists of two identical 100-kw units which can be paralleled. The fading-free area is increased by connecting the coaxial feeder across an insulator at a point 460 feet (about two-thirds of the height) up the mast radiator.

621.396.933 2543
A General Survey of Electronics in Air Transport—Jackson. (*See* 2434.)

621.396.933 2544
The M.C.A. [Ministry of Civil Aviation] V.H.F. Area Coverage Network: Audio Frequency Distribution—J. L. French. (*Electronic Eng.* (London), vol. 23, pp. 146–148; April, 1951.) General description, with block diagram, of the af equipment and its operation. See also 2545 below.

621.396.933 2545
The M.C.A. [Ministry of Civil Aviation] V.H.F. Area Coverage Network: Provision of Transmitting Station Equipment—D. H. C. Scholes. (*Electronic Eng.* (London), vol. 23, pp. 148–150; April, 1951.) General description of the modified Type-T.1131 transmitter and its temperature-controlled crystal unit. See also 2544 above.

SUBSIDIARY APPARATUS

621-526+621.316.7.076.7 2546
Control Systems and their Applications to Industry—Blunden. (*See* 2494.)

621.526 2547
Servomechanisms with Linearly Varying Elements—M. J. Kirby. (*Elec. Eng.*, vol. 70, p. 343; April, 1951.) Digest of paper presented at the AIEE Fall General Meeting, Oklahoma, 1950. An analytical method is presented for determining the stability of a servomechanism in which one or more elements vary linearly with time.

621.316.722 2548
High-Voltage Stabilization by means of the Corona Discharge between Coaxial Cylinders—S. W. Lichtman. (Proc. I.R.E., vol. 39, pp. 419–424; April, 1951.) 1950 IRE National Convention paper. The design and performance of corona-discharge voltage-regulator tubes for operation with currents of 10 to 200 μa at voltages between 700 v and 40 kv, are described. The dependence of the mode of operation and efficiency on circuit parameters is discussed.

621.396.6.017.71.012.3 2549
Estimating Temperature Rise in Electronic Equipment Cases—R. J. Bibbero. (Proc. I.R.E., vol. 39, pp. 504–508; May, 1951.) The discussion is concerned with airborne equipment. Charts are presented as an aid in calculating temperature rises. Corrections are given for pressure variations, and the effects of case color, high aircraft speed, etc., are considered.

621.396.78† 2550
Power Supplies for Large Transmitters—H. Kropp. (*Fernmeldetech. Z.*, vol. 4, pp. 25–30; January, 1951.) Review of different types of rectifiers for low- and high-voltage supplies.

TELEVISION AND PHOTOTELEGRAPHY

621.397 2551
Cathode-Ray Picture Telegraphy—F. Schröter. (*Telefunken Ztg.*, vol. 23, pp. 111–118; September, 1950.) Inherent tube and circuit factors tending to reduce the resolution attainable in practice in a cathode-ray tube are discussed. For picture telegraphy, "flying-spot"-type scanning systems appear to be most suitable, used in conjunction with a cathode-ray tube at the receiver. The electronoptical development is described of systems of this type having the following properties: ability to transmit directly, by reflection scanning, unprepared material such as manuscripts, drawings and photographs; elimination of mechanical aspects capable of affecting quality of transmission; the possibility of varying the aspect ratio and of emphasizing particular parts of the material; immediate readability at the receiver on a long-lag screen producing the same sharpness and co-ordinate fidelity as at the transmitter.

621.397.5 2552
B.B.C. Television—T. H. Bridgewater. (*Electronic Eng.*, vol. 23, pp. 120–125; April, 1951.) A brief history of the outside-broadcasts section of the B.B.C. television service, together with a comparison of the characteristics and performance of the cable and radio links used to convey pictures from the pickup point to the main transmitter at Alexandra Palace. Future developments, which will greatly extend the scope of such broadcasts, are outlined. See also 752 of April.

621.397.5:535.62 2553
Quality of Color Reproduction—D. L. MacAdam. (Proc. I.R.E., vol. 39, pp. 468–485; May, 1951; *Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 56, pp. 487–512; May, 1951.) A discussion of methods of evaluating the quality of color reproduction in television in which, as in methods already used in color photography, subjective judgments are compared with color measurements made; e.g., those made on the I.C.I. system.

621.397.5:621.317.35 2554
Notes on TV Waveform Monitor Frequency Response—Hurford. (*See* 2475.)

621.397.5:778.5 2555
American Television Film Recording Equipment—R. B. Hickman. (*Jour. Telev. Soc.*, vol. 6, pp. 167–169; October–December, 1950.) The method used with the R.C.A. Kinephoto equipment to perform the conversion from the 30 frames per second of U.S. television to the 24 frames per second of standard motion-picture projection, is described. Either an electronic or a mechanical shutter can be used to blank out the image during the pull-down interval. Details of exposure and processing of the film are given.

621.397.611.2 2556
Television Camera Tubes—E. L. C. White and J. D. McGee. (*Wireless Eng.*, vol. 28, pp. 163–164; May, 1951.) Comment on 1264 of June (Bedford).

621.397.62 2557
Some Aspects of Single Side-Band Receiver Design—W. M. Lloyd. (*Jour. Telev. Soc.*, vol. 6, pp. 135–149; October–December, 1950.) The discrepancies which appear in the response of the receiver to a unit step are discussed theoretically in relation to those features of the frequency characteristics which give rise to them. The experimentally obtained step-responses of two typical receivers are shown.

621.397.62:621.385.2:546.289 2558
An Analysis of the Germanium Diode as Video Detector—Whalley, Masucci and, Salz. (*See* 2589.)

621.397.62:621.396.67 2559
Indoor Television Aerial—H. Page. (*Wireless World*, vol. 57, pp. 168–170; May, 1951.) The antenna consists of a horizontal slot, about $\lambda/2$ long, in a vertical conducting sheet with a gain of 4 db over a vertical $\lambda/2$ dipole. Details of construction and performance are given, showing negligible change of gain, impedance, and radiation patterns for a 10 per cent frequency change.

621.397.621.2 2560
Material-Saving Picture Tube—L. E. Swedlund and R. Saunders, Jr., (*Electronics*, vol. 24, pp. 118–120; April, 1951.) The use of an es focusing system instead of the usual magnetic type, economizes in alnico-5 and copper. Performance of the new electron gun is at least equal to that of the magnetic type, and may even be the better.

621.397.645 2561
New Video Circuits in Modern TV Sets—E. M. Noll. (*Radio-Electronics*, vol. 22, pp. 26–27; April, 1951.) Video amplifier circuits, used by a number of United States manufacturers, are illustrated and briefly described.

621.397.645 2562
Shunt-Regulated Amplifiers—V. J. Cooper. (*Wireless Eng.*, vol. 28, pp. 132–145; May, 1951.) Describes circuits used for modulating television transmitters, and designed to produce across a substantially constant load, large voltage swings regulated to ensure faithful reproduction. Numerous variants of the circuit are classified and analyzed. Practical applications and experimental results are given.

621.397.8.08 2563
U.H.F. TV Propagation Measurements—K. H. Cook and R. G. Artman. (*Tele-Tech*, vol. 10, pp. 50–51, 93, and 52–54, 82; March and April, 1951.) Measurements of peak field intensity of the vision signal, and observations of relative picture quality were made at 130 locations within 25 miles of the experimental transmitter at Kansas City, under typical broadcasting conditions. Vision frequency was 507.25 mc, radiated power, 3,450 kw. Equipment is described and results are reported and discussed.

621.396.615.17 2564
Time Bases (Scanning Generators) [Book Review]—Puckle. (*See* 2492.)

TRANSMISSION

621.396.61 2565
The First High-Power Transmitter Built Since 1945 [in Germany]—K. Müller. (*Telefunken Ztg.*, vol. 23, pp. 31–38; September, 1950.) Descriptions, with block diagrams and tube details, are given for the following: (a) 100-kw broadcast transmitter, 150 to 300 kc, at Königs Wusterhausen; (b) 5-/2-kw broadcast transmitters, 545 to 1,500 kc, for north-west Germany; (c) 20-kw broadcast transmitter, 545 to 1,500 kc, at Potsdam and Hanover; (d) 100-kw broadcast transmitter, 525 to 1,610 kc, at Berlin-Britz; (e) 30-kw telegraphy transmitter, 100 to 150 kc, at Bad Vilbel, Frankfurt am Main; (f) 60-kw telegraphy transmitter, 75 to 150 kc, also at Bad Vilbel. Innovations as compared with pre-1945 practice include: thoriated instead of plain tungsten cathodes in the directly heated tubes; single-circuit cooling; ignitron protecting devices for transmitters of power >20 kw; a simple thermostat control for the quartz crystals, giving frequency constancy to within 10^{-7} over periods of 24 hours.

621.396.61:621.317.79:621.3.018.78† 2566
Measurement of Distortion in Broadcast Transmitters—H. Müller. (*Telefunken Ztg.*, vol. 23, pp. 53–66, September, 1950.) Apparatus for the measurement of nonlinear distortion, developed by Telefunken from 1946 onwards, is described. The filter method of measuring harmonic distortion is adequate for the general monitoring of transmission quality. For more stringent requirements, particularly when investigating the nature of the distortion and its frequency dependence towards the upper transmission-frequency limit, a two-tone method such as that of von Braunmühl (456 of 1935) is used, enabling symmetrical and asymmetrical distortion to be separated. For the range 30 to 150 cps, a search-tone method is used, enabling the individual harmonics to be separated.

621.396.61:621.385.4 2567
A 30-Watt Transmitter for 430 Mc/s Employing the Transmitting Valve QQE 06/40

(AX 9903)—(*Phillips Tech. Commun.* (Australia), no. 2, pp. 14–17; 1951.) See also 2062 of September (Dorgelo and Zijlstra).

621.396.78† 2568

Power Supplies for Large Transmitters—H. Kropp. (*Fernmeldetech. Z.*, vol. 4, pp. 25–30; January, 1951.) Review of different types of rectifiers for low- and high-voltage supplies.

TUBES AND THERMIONICS

537.533.8 2569

Secondary Electron Emission from Aluminum Oxide—A. R. Shul'man and I. Yu. Rozentsveig. (*Compt. Rend. Acad. Sci.* (URSS), vol. 74, pp. 497–500; September 21, 1950.) A report on an experimental investigation of the effect of temperature on the secondary-emission coefficient of Al_2O_3. No variation with temperature was observed.

537.534.8 2570

Positive Emission from Thermionic Cathodes—K. H. Steigerwald. (*Z. angew. Phys.*, vol. 2, pp. 491–493; December, 1950.) Excitation of the fluorescent screen of an electron microscope was observed even when the negative voltage applied to the control electrode was sufficient to cut off electron emission from the cathode. The effect was traced to the emission of positive ions which release secondary electrons at the control electrode, the ion current being of the order of 10^{-9}–10^{-8} A for the tungsten hairpin cathode used, and occurring only in the range of cathode temperatures 1,300°k to 1,800°k. The emission is thought to depend on a vaporization process.

537.58 2571

Elements of Thermionics—W. E. Danforth. (Proc. I.R.E., vol. 39, pp. 485–499; May, 1951.) A survey, intended primarily for workers in other fields, of the principal experimental and theoretical developments in thermionics. From a simple basis of statistical mechanics, relations are derived including the Richardson equation, the Schottky field-effect equation, and the Fowler equation for emission from a normal impurity semiconductor.

621.314.632+621.315.592† 2572

The Characteristics and Some Applications of Varistors—Stansel. (*See* 2363.)

621.385 2573

Reliability in Miniature and Subminiature Tubes—P. T. Weeks. (Proc. I.R.E., vol. 39, pp. 499–503; May, 1951.) The meaning of the term "reliability" as applied to tubes is discussed. Reliability is found to be a function not only of tube design and quality, but also of the relation between tube ratings and the operating conditions and requirements. Specific features discussed include ruggedness, operating temperature, emission stability, and life, and consideration is given to the general effect of reducing tube size.

621.385 2574

Valve Development of Telefunken since the Cessation of Hostilities (1945)—H. Rothe. (*Telefunken Ztg.*, vol. 23, pp. 93–96; September, 1950.) On account of difficulties due to the condition of the various plants, no considerable development was possible till the second half of 1948. Since then the broadcast-receiver 11 series has been completed in glass-envelope tubes and, for usw FM receivers, in metal-envelope tubes. A new series of miniature tubes ("pico" series) was marketed in mid-1949. Some half-dozen power tubes intended for communications, broadcast transmitters, and industrial generators, are also very briefly described.

621.385:621.386.1 2575

Radiographic Examination of Electronic Valves—van Wijlen. (*See* 2508.)

621.385:621.396.619.16 2576

Signal-Retardation Electron Tubes with Delay Modulation—É. Labin. (*Onde élect.*, vol. 31, pp. 82–89; February, 1951.) An electron beam, modulated in density by the signal to be retarded, is passed between a pair of deflecting plates before injection into a retarding chamber within which delay is effected by means of a magnetic field, which causes the beam to follow a helical trajectory. The amount of the delay is dependent on the angle of entry into the chamber, and this is controlled by the deflector plates, to which the required "delay" modulation voltage is applied. In the nonmodulated condition, delay may be of the order of 100 μs. Modulated, the maximum obtainable delay is an inverse function of the modulation frequency; at the limiting frequency, dispersion due to space charge may limit the output current to <1 μa. This and other limitations of the method are discussed, and illustrations are given of tubes constructed to test the validity of the principle.

621.385.012:621.317.799† 2577

Tube Characteristic Tracer Using Pulse Techniques—Wagner. (*See* 2488.)

621.385.029.6 2578

Pulse Technique in High-Power Valve Development—A. M. Hardie. (*Metrop. Vick. Gaz.*, vol. 23, pp. 350–360; April, 1951.) The trend of high-power tube development is discussed, with particular reference to the future use of demountable tubes on sw service. Design problems are enumerated. A recording technique is described for presenting positive-grid characteristics in a form suitable for engineering applications. Either point-by-point or photographic recording may be used. Some examples of the latter are presented.

621.385.029.63/.64 2579

Amplification of the Traveling Wave Tube—B. Friedman. (*Jour. Appl. Phys.*, vol. 22, pp. 443–447; April, 1951.) A simpler and more exact method is presented for solving the transcendental equation given by Chu and Jackson (3549 of 1948) for wave propagation in the helix of the traveling-wave tube. The propagation is considered as a perturbed form of that in the cold helix. The dependence of amplification factor on geometrical parameters and operating conditions is determined explicitly. The tube will not amplify if the dc beam current is too high.

621.385.029.63/.64 2580

Effect of Hydrostatic Pressure in an Electron Beam on the Operation of Traveling-Wave Devices—P. Parzen and L. Goldstein. (*Jour. Appl. Phys.*, vol. 22, pp. 398–401; April, 1951.) Small velocity spreads in the electron beam appear to cause a decrease in gain and noise figure of a traveling-wave tube. The physical explanation is that the velocity spread introduces interactions between the electrons in which the external circuit takes no part, this interaction being of the nature of a hydrostatic pressure.

621.385.029.63/.64 2581

Travelling-Wave Tubes with Dispersive Helices—F. N. H. Robinson. (*Wireless Eng.*, vol. 28, pp. 110–113; April, 1951.) Oscillation occurs in traveling-wave-tube amplifiers when reflection takes place at mismatches between the ends of the helix and the external circuit. The difficulty of obtaining good matching over the wide frequency band of normal tubes has led to the development of a dispersive helix in which the phase velocity varies rapidly with frequency. This is achieved by making the diameter of the helix very small. Amplification then occurs for only a limited range of frequencies over which correct termination of the helix is possible. By this means the beam current required to produce a given gain is much reduced. Noise factor is also comparatively low.

621.385.032.213:537.533.8 2582

Secondary-Emission Cathodes of High Stability—B. D. Tazulakhov. (*Zh. Tekh. Fiz.*, vol. 20, pp. 773–787; July, 1950.) The preparation of cathodes possessing a high stability under high temperatures and heavy current loads, was investigated experimentally. The requirements which the active and intermediate layers of the cathodes should satisfy are defined, and tables showing the properties of various suitable materials are given. The performance of complex BaO emitters deposited on Ag, Cu, Ni, nichrome, Mo, and Ta, is discussed in detail, and experimental curves showing the secondary emission from these cathodes are plotted. It is claimed that in the production technique proposed, the thickness control of the emissive layer is much simpler than in the usual methods, where it is more of the nature of an art than of a technological process.

621.385.15:621.385.831 2583

Voltage-Controlled Secondary-Emission Multipliers—A. J. W. M. van Overbeek. (*Wireless Eng.*, vol. 28, pp. 114–125; April, 1951.) Secondary-emission tubes have, in some cases, a much shorter life than normal tubes. This objectionable feature has been overcome by using a coating of Cs_2O on the dynodes and keeping their temperature below 180°C. The constructions of various experimental tubes are shown, and their characteristics described. A variable-μ tube and a very-high-slope tube with four stages of multiplication are shown. The use of grid dynodes is discussed. Some circuits in which secondary-emission tubes offer specific advantages are described, including generators of sinusoidal and nonsinusoidal oscillations, and trigger circuits.

621.385.16:537.312.5 2584

Magnetic Electron Multipliers for Detection of Positive Ions—L. G. Smith. (*Rev. Sci. Instr.*, vol. 22, pp. 166–170; March, 1951.) Two designs of 15-stage multipliers with crossed electric and magnetic fields are described. BeCu dynodes are used, of width $\frac{3}{8}$ inch for fields of 250 to 460 oersted, and $\frac{1}{8}$ inch for fields of 300 to 1,100 oersted. From their performance it is concluded that a multiplier of this type could be designed to have a rise time between 10^{-10} and 10^{-11} seconds.

621.385.2 2585

Effect of Variable Mass of the Electron on the Space-Charge Limited Current in a Diode—S. Visvanathan. (*Canad. Jour. Phys.*, vol. 29, pp. 159–162; March, 1951.) "The change in the current-potential distribution, due to the relativistic variation of the mass of the electron, has been calculated by suitable series expansions in the case of a plane parallel diode, and has been shown to be considerable in the case of large power tubes."

621.385.2 2586

The Transformation of Heat into Electrical Energy in Thermionic Phenomena—R. Champeix. (*Le Vide*, vol. 6, pp. 936–940; January, 1951.) An experiment is described and theory is adduced showing that, in a thermionic diode, the standing current vanishes when the two electrodes are at the same temperature, independently of the composition of the electrodes. Practical suggestions are made for the design of a diode without standing current, and for the determination of the actual source of emission of electrons from oxide-coated cathodes.

621.385.2:[546.27+546.289 2587

Crystal Diodes—R. W. Douglas and E. G. James. (*Proc. IEE*, vol. 98, pp. 157–168; Discussion, pp. 177–183; May, 1951.) The influence of small amounts of impurities on the electrical properties of semiconductors, and the mechanism of contact rectification, are discussed. The processing of Ge and Si for use in crystal diodes is considered in the light of the theory. The design and performance of

(a) a coaxial-type Si-crystal diode for use as a mixer at frequencies up to about 10 km, and (b) a wire-ended Ge-crystal diode are described. Particular attention is given to the frequency dependence of the rectification efficiency of the Ge diode, and to its application as a replacement for the thermionic diode.

621.385.2:546.289 2588

A New High-Conductance Crystal Diode—B. J. Rothlein. (*Sylvania Technologist*, vol. 4, p. 44; April, 1951.) The experimental Ge diode described is made by applying to the whisker contact an amount of metal paste so small that is does not add appreciably to the capacitance.

621.385.2:546.289:621.397.62 2589

An Analysis of the Germanium Diode as Video Detector—W. B. Whalley, C. Masucci, and N. P. Salz. (*Sylvania Technologist*, vol. 4, pp. 25–34 April, 1951.) Methods, including some rapid production-line tests, are discussed for the measurement of those characteristics of Ge diodes which are important in the detection of video signals. The forward and reverse conductances are assumed constant over the range of operation, and both loads with small and loads with large time constants are considered.

621.385.3+621.385.5 2590

Interelectrode Impedance in Triodes and Pentodes—E. E. Zepler and S. S. Srivastava. (*Wireless Eng.*, vol. 28, pp. 146–150; May, 1951.) Bridge measurements of capacitance and conductance were made at 1 mc and 32 mc, and the values were plotted against mutual conductance. Discrepancies between observed capacitance variations and the values indicated by North's theory (1450 of 1936) are discussed, and explanations are advanced for some of the effects.

621.385.3:546.289 2591

Effect of Auxiliary Current on Transistor Operation—H. J. Reich, P. M. Schultheiss, J. G. Skalnik, T. Flynn, and J. E. Gibson. (*Jour. Appl. Phys.*, vol. 22, pp. 682–683; May, 1951.) Transistor gain characteristics may be improved by the flow of direct current between auxiliary electrodes, one of which is placed as close as possible to the collector. The best improvement in current gain is obtained with relatively large spacing between emitter and collector.

621.385.3:546.289 2592

Crystal Triodes—T. R. Scott. (*Proc. IEE*, vol. 98, pp. 169–177; Discussion, pp. 177–183; May, 1951.) The various forms of crystal triode developed up to date are reviewed. A brief résumé is given of the various materials for the manufacture of these triodes, and the types of control used to modify their characteristics. Testing procedure is discussed. Applitions and circuit design are dealt with briefly.

621.385.3.029.64 2593

Passive Feedback Admittance of Disc-Seal Triodes—G. Diemer. (*Philips Res. Rep.*, vol. 5, pp. 423–434; December, 1950.) A discussion of the design of disk-seal triodes with a view to using the self-inductance of the grid wires to neutralize the feedback via the anode-cathode capacitance at microwave frequencies.

621.385.3.032.24 2594

Aspects in the Design and Manufacture of Planar Grids for Triodes at U.H.F.—W. J. Pohl. (*Electronic Eng.*, vol. 23, pp. 95–99; March, 1951.) Discussion, with calculations and application to practical manufacturing problems, of the relation between the grid dimensions and its ability to dissipate power. A recently developed method of producing planar ring-frame grids carrying highly tensioned wires, and the method of measurement of residual wire tension, are described. The most suitable material for constructing tensioned grids is tungsten wire with a copper coating of thickness about 15 per cent of the radius of the wire.

621.385.4:621.396.61 2595

A 30-Watt Transmitter for 430 Mc/s Employing the Transmitting Valve QQE 06/40 (AX 9903)—(*Philips Tech. Commun.* (Australia), no. 2, pp. 14–17; 1951.) See also 2062 of September (Dorgelo and Zijlstra.)

621.385.832.001.4 2596

A Note on Cathode-Resistance Stabilization of C.R.T. Gun Current—H. Moss. (*Electronic Eng.*, vol. 23, pp. 111–112; March, 1951.) An expression defining the increase in current stability produced by an autobias cathode resistor is deduced in terms of a stability factor S given by $2S=7E_c/E_d-5$, where E_c and E_d are the grid cutoff voltage and drive voltage, respectively. Graphs are drawn of S against cathode resistance for various cutoff voltages.

621.386.7 2597

Centering the Cathode in a Demountable X-Ray Tube—R. Fouret. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1651–1653; April 30, 1951.) Centering is performed while the tube is under vacuum, by an electrical method which consists of rendering minimum the capacitance between anode and cathode.

621.396.615.141.2 2598

On the Theory of the Anode Block of a Plane Magnetron—S. D. Gvozdover and V. M. Lopukhin. (*Zh. Tekh. Fiz.*, vol. 20, pp. 955–960; August, 1950.) A mathematical discussion is presented of magnetrons with anode blocks of the hole-and-slot and slot types. To simplify the discussion, the space occupied by the anode block is divided into the interaction space, which does not include the anode resonators, and the space which includes these resonators. Equations for the electromagnetic fields in the interaction space are derived and solved, and the natural frequencies of oscillation are determined by an approximate method in which the complex impedances of the interaction space are matched to those of the resonators. The discussion is limited to the case of two-dimensional (plane) magnetrons, and the end effects, as well as the effects of couplings, are neglected.

621.396.615.141.2:537.525.92 2599

The Space-Charge in a Magnetron under Static Cut-Off Conditions: Planar or Quasi-planar Magnetron—G. A. Boutry and J. L. Delcroix. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1413–1415; April 9, 1951.) The two static cutoff states [1828 of 1950 (Delcroix and Boutry)] are compared; the total space charge is the same for the two cases. The condition of lower energy level of the electron gas corresponds to the Brillouin state. The cutoff surface is the same for the two cases.

621.396.615.141.2:537.525.92 2600

The Space-Charge in a Magnetron under Static Cut-Off Conditions: Cylindrical Magnetron—J. L. Delcroix and G. A. Boutry. (*Compt. Rend. Acad. Sci.* (Paris), vol. 232, pp. 1653–1655; April 30, 1951.) The two static cutoff states are compared (see 2599 above for corresponding consideration of planar magnetrons). The total space charge is different for the two cases; the condition of lower energy level of the electron gas corresponds to the "bidromic" state. The cutoff surface is not generally in the same position for the two states.

621.396.615.141.2:537.525.92 2601

Analysis of Synchronous Conditions in the Cylindrical Magnetron Space Charge—H. W. Welch, Jr., and W. G. Dow. (*Jour. Appl. Phys.*, vol. 22, pp. 433–438; April, 1951.) "In the multianode cylindrical magnetron there exist favored phase velocities of the electromagnetic wave around the interaction space between anode and cathode These velocities are characteristic of the resonant system attached to the anode segments. In the oscillating magnetron, the electronic space charge within the interaction space is presumed to maintain synchronism with one of these velocities. Certain of the conditions of synchronism, which can be discussed analytically, are treated in this paper The results, although based on restrictive assumptions, can be used in the interpretation of magnetron operation, and in predicting regions of efficient behavior. See also 1282 of June (Welch).

621.396.615.142 2602

The Limiting Efficiency of Oscillation Generation by Means of Velocity-Modulated Electron Beams in Drift-Space Valves with Fields of Finite Length—R. Gebauer and H. Kosmahl. (*Z. angew. Phys.*, vol. 2, pp. 478–486; December, 1950.) The concept of the ideal efficiency is introduced. This quantity is a measure of the greatest possible amount of hf energy which can be extracted from the tube, neglecting the velocity modulation, and is hence also an indication of the quality of the focusing. The optimum length of drift-space for a given modulation depth and control-gap length is calculated. The focusing properties of infinitely short and finite-length fields are compared and found to be equivalent only for vanishingly small modulation. The relation of the practically attainable limiting efficiency to the ideal efficiency, is defined. The practical limiting efficiency decreases with increase of modulation depth.

621.385.032.216 2603

Die Oxydkathode: 2. Teil—Technik und Physik [Book Review]—G. Harrmann and S. Wagener. Publishers: J. A. Barth, Leipzig, Germany, 2nd edn, 1950, 284 pp. (*Fernmeldetech. Z.*, vol. 4, p. 46; January, 1951.) A modern and exhaustive exposition of the subject. Vol. 1: 2985 of 1949.

MISCELLANEOUS

621.396 2604

Radio Progress during 1950—(PROC. I.R.E., vol. 39, pp. 359–396; April, 1951.) A survey based on material compiled by the 1950 Annual Review Committee of the IRE, and including 1,084 references. The material is grouped under the following headings: antennas and waveguides; audio techniques; electroacoustics; sound recording and reproducing; circuit theory, electron tubes and solid-state devices; electronic computers; facsimile; industrial electronics; measurements; mobile radio; modulation systems; navigation aids; piezoelectric crystals; radio transmitters; receivers; standards on symbols; television system; video techniques; wave propagation.

621.396(083.72) 2605

Standards on Abbreviations of Radio-Electronic Terms, 1951—(PROC. I.R.E., vol. 39, pp. 397–400; April, 1951. Reprints of this Standard, 51 IRE 21 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.

621.396 Tesla 2606

The Life and Work of Nikola Tesla—A. Damianovitch. (*Bull. Soc. franç. Élect.*, vol. 1, pp. 85–99; February, 1951.) Lecture before the Société française des Électriciens, reviewing the pioneer work of Tesla in the field of ac and radio engineering.

621.39 2607

Electrical Engineers' Handbook—Electric Communication and Electronics [Book Review]—H. Pender and K. McIlwain. Publishers: Chapman and Hall, London, Eng., and J. Wiley and Sons, New York, N. Y., 4th edn, 1345 pp., 68s. (*Electrician*, vol. 146, p. 823; March 9, 1951.) An entirely rewritten edition, with contributions by 78 specialists. FM and pulse techniques in communications and radar are included for the first time. A bibliography is appended to each of the 23 sections.

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Thorough Aging— *all units undergo lengthy artificial aging at high temperature to insure mechanical and electrical stability of cards*

Good Frequency Characteristic — *most units can be used at frequencies as high as several hundred kilocycles, as well as at dc*

Low Temperature Coefficient *less than ±0.002% per deg. C. at room temperatures for all decades except the 0.1 ohm unit*

Excellent Electrostatic Shielding— *units mounted on 3/16th inch aluminum panels and enclosed in welded aluminum cases with 1/8th inch walls*

Low and Constant Switch Contact Resistance *between 0.002 and 0.003 ohm*

Dial Values Overlap— *eleven contact studs with ten resistance cards*

Cards Adjusted at Their Own Terminals—*resistance increments always correctly read*

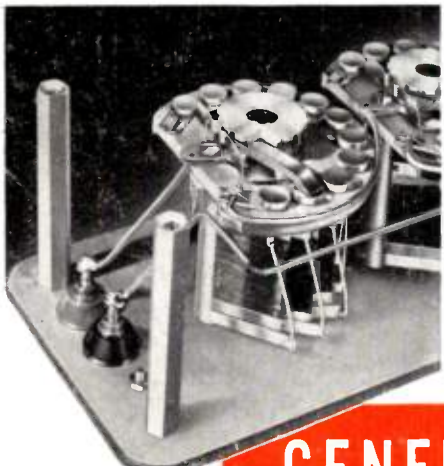
Separate Ground Terminal on Case— *resistors can be shielded whether used grounded or ungrounded*

Positive Detent Mechanisms

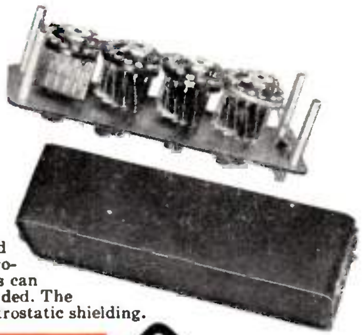
Current-Carrying Capacity Engraved on Each Decade

Ten Sizes— *between 0.1 to 111 ohms and 10 to 1,111,100 ohms*

Conservatively Priced— *between $56 and $133*



The 100,000-ohm resistance unit. Note the four-leaf off-set switch arm designed to wipe across contact studs. The 1-, 10-, and 100-ohm steps are Ayrton-Perry wound on molded phenolic forms especially shaped and heat treated to minimize aging effects. The 0.1-ohm steps are bifilar-wound with ribbon; the 1,000-, 10,000-, and 100,000-ohm steps are unifilar wound on thin mica cards.



The Type 1432-Q unit removed from its welded aluminum box. A separate ground terminal is provided on the aluminum panel so that the units can be shielded, whether used grounded or ungrounded. The aluminum panel and box provide excellent electrostatic shielding.

GENERAL RADIO Company

275 Massachusetts Avenue, Cambridge 39, Mass.

90 West Street NEW YORK 6 920 S. Michigan Ave CHICAGO 5 1000 N. Seward St. LOS ANGELES 38

new PRECISION RESISTANCE BOXES

FOR MANY YEARS the General Radio Type 602 Decade Resistance Box has been the adjustable standard of resistance in laboratories where a wide range of accurately and permanently adjusted resistance units is required.

G-R now announces a new and improved series of decade resistance units, the Type 1432. Some of the features of the new decades are:

Increased Accuracy— *basic accuracy ±0.05% — a two-fold increase over the Type 602's*

Permanency of Calibration— *stability considerably better than accuracy of adjustment — will stay within specified tolerances well beyond one-year warranty*

Thorough Aging— *all units undergo lengthy artificial aging at high temperature to insure mechanical and electrical stability of cards*

Good Frequency Characteristic — *most units can be used at frequencies as high as several hundred kilocycles, as well as at dc*

Low Temperature Coefficient *less than ±0.002% per deg. C. at room temperatures for all decades except the 0.1 ohm unit*

Excellent Electrostatic Shielding— *units mounted on 3/16th inch aluminum panels and enclosed in welded aluminum cases with 1/8th inch walls*

Low and Constant Switch Contact Resistance *between 0.002 and 0.003 ohm*

Dial Values Overlap— *eleven contact studs with ten resistance cards*

Cards Adjusted at Their Own Terminals—*resistance increments always correctly read*

Separate Ground Terminal on Case— *resistors can be shielded whether used grounded or ungrounded*

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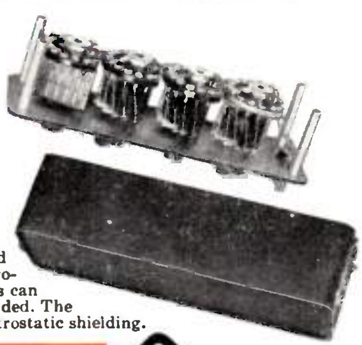

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