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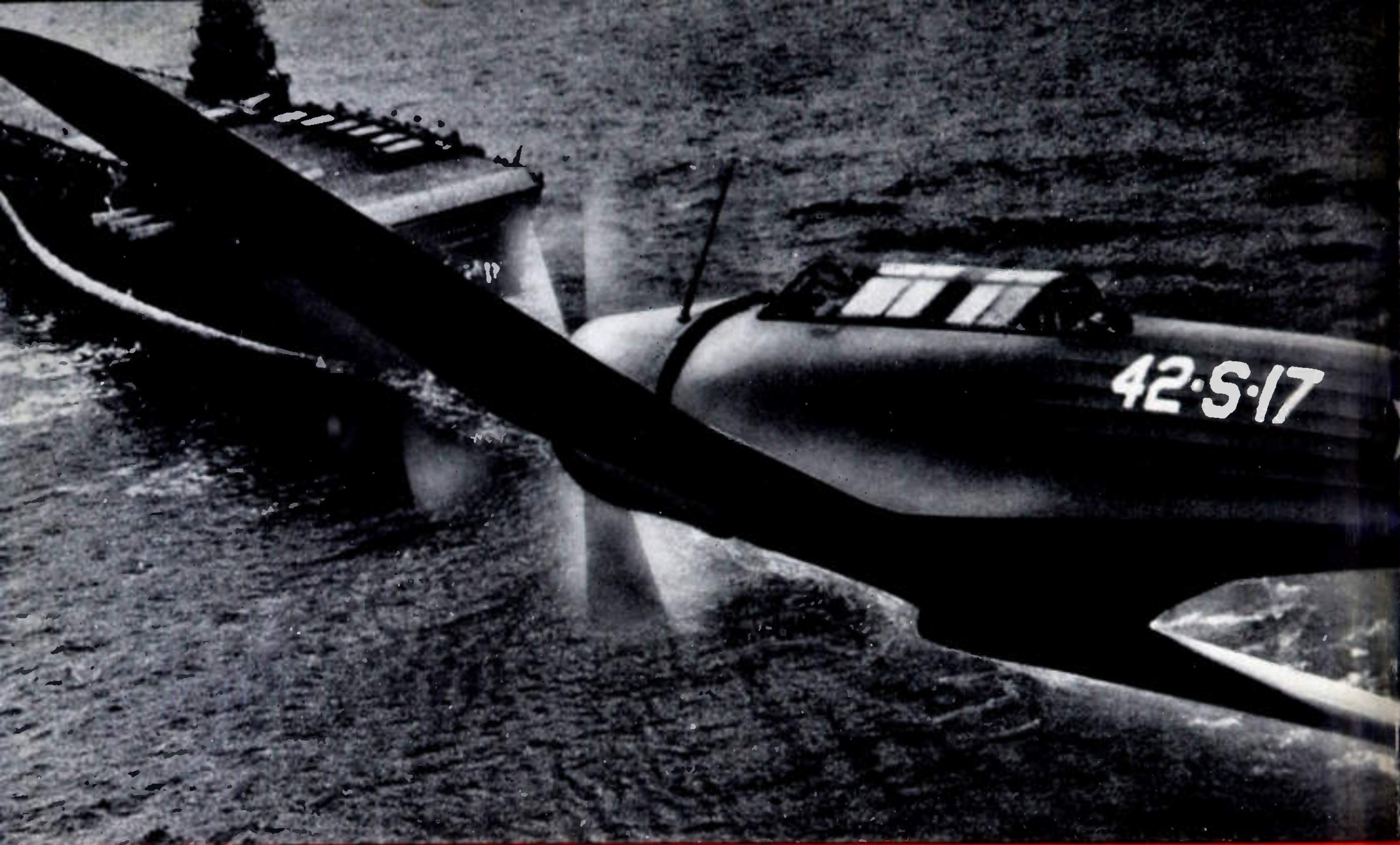
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Preparation of Technical Articles*

BEVERLY DUDLEY†, ASSOCIATE, I.R.E.

Summary—To appeal to the interests and intellectual requirements of the reader, the text of technical articles must be properly selected, arranged in logical order, and well proportioned, with essential concepts presented in proper relation to one another. To assist in meeting these requirements, a topical outline is given as a basis for the development of research or engineering articles. Recommendations are given for the typing, mailing, and proofreading of the manuscript. Suggestions are included for the preparation of diagrams and photographs so that they may be most suitable for publication.

Introduction

TECHNICAL articles, regardless of their subject, are meant to represent a contribution to knowledge. For this reason alone, if not for many others, they should be prepared with utmost care. If they are written in a highly critical time, as that into which this decade falls, they should disseminate excellence in more ways than one. The English employed should receive careful consideration. Authors who pen technical articles are not expected to delight the connoisseurs of *belles lettres*. But it is not necessary that they completely neglect their English. A spoken word may be forgiven; it may also be forgotten. But a written word spells responsibility to its author at all times. If writers would pay more attention to their language, many a reader not only would receive a far clearer meaning of the subject conveyed, but would also derive greater enjoyment from the article.

Any article intended for publication in the technical press of the United States should be written in clear, concise, good English. There are many occasions where the author has failed to make his meaning clear, because he has underestimated this simple point. The author's meaning may be entirely clear to him, but if he would take the trouble to have someone else read his article many ambiguities might be corrected in the original manuscript.

Primary Purpose of Article

To be read is the primary purpose of any article. If it is eagerly read from beginning to end, the author reaps an additional reward for his name will have imprinted itself upon the mind of the reader, and articles bearing his by-line will be looked for. Any legitimate evocation which may induce the prospective reader to follow the text attentively is permissible, provided it is done within the definite limits of good taste. Good taste never should be extenuated and should serve as a guide for all technical publications because the writer will be regarded as an authority.

It is obvious that if the article is to be read it must be interesting to the group for which it is intended. A first consideration is that the subject matter must be appropriately selected. However, the manuscript must also appeal to the intellectual and academic standards of the readers.

General Considerations Determining the Type of Writing

The type of writing which should be employed will depend upon a number of factors. If the article is to be used for reference purposes, for example, each individual section of the article should be complete in itself so that the important information is readily apparent and immediately available. On the other hand, if the article is to be used for purposes of general instruction only, the requirement that each portion of the article be an independent unit need not be so rigorously followed.

The type of writing employed will depend to a very considerable extent upon the publication in which it is to appear. Therefore it is recommended that the author study the publication to which he intends to submit his manuscript in order that he may prepare his material in conformity with its style and policies. Very frequently such publications will have certain editorial and technical requirements which are published in the journal. By following these recommendations the author will save himself a considerable amount of time and effort in the proper preparation of his article.

Ethical and Censorship Considerations

The author has definite obligations to ascertain that the statements made are as precise and accurate as possible. Errors in the published article as well as inconsistencies and mistakes will reflect upon his technical ability.

At the present time each author submitting manuscripts for publication has an additional duty imposed upon him by virtue of the fact that the United States has entered into the world conflict. Much technical literature is potentially, if not actually, of such a nature as to be of aid or comfort to the enemy. Articles falling in this classification should not be published, or at least should be modified so that they conform to the best interests of the nation. It is recognized, of course, that by withholding the publication of technical or scientific information, the allies as well as the enemies may be handicapped and there appears to be some tendency to recognize this point of view. Censorship of technical articles is the responsibility of the individual author as well as that of the editor of the publication. If there is any doubt in the author's mind as to whether or not discussion of a particular phase of his work might be contrary to the interests of the nation, he

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† Managing Editor, *Electronics*, McGraw-Hill Publishing Company, New York, N. Y.

should ascertain from his research director, or from the editor of the publication for which the article is intended, the advisability of publishing the results of his research.

Common sense demands that articles appearing in the technical press be prepared with good taste. Culture and ethics will dicate a limit to the number of times an author mentions a company's name in his article, and a reasonable degree of modesty in referring to the achievements of the persons in his organization. On the other hand, there have been cases where the work of others, who had directly or indirectly made contributions in the realm upon which the article touches, has been consciously or unconsciously overlooked. No author who submits an article for publication in one of the leading technical journals of the country should subject himself to such criticism for it is incumbent upon him to be familiar with the work of others. Fair play (if not cultural attainment) will demand that adequate recognition be given to the work of others, whether a part of his organization or not. Acknowledgment for assistance and similar courtesies may, with advantage, form a part of the published manuscript.

Five Important Points Which Help Make a Good Article

Only experience can be used to ascertain whether an article is a good one. No amount of written instructions can guarantee that an article composed in accordance with suggestions will be suitable throughout. Nevertheless, observance of the following important points will aid in assuring that the technical article is properly prepared.

(a) Selection. Include all material and data which may be necessary to bring out the main idea or message of the article. Reject all material which is irrelevant.

(b) Adaptation. The article should be adapted to the technical, cultural, academic, and intellectual attainment of the reader. As a corollary, the author should have in mind a particular type of reader for whom he is writing, even before he begins to prepare his manuscript.

(c) Arrangement. Arrange all parts of the article in a natural, logical order, keeping related topics together

(d) Articulation. The relationship of all parts of the article should be made clear and unmistakable. Much of this can be accomplished by clear, concise writing but considerable aid may also be obtained through the use of an appropriate outline in its preparation, or the use of heads and subheads in the manuscript itself to segregate appropriately the individual parts of the complete article.

(e) Proportion. The relative amount of space to be given to the various aspects of the subject should be determined by their relative importance to the article as a whole. The more important topics should receive proportionately more extensive treatment, while relatively little space should be devoted to insignificant details which the intelligent reader may be expected to supply for himself.

Writing the Manuscript

It is rather difficult to answer the question of optimum length of a manuscript since many factors enter into such a decision. The article should be long enough to treat all of the essential points adequately; no longer. The shorter an article is, while still conveying the same information, the more useful it will be. A good rule to follow in determining the length of an article is: Have something to say, say it, stop.

The author will have to determine the number of copies which should be prepared. In most cases the editor or publisher will require only one copy for his own use. But where the manuscript may be examined by several specialists in a particular field, the editor may require several copies. Publication in the Proceedings of the Institute of Radio Engineers will be expedited by submitting three copies of the manuscript and illustrations. In all cases the author should retain a copy of the manuscript for purposes of checking and proofreading.

An important point to consider in the initial preparation of an article (especially if it is mathematical) is the terms and symbolic notations employed. The author should be clear to specify at all times and in unmistakable form the terms and symbols to be employed, especially if they are used in some unusual or nonstandard form. It is advisable to employ terminology and symbolic notation which are universally recognized, and are in conformity with the standards of the various engineering and scientific societies.

No article should be written until all the necessary material which the author may require for its preparation is at hand. This will save lengthy interruptions and tend to minimize discontinuities of thought. It will be helpful if the article is planned in outline form before the text is begun. A written outline is usually most useful, but at least a mental picture of the article as a whole should preface the start.

The outline for a technical manuscript may include the following subjects:

Title
By-line
Summary or abstract
Introduction
Historical development of prior arts
Theory underlying the advances recorded in the article
Experimental method for verifying the theory
Description of experimental method and equipment
Results of experimental method
Analysis of experimental results
Correlation of theory and experiment
Account of the discrepancies between theory and experiment

Significance of discrepancies
Conclusion
Acknowledgment
Bibliography and references
Appendixes

The successful writer realizes the psychological values that an appropriate title holds. The right title is very important, yet few writers give this matter careful thought. In many cases the title sounds as if it had been tacked on hastily, just before sealing the envelope that is to carry the article to its final destination, as if the author had remembered at the last minute that the inclusion of a title was customary

The abstract or summary should be an abridgement of the essential material in the article. It enables the reader to obtain quickly an idea of the contents of the article so that he may determine whether these meet his present requirements in his search for information. The abstract must present as much information as possible as concisely as possible. The important results and conclusions should be indicated and the significance of the article always be clearly stated. It is not sufficient for the abstract or summary to state merely that certain research work has been undertaken since this conveys no information concerning the details of the research. It is much more desirable for the author to state his results and conclusions, and the significance of his work in relation to the present state of the science.

The introduction should be planned to acquaint the reader as thoroughly and as quickly as possible with the problems to come, their importance, their relation to other problems with which the reader is familiar and to pave the way for the main text which is to follow.

A sparkling introduction will immediately attract attention, but if the entire article is written in a brilliant style, it will hold the reader fascinated. Why should a technical article be unattractive? For too long a period we have seen in print uninspiring, clumsily written articles at the reading of which our eyes threatened to close several times, despite the undisputed technical contribution which they "elucidated." An article that fails to beguile us to reading it from start to finish has completely missed its aim. For a goodly number of years past we have reflected upon dullness as being commensurable with erudition; in reality we have been lacking imagination, originality, and pen appeal.

The conclusions should state clearly the deductions which are logically drawn from the treatment as developed in the article. The conclusion is mainly a statement of results achieved or new thoughts which have been derived as a result of the article.

Mechanical Considerations in Writing a Manuscript

The physical and mechanical form in which a manuscript is presented can do much to make an editor feel kindly or unkindly toward accepting an article.

It should go without saying that all copy should be typewritten with a machine having a clean ribbon and making a firm impression. All typing should be done on white paper $8\frac{1}{2}\times 11$ inches in size. It is especially desirable to avoid the use of legal-size paper since this does not conveniently fit into filing cabinets and the standard size of mailing envelopes. Margins of at least 1 inch should be left all around the paper although it is preferable to leave a margin of $1\frac{1}{2}$ inches on the left-hand side for binding.

All text should be double spaced, in order that appropriate editing and printer's marks may be inserted on the original manuscript.

Do not include drawings on pages of typed text. The text is sent to the printer who sets it in type whereas the drawings are usually sent to the engraver who makes cuts of them. If the drawings are on the same sheet with the text, it is necessary either to cut the drawings and text apart, or to have the printer and engraver work from the same sheet. Both practices are undesirable. The latter is especially bad since the copy may become easily damaged and soiled making it impossible for the engraver to produce satisfactory cuts from them.

Use of Mathematics

Sometimes the question arises as to the extent to which mathematics can be used advantageously in a technical article. The answer to this question can be given only by the author himself. He should remember that the use of an extensive amount of mathematical treatment will necessarily limit the audience which can appreciate his work since not all persons (not even all engineers and all scientists) are capable of reading and properly interpreting mathematical statements. Likewise, it should be recognized that mathematics are necessary only where quantitative considerations are involved. On the basis of these thoughts we may establish the following three points for the author to bear in mind: (1) Use mathematics only where they are needed to develop some quantitative relationships, (2) omit the use of mathematics where their employment is not essential for a clear understanding of the subject, (3) use only as much mathematics as are required, and keep these in the most simple form consistent with the development of the essential quantitative concepts.

An author can do much to simplify the mathematical form and complexity with which his article appears. This is especially true if he deals with relatively simple fractions. In such cases it is highly desirable, from a typographical point of view, to write such equations with the free use of parentheses, brackets, and fraction bars, rather than to use a long dividing line with numerator above and denominator below. This latter practice requires at least two or three lines of type and must be set by hand, whereas the recommended practice frequently makes it possible to set a relatively complicated equation of the fraction type on a typesetting machine. Likewise, complicated exponents may be greatly shortened by writing it in the form "exp" followed by the exponent. For example, it is simpler to set the equation in the form

$$I = AT^2 \exp(-b_0/T)$$

than in the form

$$I = AT^2\epsilon^{-b_0/T}.$$

Typesetters are not mathematicians and the simpler the author can make the physical form of his mathematics, the more nearly certain he can be that his text will come through on galley or page proofs in the required manner.

Illustrating the Article

Illustrations are usually a highly important part of any well-prepared technical article. They may be used to present a better visual picture than can be done with words, to further the concept of quantitative relationship (as through the employment of graphs), or to create an easier understanding of the relationships to various parts than can be done in type.

All illustrations, whether photographs or line drawings, should be clear, clean cut, and to the point. The intended information should be designated easily and clearly and any extraneous material should be eliminated. In photographic illustrations, particular attention should be given to the background which may show material in which the author is not particularly interested. It is advisable to provide sufficiently wide margins in all cases. In making a cut for purposes of reproduction, the editor can always crop an illustration if it shows unnecessary extraneous matter, but he is not in a position to supply additional information which the author does not provide for him.

It is, of course, necessary to provide adequate captions for cuts for all illustrations. These captions together with their illustrations should be sufficiently complete to convey to the reader adequate information to enable him to interpret the illustration properly. It should not be necessary to read the text completely to understand what an illustration is about.

Requirements for Photographs

A good half-tone reproduction can be obtained only from a good photoengraving, and this in turn can only be made if the original photograph is a good one. Glossy 8-×10-inch photographic prints which are clear and sharp are by far the most suitable for publication. The "soot and whitewash" type of contrasty print is undesirable. The photographic print should show all tonal ranges from clear white to black, with good gradation in varying shades of gray. Cuts from which printed reproductions are made employ a metal plate whose surface has been finely screened and appreciable detail may be lost in the screening process. Consequently, the printed reproduction will always show less detail than the photographs from which it is made.

It is well to have photographs made by a professional photographer. Do not submit snapshots made with a small camera or in which the dominant subject is such a small part of the entire picture that it is lost. If you make the photographs yourself, use a camera making negatives at least 4×5 inches in size. If possible use a camera with a focusing back screen so that as the inverted image on the ground glass of the camera is viewed; the necessary camera adjustments and arrangement of equipment and lights can be made in the desired manner. In general, it is advisable to provide an adequate amount of diffuse illumination to make a relatively long exposure with a small aperture (corresponding to a large f number). By this procedure (which is of course suitable only for still objects) the sharpest and clearest of images may be obtained with the least amount of trouble and effort.

Legends should be provided for each photograph. Each must be complete and accurate to show its intended relationship to the article as a whole. The legends may be pasted on the back of the photographs or at the bottom of the prints. For the PROCEEDINGS it is preferred that they be typed together on a separate sheet and the illustrations marked on the border with figure numbers only. Whichever method be used, the legend must be unmistakably identified with the appropriate print. Do not attempt to write or type on the back of photographic prints as the impressions may show through on the glossy surface.

A photograph many times the size of the original should clearly indicate the amount or extent of magnification in the legend. Likewise, this magnification will have to be modified by the ratio of reduction when the engraver's cut is made from the print. It would be well for the author to check this change in size when he receives cut proofs from the editor.

Do not use paper clips on pictures as they frequently dent the picture's surface. Photographs should be mailed in plain envelopes with a cardboard stiffener to prevent or minimize damage.

Requirements for Sketches and Graphs

All drawings for reproduction must be in black on white paper, or preferably, on tracing cloth or Bristol board. Black India ink (with a carbon base) should be used in all cases.

All graphs, drawings, and sketches should be in complete, finished form, properly lettered in ink so that the engraver's cut may be made directly from the drawing. Technical publications do not have the time, money, or facilities, nor can they assume the responsibility for having illustrations redrawn by a professional draftsman. It is the duty of the author to provide illustrations of sufficient quality that they may be used directly to make a half tone or line cut.

Lettering should be 1/4-inch high for graphs of approximately 8×10 inches in size. Drawings which are smaller or larger than 8×10 inches should have lettering proportionately smaller or larger respectively.

Do not letter figure numbers or captions on the drawings. These should be inserted in the margin where they will not interfere with the sketch or graph.

Do not use orange or green cross-section paper for preparing graphs for reproduction. The co-ordinates invariably show up and produce an undesirable cross-hatch backing. If the divisions in the graph are close together, these lines run together and block up the drawing producing a dirty and untidy appearance. Since blue does not photograph on emulsions used by photoengravers, blue graph paper can be used and the blue co-ordinates will be eliminated when the engravings are made. All important co-ordinates should be properly drawn with India ink. The co-ordinates should be spaced at least 1/4 inch for an 8-×10-inch graphical illustration. Most engineering graphs or curves are simply used to indicate a variation of functional relationships. They are seldom intended for the reader to obtain exact values from the curves. However, if this is the purpose, then the graphs should be as large as possible and should be very accurately drawn.

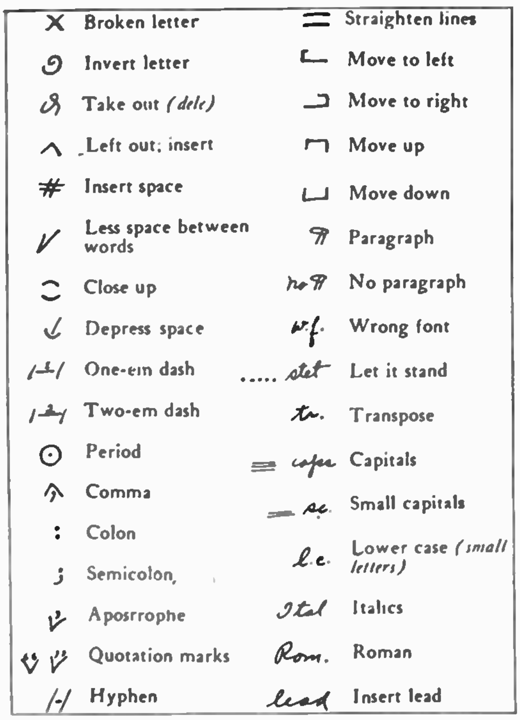
Blueprints cannot be used for purposes of reproduction although they may be used for editorial examination. It is frequently convenient to make a tracing from which the original cuts may be made and to supply the editor with a number of additional blueprints (made from this tracing) which may be used in editing the article.

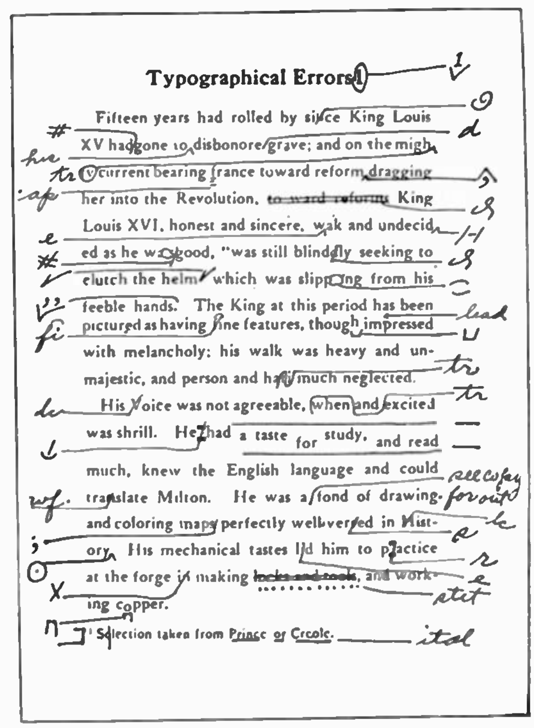
Transmitting the Article

It is advisable to send the manuscript by insured or registered mail. Insured mail enables the author to place an evaluation upon the manuscript and to collect a fee equal to this evaluation in the event the manuscript is lost. However, this is of little importance in most cases since one is hardly ever able to place a proper monetary evaluation on a manuscript. By sending the manuscript by registered mail, one is assured that everything humanly possible has been done to ascertain that the material reaches the person to whom it is directed.

Never fold any manuscript material or illustrations. If the illustrations are so large that they cannot easily be mailed flat, it is preferable to roll them up and send them through the mail in a cardboard mailing tube.

In submitting the manuscript it is advisable to accompany it by a letter of transmittal. This letter should indicate the title of the article, the author or authors, their affiliations, and to call attention to any matter which the author considers important or unusual. This letter of transmittal provides an excellent opportunity for the author to inquire about publishing procedure which may be unfamiliar to him and date of publication of his article. However, in the case of scientific or engineering organizations, where the manuscript must be read by others, it is not always possible to indicate definitely when an article may be published.





Proofreading

A manuscript which has been accepted for publication is marked for style and size of type, and is then set in type by the printer. The type is held in troughs or trays about 20 inches long and called galleys. "Galley proofs" are printed and returned for checking against the original manuscript. No new material should be added, nor should extensive changes be made in these proofs. Such changes are always expensive and wasteful, but in technical publications, the addition of new material or changes should be particularly avoided, especially if there is possibility that the article may have important patent significance.

In correcting galley proofs, the more common standard proofreaders' marks, given here, will be found useful. Corrections should be indicated on the edge or border of the galley; they should not be written in the text itself.

He who passes judgment on himself will not be judged by others. If this admonition is observed in the preparation of technical articles, it may be safely assumed that the fate of the manuscript will be a happy one.

Copper-Oxide Rectifiers in Standard Broadcast Transmitters*

R. N. HARMON†, ASSOCIATE, I.R.E.

Summary—Improvements in the processing of copper used in copper-oxide rectifiers make practical the use of this type of rectifier in modern broadcast transmitters. Features are reliability, long life, and ability to withstand surges.

THE dry-type metal rectifier most commonly known as copper oxide or Rectox is, from an operating standpoint, one of the simplest pieces of apparatus for converting alternating into direct current. It is not only a very simple design, but also easy to operate; usually with outstanding reliability. Many attempts have been made in the past to develop this type of converter into an acceptable form for converting larger blocks of power, and to put the larger types on a more competitive basis with other kinds of converters.

Recently these attempts have been successful, and by giving proper attention to all factors entering into the design of this rectifier, its field has been enlarged to include many applications formerly using hot-cathode vapor tubes. This has been particularly true of recent commercial broadcast transmitters.

Because the construction of the Rectox metal rectifier is so very simple, not many ways have been found to increase its efficiency and output. Each rectifier element consists essentially of a copper disk with a surface layer of cuprous oxide (produced by oxidizing the surface of the disk) and of a second electrode that is in contact with the layer of cuprous oxide. The stack-type rectifier unit as now made consists of a number of such elements which, separated from one another by spacers and special cooling fins, are stacked on bolts and held together by end plates.

The most important material used in the manufacture of this rectifier is the copper. A thorough investigation of the properties of copper and the unavoidable impurities which it contains has revealed that, in general, copper from which certain impurities have been removed as completely as possible to be the most appropriate grade of copper for the manufacture of these rectifiers. Of great influence on the performance of the rectifiers is also the manner in which, beginning in the copper mill and ending in the annealing furnace, the metal is worked and treated.

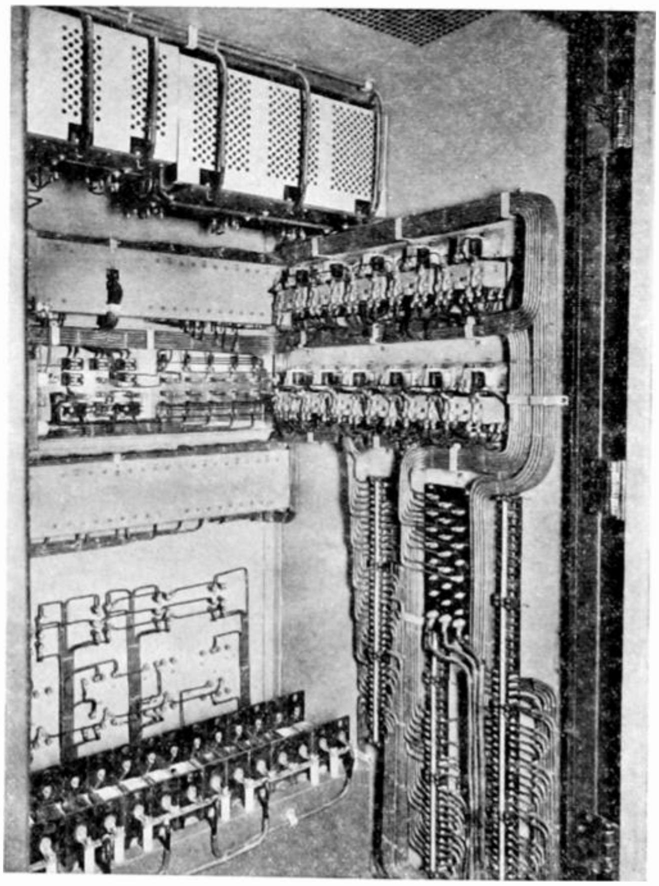


Fig. 1—Rear view of distribution cubicle for 50-kilowatt broadcast transmitter showing at bottom force-draft-cooled 3-phase full-wave 1250-volt, 0.7-ampere and 3-phase full-wave 3000-volt, 1.4-ampere rectifier for plate supplies.

By special annealing treatments, by careful choice of the copper, and by special production tests such rectifiers can be produced with great constancy, long life, and high efficiency. Further improvements in weight and size, as well as cost, have been made by forced-

* Decimal classification: R356.3. Original manuscript received by the Institute, May 13, 1942.

† Westinghouse Electric and Manufacturing Company, Baltimore, Maryland.

draft cooling which in some instances increased the output allowable more than ten times that obtainable with natural cooling.

The new forced-draft Rectox is well suited to application in broadcast transmitting apparatus. Outstanding advantages are its long life and reliability. No maintenance should be required. Even if a failure should occur, the plug-in construction permits replacement about as quickly as a tube. Its operation is instantaneous, that is, it is completely without time lag or any initial transient forming a build-up period. Hence the need for any time-delay device is eliminated, thus simplifying the control circuits.

In a standard line of 5-, 10-, and 50-kilowatt transmitters, all rectifiers except the main plate rectifiers are of this type having ratings from a few hundred volts and a quarter ampere up to 3000 volts and 1.4 amperes. Some of these units have been in operation over 15,000 hours. Nearly all of these units are forced-draft-cooled and are supplied air from the main cooling system used to cool the large radio tubes.

The Half-Wave Voltage-Doubling Rectifier Circuit*

D. L. WAIDELICH†, ASSOCIATE, I.R.E., AND C. H. GLEASON‡, ASSOCIATE, I.R.E.

Summary—*An analysis of the half-wave voltage-doubling rectifier circuit is made with the main assumption that the tube drop is zero while conducting. The performance characteristics of the circuit as predicted by the analysis are presented together with experimental verifications of several of these characteristics. Operating conditions for which polarized electrolytic condensers may be used and the currents to be expected on short circuit are discussed. The performance characteristics calculated from the analysis are presented as curves suitable for use in the prediction of the performance of an assembled circuit, and in the design of this doubler to meet specified operating conditions. A comparison is made of the performance characteristics of the half-wave and full-wave voltage doublers.*

Introduction

THE half-wave voltage doubler is being found useful as a power supply and as a component of high-voltage supplies. This circuit has several advantages over others employing input transformers. It offers economy in cost, size, and weight and hence is used in transformerless receivers. For use in radio-receiver power supplies, it has the important advantage of having a common input and output terminal.

Although no analysis of this half-wave doubler seems to have been made, several references to its operation and applications may be found.[1,2] Greinacher[3] seems to have made the first use of this circuit, employing it as the basic element of his voltage-multiplication circuit.

The results presented in this present paper on the half-wave doubler were obtained by a method of analysis similar to those employed in two previous analyses of the full-wave doubler.[4,5] The purposes of this paper are to present the results of the analysis by means of curves suitable for use in design, to compare some of the theoretical results with experimental results, and to compare the operating characteristics of the half-wave doubler with those of the full-wave doubler.

Analysis

The circuit diagram of the half-wave voltage doubler is shown in Fig. 1, and the current and voltage waveforms are shown in Fig. 2 for a complete cycle of the impressed alternating voltage e. Tube T_1 starts to conduct at $\omega t=\alpha$ and stops at $\omega t=\beta$, where $\omega/2\pi$ is the supply frequency and t is the time in seconds. Tube T_2 starts to conduct at $\omega t=\delta$ and stops at $\omega t=-90$ degrees. Condenser C_1 is charged to approximately the peak value of the alternating voltage while tube T_2 is conducting and is discharged during the rest of the cycle. The voltage of C_1 is e_c and is shown in Fig. 2. The load voltage and also the voltage on C_2 have exactly the same shape as the load current i_L flowing through the load resistance R. The tube currents i_1 and i_2 have been reduced to one fifth of their size for convenience.

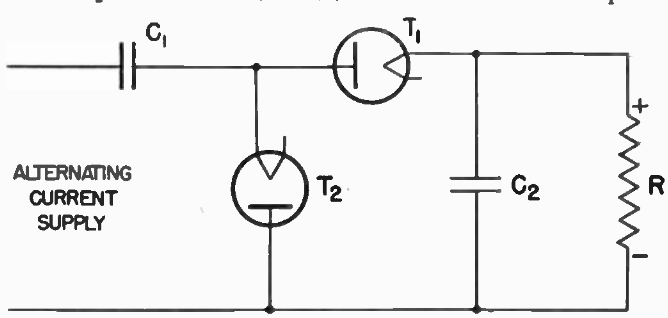


Fig. 1—Circuit diagram of the half-wave voltage-doubling rectifier.

To simplify the analysis, the following assumptions are made: (1) the applied alternating voltage is sinusoidal, and the source has no impedance; (2) when conducting the tube drop is zero, and when not conducting the tube resistance is infinite; (3) the condensers have zero power factor and are both the same size;

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‡ Westinghouse Electric and Manufacturing Company, Bloomfield, New Jersey.

[1] M. A. Honnell, "Applications of the voltage-doubler rectifier," *Communications*, vol. 20, p. 14; January, 1940.

[2] J. Millman and S. Seely, "Electronics," McGraw-Hill Book Company, New York, New York, 1941, p. 415.

[3] H. Greinacher, "Über eine Methode, Wechselstrom mittels elektrischer Ventile und Kondensatoren in hochgespannten Gleichstrom umzuwandeln," *Zeit. für Phys.*, vol. 4, pp. 195–205; February, 1921.

[4] D. L. Waidelich, "The full-wave voltage-doubling rectifier circuit," Proc. I.R.E., vol. 29, pp. 554–558; October, 1941.

[5] N. H. Roberts, "The diode as a half-wave, full-wave and voltage-doubling rectifier," *Wireless Eng.*, vol. 13, pp. 351–352; and pp. 423–430; July and August, 1936.

and (4) the load resistance has no inductance. The most serious of these assumptions is that of considering the tube drop to be zero, and this may be taken into account by an extension of this analysis.

Because the tubes act as open circuits when not conducting, the doubler circuit may be broken down into

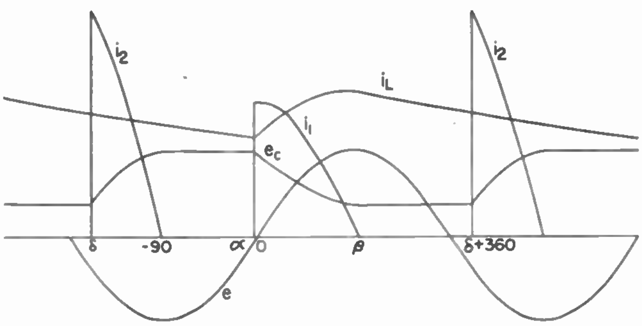

Fig. 2—Voltage and current waveforms in the half-wave voltage doubler.

three equivalent circuits which are in operation over certain portions of each cycle. While tube T_1 is conducting, from $\omega t=\alpha$ to $\omega t=\beta$, the equivalent circuit is Fig. 3(a). While neither tube is conducting, $\theta=\beta$ to $\omega t=(\delta+360$ degrees), the equivalent circuit is Fig. 3(b); and while tube T_2 is conducting, the doubler is represented by both Figs. 3(b) and 3(c). It can be shown that the tubes are never conducting at the same time, hence the equivalent circuits of Fig. 3(a) and 3(c) are never in operation simultaneously.

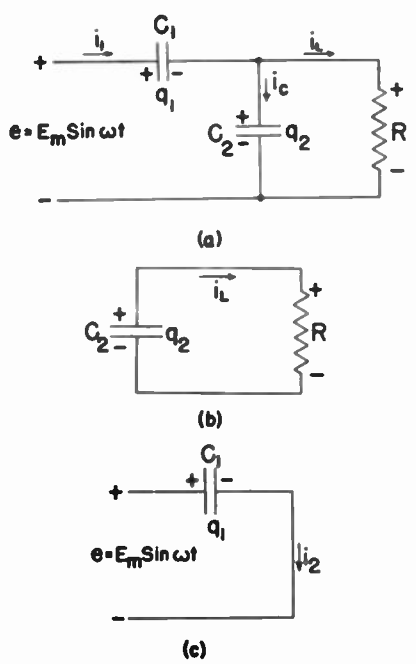

Fig. 3—(a) The equivalent circuit with tube T_1 conducting.
(b) The equivalent circuit with neither tube conducting.
(b) and (c) The equivalent circuits with tube T_2 conducting.

In spite of the simplifying assumptions made, the analysis is rather complex; therefore, only the analytical results and the more salient points of the analysis will be presented. An outline of the analysis may be found in Appendix II.

Before the voltage and current relations of the doubler can be found, the angles at which the tubes start and stop conducting must be determined. The analysis shows that all of these angles depend only upon the parameter (ωCR). In Fig. 4, for tube T_1, the angle α at which the tube starts conducting, the

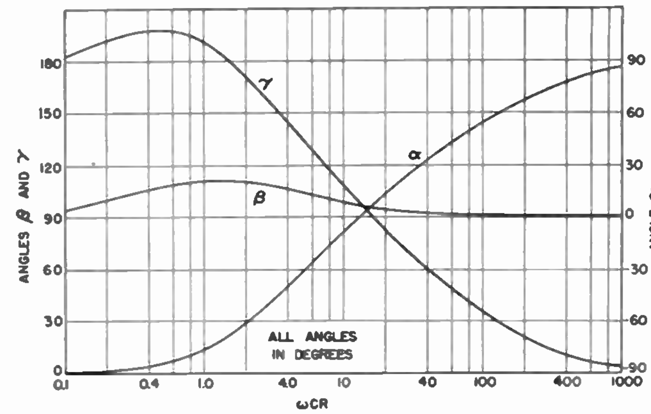

Fig. 4—The angles α, β, and γ of tube T_1.

angle β at which the tube stops conducting, and the angle γ during which the tube is conducting are shown as calculated for several values of (ωCR). For light loads (large values of ωCR) angles α and β approach 90 degrees and γ approaches zero degrees. As the load on the doubler is increased (ωCR decreasing), angle α approaches -90 degrees, and angle β rises to a maximum of 111 degrees and then decreases toward 90 degrees. The angle γ rises to a maximum of 195 degrees and then approaches 180 degrees as (ωCR) decreases toward zero.

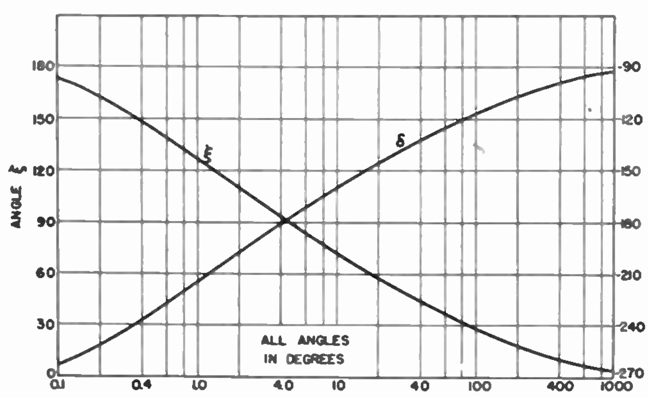

Fig. 5—The angles δ and ξ of tube T_2.

In Fig. 5 the angles of conduction for tube T_2 are also shown as functions of (ωCR) as calculated from the analysis. Angle δ at which the tube starts conducting is nearly -90 degrees for light loads and decreases toward -270 degrees as the load approaches short circuit. The angle ξ during which the tube is conducting varies from nearly zero degrees at light loads toward a limit of 180 degrees as the load is increased.

Characteristics

A most important characteristic of this circuit is the average output voltage for various loads. In Appendix II it is shown that the ratio of the average output voltage to the maximum value of the impressed alternating voltage (E_{dc}/E_m) is a function of the parameter (ωCR) alone. In Fig. 6 this ratio is shown as a function of (ωCR) by a curve calculated from the analysis. For light loads this ratio is nearly two, dropping off toward zero for very heavy loads.

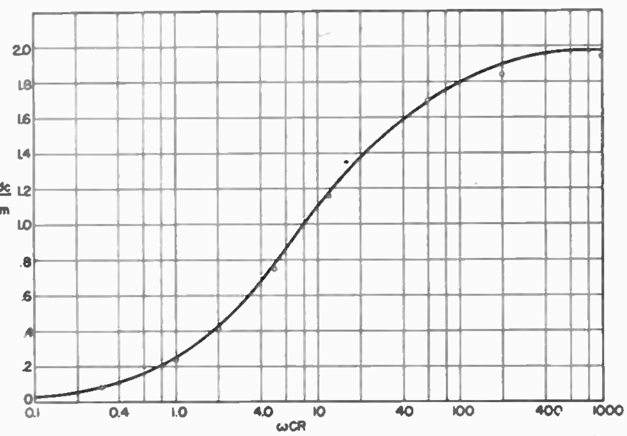

Fig. 6—The ratio of the output average voltage to the maximum value of the alternating supply voltage (E_{dc}/E_m). The circles are experimental points.

The output voltage contains some ripple voltage, with the fundamental frequency predominating. The per cent ripple r is defined as the per cent ratio of the effective ripple voltage to the average output voltage,[6] and is shown by the analysis to be solely a function of (ωCR). A curve of the per cent ripple r versus (ωCR) calculated from the analysis is shown in Fig. 7. At light loads (ωCR greater than 20) the per cent ripple is approximated by

$$r = 160/\omega CR.$$

The effect of the circuit upon the alternating-voltage supply is sometimes of interest and may be determined

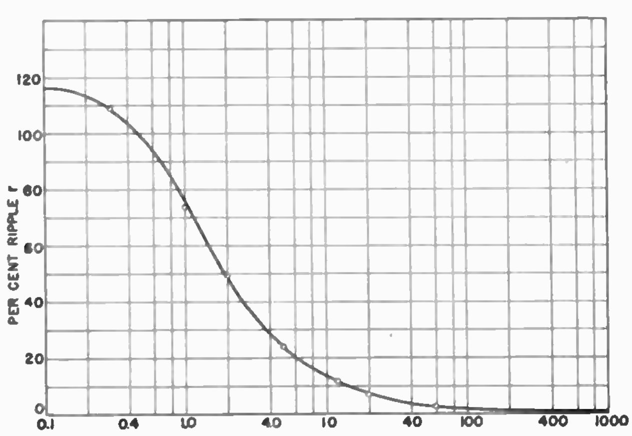

Fig. 7—The per cent ripple r. The experimental points are shown as circles.

from a knowledge of the effective current I and the input power factor. The ratio of the effective input current to the average output current (I/I_{dc}) and the power factor are also functions of (ωCR) only. Curves of these quantities calculated from the analysis are shown in Fig. 8. For light loads the tube currents are quite peaked with a correspondingly high value of (I/I_{dc}). As the load approaches short circuit, this ratio approaches 2.22. The power factor is very low for high values of (ωCR) corresponding to the very peaked currents. It rises to a maximum of 54 per cent at $\omega CR = 30$ and decreases toward zero as the load approaches

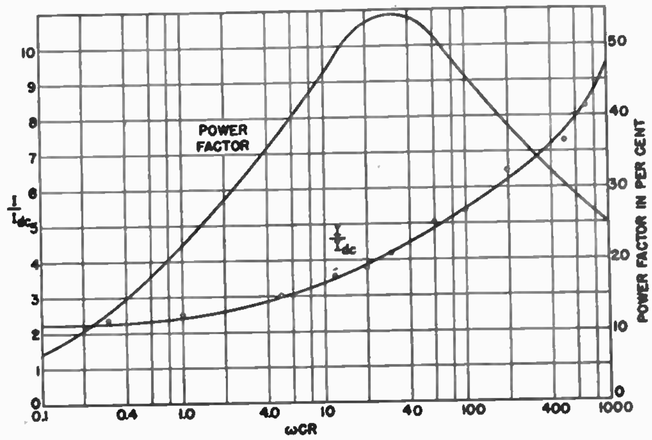

Fig. 8—The ratio of the effective input current to the average output current (I/I_{dc}) and the input power factor. The circles shown are experimental points.

short circuit, at which the doubler becomes a capacitive load on the alternating-voltage source. Examination of the doubler circuit shows that the average currents of each tube are the same and are equal to the average output current I_{dc}, which in turn may be found from Fig. 6. The ratio of the maximum tube current in tube T_1 to the average output current (i_{m1}/I_{dc}) is shown in Fig. 9 as a curve calculated from the analysis. Also shown is a similar curve for the ratio (i_{m2}/I_{dc}) for tube T_2.

The ratio of the peak inverse tube voltage to the maximum value of the applied voltage (e_p/E_m) is

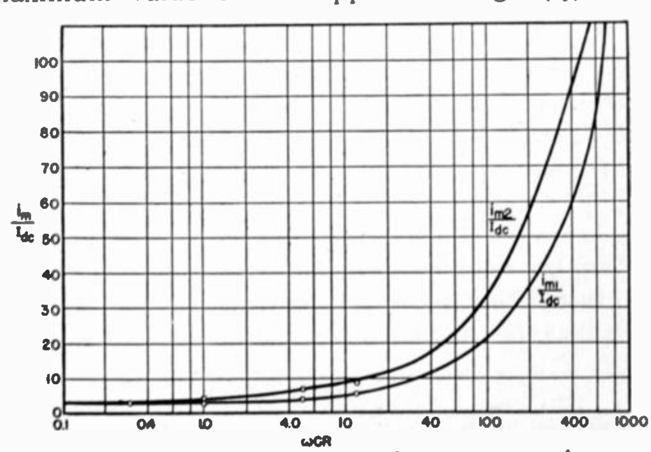

Fig. 9—The ratio of the maximum tube currents to the average output current (i_m/I_{dc}) for both tubes. The circles are experimental points.

shown in Fig. 10 for both tubes. These ratios are very nearly two at light loads and decrease toward zero as the load approaches short circuit.

As a verification of these calculated curves of the doubler's characteristics, some experimental results are shown for comparison. The ripple voltage was evaluated experimentally by measurement of the harmonic components with the aid of a wave analyzer. The

[6] Institute of Radio Engineers, "Standards on Radio Receivers, 1938," p. 6.

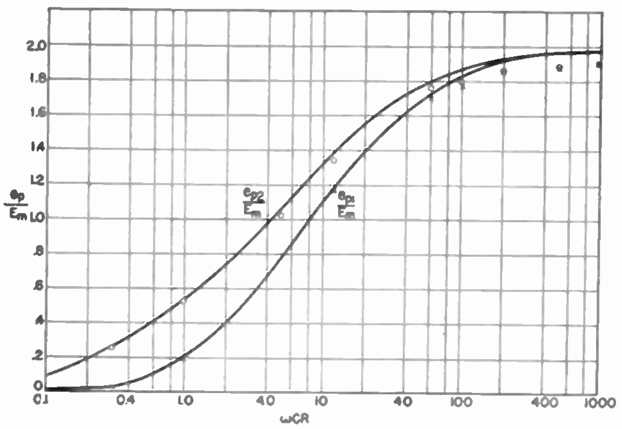

Fig. 10—The ratio of the maximum inverse tube voltage to the maximum value of the alternating supply voltage (e_p/E_m) for both tubes with experimental points shown.

maximum tube currents were measured by the use of a cathode-ray oscilloscope. The peak inverse tube voltages were measured with a peak-voltage voltmeter designed for this purpose.

Condensers

The input condenser C_1 must withstand the maximum value of the alternating-voltage supply E_m, and the load condenser C_2 must withstand a maximum voltage of $2E_m$. The analysis shows that the voltage on condenser C_1 reverses over a part of the cycle if the doubler is loaded heavily enough. This reversal occurs if the angle δ is less than -180 degrees, and from Fig. 5 this corresponds to values of (ωCR) less than 4.50. Hence a polarized electrolytic condenser may be used for C_1 only if (ωCR) is greater than 4.50. From Fig. 6 $(E_{dc}/E_m)=0.772$ for $(\omega CR)=4.50$; therefore, this value of (E_{dc}/E_m) may be used to determine whether or not this type of condenser may be used. This value of (E_{dc}/E_m) is below the usual operating values for this circuit, and hence in most practical circuits, polarized electrolytic condensers may be used without danger. In any case a polarized electrolytic condenser may be used for condenser C_2.

Operation on Short Circuit

When the doubler is short-circuited, the tubes conduct throughout alternate half cycles. The analysis shows that the effective input current on short circuit is

$$I = \omega C E_m/\sqrt{2},$$

the average output current is

$$I_{dc} = \omega C E_m/\pi,$$

and the per cent ripple is 121.2 per cent. The per cent ripple and short-circuit currents calculated from these expressions are in quite good agreement with the experimentally determined values.

Comparison of Half-Wave and Full-Wave Doublers

A comparison of the operating characteristics of the half-wave and full-wave[4] voltage doublers shows that throughout the normal operating range (ωCR greater than 10) the full-wave doubler offers a higher input power factor, lower maximum tube currents, slightly less ripple (and of higher frequency) in the output voltage, and slightly better voltage regulation; while the half-wave doubler offers lower peak inverse tube voltages, lower effective input currents, and a common input and output terminal allowing both the load and input source to be grounded if necessary.

Performance and Design

The performance of an assembled half-wave doubler may be predicted if the capacitance of the two equal condensers C_1 and C_2 and the load resistance R are known. Upon evaluation of the parameter (ωCR), the curves of Figs. 6 to 10 may be used to determine the operating characteristics of the circuit.

These curves may also be used to design a half-wave doubler to meet certain prescribed operating conditions. Often the input voltage and frequency, the output direct voltage, and the output direct current are specified. In addition, the application may restrict the per cent ripple allowable in the output voltage. Hence E_m, E_{dc}, and I_{dc} are specified together with a restriction on the per cent ripple r. From the curve of Fig. 7 the value of (ωCR) is fixed by the ratio (E_{dc}/E_m). The capacitance of the condensers may be found from (ωCR), the load resistance $R=(E_{dc}/I_{dc})$, and the supply frequency $\omega/2\pi$. The curves of Figs. 9 and 10 may be used to determine the peak inverse tube voltage and the maximum tube current, thus enabling the selection of rectifier tubes of proper inverse voltage and maximum current ratings. From Fig. 7 the per cent ripple in the output voltage may be found. If this value is greater than the per cent ripple allowable, the output voltage may be filtered, or some compromise in the specified current and voltages may be made so as to increase the value of (ωCR). The per cent ripple in the output can be materially reduced by placing a filter circuit between the output condenser and the load resistance. Insertion of a filter will cause some alteration in the operation of the doubler circuit as predicted from this analysis; however, the analysis will still afford a rather good approximation of the operating characteristics of the doubler.

Conclusions

On the basis of the experimental verifications obtained, the mathematical analysis is representative of the actual operation of the doubler circuit, and therefore, prediction of the performance of the doubler by the analysis seems justifiable. The assumption of no tube drop while conducting seems to have introduced little error, but for large tube drops an extension of this analysis would have to be used.

It has been shown that the design of this circuit and the predetermination of its performance is facilitated by the use of the results of this analysis.

Appendix I

Nomenclature

$\omega/2\pi$ = frequency of the alternating-voltage supply
t = time in seconds
$\theta = \omega t$
E_m = the maximum value of the alternating-voltage supply
e = the instantaneous value of the alternating-voltage supply
i_1 = the current flowing through tube T_1
i_2 = the current flowing through tube T_2
i = the input current, i_1+i_2
i_L = the current flowing through the load resistance R
i_c = the current through condenser C_2
e_c = the instantaneous voltage on condenser C_1
q_1 = the charge on condenser C_1
q_2 = the charge on condenser C_2
R = the load resistance in ohms
C = capacitance in farads of condensers C_1 or C_2
α = angle in radians at which the tube T_1 starts to conduct current
β = angle in radians at which the tube T_1 stops conducting
γ = angle in radians during which tube T_1 is conducting, $\beta-\alpha$
δ = angle in radians at which the tube T_2 starts to conduct current
ξ = angle in radians during which the tube T_2 is conducting, $-(\delta+\pi/2)$
$Z=\sqrt{4R^2+(1/\omega C)^2}$
$\lambda_1=\tan^{-1}(\omega CR),\ 0\leqq\lambda_1\leqq(\pi/2)$
$\lambda_2=\tan^{1}(2\omega CR),\ 0\leqq\lambda_2\leqq(\pi/2)$
i_α = current i_L at angle α
i_β = current i_L at angle β
$e_{c\beta}$ = voltage e_c at angle β
e_1 = instantaneous inverse voltage on the tube T_1
e_2 = instantaneous inverse voltage on the tube T_2
e_{p1} = peak inverse voltage on tube T_1
e_{p2} = peak inverse voltage on tube T_2
θ_p = angle at which the peak inverse voltage occurs on the tube T_2
i_{m1} = the maximum current through tube T_1
i_{m2} = the maximum current through tube T_2
θ_m = angle at which the maximum current through tube T_1 occurs
E = the effective value of the alternating-voltage supply
I = effective value of the input current
E_{dc} = the average value of the load voltage
I_{dc} = the average value of the load current
E_0 = the effective value of the load voltage
I_0 = the effective value of the load current
E_{0a} = the effective value of the alternating component of the load voltage
I_{0a} = the effective value of the alternating component of the load current
r = the per cent ripple
P = the average input power

Appendix II

Analysis

(a). Angles at Which the Tubes Start and Stop Conducting

From angle α to angle β the equivalent circuit of Fig. 3(a) is in operation. The initial conditions at angle α are substituted in the solutions of the circuit equations to give

$$i_c/C_2 = \omega E_m \cos \omega t - i_1/C_1, \tag{1}$$

$$i_\alpha R = E_m \sin \alpha + E_m, \tag{2}$$

$$i_L = \frac{E_m}{Z}\cos(\theta-\lambda_2) + \left[i_\alpha - \frac{E_m}{Z}\cos(\alpha-\lambda_2)\right]\exp[-(\theta-\alpha)\cot\lambda_2]. \tag{3}$$

At angle β, $i_L=i_\beta$; substituting in (3),

$$i_\beta = \frac{E_m}{Z}\cos(\beta-\lambda_2) + \left[i_\alpha - \frac{E_m}{Z}\cos(\alpha-\lambda_2)\right]\exp[-(\beta-\alpha)\cot\lambda_2]. \tag{4}$$

From angle β to angle (α+360 degrees) the equivalent circuit of Fig. 3(b) is in operation. For this circuit, $i_L=i_\beta$ at angle β, and

$$i_L = i_\beta \exp[-(\theta-\beta)\cot\lambda_1]. \tag{5}$$

At angle β, $i_\beta=i_L=-i_c$, and by (1),

$$i_\beta = -\omega C E_m \cos\beta. \tag{6}$$

At angle (α+360 degrees), $i_L=i_\alpha$ because of the steady-state conditions imposed, hence

$$i_\alpha = i_\beta \exp\{[\beta-(\alpha+2\pi)]\cot\lambda_1\}. \tag{7}$$

Equations (2), (4), (6), and (7) form a system of equations in the unknowns i_α, i_β, α, and β, and by the substitution of $\gamma=\beta-\alpha$ and after some manipulation, may be reduced to

$$A_1\sin\beta + B_1\cos\beta = 1 \tag{8}$$

$$A_2\sin\beta + B_2\cos\beta = 1 \tag{9}$$

where

$$A_1 = -\frac{\sin^2\lambda_2}{2}\exp(\gamma\cot\lambda_2) + \left[\frac{\sin\lambda_2}{2}\sin(\gamma+\lambda_2) - \cos\gamma\right]$$

$$B_1 = -\left[\omega CR + \frac{\sin\lambda_2\cos\lambda_2}{2}\right]\exp(\gamma\cot\lambda_2) + \sin\gamma + \frac{\sin\lambda_2}{2}\cos(\gamma+\lambda_2)$$

$$A_2 = -\cos\gamma$$

$$B_2 = \sin\gamma - \omega CR\exp[-(2\pi-\gamma)\cot\lambda_1].$$

From the relations $\sin^2\beta+\cos^2\beta=1$, (8) and (9) may be combined to give

$$(A_1 - A_2)^2 + (B_2 - B_1)^2 - (A_1B_2 - A_2B_1)^2 = 0. \quad (10)$$

These coefficients A_1, A_2, B_1, and B_2 are functions of (ωCR) and the angle γ; hence γ is defined implicitly in (10) as a function of (ωCR). For assigned values of (ωCR), the corresponding values of γ may be determined by several methods.[7,8] For a specific value of (ωCR), γ may be found from (10), angle β from (8) and (9), and $\alpha=\beta-\gamma$. Therefore, angles α, β, and γ are functions only of (ωCR).

The angle δ at which tube T_2 starts to conduct is determined by use of the fact that e_c at angle δ is equal to e_c at angle β, and from (1) is

$$e_{c\beta} = E_m(\sin\beta + \omega CR\cos\beta) = E_m\sin\delta, \quad (11)$$

and

$$\delta = \sin^{-1}(\sin\beta + \omega CR\cos\beta). \quad (12)$$

Because tube T_2 is conducting from $\theta=\delta$ to $\theta=-\pi/2$, its interval of conduction ξ is

$$\xi = -(\pi/2 + \delta). \quad (13)$$

(b). Average Load Voltage

The average load current I_{dc} is

$$I_{dc} = \frac{1}{2\pi}\int_\alpha^{\alpha+2\pi} i_L d\theta. \quad (14)$$

From $\theta=\alpha$ to $\theta=\beta$ the load current is expressed by (3), and from $\theta=\beta$ to $\theta=\alpha+2\pi$ the load current is given by (5). With the substitution of these expressions, (14) becomes, upon integration,

$$I_{dc} = \frac{\omega CE_m}{2\pi}\left[1 + \frac{\sin(\beta+\lambda_1)}{\cos\lambda_1}\right], \quad (15)$$

and from (15) the ratio of the average load voltage to the maximum value of the alternating voltage (E_{dc}/E_m) is

$$E_{dc}/E_m = I_{dc}R/E_m = \frac{\omega CR}{2\pi}\left[1 + \frac{\sin(\beta+\lambda_1)}{\cos\lambda_1}\right]. \quad (16)$$

This ratio is a function only of (ωCR).

(c). Per Cent Ripple

The effective value E_{0a} of the alternating component of the output voltage may be found from the effective value of the output voltage E_0 and the average output voltage E_{dc} by use of the relation

$$E_{0a} = \sqrt{E_0^2 - E_{dc}^2}. \quad (17)$$

The per cent ripple is then

$$r = \frac{E_{0a}}{E_{dc}} \times 100 = \frac{\sqrt{E_0^2 - E_{dc}^2}}{E_{dc}} \times 100. \quad (18)$$

[7] D. L. Waidelich, "The numerical solution of equations," *Elec. Eng.*, vol. 60, pp. 480–481; October, 1941.
[8] I. T. Whitaker and G. Robinson, "The Calculus of Observations," Blackie and Son, Ltd., London, England, 1924, pp. 78–95.

The effective output voltage E_0 is evaluated by means of the effective output current I_0, which in turn may be evaluated from

$$I_0 = \sqrt{\frac{1}{2\pi}\int_\alpha^{\alpha+2\pi} i_L^2 d\theta}. \quad (19)$$

With the substitution of (3) and (5), (19) becomes upon integration

$$2\pi\left(\frac{RI_0}{E_m}\right)^2 = \tfrac{1}{4}\sin^2\lambda_2\left\{\frac{\gamma}{2} + \tfrac{1}{4}[\sin 2(\beta-\lambda_2) - \sin 2(\alpha-\lambda_2)]\right\}$$
$$+\sin^2\lambda_2\{\cos\beta[\omega CR\cos\beta + \tfrac{1}{2}\sin\lambda_2\cos(\beta-\lambda_2)]$$
$$+\cos\alpha[(\sin\alpha+1) - \tfrac{1}{2}\sin\lambda_2\cos(\alpha-\lambda_2)]\}$$
$$+\omega CR\{-[\omega CR\cos\beta + \tfrac{1}{2}\sin\lambda_2\cos(\beta-\lambda_2)]^2$$
$$+[(\sin\alpha+1) - \tfrac{1}{2}\sin\lambda_2\cos(\alpha-\lambda_2)]^2\}$$
$$-\frac{\omega CR}{2}(\sin\alpha+1)^2 + \frac{(\omega CR)^3}{2}\cos^2\beta. \quad (20)$$

The per cent ripple r can then be found from

$$r = \frac{\sqrt{(RI_0/E_m)^2 - (E_{dc}/E_m)^2}}{E_{dc}/E_m} \times 100 \quad (21)$$

where (E_{dc}/E_m) may be obtained from (16).

(d). Effective Input Current and Power Factor

The effective input current I is evaluated from

$$I = \sqrt{\frac{1}{2\pi}\int_\delta^{\delta+2\pi} i^2 d\theta}. \quad (22)$$

From angle α to angle β the input current i is i_1, which from Fig. 3(a) is

$$i_1 = i_L + R\frac{di_L}{dt}, \quad (23)$$

and by (3)

$$i_1 = \omega CE_m\left\{\cos\theta + \frac{\sin\lambda_2}{2}\sin(\theta-\lambda_2) - \cot\lambda_2\left[\frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2) - \sin\alpha - 1\right]\exp[-(\theta-\alpha)\cot\lambda_2]\right\}. \quad (24)$$

From angle β to angle δ the input current is zero. From δ until $\theta=-\pi/2$ the input current is i_2, which from Fig. 3(c) is

$$i_2 = \omega CE_m\cos\theta. \quad (25)$$

With the substitution of (24) and (25) into (22), the effective current may be found from (26).

The elements of the doubler circuit, with the exception of the load resistance, are assumed to take no average power. Therefore, the power input to the doubler is the same as that taken by the load resistance, which is

$$P = I_0^2R. \quad (27)$$

This product I_0^2R is evaluated in (20). The power input may also be found from

$$2\pi\left(\frac{RI}{\omega CRE_m}\right)^2 = -\frac{\pi}{4} - \frac{\delta}{2} - \tfrac{1}{4}\sin 2\delta + \tfrac{1}{2}[\gamma + \tfrac{1}{2}(\sin 2\beta - \sin 2\alpha)]$$
$$+ \sin\lambda_2\cos\lambda_2\left[\frac{\sin^2\beta - \sin^2\alpha}{2}\right] - \frac{\sin^2\lambda_2}{2}[\gamma + \tfrac{1}{2}(\sin 2\beta - \sin 2\alpha)]$$
$$- 2\sin\lambda_2\cos\lambda_2\left\{[-\cot\lambda_2\cos\beta + \sin\beta]\left[\omega CR\cos\beta + \frac{\sin\lambda_2}{2}\cos(\beta-\lambda_2)\right]\right.$$
$$\left. + \left[\frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2) - \sin\alpha - 1\right][\cot\lambda_2\cos\alpha - \sin\alpha]\right\}$$
$$+ \tfrac{1}{8}\sin^2\lambda_2\{\gamma + \tfrac{1}{2}[\sin 2(\alpha-\lambda_2) - \sin 2(\beta-\lambda_2)]\}$$
$$- \sin^2\lambda_2\cos\lambda_2\left\{[\tfrac{1}{2}\sin\lambda_2\cos(\alpha-\lambda_2) - \sin\alpha - 1][\cot\lambda_2\sin(\alpha-\lambda_2) + \cos(\alpha-\lambda_2)]\right.$$
$$\left. - \left[\omega CR\cos\beta + \frac{\sin\lambda_2}{2}\cos(\beta-\lambda_2)\right][\cot\lambda_2\sin(\beta-\lambda_2) + \cos(\beta-\lambda_2)]\right\}$$
$$+ \frac{\cot\lambda_2}{2}\left\{\left[\frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2) - \sin\alpha - 1\right]^2 - \left[\omega CR\cos\beta + \frac{\sin\lambda_2}{2}\cos(\beta-\lambda_2)\right]^2\right\}. \quad (26)$$

$$P = \frac{1}{2\pi}\int_{\delta}^{\delta+2\pi} ei\,d\theta. \quad (28)$$

The input power factor is, from (27),

$$\frac{I_0^2R}{EI} \times 100. \quad (29)$$

(e). Peak Inverse Tube Voltages

1. Tube T_1

The ratio of the peak inverse voltage on tube T_1 to the maximum value of the supply voltage (e_{p1}/E_m) is

$$\frac{e_{p1}}{E_m} = \omega CR\cos\beta\exp[(\beta - \delta - 2\pi)\cot\lambda_1]. \quad (30)$$

2. Tube T_2

The ratio of the peak inverse voltage on tube T_2 to the maximum value of the supply voltage (e_{p2}/E_m) is given by

$$\frac{e_{p2}}{E_m} = \frac{\sin\lambda_2}{2}\cos(\theta_p - \lambda_2)$$
$$+\left[\sin\alpha + 1 - \frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2)\right]\exp[(-\theta_p - \alpha)\cot\lambda_2] \quad (31)$$

where θ_p is the angle at which the peak inverse voltage occurs and may be obtained from

$$0 = \frac{\sin\lambda_2}{2}\sin(\theta_p - \lambda_2)$$
$$+\cot\lambda_2\left[\sin\alpha + 1 - \frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2)\right]$$
$$\exp[(-\theta_{p-\alpha})\cot\lambda_2]. \quad (32)$$

For values of (ωCR) greater than 10, angle θ_p is very nearly equal to β and a good approximation is given by

$$\frac{e_{p2}}{E_m} = -\omega CR\cos\beta. \quad (33)$$

(f). Maximum Tube Currents

1. Tube T_1

For values of (ωCR) greater than 13.2 the maximum current through tube T_1 is

$$i_{m1} = \omega CE_m\left\{\cos\alpha + \frac{\sin\lambda_2}{2}\sin(\alpha - \lambda_2)\right.$$
$$\left. - \cot\lambda_2\left[\frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2) - \sin\alpha - 1\right]\right\}. \quad (34)$$

For (ωCR) less than 13.2 the maximum tube current is given by

$$i_1 = \omega CE_m\left\{\cos\theta_m + \frac{\sin\lambda_2}{2}\sin(\theta_m - \lambda_2)\right.$$
$$-\cot\lambda_2\left[\frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2) - \sin\alpha - 1\right]$$
$$\left.\exp[-(\theta_{m-\alpha})\cot\lambda_2]\right\}, \quad (35)$$

where θ_m is the angle at which i_1 is a maximum and may be found from

$$0 = \sin\theta_m + \frac{\sin\lambda_2}{2}\cos(\theta_m - \lambda_2)$$
$$+\cot^2\lambda_2\left[\sin\alpha + 1 - \frac{\sin\lambda_2}{2}\cos(\alpha-\lambda_2)\right]$$
$$\exp[-(\theta_{m-\alpha})\cot\lambda_2]. \quad (36)$$

2. Tube T_2

For values of (ωCR) greater than 4.50 the maximum current through tube T_2 is

$$i_{m2} = -\omega CE_m\cos\delta. \quad (37)$$

For (ωCR) less than 4.50 the maximum tube current is given by

$$i_{m2} = -\omega CE_m. \quad (38)$$

Some Characteristics of a Stable Negative Resistance*

CLEDO BRUNETTI†, ASSOCIATE, I.R.E., AND LEIGHTON GREENOUGH‡, NONMEMBER, I.R.E.

Summary—*By employing positive feedback in a two-stage amplifier it is possible to obtain an input impedance which is equal to the impedance in the feedback circuit multiplied by a negative constant. For a resistance-capacitance feedback circuit the input impedance becomes a negative resistance in series with a negative capacitance. With the proper choice of circuit constants, the negative impedance can be made to approximate closely a pure negative resistance over any given frequency range. In the apparatus described, high stability is secured by the use of inverse feedback in addition to the positive-feedback loop.*

THERE is a need for a source of negative resistance that is independent of frequency and the normal variations in tubes and supply voltages.[1] Such a device would find application in the design of oscillators,[2] the improvement of the Q of tuned parallel circuits,[3] the measurement of circuit impedances at low and high frequencies,[4] the design of special networks,[5] and other uses.

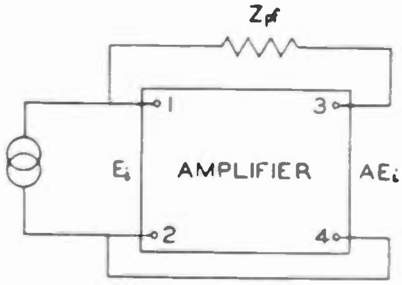

Fig. 1—Method of obtaining negative resistance.

The possibility of realizing such a negative resistance has already been pointed out in the literature.[3] If, in an ordinary amplifier, part of the output is coupled back to the input, it is possible to adjust the (positive) feedback and amplification so that the input impedance of the amplifier becomes a pure negative resisttance whose magnitude may be varied by changing either the positive feedback or the amplification. In fact, as pointed out by Crisson,[6] it is possible by this method to obtain an input impedance that is the negative counterpart of whatever impedance is used in the positive-feedback circuit. If the amplification is made substantially independent of tube and supply voltage variations by the employment of inverse feedback, the resulting negative resistance will also be found to be independent of these quantities approximately to the same degree. It is the purpose of this paper to describe the characteristics of such a negative resistance and to point out both its advantages and limitations.

Consider the circuit of Fig. 1 where an impedance Z_{pf} is connected between the input and output terminals 1 and 3, respectively, of the two-stage amplifier. The equivalent circuit (assuming the impedance of the grid circuit of the first stage of the amplifier to be infinitely large) is shown in Fig. 2. If E_i is the input voltage and A the amplification, the resulting input current will be

$$I_i = \frac{E_i - AE_i}{Z_{pf}} \tag{1}$$

from which the input impedance becomes

$$Z_n = \frac{E_i}{I_i} = \frac{Z_{pf}}{1 - A}. \tag{2}$$

Thus, if A is real (i.e., has no phase shift) and is greater than unity, Z_n will be equal to Z_{pf} multiplied by a negative number. Z_n may be varied by changing either Z_{pf} or A, or both. In practice it is found best to vary both Z_{pf} and A, the former providing large-step variations in Z_n and the latter serving as a fine control. By keeping A low, optimum stability is obtained, provided, however, A does not approach unity, in which case slight variations in the magnitude of A would be magnified in Z_n.

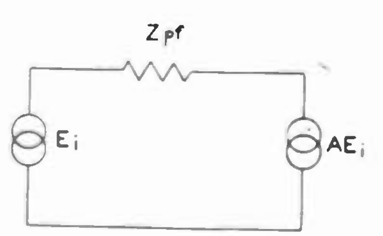

Fig. 2—Theoretical representation of amplifier as source of negative resistance.

Amplifier

The dependence of the magnitude of the negative impedance on the amplification may be found by differentiating (2) with respect to the real quantity A assuming Z_{pf} to remain constant. This yields

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[1] L. C. Verman, "Negative circuit constants," Proc. I.R.E., vol. 19, pp. 676–681; April, 1931.

[2] Cledo Brunetti, "The transitron oscillator," Proc. I.R.E., vol. 27, pp. 88–90; February, 1939.

[3] F. E. Terman, R. R. Buss, W. R. Hewlett, and F. C. Cahill, "Some applications of negative feedback with particular reference to laboratory equipment," Proc. I.R.E., vol. 27, pp. 649–655; October, 1939.

[4] H. Iinuma, "A method of measuring the radio-frequency resistance of an oscillatory circuit," Proc. I.R.E., vol. 18, pp. 537–543; March, 1930.

[5] S. P. Chakravarti, "The band-pass effect," *Wireless Eng.*, vol. 18, pp. 103–111; March, 1941.

[6] G. Crisson, "Negative impedances and the twin 21-type repeater," *Bell Sys. Tech. Jour.*, vol. 10, pp. 485–513; July, 1931.

$$\frac{dZ_n}{dA} = Z_n \frac{1}{1 - A} \qquad (3)$$

or

$$\frac{dZ_n}{Z_n} = \frac{dA}{1 - A} \cdot \qquad (4)$$

Equation (4) shows that, for A greater than unity, a given percentage variation in A will result in a larger percentage variation in Z_n. As A approaches unity the ratio of these variations increases without limit. This emphasizes the necessity of maintaining a constant amplification in the amplifier.

The circuit used in these measurements is shown in Fig. 3. It is strictly a conventional amplifier with simple positive and negative feedback applied to it as pointed out by Terman and others.[3] Amplification control is available by applying a variable amount of inverse feedback from the output back to the input and is accomplished with the variable resistor R_{nf}. This procedure yields good stability with respect to changes in supply voltages and frequency, as

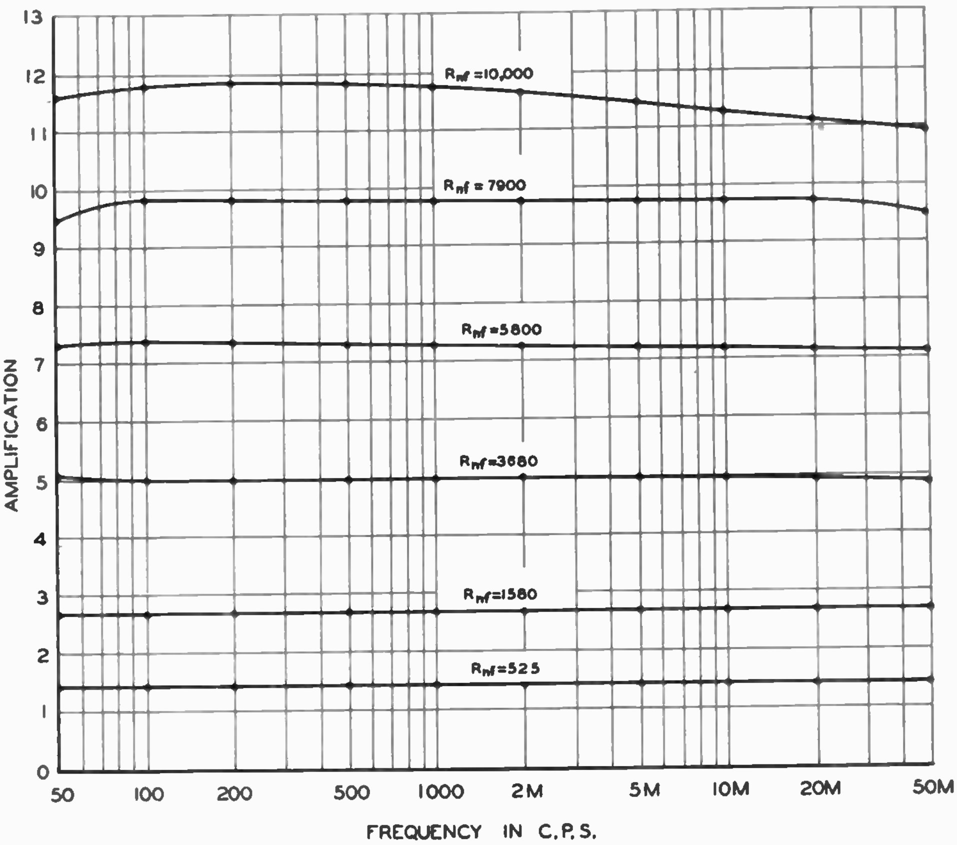

Fig. 4—Frequency response of amplifier for six different amplification settings. Load impedance is 2500 ohms.

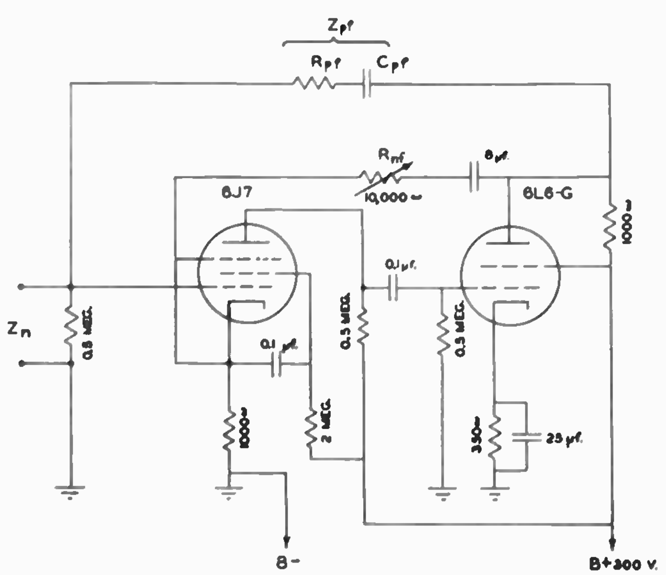

Fig. 3—Circuit for producing negative resistance.

well as a low effective internal impedance in the output stage of the amplifier. A substantial amount of inverse feedback is employed so that low amplification and high stability result.

The amplification as a function of frequency for various values of R_{nf} is shown in Fig. 4. These measurements were made with a load impedance of 2500 ohms on the amplifier. In Fig. 5 is shown how the amplification may be controlled by varying R_{nf}. For an

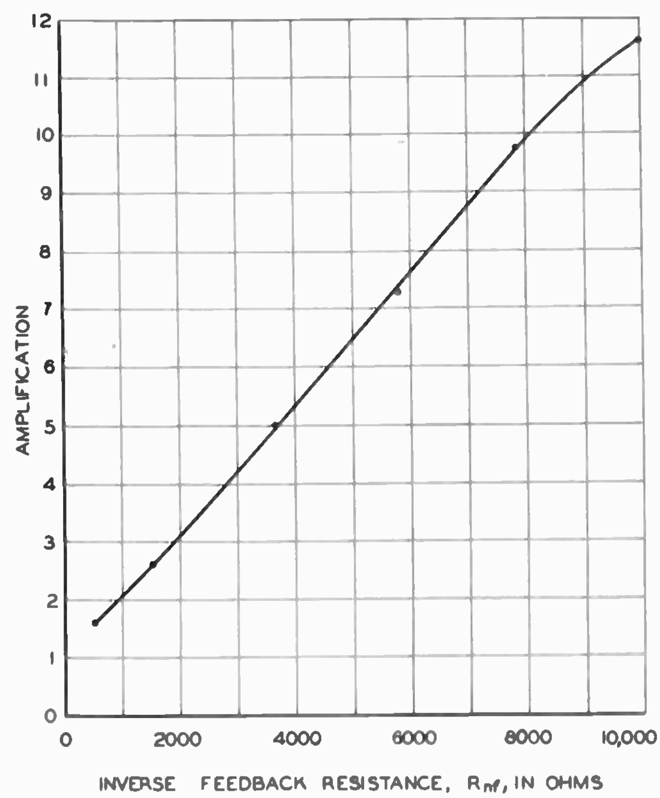

Fig. 5—Variation of amplification with inverse-feedback resistance R_{nf}.

amplification approaching a value of 10, the actual amplification variation is less than $2\frac{1}{2}$ per cent from 200 to 30,000 cycles per second. Below a value of 8 the variation is less than $2\frac{1}{2}$ per cent from 100 to 50,000 cycles.

Measurement of Negative Resistance

A simple circuit for obtaining the magnitude of negative resistance is to connect it in parallel with a

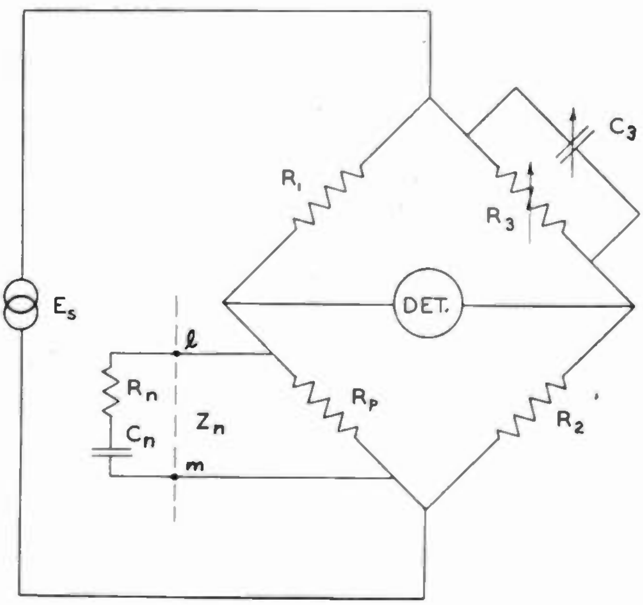

Fig. 6—Modified Maxwell bridge used for exact determination of Z_n. $R_1=R_2=500$ ohms.

known positive resistance R_p and to measure the equivalent resistance of the combination on a Wheatstone bridge. From this R_n may easily be computed. R_n is the actual input resistance of the amplifier, and will have a negative sign. In this method, it is necessary that R_p be less in absolute value than R_n for otherwise natural oscillations will take place unless arrangements are made to keep the impedance of the external measuring circuit (which is in parallel with the combination) low.

This method of measurement does not take into account the phase shift resulting from the presence of the blocking condenser C_{pf}. Although usually made as large as possible, C_{pf} will still possess some reactance at low frequencies. A better circuit for precise determination of the complex impedance, and the one employed in these measurements, is the modified Maxwell bridge shown in Fig. 6. The negative impedance to be measured is connected to the left of terminals $l-m$ and is represented by R_n and C_n in series. If $Z_x=R_x+j\omega L_x$ is the equivalent series impedance of the parallel combination of R_n, C_n, and R_p, then, for balance, we have $R_x=R_1R_2/R_3$ and $L_x=R_1R_2C_3$. Using these quantities one may then compute the resistive and reactive components of the negative impedance Z_n from the equation

$$Z_n = -\frac{Z_xR_p}{Z_x - R_p}. \tag{5}$$

If Z_{pf} is a pure resistance of magnitude R_{pf}, and for zero phase shift in the amplifier, Z_n will be a pure negative resistance whose magnitude depends on both R_{pf} and the negative-feedback resistance R_{nf}. This dependence is shown in Fig. 7 where the negative impedance is plotted against R_{pf} for four values of R_{nf}. These curves show the wide range of negative impedance that may be obtained by this method simply by varying R_{pf}. The curvature at the lower portion of the curves is the result of the reactance of the blocking condenser and the variation of the amplification for low values of R_{pf}.

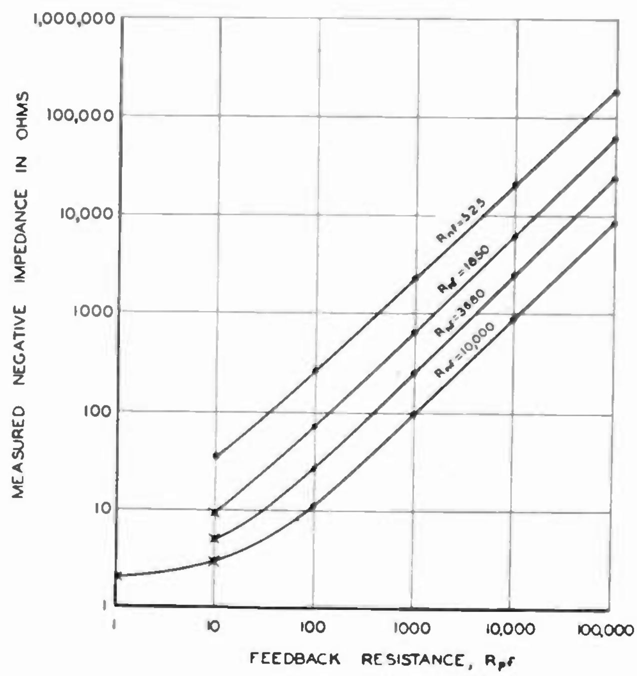

Fig. 7—Negative impedance as a function of positive-feedback resistance R_{pf} at 1000 cycles.

The ordinates of Fig. 7 represent the absolute magnitude of the negative impedance given by (2) which in general differs very little from the projection of the impedance on the negative real axis. The blocking condenser C_{pf}, as already noted, introduces some phase shift. However, with the exception of the points marked with a cross in Fig. 7, the capacitive reactance is negligible and the negative impedance and its negative-resistance component are for all practical purposes equal. For these points the impedance, measured at 1000 cycles, was accompanied by a phase shift of at most 5 to 15 degrees from a pure negative resistance. This phase shift may be reduced further if desired by increasing the size of the blocking condenser C_{pf}.

Some phase shift is also introduced in the amplifier at high frequencies by the shunting tube capacitance, and at the lower frequencies by the interstage coupling condenser.

The variation of negative resistance with frequency from 100 to 5000 cycles is shown in Fig. 8. Here also the absolute magnitude of the negative impedance is plotted for convenience. In the three upper curves for which R_{pf} equals 1000, 10,000, and 100,000 ohms, respectively, the phase angle is less than 1 degree. For the curve for which R_{pf} equals 500 ohms the largest phase angle is four degrees at 100 cycles, while for the lower curve taken with R_{pf} equal to 50 ohms a maximum phase angle of 25 degrees occurs at 100 cycles.

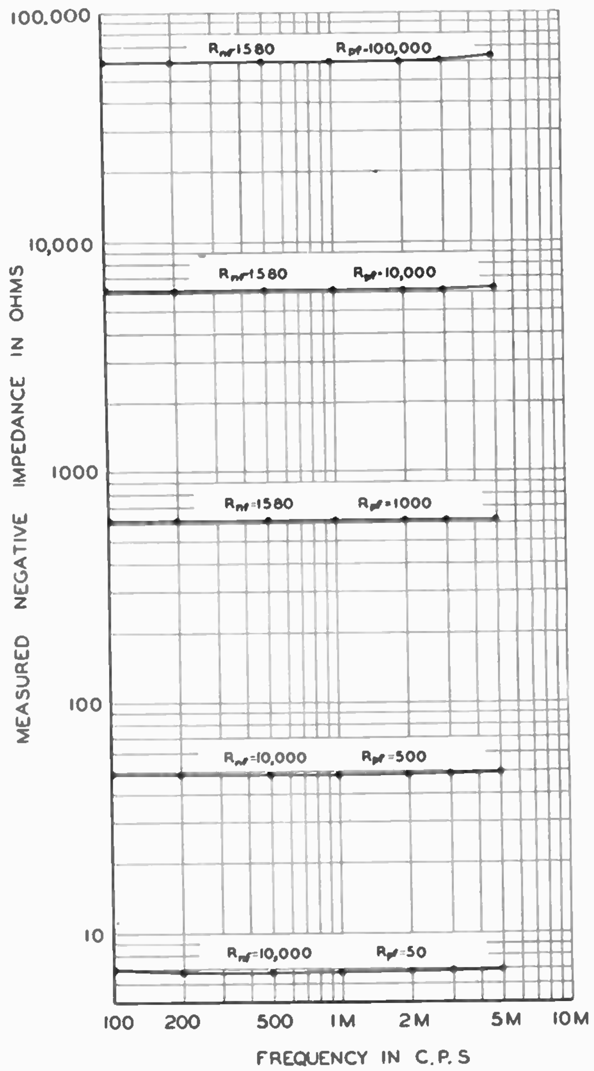

Fig. 8—Variation of negative impedance with frequency, for five representative cases.

The dependence of the negative resistance upon supply voltages is shown in Fig. 9. Here per cent changes are plotted rather than the actual value of negative resistance in order to emphasize the variations with supply voltages. These measurements were made at 1000 cycles. As is apparent, the large values of negative resistance are in general the most stable both in respect to variations in frequency as well as supply voltages.

Discussion

Low values of negative resistance may be obtained either by using low values of resistance R_{pf} in the positive-feedback circuit or by employing high amplification. Of these methods both result in unwanted phase changes produced by the blocking condenser C_{pf}. Raising the amplification is more desirable from the standpoint of keeping the phase angle of Z_n low as a higher value of R_{pf} may then be used to obtain a specified R_n. In this manner, the phase angle, being approximately proportional to the ratio of reactance in the positive-feedback circuit to the resistance in the same circuit, will be reduced. At low frequencies the reactance of C_{pf} may become appreciable. However, for most practical purposes where R_n is greater than ten ohms, the phase angle is negligible, even at a frequency of 100 cycles.

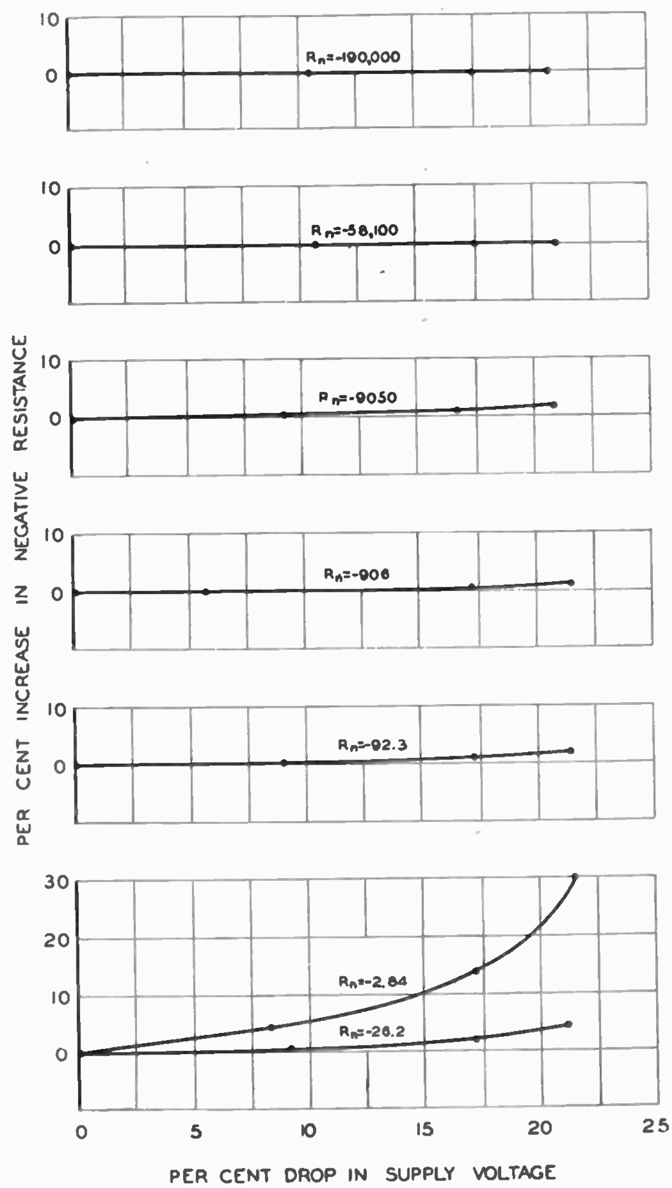

Fig. 9—Effect of supply-voltage variation on negative resistance.

A limitation in obtaining very low values of R_n is the effective internal impedance of the amplifier at its output terminals. This impedance (usually a resistance) must be included in Z_{pf}, as shown in Fig. 10. The use of sufficient inverse feedback will reduce Z_{int}, but the amplification is then also reduced. This makes it necessary to decrease R_{pf} to obtain the desired value of negative resistance. Z_{int} may vary from 5 to 20 ohms in the circuit of Fig. 3.

It should be kept in mind that R_{pf} effectively constitutes the load on the amplifier. If a low value of R_n is obtained by using a low R_{pf}, the amplification will be somewhat less stable since less voltage is available at the output terminals for feedback, hence stabilization is not as good. Distortion, also, is greater.

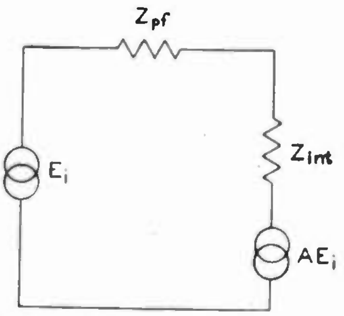

Fig. 10—Equivalent circuit including internal impedance of amplifier.

If a direct-current potential across the negative resistance is not objectionable, it is possible to modify the circuit so as to eliminate the phase angle introduced by C_{pf} as shown in Fig. 11. In this circuit the blocking condenser is placed in series with the high impedance of the grid circuit where it can do little harm.

As with all electrical elements, there is an upper limit to the alternating voltage which may be applied to the negative resistance when used as a circuit component. Because of the presence of the inverse feedback in the amplifier, alternating voltages of the order of 3 to 30 volts root-mean-square, depending upon the amplification used, may be applied to the device without introducing an appreciable amount of harmonic content in the output.

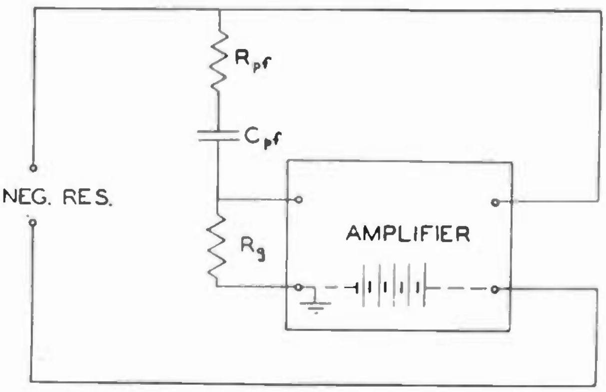

Fig. 11—Circuit to reduce effect of blocking-condenser reactance in series with R_{pf}.

Summarizing, it can be stated that with this apparatus it is possible to obtain values of negative resistance from 1 ohm up to an unlimited value. The circuit, unlike the dynatron, is not dependent upon secondary emission for its operation. Stability with respect to frequency and supply voltages is good.

Thermal-Frequency-Drift Compensation*

T. R. W. BUSHBY†, ASSOCIATE, I.R.E.

Summary—*The conditions necessary for minimizing frequency drift with variation of ambient temperature are examined for various types of circuits. In fixed-tuned circuits the drift can be eliminated by a comparatively simple adjustment of the temperature coefficient of capacitance. In variable-tuned circuits, expressions for coefficient adjustment resulting in minimum integrated drift are given, together with simpler expressions resulting in an approximate minimum. It is shown to be practicable in some instances to design circuits in which the drift is better than ten parts per million per degree centigrade, over normally used frequency ranges, by complementary adjustment of the temperature coefficients of inductance and capacitance. Frequency drift in superheterodyne receiver circuits is discussed, and it is found that the local-oscillator circuit is peculiarly adaptable to drift correction by reason of the complex nature of its capacitance network. Such padded circuits can be very effectively corrected by simple adjustment of the various capacitive coefficients, the drift factor being tracked in a manner analogous to the frequency tracking. The padding of variable-capacitance-tuned circuits for the express purpose of drift correction presents the simplest means of minimizing thermal drift in such circuits. The necessary expressions are given, together with the results of some experimental work.*

I. Introduction

IN THE parallel resonant circuit of Fig. 1, the frequency stability is dependent, among other things, on the temperature of the various circuit components.

For instance, when such a circuit is used to control the frequency of a valve generator, during the first few minutes after switching on the valve, there is a frequency change caused mainly by capacitance increase, which is due to the expansion of the valve electrodes.

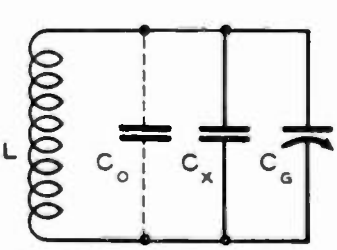

Fig. 1—Variable-capacitance-tuned circuit having thermal-drift-compensating condenser C_x.

Methods for reducing such drift have already been discussed[1,2] and it is not the purpose of this paper to deal further with it.

After a sufficient lapse of time at a constant ambient temperature, the frequency becomes stable, provided that other factors introduce no drift. Subsequently, changes in ambient temperature will cause changes in

* Decimal classification: R381×536. Original manuscript received by the Institute, April 7, 1942.

† Amalgamated Wireless, Australasia, Ltd., Ashfield, N. S.W., Australia.

[1] M. L. Levy, *Electronics*, vol. 12, p. 15; 1939.
[2] Gramophone Company, *Electronics and Television and Short-Wave World*, vol. 13, p. 407; 1940.

circuit inductance and capacitance, resulting in further frequency drift.

It is usual to express the temperature coefficients of capacitance and inductance as a number of parts per million per degree centigrade, referred to a temperature of 20 degrees centigrade, so that a coil having a temperature coefficient of +100 increases its inductance by 100 parts in a million, when raised in temperature from 20 to 21 degrees centigrade.

Ordinarily, the reference point may be arbitrarily selected with negligible error, but for extremes of temperature and/or large coefficients, the reference temperature of 20 degrees centigrade should be adhered to.

In Fig. 1, let $\alpha=(dC/dT)(10^6/C)$ be the numerical value of the temperature coefficient of the total capacitance C (where $C=C_0+C_s+C_x$), and $\beta=(dL/dT)(10^6/L)$, the corresponding coefficient for the inductance.

The resonant frequency at the reference temperature is $f=1/2\pi\sqrt{LC}$, from which it may be shown that the frequency drift due to changing temperature is given by

$$\theta = -0.5(\alpha+\beta) \tag{1}$$

where $\theta=(df/dT)(10^6/f)$, the temperature coefficient of frequency, in parts per million per degree centigrade.

II. Fixed-Tuned Circuits

In a fixed-tuned circuit $\theta=0$ when $\alpha+\beta=0$, and the resonant frequency will then be independent of temperature.

Inductances have, in general, a positive coefficient, so that the total capacitance must usually have a negative coefficient of the same numerical value, in order that there may be no frequency drift with temperature.

The availability of ceramic materials having dielectric constants which vary inversely with temperature, has made possible the manufacture of condensers having negative temperature coefficients. The capacitance of such condensers decreases with increasing temperature, and in this respect they differ from almost all other circuit components.

As it is unlikely that a condenser having the exact numerical value of coefficient will be available, it will be necessary to use a parallel combination of two condensers, one of positive coefficient, and one of negative coefficient, to obtain the required result. Therefore, let $C=C_1+C_2\cdots+C_n$, and let $\alpha_1, \alpha_2\cdots, \alpha_n$ be the respective temperature coefficients. Then α_c, the temperature coefficient of C, is given by

$$\alpha_C = \frac{\alpha_1\cdot C_1+\alpha_2\cdot C_2\cdots+\alpha_n\cdot C_n}{C}. \tag{2}$$

To take a fairly general practical case, let

C_0 = self-capacitance of the inductance
C_x = drift-compensating condenser
C_p = portion of C_x having α positive
C_n = portion of C_x having α negative
C_t = capacitance of the trimmer condenser
C_s = stray capacitance
α_0, α_p etc. = respective temperature coefficients
β = temperature coefficient of inductance
θ = temperature coefficient of frequency

All temperature coefficients are in parts per million per degree centigrade, and all capacitances are in micromicrofarads.

As the self-capacitance is inseparable from the inductance, it is not possible to measure either β or α_0 separately. The observed value β' for the temperature coefficient of inductance is

$$\beta' = \beta + \frac{\alpha_0\cdot C_0}{C} \tag{3}$$

where C is the total capacitance of the test circuit.

Since β varies with frequency, it must be measured at the frequency at which the inductance is to be used; therefore the test circuit will necessarily have the same total capacitance as the circuit undergoing analysis. Then θ will be zero when

$$\beta' = -\left[\frac{\alpha_p\cdot C_p+\alpha_n\cdot C_n+\alpha_t\cdot C_t+\alpha_s\cdot C_s}{C}\right] \tag{4}$$

which is equivalent to saying that α_0 automatically becomes part of β', and can be disregarded, provided that C_0 is not omitted from the total circuit capacitance C.

The apparatus used for the measurement of individual component coefficients is substantially that described by Leonard.[3] The component under test is placed in a small temperature chamber and connected to a two-terminal oscillator of good short-period stability.

This oscillator is initially set to beat with a substandard crystal-controlled oscillator, and the beatnote observed with changing temperature of the component. An ordinary receiver, cathode-ray oscilloscope, and beat-frequency oscillator is used to measure the beat frequency.

Then, for an inductance, $\beta=-2\theta$, and for a capacitance, $\alpha=-2C\theta/C_c$, where C_c is the capacitance of the condenser under test, and C is the total circuit capacitance of the test oscillator.

Complete assemblies are checked for thermal drift by being placed in a larger temperature chamber, and loosely coupled to the receiver. With the latter at some feet distant, deliberate coupling is usually unnecessary.

As will be seen throughout this paper, a knowledge of individual coefficients is not, in general, necessary for

[3] S. C. Leonard, *Electronics*, vol. 11, p. 18; 1938.

drift compensation, but it is useful for checking purposes and for *ab initio* design.

All components except C_s can be dissociated from the circuit for individual measurement, but the value of α_s can only be determined indirectly. The method used is to place in the circuit temporarily a condenser C_z of known temperature coefficient α_z and check the thermal drift for the complete circuit. α_s is then determined from (5).

$$\alpha_s = -\left[\frac{C(2\theta + \beta') + \alpha_z \cdot C_z + \alpha_t \cdot C_t}{C_s}\right]. \quad (5)$$

Then for $\theta = 0$

$$\alpha_{z'} = -\left(\frac{C\beta' + \alpha_s \cdot C_s + \alpha_t \cdot C_t}{C_z}\right) \quad (6)$$

where $\alpha_{z'}$ is the value of α_z for $\theta = 0$. Combining (5) and (6) to eliminate α_s gives

$$\alpha_{z'} = \frac{2C \cdot \theta}{C_z} + \alpha_z \quad (7)$$

and from (7) and (2)

$$C_p = \frac{2C \cdot \theta + C_z(\alpha_{z'} - \alpha_n)}{\alpha_p - \alpha_n}. \quad (8)$$

Only two heat runs are necessary, one for the complete circuit and one for C_z.

Therefore, in fixed-tuned circuits, thermal frequency drift may be eliminated by a simple adjustment of the temperature coefficient of a portion of the capacitance, subject to a realizable value for $\alpha_{z'}$. When the figure found for $\alpha_{z'}$ is not realizable, some circuit change must be made, either in electrical values or in temperature coefficients.

III. Variable-Capacitance-Tuned Circuits

When the circuit has to be designed to cover a range of frequencies, exact compensation at all points within the range is not possible, since the temperature coefficient of a variable condenser varies with setting, while the temperature coefficient of inductance alters with frequency.

The following notation applies to the circuit of Fig. 1.

f_L = lowest resonant frequency
f_H = highest resonant frequency
θ = temperature coefficient of frequency
$\theta_L = \theta$ at f_L in a test heat run of the complete circuit
$\theta_{L'} = \theta$ at f_L after circuit correction
$\theta_H = \theta$ at f_H in a test heat run of the complete circuit
$\theta_{H'} = \theta$ at f_H after circuit correction
C_L = total circuit capacitance at f_L
C_H = total circuit capacitance at f_H
C_z = capacitance of the drift-compensating condenser
α_z = temperature coefficient of C_z in the test heat runs
$\alpha_{z'}$ = temperature coefficient of C_z after circuit correction
C_{max} = capacitance of C_g at f_L
α_{max} = temperature coefficient of C_{max}
C_{min} = capacitance of C_g at f_H
α_{min} = temperature coefficient of C_{min}
β_L = temperature coefficient of inductance at f_L
β_H = temperature coefficient of inductance at f_H
$k = C_L/C_H$

All temperature coefficients are in parts per million per degree centigrade, and all capacitances are in micromicrofarads.

If we make the simplifying assumption that θ varies linearly with frequency, it can be shown that it will have its minimum value, integrated throughout the range, when $\theta_{L'}^2 + \theta_{H'}^2$ is a minimum, and that $\theta_{L'}$ and $\theta_{H'}$ will then have opposite signs. That is, the drift will be in opposite directions at the frequency extremes, and will therefore be zero at some frequency within the range. The assumption of linearity has been found to be sufficiently true for all practical purposes, except when the integrated drift is very low.

From (1) and (2),

$$\theta_{L'}^2 + \theta_{H'}^2 = \left(\beta_L + \frac{\alpha_{z'} \cdot C_z + \alpha_{max} \cdot C_{max}}{C_L}\right)^2 + \left(\beta_H + \frac{\alpha_{z'} \cdot C_z + \alpha_{min} \cdot C_{min}}{C_H}\right)^2 \quad (9)$$

which has its minimum value when

$$\alpha_{z'} = \frac{2C_L(\theta_L + k \cdot \theta_H)}{C_z(k^2 + 1)} + \alpha_z \quad (10)$$

which reduces to (7) for the fixed-tuned circuit. When $\alpha_{z'}$ has its optimum value, then

$$\theta_{L'} = \frac{k^2 \cdot \theta_L - k \cdot \theta_H}{k^2 + 1} \quad (10a)$$

and

$$\theta_{H'} = \frac{\theta_H - k \cdot \theta_L}{k^2 + 1}. \quad (10b)$$

$\theta_{H'}$ will usually have a small positive value and $\theta_{L'}$ a larger negative one, which is a desirable condition. Therefore, a close approximation to optimum drift is had by putting $\theta_{H'} = 0$ either experimentally, or by (10c), where

$$\alpha_{z'} = \frac{2C_H \cdot \theta_H}{C_z} + \alpha_z \quad (10c)$$

which is the same expression as (7). When $\alpha_{z'}$ has this value, then

$$\theta_{L'} = \theta_L - \theta_H/k. \qquad (10d)$$

It will be noted that the terms of the expressions are such that the correction possibilities can be fully explored after the initial experimental work is done. The latter consists only of three heat runs to determine θ_L, θ_H, and α_z, and three capacitance measurements for C_L, C_H, and C_z.

The drift at other frequencies within the range will usually be less than at the extremes, depending upon the linearity of the curve. The deviation may amount to about ten parts per million, so that when the corrected drift is of this order, the assumption of linearity is no longer justified, and the final correction is best made experimentally. This necessitates taking some heat runs at intermediate frequencies, unless the individual coefficients are known for such frequencies.

In the latter case, the drift throughout the range can be determined graphically, by plotting the effect of each circuit component individually and summing them as shown in Fig. 2, which represents a practical example where the approximate values are as in Table I.

TABLE I

$C_L = 360\mu\mu f$	$C_{max} = 310\mu\mu f$	$\beta_L = +20$
$C_H = 90\mu\mu f$	$\alpha_{max} = +130$	$\beta_H = +35$
$C_z = 50\mu\mu f$	$C_{min} = 40\mu\mu f$	$\theta_L = -70$
$\alpha_z = +40$	$\alpha_{min} = +250$	$\theta_H = -85$

α_z' is found to be -300, when θ_L' becomes -45 and θ_H' $+10$.

In the figure, curve A shows the effect on the frequency drift of the sum of the coefficients of the various capacitances, plotted against resonant frequency. The individual curves are omitted in the interests of clarity. Curve B is a similar plot for the capacitance, after the inclusion of the required negative-coefficient condenser for C_z, and curve C is the final curve, including the coil coefficient, being the algebraic sum of -0.5α and -0.5β.

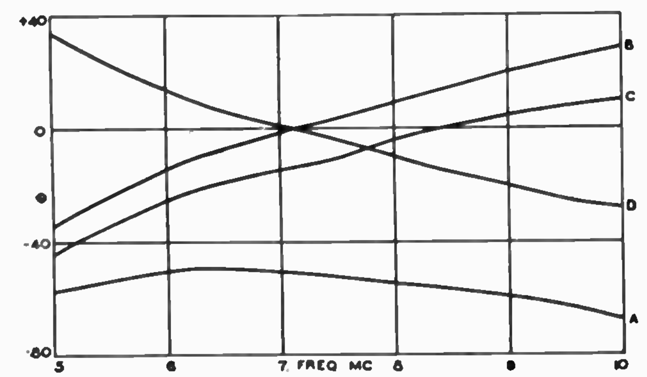

Fig. 2—Illustrating a practical example of optimum drift correction in a circuit using stock broadcast receiver parts.

From the final curve it is seen that $\theta_{L'}$ is -45 and $\theta_{H'}$ is $+10$, as already found, and in addition, the drift at intermediate frequencies is readily observed. The circuit used for Fig. 2 is fairly typical of those encountered when using average grade components designed for commercial receiver use.

Thus a considerable reduction in drift, at all frequencies, is possible by a comparatively simple adjustment of the temperature coefficient of a portion of the fixed capacitance. No further improvement can be had unless some change is made in the temperature coefficient of inductance.

IV. Low-Drift Variable-Tuned Circuits

Referring again to Fig. 2, in which curve B represents the drift due to the capacitance coefficient, it is seen that a coil having a coefficient such that -0.5β is represented by curve D, will result in zero drift at all frequencies.

Zero drift in variable-tuned circuits therefore depends entirely on the practical possibilities of so arranging the individual coefficients, that the algebraic sum is zero for all required frequencies. A coil is required having coefficients at the frequency extremes which are related by

$$\beta_L = \frac{C_H \cdot \beta_H - (\alpha_{max} \cdot C_{max} - \alpha_{min} \cdot C_{min})}{C_L} \qquad (11)$$

which is derived from (9) when $\theta_{L'} = \theta_{H'} = 0$.

In some cases, the design of inductances which will meet the requirements to a sufficient degree of accuracy, is not unduly difficult, since it is fortunate that, in general, β increases with increasing frequency, while α can be made to decrease with increasing frequency, by adjustment of the temperature coefficient of a portion of the capacitance.

A practical example (Figs. 1 and 3) will illustrate this. The circuit constants were approximately as in Table II.

TABLE II

$C_L = 192\mu\mu f$	$\alpha_z = 0$	$C_{min} = 36\mu\mu f$
$C_H = 48\mu\mu f$	$C_{max} = 180\mu\mu f$	$\alpha_{min} = +165$
$C_z = 12\mu\mu f$	$\alpha_{max} = +65$	

Fig. 3 is constructed in the same manner as Fig. 2. A is the drift attributable to the total capacitance before correction, and B the drift due to capacitance after correction. In this case, the correction is arbitrary, as the coil design is not considered until the capacitance drift has been plotted. A number of B curves can be laid out, according to the available choice of

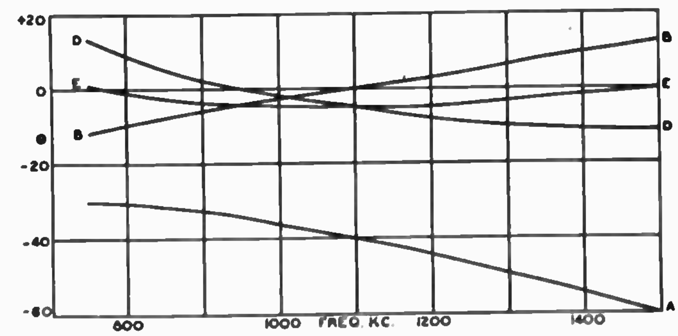

Fig. 3—Illustrating a practical example of drift correction by complementary adjustment of the temperature coefficients of inductance and fixed capacitance.

temperature coefficients for C_x. Only the one finally chosen is shown. It corresponds to $\alpha_{x'} = -600$.

The mirror images of the B curves about the zero θ axis will then provide design data for a number of possible coils, any one of which can be used, with its appropriate value for $\alpha_{x'}$, to provide a variable-frequency zero-drift circuit. Curve D gives the characteristic of the chosen coil, and E shows the drift for the whole circuit throughout the range.

The latter curve is plotted from the known coefficients of the various components. Experimental checks showed the drift throughout the range to be very small.

The inductance in this instance consisted of four litz pies on a bakelized-paper former, and it had zero temperature coefficient at about 950 kilocycles. The negative coefficient at the lower frequencies is ascribed mainly to the axial expansion of the former, whereby the pies are moved further apart with increasing temperature.[4] Other components of the coefficient become increasingly significant at higher frequencies, and the effect of the expansion of the former is more than offset by them.[5]

For the example given, it is obvious that the drift throughout the tuning range will be due entirely to practical considerations of coefficient tolerances in the various components.

The necessity for adjusting the temperature coefficient of inductance may be avoided by padding the circuit, as is done in the case of the superheterodyne local-oscillator. These possibilities are explored in Section VII, which deals with the latter circuits.

V. Variable-Inductance-Tuned Circuits

Although (10) is arrived at on the basis of a variable-capacitance circuit, it is seen to be general and applies equally well to the case of the variable-inductance-tuned circuit.

Where the method of varying the inductance does not alter the self-capacitance of the inductance, or the stray capacitance, (10) reduces to

$$\alpha_{x'} = \frac{C(\theta_L + \theta_H)}{C_x} + \alpha_x \tag{12}$$

whence

$$\theta_{L'} = -0.5(\theta_H - \theta_L) \tag{12a}$$

and

$$\theta_{H'} = 0.5(\theta_H - \theta_L). \tag{12b}$$

Usually C_0 (and also possibly C_s) does vary with inductance variation, but if the change is small in relation to the total capacitance, the error involved in using (12) instead of (10) will also be small.

The correction procedure of Section IV cannot be applied to the variable-inductance-tuned circuit, since α is constant throughout the range; therefore, for $\theta = 0$ at all frequencies, β must also be constant with varying frequency, which is not possible. The best correction for such circuits is obtained when (10) or (12) is applied, after the variation of β is reduced to a minimum. This is seen from (11), which becomes $\beta_L = \beta_H$.

A method exists for rendering such circuits effectively driftfree, by a combination of capacitances of different coefficients, connected to different points on the inductance.[6]

Variable-inductance tuning has a very useful application to circuits which are required to have a minimum of drift with "warm-up" period. The total capacitance in the circuit can be fixed at a sufficiently high value to reduce the short-period drift to negligible proportions, avoiding the necessity for special measures to overcome such drift.[7]

VI. The Superheterodyne Receiver

A superheterodyne receiver is correctly aligned when the frequency of the local oscillator is equal to the sum of the intermediate and signal frequencies, provided that the usual practice of having the oscillator frequency higher than the signal frequency is employed.

Changes in one or more of the involved frequencies will result in attenuation or loss of the signal. Assuming that the signal frequency remains constant, and disregarding the possibilities of automatic frequency control of the local oscillator, circuit drift of the various receiver channels will be separately considered.

Drift in the intermediate-frequency channel results in reduction of the signal amplitude subsequent to conversion, the attenuation being dependent on the amount of drift, and the selectivity of the channel. Being fixed-tuned circuits, they are readily corrected by the method given in Section II.

Designing the intermediate-frequency channel so as to pass a band of frequencies without attenuation, permits some drift in either the signal or oscillator frequencies, or both, without effect on the signal strength, and is a useful means of allowing for the residual drift, after other means of circuit correction have been exhausted.

Drift in the signal-frequency circuits attenuates the incoming signal prior to conversion. As in the case of the intermediate-frequency circuits, the degree of attenuation is dependent on the amount of drift, and the circuit selectivity. Though often neglected in practice, the effect can be quite serious with highly selective preconverter circuits. Methods for reducing this drift have also been discussed.

VII. Padded Circuits

The local-oscillator circuit deserves special attention, since the capacitance network is complicated by the

[4] W. H. F. Griffiths, *Wireless Eng. and Exp. Wireless*, vol. 11, p. 305; 1934.
[5] Tj. Douma, *Philips Transmitting News*, vol. 5, p. 20; 1938.
[6] Australian Patent Application No. 3070/41.
[7] P. Ware, *Electronics*, vol. 10, 12, p. 22; 1937.

necessity for frequency tracking with the signal-frequency circuits, the circuit being usually as shown in Fig. 4, to which circuit the following notation applies:

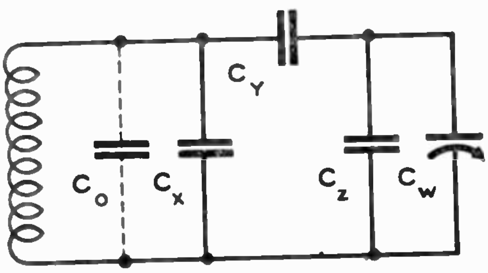

Fig. 4—Padded circuit as used in the superheterodyne local oscillator. C_y and C_z or C_y and C_z may be temperature compensated to give the circuit a very low drift factor over a range of frequencies.

f_L = lowest resonant frequency
f_H = highest resonant frequency
θ = temperature coefficient of frequency
$\theta_L = \theta$ at f_L in a test heat run of the complete circuit
$\theta_{L'} = \theta_L$ after circuit correction
$\theta_H = \theta$ at f_H in a test heat run of the complete circuit
$\theta_{H'} = \theta_H$ after circuit correction
β = temperature coefficient of inductance
C_0 = self-capacitance of the inductance L
$C_g = C_w + C_z$
C_{max} = capacitance of C_g at f_L
C_{min} = capacitance of C_g at f_H

$$C_L = C_0 + C_x + \frac{C_y \cdot C_{max}}{C_y + C_{max}}$$

$$C_H = C_0 + C_x + \frac{C_y \cdot C_{min}}{C_y + C_{min}}$$

α_z, α_y, etc. = respective temperature coefficients during test heat runs
$\alpha_{z'}$ $\alpha_{y'}$, etc. = respective temperature coefficients after circuit correction

$$a = \frac{C_{max}^2}{(C_y + C_{max})^2} \qquad b = \frac{C_{min}^2}{(C_y + C_{min})^2}$$

$$k = C_L/C_H = (f_H/f_L)^2 \qquad p = a/b$$

$$m = \frac{C_y^2}{(C_y + C_{max})^2} \qquad n = \frac{C_y^2}{(C_y + C_{min})^2}$$

All temperature coefficients are in parts per million per degree centigrade, and all capacitances are in micromicrofarads.

The coefficient α_C of a series combination of two condensers, C_y having coefficient α_y, and C_g having coefficient α_g can be shown to be

$$\alpha_C = \frac{\alpha_y \cdot C_g + \alpha_g \cdot C_y}{C_y + C_g}. \tag{13}$$

Therefore, θ for this circuit is given by

$$\theta = -0.5\left[\beta + \frac{C_z \cdot \alpha_z}{C} + \frac{C_g^2 \cdot C_y \cdot \alpha_y + C_y^2(\alpha_w \cdot C_w + \alpha_z \cdot C_z)}{C(C_g + C_y)^2}\right]. \tag{14}$$

The determination of θ is possible when all of the individual coefficients are known, but as in the previous case, the coefficient of the stray capacitance cannot be measured directly, therefore experimental heat runs of the complete circuit are essential.

This circuit opens up some very interesting possibilities for temperature compensation. In Fig. 1, there are only three possible variables, the temperature coefficients of inductance and of variable and fixed capacitance. The adjustment of the coefficients of the inductance or the variable capacitance is a comparatively difficult matter, and if these are eliminated from consideration, only C_x remains, so that (10a) and (10b) give the optimum correction figures.

But in Fig. 4, different effects are had according to which particular coefficient is adjusted. The possible variables are now five in number, and they can be operated on either singly, or in any convenient combination.

Eliminating the coefficients of the inductance and variable capacitance, as in the previous instance, still leaves the three condensers C_x, C_y, and C_z available for correction purposes. Considering first the use of these condensers individually, the necessary information for C_x alone is given by (10). Expressions (15) and (16), respectively, have been developed using C_y and C_z alone.

Equation (15) is obtained by setting up expressions for $\theta_{L'}$ and $\theta_{H'}$ from (14), summing their squares, and finding a value for $\alpha_{y'}$ which makes the sum a minimum. This gives

$$\alpha_{y'} = \frac{2C_L(a \cdot \theta_L + k \cdot b \cdot \theta_H)}{C_y[a^2 + (kb)^2]} + \alpha_y. \tag{15}$$

When $\alpha_{y'}$ has its optimum value, then

$$\theta_{L'} = \frac{k^2 \cdot \theta_L - k \cdot p \cdot \theta_H}{p^2 + k^2} \tag{15a}$$

and

$$\theta_{H'} = \frac{p^2 \cdot \theta_H - k \cdot p \cdot \theta_L}{p^2 + k^2}. \tag{15b}$$

The effect of the padder on the temperature coefficient is usually such that $\theta_{L'}$ has a small positive value and $\theta_{H'}$ a larger negative value, so that it is not as suitable as C_x or C_z for compensation purposes by itself. A close approximation to minimum integrated drift results when $\theta_{L'}$ is made zero, either experimentally or by means of (15c) where

$$\alpha_{y'} = \frac{2C_L \cdot \theta_L}{a \cdot C_y} + \alpha_y \tag{15c}$$

then

$$\theta_{H'} = \theta_H - k\cdot\theta_L/p. \tag{15d}$$

Equation (16) is obtained similarly to (15) and gives

$$\alpha_{z'} = \frac{2C_L(m\cdot\theta_L + k\cdot n\cdot\theta_H)}{C_z[m^2 + (kn)^2]} + \alpha_z \tag{16}$$

when

$$\theta_{L'} = \frac{(kn)^2\theta_L - k\cdot m\cdot n\cdot\theta_H}{m^2 + (k_n)^2} \tag{16a}$$

and

$$\theta_{H'} = \frac{m^2\cdot\theta_H - k\cdot m\cdot n\cdot\theta_L}{m^2 + (kn)^2}. \tag{16b}$$

As in the case of the adjustment of $\alpha_{x'}$, a close approximation to minimum integrated drift results from putting $\theta_{H'} = 0$, when

$$\alpha_{z'} = \frac{2C_H\cdot\theta_H}{n\cdot C_z} + \alpha_z \tag{16c}$$

and

$$\theta_{L'} = \theta_L - m\cdot\theta_H/k\cdot n. \tag{16d}$$

VIII. Temperature-Coefficient Tracking

Either C_x, C_y, or C_z can be used to obtain some measure of drift reduction, but a combination of C_x and C_y (expression (17)), or C_z and C_y (expression (18)) can be so adjusted as to give a very low coefficient throughout a range of frequencies. Some work of this nature has already been carried out in a specific instance.[8]

Equations (17) and (17a) are obtained by equating the expressions for $\theta_{L'}$ and $\theta_{H'}$ to zero, when

$$\alpha_{y'} = \frac{2C_H(k\cdot\theta_L - \theta_H)}{C_y(a - b)} + \alpha_y \tag{17}$$

and

$$\alpha_{x'} = \frac{2C_H(p\cdot\theta_H - k\cdot\theta_L)}{C_x(p - 1)} + \alpha_x. \tag{17a}$$

Subject to realizable values for $\alpha_{y'}$ and $\alpha_{x'}$, the drift can be zero at both extremes of the range, and will depart from zero only in so far as the assumption of linearity of θ versus frequency is unjustified. In practice, deviations of about ten parts per million may be expected. If the departure is too great, putting $\theta = 0$ at two suitable frequencies within the range, instead of at the extremes, results in further improvement.

In normal circuits, if realizable values for $\alpha_{y'}$ and $\alpha_{x'}$ are not obtained, it will be found that some of the individual coefficients are unnecessarily high. Best results usually accrue when the "uncorrected" drift has been reduced to a minimum by careful selection and design, so that individual components have low coefficients, although there are often limits to such a procedure.

For instance, in the adjustment of the temperature coefficient of an inductance, zero or a very low value is obtained only by cancellation of certain components of the coefficient against others,[5] and this practice may be no better, and is often worse than correcting for the original inductance coefficient by adjustment of the capacitance coefficient. A cyclic and reproducible coefficient is to be preferred to one which, though lower, is erratic.

Equation (18) is obtained in similar fashion to (17) is and gives

$$\alpha_{z'} = \frac{2C_H}{C_z(C_{\max}^2 - C_{\min}^2)}\left(\frac{C_{\max}^2\cdot\theta_H}{n} - \frac{k\cdot C_{\min}^2\cdot\theta_L}{m}\right) + \alpha_z \tag{18}$$

and

$$\alpha_{y'} = \frac{2C_H\cdot C_y}{C_{\max}^2 - C_{\min}^2}\left(\frac{k\cdot\theta_L}{m} - \frac{\theta_H}{n}\right) + \alpha_y. \tag{18a}$$

Practical examples of the use of these expressions are shown in Fig. 5, the lower curves giving the drift prior to correction, and the upper ones the most probable value of the corrected drift. The lower curves have been assumed linear, as observations were made only at the frequency extremes.

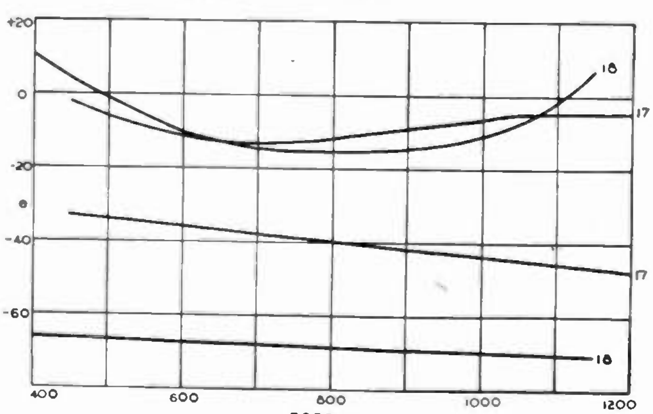

Fig. 5—Illustrating practical examples of drift correction by use of the padded circuit with coefficient tracking.

The upper curve for (17) gives a very good experimental check, since the exact values of coefficients required by the expressions were not readily obtainable. Values somewhat lower were used, and it was expected that the drift would be a few parts per million negative throughout the range. The inability to obtain the exact values for the coefficients, as determined by the expressions, will sometimes be the limiting factor in design.

The circuit used for (18) covers a frequency range of 2.9 to 1. In this case θ was made zero at two frequencies within the range, so as to reduce the drift at 800 kilocycles.

The examples are representative of what may be done with the use of stock broadcast receiver parts, having existing tolerances on the various electrical

[8] A. G. Manke, *RMA Eng.*, vol. 2, p. 18; 1938.

values and coefficients. It should be possible to have the average value of the corrected drift not more than ten parts per million per degree centigrade throughout a range of nearly 3 to 1.

Further improvement results from the restriction of the frequency coverage, and/or the use of components in which thermal drift has been a specific factor in design.

The adjustment of C_z and C_z was tried in a few cases, and it was found that the correction of such circuits called for coefficient values in excess of those possible. In any case, such expressions would have only limited application, since usually only one or the other is present in a circuit.

IX. Three-Point Tracking

The possibilities of three-point tracking have been briefly examined. It is evident that by operating on all three coefficients (α_x, α_y, and α_z), a third expression for θ at some other frequency could be set up and equated to zero. This would give a third tracking point at any suitable frequency within the range.

Such a third tracking point would be useful if there is considerable curvature in the two-point characteristic, as, e.g., in the examples of Fig. 5. Three-point tracking also gave unattainable values for $\alpha_{z'}$ and $\alpha_{z''}$, and it is not, in general, practically possible to have $\theta=0$ at more than two frequencies in the range, so that (17) or (18) gives the best possible compensation for any circuit.

The alteration of the third coefficient will have some effect on the shape of the curve, and may therefore give a smaller departure from zero at mid-frequencies.

IX. Conclusions

In circuits intended for operation at one frequency, thermal drift can be eliminated by adjustment of the temperature coefficient of a portion of the capacitance, as determined by (7).

In variable-capacitance-tuned circuits, the integrated drift is a minimum when $\theta_{L'}^2+\theta_{H'}^2$ is a minimum, the design information being given by (10). A close approximation to optimum conditions obtains when $\theta_{H'}$ is made zero, either by the use of (10c) or experimentally.

Circuits having θ less than ten parts per million per degree centigrade over normally used frequency ranges, are sometimes possible by the complementary adjustment of the coefficients of inductance and capacitance, as given in (11) and in the graphical construction of Fig. 3.

Variable-inductance-tuned circuits are a special case of the general formulas obtained for the variable-capacitance-tuned circuits. Minimum integrated drift occurs when $\theta_{L'}+\theta_{H'}=0$. These circuits have the advantage of being very suitable for "warm-up" drift reduction.

The padded circuit of the superheterodyne local oscillator is peculiarly adapted to thermal-drift correction by reason of its complex capacitance network. Suitable adjustment of two of the condenser coefficients, giving two-point coefficient tracking, results in a very small drift throughout the range, when the conditions of (17) or (18) are met.

Experimental work has confirmed the correctness of the expressions in so far as the experimental errors permit. It appears that the average value of the drift factor in the padded circuits is readily reduced to ten parts per million per degree centigrade for frequency ranges of up to 3 to 1.

Three-point tracking of the temperature coefficient of frequency, by the adjustment of three condenser coefficients, has been found to be impractical, due to the large values of coefficient called for in normal circuits.

Acknowledgment

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Attenuation of Electromagnetic Fields in Pipes Smaller Than the Critical Size*

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Summary—*A theoretical and experimental discussion is given of electromagnetic fields in pipes smaller than the critical size, especially with regard to attenuation, and based upon wave-guide theory. It is shown that the rate of attenuation, as the wavelength increases and passes through the critical value, approaches a high asymptotic value. Confirmatory experimental data are given. Simple formulas for attenuation are included.*

RECENTLY several papers[1-3] have appeared in which the propagation of electromagnetic waves in pipes has been discussed. These have indicated that for a given frequency there is a critical pipe size below which wave propagation is not possible, or in other words, for a pipe of a given size there is a cutoff frequency f_0 below which wave propagation does not occur. Aside from this, so far as the writer is aware, no discussion has previously been made of phenomena below cutoff or very near to it. In fact most of the present published theory is not valid for frequencies very near to f_0. It is the purpose of the present paper to discuss this case somewhat more completely.

The general expressions for the fields in cylindrical tubes may be written in the form[1,2]

$$\left.\begin{aligned} E &= E_1(r, \theta) \exp(jwt \pm \gamma z) \\ H &= H_1(r, \theta) \exp(jwt \pm \gamma z) \end{aligned}\right\}, \tag{1}$$

whence it is seen that the field along the axis of the tube is controlled by the factor $\exp(jwt \pm \gamma z)$, where γ, the propagation constant, may be written $\gamma = \alpha + j\beta$. The real part of γ is the attenuation constant and the imaginary part is the phase constant.

The forms of propagation constant discussed in the above references[1-3] are not valid for $f \leq f_0$, except in Barrow's paper. He gives a form of γ which is valid for the E-type wave, over the whole frequency range, but does not discuss it, putting it immediately into a simpler form which does not hold over the whole range. Barrow's general expression is limited only by the assumption that the conductivity of the tube is very large but finite. After making several changes of notation it may be written

$$\gamma = \frac{2\pi}{\lambda_0}\left\{\left[1 - \left(\frac{\lambda_0}{\lambda}\right)^2 - \frac{\lambda_0{}^2 w}{4\pi^2}\right] + j\,\frac{\lambda_0{}^2 w}{4\pi^2}\right\}^{1/2}, \tag{2}$$

where λ_0 is the free-space wavelength corresponding to f_0, i.e., the cutoff wavelength, and

$$w = \frac{\sqrt{2}\,\omega\epsilon_1}{a}\sqrt{\frac{\omega\mu_2}{\sigma_2}}, \tag{3}$$

where

ϵ_1 = dielectric constant of the medium ($10^{-11}/36\pi$ farad per centimeter for air)
μ_2 = permeability of the tube (henries per centimeter)
σ_2 = conductivity of the tube (mhos per centimeter)
a = radius of the tube.

For copper pipes of practical sizes, the w term is negligibly small except very near cutoff.

The real part of (2) determines the attenuation. It is of interest to consider its variation as λ varies from below to above λ_0. This has been plotted in Fig. 1,

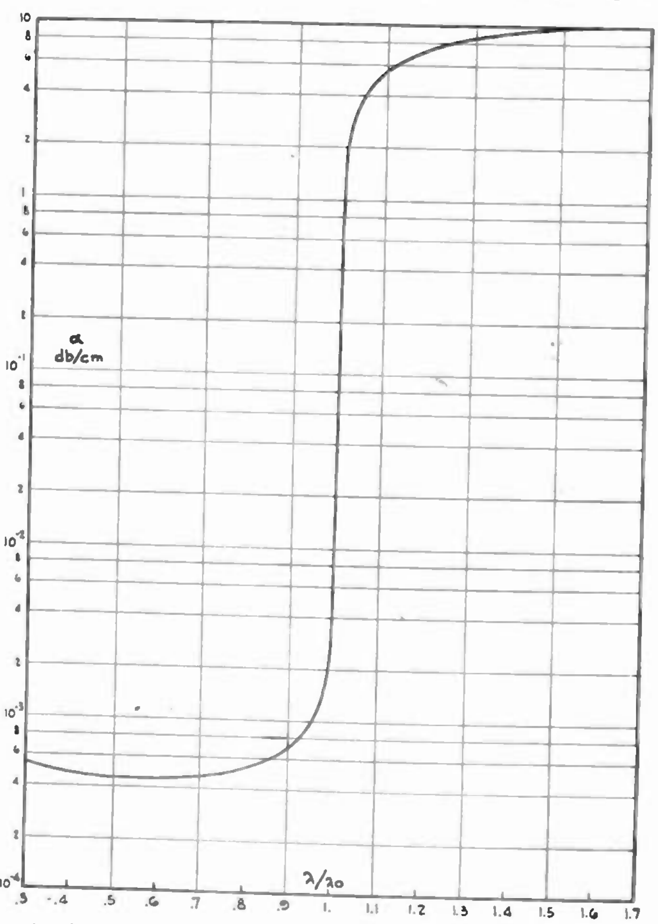

Fig. 1—Attenuation for the case of an $E_{0,1}$ type of field in a copper pipe of 1.58 centimeters radius, showing the variation through the cutoff point.

for the case of $E_{0,1}$ type of field in a copper tube of radius $a = 1.58$ centimeters. For $\lambda < \lambda_0$ the curve is identical with those published previously[1-3], except as

* Decimal classification: R110. Original manuscript received by the Institute, April 6, 1942.
† RCA Laboratories, Princeton, New Jersey.
[1] W. L. Barrow, "Transmission of electromagnetic waves in hollow tubes of metal," Proc. I.R.E., vol. 24, pp. 1298–1329; October, 1936.
[2] J. R. Carson, S. P. Mead, and S. A. Schelkunoff, "Hyper-frequency wave guides," *Bell Sys. Tech. Jour.*, vol. 15, pp. 310–333; April, 1936.
[3] G. C. Southworth, "Hyper-frequency wave guides," *Bell Sys. Tech. Jour.*, vol. 15, pp. 284–309; April, 1936.

$\lambda \to \lambda_0$ the attenuation does not become infinite as previously indicated, but rises very steeply to a comparatively high value, and for $\lambda > \lambda_0$ it approaches an asymptotic value of 13.2 decibels per centimeter.

The imaginary part of γ determines the form of the field along the z axis. For $\lambda < \lambda_0$, it is seen from (2), the imaginary part is very large compared to the real part; i.e., $\beta \gg \alpha$. Hence we see from (1),

$$E = E_1(r, \theta)\exp(-\alpha z)\exp(-j\beta z),$$

that the field is simple harmonic with small attenuation. However, for $\lambda > \lambda_0$, we have $\beta \ll \alpha$. The factor $\exp(-j\beta z)$ is very near unity. The harmonic variation disappears and the field decreases at a high exponential rate as determined by $\exp(-\alpha z.)$

At the cutoff point $\lambda = \lambda_0$ and $\gamma = \sqrt{w(-1+j)} = \sqrt{w}(\cos 67.50 + j \sin 67.50)$. Thus α and β are of similar magnitude.

It is evident that there is a continuous transition through the cutoff point, from a slightly attenuated simple harmonic wave to a highly attenuated exponential field.

It is possible to derive very simple expressions for the attenuation which are accurate except close to λ_0. To do this write γ^2 in the form

$$\gamma^2 = \rho\epsilon^{2\phi},$$

whence

$$\gamma = \sqrt{\rho}\,\epsilon^{\phi},$$

$$\alpha = \sqrt{\rho}\cos\phi,$$

and

$$\beta = \sqrt{\rho}\sin\phi,$$

where, from (2)

$$\rho = \frac{4\pi^2}{\lambda_0^2}\sqrt{\left(1 - \frac{\lambda_0^2}{\lambda^2} - \frac{\lambda_0^2 w}{4\pi^2}\right)^2 + \left(\frac{\lambda_0^2 w}{4\pi^2}\right)^2}.$$

Except very near to λ_0 the w term is negligibly small and this is closely approximated by

$$\rho = \frac{4\pi^2}{\lambda_0^2}\left(\frac{\lambda_0^2}{\lambda^2} - 1\right), \quad \text{for} \quad \lambda < \lambda_0,$$

and

$$\rho = \frac{4\pi^2}{\lambda_0^2}\left(1 - \frac{\lambda_0^2}{\lambda^2}\right), \quad \text{for} \quad \lambda > \lambda_0.$$

Consider the two cases:

1. $\lambda < \lambda_0$.

Here γ^2 has a large negative real part, and a small positive imaginary part, hence 2ϕ is almost 180 degrees and ϕ is almost 90 degrees. Hence we may write $\cos\phi = w/2\rho$, and $\sin\phi = 1$.

Therefore,

$$\alpha = \sqrt{\rho}\cos\phi = \frac{\lambda_0 w/4\pi}{\sqrt{\dfrac{\lambda_0^2}{\lambda^2} - 1}},$$

or, multiplying numerator and denominator by λ/λ_0,

$$\alpha = \frac{\lambda w/4\pi}{\sqrt{1 - \dfrac{\lambda^2}{\lambda_0^2}}}, \tag{4}$$

and

$$\beta = \frac{2\pi}{\lambda}\sqrt{1 - \frac{\lambda^2}{\lambda_0^2}}. \tag{5}$$

These are identical with the expressions derived by Barrow;[1] and Carson, Mead, and Schelkunoff[2], for this case.

2. $\lambda > \lambda_0$.

Here γ^2 has a large positive real part, and a small positive imaginary part. Hence, 2ϕ is nearly zero, $\cos\phi = 1$, and $\sin\phi = w/2\rho$. Therefore,

$$\alpha = \frac{2\pi}{\lambda_0}\sqrt{1 - \frac{\lambda_0^2}{\lambda^2}}, \tag{6}$$

and

$$\beta = \frac{\lambda_0 w/4\pi}{\sqrt{1 - \dfrac{\lambda_0^2}{\lambda^2}}}. \tag{7}$$

It is of interest that the attenuation in this case depends only on λ and λ_0 and is independent of the tube material, except for the assumption that it is of high conductitivy.

The above discussion applies only to E waves, since the basic expression (2) was derived for that case only. The case of H waves for $\lambda < \lambda_0$ has been discussed in detail by Barrow; Carson, Mead, and Schelkunoff, etc. The case $\lambda > \lambda_0$ may be handled by noting that the propagation constant for H waves may be written[4]

$$\gamma = \frac{2\pi}{\lambda_0}\sqrt{1 - \frac{\lambda_0^2}{\lambda^2}},$$

except for λ very near to λ_0. For $\lambda > \lambda_0$ this is real and thus represents the attenuation, which is seen to be identical with (6). The assumptions are the same in both cases. Hence (6) is valid for both E waves and H waves.

Equations (4) and (5) have been discussed in previous publications,[1-3] and will not be dealt with further here. Equations (6) and (7) however appear not to have been mentioned previously.[5] Equation (6) has been checked over the wavelength range from 7 to 10 centimeters for a tube of 1.58 centimeters radius, and for a field configuration of the $H_{1,1}$ type. These data, which were obtained by G. Fernsler, are plotted

[4] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Company, New York, N. Y., 1941, p. 539.

[5] R. A. Braden of this laboratory first noticed that our experimental measurements of attenuation rate followed an empirical law similar to equation (6). The similarity of this to expressions occurring in wave-guide theory led to the present investigation.

in Fig. 2. There is agreement within the accuracy of the measurements.

The extension or fringing of fields into tubes has been discussed in a number of instances[6,7] for static or low-frequency fields. These represent limiting cases given by $\lambda \to \infty$, and for which, from (6),

$$\alpha = 2\pi/\lambda_0. \tag{8}$$

Harnett and Case[7] have described three examples, to which this result is applicable, in which tubes are used

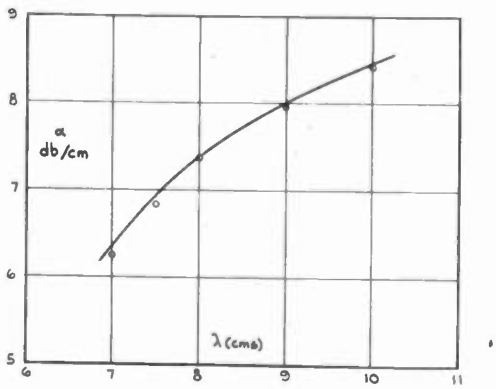

Fig. 2—Attenuation for the case of an $H_{1,1}$ type of field in a pipe of 1.58 centimeters radius. The circles are experimental points. The line is computed from equation (6).

as attenuators for signal generators, and in which three different field configurations are employed. The first type has an input electrode consisting of a small circular disk centrally located in the tube. This produces a radial electric field similar to that of the $E_{0,1}$ wave. Hence from wave-guide theory, $\lambda_0 = 2.62a$. Therefore, from (8),

$$\alpha = \frac{2\pi}{2.62a} \text{ napiers per centimeter}$$

$$= \frac{20.9}{a} \text{ decibles per centimeter}$$

$$= 20.9 \text{ decibles per radius,}$$

which is in agreement with previous results.[6,7] The second type has an input electrode consisting of a coil whose axis is normal to the axis of the tube. This produces a field configuration similar to that of the $H_{1,1}$ wave. The limiting value of α is found to be 16.0 decibels per radius. The third type consists of an input electrode in the form of a coil whose axis coincides with that of the tube. The field is analogous to that of the $H_{0,1}$ wave, and the limiting attenuation is found to be 33.3 decibels per radius. These three cases are in agreement with the limiting values found by H. A. Wheeler, and given in the Harnett and Case paper.

The formulas for $\lambda > \lambda_0$ for these three cases may be written as follows:

$$E_{0,1} \text{ type, } \alpha = 20.9\sqrt{1 - \frac{\lambda_0^2}{\lambda^2}} \text{ decibels per radius,} \tag{9}$$

$$H_{1,1} \text{ type, } \alpha = 16.0\sqrt{1 - \frac{\lambda_0^2}{\lambda^2}} \text{ decibels per radius,} \tag{10}$$

$$H_{0,1} \text{ type, } \alpha = 33.3\sqrt{1 - \frac{\lambda_0^2}{\lambda^2}} \text{ decibels per radius.} \tag{11}$$

Formulas for other types of waves may be derived by inserting the proper value of λ_0 in (6).

The accuracy of (6) and (7), and also the derived formulas (9), (10), and (11) may be estimated as follows: From the accurate expression (2) we see that the terms involving w (which were neglected in obtaining (6) and (7)) affect the order of magnitude of the result only when

$$1 - \frac{\lambda_0^2}{\lambda^2} \doteq \frac{\lambda_0^2 w}{4\pi^2} \doteq 10^{-4}$$

for copper pipes. This may be written

$$\lambda^2 - \lambda_0^2 = 10^{-4}\lambda^2,$$

or

$$(\lambda - \lambda_0)(\lambda + \lambda_0) \doteq 10^{-4}\lambda^2,$$

or

$$(\lambda - \lambda_0) = \Delta\lambda \doteq \frac{10^{-4}\lambda^2}{2\lambda_0},$$

or

$$\Delta\lambda \doteq 10^{-4}\lambda_0.$$

Thus, λ must be within 0.01 per cent of λ_0 before the w term is equal in magnitude to the sum of the other terms.

[6] I. Langmuir and K. T. Compton, "Electrical discharges in gases, Part II," *Rev. Mod. Phys.*, vol. 3, pp. 212–213; April, 1931.

[7] D. E. Harnett and N. P. Case, "The design and testing of multirange receivers," Proc. I.R.E., vol. 23, pp. 578–594; June, 1935. The attenuation formulas given in this paper were derived by H. A. Wheeler, but his derivations have not been published.

Institute News and Radio Notes

WINTER CONFERENCES—1943

The Board of Directors of the Institute, co-operating in the program of reducing long-haul railroad passenger traffic, has acted to eliminate the usual three-day convention and exhibition of component radio parts in New York this forthcoming January. To replace it, the Board has named Thursday, January 28, 1943, as a nationwide Winter Conference Day, on which as many Sections of the Institute as can possibly do so will hold simultaneous technical meetings. In this ambitious plan to "bring the mountain to Mahomet," the Directors believe that the Institute will achieve many of the more important objectives of a national convention, and, if Section officers enter into the spirit of the idea, will arouse local interest to an extent not possible through gathering at one point.

In New York City, the Board has accepted the kind invitation of the American Institute of Electrical Engineers, which during the week of January 25–29 is holding its National Technical Meeting there, to join with that society on January 28. Morning and afternoon sessions of the I.R.E. will be open to its own members and also to registrants at the A.I.E.E. technical meeting. The A.I.E.E. has courteously invited I.R.E. members to its communication and industrial electronics sessions, particulars of which will appear in the January PROCEEDINGS. The day's activities will culminate in the joint A.I.E.E.-I.R.E. evening meeting, with an address on the subject of Ultra-High Frequencies by Dr. George C. Southworth of Bell Telephone Laboratories.

Technical Papers

It is expected that technical papers and material used at the New York Conference will be made available, upon request, for simultaneous presentation at other I.R.E. Sections on the date named. Steps are being taken to make possible the exchange of papers among Sections so that each may contribute and all may benefit. If desired by the authors, papers so contributed will be considered by the Papers Committee and Board of Editors as to subsequent publication in the PROCEEDINGS. Papers for presentation at the conference must be submitted to the Institute office, 330 West 42nd Street, New York, N. Y., not later than January 1, 1943.

Other Features

Other features on the New York Conference program are: The Annual Meeting of the Institute; induction of officers for the year 1943; presentation of the Medal of Honor; award of Fellowships and other special awards. There will be no social activities at the New York Conference and no ladies' program. There will be no meeting of the Sections Committee.

Board of Directors

The Board of Directors met on October 7, 1942. Those present were A. F. Van Dyck, president; C. C. Chambers, I. S. Coggeshall, W. L. Everitt, Alfred N. Goldsmith, editor; L. C. F. Horle, C. M. Jansky, Jr., F. B. Lewellyn, Haraden Pratt, treasurer; F. E. Terman, B. J. Thompson, H. M. Turner, H. A. Wheeler, L. P. Wheeler, and H. P. Westman, secretary.

Institute representatives at 76 colleges and universities were appointed to serve until June 30, 1943.

A new Bylaw to follow present Bylaw Section 45 was adopted and reads as follows:

"The Board shall make appointments to the following Committees: Annual Review, Electroacoustics, Electronics, Facsimile, Frequency Modulation, Radio Receivers, Radio Wave Propagation, Standards, Symbols, Television, and Transmitters and Antennas, each year between January 1 and May 1 and the terms of appointments shall be from May 1 of the year when the appointments are made until April 30 of the following year. Additional appointments may be made to fill vacancies or to care for special cases as the need arises, with the term of the appointment expiring April 30."

The report of a special committee appointed to present to the Board a report on the advisability of, and on ways and means of, changing the membership-grade structure to permit the Institute to serve better a larger and more diversified membership was considered. The Constitution and Laws Committee was directed to prepare the necessary changes in the Constitution and Bylaws to put the proposals of the Committee into effect.

The Executive Committee was directed to appoint a Conference Committee to arrange for a technical conference in New York City during the latter part of January, 1943. This conference, which will be of one day's duration, will replace the normal Winter Convention which was canceled some months ago.

The regular November meeting of the Board of Directors was held on the fourth of the month and was attended by A. F. Van Dyck, president; Austin Bailey, C. C. Chambers, I. S. Coggeshall, W. L. Everitt, Alfred N. Goldsmith, editor; C. M. Jansky, Jr., F. B. Llewellyn, Haraden Pratt, treasurer; F. E. Terman, B. J. Thompson, H. M. Turner, L. P. Wheeler, and H. P. Westman, secretary.

The report of the Tellers Committee was received and the following new officers were declared elected:

President, 1943
L. P. Wheeler
Vice President, 1943
F. S. Barton
Directors, 1943–1945
W. L. Barrow
F. B. Llewellyn
H. A. Wheeler

On recommendation of the Awards Committee the following members will be advanced to Fellow at the time of the Annual Meeting which will be held in New York City in January: Andrew Alford, I. S. Coggeshall, J. B. Dow, L. E. DuBridge, P. C. Goldmark, D. E. Harnett, D. D. Israel, A. G. Jensen, G. F. Metcalf, and Irving Wolff.

The membership of the Appointments Committee was decided on and it will be comprised of L. P. Wheeler, chairman;

W. L. Everitt, F. B. Llewellyn, B. J. Thompson, and H. A. Wheeler.

A new Bylaw was adopted to follow the present Bylaw Section 44 and reads as follows:

"The Board of Directors is authorized to waive, in whole or in part, the application in any particular case of the contents of Bylaw Section 44 during the period ending December 31, 1943."

Section 44 of the Institute Bylaws prescribed the minimum requirements for the operation of sections. This Bylaw was adopted in anticipation that some sections may have difficulty in maintaining these minimum requirements basically as a result of general conditions and not through lack of interest or initiative on the part of the management groups.

A New York Section having been established, no further need exists for the New York Program Committee which formerly prepared the programs for New York meetings. Accordingly, Section 45 of the Institute Bylaws was amended to delete from the list of standing committees the name of the New York Program Committee.

Alfred N. Goldsmith was named to serve as a representative of the Institute on the Standards Council of the American Standards Association for the period 1943–1945, the Secretary being designated to serve as alternate for the same period.

Professor Everitt announced that there were no plans being made for a Broadcast Engineering Conference at Ohio State University during 1943.

The report of the Constitution and Laws Committee on proposed revisions of the Constitution and Bylaws submitted in accordance with the instructions issued at the previous meeting of the Board of Directors, was considered. A number of modifications were made in the proposals and the report returned to the Committee for further consideration by it.

Executive Committee

A meeting of the Executive Committee was held on September 18 and was attended by A. F. Van Dyck, chairman; Alfred N. Goldsmith, editor; Haraden Pratt, treasurer; B. J. Thompson, and H. P. Westman, secretary.

A memorandum on broadening the scope of the Institute was prepared for submission to the Board of Directors.

W. B. Cowilich was employed to serve as Assistant Secretary.

On October 2, A. F. Van Dyck, chairman; I. S. Coggeshall, Alfred N. Goldsmith, editor; R. A. Heising (guest); F. B. Llewellyn, Haraden Pratt, treasurer; B. J. Thompson, and H. P. Westman, secretary, attended a meeting of the Executive Committee.

It was agreed that the memorandum on "Broadening the Scope of the Institute" which was prepared at the previous meeting of the Executive Committee be forwarded to the Constitution and Laws Committee together with certain other information and views for its information and guidance in preparing proposed modifications of the Constitution and Bylaws. The proposed changes were requested in time for consideration by the Board of Directors at its November meeting.

Approval was granted of 88 applications for Associate, 41 for Student, and 4 for Junior membership.

A cordial invitation from the Institution of Electrical Engineers of London for Institute members visiting England to make use of its library facilities and to attend its meetings was accepted with thanks. A reciprocal privilege is afforded to members of the I.E.E. on visit to this country to make use of such facilities of our Institute as may be useful to them.

On October 30, a meeting of the Executive Committee was held and was attended by A. F. Van Dyck, chairman; I. S. Coggeshall, Alfred N. Goldsmith, editor; R. A. Heising (guest); F. B. Llewellyn, Haraden Pratt, treasurer; B. J. Thompson, and H. P. Westman, secretary.

J. B. Atwood, Stewart Becker, J. F. Johnson, H. T. Maser, M. D. McFarlane, E. R. Piore, and C. H. Wesser were transferred to Member grade. F. B. Bramhall, D. D. Carpenter, A. V. Dubinin, D. H. Marathe, E. D. McArthur, M. S. Neiman, C. T. Scully, A. H. Simons, G. F. Van Dissel, C. M. Wallis, and Michael Wysotzky were admitted to Member.

Approval was granted of 112 applications for Associate, 99 for Student, and 3 for Junior grade.

The petition for the establishment of a New York Section which was received on October 7 and signed by 5 Fellows, 13 Members, and 64 Associates in good standing was approved.

A report of the Constitution and Laws Committee on proposals to modify the Constitution and Bylaws of the Institute was reviewed.

Books

A-C Calculation Charts, by R. Lorenzen

Published by John F. Rider Publisher, Inc., 404 Fourth Avenue, New York, N. Y. 160 pages. 144 charts. 9×12 inches. Price $7.50.

After a brief historical and explanatory introduction, this large volume comprises 144 full-page reactance charts like those commonly used by communication engineers. Each chart is one cycle square, printed in two colors. All charts are alike except for the numbering of the scales. They cover the range of 10 cycles to 1000 megacycles, 0.01 ohm to 10 megohms, 0.1 micromho to 100 mhos.

Each of the 144 charts is accompanied by two conversion scales as an aid in impedance computations. One is a reciprocal conversion between ohms and mhos. The other is a square or square-root conversion for solving right triangles. There are two additional charts, one giving the relation $Q=X/R$ or B/G and the other giving $Q=\tan\theta$.

This reference book is recommended for libraries and laboratories where many reactance computations in various frequency ranges have to be made with accuracy nearly as great as that of a ten-inch slide rule.

H. A. Wheeler
Hazeltine Service Corporation
Little Neck, L. I., N. Y.

Handbook of Technical Instruction for Wireless Telegraphists, by H. M. Dowsett and L. E. Q. Walker

Published by Iliffe and Sons, Ltd., Dorset House, Stamford Street, London, S.E. 1, England, Seventh Edition, 1942. 664 pages. 618 figures. $5\frac{1}{2}\times 8$ inches. Price, 25 shillings.

This Handbook aims chiefly to provide instruction and reference material for seagoing operators and others, relative to the general practice of marine wireless communication, illustrated by apparatus developed by British wireless companies.

About one fourth of the Handbook is devoted to the basic theoretical considerations which are applicable in all fields of radio communications. This material is clearly presented and numerous mathematical formulas most likely to be useful to the operator in the field are included.

Circuit diagrams are presented for a great number of different types of British shipboard apparatus, including transmitters, receivers, direction finders, auto alarms, emergency sets, and lifeboat sets, making the Handbook a convenient reference source for anyone working in this field.

H. O. Peterson
R.C.A. Communications, Inc.
Riverhead, L. I., N. Y.

The Radio Amateur's Handbook (Nineteenth Edition), 1942

Published by the American Radio Relay League, Hartford, Conn. 446 pages +8-page index+96-page catalog section. 680 figures. $6\frac{1}{2}\times 9\frac{1}{2}$ inches. Price $1.00.

This is the nineteenth edition of the Radio Amateur's Handbook. Its arrangement and selection of material reflects the experience gained in the production of eighteen preceding editions.

About one third of the Handbook is devoted to the presentation of the basic theory of the components used in radio communication. This section includes a good chapter on practical antenna design.

There is also a section relative to the practices of station operation and traffic handling.

About one third of the Handbook pertains to construction plans and data for transmitters, receivers, transceivers, modulation equipment and other items of general use in the amateur's station. A comprehensive selection of tube characteristics and miscellaneous data are also included.

The catalog section is of interest and value in that many of the component parts illustrated may be applicable to the radio problems incidental to the war effort.

H. O. Peterson
R.C.A. Communications, Inc.
Riverhead, L. I., N. Y.

Acoustic Design Charts, by Frank Massa

Published by the Blakiston Company, Philadelphia, Pa. 219 pages+xiv pages+8-page index. 6¼×9¼ inches. Price $4.00.

Those engaged in the design of electro-acoustical apparatus will welcome the publication of this volume. It comprises a collection of charts, mostly in the form of straight line graphs on log-log co-ordinates, of quantitative data pertaining principally to acoustics and to mechanical vibrating systems. In many cases families of curves are given, which permit the effect of varying the parameters of a system to be evaluated without computation. Even a mere inspection of the charts is instructive, in that the ranges of values to be encountered in practical problems are indicated. A convenient feature is the use of both English and metric units in appropriate cases; for example, the length of a tube as a function of its resonance frequency is given in feet, in inches, and in centimeters. In most cases the scales chosen permit a constant percentage accuracy of reading over the large range of values covered. While the charts are not large in size, the precision is sufficient for most preliminary design calculations.

A wide range of material is covered, most of which is particularly applicable to the design of loudspeakers and microphones. The charts are arranged in ten sections covering fundamental relations in plane and spherical sound waves; attenuation of sound and vibrations; acoustical elements; mechanical and acoustical vibrating systems, including clamped, stretched, and piston diaphragms; design factors involved in direct radiator and horn-type loudspeakers; sound reproduction in rooms and in free space; design of electromagnetic systems; and, miscellaneous other data of use to acoustical engineers. There is a complete index with adequate cross references.

In few branches of engineering other than acoustics is a knowledge of fundamental principles more essential to the safe application of data such as are presented here. It is, therefore, unfortunate that neither the equations of the graphs nor references locating them in the literature can be found anywhere in the book. This is an omission which it is hoped will be rectified in future editions of this very useful volume.

Academicians may argue that graphic aids such as these are crutches for mental cripples. To this reviewer, however, the Massa charts seem more like a comfortable wheel chair in which one may, with little effort, glide smoothly along paths formerly negotiated only with painful effort.

Benjamin Olney
Stromberg-Carlson Telephone
Manufacturing Co., Rochester, N. Y

Contributors

Cledo Brunetti

Cledo Brunetti (A'37) was born on April 1, 1910, at Virginia, Minnesota. He received the B.E.E. degree from the University of Minnesota in 1932 and the Ph.D. degree in 1937. From 1932 to 1936 he was a Teaching Fellow in the department of electrical engineering at the University of Minnesota and from 1936 to 1937, an instructor. From 1937 to 1939 Dr. Brunetti was an instructor in electrical engineering at Lehigh University and from 1939 to 1941, assistant professor of electrical engineering. During the summers of 1939 and of 1940 he was a research associate at the radio laboratory, National Bureau of Standards, Washington, D. C. Since May, 1941, he has been a radio physicist at the National Bureau of Standards, Washington, D. C., working on problems in connection with the war effort. He is a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

Beverly Dudley (J'24–A'27) was born on April 2, 1906, at Chicago, Illinois. He attended Armour Institute from 1926 to 1928. In 1935, he received the B.S. degree from Massachusetts Institute of Technology; from 1935 to 1939 he did graduate work in electrical engineering and physics at Columbia University. Mr. Dudley was assistant technical editor of *QST* at the American Radio Relay league from 1929 to 1930 and assistant secretary of the Institute of Radio Engineers from 1930 to 1932. He did technical editorial work at the General Radio Company during the summer of 1934 and experimental and factory production, test, and specifications on cathode-ray tubes and metal receiving tubes at the RCA Manufacturing Company, 1935–1936. Since 1936 he has been with the McGraw-Hill Publishing Company as assistant editor, associate editor, and managing editor of *Electronics*. From 1939 to 1941 he was managing editor of *Photo Technique*. At the present time he is the consulting editor for a new series of texts to be known as the Radio Communication Series. Mr. Dudley is the author of "Handbook of Photography; Making your Photographs Effective" and "Introduction to Radio Engineering," which is now in preparation. He is a member of the Optical Society of America and the Acoustical Society of America.

Beverly Dudley

C. Herbert Gleason

C. Herbert Gleason (A'42) was born at Halstead, Kansas on October 1, 1918. He received the B.S. degree in electrical engineering from the University of Kansas in 1940 and the M.S. degree in E. E. from the University of Missouri, 1942. He was a teaching assistant at the University of Missouri during 1940 and 1941 and an instructor at that University in 1941 and 1942. At present, Mr. Gleason is a junior

LEIGHTON GREENOUGH

engineer with the electronics engineering division of the Westinghouse Electric and Manufacturing Company, Bloomfield, New Jersey. He is a member of Sigma Xi and Tau Beta Pi.

⁂

Leighton Greenough was born on December 30, 1918. He received the B.S. degree in electrical engineering from Tufts College in 1940. In 1942, he received the M.S. degree in electrical engineering from Lehigh University, after two years as half-time Research Fellow. His work was with negative circuit elements. Since his graduation Mr. Greenough has been with the RCA Laboratories.

RALPH N. HARMON

⁂

Ralph N. Harmon (A '30) was born on August 29, 1906, at California, Pennsylvania. He received the B.S. degree in electrical engineering in 1928 from the Carnegie Institute of Technology. From 1925 to 1929, Mr. Harmon was with the Duquesne Light Company; since 1929, he has been with the Westinghouse Electric and Manufacturing Company.

⁂

E. G. LINDER

Ernest G. Linder (A '40) was born in 1902 at Waltham, Massachusetts. He received the B.A. degree from the University of Iowa in 1925, and the M.S. degree in 1927. He was an instructor in physics at the State University of Iowa from 1925 to 1927; an instructor at California Institute of Technology from 1927 to 1928; a Detroit Edison Fellow at Cornell University from 1928 to 1932, where he received his Ph.D. degree in 1931. Dr. Linder was employed in the Research Division of the RCA Victor Company from 1932 to 1935; and the RCA Manufacturing Company, RCA Victor Division from 1935 to 1942. At present he is with the RCA Laboratories at Princeton, New Jersey. He is a member of the American Physical Society.

⁂

D. L. WAIDELICH

D. L. Waidelich (S'37–A'39) was born on May 3, 1915, at Allentown, Pennsylvania. He received the B.S. degree in electrical engineering in 1936 and the M.S. degree in 1938 from Lehigh University. From 1936 to 1938 he was a teaching assistant at Lehigh University; in 1938, he became an instructor at the University of Missouri and he is now an assistant professor of electrical engineering there. Mr. Waidelich is an associate member of the American Institute of Electrical Engineers and a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

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GENERAL INFORMATION

The Institute

The Institute of Radio Engineers serves those interested in radio and allied electrical-communication fields through the presentation of publication and technical material.

Membership has grown from a few dozen in 1912 to more than eight thousand. There are several grades of membership, depending on the qualifications of the applicant, with dues ranging from \$3.00 per year for Students to \$10.00 per year for Members and Fellows.

PROCEEDINGS, Standards Reports, and any other published material are sent to members without further payment.

The PROCEEDINGS

The PROCEEDINGS has been published without interruption from 1913 when the first issue appeared. Over 2100 technical contributions have been included in its pages and portray a currently written history of developments in both theory and practice. The contents of every paper published in the PROCEEDINGS are the responsibility of the author and are not binding on the Institute or its members. Text material appearing in the PROCEEDINGS may be reprinted or abstracted in other publications on the express condition that specific reference shall be made to its original appearance in the PROCEEDINGS. Illustrations of any variety may not be reproduced, however, without specific permission from the Institute.

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Back Copies

The Institute endeavors to keep on hand a supply of back copies of the PROCEEDINGS for sale for the convenience of those who do not have complete files. However, some issues are out of print and cannot be provided.

Standards

In addition to the material published in the PROCEEDINGS, Standards on various subjects have been printed. These are available to members free of charge as long as the supply lasts; others may purchase them at the prices listed below.

	Price
Standards on Electroacoustics, 1938	\$0.50
Standards on Electronics, 1938	Out of Print
Standards on Transmitters and Antennas, 1938	Out of Print
Standards on Transmitters and Antennas: Definitions of Terms, 1938	\$0.20
A Reprint (1942) of the like-named section of "Standards on Transmitters and Antennas, 1938.	
Standards on Transmitters and Antennas: Methods of Testing, 1938	\$0.50
A Reprint (1942) of the like-named section of "Standards on Transmitters and Antennas, 1938."	
Standards on Radio Receivers, 1938	Out of Print
Note: Two sections of this report were reprinted as individual booklets: "Definitions of Terms" and "Methods of Testing Broadcast Radio Receivers."	
Standards on Radio Receivers: Definitions of Terms, 1938	\$0.20
A reprint (1942) of the like-named section of "Standards on Radio Receivers, 1938."	
Standards on Radio Receivers: Methods of Testing Broadcast Radio Receivers, 1938	\$0.50
A reprint (1942) of the like-named section of "Standards on Radio Receivers, 1938."	
Standards on Radio Wave Propagation: Definitions of Terms, 1942	\$0.20
Standards on Radio Wave Propagation: Measuring Methods, 1942	\$0.50
Standards on Facsimile: Definitions of Terms, 1942	\$0.20

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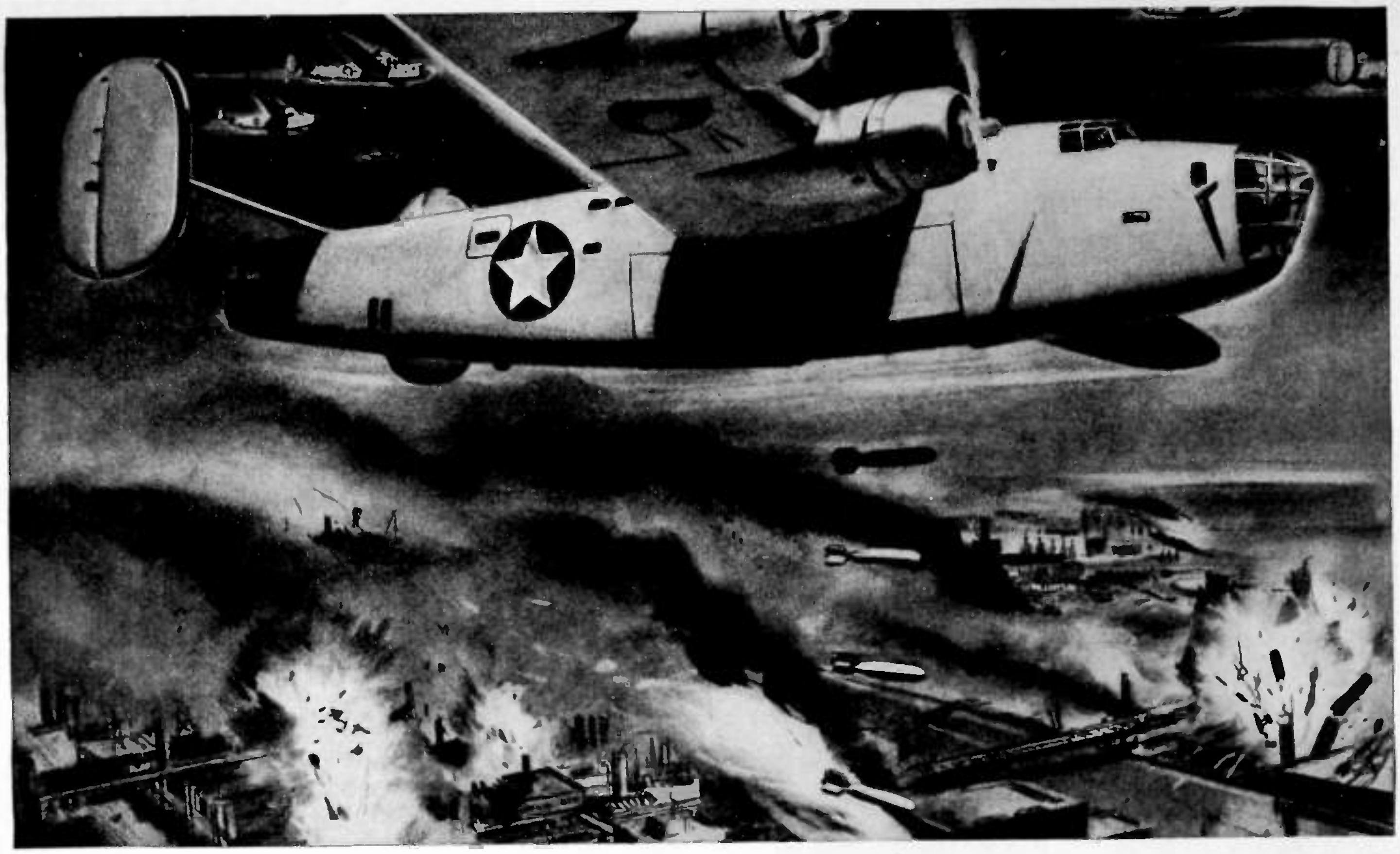
Numbers refer to the chronological list. Light-face type indicates papers, **bold-face** type indicates discussions, and *italics* refer to books and book reviews.

SUBJECT INDEX

NONTECHNICAL INDEX



More bombs for Berlin . . . more terror for Tokio

IN TOTAL WARFARE, every pound of metal we can save means just so much added destruction we can hurl at the enemy.

7000 tons of steel saved means approximately 70,000 more heavy bombs to drop on Berlin and Tokio.

2500 tons of copper saved means nearly 1,000,000 additional 3 in. A. A. cartridge cases for blasting Zeros and Messerschmitts out of the sky.

Westinghouse will save these huge amounts of steel and copper during 1942 in its Transformer Plant alone . . . because of *radical improvements in transformer materials and design* that have come out of the Westinghouse Research Laboratories and Westinghouse engineering:

An entirely new kind of transformer steel wherein the crystals line up like tenpins in a row, instead of in helter-skelter fashion as in ordinary silicon steel. This new grain-oriented steel . . . called Hipersil . . . is the result of nine years of intensive study and experiment by Westinghouse Research Engineers.

Hipersil has *one third* more flux-carrying capacity than the best grade of ordinary silicon steel. Because of its higher permeability, less Hipersil core material is needed. Transformer weight may be reduced *25 to 30 per cent*. Hipersil, alone, will save thousands of tons of critical steel and copper in the vast number of transformers Westinghouse will make during 1942.

Forced oil-cooled transformers in which the cooling medium is rapidly circulated inside the transformer shell. This means quicker heat removal. Smaller cores and coils can do a bigger job. Strategic steel and copper are saved to help win the war.

Copper temperature control. Here a liquid-cooled transformer carries a load up to the point where the copper reaches a critical temperature, beyond which the insulation may be damaged. By the use of this thermal control, higher safe loads can be carried. A transformer of a given size, therefore, does *more work* . . . another way of saving steel and copper for more shells and tanks and ships to hurl at the Axis.

Better transformers go hand in hand with these major achievements of Westinghouse "know how" in saving strategic materials for our all-out drive to victory.

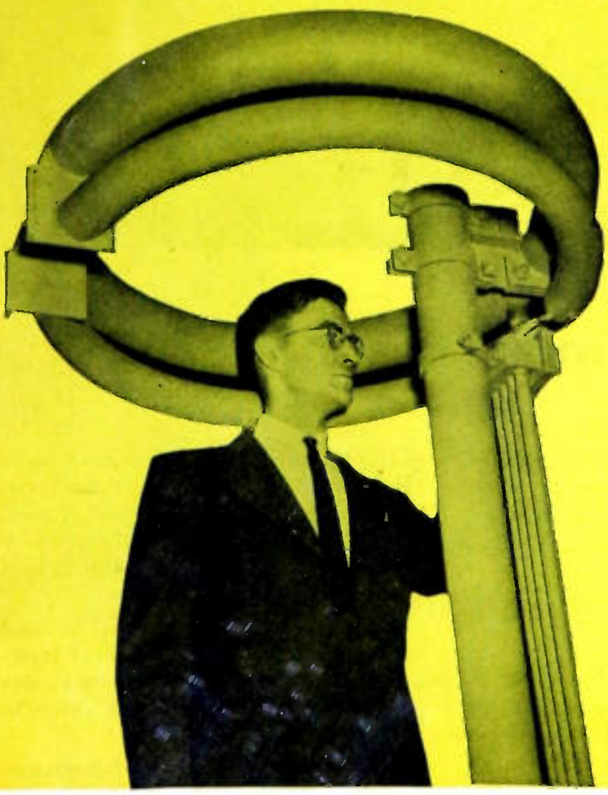
Westinghouse Electric & Manufacturing Company, Pittsburgh, Pennsylvania. Plants in 25 cities. Offices everywhere.

You are cordially invited to visit the Westinghouse Electrical Exhibit at the Annual Convention of the American Association for the Advancement of Science . . . December 28-31st . . . at the Commodore Hotel, New York.

Westinghouse

. . . making Electricity work for Victory

What do THEY think of FM?



This new antenna by G. E., with circular bays, eliminates the usual complex, costly structure, yet radiates energy uniformly. It is an example of General Electric engineering leadership in FM equipment.

G.E. WENT TO THEM AND FOUND OUT!

AMONG owners of frequency-modulation receivers, a large majority like the quality of FM reception. For example, 85 per cent say it is better than regular broadcast reception, and 91 per cent would recommend it to their friends!

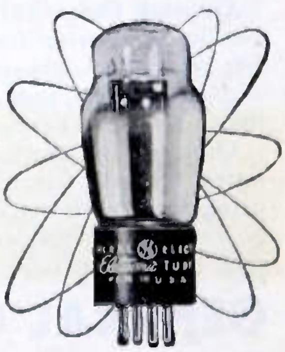
These are facts and figures taken directly from a survey made for General Electric in 14 cities by an independent research organization. Among owners of General Electric FM receivers, the approval registered was even greater.

FM Receivers
NO OTHER

The research organization went directly to private homes for its findings. It sought and obtained answers from both FM and non-FM owners of high, medium, and low cost sets. The answers took on a pattern of telling significance.

Seventy-eight per cent of the non-FM owners rated virtual freedom from static and better tone quality as the outstanding FM advantages. Eighty per cent of FM owners emphasized these same advantages also.

Today G. E. is building FM transmitting and receiving equipment for war purposes only, with the same precision and skill that characterize all of its electronic devices. When peace comes, General Electric FM equipment will be more than ever the best that money can buy! . . . For detailed information on the FM survey, write for booklet, entitled "What the Consumer Thinks of FM," to Radio, Television, and Electronics Department, General Electric, Schenectady, N. Y.



FM Broadcast Apparatus • FM Broadcasting • FM Police Radio • FM Military Radio

MANUFACTURER OFFERS SO MUCH FM EXPERIENCE

FM

GENERAL ELECTRIC

180-A2-6910

FOR *Better* RESULTS USE TUBES WITH GENERAL CERAMICS STEATITE INSULATORS

For long tube life at high frequency and temperatures, depend on tubes with

GENERAL CERAMICS STEATITE INSULATORS

Two major advantages over ordinary commercial steatites are:

1. **Low surface conductivity and contamination.**
2. **Good insulating quality at high operating temperatures.**

As vacuum tubes increase in power, more and better insulation becomes necessary. Glass insulation is unsatistactory at high temperatures and high frequencies. Natural lava from Sicily has the essential insulating qualities but is unobtainable.

General Ceramics has succeeded in producing a steatite that, as an insulator, is equal to natural lava, but requires no expensive Carboloy tools for machining. In addition, these insulators are made of domestic materials and therefore are available in quantities.

Outstanding electronic tube manufacturers have tested General Ceramic's Steatite Insulators and are using them with marked success. These insulators can be supplied in pressed or extruded shapes for every type of vacuum tube requirement.

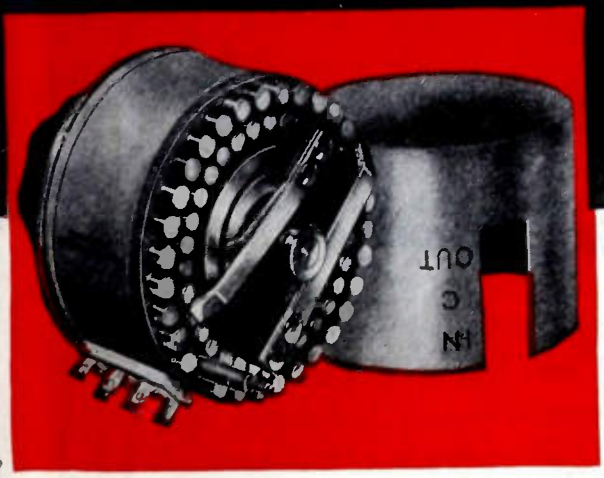
GENERAL CERAMICS AND STEATITE CORPORATION

KEASBEY NEW JERSEY

3346


RCA
813
Radiotron

U. S. Marine Corps Photo



STURDY

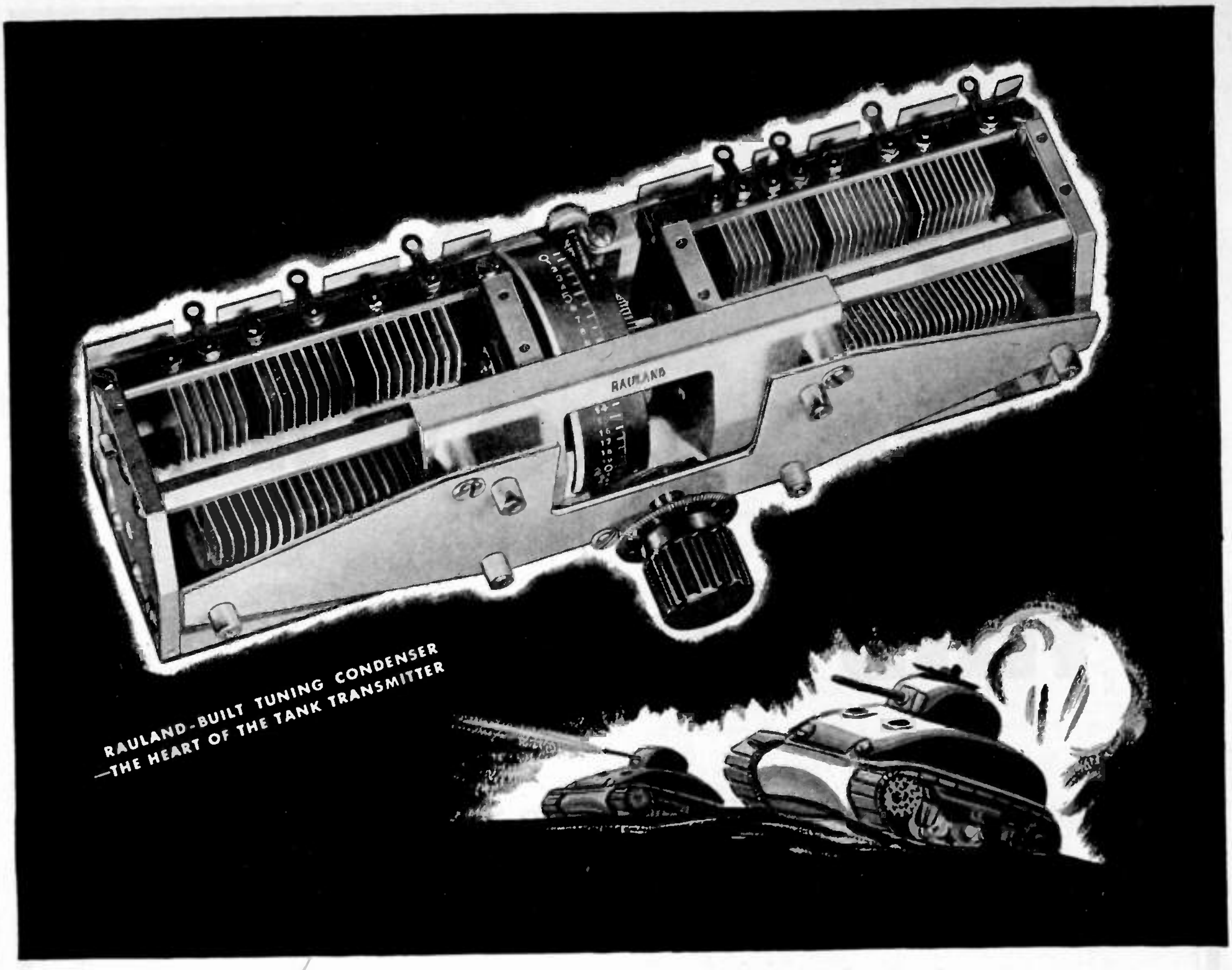
Our steel-nerved fighting men are making valuable contributions to world freedom in the strengthened battle lines of the United Nations.

With equal ruggedness, DAVEN attenuators, in actual combat zone equipment or in war production operation, are meeting the most critical standards for accurate and consistent performance.

A DAVEN catalog should be in your reference files. We list the most complete line of precision attenuators in the world; "Ladder," "T" type, "Balanced H" and potentiometer networks—both variable and fixed. Also, more than 80 models of Laboratory Test Equipment as well as Super DAVOHM precision type wire-wound resistors, with accuracies from ±1% to ±0.1%. A request will bring this catalog to you.

THE DAVEN COMPANY

158 SUMMIT STREET NEWARK, NEW JERSEY



"Taking it" . . . together

Communications...are key factors in every movement and action of the tank in combat. Orders must be received and sent right through a tortuous mixture of mechanical noise and artillery thunder . . . and these orders *must get through!* Time, at such moments is precious as life and RAULAND shortwave equipment is *Electroneered* to meet Uncle Sam's demand that communications for tanks be as "tough and dependable as the tank itself." They must have "the stuff" to stand up under all the shocks and hard treatment sustained in battle action.

To make RAULAND communication transmitters even more dependable, only RAULAND *Electroneered* tuning condensers are used. They are designed and built to minutely controlled variations and a fine degree of tuning . . . and this is maintained through the roughest periods of tank maneuvers and battle operations.

• *Electroneering* is our business •

Rauland

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The Rauland Corporation . . . Chicago, Illinois

Buy War Bonds and Stamps! . . . Rauland employees are all doing their part as members of the 10% savings club.

THE INSTITUTE serves those interested in radio and allied electrical communication fields through the presentation and publication of technical material.

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The Institute of Radio Engineers, Inc.
Harold P. Westman, Secretary
330 West 42nd Street
New York, N.Y.

The Institute of Radio Engineers

Incorporated

330 West 42nd Street, New York, N.Y.

To the Board of Directors

Gentlemen:

I hereby make application for ASSOCIATE membership in the Institute of Radio Engineers on the basis of my training and professional experience given herewith, and refer to the sponsors named below who are personally familiar with my work.

I certify that the statements made in the record of my training and professional experience are correct, and agree if elected, that I shall be governed by the Constitution of the Institute as long as I continue a member. Furthermore I agree to promote the objects of the Institute so far as shall be in my power.

....
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(Date)

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(Signatures not required here)

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Month During Which Application Reaches I.R.E. Headquarters	Amount of Remittance (=Entrance Fee+Dues) Which Should Accompany Application	
Grade →	Associate	Period Covered by Dues Payment
Jan., Feb.	\$7.50 (=\$3+\$4.50*)	Apr.-Dec. (9 mo. of current year)†
Mar., Apr., May	6.00 (= 3+ 3.00*)	July-Dec. (6 mo. of current year)†
June, July, Aug.	4.50 (= 3+ 1.50*)	Oct.-Dec. (3 mo. of current year†
Sept., Oct., Nov.	9.00 (= 3+ 6.00*)	Jan.-Dec. (entire next year)
Dec.	7.50 (= 3+ 4.50*)	Apr.-Dec. (9 mo. of next year)†

† You can obtain the Proceedings for the entire year by including with your application a request to that effect and a remittance of \$9.00.

* Associate dues include the price of the Proceedings, as follows: 1 year, \$5.00; 9 months, \$3.75; 6 months, \$2.50; 3 months, \$1.25. This may not be deducted from the dues payment.

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To Qualify for Associate membership, an applicant must be at least 21 years of age, of good character, and be interested in or connected with the study or application of radio science or the radio arts.

An Application should be filed, preferably on blanks obtainable on request from I.R.E. Headquarters or from the secretary of your local Section. If more convenient, however, the accompanying abbreviated form may be submitted. Additional information will be requested later on.

Sponsors who are familiar with the work of the applicant must be named. There must be three, preferably Associates, Members, or Fellows of the Institute. Where the applicant is so located as not to be known to the required number of member sponsors, the names of responsible nonmembers may be given.

Entrance Fee and Dues: The Associate entrance fee is \$3.00. Annual dues are \$6.00 per year, which include the price of the Proceedings as explained in the accompanying remittance schedule.

Remittance: Even though the I.R.E. Constitution does not require it, you will benefit by enclosing a remittance with your application. We can then avoid delaying the start of your Proceedings.

Your Proceedings will start with the next issue after your election, if you enclose your entrance fee and dues as shown by the totals in the accompanying remittance schedule. Any extra copies sent in advance of the period for which you pay dues (see last column) are covered by your entrance fee.

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Receipt Acknowledged .... Elected .... Notified ....

4111

Product of

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Skills in Electronics

"UNITED" electronic power tubes cannot be spun out on swift, automatic assembly lines. The painstaking manufacturing of these sensitive devices requires the skill of human hands.

Here at the "United" Plant, incredibly accurate hands perform under a system of personal supervision by electronic engineers. One by one, the steps of forming and fitting the stems, leads, plates, grids, wires and rods combine to produce transmitting tubes of such flawless precision that they consistently win top rating for performance. Never before were the hands of craftsmen and the brains of scientists so superbly "United" in advancing the scope and purpose of electronics.

Consistent technical advances in tubes, now required for war, some day will be more readily available to you for radio communication, physiotherapy and industrial electronics. Remember to look for "United" on the tubes.

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Johnson, J. F., 4316 Whitman Ave., Seattle, Wash.
Maser, H. T., General Electric Co., Schenectady, N. Y.
McFarlane, M. D., Massachusetts Institute of Technology, Cambridge, Mass.
Piore, E. R., 325 Central Park West, New York, N Y.
Wesser, C. H., 3441—21 St., Wyandotte, Mich.

Admission to Member

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Carpenter, D. D., 269 Somerset West, Ottawa, Ont., Canada
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Marathe, D. H., Radio Electric Ltd., Lamington Chambers, Bombay, No. 4, India
McArthur, E. D., Electronics Lab., General Electric Co., Schenectady, N. Y.
Neiman, M. S., 1026—17 St., N.W., Washington, D. C.
Scully, C. T., 59 Sevenoako Way, St. Pauls Cray, Kent, England
Simons, A. H., 54 Western Elms Ave., Reading, Berks, England
Van Dissel, G. F., 30 Fifth Ave., New York, N. Y.
Wallis, C. M., University of Missouri, Columbia, Mo.
Wysotzky, M. Z., 3355—16 St., N.W., Washington, D. C.

Admission to Associate

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Amsler, E. F., 861 University Ave., Palo Alto, Calif.
Arnold, H. H., Keota, Iowa
Betterton, A. R., 2021 Longwood St., Baltimore, Md.
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Bodle, D. W., 463 West St., New York, N. Y.
Brandt, R. H., Newman Springs Rd., Red Bank, N. J.
Breetz, L. D., Naval Research Laboratories, Anacostia Station, D. C.
Brodsky, S., 507 Crawford St., Toronto, Ont., Canada
Broughton, G. E., Sixth Signal Service Co., Fort Custer, Mich.
Cameron, E. G., 336—23 Ave., San Mateo, Calif.
Cantelon, E. A., 14833 Cloverdale Ave., Detroit, Mich.
Carroll, R. L., West Virginia University, Morgantown, W. Va.

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Membership

(Continued from page xviii)

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Kercher, T. C., Atlantic Fleet Schools—Radio—Naval Operating Base, Norfolk, Va.
Knoop, W. A., Jr., 2 Main Ave., Passaic, N. J.

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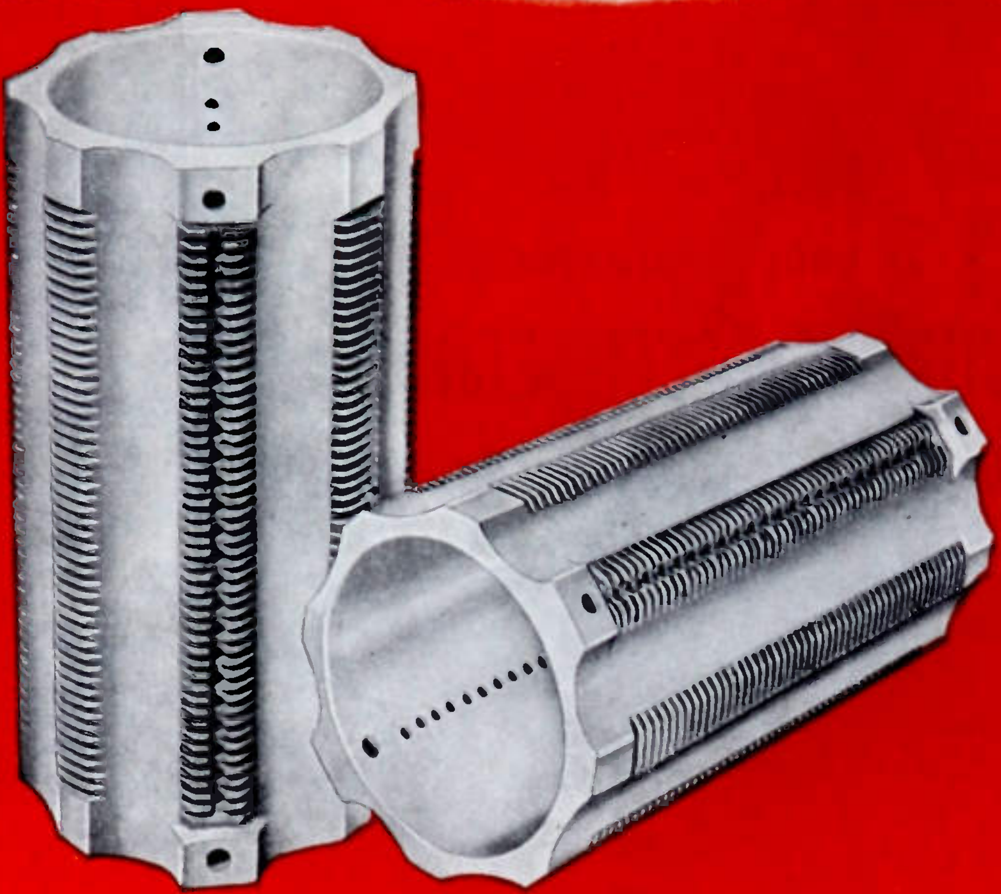

Pressed—for time

AND TIME SAVED BY PRESSING

A customer needed a large quantity of these 7½-inch coil forms — in a hurry. The normal processing called for extruding, cutting-off, threading, drilling and other machining that would have made "on time" delivery impossible as all equipment necessary for these processes was tied up for months ahead.

We could have thrown up our hands and said "Sorry" We could have found plenty of alibis. But that is not our way.

Our Engineering Department went to work. "What about pressing?" asked someone. Pressing? A piece 7½ inches long with 52 holes and eight flutes and 52 threads on each flute? A stiff problem. It had not, to our knowledge, been done before.

"All right, let's try it!"

The die was probably the most complicated one that ever came out of our tool shop.

To make a long story short — we did it, and, pardon us for saying so, we are rather proud of this achievement.

If *you* have any special steatite problems, we would like to have a shot at them.

GENERAL CERAMICS AND STEATITE CORPORATION

KEASBEY, NEW JERSEY

GENERAL

STEATITE INSULATORS

3330

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(Continued from page xx)

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Roberts, W. W., 101 Clark St., Brooklyn Heights, N. Y.
Rose, S. P., "Corton," Chignal Road, Chelmsford, Essex, England
Rosenberg, J. D., Capitol Radio Engineering Institute, 3224—16 St., N.W., Washington, D. C.
Rudner, M. A., 6236 S. Harper Ave., Chicago, Ill.
Sakoski, C., 395 Reynolds St., Kingston, Pa.
Sanderson, J. K., Dodge Radio Institute, Valparaiso, Ind.

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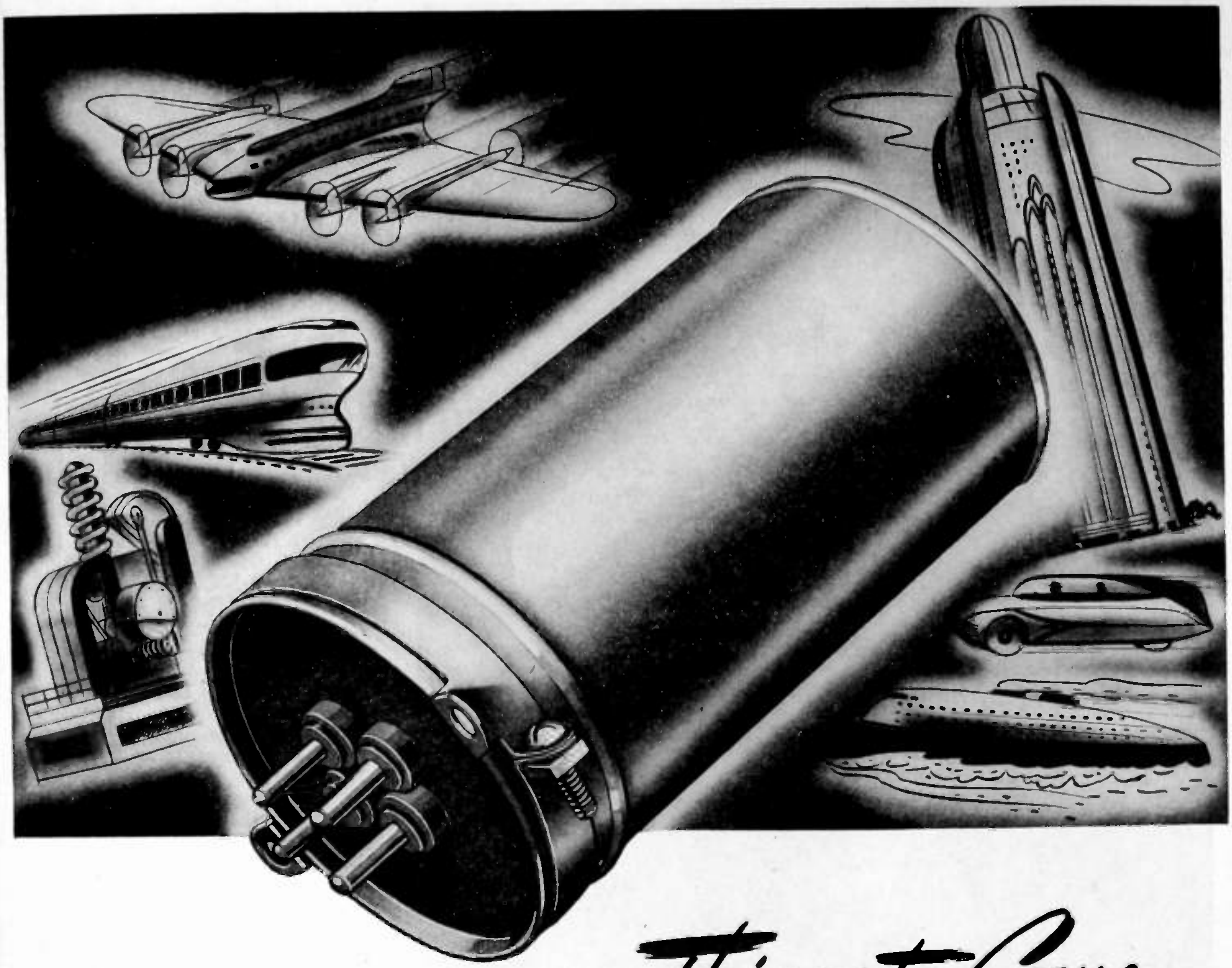

THE SHAPE OF *Things to Come*

● Simply a plug-in capacitor. True. The fact that Aerovox spent months perfecting the corrosion-proof base is beside the point here. Likewise that such capacitors—in the electrolytic, wax-filled and oil-filled types—are standard in essential wartime equipment.

The vital point is that this capacitor symbolizes "The shape of things to come." The plug-in feature denotes ready checkup and replacement. That in turn signifies continuous, gruelling, accelerated-wear service that wears out the best capacitors in months instead of in years under usual operating conditions. Just as the demountable-rim wheel marked the transition of the automobile from Sunday pleasure rides to everyday essential transportation, so this plug-in capacitor spells an infinitely expanded usage of radio technique, radio components, radio manpower.

Our first job is to win the war. Aerovox is now concentrated on just that. And while tens of thousands of radio men are engaged in waging this war, gaining invaluable training and experience and, indeed, compressing decades of normal progress into as many years, so we at Aerovox are laying the foundation for greatly expanded radio and electronic opportunities in the coming days of peace. Thus "The shape of things to come."

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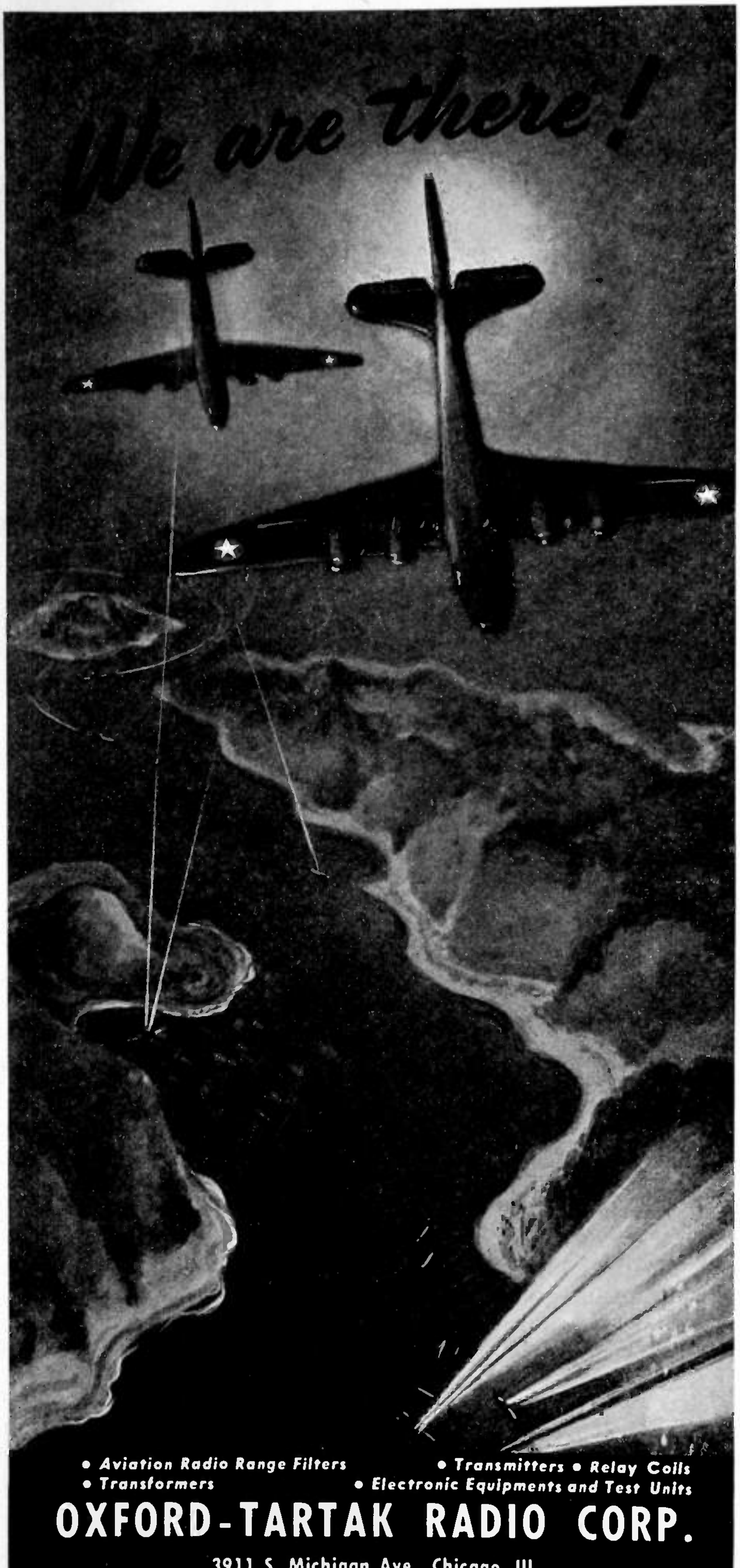

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Sigman, H. M., 6376 Sutherland, St. Louis, Mo.
Skorupski, L. J., 58-09 Van Horn St., Elmhurst, L. I., N. Y.
Smart, M. E., 228 Hawthorne Ave., North Sacramento, Calif.
Snoddy, R. E., Box 336, Raytown, Mo.
Sorensen, E. M., 6 Deer La., Stamford, Conn.
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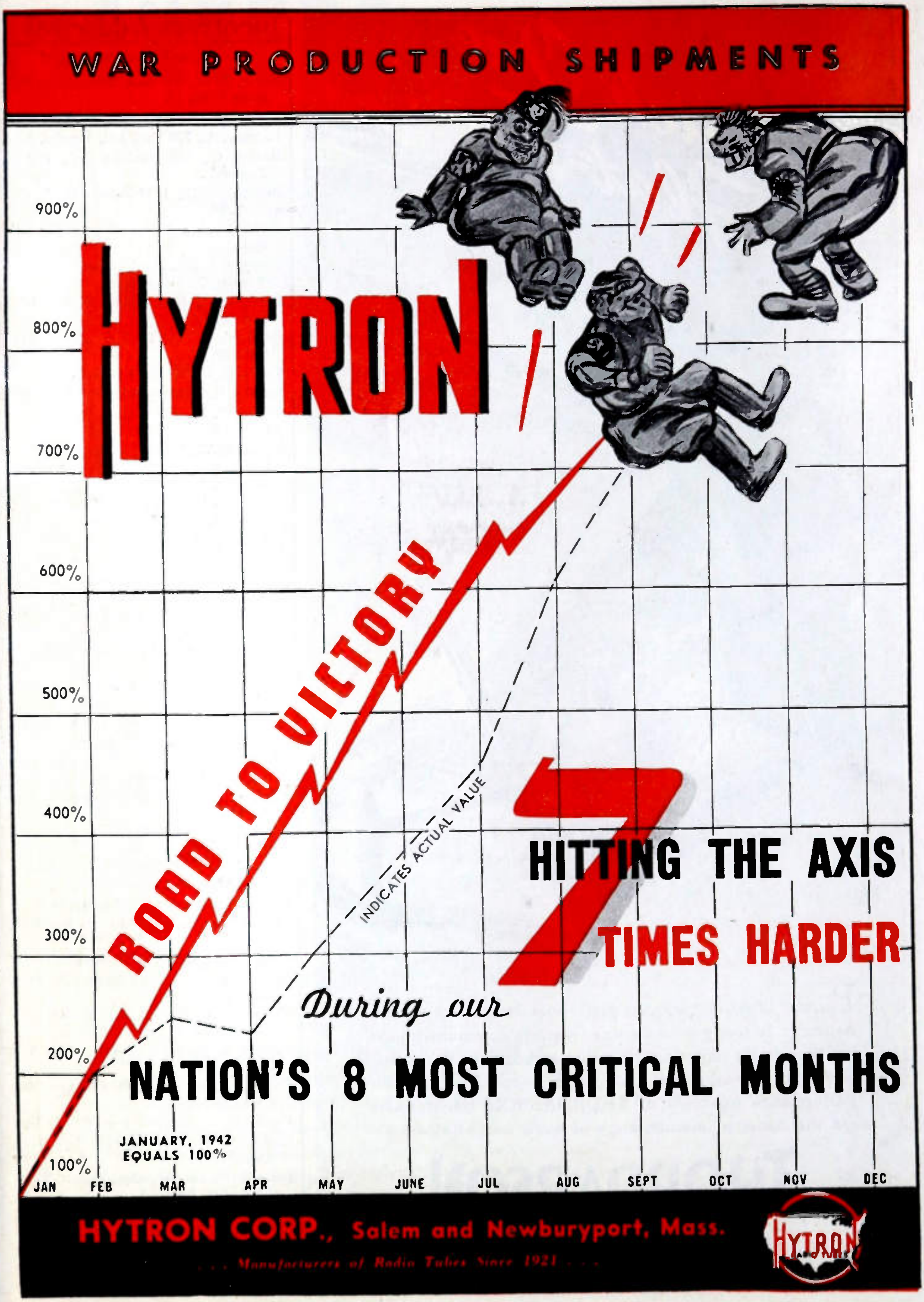
Valinet, S., 1828 Central Ave., Indianapolis, Ind.
Wagner, F. W., 1161 Cumberland Rd., Jacksonville, Fla.
Wakefield, E. H., 142 Main St., Vermilion, Ohio
White, A., Box 71, West Lynn, Mass.
Wiggins, A. M., Pens Neck Cir., R. F. D. 1, Princeton, N. J.
Wintermute, C. H., 1451 Compton Ter., Hillside, N. J.
Witschen, R. B., 824 Sixth Ave., N., St. Cloud, Minn.
Work, W. R., Carnegie Institute of Technology, Pittsburgh, Pa.
Zobel, E. R., Y.M.C.A., Grand Forks, N.D.

Incorrect Addresses

Listed below are the names and last-known addresses of the members of the Institute whose correct addresses are unknown. It will be appreciated if anyone having information concerning the present addresses of the persons listed will communicate with the Secretary of the Institute.

Alcorn, E. B., 7807 Walinca Ter., Clayton, Mo.
Armstrong, H. W., Hazeltine Service Corp., 58-25 Little Neck Pkwy., Little Neck, L. I., N. Y.
Ayer, W. E., Box 1006, Stanford University, California
Ayres, W. R., 2014 S. Duluth, Sioux Falls, S. D.
Bailey, W. B., Norfolk Navy Yard, Portsmouth, Va.
Baird, G. N., Airway Communication Station, Box 349, Rt. 1, Galveston, Tex.
Baird, H., 58 Montclair Ave., North Quincy, Mass.

(Continued on page xxvi)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Baker, A. L., 560 Tech. School Squadron, Scott Field, Ill.

Barkley, H. F., 21 Royall, Medford, Mass.

Barstow, A., 700 Park La., Topeka, Kan.

Baxter, C., 298 Banbury Rd., Oxford, England

Benson, A. E., 1532 Austin Blvd., Oak Park, Ill.

Bernier, V. A., 5211 E. 10, Indianapolis, Ind.

Bissmire, E. M., Ngaio Road, Takapuna Auckland, New Zealand

Blum, P. F., 2634 Quincy, Kansas City, Mo.

Bradford, R. C., 5517 Glenwood Ave., Chicago, Ill.

Brady, C. L., WJIM, Lansing, Mich.

Brinson, R. S., 710—21 St., Columbus, Ga.

Brown, C. R., Phi Kappa Sigma House, Orono, Me.

Brown, H. W., 726 Dudley St., Dorchester, Mass.

Burley, R. M., 1368 Academy La., Teaneck, N. J.

Buchter, T. W., Fort Monmouth, N. J.

Caird, K., 2431 N. Tripp Ave., Chicago, Ill.

Camras, M., 1418 S. Karlov Ave., Chicago, Ill.

Carpenter, R. E. H., 35 Croham Valley Rd., Croydon, Surrey, England

Carr, L. H., WBBM Transmitter, Glenview, Ill.

Carter, R. H., 924 Broad St., Mount Vernon, Wash.

Chapman, A. B., 5927 Lewis, Dallas, Tex.

Chauvin, D. M., Box 6845, University Station, Baton Rouge, La.

Chelgren, A., 2455 N. Eighth St., Milwaukee, Wis.

Chisholm, E. E., Calle 25, No. 339. Vedado, Habana, Cuba

Chow, W. T., M. I. T. Graduate House, Cambridge, Mass.

Clement, P. F., Box 627, Fresno, Calif.

Coleman, L. E., 115 Alder St., Bluefield, W. Va.

Cook, D. C., 546 N. 79 St., Seattle, Wash.

Cotsworth, A., III, 422 N. Scoville Ave., Oak Park, Ill.

Cramer, K. H., 239 Alameda St., Klamath Falls, Ore.

Cullum, A. Earl, Jr., Highland Park Village, Dallas, Tex.

Cummings, H., 505 E. Washington St., Iowa City, Iowa

Currie, W. O., Box 332, Fort Worth, Tex.

Dayot, V. D., 107 Cayuga Park Cir., Ithaca, N. Y.

Dawe, F. W., 31, Arundel Gardens, Westcliff-on-Sea, England

Debs, R. J., 113 Biddle St., Wilkinsburg, Pa.

Deisinger, D. A., 26 Wallace St., Red Bank, N. J.

(Continued on page xxviii)

Incorrect Addresses

(Continued from page xxvi)

Devaney, R. G., 216 N. Second St., Camden, N. J.

Dixon, F. C., Communications, c/o Pan American Airways, La Guardia Airport, New York, N. Y.

Doll, E. B., 2415 Good Hope Road, S. E., Washington, D. C.

Doll, H.-G., 2720 Leeland Ave., Houston, Tex.

Dorsa, S. C., 536 John Jay Hall, Columbia University, New York, N. Y.

Dryden, E. R., Jr., Y.M.C.A., Omaha, Neb.

Eames, A. M., Jr., Boothbay Harbor, Maine.

Easton, Allan, 45–40—43 St., Long Island City, L. I., N. Y.

Ehrlich, C. N., 304 W. 92 St., New York N. Y.

Estep, C. R., 5503 Harcourt Ave., Los Angeles, Calif.

Estes, R. A., Consolidated Aircraft Corp., San Diego, Calif.

Fariss, G. W., WGPC, Albany, Ga.

Fathauer, G. H., 106 E. Daniel, Champaign, Ill.

Fausett, F., 441 Hopkins St. S.W., Atlanta, Ga.

Fischer, C. F., California Institute of Technology, Pasadena ,Calif.

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Fisher, Lt. B. A., 2049 Broadway, New Orleans, La.

Fisher, G., 1961 University Ave., Palo Alto, Calif.

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Fritzinger, L. E., 1402 W. 81 St., Los Angeles, Calif.

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Gamara, N. J., 8612 Halls Ferry Rd., St. Louis, Mo.

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Granros, G. J., Box 176, Pioneer Hall, Minneapolis, Minn.

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Hamilton, Lt. O. W., Box 84, Chilmark, Mass.

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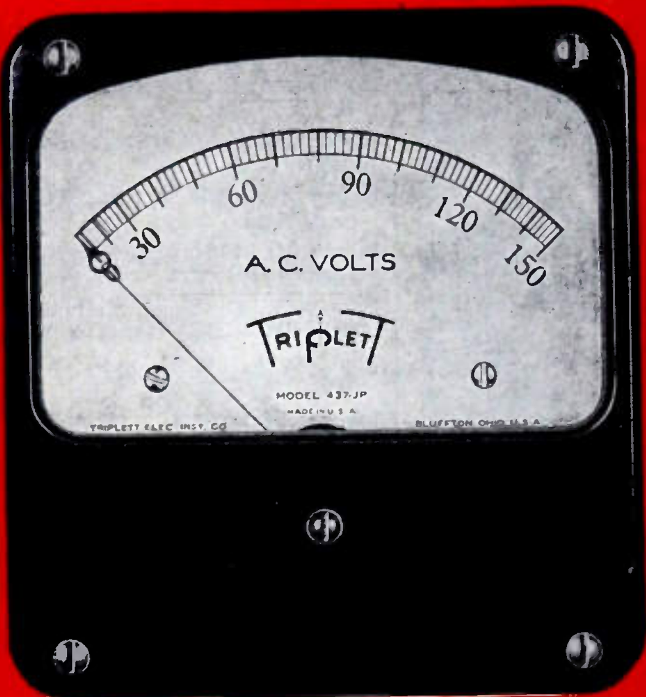

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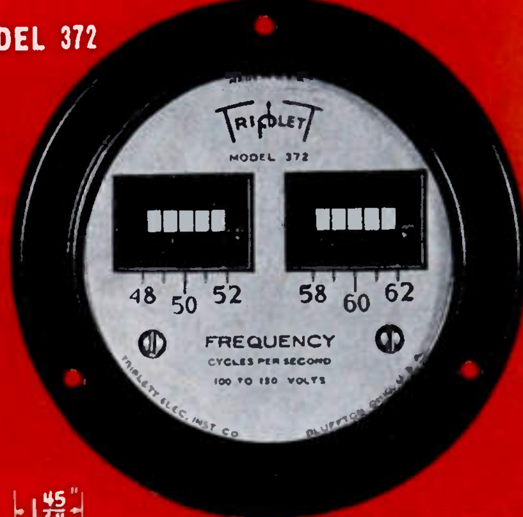

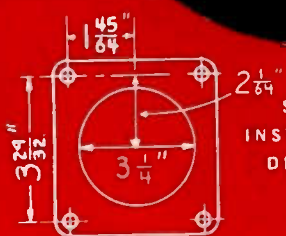

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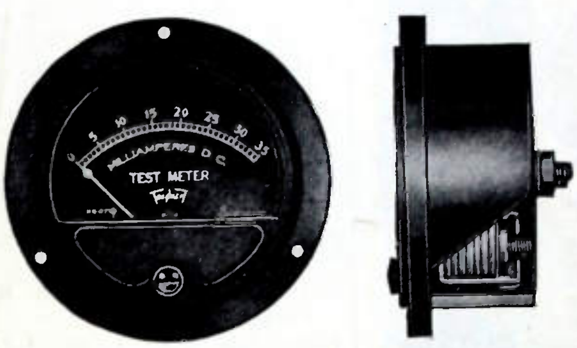

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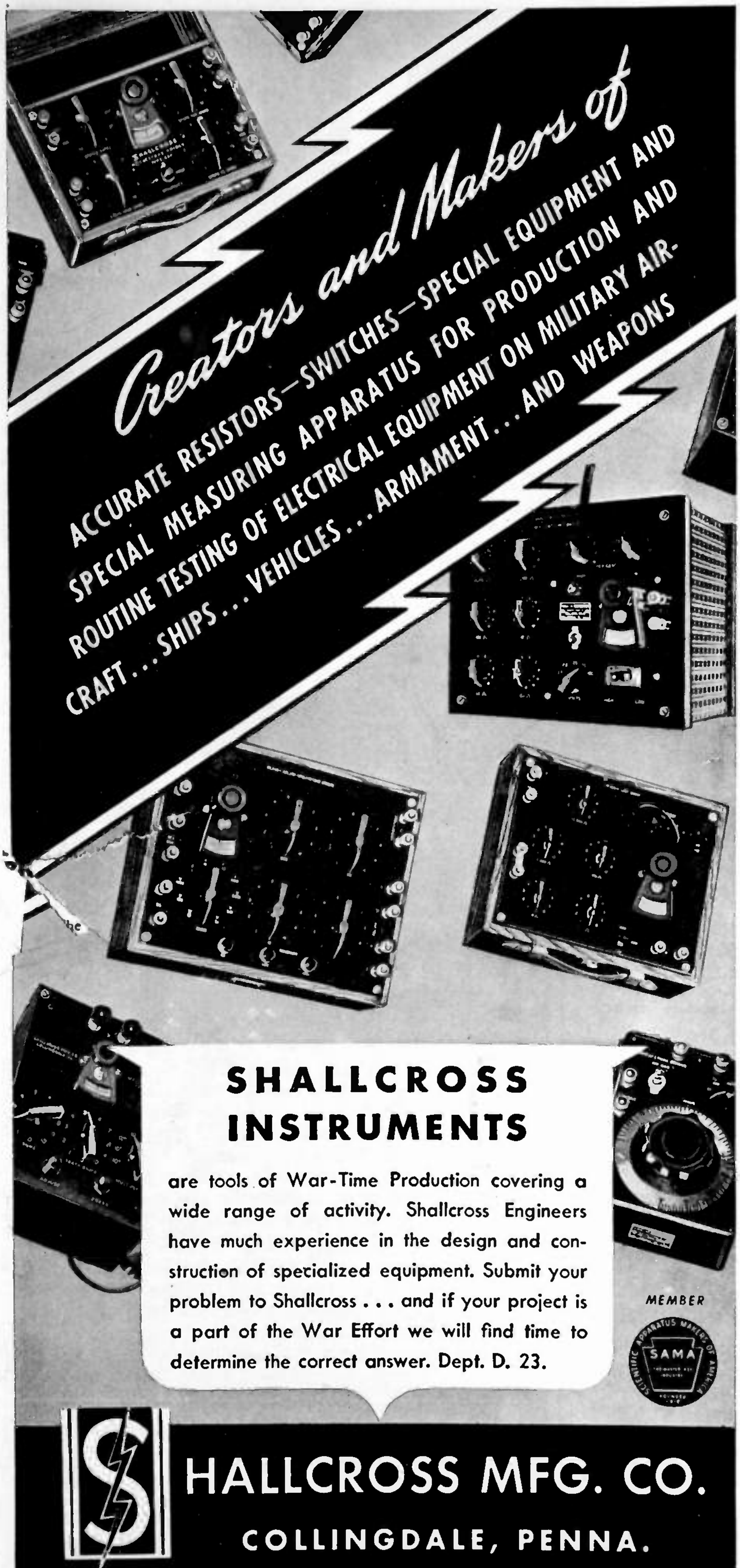

Incorrect Addresses

(Continued from page xxviii)

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McCallum, R., c/o F.M.O., Halifax, N. S., Canada

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McKinney, A., 34th School Squadron Scott Field, Ill.

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Miller, H. C., Box 157G, Rt. 1, Los Altos, Calif.

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Perry, C. S., Jr., 40 Jefferson Ave., Haddonfield, N. J.

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Radcliffe, J. C., Eagle River, Wis.

Renhard, J. A., 25 Lafayette Ave., Haddonfield, N. J.

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Respondek, A. M., The Colony House Hotel, Great Neck, L. I., N. Y.

Rettenmeyer, R. D., 380 E. 154 St., New York, N. Y.

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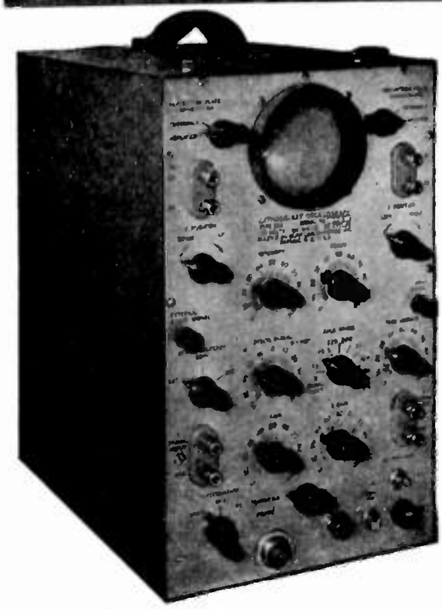

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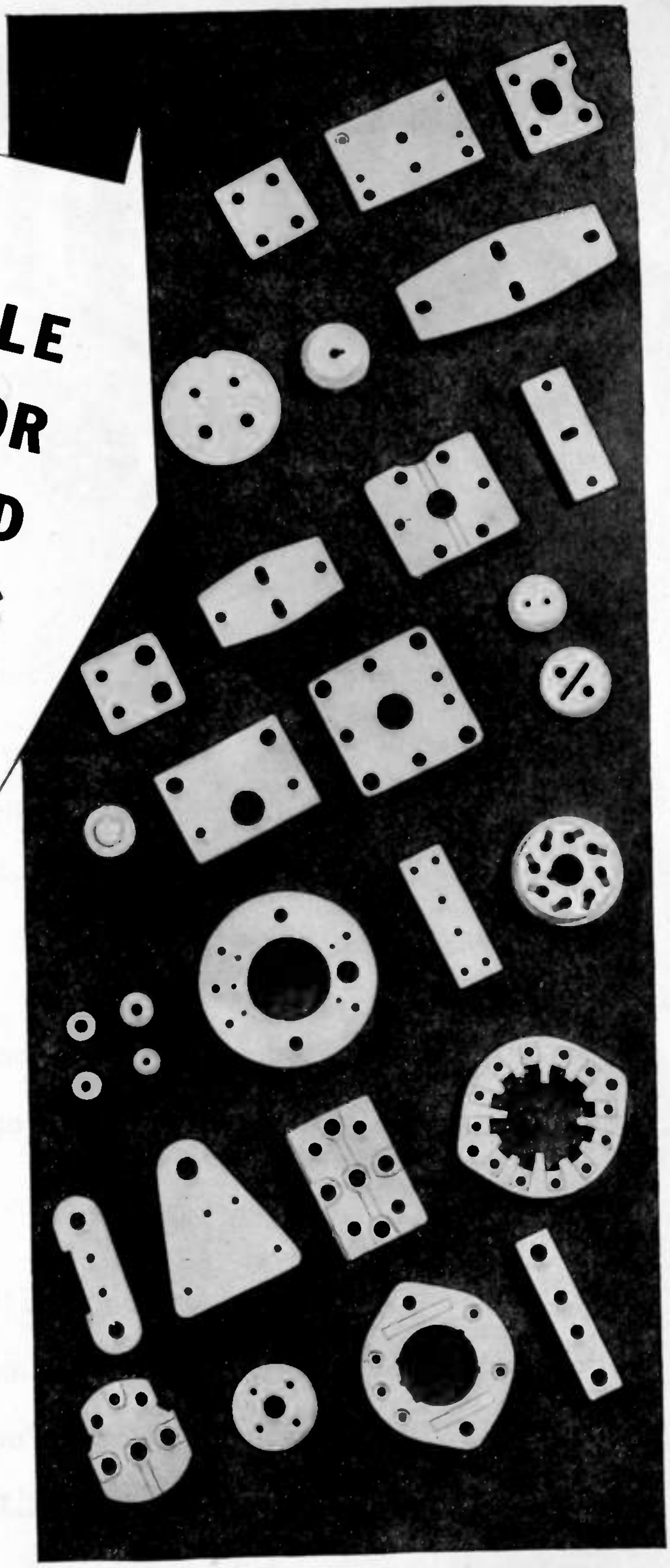

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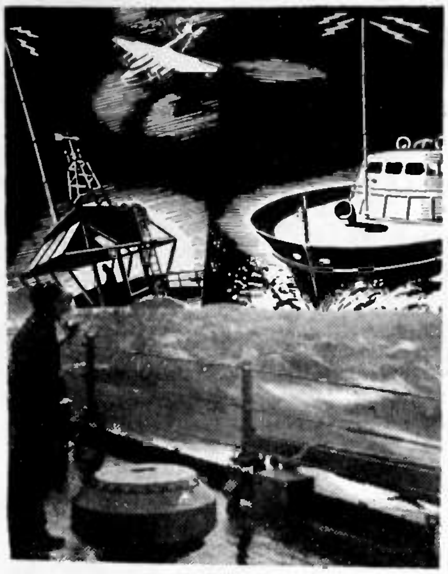

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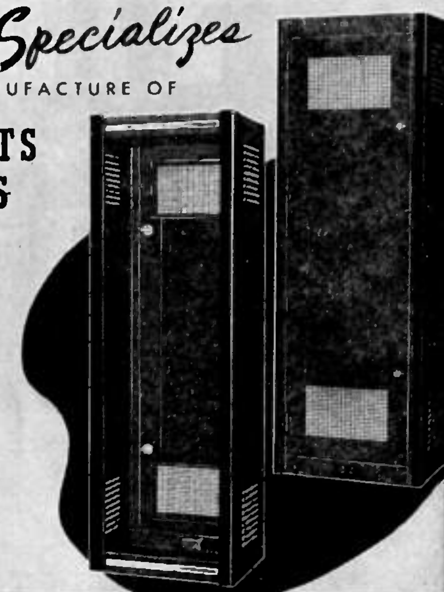

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The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

Incorrect Addresses

(Continued from page xxxii)

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Wood, W. L., 67th C. A. "AA," Fort Bragg, N. C.
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The following commercial literature has been received by the Institute.

NEW MATHEMATICAL TYPE-WRITER TYPE • • • *Ralph C. Coxhead Corporation, 333 6th Avenue, New York, N. Y.* Description given in the house organ "Vari-Typer Between Us" of a new type face available on vari-type typewriters which permit the use of 90 characters and symbols necessary in the practical application of mathematics to industry. This type face was planned for the use of Signal Corp engineers and their defense manufacturing corporations.

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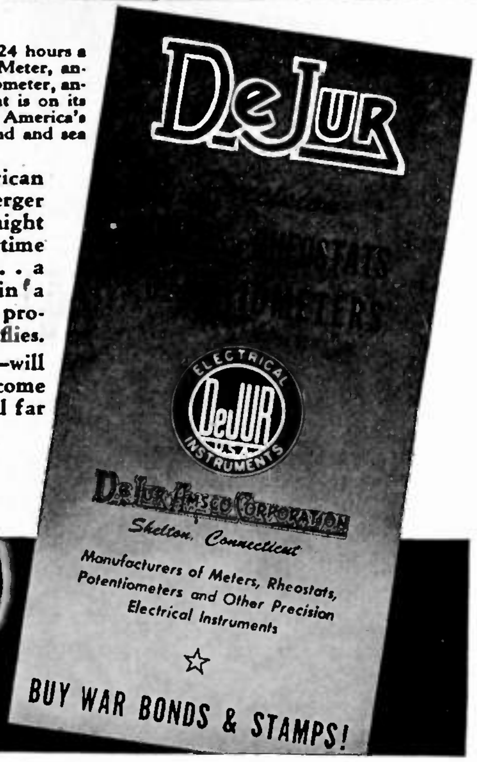

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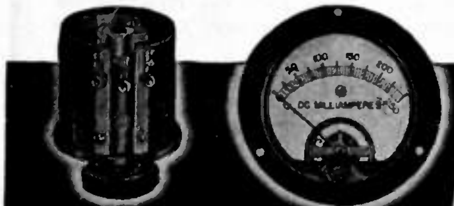

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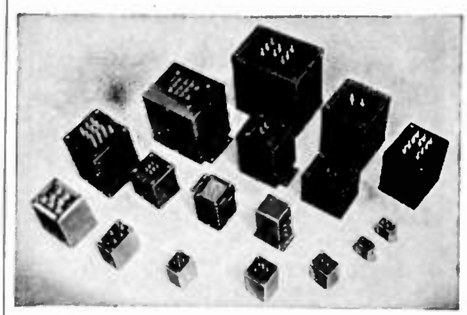

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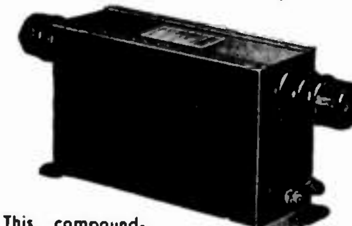

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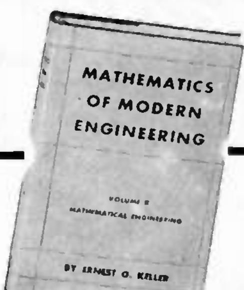

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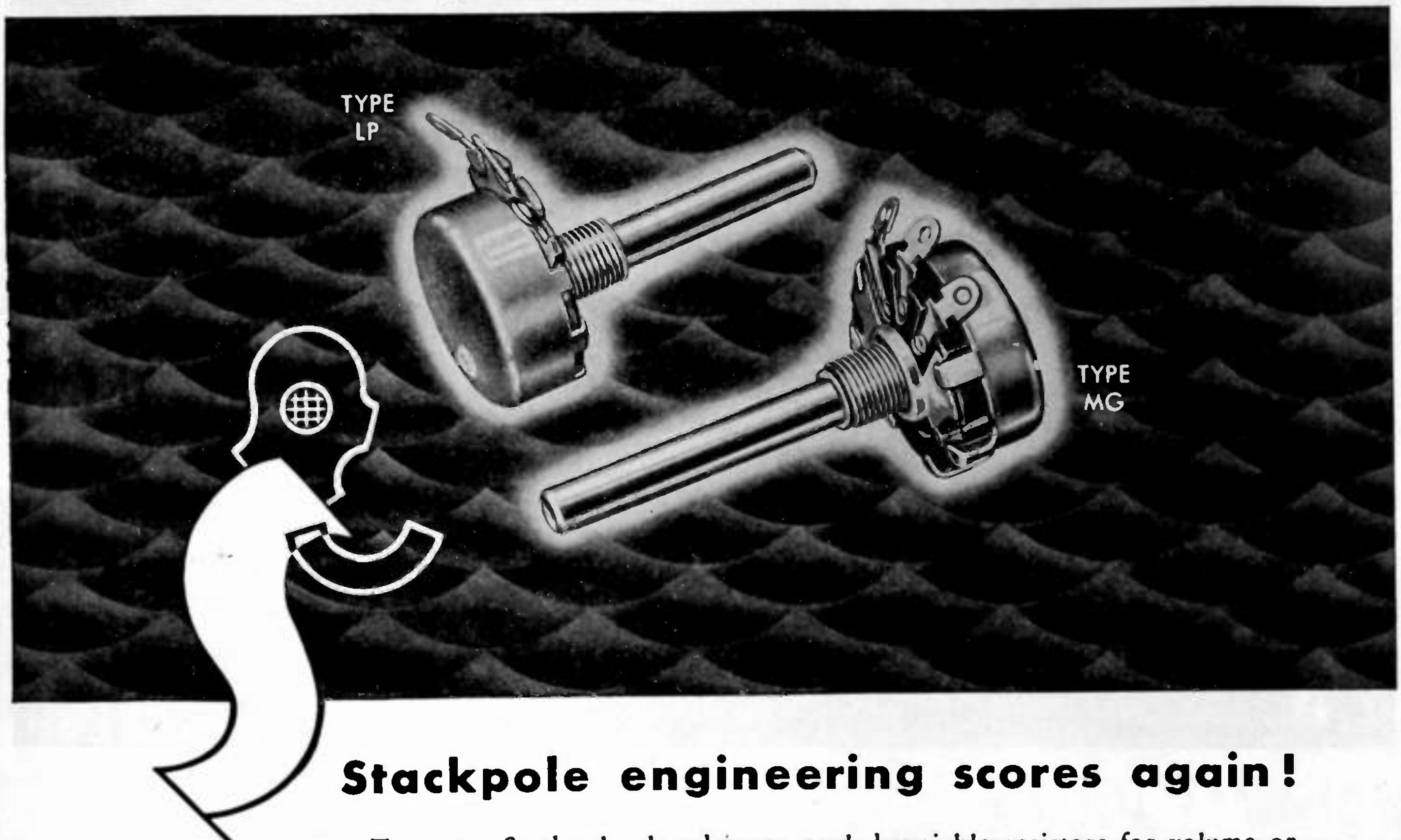

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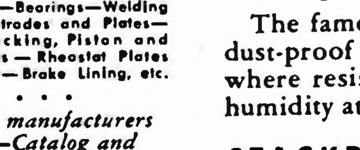

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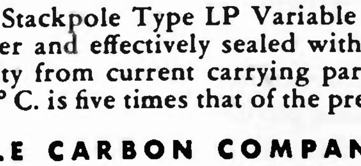

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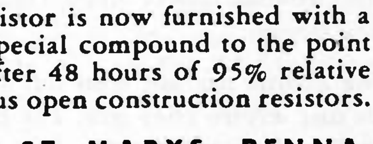

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