RCA REVIEW

A Quarterly Journal of Radio Progress

Published in July, October, January and April of Each Year by
RCA INSTITUTES TECHNICAL PRESS
A Department of RCA Institutes, Inc.
75 Varick Street New York, N. Y.

VOLUME III January, 1939 NUMBER 3

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SUBSCRIPTION:

United States, Canada and Postal Union: One Year $1.50, Two Years $2.50, Three Years $3.50
Other Foreign Countries: One Year $1.85, Two Years $3.20, Three Years $4.55
Single Copies: 50¢ each

Entered as second-class matter July 17, 1936, at the Post Office at New York, New York, under the Act of March 3, 1879.

Printed in U.S.A.

NETWORK BROADCASTING

By

David Sarnoff

President, Radio Corporation of America and Chairman of the Board, National Broadcasting Company

A Statement before the Federal Communications Commission, Washington, D. C., November 14, 1938.

MR. CHAIRMAN and Gentlemen of the Commission:—I appear before you today as President of the Radio Corporation of America and Chairman of the Board of the National Broadcasting Company. I appear in both capacities because the National Broadcasting Company is wholly owned by the Radio Corporation of America. RCA itself—and, consequently, NBC—is owned by a quarter of a million stockholders. No single stockholder, whether an individual, a corporation or a financial institution, owns as much as ½ of 1 per cent of RCA stock.

The by-laws of the corporation provide that at least 80 per cent of the voting stock shall be held by American citizens. Approximately 95 per cent of all the outstanding stock of RCA is held in the United States.

RCA and NBC Welcome Investigation

Speaking for both the companies and the stockholders I represent, we welcome this hearing and the opportunity it provides to review and appraise network broadcasting in the United States. Proud of the tremendous developments in radio during the nineteen years since the RCA was formed, we welcome this occasion when the government, the industry, and the public may jointly take stock of the present, and chart the course into a vast and unexplored future.

The questions you have presented to the National Broadcasting Company covering the specific items of this investigation will be answered factually and in detail by the executives and operating officials of NBC, who will appear before you.

Let me say at the outset that I am not here to advocate a "status quo" of broadcasting, or to oppose changes in a changing art. I do not come here to say that broadcasters are infallible, that no improvements are possible, that we have no problems, or that the Commission cannot help in their solution. I am here to aid this investigation in any way I can. I propose to describe briefly the objectives of the Radio Corporation of America which led to the formation in 1926 of the National Broadcasting Company, and to discuss the part we have taken as a pioneer in the development of the American system of broadcasting. I should like also to present my views on certain problems which at present face the broadcasting industry.

SERVICES OF RCA

The Radio Corporation of America is engaged in every field of radio. The integration of our activities provides a broad basis for radio research and experiment, for development of new products, services and methods, for cooperation with the government in times of peace or war, and for service to other radio organizations and to the public.

If I were an artist and wished to describe the business of RCA, I would draw a picture of a tree with three branches. The branches grow out of a common trunk, and their smaller branches and leaves intermingle. Then I would sketch in the part of the tree that is ordinarily unseen, the part that gives it life and growth—the root.

The three branches of RCA represent communications, manufacturing, and broadcasting. The root of the tree is research.

Research stimulates the development of radio in all its phases, just as the root provides sustenance for all branches of the tree. The ultimate results of an experiment in radio research cannot be foreseen when it is started. As a matter of fact, the principal beneficiary often turns out to be a different branch from the one for which the experiment was begun. In addition to furthering radio communications, manufacturing and broadcasting, radio research has created countless devices that are today indispensable in many fields outside the radio industry: in almost every scientific field, in numerous phases of manufacturing, in telephony, in medicine, and in public safety.

RCA RESEARCH BENEFITS ENTIRE INDUSTRY

RCA leadership in radio research is recognized in the United States and throughout the world. Our organization encounters in the field of research nothing like the consistent, able competition it meets in the operating fields of communications, manufacture, and broadcasting. Yet, the benefits of RCA research are made as readily available to competitive services as to our own, and it must be borne in mind that there is no profit in research until after its discoveries have been translated into products and services.

A case in point is television—which will add the services of radio sight to those of sound. RCA has been actively engaged in television research for more than ten years, and has spent millions of dollars to bring it to the verge of useful public service. Financially, this development is still a subject of outgo rather than income.

Nevertheless, sixty-five radio manufacturers in the United States already have been licensed by the RCA to make use of its inventions in this field. When commercial television licenses are granted by this Commission, a competitor, the Columbia Broadcasting System, using an RCA transmitter, will be able to go on the air with commercial television programs as promptly as NBC. RCA television transmitters are equally available for purchase by others who may desire to enter this field.

The many-sided developments which have grown out of radio research emphasize the interrelationship and interdependence that exist between all the services of radio. Good radio programs, for example, have created a market for six to eight million receivers a year in the United States. These receivers in turn have created the national audience which makes good programs possible. The broadcaster and the radio set manufacturer are partners in the same enterprise, and each depends on the other for his existence and his earnings.

Underlying everything we know about the transmission and reception of radio programs are the lessons we learned in the field of point-to-point radio communications. A quarter of a century of radio experience lay behind the first broadcast program, and made it possible.

When a home service of radio facsimile proves economically practical it will furnish another example of the parent-and-child relationship between our narrowcasting services and those of broadcasting. The transmission of photographs and other facsimile material by RCA Communications between the United States and Europe has become an old story; facsimile broadcasting by domestic stations for home reception promises to be one of the fascinating new chapters of radio history.

Our policies are based on the belief that the public interest, the development of the radio art, and the usefulness of the Radio Corporation of America will best be served by a strong, prosperous, and growing radio industry, and by vigorous competition which results in better service to the public and greater stimulus to the industry.

Early Days of Broadcasting

When the Radio Corporation of America was formed, nineteen years ago this fall, its immediate object was to provide an American-owned system of international communications—organization, personnel, research and technical facilities.

Soon after the formation of RCA, broadcasting made its start, and we pioneered in this new field. It captured the imagination of the public almost overnight. The devices and patents which made it possible for RCA to operate an international communications system also were required to make home receiving sets and broadcast transmitters. Experienced radio engineers were needed, and these too were available in the RCA organization.

Following the historic election broadcast in 1920, by Station KDKA in Pittsburgh, radio stations were established rapidly in other important cities. The public sat up late at night to capture the faint, elusive call letters of distant stations. It was the thrill of listening to far-off places that gave radio broadcasting its first impetus—a thrill, by the way, which reached another dramatic climax only a few short weeks ago when history was being written in Europe.

In those early days of broadcasting various organizations entered the field for the incidental advertising that could be obtained—flour mills, department stores, music shops, and even garages. Newspapers, too, were early entrants, foreseeing new possibilties for speedier communication.

RCA, and its associates in the electrical field, had a broader purpose—that of pioneering an art which would create markets for new types of radio equipment, new services, and new avenues of employment for labor and capital.

It soon became evident that the growth and permanence of radio broadcasting depended primarily on the quality and variety of programs. The novelty of tuning in distant call letters quickly wore off. It was not enough for the listener to hear a solitary piano tinkling away in a make-shift studio. Second-rate musicians began to pall, and amateur singers wore out their brief welcome.

The operation of a broadcasting station, at that time, was a matter of expense with no corresponding revenue. There was nothing to induce station owners to employ expensive professional talent, or to improve studio and station facilities for better transmission and reception.

During those early years of broadcasting, RCA operated local broadcasting stations, of which the most important was WJZ, in New York City. RCA also experimented with various station hook-ups, using telegraph lines.

After five years of hectic development, broadcasting stood at the crossroads. The alternatives were either to evolve a basis of support by private enterprise, or to seek a government subsidy, with an attendant tax on receiving sets and the natural consequence: government broadcasting.

CREATION OF NETWORKS

Fortunately for the United States, the democratic answer was found by private enterprise. In 1926, RCA purchased Station WEAF from the American Telephone & Telegraph Company, arranged to lease AT&T wire lines for interconnection with other stations, and organized the National Broadcasting Company. NBC then took over the experimental program service which the telephone company had instituted, and extended it to a group of independent stations, which—with WEAF as the key station—became the Red Network.

Network broadcasting provided greatly improved programs by tapping the talent centers of the nation and syndicating these programs over telephone lines to local, independent stations. Not only did the network system appeal to the listeners and the independent station owners, but it also attracted the business interests of the nation to the use of radio broadcasting as an advertising medium. The economic support thus developed met the needs of the three parties whose interests were at stake: the public, the station owner, and the advertiser.

To the public, the network brought a new world of ideas, of music, of enjoyment centered in the home. It turned the page to a new chapter of America's social history.

For the station owner, the network provided programs—both commercial and sustaining—of a quality he could not individually afford, and with talent not physically accessible to his station. It brought him revenue from national as well as local commercial sponsors.

To the advertiser, the network furnished a large circulation spread over a wide area. Such circulation justified, over and above the cost of station time, the talent expense of high-quality programs. It is worthy of remark that the enterprise which broadcasters have displayed in building the American system of broadcasting has been paralleled by the enterprise of the business men who so quickly recognized the advertising power of the new medium.

WHY THE NBC WAS FORMED

I cannot better describe the reasons which led to the formation of the National Broadcasting Company than to read excerpts from the announcement of its creation by RCA, published as a newspaper advertisement on September 14, 1926:

"The Radio Corporation of America is the largest distributor of radio receiving sets in the world. . . . It is more largely interested, more selfishly interested, in the best possible broadcasting than is anyone else.

"The market for receiving sets in the future will be determined largely by the quantity and quality of the programs broadcast. Today the best available statistics indicate that 5,000,000 homes are equipped, and 21,000,000 remain to be equipped. . . . Any use of radio transmission which causes the public to feel that the program is not the highest, that the use of radio is not the broadest and best use in the public interest, that it is used for political advantage or selfish power, will be detrimental to public interest in radio, and therefore, to the Radio Corporation of America.

"To insure, therefore, the development of this great service, the Radio Corporation of America has purchased for one million dollars Station WEAF from the American Telephone and Telegraph Company, that company having decided to retire from the broadcasting business.

"The Radio Corporation of America has decided to incorporate that station, which has achieved such a deservedly high reputation for the quality and character of its programs, under the name of the National Broadcasting Company, Inc. The purpose of that company will be to provide the best programs available for broadcasting in the United States. The National Broadcasting Company will not only broadcast these programs through Station WEAF, but it will make them available to other broadcasting stations throughout the country so far as it may be practical to do so, and they may desire to take them. It is hoped that arrangements may be made so that every event of national importance may be broadcast widely throughout the United States.

"The Radio Corporation of America is not in any sense seeking a monopoly of the air. If others will engage in this business we will welcome their action, whether it be cooperative or competitive. The necessity of providing adequate broadcasting is apparent. The problem of finding the best means of doing it is as yet experimental. The Radio Corporation of America is making this experiment in the interest of the art and the furtherance of the industry."

I would call your attention particularly to two sentences in this announcement, written twelve years ago:

First, "The National Broadcasting Company will not only broadcast these programs through Station WEAF, but it will make them available to other broadcasting stations throughout the country."

Second, "The RCA is not in any sense seeking a monopoly of the air. If others will engage in this business we will welcome their action, whether it be cooperative or competitive."

Growth of Networks

As soon as our formation of a national broadcasting company was announced, independent station owners, local civic organizations, and community leaders from every section of the United States wrote, telephoned or called in person to ascertain how soon network programs would be brought to their communities. To meet the popular demand represented by these requests, NBC rapidly expanded the experimental hook-ups of the Red Network into a regular service arrangment, providing programs to leading cities of the United States.

It quickly became apparent that a single network service was not enough to satisfy the demands of the radio audience for diversified programs of national interest and importance; that if broadcasting were to be popularized to all, there should be more than one type of program simultaneously available to listeners. Other station owners, particularly in the cities where their competitors had made program service arrangements with the Red Network, pressed for network affiliations. Therefore, in less than two months after the first NBC network began service, we created a second network—the Blue—with WJZ, New York, as the key station. As the networks were expanded, stations in remote, thinly populated areas, that could not be expected to bring the NBC a profit, were added, in the interests of a truly comprehensive national service.

Competitors followed NBC into what is now a leading industry. In addition to the network services offered by NBC, other companies later established transcontinental and regional networks, which now serve the nation.

Looking back at the amazing development of network broadcasting in a little more than a decade, it can be seen that our pioneering undertakings in 1926 were fully justified. While we were confident of the public demand for good programs, and were successful in creating a market for radio receivers, we could not be certain of meeting the cost of these programs by radio advertising. The substantial investments which have made the American broadcasting system what it is were made because radio proved itself an important medium in the highly competitive field of advertising.

It is largely owing to network broadcasting that radio in the United States has grown into a billion dollar industry, which today provides employment for hundreds of thousands of persons. It is estimated that the American public has invested more than three billion dollars in receiving apparatus and spends a billion hours a week listening to the radio.

But the importance of broadcasting cannot be measured in statistics or dollars and cents. It must be appraised by the effect it has upon the daily lives of the people of America—not only the masses who constitute a listening audience numbered in tens of millions, but the sick, the isolated, and the under-privileged, to whom radio is a boon beyond price. The richest man cannot buy for himself what the poorest man gets free by radio.

Freedom of Radio

The American people have a free radio because they have a broadcasting industry that pays its own way. Those who object to commercial announcements on the air are apt to forget that it is the revenue from these announcements which makes it possible for them to hear regularly a symphony orchestra conducted by Toscanini, the broadcasts of the Metropolitan Opera, America's Town Meeting of the Air, the National Farm and Home Hour, the Damrosch Music Appreciation Hour, and many other costly sustaining features of the networks. A single radio performance of any one of these programs would be an event of outstanding importance in other countries. I think I am making a simple statement of fact when I say that the people of the United States are provided with the finest and most varied radio programs produced anywhere in the world. And our traditional liberties have been fortified with a new freedom—freedom of radio—which takes its place with our older freedoms, of religion, speech, and press.

These are days when democracy is being subjected to attack from without, and to doubt from within. Yet there is no other form of government under the sun in which the elements which we consider the most essential and the most precious to American life are allowed to exist. In the dictatorships of the world the freedoms of religion, of speech, of the press, and of radio have been destroyed.

If any illustration is needed of the effect of undemocratic controls over broadcasting, it is supplied by the autocracies of the Old World where broadcasting has been converted into the most powerful instrument of dictatorship.

There, certain governments now tell their people which programs they may hear and which they may not. In some parts of Europe, to listen to a radio program originating in another country is to invite a jail sentence. I am told that in one country authorities are now discussing the prohibition of all radio reception by their citizens. The only type of receiver that would then be allowed in the home would be one limited to receiving government programs transmitted over telephone or electric light wires. On the other hand, it is apparently not contemplated that transmission of propaganda programs originating in that country would stop. It would go on—aimed like a machine-gun at the people of other nations. The objective is to permit export, but not import of ideas.

I was in Europe during the first half of the recent crisis; then, after a week on the Atlantic, I was at home during the final critical days. In order to get the full news of Europe while abroad I had to listen to programs sent by short-wave from the United States. American listeners were better and earlier informed on events in Europe than the Europeans themselves. Seventeen minutes after the completion of the Munich Conference, NBC had the terms of the Four-Power agreement on the air. I have read grateful letters from citizens of European countries commenting on this historic broadcast, saying that it brought them their first relief from fearful tension.

Human liberty was not lost in those countries through any lack of desire on the part of the individual to be free. It was lost through his blindness to the forces that enslaved him, and through his failure to cherish and protect the institutions that would have kept him free.

In this time of world crisis, it is of vital importance that every American citizen should recognize, in the freedom of our American system of broadcastng, one of the essential guarantees of his own personal freedom.

Regulation of Broadcasting

The creation of this American system of broadcasting, however, has not been achieved without difficulties and problems. The problems that touch the public interest are of two kinds: those relating to technical facilities, and those relating to programs. With respect to the regulation of facilities, the powers of the Commission are adequate, clearly expressed by law, and understood by broadcasters.

When we consider the technical development of radio we must remember that radio has never ceased to be a pioneer. The day may come eventually when its pioneering work is over, but it is a day I do not expect to live to see. Whenever we seem to have learned to extract the utmost usefulness from one portion of the radio spectrum, another part of the band looms up—first in theory, then as a subject for experiment, and finally as a practical medium of public service.

If wavelengths were now available for an unlimited number of broadcasting stations, the only limitation would be that of public acceptance. The same holds true of networks. As radio science learns to employ new channels in the ether—to use waves measured in centimeters and millimeters—the day will come when there will be more wavelengths available than stations and networks to use them.

The time is coming—and it may come sooner than anyone expects —when the present-day facilities and services of radio will prove small in comparison with the unlimited technical and artistic achievements possible in this young and swiftly-moving industry. Television, to name but a single example, stands today where sound broadcasting stood 18 years ago. With all that we have learned, is there any man who would say that television will not go farther in the next 18 years than sound radio has gone up to the present day?

With whatever technical controls broadcasting is clothed, they must be kept as flexible, as capable of expansion, as the industry itself. The situation is like that of a growing boy and his breeches. The breeches have got to have wide seams, so they can be let out when they get tight. Otherwise something is going to give way, and it seems to be a law of Nature that it won't be the boy. He just keeps growing.

When we turn to the realm of program service, however, we meet a broader question than is involved in the regulation of technical facilities. Here we deal with a vital force, a great servant of mankind when used properly, but, when abused, capable of destroying human rights. It is the social impact of radio which has raised the all-important question of social responsibility.

The Communications Act provides that your Commission shall have no power of censorship over radio programs, and that you shall adopt no regulation which interferes with the right of free speech. Therefore we must find within the broadcasting industry itself a solution which will adhere to American traditions, and at the same time meet this social responsibility.

The broadcasting industry was gratified to hear Chairman Frank R. McNinch state so clearly in his nationwide broadcast last Saturday evening—and I quote his words—"Obviously the power of censorship and selection must be lodged somewhere; and the broadcaster is the one to exercise this power and answer to the public for the manner in which he exercises it."

Recommendation for a Voluntary System of Self-Regulation

The record of network broadcasting in America proves the efforts that have been made here to safeguard public interest, to advance culture, and to provide unbiased news and wholesome entertainment. In spite of its youth and the great complexity of its problems, the industry can take pride in its accomplishments in this respect.

In the National Broadcasting Company we have our own code of program policies, formulated over a period of twelve years. It is based not only on our own operating experience, but also on the wisdom and advice of the Advisory Council of NBC. This council is composed of public-spirited men and women of high standing and wide experience. They represent education, religion, social welfare, music, labor and industry. The Council was formed at the time of the organization of the company, and has been in existence ever since. In following this code, the NBC has had to face objections from groups and individuals whose ideas and wishes ran counter to its standards. Living up to the code has also entailed the sacrifice of commercial revenue.

Other networks, and individual stations as well, have program codes of their own. The National Association of Broadcasters has a Code of Ethics adopted in 1935.

But the time has come for more positive action.

The fate of broadcasting in other nations and the attacks on democracy throughout the world clearly indicate the necessity for finding a democratic solution for the problems of the American system of broadcasting,—a solution which on the one hand, will enable us fully to meet the social obligations of radio, and on the other, will protect our traditional freedoms.

I would therefore like to take this opportunity to advocate to the broadcasting industry that it establish a voluntary system of self-regulation in its field of public service, and that it take the necessary steps to make that self-regulation effective.

My recomendation is that the experience of the different groups within the industry should now be combined and correlated. An industry code should emerge that advances beyond all previous standards. Such a code should be an act of voluntary self-regulation on the part of the entire broadcasting industry in the United States.

In writing this code, the industry should gather the views of broadcasters, of groups representative of public opinion, and of this Commission. After the code is formulated the public should be made thoroughly familiar with it. All broadcasting networks and stations should be invited and encouraged to adopt it. The code should be subject to periodic review by the industry, and kept up to date. It should be administered by a suitable agency representative of the entire industry.

I make this recommendation in the belief that such self-regulation is the American answer to an American problem. In every consideration of radio broadcasting, the "public interest" we are pledged to serve is that of the entire nation. This public interest is reflected directly by the 27,000,000 receiving set owners who represent an overwhelming majority of the country's homes. By their control of the nation's radio dials they give approval or disapproval to radio programs, and decide the ultimate fate of the broadcaster. Here we find legitimate censorship by public opinion.

It is the democratic way in a democratic country.

NEW TELEVISION AMPLIFIER RECEIVING TUBES*

By

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Summary: *Television circuits require amplifying tubes of high grid-plate transconductance and low input and output capacitances to realize a voltage gain per stage sufficient to keep the amplifying stages at a reasonable number. To this end the 1851 and 1852, sharp cut-off, 9000-micromho grid-plate transconductance tubes—and the 1853, semi-remote cut-off, 5000-micromho grid-plate transconductance tube were developed. The improvements are the result of decreasing the control-grid-to-cathode spacing, and at the same time decreasing the pitch and diameter of the control-grid wires.*

The maximum allowable resistance in the control-grid circuit is determined from the grid-plate transconductance of the tube, the cathode-bias resistor, and the screen voltage-dropping series resistor. Also, the use of a small unby-passed resistor in the cathode circuit to neutralize the changes in input capacitance and input loading with varying plate current is presented.

The 1851 and 1852 have the highest ratios of grid-plate transconductance to plate current of any commercially available tubes with the result that they have high signal-to-noise ratios. The high grid-plate transconductances also result in high-conversion transconductance when these tubes are used as mixer tubes; the 1851 and 1852 give a maximum of 3500 micromhos, and the 1853 a maximum of 1500 micromhos. With practical circuits the 1851 and 1852 have produced gains per stage of 3.5 to 7.0 at 50 megacycles, and of 20 to 45 at 11 megacycles. Similarly, the 1853 has produced gains per stage of 2 to 4 at 50 megacycles, and of 6.5 to 13 at 11 megacycles. All of these values are for a band-pass of 2.5 megacycles.

INTRODUCTION

PRESENT television receiving circuits, since they must be capable of passing a very wide frequency spectrum at high frequencies, have made it necessary to improve the associated tubes if the number of amplifying stages are to be kept down to a reasonable number. In a typical broadcast receiver, the output impedance of the intermediate-frequency stage is of the order of 100,000 to 200,000 ohms so that with a tube having a grid-plate transconductance of 1000 micromhos a gain per stage of 100 to 200 can be realized. In television receivers, the corresponding output impedances drop to 1000 to 2000 ohms so that with the same tubes the gain would be only 1 or 2. The situation is even worse when high-frequency stages are considered. Here the maximum impedance that can be built up is only a few hundred ohms. These limitations for both the

* Presented at I.R.E. Convention, Rochester, N. Y., Nov. 14-16, 1938.

high-frequency and intermediate-frequency impedances are the result of the shunting capacitances being unfavorably high. Therefore, it is of the utmost importance that the fixed shunting capacitances of the tube and wiring be minimized. Since this shunting capacitance consists of the output capacitance of the first stage plus the input capacitance of the succeeding stage, the relative merit (M) of an amplifier tube is conveniently designated as the ratio of its grid-plate transconductance (g_m) to the sum of its input and output capacitances. Thus,

$$M = \frac{g_m}{C_{\text{input}} + C_{\text{output}}}$$

H. A. Wheeler[1] has shown that if the tubes are connected by proper filter networks the impedances that can be built up are inversely proportional to the square root of the sum of the input and output capacitances. With these special filter networks, the figure of merit becomes,

$$M = \frac{g_m}{\sqrt{C_{\text{input}} + C_{\text{output}}}}$$

Any solution to the problem of improving the present receiving tubes must, therefore, provide a substantial increase in grid-plate transconductance without too much capacitance increase. This requirement immediately rules out the simple "brute-force" method of increasing the size of the cathode and maintaining the present spacings, since increasing the transconductance by a factor n will also increase the input capacitance by the same factor n. The output capacitance would probably also increase by a factor approximately equal to n due to the larger plate required to handle the proportional increase in plate dissipation. Electron-multiplier tubes, as developed by Dr. V. K. Zworykin, obviously fill these requirements nicely. It is theoretically possible to obtain an enormous transconductance and by proper design to provide lower values of input and output capacitances than those of conventional tubes. The chief obstacles to the use of the electron-multiplier tube are the elaborateness of the multistage mechanical structure and the need of providing the high voltage required to get high ratios of secondary emission to primary current. Other secondary-emission schemes for amplifying controlled-cathode currents have also been used successfully in the laboratory, but these are too critical of adjustment for general application.

Another method is to reduce the spacing between the cathode and control grid. As a result of using small spacing between these two

[1] H. A. Wheeler, "Wide-Band Amplifiers for Television", paper given at the I.R.E. Convention, June 18, 1938, New York City.

electrodes, B. Salzberg succeeded in obtaining a transconductance of 7000 micromhos in a tube similar to the 57 type. Use of this principle has also made possible the acorn types 954, 955, and 956. If the control grid is brought closer to the cathode, the permeance of the tube, and consequently its transconductance, goes up inversely as the square of the spacing. The grid-cathode component of input capacitance on the other hand increases inversely only as the spacing. If the spacing of control grid to cathode for a specified tube were halved, the tube transconductance would be increased four times for the same cathode current, the capacitance doubled, and the figure of merit would be slightly better than twice that of the original tube, provided the output capacitance were reasonably small in comparison to its input capacitance. This was the reasoning which prompted the development of the

Fig. 1—Photograph of the 1852 tube. The 1853 has the same appearance.

Fig. 2—Photograph of the 1852 mount. The 1853 mount has a similar appearance.

Fig. 3—Photograph of Type 1851 tube.

Type 1851 tube and which was carried over to the newer single-ended Types 1852 and 1853, illustrated in Figures 1 and 2.

Mechanical Features of Type 1851, 1852, and 1853 Tubes

The major problems encountered in developing the 1851 and 1852 were nearly all mechanical. Since the clearance between cathode and control grid was very close (0.005 inch between cathode surface and grid wire), only a flat structure was considered feasible. This restriction results from the present technique of winding grids where variations as great as plus or minus 0.001 inch can be expected and where, after heating the tube, on exhaust the grid wires may "reset" to give even greater variations. These variations occur with flat, round, or elliptical grids. However, of these grid forms, flat grids can most easily be stretched after winding on a flat mandrel so that the size variation even after heating is less than plus or minus 0.0005 inch. Round grids must be concentric with a round cathode. Consequently the side rods must be placed on the outside of the grid which is impractical. Elliptical grids requiring elliptical cathodes are much more difficult to maintain to size than the flat structures and offer very little improvements in performance over the flat structures.

For optimum control (that is, maximum grid-plate transconductance), the electric field before the cathode should be uniform even though there is a nonuniform conducting plane consisting of the individual grid wires in this field. In conventional tubes this condition is approached if the distance between the centers of the individual grid wires is 0.6 or less of the spacing from cathode surface to the plane of the grid-wire centers. However, if this ratio becomes much greater than unity, the ratio of transconductance to plate current falls off rapidly and the cut-off characteristic of the control grid becomes poor. Thus, a clearance of 0.005 inch should theoretically require a grid having over 300 turns per inch. It is impractical to wind such a grid, but a test was made using 250 turns per inch with a grid wire of 0.0012 inch in diameter (a human hair is approximately 0.003 inch) to get some idea of the magnitude of transconductance obtainable with fine grid pitch. With this grid, a transconductance of 11,000 to 12,000 micromhos at 10 milliamperes plate current was obtained. The grid used in the 1851 and 1852 is made with 0.002-inch wire wound at 142 turns per inch. The resulting transconductance is 9000 micromhos at 10 milliamperes plate current.

The positioning of the No. 2 grid (screen) should be as far away from the No. 1 grid (control grid) as feasible in order to minimize that part of the input capacitance due to the screen. Optimum positioning would result in the necessity for excessively high-screen voltage, since for a given plate current, the screen voltage will be approximately directly proportional to the spacing between control grid and screen. For example, if the distance between these two grids were halved, then halving the screen voltage would give the same plate current and only the input capacitance would be increased. The final positioning used in the 1852 and 1851 was a compromise based on the desirability of obtaining a semi-remote cut-off characteristic. The spacing between control grid and screen is such that with 150 volts on the screen (and normal control-grid bias) the tube operates at maximum transconductance and plate current. If the screen voltage is supplied from a 300-volt supply through a dropping resistor of proper value, the cut-off voltage point of the control grid will be just double that for a fixed 150-volt screen potential. This compromise resulted in an increase in the input capacitance from 9 to 11 micromicrofarads as compared with a screen positioned for 250-volt operation.

Since it is so desirable to minimize the 3 micro-microfarads of capacitance contributed by the conventional top seal such as found in the Type 6K7, the construction of the top lead-out was altered in the 1851 (see Figures 3 and 4). This change resulted in a saving of approximately 2 micro-microfarads, and was accomplished by providing a large hole in the metal insert at the top of the envelope and using a flared glass to lengthen the distance through the glass from the lead-out to the metal shell. This expensive and unconventional construction has been now superseded by the simpler construction of the single-ended 1852. The 1852 is identical with the 1851 electrically and mechanically except that in the 1852 all the leads are brought out through the stem.

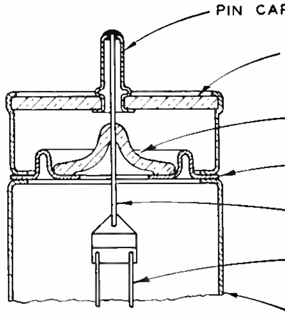

Fig. 4—Cross section at top of the 1851 tube.

The output capacitance of the 1851 design was reduced from 10 micro-microfarads in developmental types to 5 micro-microfarads by replacing the usual cylindrical plate with two narrow flat plates (see Figure 2). The side rods of the control grid, being large in diameter, have a focusing effect which even at normal bias confines the plate current to a strip about one-half the width of these plates.

Grid emission can be quite serious in producing noise in some applications of high-transconductance tubes. Furthermore, grid emission, if large resistances are placed in the grid circuit, may cause the tube to "run away" due to loss of negative bias. The usual radiators and materials used to control grid emission were all insufficient to keep the grid cool enough or to prevent building up of grid-emitting surfaces during the life of the tube. It was necessary to resort to cooling the grid by conducting the heat away from the side rods to the outside of the envelope. The usual heat-conducting side rods of copper proved to be too soft to maintain the close tolerances necessary, and so an alloy of copper and silver is used. In the case of the 1851 the heat is conducted from this grid by a short piece of nickel strap fastened to the tops of both grid legs and thence through a

0.030-inch Dumet lead to the glass seal and top cap (see Figure 4). In the case of the 1852 and 1853, the heat is similarly carried out to the glass in the bottom seal and the socket pin (see Figure 5). In all cases it is necessary to darken the inside of the metal envelope so that the heat radiated from the cathode, No. 2 grid, and plates will not be reflected and heat parts including the control grid excessively. Darkening is done on the exhausting machines by admitting air to the system while the envelope is receiving its first heating, and oxidizing the inside of the envelope.

The 1853 is a remote cut-off r-f pentode similar to the 1852 in construction. Both types are single-ended, that is, all the leads including control grid and plate are brought out at the base. The method of shielding these two leads within the tubes is the same as used in the new series of receiver tubes (6SK7, 6SJ7, 6SQ7, etc.), and has been described recently by Kelley and Miller.[2]

Electrical Characteristics of the 1851, 1852, and 1853 Tubes

The static characteristics of the 1851, 1852, and 1853 have the generalized forms of any typical pentode. Typical operating voltages, currents, and capacitances are given in the table below:

Type	1851 1852		1853	
Heater voltage	6.3	volts	6.3	volts
Plate voltage	300	max. volts	300	max. volts
Suppressor voltage	0	volts	0	volts
Screen voltage	150	max. volts	200	max. volts
Grid voltage	—		—3	min. volts
Cathode-bias resistor	160	ohms	—	ohms
Heater current	0.45	ampere	0.45	ampere
Plate current	10.0	milliamperes	12.5	milliamperes
Screen current	2.5	milliamperes	3.2	milliamperes
Transconductance	9000	micromhos	5000	micromhos
Plate resistance	0.75	approx. megohm	0.7	approx. megohm
Grid bias for cut-off	—6	volts	—15	volts
Triode amp. factor	40		32	
Input capacitance { cold	11.0	$\mu\mu$f	8.0	$\mu\mu$f
Input capacitance { hot	13.5*	$\mu\mu$f	9.3**	$\mu\mu$f
Output capacitance (cold)	5.0	$\mu\mu$f	5.0	$\mu\mu$f
Grid-plate capacitance	0.015	max. $\mu\mu$f	0.015	max. $\mu\mu$f
Input loading at 40 Mc.	4300*	ohms	8500**	ohms

* With plate current of 10 ma.
** With plate current of 12.5 ma.

[2] R. L. Kelley and J. F. Miller, "Single-Ended R-F Pentodes", *Electronics,* September, 1938.

It will be noted from the above table that the Types 1851 and 1852 are shown for cathode-bias operation. The 160-ohm cathode-bias resistor and nominal cathode current of 12.5 milliamperes flowing through it, develop a bias of —2.0 volts. This value of fixed bias cannot be recommended since variation in tube contact potential may change the grid bias by as much as ±0.5 volt during tube life. Also, mechanical variations between tubes account for about ±0.25 volt in equivalent bias variation. These two effects preclude the use of fixed bias if tubes are to be interchangeable. The use of the 160-ohm cathode-bias resistor cuts the effect of these variations to about one-third of that obtained if fixed bias were used. The effect can be made smaller by using cathode resistors of still higher value. The 1853 has roughly one-half the transconductance and about twice the clearance between

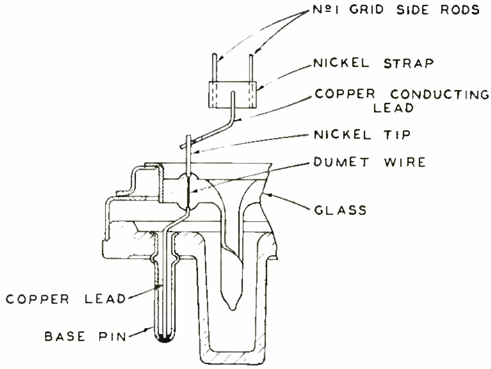

Fig. 5—Cross section through base of the 1852 and 1853 tubes. This figure shows the method of conducting heat from No. 1 grid to glass seal in stem and to pin in base.

cathode and No. 1 grid in comparison with the 1852, and is not so critical to mechanical and electrical variations. The 1853, therefore, does not require the use of the cathode resistor for obtaining bias.

A tube critical in its bias adjustment must have limitations as to the maximum allowable resistance which can be used in the input-grid circuit, because small gas, leakage, or grid-emission currents flowing through this resistance will alter the effective grid bias. The following relationships between the various resistances and the d-c currents flowing in typical grid-controlled vacuum-tube triode, tetrode, and pentode circuits hold so long as the No. 1 grid current (I_{c1}) is negligible in comparison with the total cathode current (I_k).

For pentodes and tetrodes,

$$R_{c1} = \frac{\triangle I_k}{\triangle I_{c1}} \left[\frac{1}{g_k} + R_k \left(1 + \frac{1}{u_{1-2}} \right) + P \frac{R_{c2}}{u_{1-2}} \right]$$

For triodes,

$$R_{c1} = \frac{\triangle I_b}{\triangle I_{c1}} \left[\frac{1}{g_m} + R_k \left(1 + \frac{1}{u_{1-p}} \right) + \frac{R_p}{u_{1-p}} \right]$$

where,

$\triangle I_k$ = Change in cathode current
$\triangle I_b$ = Change in plate current
$\triangle I_{c1}$ = Change in No. 1 grid current
R_{c1} = Grid resistor in ohms
R_{c2} = Series-screen resistor in ohms
R_k = Cathode resistor in ohms
R_p = Series-plate resistor (load) in ohms
u_{1-2} = Triode-amplification factor of pentode or tetrode
u_{1-p} = Amplification factor of triode
P = Ratio of screen current (I_{c2}) to cathode current

$$= \frac{I_{c2}}{I_{c2} + I_b}$$

g_m = Grid-plate transconductance in mhos
g_k = Grid-cathode transconductance in mhos

$$= g_m \frac{I_k}{I_b}$$

These equations can be put in more useful form if we specify that for a grid current of one microampere a change in cathode (or plate) current of one milliampere can be tolerated. This is a reasonable value to assume since good tubes should never have a gas current of more than one or two microamperes. Thus, the ratio $\triangle I_k$ to $\triangle I_{c1}$ will be 1000. Substituting in the pentode formula this value and the normal values of g_k, P, and u_{1-2}, we have:

for 1851 and 1852,

$$R_{c1} \text{ (max.)} = 10^3 \, [88 + 1.025 \, R_k + 0.005 \, R_{c2}]$$

and for the 1853,

$$R_{c1} \text{ (max.)} = 10^3 \, [160 + 1.03 \, R_k + 0.006 \, R_{c2}]$$

It is interesting to note that the 1851 and 1852 with screen voltage obtained from a fixed supply and the recommended 160-ohm cathode-bias resistor can tolerate a maximum R_{c1} of 250,000 ohms; and with the screen voltage obtained through a 60,000-ohm dropping resistor from a 300-volt supply, a maximum of about 550,000 ohms. The 1853 with fixed screen voltage and fixed bias voltage gives a maximum R_{c1} of only 160,000 ohms. If a cathode-bias resistor of 190 ohms is used, the maximum R_{c1} is 350,000 ohms; and if a series-screen dropping resistor of 30,000 ohms for a 300-volt supply source is used with the cathode-bias resistor, a maximum R_{c1} of 530,000 ohms may be used.

However, if larger cathode-bias resistors are used and are tied back to a positive-voltage point on the supply bleeder, the regulating action is so improved that input-series resistances having maximum values of several megohms may be used. It is obvious, since these considerations are for d-c effects only, that the cathode-bias resistor and the series-screen resistor may be by-passed without impairing the a-c amplifying characteristics of the amplifier stage.

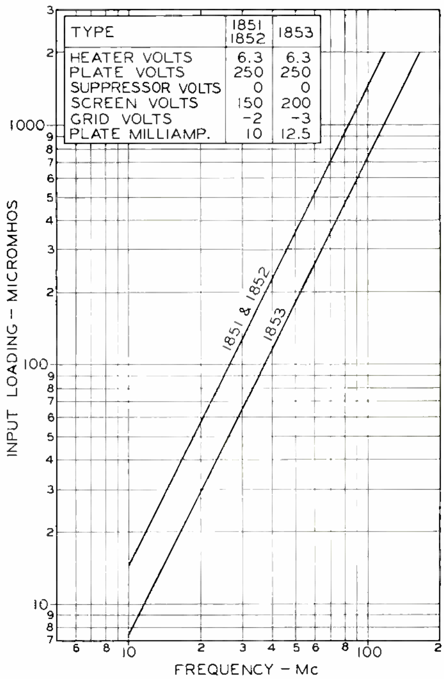

Fig. 6—Total input-loading vs. frequency at constant plate-current for Types 1851, 1852, and 1853.

The input impedances of these three tubes is of such a nature that they can seriously hamper the performance of a television receiver. The input impedance consists essentially of two parts—an ohmic component usually designated as "loading", and the reactive component, input capacitance. The ohmic component varies with the square of the frequency whereas the reactive component varies inversely with frequency. These components also vary with cathode current at constant frequency. Figure 6 shows the loading vs. frequency characteristic (at normal-plate current) for the 1851, 1852, and 1853, and Figures 7 and 8, under the curves marked $R_k = 0$, show the changes in input capacitance and input loading for change in plate current (at 40 megacycles). For intermediate frequencies of the order of 10 megacycles neither of the effects shown in Figures 7 and 8 is serious, but if a high-frequency stage tuned for a frequency of 50 megacycles is investigated, both effects are found to be very serious. This result is shown in Figure 9, where the response at the grid of the high-frequency amplifier is plotted against frequency for various control-grid biases. When the bias is low, and the plate current

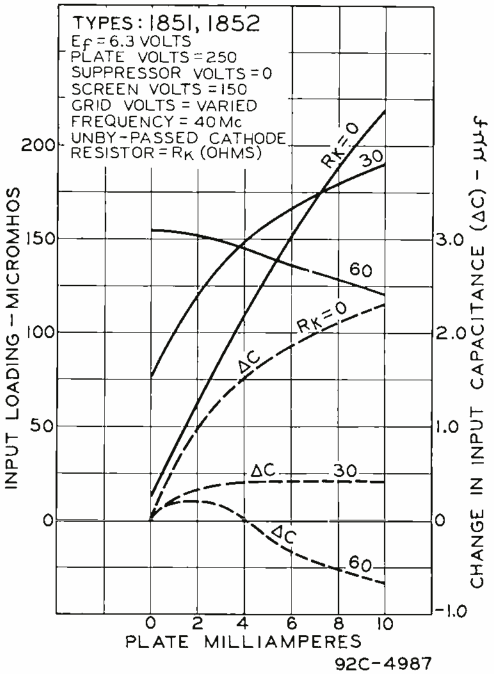

Fig. 7—Input-capacitance change and input-loading vs. plate-current at 40 megacycles for the 1851 and 1852.

therefore is high, the loading is high. High loading has the same effect as adding shunt resistance across the tuned circuit; it widens the band-pass and lowers the maximum response. As the bias is made more negative, the loading decreases, the response improves, the band-pass narrows and, due to the change in input capacitance of the tube, the peak of the band shifts to a higher frequency. These effects are shown in the curves of Figure 9. The —3-volt bias curve has been shown with increased input (approximately the same peak response as the —14-volt curve) so that the broadening of the band-pass can be more readily observed. In Figure 10 are shown the results of applying the proper corrective measures. It will be observed that there is no frequency shift and practically no change in band-pass width. The results were obtained by a method originally suggested by

Strutt and Van der Ziel[3] in which a small unby-passed resistor is placed in the cathode circuit. They show that the input admittance of the tube (from ground to control grid) is approximately,

$$Y_{\text{input}} = \frac{Y_g}{1 + g_k R_k} = \frac{g_g + j\omega C_{\text{input}}}{1 + g_k R_k}$$

where Y_g is the admittance of the paralleled input-loading admittance (g_g) and input capacitance between cathode and control grid only

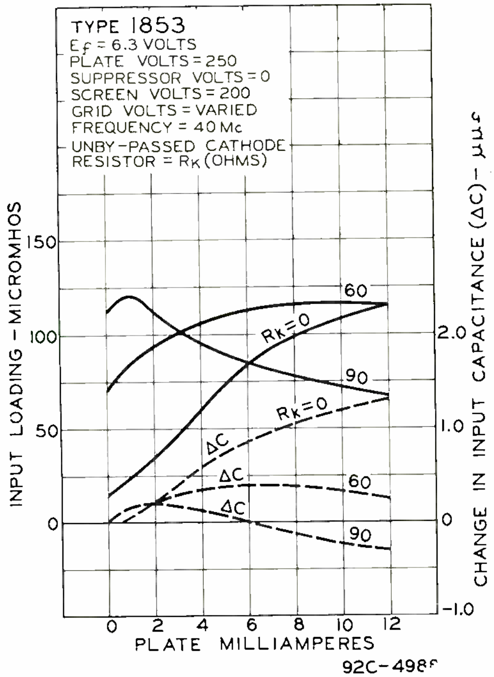

Fig. 8—Input-capacitance change and input-loading vs. plate-current at 40 megacycles for the 1853.

(C_{input}), R_k the unby-passed cathode resistance, and g_k is the grid-cathode transconductance which is simply the grid-plate transconductance g_m corrected for the current absorbed by the screen grid. The grid-cathode transconductance is,

[3] M. J. O. Strutt and A. Van der Ziel, "Einfache Schaltmassnahmen zur Verbesserung der Eigenschaften von Hochfrequenzempfangsröhren im Kurzwellengebiet," *Elek. Nach. Tech.*, Vol. 13, pp. 260-268; August (1936).

M. J. O. Strutt and A. Van der Ziel, "Die Ursachen für die Zunahme der Admittanzen moderner Hochfrequenz — Verstärkerröhen im Kurzwellengebiet," *Elek. Nach. Tech.*, Vol. 14, pp. 281-293; Sept. (1937).

"Simple Circuit Arrangements for Improving the Operation of H-F Amplifying Valves in the Short-Wave Band," *Philips Setmakers' Bulletin*, No. 42, pp. 53-61; May-June, 1937.

M. J. O. Strutt and A. Van der Ziel, "The Causes for the Increase of the Admittances of Modern High-Frequency Amplifier Tubes on Short Waves," *Proc. I.R.E.*, Vol. 26, pp. 1011-1032; Aug. (1938).

$$g_k = g_m \frac{(i_p + i_{c2})}{i_p}$$

Separating the real and imaginary components of the input admittance, Y_{input}, and analyzing the imaginary component for capacitance changes, we may immediately write:

$$C_{input} = \frac{C + \Delta C}{1 + g_k R_k}$$

where C is the input capacitance (grid to cathode only) at cut-off, and ΔC is the increase in this capacitance due to electron flow. Therefore,

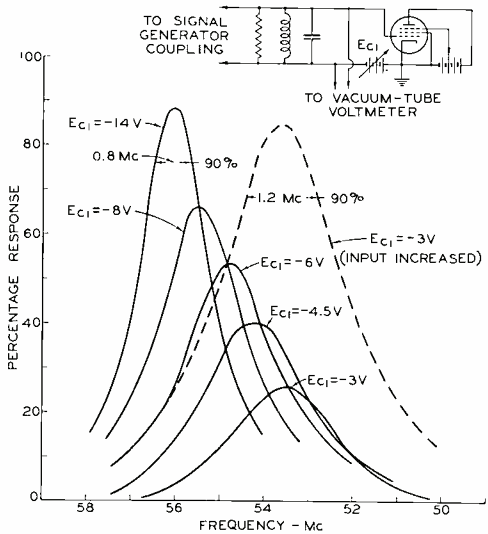

Fig. 9—Effect of input-capacitance change and input-loading on response vs. frequency for various control-grid biases.

for example, to have the same input capacitance at cut-off as at maximum current it is only necessary to make,

$$C_{input} = \frac{C + \Delta C}{1 + g_k R_k}$$

Choosing a value of g_k corresponding to maximum plate current and solving for R_k, we obtain

$$R_k = \frac{\Delta C}{C} \frac{1}{g_k}$$

In a similar manner we investigate the real component for input-loading effects. The effective-input conductance is

$$g_{\text{input}} = \frac{g_g}{1 + g_k R_k}$$

where g_g is the electron-transit time and, as will be discussed later, is also the equivalent effect introduced from the cathode-lead inductance. This g_g varies with the square of frequency (see Figure 6) and, therefore, we may state

$$g_g = K\omega^2$$

where K is a constant and $\omega = 2\pi f$, where f is the frequency in cycles per second. When plate current does not flow as a result of the tube

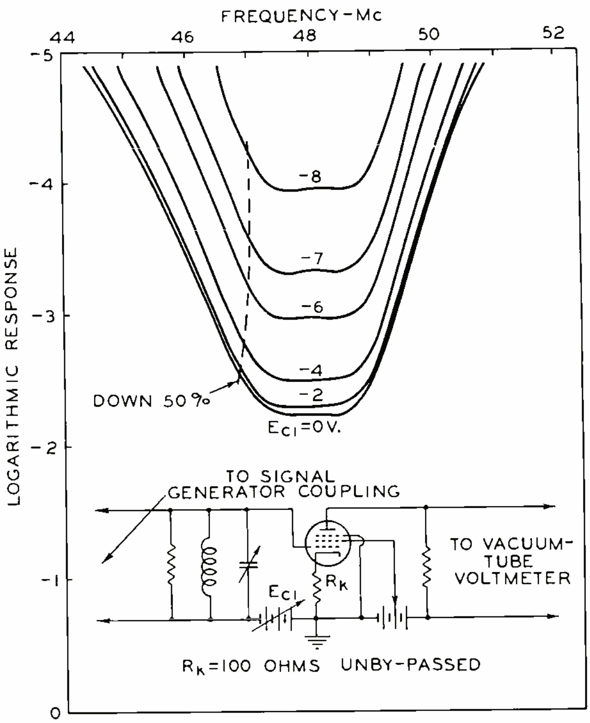

Fig. 10—Effect of corrective measures on response vs. frequency for various control-grid biases.

working at cut-off bias, and if we assume that R_k is small compared to $1/g_g$, the equivalent parallel input conductance, g_{input}, of the input capacitance C and cathode resistor R_k in series is approximately

$$g_{\text{input}} = \omega^2 C^2 R_k \text{ (with no plate current)}$$

Equating the no-plate-current condition to the plate-current-flow condition gives,

$$\frac{K\omega^2}{1 + g_k R_k} = \omega^2 C^2 R_k$$

Solving for R_k, we obtain

$$R_k = \frac{1}{2g_k}\left[-1 + \sqrt{\frac{1 + 4Kg_k}{C^2}}\right]$$

In general, loading variations and capacitance variations cannot both be compensated for at the same time, unless

$$K = \frac{C(C + \triangle C)}{g_k}$$

Fortunately, however, fulfilling one condition substantially improves the other because both conditions are independent of frequency. Check

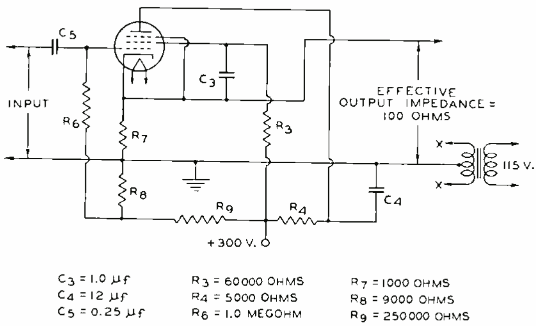

Fig. 11—Impedance transformer for use between high-impedance circuit and low-impedance circuit.

tests showing the compensating effects for the 1851, 1852, and 1853 are shown graphically in Figures 7 and 8. The computed value of R_k for the 1851 and 1852 for no capacitance change was 44 ohms, and for no-loading change was 77.5 ohms. A value providing a good compromise is probably near 40 ohms. For the 1853 a good value is approximately 70 to 80 ohms.

It should be noted that the unby-passed cathode resistor, because it produces degenerative amplification, reduces the gain to $1/(1 + g_k R_k)$ of its value when $R_k = 0$. For the 1851 and 1852 this value would be about 70 per cent and for the 1853 would be about 80 per cent. It is consequently desirable to work with as low a value of R_k as possible. Also, an additional precaution in the circuit design is necessary, i.e., the screen and suppressor grids should be grounded and not tied to the cathode, or else feedback may result. The plate-to-cathode capacitance, therefore, in the tube and wiring layout becomes important since the cathode is not at ground potential. The 1851, 1852, and 1853 have plate-cathode capacitances of about 0.0055 micro-microfarads.

Strutt and Van der Ziel[4] have shown that the common-lead inductance of the input and output circuits, i.e., the inductance of the cathode lead, contributes loading to the input circuit, which since it also varies with the square of frequency cannot be distinguished from the electron transit-time loading. It is quite probable that the greater part of the input loading of these tubes is caused by the cathode-lead inductance. Using the results of the work by North and Ferris[5], we find that the 1851 and 1852 should have electron-transit-time loadings of the order of magnitude of 3 to 12 micromhos at 40 megacycles

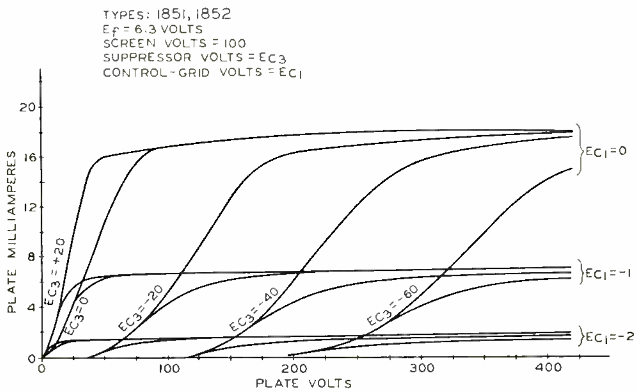

Fig. 12—Suppressor characteristics of the 1851 and 1852.

whereas the observed value is 230 micromhos. Furthermore, a test in which a short piece of wire (0.025-inch diameter and 20-mm long) having a computed inductance of only 0.02 microhenries was made part of the common cathode lead external to the tube showed that the input loading was increased to 330 micromhos. The common cathode-lead inductance of the tube alone is of this same magnitude. In another test with the 1853, two cathode leads were brought out to separate base pins and one of these was used for the grid-return

[4] loc. cit.

[5] W. R. Ferris, "Input Resistance of Vacuum Tubes as Ultra-High-Frequency Amplifiers," *Proc. I.R.E.*, Vol. 24, pp. 82-105; Jan. (1938).

D. O. North, "Analysis of the Effects of Space Change on Grid Impedance," *Proc. I.R.E.*, Vol. 24, pp. 108-136; Jan. (1938).

circuit and the other for the plate and screen circuits. With this connection, the value of loading dropped from 115 to 85 micromhos. This result indicates the necessity of making the cathode-return connections as close to the cathode pin as possible. It is also apparent that if further improvement in tubes working as amplifiers at high frequency is to be realized the inductance of the common-cathode lead must be minimized.

Circuit Applications

Due to their high transconductance and comparatively low-plate current, the 1851 and 1852 have given the highest signal-to-noise ratio of any commercially available tube types. This feature makes the 1851 or 1852 useful in the stage following an Iconoscope in a television transmitting system. In this stage, noise or "snow" has always been a limiting factor.

In the following tables, a comparison is made of the equivalent noise resistance of several tube types. This equivalent noise resistance is defined as that resistance which, if placed in series with the control grid of the tube, will produce noise equal to that introduced by the tube itself. The voltages chosen were those giving optimum results.

Noise Equivalents

Type	6C6	954	6L6-G	6Y6-G	6V6-G	1851	
Plate voltage	300	300	300	300	300	250	Volts
Screen voltage	50	50	50	25	50	90	Volts
Grid voltage	—1	—1	—1	—1	—1	—1	Volts
Plate Current	2	2	20	22	13	10	Ma.
Screen Current	0.5	0.4	2	2	1.5	3	Ma.
Equivalent noise†	6000	5500	1050	600	1800	520*	Ohms

Type	6C6°	955	6L6-G°	6Y6-G°	6V6-G°	1851°	
Plate voltage	50	50	50	40	50	90	Volts
Grid voltage	—1	—1	—1	—1	—1	—1	Volts
Plate Current	2	2	15	22	10	10	Ma.
Equivalent noise†	1600	1700	600	550	800	250*	Ohms

† Equivalent resistance introduced into control-grid circuit
* Approximate
° Connected as triodes with G_1 as control grid.

In the above tables, it will be noted that when tetrodes or pentodes are connected as triodes, there is a considerable improvement because noise introduced by the screen current is eliminated.

Another useful application of the 1851 and 1852 is that either can be employed to couple the output of the Iconoscope to the input of the first video amplifier. Since the Iconoscope is a low-current device, it should be worked preferably into loads of the order of 30,000 to 300,000 ohms. This load cannot be used for the usual first video-amplifier tube because its input capacitance, being of the order of 10 micro-microfarads, effectively by-passes this input load for high

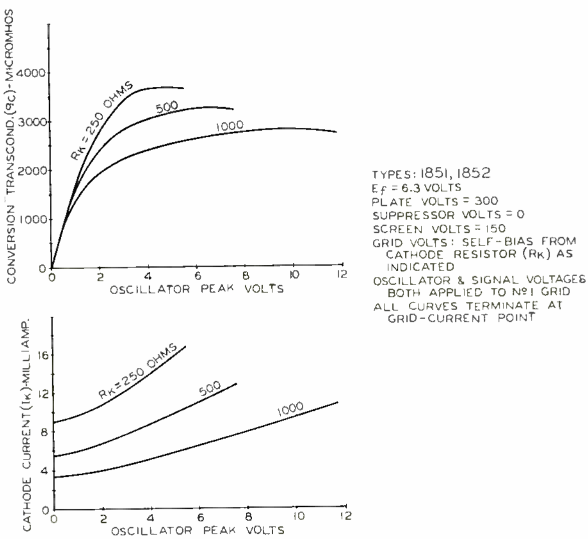

Fig. 13—Converter characteristics of the 1851 and 1852 for various values of by-passed cathode resistor.

frequencies. However, by using the cathode-coupling circuit shown in Figure 11, the output of the Iconoscope can be fed into an 1851 or 1852 connected as a triode or pentode to feed the first video stage, with the benefit that the effect of the shunt-input capacitance of the coupling tube is made negligible and the signal-to-noise ratio is increased. The cathode-coupling stage has a gain slightly less than unity.

For experimental applications or for applications where uniformity of tube characteristics during life is not important, the transconductance of the 1851, 1852, and 1853 can be increased about 25 per cent by connecting the No. 3 grid as an accelerator instead of as a suppressor. The No. 3 grid should have a voltage about 50 volts more positive than the normal No. 2 grid voltage. Under these conditions, the secondary emission from the screen grid adds to the normal-plate current. This secondary emission is large enough to cause negative-screen current. The sum of the currents of the No. 2 and No. 3 grids is close to zero. The plate current of the 1851 and 1852 will, therefore, be 12.5 milliamperes instead of 10.0 milliamperes and the transconductance will be about 11,250 micromhos instead of 9000 micromhos. Likewise for the 1853, the plate current will be raised to about 15.5 milliamperes and the transconductance will be raised from 5000 to 6200 micromhos.

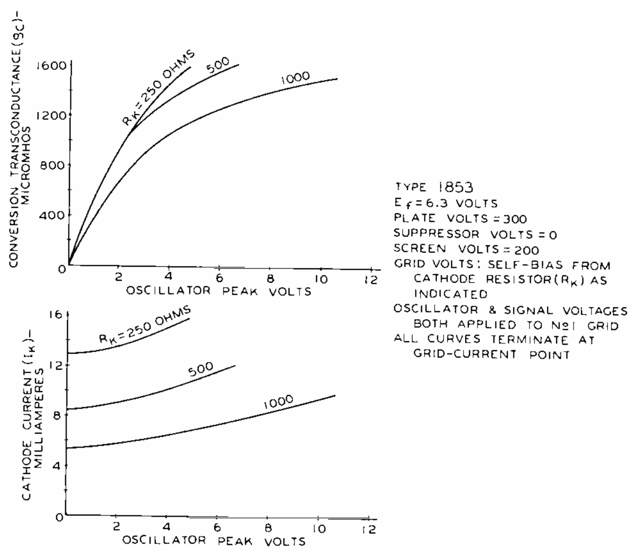

Fig. 14—Converter characteristics of the 1853 for various values of by-passed cathode resistor.

In Figure 12 the suppressor characteristics are shown. A convenient use for these curves is in connection with the operation of the video-amplifying stage which follows the Iconoscope. This stage is used for the introduction of the blanking signal. With a blanking signal of the order of 50 volts peak or more supplied to the suppressor neither the input nor output-video circuits has to be altered.

When the 1851, 1852, or 1853 is used as a mixer or converter tube, the signal and local-oscillator voltages are both coupled to the control grid. Coupling to the cathode is not recommended since, as was discussed earlier in this paper, any inductance in the cathode lead produces serious increases in input loading. Figures 13 and 14 show the converter characteristics of the 1851, 1852 and 1853, respectively, for various values of cathode resistance (by-passed by a sufficiently large condenser). The measurements were made by the R.M.A. method in which sixty-cycle signal and oscillator voltages are used. These curves indicate that if large variations in oscillator voltage are encountered, fairly uniform-conversion transconductance (g_c) may be obtained by using a 1000-ohm cathode-bias resistor. If small variations are encountered, a higher g_c may be obtained by the use of smaller values of cathode-bias resistors. The effect of adding a by-passed dropping resistor in the screen circuit does not greatly change the general shape of the curves. However, a 100,000-ohm screen resistor, for example, has the advantage of approximately halving the changes in cathode current and may be desirable for minimizing the tube input-loading and capacitance changes. Its undesirable feature is that the maximum oscillator-peak voltage which may be used is made approximately 2 volts less for the 1851-1852 types and approximately 1 volt less for the 1853.

It may be in order to indicate a few simple expedients to prevent the inevitable parasitic oscillations which occur when one of these tubes is placed in a tube-test set not wired for reading high-transconductance tubes. Usually resistors (quarter-watt size) of 100 to 500 ohms placed close to and in series with the control-grid terminal will stop this undesirable effect. If the tube and circuit still persist in oscillating, about 100 ohms placed in series with and close to the screen terminal together with a 0.001 microfarad condenser which shunts this grid to ground will stop the most obstinate cases. In some practical television circuits, tying the suppressor grid to a positive voltage of +10 to +40 volts has also worked successfully.

Although these tubes may be used in various applications as discussed above, their prime use is as tuned high-frequency amplifiers and intermediate-frequency amplifiers. With practical circuits for the 1851 or 1852, the gain per stage is 3.5 to 7 at a frequency of 50 megacycles and a band-width of 2.5 megacycles. At an intermediate frequency of 11 megacycles and a band-width of 2.5 megacycles, the gain per stage is 20 to 45. Similarly, for the semi-remote cut-off 1853, the gain per stage is 2 to 4 at a frequency of 50 megacycles and a band-width of 2.5 megacycles. At an intermediate frequency of 11 megacycles and the same band-width, the gain per stage is 6.5 to 13. These values apply to circuits of somewhat lower band-pass than is desirable for the present 441-line picture.

ANALYSIS AND DESIGN OF VIDEO AMPLIFIERS

By

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Part II

Introduction

THIS paper is an extension of a previous one[1] which contained an elementary discussion of the factors influencing the performance of wide-band video amplifiers. The principal items of interest were an analysis of the effects of variable time delay and amplitude response on the reproduced picture, and a discussion of one method of high-frequency compensation. Notes on certain measurement techniques, and a discussion of low-frequency requirements were also included.

It is sufficient, in the present case, to repeat, as explained in the previous paper[1], that the ideal video amplifier should have flat frequency response and constant time delay over the band of frequencies required for adequate reproduction of the transmitted picture.

The importance of maintaining the characteristics of individual video stages as close to the ideal values as possible is accentuated in cases where numerous stages are connected in cascade. This is true because the overall gain is equal to the product of the individual stage gains, while the net time delay is equal to the sum of the time delays of the individual stages. It is interesting to note that this applies particularly to video-amplifier chains, where thirty or more stages may be used in a television transmitter.

It may be said, in general, that it is quite difficult to maintain both time delay and gain constant over a wide band. Generally a compromise is made, with neither the gain or delay exactly constant, but with both satisfactorily close to optimum values.

Furthermore, the correction expedients which are applicable at one end of the video band have no effect at the other end. Thus, the use of peaking coils for maintenance of constant gain at high frequencies does nothing whatsoever to the low-frequency performance.

Because of this segregation of the video band into two distinct regions, it is desirable to treat the high- and low-frequency characteristics as separate problems. This procedure will be followed in this article, with the first section containing a discussion of the high-frequency aspects of the problem, and the last section dealing with the low-frequency performance of the amplifier.

SECTION I

HIGH-FREQUENCY CONSIDERATIONS

All of the video-amplifier circuits which have appeared to date consist essentially of resistance-coupled stages, each provided with some form of high-frequency gain and phase correction. The decrease in amplification and introduction of phase distortion at the higher frequencies in an uncompensated amplifier is a direct result of the existence of unavoidable shunt capacitances, which are found in any circuit containing vacuum tubes and associated components (resistors, wiring, sockets, etc.). The reactance of these shunt capacitances appears as part of the plate-circuit load, and causes its impedance to decrease as the frequency is increased, resulting in a decrease in gain at the higher frequencies. The loss in gain is accompanied by a phase delay, which, with no shunt reactance in the plate load, would normally be zero. The manner of variation with frequency of this phase delay is important in determining the constancy of time delay over the video band.

There are several ways to reduce the effect of the load-circuit capacitance. One method involves the use of a very small load resistor, whose resistance is so low compared to the reactance of the shunt capacitance at the highest video frequency that the reactance has no effect on the gain or phase characteristics. This arrangement would possess no practical advantages, because of the great loss in gain per stage entailed by the use of a small plate resistor. (The gain in a video-amplifier stage may be taken generally as the product of the tube's mutual conductance and the plate-load impedance, since pentodes are used almost universally.)

A more practical way to obtain adequate high-frequency performance is to employ a circuit containing inductance to offset the loss in gain due to shunt capacitance. In this way essentially constant gain may be obtained, without resorting to abnormal reductions in the value of plate-load resistor.

The expedients employed to extend the frequency band in which constant gain obtains are described variously as correction circuits or peaking circuits. These may take any of several forms, depending upon whether the wide-band characteristics are obtained by inserting a peaking coil in the load circuit, to maintain the load impedance at a constant value, or whether the desired effect is obtained by the use of a coupling circuit, such as a low-pass filter, between successive stages of the amplifier.

Four types of high-frequency video load circuits will be discussed here: (1) Uncompensated load circuit. (2) Compensated circuit containing a peaking coil in series with the load resistor—known as shunt peaking. (3) Compensated load circuit in which a π-type low-pass filter is employed as the coupling element—known as series peaking. (4) Combination of shunt and series peaking.

The analysis of these various types of load circuits and the evaluation of their relative merits is somewhat simplified and made more readily adaptable to direct comparison by the use of the following list of symbols and definitions.

T_1 and T_2 = Two successive tubes of a video amplifier circuit.

R_L = Load resistor in plate circuit of T_1.

C_T = Total capacitance shunting the load circuit. This includes tube and wiring capacitances.

C_1 = Total output capacitance of T_1.

C_2 = Total input capacitance of T_2.

$C_2/C_1 = m$.

L_1 = Inductance of peaking coil in series with plate-load resistor of T_1 (shunt peaking).

L_2 = Inductance of peaking coil connected between plate of T_1 and grid of T_2 (series peaking).

f_o = Top frequency in the video band.

f = Any frequency in video band above 1 kc.

Φ = Phase delay in radians (caused by reactance in plate-load circuit).

$T = \dfrac{\Phi}{\omega}$ = Time delay in seconds (due to reactance in plate-load circuit).

$\triangle_T$ = Departure from constant time delay (seconds).

It should be noted at this point that, in general, maintenance of a flat frequency-response characteristic (at high frequencies) in a video amplifier stage usually will result in sufficiently uniform high-frequency time delay so that correction expedients which might be applied to produce an entirely uniform delay (and which might alter the response somewhat) are not usually necessary or desirable. This, of course, depends largely on the total number of stages in cascade to be employed for a given purpose, for, as pointed out previously, the overall gain characteristic is the product of the individual stage gains, whereas the total time delay is the sum of delay characteristics of each stage.

In this connection it is important to observe that the high-frequency performance of the amplifier determines the quality of the picture along any horizontal line, i.e., the horizontal detail and resolution. If both gain and delay characteristics are flat the picture is reproduced exactly. If the gain is constant in the video band and the time delay varies with frequency, all the high-frequency components are reproduced precisely in their proper relative amplitudes, but the location of the various picture elements is not correct, because of the different amounts of time taken for passage of the different frequencies. This results in inferior reproduction of horizontal detail.

It is difficult to determine precisely the maximum permissible variation in time delay in a complete television system. Some authors[2] suggest limiting the total variation in time delay to 0.1 μsec. Data calculated for a typical case shows that, for a 441-line picture (10-inch horizontal dimension on a 12-inch tube) a variation in time delay of 0.1 μsec. up to a top frequency of 2.5 Mc would cause the 2.5 Mc component to be displaced laterally by 0.015-inch with respect to the low-frequency components. This would amount to about one picture element (at 2.5 Mc) in the horizontal direction. It should be obvious that, for a given total tolerable time-delay variation, the permissible departure from constant delay-time per stage decreases as the number of stages increases, since the total delay variation is the sum of the individual-stage delay variations. Therefore, in receivers, where at the most, only three or four video stages would be employed, the permissible variation in time delay per stage is greater than in cases where a large number of stages are used in an amplifier chain. The line amplifiers in television transmitters fall in the latter category. Furthermore, in receivers, the delay variations in the i-f circuits are generally much greater than the delay variations in the video amplifiers, consequently attention is generally directed toward minimizing departure from constant delay in the i-f circuits. The procedure in designing video amplifiers for receivers consists principally, therefore, in obtaining a flat gain characteristic over the video band, while the time delay is permitted to depart from a constant value, within reasonable limits.

The magnitude of the time-delay variations, Δ_T (departure from the desired constant value) in an amplifier stage may be written in a number of ways. One method of expressing Δ_T, which will be used in this article, evaluates the departure from constant time of transmission as a fractional part of a period at the top video frequency, i.e., $\Delta_T = K/f_o = KT_o$.

Circuit 1—Resistance-Coupled Video Amplifier Utilizing No High-Frequency Peaking Expedients

The plate load Z_L comprises the load resistor R_L in parallel with the total shunt capacitance, C_T. The gain, which is equal to $Gm\ Z_L$, falls off as the frequency is increased according to

$$\text{gain} = Gm\ Z_L = \frac{Gm\ R_L}{\sqrt{1 + (2\pi f C_T R_L)^2}}$$

If we let R_L equal the reactance of C_T at the top frequency, f_o, we have

$$R_L = \frac{1}{2\pi f_o C_T}$$

and

$$\text{gain} = \frac{Gm\ R_L}{\sqrt{1 + (f/f_o)^2}}$$

At this frequency where $R_L = \dfrac{1}{2\pi f_o C_T}$, the gain is 70.7 per cent of the gain at low frequencies ($f = 10$ kc, for instance). The departure from constant time delay at f_o is $0.034/f_o$, i.e., 3.4 per cent of the period T_o at the top frequency, f_o. With $f_o = 3$ Mc, Δ_T is 0.011 microseconds. This is the difference in time delay caused by the presence of shunt capacitance in the plate-load circuit.

It should be evident that the gain of this type of load circuit is not sufficiently constant to permit its use in a video amplifier, unless the load resistor is made small compared to the total shunt-load reactance at the top video frequency.

While this analysis is included primarily to demonstrate the behavior of an uncompensated circuit, and as a basis of comparison for other compensated circuits to follow, it can be put to use as a means for measuring the total load-circuit capacitance of a video stage. The method, described in detail in Part I of this article, makes use of the fact that the gain of an uncompensated stage falls to 70.7 per cent of its low (10 kc) frequency value at a frequency for which the reactance of the capacitance in question is equal to the plate-load resistance. The measurement of the point of 0.707 response may be determined by noting the frequency at which the input to the stage under test must be increased to $\sqrt{2}$ times its low-frequency value, to maintain constant stage output.

The indicating device may include the following tube in the chain, which should have a low (100-ohm) resistor connected in its plate circuit to provide a voltage drop which can be read on a vacuum-tube voltmeter. The bias of this second tube should be maintained at its operating value, to preclude any error due to input capacitance variation with bias. A variation of this connection applies the vacuum-tube voltmeter across the load resistor R_L (with the following tube in circuit) and measures the output across R_L. The capacitance contributed by the vacuum-tube voltmeter must be known and taken into account in this measurement.

The total output-circuit capacitance can also be measured by a substitution method, in which a "Q" meter may be employed. A coil is selected to resonate with 100 $\mu\mu f$ or so on the "Q" meter at some frequency between 500 and 2000 kc. The circuit is resonated and then the capacitance terminals of the "Q" meter are connected across the output circuit of the stage under test. The plate-load resistor is removed and the plate-supply voltage of T_1 is turned off. The second tube operates normally, with its bias fixed at the operating point. The amount by which the "Q"-meter calibrated-capacitance must be changed, in order to re-establish resonance in the "Q"-meter circuit, is equal to C_T, the total shunt capacitance in the video stage.

Note that the resonant voltage which appears across the "Q"-meter tank circuit must be limited in amplitude to prevent rectification in the second tube's grid circuit, which would result in a change in bias and in second-tube input capacitance.

Circuit 2—Video Stage Compensated by a Coil in Series With the Load Resistor (Shunt Peaking)

This type of video stage may be compensated (the plate load-circuit impedance made essentially constant over the required frequency band) by inserting a properly proportioned inductance in series with the load resistor. The peaking-coil inductance is determined by the values of R_L, C_T, and the top video frequency, f_o.

R_L is chosen to equal the reactance of C_T at the top frequency, f_o, (C_T is measured with L_1 not in circuit, by either of the methods previously described). Therefore,

$$R_L = \frac{1}{2\pi f_o C_T}$$

The value of L_1 is determined from $2\pi f_o L_1 = \frac{R_L}{2}$ at the top frequency, f_o. Hence,

$$L_1 = \frac{R_L}{4\pi f_o}$$

The resonant frequency of L_1 and C_T is seen to be $\sqrt{2}$ times the top video frequency, f_o.

The gain is essentially constant, up to the frequency f_o, and is equal to GmR_L.

The time delay, in terms of f and f_o is

$$T = \frac{1}{2\pi f} \tan^{-1} \left[\frac{1}{4} \left(\frac{f^3}{f_o^3} + 2\, f/f_o \right) \right]$$

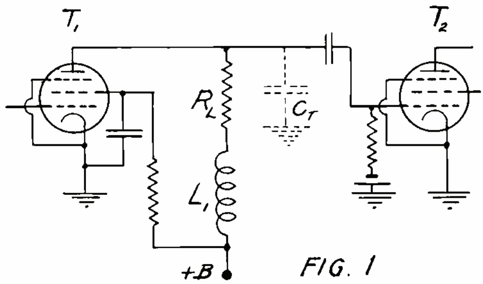


FIG. 1

The difference in time delay over the video band (from 1 k_c to f_o) is $0.0231/f_o$ seconds. With $f_o = 3$ Mc, this time-delay departure corresponds to 0.0077 microseconds. Note that the gain over the video band, GmR_L, is equal to the gain which would be experienced in an uncompensated stage having zero shunt-load capacitance, and the same value of plate-load resistor as is used here. This improvement in gain characteristic is achieved with no increase in the departure from constant time delay; in fact an improvement of 30 per cent in the approach to constant time delay has been obtained.

The values selected for compensating the circuit

$$R_L = \frac{1}{2\pi f_o C_T} \text{ and } L_1 = \frac{R_L}{4\pi f_o}$$

are not necessarily productive of the best phase and amplitude response. Other authors[2] have shown that more nearly constant time delay and amplitude response may be obtained by using slightly different values of R_L and L_1.

If we designate the ratio of load resistance, R_L to capacitive reactance X_c at the top frequency by p, and the ratio of inductive to capacitive reactance at f_o by s, we have

$$p = \frac{R}{X_c} = 2\pi f_o C_T R_L$$

$$s = \frac{X_L}{X_c} = (2\pi f_o)^2 L_1 C_T$$

The values chosen in the preceding case are $p = 1.0$ and $s = 0.5$. If, instead, we use $p = 0.85$ and $s = 0.3$, the time-delay curve is almost precisely flat, and the gain variation over the frequency band is slightly less than in the case previously described. However, this latter arrangement entails the use of a lower value of load resistor, so that the gain is decreased 15 per cent at all frequencies.

As a typical case, consider a video amplifier employing Type 1851 tubes. The total load-circuit capacitance ($C_{in} + C_{out}$ plus wiring and strays) is about 25 $\mu\mu f$. Let the top video frequency be 3 Mc, in which case $X_c = 2120$ ohms. If $p = 1$ the load resistor would also be 2120 ohms, and the coil inductance (for $s = 0.5$) would be

$$\frac{2120}{2 \times 2\pi f} = 56\ \mu h$$

The results would be satisfactory on a basis of constant gain and time delay, with the actual gain equal to 19 per stage, for a tube having a mutual conductance of 9000 $\mu mhos$.

Use of $p = 0.85$ and $s = 0.3$ would require a resistor of 1800 ohms and a coil inductance of 33.5 microhenries, with a gain of about 16 per stage. In general, in practical cases, the value of L_1 may vary somewhat from the prescribed values, as may also the value of R_L. When several stages are used in cascade it is very important, however, to obtain as nearly uniform gain and time-delay curves as possible for each stage, since, if L_1 and R_L are not chosen according to the prescribed values, it will be difficult to make counteracting changes in alternate stages to produce an overall characteristic which does not depart too much from the theoretical values (i.e., over and under compensation in successive stages).

CIRCUIT 3—π-TYPE LOW-PASS FILTER EMPLOYED AS COUPLING ELEMENT BETWEEN PLATE OF T_1 AND GRID OF T_2 (SERIES PEAKING)

This type of circuit possesses certain advantages over the shunt-peaking arrangement because it effectively separates C_2 from C_1 by means of L_2. (They would normally be in parallel in the shunt peaking circuit, appearing across R_L and L_1.) It affords a greater gain per stage with a smaller departure from constant time delay. The action of the circuit in preserving the high-frequency response of the amplifier may be described briefly as follows: A voltage $eg\,GmR_L$ is considered to exist across R_L (with C_1, L_2 and C_2 removed). C_1 is next considered to exist across R_L, which causes attenuation of the higher frequencies and produces a voltage

$$\frac{eg\,GmR_L}{\sqrt{1+(2\pi fC_1R_L)^2}}$$

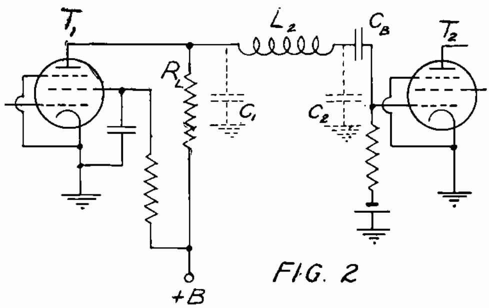


FIG. 2

across R_L and C_1, in parallel. This voltage is applied to the voltage divider consisting of L_2 and C_2 in series, and the resultant drop across C_2 is maintained constant by resonant rise effects in L_2C_2, which counteract the attenuation produced by C_1.

The performance of the circuit depends upon a number of factors. One of these is the ratio of the two capacitances, C_1 and C_2, which appear at the terminals of the low-pass filter. Let this ratio of C_2/C_1 be m. C_1 includes the output and stray wiring capacitances associated with tube No. 1. C_2 includes the input and wiring capacitances of tube No. 2, as well as the stray capacitance between the blocking condenser, C_B, and ground. Note, from Figure 2, that the blocking condenser may be connected at either end of L_2 to assist in adjustment of the value of m.

The value of total capacitance $(C_1 + C_2)$ may be determined experimentally by the methods described for use with the shunt-peaking circuit (L_2 is shorted in this measurement). To measure C_1 open L_2

and find the frequency at which the gain of T_1 is 70.7 per cent of its low-frequency value. A vacuum-tube voltmeter of known input capacitance may be used across R_L as an indicating device, and its contribution to C_1 must be taken into account.

An alternative method of measuring C_1 makes use of the "Q" meter, as described in connection with shunt-peaking circuits.

It has been pointed out by Albert Preisman of R.C.A. Institutes that, for best performance, C_2 should be at least twice C_1, i.e. $m \geqq 2$. This condition is fulfilled in most practical applications. If m is found to be less than 2, the ratio may be adjusted by proper disposition of the circuit components (such as the d-c blocking condenser for T_2) or by connecting small capacitances across the filter input or output terminals. It should be noted, however, that the use of additional capacitance at either end of the filter to produce the desired value of m will result in a loss in gain, for the absolute gain is inversely proportional to the total capacitance in the load circuit.

With C_2 and C_1 known from measurement, the first step in designing the coupling network is to select f_o, the top frequency in the video band for which constant gain is desired. This value of f_o, in conjunction with C_1 determines the inductance L_2 of the series-peaking coil.

To find L_2, let f_r be the resonant frequency of L_2 and C_1, i.e.,

$f_r = \dfrac{1}{2\pi\sqrt{L_2C_1}}$. The value f_r is chosen to be $\sqrt{2}$ times the top video frequency, f_o. Therefore, L_2 is determined from

$$L_2 = \frac{1}{2(2\pi f_o)^2 C_1}$$

The inductive reactance of L_2 at f_o is equal to one-half the reactance of C_1 at the same frequency.

The value of R_L, the plate-load resistor, with $m = 2$ and $f_o = f_r/\sqrt{2}$, is equal to one-half the reactance of C_1 at the top video frequency, f_o. Since $m = 2$, R_L also equals one and one-half times the total load-circuit, capacitive reactance at f_o.

The procedure for compensating a stage may be itemized as follows: (1) Measure C_1 and C_2 and, if necessary, adjust C_2/C_1 to be at least 2; (2) make the terminating resistor R_L equal to one and one-half times the reactance of $(C_1 + C_2)$ at the top video frequency, f_o and connect the resistor across the plate end of the filter network; (3)

obtain a coil which resonates with C_1 at $\sqrt{2}$ times the top video frequency, or use the relation $L_2 = \frac{2}{3} (C_1 + C_2) R_L^2$. The resistance of coil L_2 is immaterial as long as the coil Q is greater than 20.

Under some conditions it might be necessary to work out of a high plate-circuit capacitance into a low grid capacitance. In such a case the value of C_2/C_1 may be more nearly ½ instead of 2. In that event, the values of L_2 and R_L are the same as those calculated for $m = 2$, but the load resistor is connected across the output terminals of the network, i.e., across the smaller terminating capacitance. A reciprocal action permits interchanging the point of resistor termination in this special case, and results in operating characteristics which are the same as for the more likely case discussed previously. Specifically, the coupling network may be turned end for end without affecting its operation.

The basic design equations, to be used for any value of m, with the top video frequency chosen to be 0.707 times the resonant frequency of L_2 and C_1, are

$$R_L = \frac{1}{\sqrt{2m}\,\omega_0 C_1}$$

$$L_2 = \frac{1}{2\omega_0^2 C_1}$$

where $\omega_0 = 2\pi$ times the top video frequency. If the values suggested above $\left(R_L = \frac{3}{2}\,\frac{1}{(C_1 + C_2)\,\omega_0} \text{ and } L_2 = \frac{2}{3}\,(C_1 + C_2)\,R_L^2\right)$ are used in the video stage, the gain and time-delay characteristics are essentially flat out to f_0. The absolute value of gain is 50 per cent greater than the gain experienced in a shunt-peaking circuit having the same total load-circuit capacitance and the same value of f_0. The departure from constant time delay is $0.0113/f_0$ seconds. For a 3-Mc band the variation in time delay is 0.004 microseconds, which is somewhat smaller than Δ_T in the shunt-peaking case. The total time delay is greater with series peaking, but, of course, this is relatively unimportant, since, within reason, the magnitude of T is of no consequence, provided the departure from a constant value is small.

The series-peaking circuit merits serious consideration on the basis of these results. It may be expected to exceed the shunt-peaking circuit in performance in cases where the capacitance distribution is favorable, or when the ratio of capacitances can be adjusted to the desired value without causing a decrease in gain below the value experienced with shunt peaking. Note that operation with values of m less than 2 will cause the gain characteristic to peak at the high end. While this effect is not desirable generally, it may find some utility for peaking purposes in amplifiers in which the high-frequency gain in other stages of the chain is deficient. Such a condition might exist by virtue of the high-frequency attenuation experienced in a concentric transmission line, wherein a drop at the top end of the video band must be overcome by subsequent peaking stages.

Circuit 4—Combination of Circuit 2 and Circuit 3

This circuit provides certain advantages over either No. 2 or No. 3 used singly. As described by E. W. Herold[3] it has the following char-

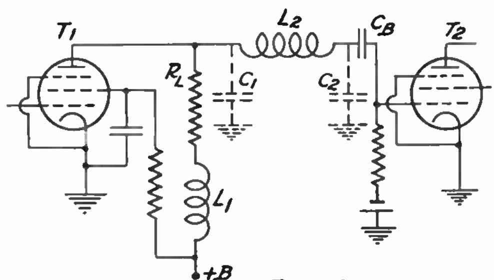


FIG. 3

acteristics: for a given total load-circuit capacitance C_T and prescribed top video frequency f_o, the load resistor which may be used (maintaining constant gain up to f_o) is approximately 80 per cent greater than in the case of simple shunt peaking. This means, of course, 80 per cent higher gain per stage, for the gain is $Gm\,R_L$, when the circuit is properly compensated. The departure from constant time delay is roughly equal to that experienced in a simple series-peaking circuit.

The disposition of circuit components required to produce the 80 per cent increase in over-all gain are as follows:

$$m = C_2/C_1 = 2 \qquad L_1 = 0.12\ (C_1 + C_2)\ R_L^2$$

$$R_L = \frac{1.8}{\omega_0 (C_1 + C_2)} \qquad L_2 = 0.52\ (C_1 + C_2)\ R_L^2$$

To design a stage similar to that shown in Figure 3 the procedure is as follows: (1) Select the top frequency f_o to be passed with uniform gain; (2) make $m = C_2/C_1$ equal to 2; (3) determine the total load-circuit capacitive reactance at the top frequency; (4) choose a load resistor equal to 1.8 times this total load-circuit reactance at f_o; (5) calculate L_1 and L_2 from the formulas given above.

The following table itemizes the performance characteristics of the several types of circuits described in this section.

Cir. No.	Type of H.F. Comp.	$\frac{R_L}{2\pi f_o C_T}$	ΔT μsecs.	L_1	L_2	$\frac{C_2}{C_1}$
1	none	1	$\frac{.035}{f_o\ \text{Mc}}$			
2	shunt	1	$\frac{.0231}{f_o\ \text{Mc}}$	$.5C_TR_L^2$		
3	series	1.5	$\frac{.0113}{f_o\ \text{Mc}}$		$.67C_TR_L^2$	2
4	shunt and series	1.8	$\frac{.015}{f_o\ \text{Mc}}$	$.12C_TR_L^2$	$.52C_TR_L^2$	2

SECTION II

Low-Frequency Considerations

The presence of the unavoidable shunt capacitance in the output circuit of a video stage impairs the operation of the device only at high frequencies. Below 100 kilocycles the shunt reactances have negligible effect. Consequently, for frequencies ranging between 100 kilocycles and 200 cycles, probably all types of standard video stages perform creditably whether they are compensated or not.

In the frequency range extending below 200 cycles, the gain and time-delay characteristics of a video stage are also subjected to variations from the ideal conditions. These are caused by the inability of the d-c blocking condenser in the grid circuit, in combination with the grid leak, to pass the low video-frequency signal components with their proper amplitude and phase composition, or they may be due to inadequate by-passing of a cathode bias resistor.

It will be recalled that departure from constant gain and time delay in the high-frequency portion of the video band causes imperfect reproduction of horizontal detail. Insofar as the horizontal detail is concerned the video amplifier could be cut off at 10 kc (passing no signals below this frequency), since the lowest frequency involved in reproducing a picture along any horizontal line is equal to the line repetition rate, 13,230 cycles per second.

The function of the very low video frequencies is to supply the background of the reproduced picture. Failure of a video stage to pass these frequencies in their original wave form will generally cause the background to vary in intensity from top to bottom of the picture. As an example of this effect, consider the situation existing when an all-white screen is to be transmitted. If the low-frequency characteristics of the amplifier are inadequate, the background of the picture will be non-uniform, i.e., there will be a gradual variation in shading in the vertical direction. This effect is least pronounced when an all-white or black screen is transmitted. Maximum departure from the desired background conditions occurs when the screen is half-black and half-white, about a horizontal center line. This matter will be discussed in greater detail later in this section.

Insofar as circuit performance at the low-frequency end of the band is concerned, it can be said that maintenance of proper phase characteristics is more important than maintenance of constant gain. However, even though the phase and gain characteristics are known, it is only with considerable difficulty that the response of the system to a low-frequency pulse may be predicted; that is, the performance of a video stage at low frequencies cannot be judged readily from measurements taken on a monotone basis. Consequently, the low-frequency performance of the amplifier may be more readily evaluated on a square-wave basis. This can be accomplished experimentally by applying a low-frequency square wave and by observing the amount of distortion of the output wave form. Or, the problem can be approached analytically, as shown below.

Let Figure 4 represent a grid-coupling circuit and Figure 5 a square wave (60 cycles base frequency) to be passed through it. We wish to establish some means of determining the effect of the grid-circuit time constant on the tilt appearing in the square wave.

The voltage drop across C, due to the application of a voltage E, may be written rigorously as $E_c = E(1 - \varepsilon^{-t/CR})$, where t is the time interval following the application of E to C and R.

For relatively large time-constant circuits, in which the current through R is essentially constant for a short interval following the application of the pulse, we may write

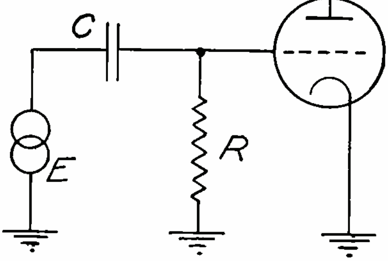
$$E_c/E = t/CR$$

$E_c/E \times 100$ is the percentage drop in amplitude of the rectangular wave during the duration of the pulse, t. E is measured from average value to peak value of the wave. Figure 6 shows the wave after having passed through the grid-coupling circuit. A 10 per cent drop in voltage amplitude is assumed for illustrative purposes.

Note that the amplitude of the pulse approaches the average value of the wave, that is, if the first pulse were allowed to decay indefinitely the total fall in voltage would not exceed the peak value of the wave. If the wave form of the pulse is changed, so that the positive and negative loops are of unequal time duration, the slope of the wave top becomes less pronounced for the long pulse and more pronounced for the short one. Equal positive and negative pulses in a square wave impose the most severe requirements on the grid-coupling circuits, for a given permissible wave-top tilt.

The formula given above may be used to advantage in determining the values of C and R in the grid circuit for a given percentage drop

FIG. 4



in amplitude of a square wave, by letting t equal the time duration of the pulse and by calculating C and R. Note that any values of C and R which produce a given time constant will result in the same low-frequency response. Practically however, extreme values of C and R should not be used, for large grid resistors will influence the d-c bias considerably in the event of grid current (due to gas or grid emission) in the tube. On the other hand, low values of R will require large values of C, which may cause an increase in total load-circuit capacitance (effective at the high frequencies) due to the stray capacitance from the physically larger blocking condenser to ground.

As an example of the application of the square-wave analysis to a grid-coupling stage, consider a square pulse, of 60 cycles base frequency, applied to a grid circuit containing a 0.25-μf blocking condenser and 0.5-megohm resistor. Since the duration of a single pulse is 1/120 second, the slope in wave top may be computed from

$$\frac{E_c}{E} = t/CR = \frac{1}{120} \times 0.25 \times 0.5 = 6.7 \text{ per cent.}$$

If, on the other hand, we calculate the relative voltage response at 60 cycles (on a monotone basis), it is found that the response at 60 cycles is better than 99.9 per cent of the response with an infinite time constant. This indicates the necessity for examining the low-frequency characteristics of a video stage on a rectangular pulse basis.

There are several arrangements available for compensating for deficiencies in the grid-coupling circuits of video amplifiers. Some involve rather complicated resistor-capacitor networks, placed in the plate and grid circuits, which provide equalization of frequency response and cancellation of phase shift down to quite low frequencies. The intricacies and mathematical analysis of these circuits will not

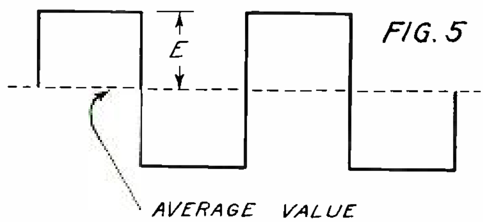


FIG. 5

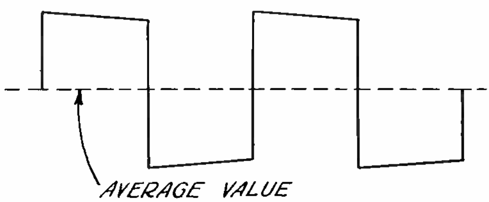


FIG. 6

be included here, for it is felt that one simple form of correction circuit, used widely in video amplifiers, and discussed here, should be all that is required for proper operation.

The simplest arrangement includes a resistance and capacitance in parallel, connected as shown in Figure 7. It can be shown that satisfactory low-frequency response can be achieved with this type of compensation provided the time constant in the grid circuit is approximately equal to the time constant of the video load resistor, R_L, and the decoupling condenser, C_F. This is very nearly true for all frequencies at which the value of R_F, the decoupling resistor, is greater than ten times the reactance of C_F. As a practical example, let R_L be 2000 ohms (video load), $C_F = 16$ μf and R_F 2500 ohms. Then the grid time constant must be equal to $R_L C_F = 0.032$ seconds. This would require a 0.25-megohm leak and only 0.125-μf grid-blocking condenser.

Note the appreciable reduction in grid-circuit time constant below the uncompensated value. The low-frequency response, in this case, is satisfactory down to 60 cycles. To extend the range to 30 cycles, the only change required is to double the size of R_F, the decoupling resistor.

This circuit has advantages over and above its low-frequency-response compensation. One of these is its filtering action against hum originating in the *B* supply. Another advantage, also due to filtering action, is the suppression of motor-boating tendencies at very low frequencies.

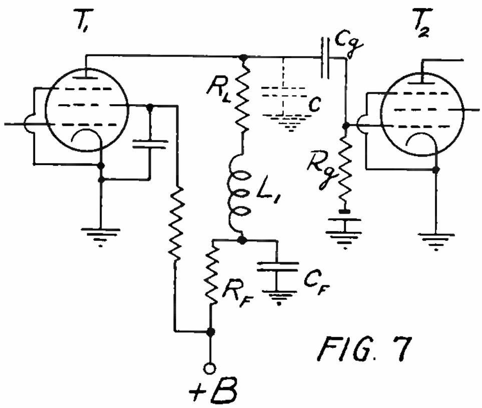


FIG. 7

The general procedure in compensating a stage at low frequencies is to select a value of C_F, say 16 μf. This, in combination with R_L, which is determined by high-frequency considerations, gives the time constant which must be obtained in the grid circuit. The decoupling resistor should be made as large as possible, consistent with obtaining the required d-c plate voltage from a normal *B* supply.

It should be noted that this type of plate compensation, when used to counteract deficiencies in the grid-coupling circuit, will not give perfect response down to very low frequencies (and to d.c.) except in the theoretical (and, practically, not applicable) case in which the decoupling resistor is infinite.

One of the major problems encountered in video-amplifier design is that of obtaining the required d-c grid biases. Three methods are available: (1) Battery bias; (2) bias obtained from a bleeder resistor, whose voltage is obtained from the plate power supply, generally by the insertion of a small resistor in the *B*-return lead; (3) cathode or self-bias. Cathode bias is, for several reasons, to be preferred. One of its advantages is that it permits the use of larger grid-leak resistors than in the case of fixed bias. A second advantage is that it involves only one small resistor and a by-pass condenser. Principally, however, its utility lies in the fact that deficiencies in the cathode by-pass condenser may be compensated in the plate circuit by the insertion of a parallel *RC* network at the low-potential end of the video load. In this case, the compensation may be made exact at all frequencies down to direct current, with practical values of circuit components. This is to be contrasted with the compensating effect of the same type of plate network when used to counteract deficiencies in the low-frequency response of grid-coupling circuits. In this case compensation is exact only for frequencies at which the feeding or decoupling resistor is very large compared to the reactance of the plate-filter condenser.

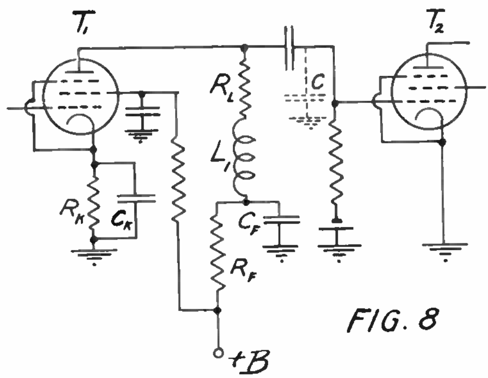


FIG. 8

The circuit for compensating for cathode-bias-network effects is that of Figure 8. In the cathode-bias case, the relation between the various resistors and condensers for compensation at all low frequencies is as follows:

$$C_K R_K = C_F R_F$$
$$R_F = R_K (Gm\, R_L)$$
$$C_F = C_K / Gm\, R_L$$

Note the presence of the GmR_L term. This product is the gain of the stage at frequencies in the mid-range of the video band. Its presence suggests a method of compensating the stage at low frequencies without a specific knowledge of *Gm* or the high-frequency gain. The procedure is as follows:

Connect the stage as shown in Figure 8 with both C_K and C_F sufficiently large to have negligible reactance at a frequency of 10 kc. Adjust the cathode bias resistor to produce the desired grid bias. Apply to the grid a voltage having a 10-kc frequency and of sufficient magnitude to produce a readable deflection on an oscilloscope or vacuum-tube voltmeter connected from plate to ground. With R_L set at its proper value to give the desired high-frequency performance note the reflection of the indicating device. This is a measure of GmR_L, the high-frequency gain. Now remove both C_K and C_F and adjust R_F (with R_K fixed at its correct bias value) to produce the same indicator deflection (constant-input volts maintained at the grid). Then shunt the cathode-load resistor with a by-pass condenser of the value to be used in circuit (say 25 μf). This being done, the only remaining step to achieve complete compensation is to shunt the decoupling-resistor R_F with a condenser which makes the time constant of the cathode circuit equal to that of the plate-filter circuit, i.e., $C_F R_F = C_K R_K$.

Values which might be employed in an 1851 video stage are:

$R_L = 2000$ ohms (depending upon video-band width)

$R_K = 150$ ohms

$R_F = 2500$ ohms

$C_K = 25\ \mu f$ electrolytic

$C_F = 1.5\ \mu f$.

It will generally not be necessary to compensate in the plate circuit of one stage, for deficiencies in grid coupling and cathode-bias circuits employed in the same stage, for the cathode-bias operation permits the use of a larger grid leak than in the case of fixed bias. This will aid in preserving the low-frequency characteristics of the grid-coupling circuit.

It is difficult to prescribe exactly the minimum values of grid-circuit time constants which may be used in a video stage, for the choice of time constant will depend upon the permissible slope of the wave tops, and upon the number of stages in the chain. Generally, in video stages containing no low-frequency plate compensation (for deficiencies in grid circuits) the time constant of each grid circuit should be from 10 to 15 times the period of the lowest frequency to be transmitted.

One should not attempt to compensate for deficiencies in a number of grid or cathode-circuit time constants in one plate-compensating network, because the results will be unfavorable unless the departure from flat-top performance on a square-wave basis is small in each stage. Best results are obtained, in a multi-stage amplifier, by compensating in each successive plate circuit.

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[1] "Analysis and Design of Video Amplifiers," S. W. Seeley, C. N. Kimball, R.C.A. REVIEW, Vol. II, No. 2, Oct. 1937.

[2] Freeman and Schantz, "Video Amplifier Design," *Electronics*, August, 1937.

[3] "High Frequency Correction in Resistance-Coupled Amplifiers," E. W. Herold, *Communications*, August, 1938.

OBSERVATIONS ON SKY-WAVE TRANSMISSION ON FREQUENCIES ABOVE 40 MEGACYCLES*

By

D. R. GODDARD

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Summary.—The results of daily observations at Riverhead, N. Y., since September, 1937 of European 40 to 45-megacycle-per-second transmitters are reported. Measurements of field strength were made on English, French, and German television signals. Multipath propagation of the English video channel was observed optically and the difference in path length determined.

THIS paper deals briefly with the results obtained by systematically observing and measuring the field strength of television signals from England, France, and Germany.

The English transmitters located at Alexandra Palace, London, operated on 41.5 megacycles per second for the audio channel and 45 megacycles for the picture channel. The frequency of the French audio transmitter at the Eiffel Tower was 42 megacycles and the Berlin audio transmitter frequency was 42.5 megacycles.

A rhombic antenna 45 feet above ground and directed towards London was used for these measurements. Its length was 400 feet per side and it was arranged so that the dimensions of the major and minor diagonals could be readily changed. This was done so as to facilitate matching the antenna to the vertical arrival angle of the signal. The effective height of the antenna system was about 20 meters.

Antenna adjustments were made by comparing various settings of the rhombic antenna to a reference dipole 45 feet above ground. As comparisons were made on the London 41.5 megacycle signal the results gave the optimum setting for that signal. This setting corresponded to a vertical arrival angle of roughly seven degrees.

Figure 1 shows the receiving equipment used. In the foreground is a television receiver with a small camera mounted over the Kinescope. Only the video amplifier and Kinescope controls were used as it was thought desirable for these experiments to have available greater flexibility than the radio circuits of this set provided. Therefore, the receiver standing to the left of the one just described was designed by Mr. Trevor of this department. This set provided automatic or manual volume control, a minimum noise equivalent of about 30 microvolts, with a band width somewhat less than 5 megacycles,

* Based on a paper presented at the Joint U.R.S.I.-I.R.E. Meeting, Washington, D. C., April 29, 1938, and published in *Proc. of I.R.E.*, January, 1939.

and two diode outputs, one giving a "positive" and one a "negative" image. On the bench is the signal generator and receiving equipment used for signal strength measurements.

Most of the observations took place between 9:45 A.M. and 11:30 A.M., E.S.T. as that interval appeared to correspond approximately to the afternoon schedules of all three countries. On several occasions, however, the transmitters continued on into the afternoon, usually on tone modulation. On November 19 and 20 the English audio trans-

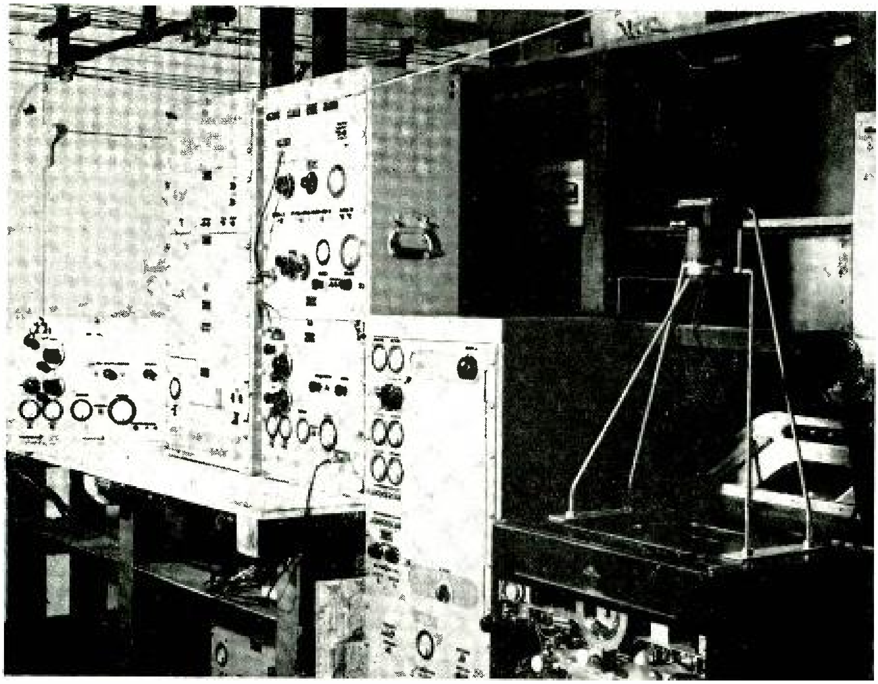

Fig. 1—Ultra-high-frequency receiving equipment used for field-strength determinations and path-delay measurements.

mitter was operated from 5 A.M. to 8 A.M., and 9 A.M. to 11 A.M., E.S.T. On both occasions the signal was first heard at Riverhead a few minutes before seven. On the latter date, however, the signal disappeared at 7 A.M. and did not re-appear until 9:20 A.M. October 16 from 4:00 P.M. to 4:30 P.M. was the only occasion on which the 41.5 megacycle English signal was heard on the evening schedule corresponding to 3:45 P.M. to 5:00, E.S.T.

Figure 2 shows the peak signal strength in decibels above or below one microvolt per meter of the English audio (41.5 megacycles) and video (45 megacycles) signals for every day during the winter of 1937-8 that either or both were heard. The small crosses indicate

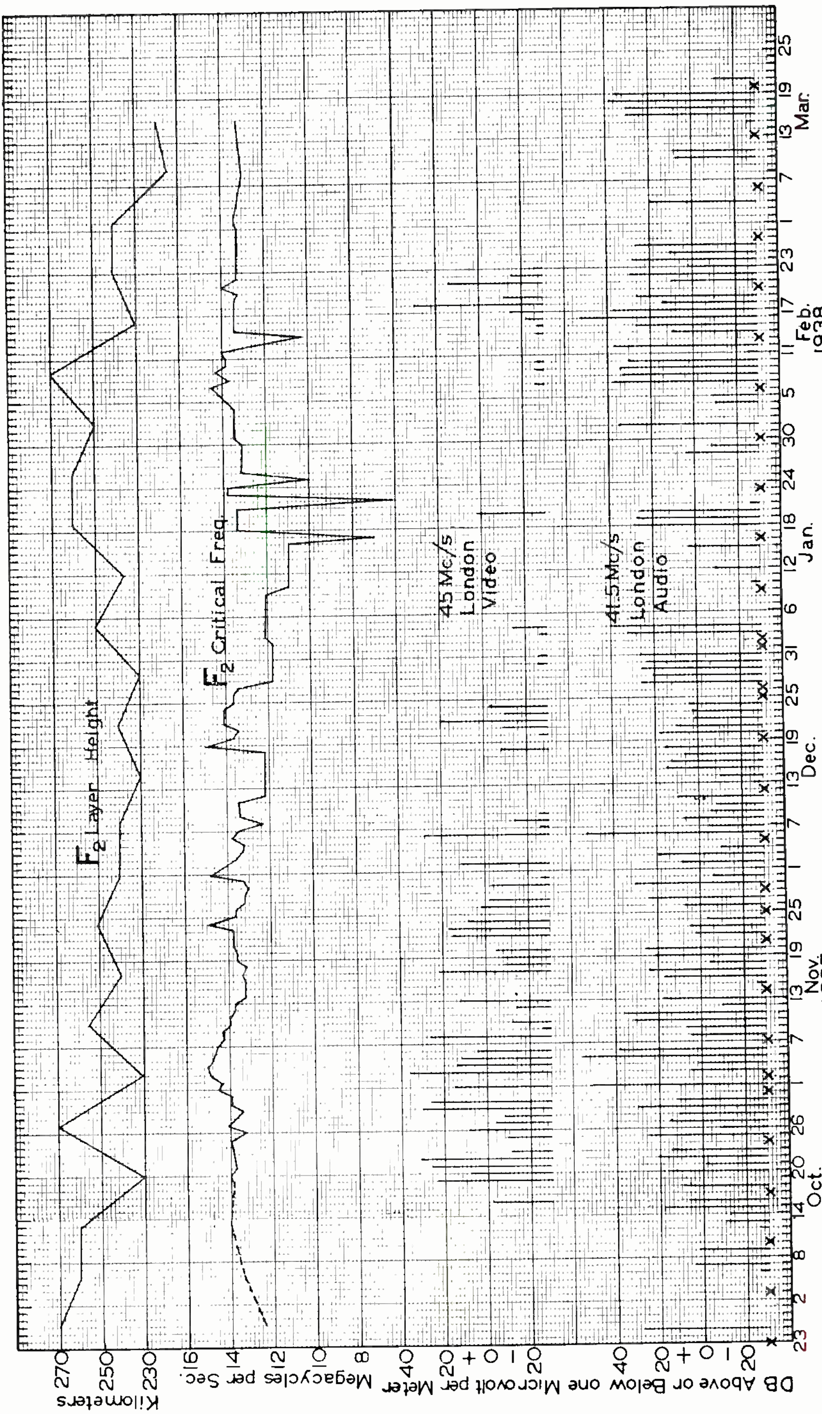

Fig. 2—Comparison of F_2 layer height and critical frequency to observed maximum signal strengths from the London television audio and video channels. The crosses indicate days on which no observations were made.

holidays on which no observations were made. The uppermost curve is a plot of F_2 layer virtual height as broadcast by the National Bureau of Standards. These measurements are taken each Wednesday at noon E.S.T. Directly below this curve is a plot of the critical frequency of the F_2 extraordinary ray. The data for this curve were supplied by Dr. Dellinger of the National Bureau of Standards. Some of these values were not obtained by actual critical frequency measurements, but are close approximations.

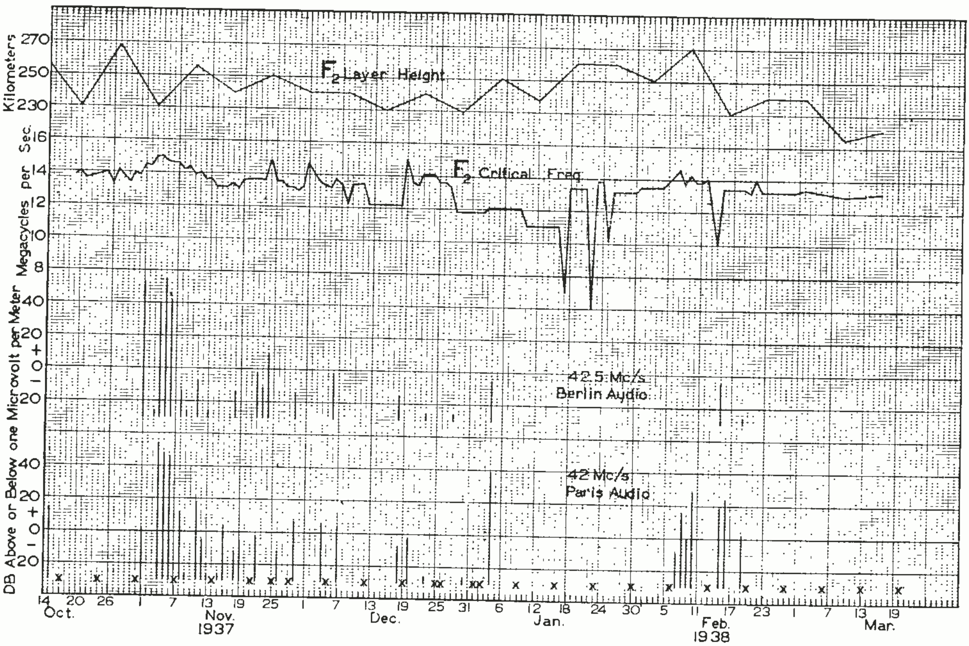

Fig. 3—Comparison of F_2 layer height and critical frequency to observed maximum signal strengths from the Berlin and Paris television audio channels. The crosses indicate days on which no observations were made.

Figure 3 shows the peak signal strength in decibels above or below one microvolt per meter for the French audio (42 megacycles) channel and the German audio (42.5 megacycles) channel for every day during the winter of 1937-8 that either or both were heard. The upper two curves of this figure are plotted from the same data as the corresponding curves on Figure 2.

Inspection of Figures 2 and 3 indicate that a strong signal is not necessarily accompanied by a high critical frequency or low layer height. However, the month of November produced the most consistently strong signals and was characterized by a uniformly high critical frequency. One interesting case was that of February 14. On this day, probably due to a magnetic storm, the noontime critical frequency dropped to 10,150 kilocycles and yet the English 45-megacycle channel was heard faintly and the English 41.5-megacycle channel was quite strong. On December 1, however, the critical frequency rose to 14,700 kilocycles and only a weak signal was observed on 41.5 megacycles and the 45-megacycle channel went unheard.

Of course, it should be pointed out that the critical frequency and layer-height measurements were made at Washington, D. C. at noon while most of the signal-strength measurements were made from one

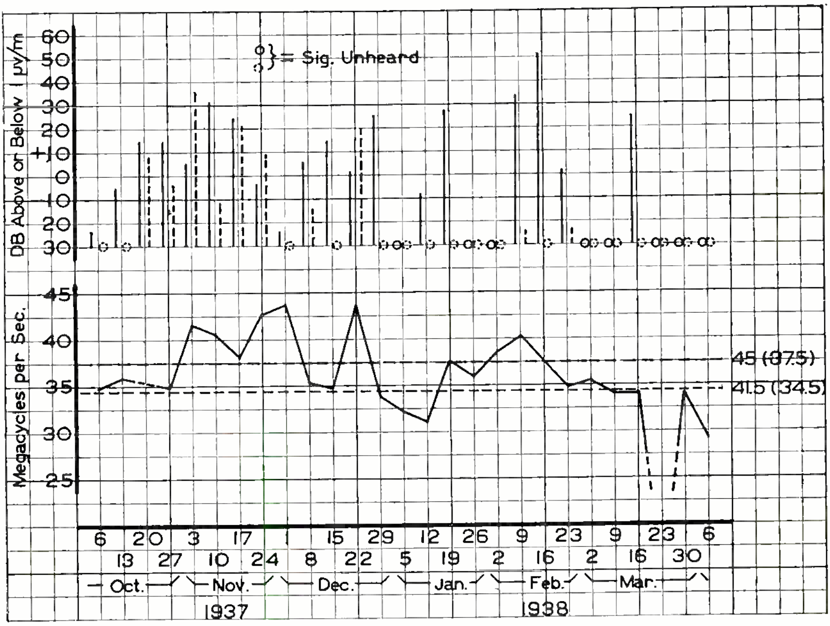

Fig. 4—Comparison of observed signal strengths to maximum usable frequencies interpolated for 2700 kilometers. Vertical solid lines of the upper plot represent maximum signal strengths observed from London on 41.5 Mc. Broken lines represent the same for 45 Mc.

to two hours earlier in the day. Furthermore, as the signals coming from Europe probably traversed the Atlantic ocean in two hops, the places in the ionosphere that caused the signals to return to the earth were something like 1800 and 4500 kilometers northeast of Washington.

An attempt to correlate the maximum usable frequencies taken from the weekly broadcast of ionosphere data from station WWV appears in Figure 4. The lower half of this figure represents the maximum usable frequency interpolated for a distance of 2700 kilometers plotted for each Wednesday during the winter of 1937-8. A wavelength of 2700 kilometers was used as it represents half the distance between Riverhead, N. Y. and London. The upper half of the figure is a plot of maximum signal strength observed on the two English channels for the same days that the ionosphere measurements were made. The broken lines represent the 45-megacycle signal and the solid lines represent the 41.5-megacycle channel. The small circles indicate no signal heard for that day.

From these data it may be seen that at no time was 45 megacycles indicated as useful, the highest values being 43.4 and 43.7 megacycles on December 1 and 22 respectively. On December 1 the voice channel was heard faintly and the video channel not at all, while on December 22 both frequencies were quite strong. Another interesting case is that of October 27, a day having the relatively low maximum usable frequency of 35 megacycles. On that occasion both the 41.5 and the 45-megacycle signals were fairly strong. On March 23, there was a severe magnetic storm decreasing the usable frequency to 15.8 Mc.

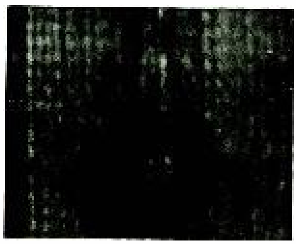

Fig. 5—Photograph of television image received at Riverhead from London showing displacement due to multipath propagation.

These values of maximum usable frequencies were obtained, it is understood, without including the Lorentz polarization term. Application of the Lorentz term would in this case increase the predicted maximum usable frequencies by about 20 per cent. This may be shown graphically either by replotting the maximum usable frequency curve or, as is done in Figure 4, by drawing the horizontal solid and dotted lines opposite 34.5 and 37.5 megacycles of the ordinate scale. These values represent 20 per cent less than the voice and video frequencies of the London transmitters.

Now the correspondence between predicted maximum usable frequencies and observed signal conditions is somewhat improved. Using these horizontal lines as references, prediction for the 45-megacycle channel rises from 38 per cent correct to 73 per cent correct, and for the 41.5-megacycle signal increases from 46 per cent correct, to 77 per cent correct.

On November 5 the signal strength of the English voice channel rose to about 56 decibels above one microvolt per meter. Computation indicates that, neglecting the effect of the ground near the receiving and transmitting antennas, the field strength at Riverhead from the 3-kilowatt English transmitter should have been about 40 decibels above one microvolt per meter. At most, the effect of the ground at both antennas would have increased the field at Riverhead by 12 decibels making the expected field 52 decibels above one microvolt per meter or four decibels less than the peak values actually measured. Of course, the field strength measurements probably include an error of a few decibels, but even so indications were that the attenuation over the path must have been at times nearly zero. Possibly there had been a concentrating or focusing effect of some nature.

The fading on all four signals observed was usually very deep and rather rapid. During days of very strong signals, however, the fading was quite slow, occasionally remaining constant for nearly a minute at a time. Selective fading on the voice channels occurred rarely and was invariably accompanied by a deep dip in signal strength. Two-receiver diversity reception very effectively removed the distortion produced by this selective fading.

These European television signals have been reported heard on a number of occasions from as far west as Phoenix, Arizona. Mr. Clyde Criswell, located near Phoenix has reported hearing all the aforementioned signals during the winters of 1936-7 and 1937-8. Dr. G. W. Kenrick at San Juan, Puerto Rico reported hearing the French, German, and English voice channels several times during the past winter. On most of these occasions the signals were also heard at Riverhead.

"Around the world" echo has been observed on several occasions. A particularly interesting example occurred about 10:45 A.M. on February 17 when "Around the world" echo was heard on the second harmonic (37.8 megacycles) of a Rocky Point, New York transmitter operating on 18.9 megacycles.

Before closing, mention should be made of the results obtained with the Kinescope shown in Figure 1. On February 18 the English video channel became strong enough to synchronize the Kinescope sweep circuits and allow glimpses of the picture being transmitted. Usually these pictures consisted of numerous images superimposed one on another indicating two or more paths of propagation. The path conditions were continually changing and occasionally a single picture would appear quite plainly and with good detail. Figure 5 shows an attempt to photograph this multipath phenomenon. It shows the front view of a man's head and shoulders. As can be seen there are two images and computation shows that the horizontal displacement represents a time delay of about 3.5 microseconds which corresponds to a difference in total length of the two paths from London of something less than 3000 feet.

The helpful suggestions of Mr. Martin Katzin are acknowledged.

RECENT DEVELOPMENTS IN RADIO TRANSMITTERS

By

J. B. COLEMAN AND V. E. TROUANT

RCA Manufacturing Company, Inc., Camden, N. J.

Summary—*This paper outlines and illustrates the mechanical and electrical developments that have been incorporated in the design of radio transmitting apparatus. Various systems of modulation, the application of feed-back and the modulation requirements for broadcast service are discussed. Diagrams and photographs are included to indicate the extent to which multi-element tubes are being applied in high frequency apparatus. Numerous photographs indicate the trend in industrial styling.*

DURING recent years the concentration of engineering ability and skill on transmitter development has resulted in remarkable changes in design technique. This is reflected in reduction in size and cost of equipment, improved efficiency, economy, and performance, and simplification of adjustment and maintenance. The development of new special-purpose tubes has given the designer opportunities to simplify equipment of which he has taken full advantage. The performance of transmitters with respect to fidelity has been kept well ahead of the requirements of the demands of the receiver manufacturer and the public, with the result that the purchaser of modern equipment is assured that he will be protected from obsolescence due to improving standards. The development of new materials and processes has also contributed to equipment value, from the standpoint of durability, reliability, and appearance.

BROADCAST TRANSMITTERS

Many broadcast station owners are replacing their transmitting apparatus to take advantage of the improved performance, simplicity, and low operating expense of modern designs. Five years ago most commercial equipments used low-level modulation followed by linear amplifiers. Today, the trend is toward high-level modulation or high efficiency linear amplifiers to reduce operating expense. There has also been a distinct improvement in performance. Distortion in the order of 2 or 3 per cent rms over the audio range at 100 per cent modulation is not unusual. The carrier noise level has also been reduced to about 60 db below the level for 100 per cent modulation. The operation of tube filaments directly from a-c supplies and the development of efficient high-voltage rectifiers has eliminated rotating equipment.

Mechanically, the designs have been simplified and steel has almost entirely replaced aluminum for frames and panels. The industrial styling trend is reflected in two-tone finishes, smooth contours, and unified appearance. Figure 1 is a photograph of a 50-kw broadcast transmitter which is typical of recent designs.

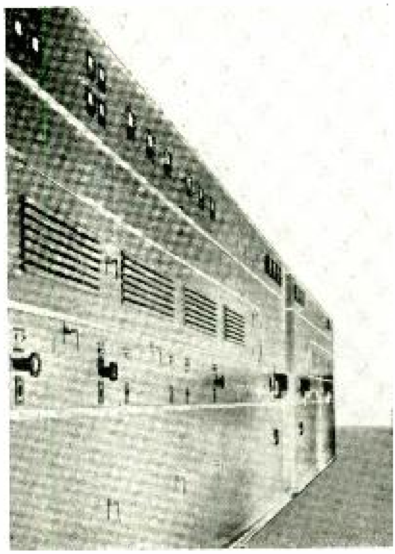

Fig. 1—50-kw broadcast transmitter.

Mechanical Design

The chassis type of construction has proved more economical than the frame type, largely because it facilitates assembly and wiring. Both vertical and horizontal chassis are used; the former where maximum accessibility is required and the latter for optimum utilization of space. The basic exciter unit shown in Figure 2 is an example of the vertical chassis construction. The apparatus is assembled and wired with the "L" chassis in a horizontal position. The chassis are then installed vertically in the frame and the interconnections made. The apparatus is arranged for convenient wiring and accessibility. Studs welded to the chassis are provided to mount the heavier parts, while small pieces are secured by screws fitting tapped holes in the chassis so that any piece may be removed without removing the side shields. Another advantage of this construction is that the tubes are

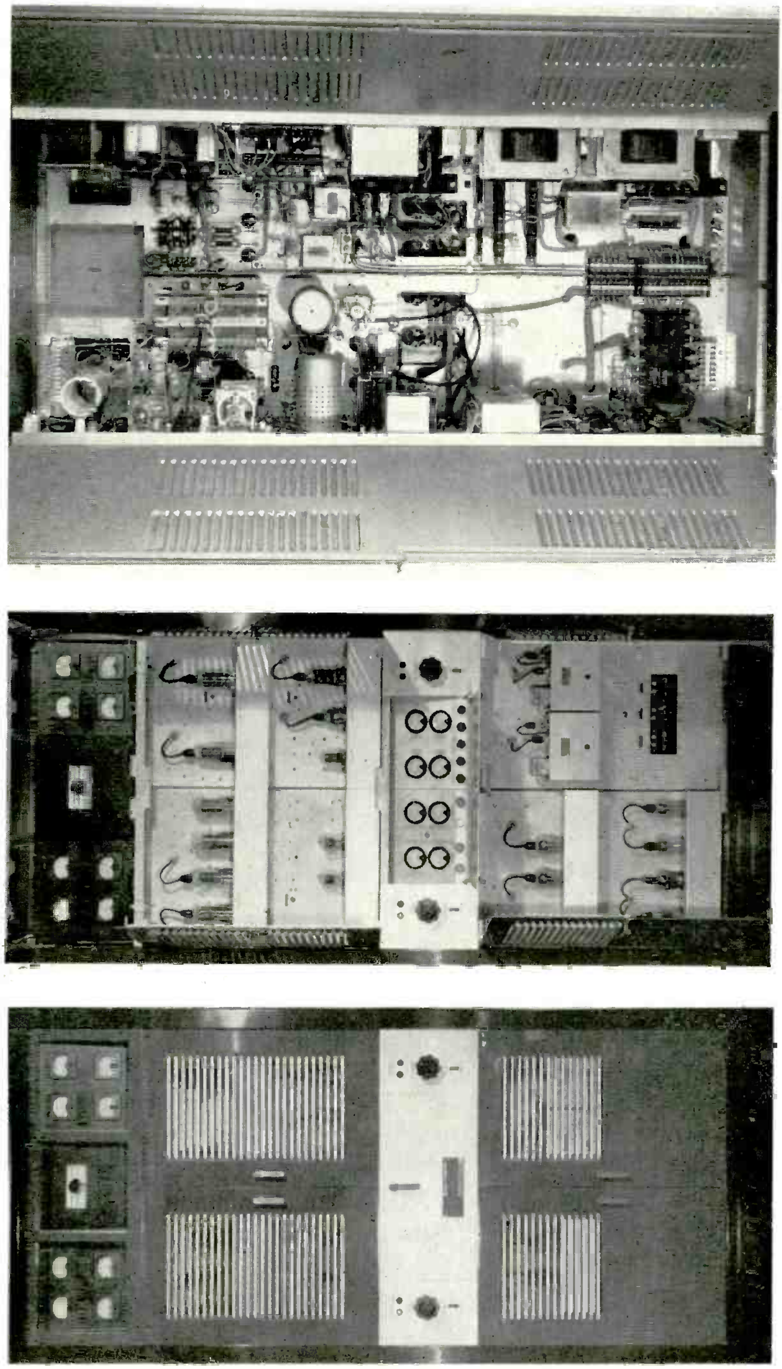

Fig. 2—Front and rear views of basic exciter unit.

mounted toward the front of the cabinet which provides electrical shielding and also protects the apparatus from the heat generated by the tubes. Unusually free circulation of air is provided around the tubes as well as the apparatus by this method of construction.

An example of the horizontal chassis construction is the low-power transmitter shown in Figure 3. This type of construction is easily handled in the factory and offers flexibility in design.

Fig. 3—Low-power transmitter, 100-250 watts, front and rear views.

With either type of construction the frames and chassis are protected against corrosion by either copper plate, or zinc chromate, and a lacquer finish. Copper plate is used wherever ferrous materials are used in radio-frequency fields. All of the coil forms and insulators are ceramic materials.

Electrical Design

The selection of the modulation system is probably the most important factor in the electrical design. Linear amplifiers are being superseded by more efficient circuits which are more readily maintained in adjustment. The following tabulation outlines the tube requirements of the various systems, now in general use, for one power rating.

5-kw Carrier[1]

15% Modulation (Average Program Level)

	Power Input	Efficiency	Total Plate Dissipation
Linear Amplifier	16.70 kw	30.5%	11.60
High Efficiency Linear Amplifier	8.20	61	3.20
High Level Class B	7.85	68	2.85

The integrated percentage of modulation over long periods results in an average percentage of modulation even lower than 15.

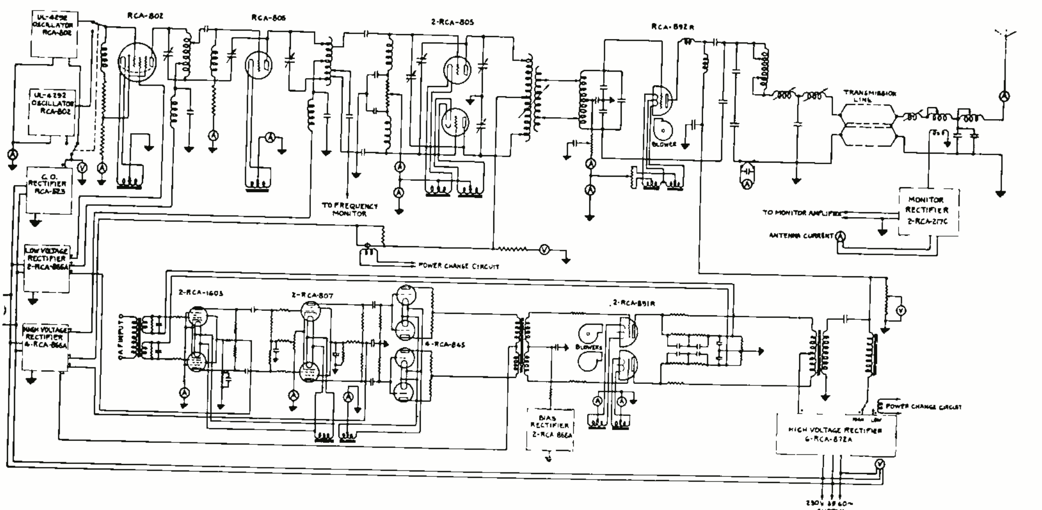

Fig. 4—Simplified schematic of 5-kw broadcast transmitter.

The final choice of the modulation system must also include consideration of the hourly tube cost as an important part of the operating expense.

The development of circuits using controlled degeneration is of considerable importance, particularly in application to transmitters of 5 kw or higher rating, since it permits of economically obtaining satisfactory performance, and results in large savings in first cost and operating cost. In the design of a high-level modulated 5-kw transmitter the application of degeneration permitted a reduction in the idle modulator plate current from 300 to 30 ma per tube, while maintaining the distortion at 2 to 3 per cent rms. With degeneration the

[1] The figures are based on incorporating feed-back in the design.

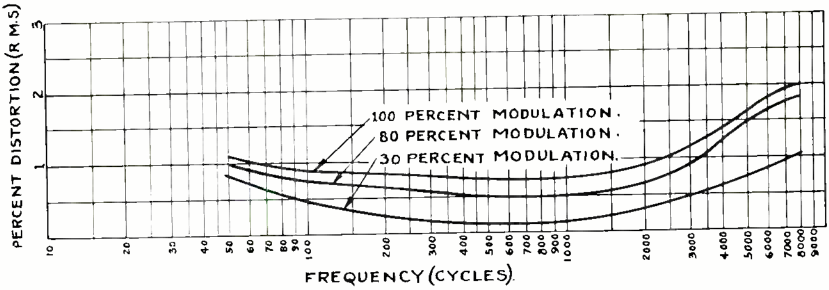

Fig. 5—Distortion characteristic of 5-kw transmitter.

carrier noise level was 65 to 70 db below the level of 100 per cent modulation even with a-c operation of tungsten filaments. Elaborate precautions must be taken to control the phase rotation with respect to frequency characteristic in the circuits to which degeneration is to be applied. For this reason, it is generally impracticable to realize any advantage from the application of degenerative feed-back to existing designs.

The simplified schematic of the 5-kw transmitter, Figure 4, shows the feed-back circuit. A potentiometer on the primary of the modulation transformer provides the voltage that is introduced, out of phase, on the input of the audio system. The hum or noise generated in the r-f power amplifier appears across the modulation transformer and is introduced, out of phase, on the input of the audio system. Thus hum or noise generated in the modulated amplifier is also reduced by the feed-back. The distortion vs. modulating frequency and the audio-frequency response are shown in Figures 5 and 6.

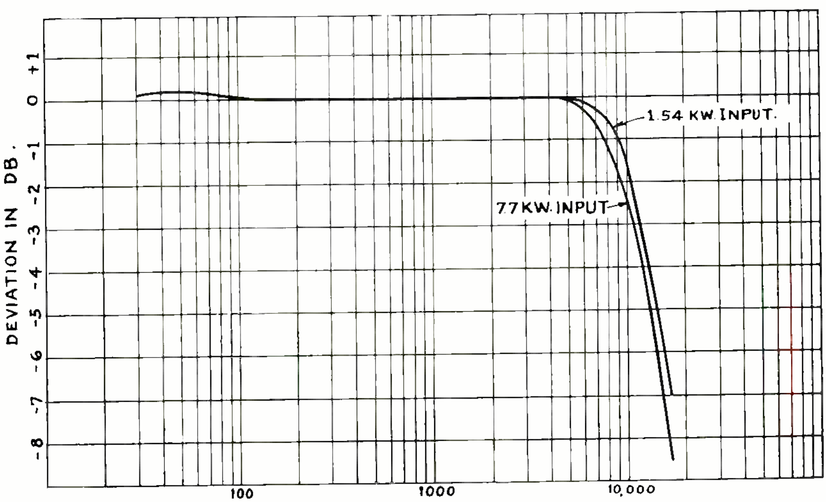

Fig. 6—Audio-frequency response of 5-kw broadcast transmitter.

The tabulation shows that for 5-kw carrier Class B high-level modulation required relatively little dissipation, in fact, less than some 1-kw designs. This permitted advantage to be taken of a radically new development—the application of direct air cooling to high-power tubes, which had previously required a circulating water system for adequate cooling. Preliminary tests indicated that a relatively small radiator and blower system would maintain the anode at a safe

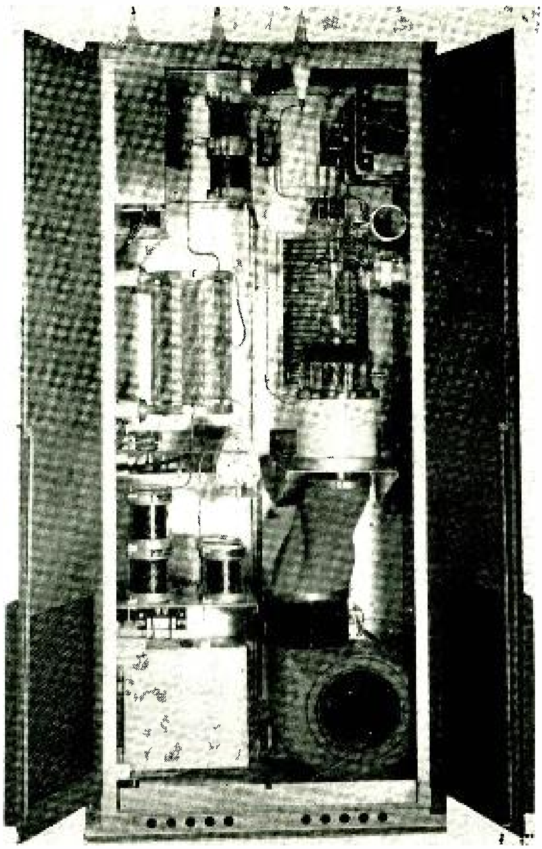

Fig. 7—Rear view of radio frequency unit of 5-kw broadcast transmitter.

operating temperature. The advantages are obvious; no cooling unit required, no danger of freeze-up, lower installation cost, elimination of electrolysis, no expensive insulating hose column, no flow and temperature indicators. The air-cooling equipment in the 5-kw transmitter is shown in Figure 7. A low-speed blower flexibly mounted provides the required air flow. The figures also show the adaptability of the vertical chassis to air cooling.

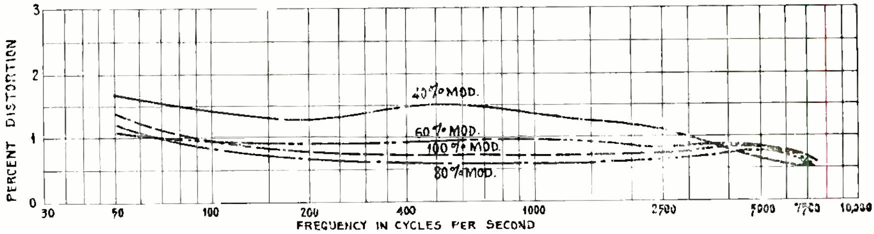

Fig. 8—Distortion characteristic of basic exciter unit operated as 250-watt transmitter.

The selection of a modulation system is largely controlled by the required distortion limits. It is generally conceded that most observers cannot detect 3 per cent rms distortion and that very few observers can detect 2 per cent, particularly if the harmonic content is confined largely to low-order harmonics, second, and third. Furthermore, the entire system, from microphone to loudspeaker, must be considered in determining a practical value. Doubtless, as the art progresses lower distortion will be required. For these reasons, 3 per cent rms distortion is being accepted as a practical and economical limit. Figure 8 shows a typical distortion curve for the low-power basic exciter unit shown in Figure 2 when operated as a 250-watt transmitter.

Consideration is also given to the fact that 100 per cent modulation at the higher audio frequencies is never required. Assuming that a program consists of speech, orchestral numbers, or instrumental solos of sufficient duration for the gain to be accurately adjusted for 100 per cent modulation, the maximum percentage of modulation for any frequency is shown in Figure 9. It is interesting to note that a flute solo

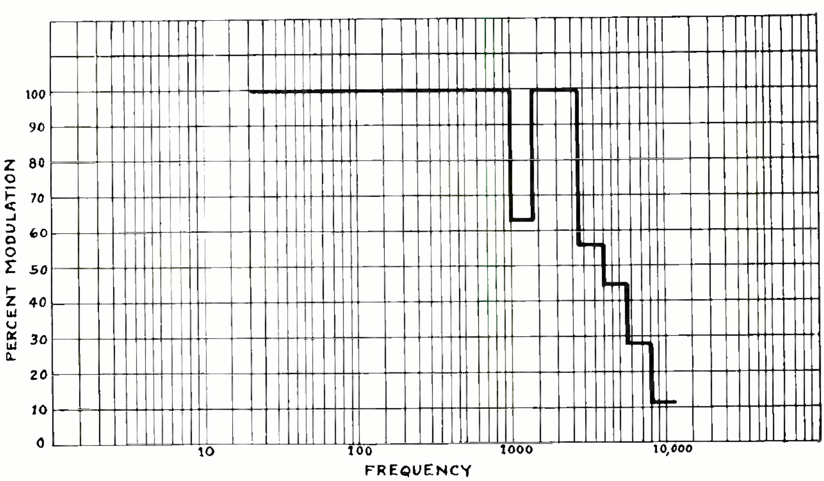

Fig. 9—Maximum percentage of modulation for any type of program.

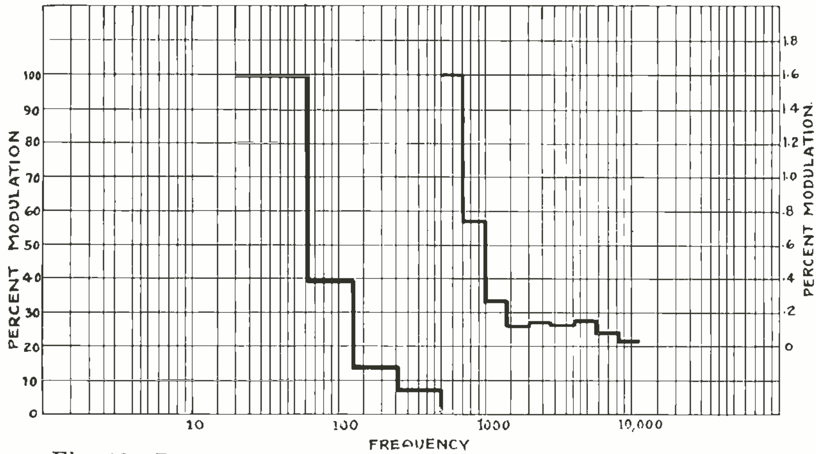

Fig. 10—Percentage of modulation vs. frequency likely to occur in a program.

imposes the most stringent requirements. If we assume that the transmitter is adjusted for 100 per cent modulation for the highest absolute sound pressure that is likely to be broadcast, regardless of its source, we obtain the results shown in Figure 10. For purposes of comparison, the modulation percentages for a seventy-five piece symphony orchestra and for normal speech are shown in Figures 11 and 12. From these data it is obvious that more than 50 per cent modulation above 4000 cycles will never be required and that even 10 per cent modulation at 2500 cycles will be extremely rare. It would probably be satisfactory to base performance (distortion) on 100 per cent modulation at 30 to 1000 cycles, 50 per cent at 2500 cycles, 25 per cent at 5000 cycles and 10 per cent at frequencies above 7500 cycles.

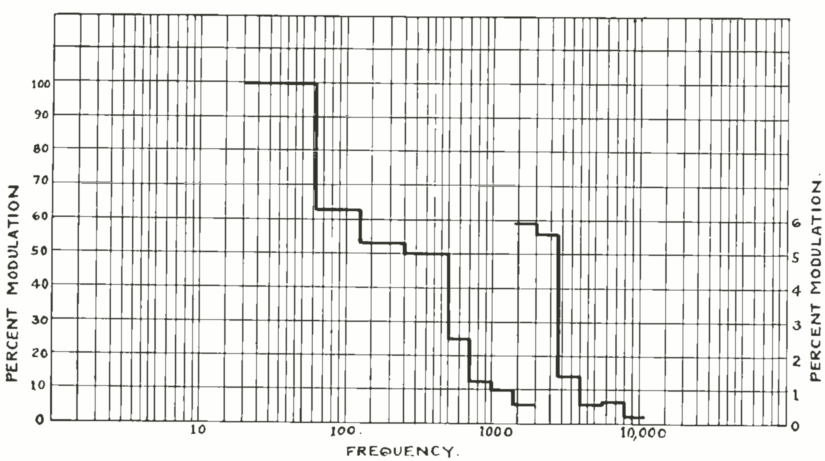

Fig. 11—Percentage of modulation vs. frequency of symphony orchestra.

During the last few years, rotating machines as a source of filament excitation, bias or plate supply have been eliminated. A technique has been developed for the use of a.c. in filament heating, whereby degeneration and multi-phase filament heating are employed to minimize the noise generated in tungsten-filament tubes. A satisfactory noise level is obtained in the case of small transmitters, using thoriated-filament tubes, by using center-tapped resistors, and carefully choosing the type of tube used for each application. It is necessary to investigate each design to be certain that the carrier noise does not change with the percentage of modulation, and that no objectionable sum and difference frequencies are generated by combinations of carrier noise frequencies and modulation frequencies. It has been

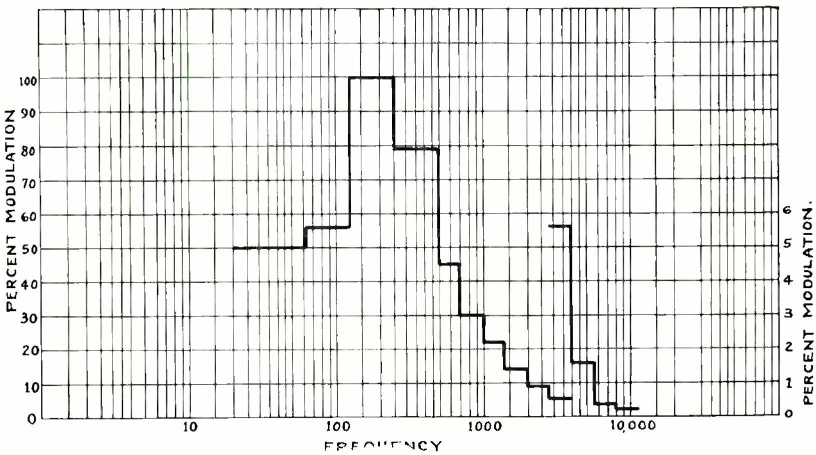

Fig. 12—Percentage of modulation vs. frequency for speech.

found that sum and difference frequencies of this type, having an amplitude greater than 0.1 per cent of the maximum signal level are definitely objectionable.

Another important consideration for a broadcast transmitter is the frequency stability of the emitted wave. The development of quartz crystals having a temperature coefficient of the order of $1/10^6/C^0$ makes it possible to eliminate elaborate constant-temperature chambers and substitute relatively simple temperature-control devices. Such a holder was recently described by Mr. W. F. Diehl.[2] The air-gap crystal mounting is surrounded by a heater which is in close thermal contact with the bimetallic thermostat. A "compensator" in contact with the thermostat reduces the effect of changes in the ambient temperature.

[2] A new Piezo Electric quartz crystal holder with thermal compensator by W. F. Diehl, RCA REVIEW for October, 1936.

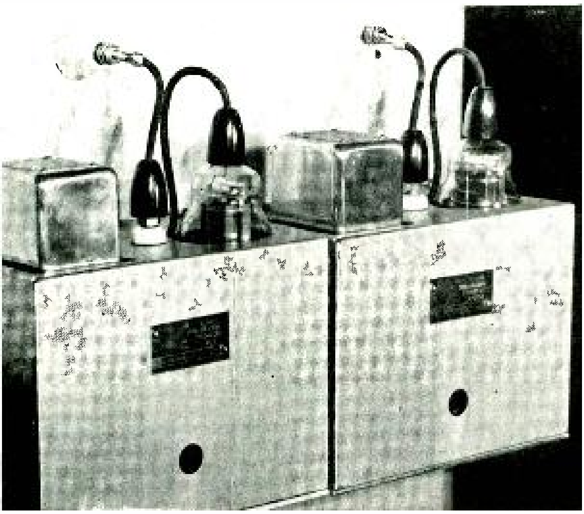

Fig. 13—Crystal oscillators in basic exciter unit.

To take full advantage of the stability of the crystal and holder, it is desirable to use a crystal-oscillator circuit that is inherently stable. The crystal oscillator shown in Figure 13 was developed for this purpose. From the schematic diagram shown in Figure 14 will be noted that there is no tuned circuit associated with the crystal and that it is effectively "electron-coupled" to the output. A small capacitor is provided in shunt with the crystal to adjust it exactly to the desired fre-

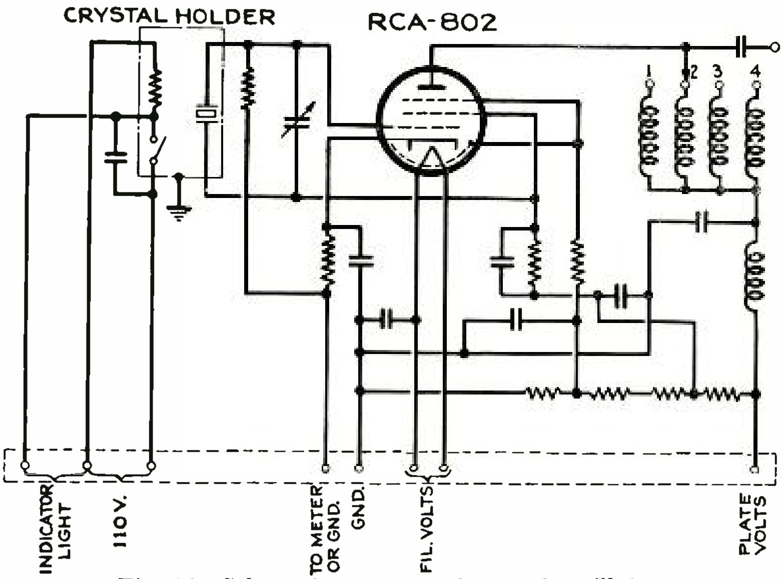

Fig. 14—Schematic diagram of crystal oscillator.

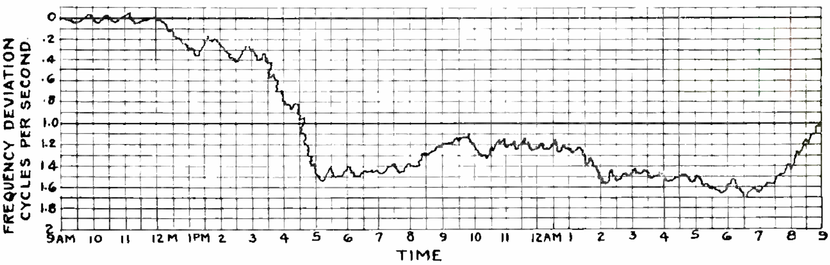

Fig. 15—Frequency stability of broadcast crystal oscillator.

quency. Figure 15 shows the frequency deviation during a typical 24-hour period for a 1000-kc crystal.

Considerable progress has been made in antenna systems and coupling apparatus. A better understanding of the radiation characteristics of lines and the commercial availability of coaxial lines of various sizes, has resulted in more general application of grounded lines, both of the open-wire and coaxial types. The schematic of the 5-kw transmitter, Figure 4, shows a circuit, which with proper circuit elements, will feed a grounded line and which is also a low-pass

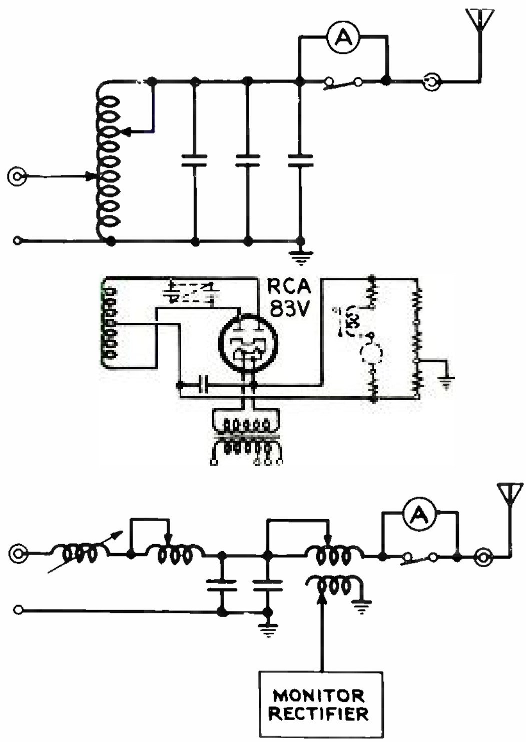

Fig. 16—Typical line-terminating and antenna-tuning circuit.

filter to attenuate the r-f harmonics. The use of grounded lines has simplified the line-terminating and antenna-tuning circuit to a simple impedance-matching network. Figure 16 shows the schematic circuits for the typical line-terminating and antenna-tuning units. The T-network may also be used to obtain the desired phase rotation for a directional array. The monitor rectifier introduces very little distortion and provides current to operate a remote antenna-current indicator and also an electrical interlock which removes plate voltage in case of a flash-over in the antenna system, and thus prevents the possibility of damage to the equipment due to a sustained arc.

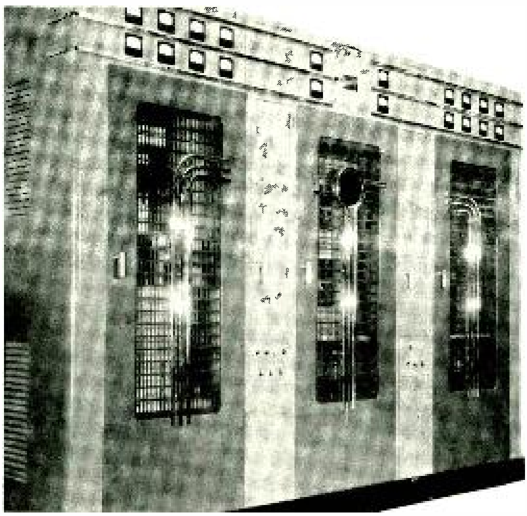

Fig. 17—10-kw high-frequency transmitter.

HIGH-FREQUENCY BROADCAST AND COMMUNICATIONS

For international broadcasting, certain frequencies have been made available by the Federal Communications Commission in the high-frequency band between 6 Mc and 26 Mc. For this service the minimum power permitted is 5 kw. A transmitter developed specifically for this service is shown in Figure 17 and Figure 18. This transmitter has provision for rapidly switching between two frequencies in the band from 6 Mc to 21.6 Mc. Normally one of the lower frequencies between 6 Mc and 10 Mc is used for night-time transmission and one of the higher frequencies between 10 Mc and 21.6 Mc for day-time. In order to reduce the program time loss, electrically operated solenoids operate to transfer the tubes in the exciter from one set of pretuned radio-frequency circuits to another, and to select the proper crystal. In the power amplifier, which employs water-cooled tubes, solenoids operate switches to select separate pre-tuned grid circuits and the proper tank circuit capacity and taps on a single tank coil. Four quartz crystals are employed, one active and one spare for each frequency. These crystals operate at frequencies between 3 and 5.4 Mc. Two Type 207 water-cooled tubes are employed in the power amplifier. These tubes are plate modulated by two Type 891 water-cooled tubes operating as Class B modulators. The transmitter is capable of delivering 10 kw at frequencies up to 10 Mc and the power output

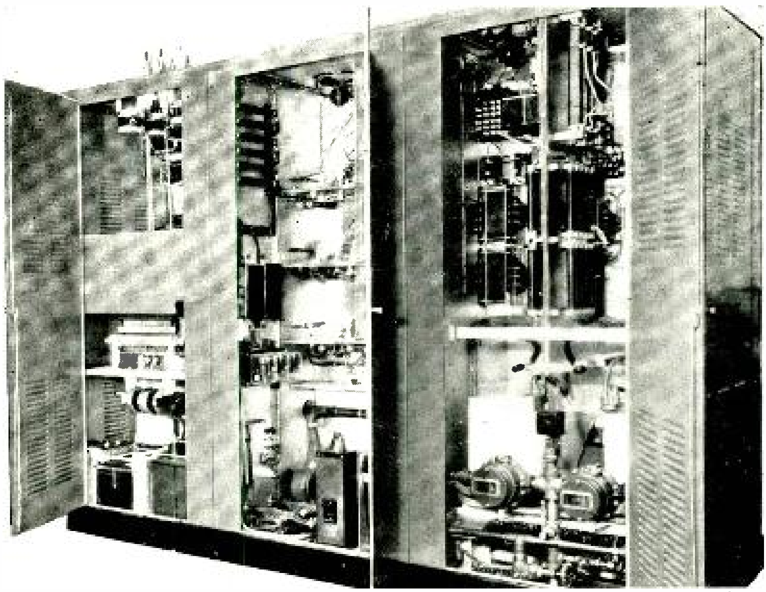

Fig. 18—10-kw high-frequency transmitter.

decreases to 6 kw at 21.6 Mc. The efficiency of the output stage from plate input to transmission-line output varies from 70 per cent at the lower frequencies to 60 per cent at the highest frequency.

In the mechanical construction of this transmitter, many departures have been made from previous practice in high-frequency transmitter construction. Welded-sheet steel has been used throughout for the cabinet and supporting shelves. All steel parts are heavily copper plated. Since the skin effect at high radio frequencies results in small penetration, the plating serves to reduce shielding losses to a low value. This use of steel has permitted the construction of a cabinet of pleasing appearance which is extremely rugged and yet not of complicated mechanical construction. In addition to the transmitter unit, the complete equipment includes a closed-radiator-type cooling unit, voltage regulator to correct for line-voltage variations in the order

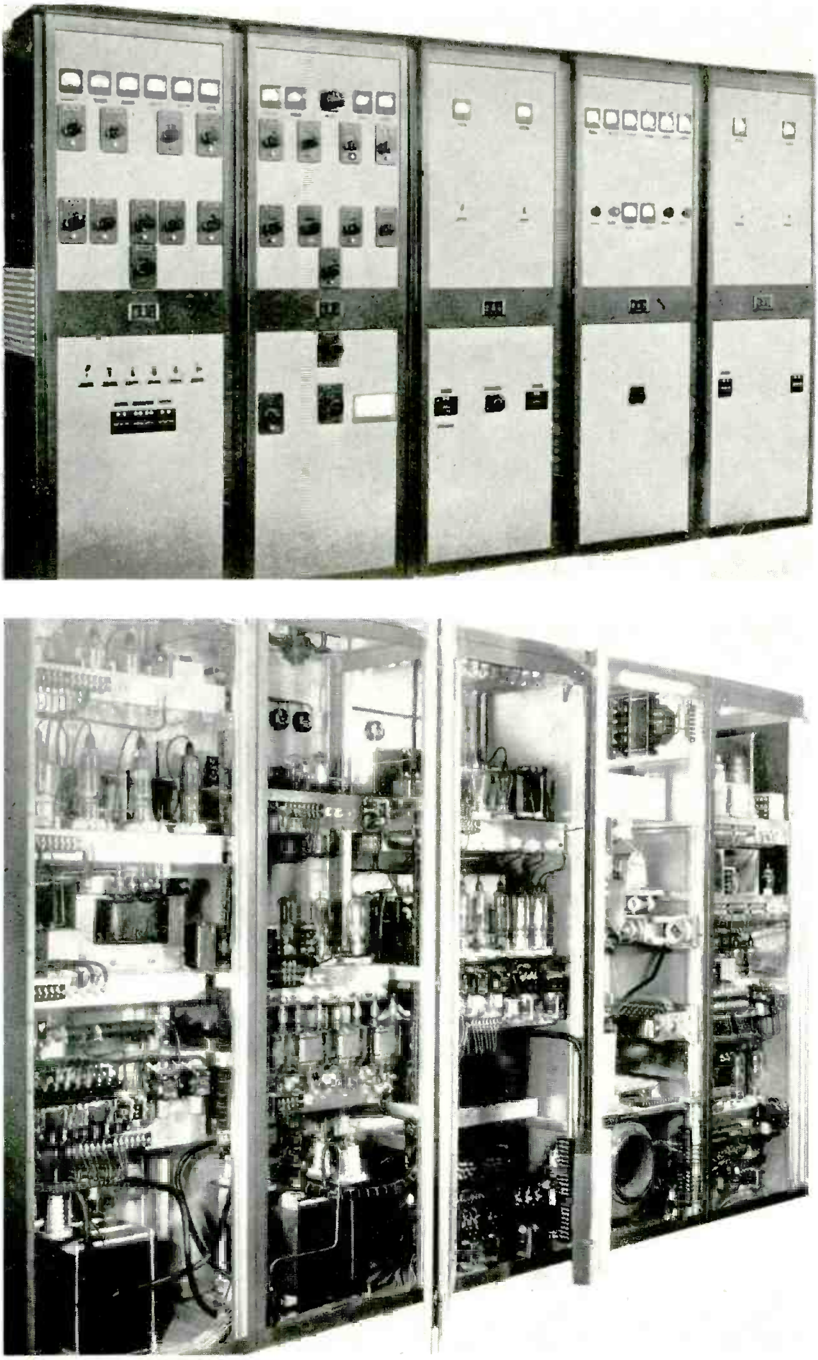

Fig. 19—Medium-power, high-frequency communication transmitter.

of 10 per cent above or below normal, an oil-cooled modulation transformer, and coupling reactor.

For general high-frequency communication applications for fixed stations, there has been need for a medium-powered transmitter that can be used for either telephone or telegraph service and capable of being quickly adjusted to any frequency in the normal high-frequency band. Such a transmitter is shown in Figure 19. This transmitter is continuously variable over a frequency range from 3 to 20 Mc, adjust-

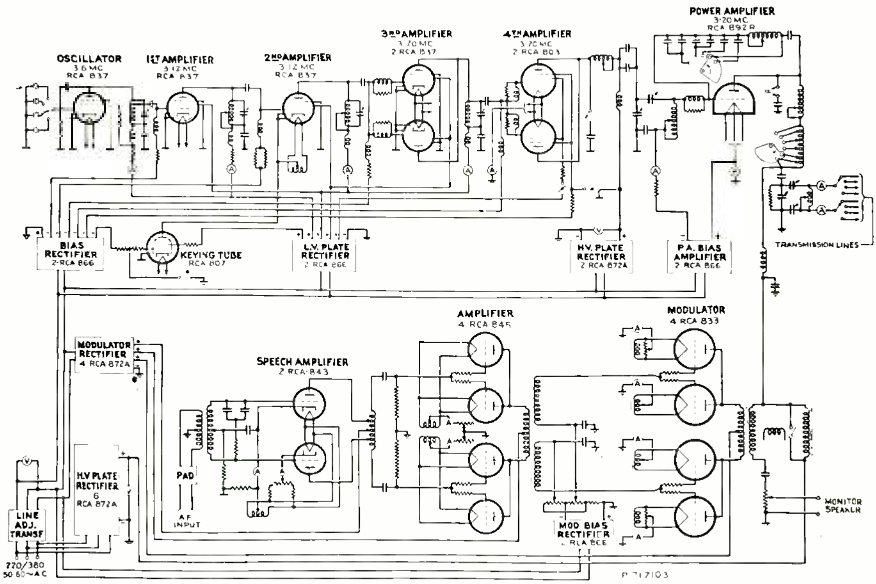

Fig. 20—Simplified schematic diagram of medium-power, high-frequency communication transmitter.

ment being made to any frequency for which a crystal is provided by means of front panel controls. Provision is made for four crystals and four balanced transmission lines to feed separate antennas. The power output is 4 kw on telegraph and 2 kw on telephone at the highest frequency. Air-cooling only is used and the output radio-frequency amplifier uses an RCA-892R tube. The general circuit arrangement of the transmitter is shown in the simplified schematic diagram, Figure 20. The four crystals operating at frequencies between 3 Mc and 6 Mc are selected by a front panel switch. The crystal stage is followed by two successive intermediate amplifier stages, each of which employs RCA-837 tubes. The first stage may serve as an amplifier or frequency multiplier. A third intermediate stage employing two RCA-837 tubes may either amplify or double the frequency. A fourth stage employing two RCA-803 tubes always serves as an amplifier and provides excitation for the RCA-892R tube in the power amplifier. To provide high-speed keying and not be limited by relay speed, a vacuum-tube keyer employing an RCA-807 operates in the plate circuit of the second intermediate amplifier. For telephone operation a separate modulator unit is used which contains two successive stages followed by a Class B modulator stage employing four RCA-833 tubes which feeds audio power through the output transformer across the modulation reactor in the plate circuit of the power amplifier. The overall audio performance when operating on telephone

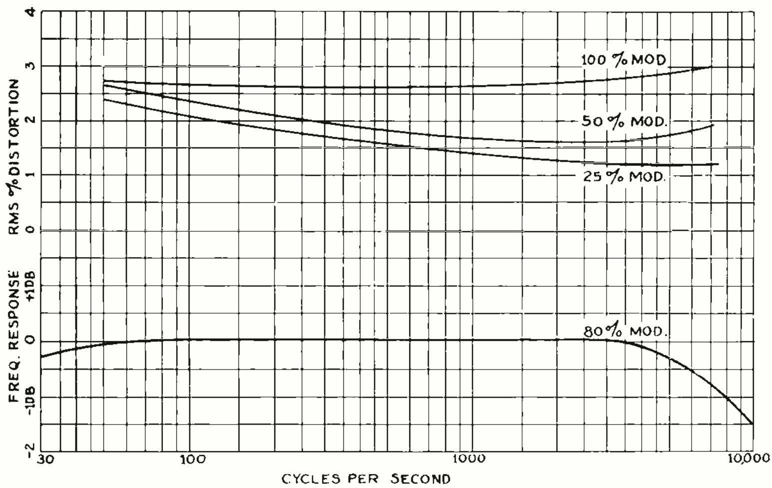

Fig. 21—Over-all audio performance of transmitter shown in Figure 19.

with 2 kw output is shown in Figure 21. For convenience in shipping and installation, the transmitter is divided into five separate units: exciter, power amplifier, high-voltage rectifier, modulator, and low-voltage rectifier. Access to the tubes and circuit components is through doors in the rear of the units. All ventilating openings in the cabinet are carefully covered with fine mesh screen. A glass-wool filter is provided in the air intake to the blower which cools the RCA-892R tube in the power amplifier.

When it is desired to employ this equipment as part of a private two-way telephone system, a separate unit as shown in Figure 22 is provided. This unit contains a transmitting speech inverter, receiving speech inverter, combined with the necessary amplifiers for both channels, and voice-operated relays.

Ultra High Frequency

The transmitter shown in Figure 23 was designed for police and broadcast pick-up service operating in the frequency band from 30

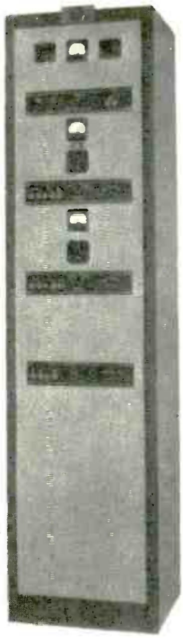

Fig. 22—Privacy equipment.

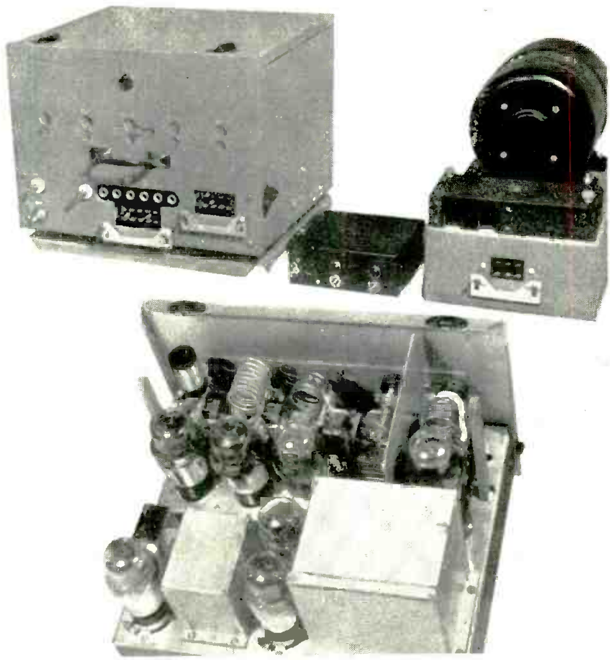

Fig. 23—30 to 42-Mc pick-up transmitter.

to 42 Mc. This transmitter is primarily intended for installation in automobiles and operates from the six-volt car battery. The equipment consists of four main units: the transmitter, the dynamotor, the control unit, and the telephone handset, which are interconnected by cables and plugs. The transmitter employs a V-cut crystal operating at one-

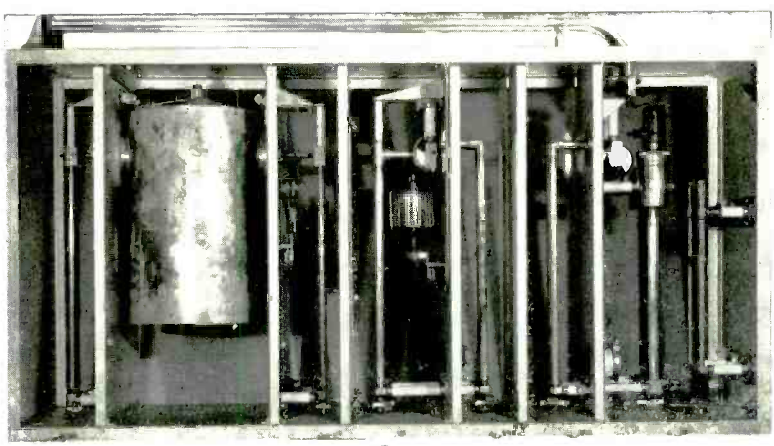

Fig. 24—Experimental 4-kw, 100-Mc equipment.

quarter the output frequency. The crystal oscillator tube is an RCA-1610 which drives a 1610 intermediate-amplifier frequency doubler. A second intermediate-amplifier doubler drives the output amplifier, both stages using the RCA-1608 tubes. The audio system consists of a Type 46 tube driver and two 46 tubes operating as Class B modulators which plate modulate the output r-f amplifier. The frequency characteristic is flat from 70 to 7000 cycles and the rms distortion is less than 7 per cent at 100 per cent modulation at any frequency between 70 and 7000 cycles. The power output is 15 watts and at this output the drain from the 6-volt battery is 35 amperes. When it is desired to operate the transmitter at fixed locations a separate rectifier power unit operating from 110 volts a.c. replaces the dynamotor unit.

It is expected that rapid progress will be made in the field of ultra-high-frequency transmitters, not only for higher power, but for higher frequency. Assignments have now been made for various communications services to operate at frequencies up to 200 Mc. Special tubes and circuits necessary to obtain proper operation at these ultra-high frequencies will lead to many design features which will radically change our conception of transmitter construction. An example of this trend is shown in Figure 24 which shows an experimental set-up which will deliver 4 kw peak power at 100 Mc.

The authors wish to acknowledge use of Figures 9, 10, 11, and 12, and text relating thereto, which were taken from the article entitled "Probable Percentage Modulation at Various Audio Frequencies," by R. Serrell, which appeared in *Broadcast News* in December, 1936, and express their appreciation for permission to use this material.

RADIOTELEPHONE FOR SMALL YACHTS

By

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IN the design of radiotelephone equipment to provide two-way communication with harbor telephone stations one of the important factors to be considered is the available power supply on the average small yacht or cruiser. Radio transmitters having a power output of 15 watts or greater have been used for some time on vessels where a large 12, 32 or 110-volt battery is the primary power source.* On the other hand there are large numbers of privately owned pleasure craft whose only source of electrical energy is the standard 6-volt battery used for starting and lighting. Such craft operate mainly in sounds, bays, or at moderate distances off shore, and accordingly do not require long-distance two-way communication. For such applications, a compact two-way radiotelephone has been developed recently. Of interest to the engineer are various aspects of design and performance which are covered in this paper.

Several frequencies are allocated by the Federal Communications Commission for harbor radiotelephone service. At present for transmission from the vessel to the coastal harbor station, there are 11 channels, spaced 8 kc apart between 2110 and 2206 kc. There is also the intership frequency of 2738 kc to provide intercommunication between small vessels. The United States Coast Guard maintains a watch on the frequency of 2670 kc and this frequency is used for safety purposes, to summon aid in case of distress, or in other emergencies. The coastal harbor stations transmit to the vessels on a frequency between 2500 and 2600 kc. In order to minimize interference, the radio equipped vessel must be able to transmit on a particular frequency that is monitored by the harbor station, and in turn be able to receive the transmitting frequency assigned to the harbor station. A minimum of three frequencies is therefore required on the vessel, one for communication with the harbor station, one for intership operation, and one for the Coast Guard frequency.

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* "Ship to Shore Harbor Telephone Equipment" by H. B. Martin, RCA Review, July, 1938.

It is necessary to provide minimum current drain from the 6-volt battery that is used for power supply on the average small craft. It was therefore decided to use a vibrator unit to provide plate and screen supply for the transmitting and receiving tubes. The drain on the battery when the equipment is standing by to receive a call is approximately 6 amperes. When the transmitter is turned on the load on the battery is approximately 11 amperes. A simplified schematic circuit diagram is shown in Figure 1. A 6C5 oscillator, crystal controlled, is

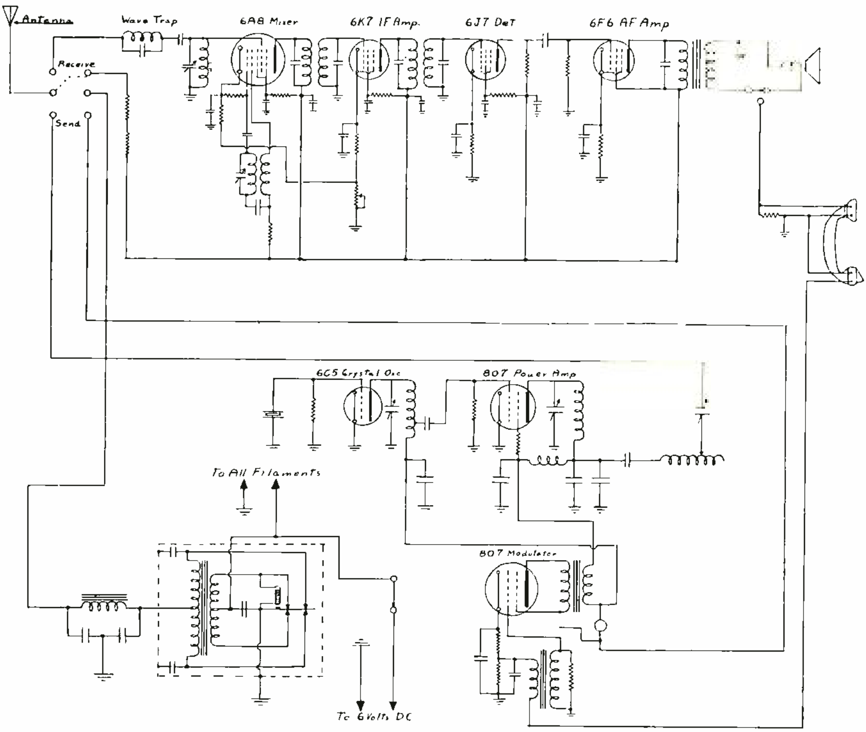

Fig. 1

employed in the transmitter to insure accurate maintenance of frequency. The crystals are clamped in ceramic holders of the type used for aircraft service so that vibration will not affect their operation. The output of the crystal oscillator stage is coupled to the control grid of an 807 beam power-amplifier tube. The beam power-amplifier tube is provided with a tuned tank circuit which is loosely coupled to the antenna circuit. Since the natural frequency of the small antennas that are used is considerably higher than the transmitting frequency, it is necessary to provide inductive loading in the antenna circuit so that the antenna operates as a loaded quarter-wave "Marconi" antenna.

A variable series antenna condenser is also provided to permit the antenna to be accurately resonated to the output frequency. Plate and screen modulation of the power-amplifier tube is used and the audio energy necessary for this purpose is obtained from an 807 tube whose output is transformer-coupled to the power-amplifier tube. The microphone on the handset is coupled to the control grid of the modulator tube through a step-up transformer in the conventional manner.

A compact superheterodyne receiver is used. The receiver is pretuned to three frequencies and consists of a mixer-oscillator, a 465-kc intermediate frequency amplifier, a second detector, and an audio amplifier. The output circuit of the audio amplifier is arranged so that either loudspeaker or headphone reception may be used. Adequate

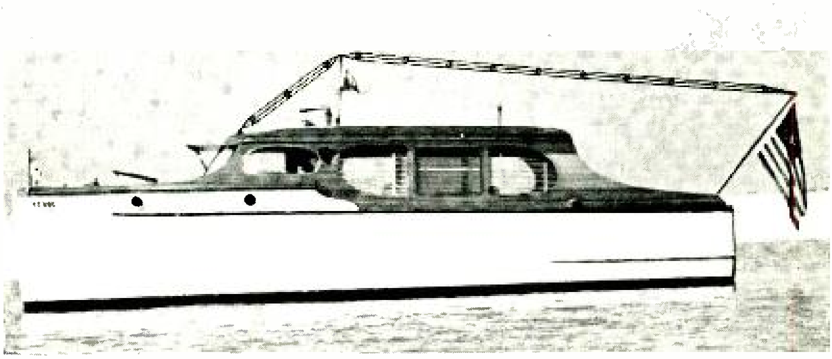

Fig. 2

audio output is obtained from the receiver with an input signal on the order of 20 microvolts.

The effectiveness of a radiotelephone on a small vessel is determined to a considerable extent by the type of antenna which may be erected. Masts of appreciable height are not usually available and although the equipment may be designed to work into antennas of low capacitance, this may result in excessive losses in the loading coil that is used. As an example of an effective antenna, the arrangement shown in Figure 2 may be considered. This vessel, the *Phantom*, owned by Mr. O. B. Hanson, utilizes a small cage structure in conjunction with the radiotelephone set. An antenna of this type has an equivalent capacitance of about 160 $\mu\mu f$. and accordingly does not require a large amount of loading for the 2000 to 3000 kc band. In order to insure an adequate ground, the *Phantom* uses a copper plate fastened to the outside of the hull below the water line. While it is possible to use the propeller shaft and the engine structure as a ground system on small vessels, such practice does not result in effective low-resistance grounds and is subject to the additional disadvantage of introducing noise in the radio receiver when the vessel is under way.

The nominal daytime range that is obtained with a small radiotelephone such as described in this paper is approximately 25 miles under average conditions. Under favorable conditions, in the daytime, when the noise level is low, a range of 55 miles has been obtained with the *Phantom*. Since the power of the coastal-harbor transmitter is many times greater than that of the equipment on small vessels the received signal on the vessel will be greater than the signal received at the harbor station. In other words, the transmitting range of the

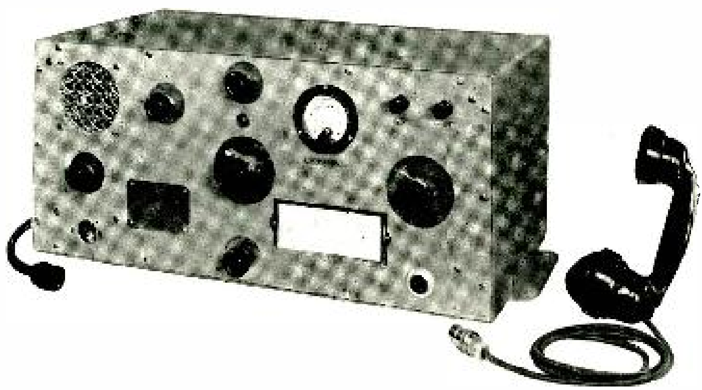

Fig. 3

shipboard equipment is apt to be less than the receiving range. This difficulty may be largely overcome through the use of remote receiving stations on shore, which are connected through landlines to the harbor transmitting station. In the vicinity of Long Island Sound this is accomplished by receiving stations located at Whitestone, Port Jefferson and Bay Shore. The harbor transmitting station is located on Staten Island.

In general the night-time range over moderate distances is less than that obtained in the daytime due to the prevalence of a higher noise level at night. Ranges of several hundred miles are sometimes obtained at night with low-powered transmitters, on account of skywave transmission, although at the same time it might not be possible to communicate locally 25 to 50 miles due to noise or skip distance effect.

A photograph of the small-vessel radiotelephone is shown in Figure 3. A standard telephone handset is plugged into a suitable jack and the 6-volt power supply into a second jack. Connections to the antenna and ground posts complete the installation. The three transmitting and receiving channels are selected simultaneously by means of a switch in the center of the panel. This switch, which also turns the power on the set selects the appropriate pretuned circuits in the transmitter and receiver for the channel which is to be used. The antenna circuit is then resonated by means of the antenna tuning control and a small meter is provided to facilitate this adjustment. Send-receive operation is accomplished by means of a small two-position switch which connects the antenna to the transmitter or the receiver, and at the same time connects the plate supply to the transmitting or receiving tubes. Due to the portable nature of the set it also finds application in small motor boats which are carried on large yachts. In such cases the small radiotelephone may be quickly installed in the motor boat and used to maintain communication with the yacht on the intership frequency. A range of 5 to 10 miles is usually adequate between the motor boat and the yacht, and permits the use of a small vertical antenna on the motor boat. A satisfactory vertical antenna is the plug-in telescope type which may be quickly fastened in place, extended to a suitable height, and later removed when the motor boat is stowed on the davits of the yacht.

Before the two-way radiotelephone may be used, it is necessary for the owner to obtain suitable licenses from the Federal Communications Commission. A "Ship Station" license is required to cover the particular type of equipment that is to be used. A Third-Class Radiotelephone Operator's license is also necessary. Such a license is obtained by passing a suitable examination which covers the general rules and regulations applicable to mobile communication services.

MEASUREMENT OF BROADCAST COVERAGE AND ANTENNA PERFORMANCE

BY

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PART II

Summary.—*The first part of this paper was presented in an earlier issue of the* RCA REVIEW[21] *and covered the general aspects of "Daytime and Night-time" coverage. This second part of the paper covers the determination of the efficiency of the antenna system. It describes in detail the measurements necessary to determine the effectiveness of the antenna in converting the power into radiated energy.*

ANTENNA MEASUREMENTS

THE antenna resistance is that component of the antenna impedance which when multiplied by the current squared gives the input power to the antenna. It is important that the value of antenna resistance and current be accurately known in order to ascertain (a) if full licensed power is being fed into the antenna and (b) if with full power whether normal radiation is being obtained at say, one mile from the antenna, as expressed in millivolts per meter.

RESISTANCE AND REACTANCE MEASUREMENTS

In general if the antenna reactance and resistance are both below a hundred ohms it is comparatively easy to measure them with an error very little greater than that found in the standards used in the measurement. However, in the measurement of resistance and reactance of higher value, considerable care must be exercised to obtain this order of accuracy. It is not the intent of the above statements to imply that an arbitrary value of a hundred ohms marks a sharply defined limit, but that the making of accurate measurements necessitates more precautions when the higher impedances are measured.

In all measurements of antenna impedance it must be remembered that each unit of length of lead-in represents an additional shunt capacitance.*

[21] W. A. Fitch and W. S. Duttera, "Measurement of Broadcast Coverage and Antenna Performance," Part I, RCA REVIEW, Vol. II, No. 4, April, 1938.

* It also represents added series inductance, but for most measurements this is of little importance.

be used indirectly to measure higher values of resistance. As a result of many measurements by the various methods, the authors have found the bridge and substitution methods to be most reliable and satisfactory. The three methods will now be discussed in detail.

Bridge Method

The radio-frequency bridge with suitably compensated arms operates as an equal-arm capacitance bridge and measures directly capacitances of from 40 $\mu\mu f$ to 1250 $\mu\mu f$ and resistances of from 0.1 ohm to about 100 ohms. For the direct measurement of resistance and inductive reactance it is necessary to use a series condenser of sufficient capacitance to produce a net capacitive reactance that is within the above mentioned values. The bridge requires a shielded signal generator with an output voltage of about five volts which may or may not

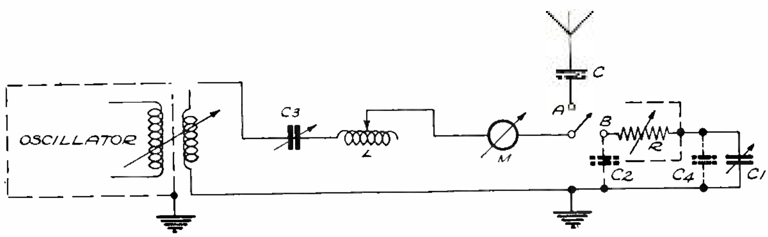

Fig. 13—Measurement of reactance by the substitution method.

be modulated. The signal generator should be free of frequency drift when the impedance of the load is varied. A detector is required to indicate the degree of balance of the arms of the bridge. In order to obtain a satisfactory balance a maximum sensitivity of about 10 microvolts is necessary. The type of detector will of course depend upon whether the signal generator is modulated or unmodulated. If the signal generator is unmodulated, the detector should have a suitable gain control and output meter. A meter is necessary to indicate that the arms of the bridge are approaching a balance. In order to extend the sensitivity of the meter, as a better balance of the bridge is obtained, it is necessary to use a suitable attenuator. If it is modulated the detector may be an ordinary receiver. As the null is approached the sensitivity may be raised by using the regular gain control. The operation for direct measurement of resistance and reactance is similar to low-frequency measuring procedure[23].

[23] Laws "Electrical Measurements," (McGraw Hill).

SUBSTITUTION METHOD

The substitution method is shown schematically in Figure 13. If the antenna has inductive reactance it is necessary to insert a series low-loss condenser, C, of sufficient capacitance so that the net capacitive reactance of the antenna is within the range of C_1. In Figure 13, C_1 is a low-loss standard variable condenser, C_3 is a large variable condenser (if a large frequency range is to be covered) and the loss of this condenser is unimportant, L is a large inductance (if a large frequency range is to be covered) with taps, and the loss of this coil is also unimportant, M is a radio-frequency milliammeter preferably in the order of 100 ma full scale. C_2 is the stray capacity to ground of the connection between the point B and the decade resistance box. C_4 is the stray capacitance (to ground) of the connection between the resistance box and C_1. The capacitances C_2 and C_4 should be kept at a minimum. If a more sensitive meter is used, the meter will be subject to failure from static or current induced in the antenna from other radio-frequency sources and in addition may be so sluggish that much greater time will be required in making the measurements. The time required for a measurement is especially important where the oscillator is fed from a varying a-c supply. R is a decade box, preferably of the constant-reactance type, suitable for radio-frequency measurements.

In order to measure antenna impedance by the substitution method the meter is first connected to the antenna (Point A) and C_3 is varied until a maximum current is obtained in M. If this occurs with maximum or minimum capacitance of C_3 the inductance L is increased or decreased until by tuning C_3 a maximum current is obtained somewhere in the middle range of the condenser. The coupling with the oscillator should be adjusted so that a current, I, is obtained on the meter. This current should be between half- and full-scale reading of the meter. The meter is then connected to the substitution circuit (Point B) and, always tuning C_1 for maximum current, R is varied until the previous value of I is obtained. The meter should then be rapidly switched back to A and then again to B and the same value of current I should always be obtained.

The antenna impedance is then

$$Za = R - jX_{c1} \tag{4}$$

If C has been used, the antenna impedance is,

$$Za = R - j(X_{c1} - X_c) \tag{5}$$

The oscillator should preferably have an output in the order of 10 watts. If the oscillator has greater output it becomes more difficult to get sufficient shielding while much less output will result in poor regulation. It should be well shielded, have low-frequency drift, and should have low-harmonic content. The connection between the meter and the antenna should have exactly the same capacity to ground as the connection between the meter and the substitution circuit. By keeping the connection from B to R and R to C_1 short, as well as A to C and C to the antenna, this effect is minimized. The accuracy of the resistance measurement, except for the accuracy of the standard decade box R, depends largely on keeping the stray capacitance C_2 low, if large values of resistance or reactance are to be measured. The accuracy of the reactance measurement, except for the accuracy of the standard condenser C_1, depends largely on keeping C_4 and C_2 small in comparison with C_1. The overall accuracy of the measurement then

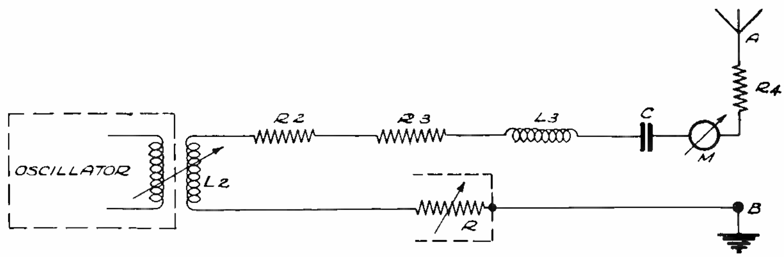

Fig. 14—Measurement of resistance by the resistance-variation method.

is dependent upon (a) the accuracy of the standards, (b) the shielding and stability of the oscillator, (c) the minimization of capacitances C_2 and C_4, and (d) the personal element in making the measurement. If we neglect the personal element, it should be possible to measure resistances and reactances within 1 to 2 per cent or plus or minus 0.2 ohm whichever is the larger. This is the direct measurement of resistance and reactance.

Resistance Variation

It will be noted that the above methods measure both resistance and reactance. The resistance-variation method is perhaps best adapted for the measurement of resistance only. This method is shown schematically in Figure 14. In this measurement the coupling coil resistance plus the resistance of L_3 plus the resistance of the meter (M) must be known or must be negligible with respect to the antenna resistance. The condenser C should have low loss. The measurement is made by first tuning the antenna to resonance with L_3 and C, as indicated by a maximum current in M. This should be nearly full-scale reading of the meter (say 80 ma on a 100-ma r-f meter) the resistance R meanwhile is zero. The resistance is then increased until at a value R_1 a current of I_1 is obtained. The resistance of the antenna is then,

$$Ra = \left[\frac{I_1}{I - I_1} \right] R_1 - R_2 - R_3 - R_4 \tag{6}$$

Where R_2 = Resistance of the coupling coil (L_2)

R_3 = Resistance of the loading coil and condenser (L_3 and C)

R_4 = Resistance of the meter (M)

This method of resistance measurement depends for its accuracy upon: (a) A radio-frequency source having perfect regulation and thus having very much greater power capabilities than the power actually used in the measurement; (b) the accurate knowledge of resistance of L_2, L_3, C, and M; (c) a meter having good relative accuracy and having a scale sufficiently large that it can be read with good accuracy; (d) and of course upon the accuracy of the decade box R_1. Because all these requirements are necessary to obtain an accurate resistance measurement, this system of measurement is most conveniently used on an existing installation to obtain only the approximate value of antenna resistance where the transmitter may supply the necessary radio-frequency voltage and the only equipment needed is an r-f milliammeter and a decade-resistance box. Where measurements of resistance and reactance are to be made accurately over a band of frequencies, this method is undesirable. Reactance may be measured by connecting points A and B (when C is a calibrated variable condenser). When C is tuned for maximum current, a reading of C_1 is obtained. When point A is connected to the antenna and C is again tuned for maximum current, a value C_2 is obtained. The antenna reactance is then,

$$Xa = X_{c1} - X_{c2} \tag{7}$$

This method of measurement is not desirable as it is often open to unavoidable errors.

Measurement of High Antenna Impedances

All of the measurements described above were made by the so-called direct method. The maximum value of antenna impedance that should be so measured is in the order of 100 ohms. If that or higher antenna-

This capacitance may have an important influence on the measured resistance if the antenna has either high resistance or high reactance, or both. Consequently it is of prime importance, in determining antenna power, that if the resistance is measured at say Point 3, Figure 11, the current should be measured at that same point. Otherwise there may be appreciable difference between the resistance at the resistance measuring point and the resistance at the point where the current is measured.[22] By measuring the resistance at Point 3, it is not meant that a four-foot, or three-foot or even 6-inch connection is made between the terminals of the measuring equipment and Point 3, but that Point 3 is at the terminals of the measuring equipment or that by the use of Equations (8) and (9) a suitable correction is made for the shunt capacitance. This, of course, necessitates measuring the

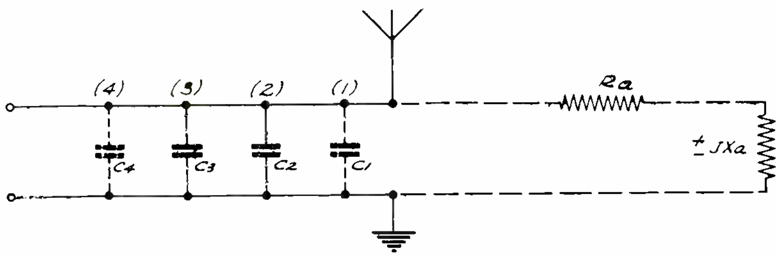

Fig. 11—Schematic of an antenna and its feeding conductor.

capacitance of the connection. Quite often it is inconvenient to put the measuring equipment very close to the point at which measurements are desired. In this case the shunt capacitance and series inductance should be measured and proper corrections made. This procedure can hardly be overstressed in high impedance antennas and the magnitude of the new resistance value may be expressed by

$$Rn = Ra \left[\frac{Xm^2}{Ra^2 + (Xm - Xa)^2} \right] \tag{3}$$

where $Xm = \dfrac{1}{2\pi f\ (C1 + C_2 + C_3 + C4 + Cn)}$ when Rn is the resistance at the point n.

This means that all stray capacitances and long connections to the antenna circuit have to be taken into consideration.

[22] E. A. Laport, "Some Notes on the Influence of Stray Capacitance Upon the Accuracy of Antenna Resistance Measurements," *Proc. I.R.E.*, May, 1934.

Figure 12 is a representative diagram of the line-termination equipment found in many stations using an open-wire balanced line. The capacitances of the antenna lead-in are shown as C_s and Point (A) is the point at which the antenna resistance and current is measured and may be just inside the tuning house.

It is necessary to have a meter in the transmitter room which reads proportional to antenna current or voltage. This permits the operator to tell at a glance whether the antenna output is normal. Due to high voltages at Point (A) it is difficult to insert a current monitoring device which will operate a meter at the transmitter. The antenna current meter located in the transmitter building is generally fed from a thermocouple and in high-power installations this is through a current transformer. The current transformer or thermocouple is usually placed at Point (B), just adjacent to the ground. This is due to the fact that it is very difficult to insulate a current transformer for all

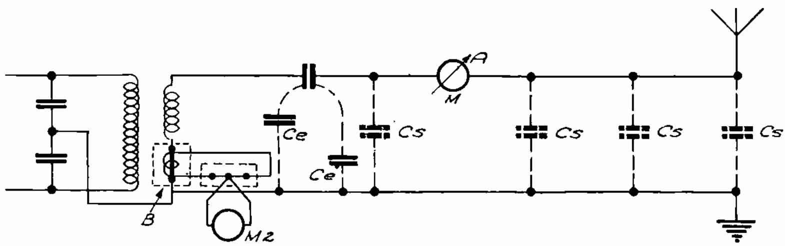

Fig. 12—Balanced transmission—line transmission.

voltages that are likely to be induced in the antenna. In this case it is very important that the resistance measured at Point (A) be not used with the current at Point (B) to obtain antenna power. If the meter M_2 is to be used alone the resistance should be measured at the Point (B). It is sometimes simpler to measure the resistance at Point (A) and insert an accurate meter there to determine the output power; then, from this power the resistance may be computed at Point (B) from the indicated current at the Point (B). For all practical purposes the accuracy of output power measurement is then the same as the accuracy of resistance and current measurement at Point (A) provided Meter *B* has good relative accuracy.

There are three methods of resistance measurements which may conveniently be used. They are (a) the radio-frequency bridge, (b) the substitution method, and (c) the resistance-variation method. Each of these methods is satisfactory if the proper precautions are observed when each is used. These methods are applicable to the direct measurement of resistances up to about 100 ohms. They may impedance measurements are to be made it is necessary to shunt the antenna with a known capacity reactance, *Xm*, having a good power factor. The value of resistance and reactance then measured will not be those of the antenna. The actual antenna resistance and reactance may be computed from the following.

$$Ra = \frac{R\,Xm^2}{(Xm - X)^2 + R^2} \tag{8}$$

and

$$Xa = \frac{Xm\,[X(Xm - X) - R^2]}{(Xm - X)^2 + R^2} \tag{9}$$

where

Xm = Reactance of the shunting capacitance.

X = Measured reactance of the antenna when shunted by Xm.

R = Measured resistance of the antenna when shunted by Xm.

Xm should be of such a value that when shunted across the antenna terminals it gives a resistance of less than 100 ohms for R.

In all measurements of impedance it is necessary to have an accurate knowledge of the frequency. For the direct and indirect measurement of impedances the frequency should be known to within approximately plus or minus 0.1 per cent. This frequency accuracy is necessary to obtain the absolute accuracies mentioned later in this paper. This accuracy of frequency may be obtained by calibrating the r-f signal generator against broadcast stations during the late evening, and plotting dial settings against frequency. The r-f signal generator drift until the time of making the measurement may then be corrected by zero beating, with only one or two of the previous calibration frequencies, and replotting the curve from these points and the slope previously determined. A piezo-electric calibration used in conjunction with an oscillator having a finely subdivided scale may be substituted for the above procedure.

These measurements of resistance and reactance are of importance in obtaining antenna power and in designing the matching circuit between the transmission line and antenna. It should be noted that these measured values are influenced by, (a) the height and configuration of the antenna, (b) the shunting capacitance at the base due to the capacitance of the base insulators and the capacitance of the "lead-in", (c) losses in the antenna and ground and (d) absorption or re-radiation by conducting structures in the vicinity of the antenna. Thus, it is seen, that in the measured values of antenna impedance, there may be a reflection of many factors. Some of these factors are non-existent in some cases and in other cases a suitable correction factor may be applied.

The shunting capacitance mentioned above in (b) is of importance only if the measured values of resistance and reactance are to be compared with the theoretical values or if the measurements are to be used as a basis of determining the probable impedance of another antenna with a different mode of operation. If the shunting capacitance is known, the results may be corrected by Equations (8) and (9).

The losses in the antenna and ground, mentioned in (c) above, cannot be exactly computed by assuming the loss resistance to be the difference between the measured and calculated resistance, because the calculated resistance cannot generally be determined with sufficient accuracy. This is the case even if all the possible irregularities mentioned above, have a negligible effect on the resistance. This loss resistance may best be determined by the measurement of field intensity as covered in a later part of this paper.

If, as mentioned in (d) above, there are any near-by conducting structures, in which there is an appreciable induced current flow, the impedance of the antenna will be influenced. This impedance change will be proportional to the mutual impedance between the other structure and the antenna, as well as the tuning of the other structure. This will result in some loss of energy as well as distortion of the field of the antenna.

Preparation of Measurement Data

If it is proposed to obtain the impedance of an antenna at only one frequency, it is desirable to make a series of measurements in three to five steps of about 5 kc on each side of the desired frequency and to plot these measurements as shown in Figure 15. It may be stressed here that, any one unfamiliar with antenna measurements, should repeat the measurements several times and if possible should check the results by another system of measurement. These measurements may be considerably in error in absolute value, but when plotted as in Figure 15 may give a smooth and appropriate appearing curve. Thus by plotting the values as in Figure 15, only the differences in the handling of the equipment between one observation and another become apparent while the system or method may be faulty without becoming evident in the curve. The reactance at 660 kc is inductive and of a value of 330 ohms. If a shunt capacitance of 12 $\mu\mu f$ were present, the resistance would be as shown by the dashed line (B) in Figure 15, and if not considered would represent an error of 3.5 per cent. A connection about 4 feet long, if not considered, would cause such an error.

For a matter of record this curve and the data sheet should contain such pertinent information as (1) date, (2) observer, (3) station, (4) a brief description of the antenna with special mention of anything that may be changed such as towers grounded, etc., (5) such

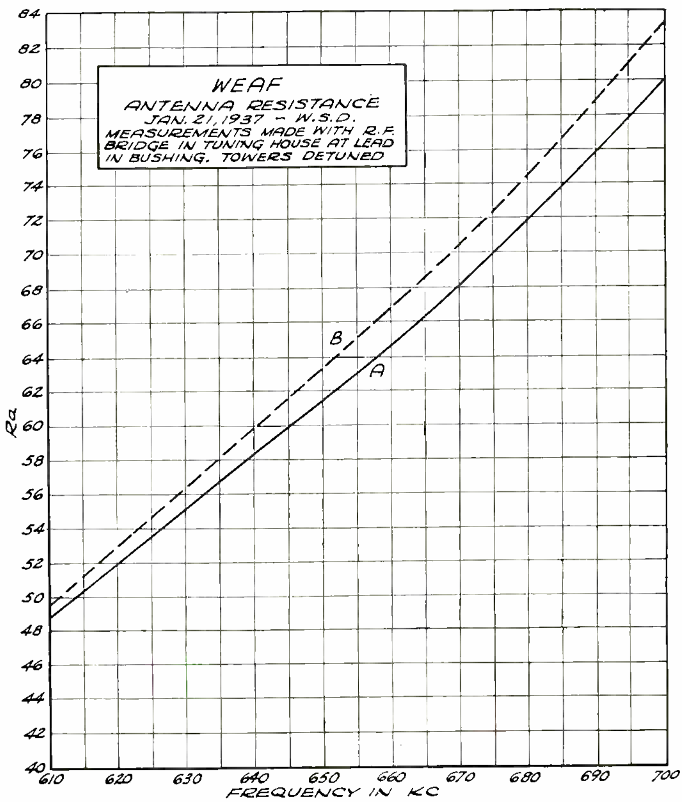

Fig. 15

points as are questionable because of interference or other difficulties should be noted as questionable, (6) point in the antenna circuit at which the measurements were made, and (7) the method of measurement used.

Measurement of Power

By measuring the current and the antenna resistance the power input is obtained from,

$$P = I_a^2 Ra \tag{10}$$

The current should preferably be measured with an internal-thermocouple meter or if with an external thermocouple, the leads from the couple to the meter should be as short as possible so that the added capacitance does not detune the antenna circuit. An internal-thermocouple meter should have the case connected to the power-supply side of the meter. This prevents the meter from erroneously reading higher due to capacity currents. An external-thermocouple meter should be provided with shielded leads and similar precautions should be observed. The antenna current should give nearly full-scale indication on the meter. That is, the full-scale reading of the meter should be less than two times the antenna current. The meter should have a sufficiently large scale, so divided and calibrated that an absolute accuracy of better than about two per cent is obtainable. This

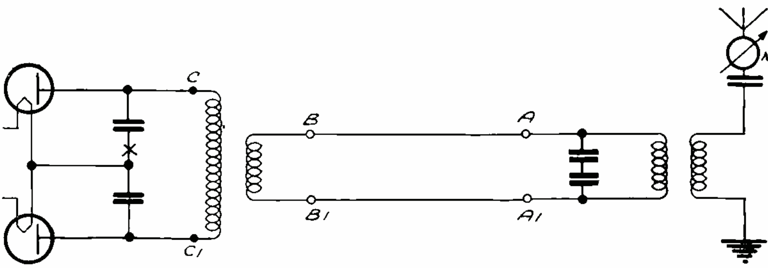

Fig. 16—Typical balanced-line layout.

means a meter having an accuracy of 1 per cent of full-scale reading and the deflection being more than half-scale. The above measurements indicating the power in the antenna may not necessarily be used to indicate the power out of the tubes in the last stage of the transmitter. The antenna-coupling equipment, the transmission line, and the circuits on the plates of the tubes have losses. Under some conditions these various components may account for an appreciable loss. It is generally desirable to measure the output power of the last stage. This serves as a check on the antenna measurements and also on the operation of the transmitter.

The system may be checked in steps. This may be quite difficult to do under some circumstances, especially if a balanced line is used. If a set-up, such as shown in Figure 16, is used, it is hard to measure the input resistance at points AA_1 or BB_1 or CC_1 due to the difficulty of measuring the r-f resistance of a balanced circuit. However fairly accurate measurements may be made by opening the circuit at X *with all transmitter voltages off* and measuring the resistance as outlined above, and then measuring the current at this point. If the power at this point is substantially higher than that measured in the antenna circuit the losses may be found by checking the Q of the coils, condensers and all the connections. Such a check will generally show the source of power loss.

A somewhat simpler measurement of power into the transmission line may be made by measuring the current at points AA_1 and BB_1, when those points are outside of buildings. If the transmission line is perfectly terminated and there is no voltage induced in it from the antenna, these values of current should be the same and when multiplied by the surge impedance give a fairly accurate determination of transmission-line power.

The surge impedance of an open balanced-two-wire line, well above ground, may be computed from

$$Zo = \frac{120}{\cos h \dfrac{s}{2r}} = 276 \log_{10} \left[\frac{s}{2r} - \sqrt{\frac{s^2}{2r} - 1} \right] \qquad (11)$$

where,

$s =$ Spacing between the wires in inches.

$r =$ Radius of the wires in inches.

However, if the points AA_1 or BB_1 or both are inside buildings, or if there are any irregularities in the line, or if there is a substantial induced in-phase voltage from the antenna, this procedure may be subject to serious errors.

The surge impedance of a four-wire line having diagonally opposite wires grounded and the other diagonally opposite wires connected together and fed with respect to the grounded wires is

$$Zo = 69 \left[\log_{10} \left(\frac{1.414\ (h+d)^2}{rd} \right) - \frac{\log_{10}{}^2 \dfrac{2h^2}{d^2} (h+d)}{\log_{10} \dfrac{1.414\ (h+d)^2}{rd}} \right] \qquad (12)$$

where

$h =$ Height of the center of the transmission line above ground in inches.

d = Spacing of the wires in inches, with the wires placed on the corners of a square and is large with respect to r.

r = Radius of the wires in inches.

Since the above line has one side grounded, the resistance and current may be measured at the transmitter as outlined above and the line-input power may be accurately determined.

Of course, the same measurements can be made on a coaxial transmission line. In this case the surge impedance may be closely computed from

$$Zo = 138.4 \log_{10} \frac{R}{r} \tag{13}$$

where,

R = Inner radius of the outer cylinder in inches.

r = Outer radius of the inner cylinder in inches.

The surge impedance found by Equation (13) will be slightly high, depending upon the capacitance introduced by the line insulators.

The surge impedance of a transmission line may be determined from measurements of impedance (as outlined above for antennas) at one end of the line when the other end is shorted and when the other end is open. From these two measurements, the surge impedance is

$$Zo = \sqrt{Zoc\, Zsc} \tag{14}$$

where,

$Zsc = \sqrt{R_{sc}^2 + X_{sc}^2}$ = Impedance of the line when the far end is short-circuited.

$Zoc = \sqrt{R_{oc}^2 + X_{oc}^2}$ = Impedance of the line when the far end is open-circuited.

If a balanced line is used it is relatively difficult to measure these impedances.

If the transmission line is not properly terminated, this method of power determination will give erroneous results. There are two possible ways of determining if the line is properly terminated. The first is to measure the surge impedance of the line, as outlined above and then measure the impedance with the antenna and tuning equipment connected normally. If the impedance thus measured is purely resistive and of the same magnitude as the surge impedance of the line, it may be safely assumed to be terminated properly. The second method is by measuring the voltage across the line or the current in the line at the various points along the line. If these voltages or currents are uniform the line may be assumed to be properly terminated.

By measuring the power at three points as outlined above it is possible to determine just where the losses are in the system, and also to determine if the last stage of the transmitter is delivering the proper power into its load. After locating any source of power loss the remedy is obvious.

Determination of Antenna Efficiency

Antenna efficiency may properly be defined as the ratio of the radiated power to the antenna-input power. It is to be noted that by this definition the criteria of high efficiency is in reducing the losses in the antenna system to a minimum. The antenna performance with regard to the fading free-service area of the station is not involved in this definition. In other words the problem may be expressed as, how much power is actually radiated and what percentage is this of the input power to the antenna? This is the practical problem that the station engineer has to face in deciding, (a) if the station is performing near maximum efficiency, (b) what if anything may be done to reduce existing losses, (c) if when improved, will the improvement provide sufficiently increased service to be justified, and (d) if entirely new equipment or antenna system is justified.

Radiated Power

The radiated power may best be determined by the measurement of field intensity near the antenna. If the soil is a perfect conductor; if the pattern of the antenna is perfectly circular; if all measuring points give absolutely reliable results; and if the distance of the measuring point from the antenna can be very accurately determined; the radiated power can be obtained by the measurement of the field intensity at a single point. This field intensity can then be converted into radiated power from Figure 17 and from the following general relation between input power and measured-signal intensity (E) at one mile,

$$P_1 = P_2 \frac{E_1^2}{E_2^2} \text{ watts} \tag{15}$$

Since $P_2 = 1$ watt in Figure 17

$$P_1 = \frac{E_1^2}{E_2^2} \text{ watts} \tag{16}$$

where

E_1 = Measured field strength at one mile in mv/m.

E_2 = Ideal field strength at one mile in mv/m.

P_1 = Radiated power necessary to produce a signal E_1 at one mile from a no-loss antenna of the same electrical height as the actual antenna.

P_2 = 1 watt.

Let us assume an antenna of constant cross section 400 feet high, operating on 1000 kc with say 5 kw input. The unattenuated signal intensity at one mile is say 430 mv/m. From Equation 2[21] the elec-

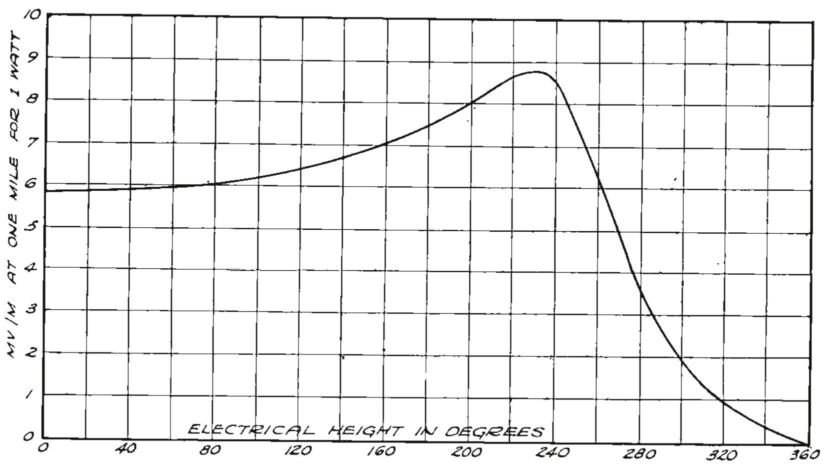

Fig. 17

trical height of this antenna is $\frac{400\ (1000)}{2600}$ or 154 degrees. From Figure 17 the theoretical field intensity from an antenna of that height is 6.9 mv/m. From Equation (16) the power necessary to obtain 430 mv/m from a no-loss antenna of that height is

$$P_1 = \frac{E_1^2}{E_2^2} = \frac{430^2}{6.9^2} = 3{,}880 \text{ watts}$$

The efficiency of the radiating system is then,

$$\frac{3880}{5000} = 77.6 \text{ per cent.}$$

In Part III which will follow in a later issue of the RCA REVIEW the method of determining the one-mile field intensity will be outlined and possible sources of error pointed out. A section will also be devoted to the selection of a transmitter site and a description of the procedure necessary to check the suitability of the proposed site.

CHARTS FOR TRANSMISSION-LINE MEASUREMENTS AND COMPUTATIONS

By

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Summary—Charts have been prepared which, when used with data obtained from simple measurements along a transmission line, enable one to quickly determine the load impedance (resistance and reactance) at the far end, at the transmitter, or, for that matter, at any desired position along the line. Conversely, if the impedance at any position is known the impedance at any other position may be rapidly obtained from these charts. Also when the impedance at a particular location is known a complete knowledge of the voltage and current distributions is easily obtained.*

These charts were developed primarily for use with antenna tests where many frequencies are involved, but an almost unlimited number of problems may be solved with a minimum of labor by their use. The solutions of a few typical problems are shown to illustrate the methods of use.

A brief simplified transmission-line theory is developed. Anyone familiar with this theory should be able to solve rapidly almost any transmission-line problem.

The method of using the chart when a line has substantial attenuation is described.

The mathematical theory of the chart construction is given in an Appendix.

Definitions, Symbols and Relations

$\underset{.}{Z}$ = Impedance in the vector sense

$\underset{.}{Z} = |Z| \underline{/\phi} = |Z|\varepsilon^{j\phi} = |Z| \cos\phi + |Z| \sin\phi = r + jx$, where:

$|Z| = \sqrt{r^2 + x^2}$ = magnitude of impedance

$\phi = \tan^{-1}\dfrac{x}{r}$ = phase angle of the impedance

$r = |Z| \cos\phi$ = resistance component of impedance

$x = |Z| \sin\phi$ = reactive component of impedance

$j = \sqrt{-1}$

Z_0 = surge impedance of line in ohms

$\underset{.}{K}$ = coefficient of reflection in the vector sense

$\underset{.}{K} = |K| \underline{/\psi} = |K|\, \varepsilon^{j\psi}$ where:

* The chart included in this paper is available in a larger size which is more finely divided for engineering use. Copies may be obtained without charge by addressing RCA Review.

$|K|$ = the magnitude of the reflection coefficient which is the ratio of the amplitude of the reflected voltage wave to the main voltage wave

ψ = phase angle of K and is the angle of rotation of the reflected voltage vector with respect to the main voltage vector at the load impedance

E_{max}, E_{min}, I_{max}, I_{min} are the maximum and minimum effective voltages or currents as read by shifting the position of a meter along the line

$Q = I_{max}/I_{min} = E_{max}/E_{min}$

d = distance along line from load to a voltage *maximum* or current *minimum*

λ = wave length

$|K| = (Q-1)/(Q+1)$ By this relation $|K|$ is determined from measurements

$\psi = 2 \times 360° \times d/\lambda$ By this relation ψ is determined from measurements

$\theta = 360° \times S/\lambda$ = line angle corresponding to any distance S

E_m, I_m = main wave voltage, current

E_r, I_r = reflected wave voltage, current

E_s, I_s = total voltage, current at any position

For concentric line:

$Z_o = 60 \log_\epsilon \dfrac{b}{a}$ ohms, where b and a are radius of outside and inside conductors respectively

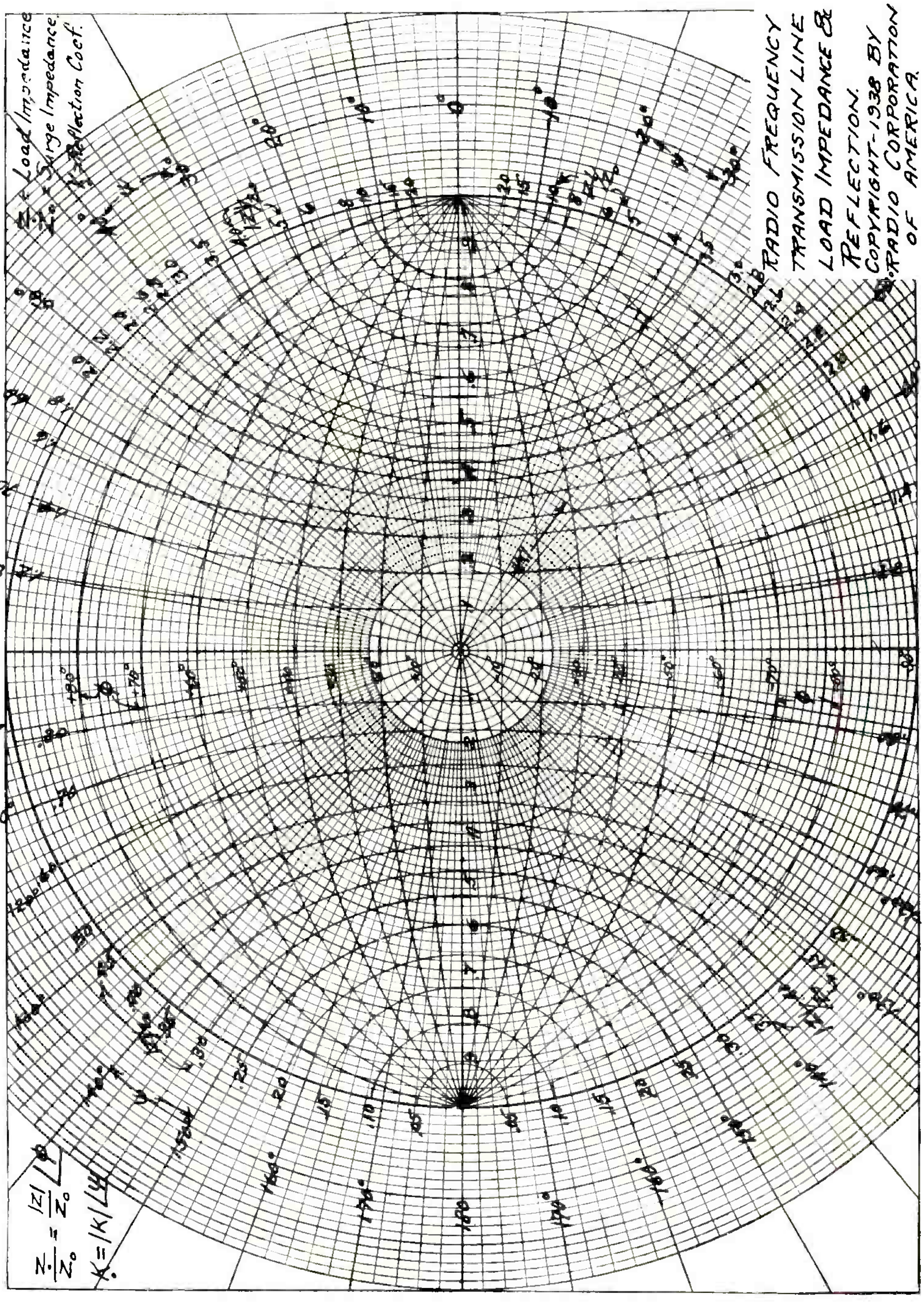
$= 138 \log_{10} \dfrac{b}{a}$ ohms

b and a to be measured on inside of outside conductor and outside of inside conductor respectively

For open wire pair:

$Z_o = 120 \log_\epsilon \dfrac{D}{a}$ approximately, where D is spacing between wire centers and a is the wire radius. Formula is quite accurate so long as spacing is several times wire radius.

$= 276 \log_{10} \dfrac{D}{a}$ ohms


RADIO FREQUENCY TRANSMISSION LINE LOAD IMPEDANCE & REFLECTION.
COPYRIGHT-1938 BY RADIO CORPORATION OF AMERICA.
N = Load Impedance / Surge Impedance
K = Reflection Coef.

Practical Problems

While various transmission-line problems may be solved purely by rule of thumb together with the chart the reader is urged to read the discussion under "Transmission-Line Theory". By so doing he should obtain a vector picture and method of reasoning, a lack of which oftentimes leads to grave mistakes.

The two problems treated in detail below illustrate the pure rule of thumb procedure. Under (3) the method of solving other types of problems is discussed.

1. Given: An antenna loading a transmission line.

Question: What is the impedance presented by the antenna?

Let $\lambda = 10$ meters. Assume that we use a sliding thermocouple galvanometer reading relative current squared (I^2) for measurements along the line. We need the following data:

(1) Ratio $Q = \dfrac{I_{max}}{I_{min}}$ and (2) Distance "d" from load terminals to current minimum. Assume that by sliding the meter we obtain $I^2_{max} = 87$.

$I^2_{min} = 16$ from which we obtain $Q = \sqrt{\dfrac{87}{16}} = 2.23$ and

$$|K| = \frac{Q-1}{Q+1} = \frac{2.23-1}{2.23+1} = 0.551.$$

The I_{min} position is most accurately determined by first locating the two positions, one in each direction from I_{min}, where equal arbitrary meter readings are obtained, the arbitrary reading being roughly the average of I^2_{max} and I^2_{min}. Assume that in the present case we choose $I^2 = 50$ for the arbitrary reading and obtain distances of 7.32 meters and 9.88 meters. The distance d is then $d = (9.88 + 7.32)/2$

$= 8.60$ meters. Hence $\psi = 2 \times 360° \times \dfrac{8.60}{10} = 620°$. We may add or

subtract any multiple of 360°. Subtracting 720°, $\psi = -100°$. Our data now are $\dot{K} = 0.551 \underline{/-100°}$. Now we locate the vector $\dot{K}$ by a pencil point on the polar diagram of the chart, the radius $|K|$ being 0.551 units and the angle $\psi = -100°$. Laid on the polar diagram are impedance coordinates $|Z|/Z_o$ and ϕ. We find that this point corresponds to $|Z|/Z_o = 0.865$ and $\phi = -57.3°$. Thus, the impedance

Z presented by the antenna is $0.865\ Z_o \underline{/-57.3^\circ}$ ohms. If we wish to separate Z into its resistance and reactance components we may do this by projecting the impedance vector upon the horizontal and vertical axes either graphically or by multiplication by the cosine and sine of 57.3°. We then have $Z = 0.467\ Z_o - j\ 0.727\ Z_o$ ohms or a resistance of $0.467\ Z_o$ ohms in series with a capacity reactance of $0.727\ Z_o$ ohms.

2. Given: The same line and antenna as in Problem 1.

Question: What is the load presented to the transmitter by the line when its total length is 125 ft. or 38.1 meters?

The procedure for this case is similar to that in Problem 1 excepting that we use a new phase angle ψ' where $\psi' = \psi - 2\theta$ and θ is the line angle. Since $\lambda = 10$ meters, the length is $\dfrac{38.1}{10} = 3.81$ wave lengths. θ is then $3.81 \times 360^\circ$ and $2\theta = 7.62 \times 360^\circ$. Subtracting the eighth or nearest multiple of 360° we obtain $2\theta = -0.38 \times 360^\circ = -137^\circ$. In Problem (1) we found ψ to be -100°. Hence: $\psi' = \psi - 2\theta = -100^\circ - (-137^\circ) = +37^\circ$. $|K|$ having already been found to be 0.551, we find from the chart the value of impedance corresponding to $|K| = 0.551$ and $\psi = +37^\circ$ and obtain $|Z|/Z_o = 2.60$ and $\phi = +41.2^\circ$. The input impedance Z is therefore:

$$Z = 2.60\ Z_o \underline{/+41.2^\circ} \text{ ohms}$$

3. Other Types of Problems

(a) Suppose we know the input impedance of a line and wish to find the load impedance. This is the reverse of Problem (2). Here we first find $|K|$ and ψ on the chart corresponding to the known impedance. Then we use a new phase angle ψ' where $\psi' = \psi + 2\theta$. In other words we rotate the phase in a leading (counter-clockwise) direction equal to *twice* the line angle in proceeding from transmitter to load.

(b) Assume we have an antenna terminating a long transmission line to be used for television purposes and wish to know the impedance vs. frequency characteristic at the transmitter. Here we have a line of constant physical length, but whose angular length is proportional to frequency. If the line were ten wave lengths long the line angle would be $10 \times 360^\circ$. A one per cent change in frequency would change the angular length 36° and produce a change in ψ' of 72°. To solve the problem we may locate the points of the supposedly known impedance vs. frequency characteristic on the chart, rotate each point through an arc 2θ for the particular frequency, connect the resulting points by a smooth curve and then read off the corresponding input impedances.

(c) Under the conditions discussed under (b) we may wish to know the effect of the addition of various short lengths of line upon the input impedance vs. frequency characteristic. Such effects may be rapidly evaluated by proceeding as under (b), laying over the chart a piece of transparent paper upon which is traced the final curve of (b) and then rotating this tracing clockwise *180°* for each *quarter* wave-length increase in line length or vice versa and reading off the new impedance vs. frequency curve. This procedure is not exact, but is sufficiently accurate for most purposes when the total length of line is ten or more times the length of the added or subtracted lengths considered. It would be strictly accurate if the rotation of the tracing paper were changed for each different frequency of the band to exactly correspond to twice the line angle of the added line section at the particular frequency.

TRANSMISSION-LINE THEORY

(1) Smooth Lines with Negligible Attenuation

The voltage along any transmission line may be considered as due to the superposition of two waves traveling in opposite directions. We shall speak of the wave traveling from the transmitter toward the load as the main wave and that traveling from the load toward the transmitter as the reflected wave. For purposes of analysis these two waves may be represented by vectors rotating in opposite directions and the voltage at any position as the vector sum of these two vectors. The main wave vector lags, i.e., rotates in a negative (clockwise) direction, by 360° for each wave length as we proceed from the transmitter toward the load. The reflected wave vector rotates in a negative direction as we go from load toward transmitter. The current corresponding to the above-mentioned voltage vectors is the vector *difference* of two oppositely rotating similar vectors. The ratio between the main wave voltage and main wave current and also between the reflected wave voltage and the reflected wave current is equal to the surge, or characteristic, impedance of the line. The phase and amplitude relations between the reflected and main wave vectors are determined by the characteristic of the load at the end of the line. If the line is open at the far end the reflected voltage vector is equal and in phase with the main voltage vector at the end resulting in a voltage maximum. If the line is closed at its far end the main and reflected voltage vectors are equal, but in phase opposition at the far end, the resultant voltage being zero. If the load is a pure resistance and its value is higher than the surge impedance of the line the reflected voltage vector is in phase with the main voltage vector, but of a lesser amplitude. If the load is a pure resistance of a value lower than the surge impedance the reflected voltage vector is in phase opposition to the main wave vector, but of a lesser amplitude. When the load is a pure resistance equal to the surge impedance the amplitude of the reflected vector is zero, that is, the wave along the line is a pure traveling wave. The load conditions so far discussed are the only load conditions where the reflected vectors are in phase or in phase opposition to the main wave vectors at the load. When the load

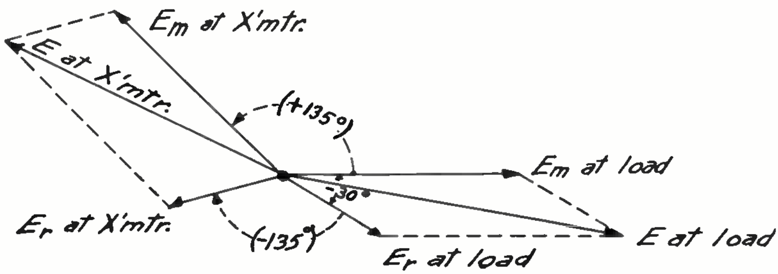


Fig. 1

is an impedance having a reactive component the vector representing the reflected voltage wave is always rotated with respect to the main wave vector at the load.

When both the amplitude and angle of rotation of the reflected voltage vector are known at the load the voltage and current relationship for all points along the line may be easily obtained graphically by drawing a vector diagram. The coefficient of reflection is defined as the vector ratio of the reflected voltage wave to the main voltage wave at the load. In other words the magnitude of the reflection coefficient is equal to the ratio of the numerical values of main and reflected voltage waves and the angle of the coefficient is equal to the angle of rotation of the reflected voltage vector with respect to the main voltage vector at the load.

To show the principles which have been discussed let us draw a vector diagram for a line where the load is such that the reflected voltage wave is rotated by $-30°$ and its magnitude is 50 per cent of that of the main wave voltage. Under these conditions we may write,—

$$\dot{K} = 0.50 \, \underline{/-30°}$$

Referring to Figure 1 we first draw *Em* at an angle of zero degrees and then *Er* with an amplitude of 0.5 at an angle of $-30°$. Let us assume this line to be $3/8$ of a wave length long and that we wish to find the conditions at the generator. We then rotate the vector *Em* in a positive direction by an angle equal to $3/8 \times 360°$ or $135°$. Similarly, we rotate *Er* in a negative direction by $135°$ from its initial position, making the final angle for *Er* equal to $-135° - 30° = -165°$. The vector sum of the two vectors, after these rotations, is then equal to the voltage at the transmitter.

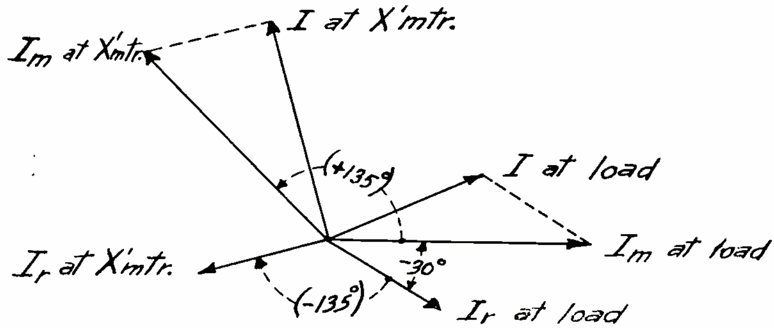


Fig. 2

Figure 2 shows a similar diagram for the current. The procedure is the same as that for the voltage with the exception that the desired current at the transmitter becomes the vector difference between the main and reflected waves rather than vector sum.

These two diagrams can, of course, be combined. The same vectors may be used to represent both current and voltage waves if it is kept in mind that the ratio between voltage and current is equal to the surge impedance of the line and that we must take the vector sum for the voltage condition and vector difference for the current condition. The input impedance at the transmitter is, of course, the input voltage divided by the input current in the vector sense. In other words the magnitude of the impedance is equal to the numerical value of the input voltage vector divided by the magnitude of the input current vector and the phase angle of the impedance is equal to the angle between these two vectors.

Such diagrams as those in Figures 1 and 2 show all voltages and currents for any position along the line in correct phase relationship with respect to each other. If we are interested only in the phase relation between voltage and current at one position along the line these diagrams may be combined and further simplified. Figure 3 shows how this may be accomplished, where instead of rotating each vector we hold the main vector still and rotate the reflected vectors by *double* the angle corresponding to line length. Here we let a single vector represent both the main wave voltage and current. The reflected wave voltage and current vectors are then drawn diametrically opposite. In this way we may immediately obtain both the voltage and current for any position by drawing lines from the origin to the terminii of the reflected wave vectors. The phase angle between voltage and current is then equal to the angle between these two resultant voltage and current vectors.

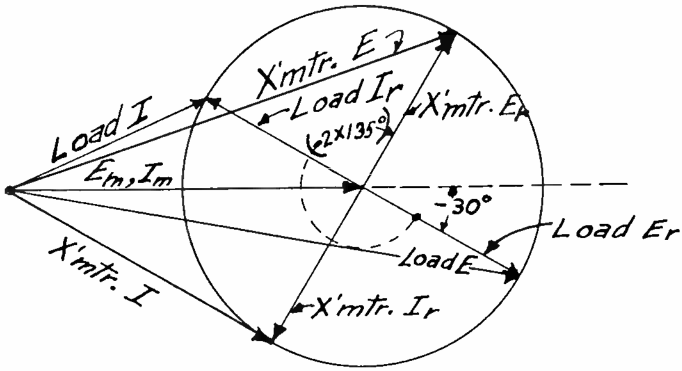


Fig. 3

If desired we may carry the process outlined further and make up a diagram by which we may quickly read off the values of current and voltage for any point along the transmission line once we know the coefficient of reflection. However, such a diagram is not usually of great utility since the impedance is not given directly.

From the discussion given it should be apparent that the conditions of voltage and current along any line are completely and simply described by the coefficient of reflection when it is considered as a vector quantity. Such distributions along any line may also be described by giving the ratio of maximum to minimum voltage or current and the distance from some reference point to the position of voltage maximum or current minimum. Each method of description has its advantages and disadvantages. The chart in this report is based upon the coefficient of reflection. To one constructing such a chart the reasons for using the coefficient of reflection to describe line conditions will be apparent, but these reasons will not be discussed here.

As stated before, the coefficient of reflection in its vector sense is completely determined by the characteristics of the load at the end of the line. The load may be expressed in terms of resistance and reactance or as a vector impedance with a certain phase angle. No matter how we describe the line or how we describe the load impedance we have four quantities depending upon one another. The magnitude alone of the coefficient of reflection depends upon both the magnitude and phase angle of the load impedance. This is also true with respect to the angle of the coefficient alone. The purpose of the chart is to show the variations graphically and enable one to obtain quickly values without computation.

2. *The Chart*

The chart was made up by first setting up a system of polar coordinates representing the complex coefficient of reflection $\underset{\cdot}{K}$, represented by its magnitude $|K|$ as radii and its angle ψ. Then two additional families of curves were drawn over the polar coordinates to represent the complex terminating impedance $\underset{\cdot}{Z}$ which will produce any particular coefficient of reflection. This impedance is represented as regards its magnitude $|Z|$ by one family of curves, extending from zero to infinity and as regards its phase angle ϕ by the other family of curves extending from $-90°$ to $+90°$. All of the curves in both of these families are arcs of circles which materially reduced the labor of plotting them.

It will be seen that this chart enables one to obtain readily the impedance in which a line has been terminated when the reflection set up has been measured and, vice versa, to determine the reflection which will be set up when the terminating impedance is known.

A second valuable feature of this chart is that it enables one to determine readily conditions of reflection $\underset{\cdot}{K}$ and effective impedance $\underset{\cdot}{Z}$ at any point along a line after the reflection coefficient or terminating impedance has been determined at one particular point on the line. This is done by locating a point on the chart corresponding to the known $\underset{\cdot}{K}$ or $\underset{\cdot}{Z}$ and then, holding a constant radius, i.e. a constant $|K|$, following around the chart to the point corresponding to the point in question on the line. A rotation on the chart in a clockwise direction corresponds to moving along the line in a direction from the load toward the generator, and vice versa. A rotation on the chart of 180° corresponds to a distance along the line of a quarter-wave length, i.e. 90° of line angle.

3. *Lines Having Substantial Losses*

In order to apply the theory which has already been developed to lines having appreciable attenuation only minor modifications are necessary. We may still use two rotating vectors to represent the main and reflected voltage wave. However, we must now attenuate the main wave vector and increase the length of the reflected wave vector as we proceed from the transmitter toward the load. The velocity on the line may be somewhat less than that for a wave in free space, but the correction is usually small. Regardless of what the actual velocity may be it introduces no complication so long as we rotate our vectors through an angle corresponding to the wave length along the line rather than the wave length in space. Suppose we have a line where the total attenuation for a traveling wave is 1 *db*. Assuming that we wish to determine the input impedance on this line when we have a load causing reflection, we might increase the main wave vectors of current and voltage at the input end of the line by 1 *db* or to 112 per cent of their lengths at the load and decrease the reflected wave vectors by 1 *db*, or to 89.1 per cent of their lengths at the load. However, insofar as the relationship between voltage and current at one position on the line is concerned we would obtain the same result if we held the main wave vector constant and decreased the length of the reflected wave vector by twice the preceding amount or 2 *db*, or to 79.4 per cent of its original length at the load.

From the above reasoning it should be apparent that in order to use the chart in connection with lines having substantial attenuation it is necessary only to increase or decrease the reflected wave vector to a length corresponding to *twice* the attenuation for the particular length of line. Thus if we wish the input impedance for a certain line having some particular load impedance at its far end we first find the value of $\dot{K}$ corresponding to the load impedance, shorten this vector to a length corresponding to *twice* the attenuation on the line, rotate this vector in a negative or counter-clockwise direction, and read off impedance from the coordinates on the chart in the same manner as for a line having negligible losses. If we wish to determine the impedance of the load when the input impedance is known the process is the same as that previously discussed excepting that we must now increase the reflected wave vector to a length corresponding with twice the attenuation on the line.

Acknowledgment

The writer wishes to acknowledge the painstaking care of Mr. E. D. Thorne in the construction and drawing of the charts.

Appendix

Theory of Chart

The chart represents a mapping of the complex $\dot{Z}$ plane upon the complex $\dot{K}$ plane. Figure 4(a) shows the $\dot{Z}$ plane. On it $\dot{Z}$ lies in the

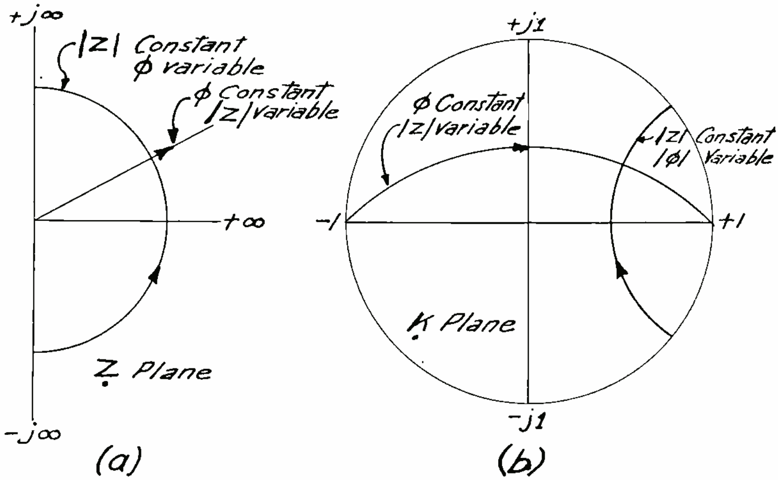


Fig. 4

1st and 4th quadrants only, since we do not consider negative resistance components, and its magnitude $|Z|$ may range in value from zero to infinity. Figure 4(b) illustrates the K plane. Its magnitude $|K|$ ranges from zero to 1 while its phase may have any values from zero to 360 degrees.

If the magnitude of Z is held constant while its phase angle is allowed to vary it traces the half circle shown in 4(a). When this circle is mapped upon the K plane it becomes the circular arc labelled $|Z|$ constant ϕ variable in Figure 4(b). Similarly if we hold the phase ϕ of Z constant and let $|Z|$ vary the straight line on the Z plane Figure 4(a) becomes the circular arc ϕ constant $|Z|$ variable extending from -1 to $+1$ on the K plane of Figure 4(b).

Figure 5 shows the method of construction for the curve $|Z|$ constant ϕ variable. In the diagram $\overline{bo}$ equals the vector 1, $\overline{bP} =$ vector $1 + K$ and $\overline{Pd} =$ vector $1 - K$. Also $\overline{bd} = 2 = 1 + K + 1 - K$. The

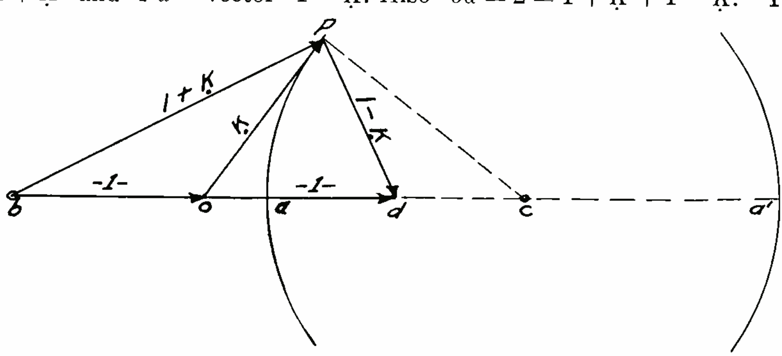


Fig. 5

geometry is as follows: Let b and d be reciprocal points to a circle of radius $\overline{ac}$ with center at c. Then triangles dPc and Pcb are similar

and $|Z| = \dfrac{bP}{Pd} = \dfrac{bc}{ba} = \dfrac{ba}{ad}$ and $\overline{oc} = \dfrac{1}{2}(oa + \overline{oa'}) = \dfrac{1}{2}\left(|K|_o + \dfrac{1}{|K|_o}\right)$

where K_o is the value of $\underset{\cdot}{K}$ for $\phi = o$. The radius of circle $ac = oc$

$- |K|_o = \dfrac{1}{2}\left(\dfrac{1}{|K|_o} - |K|_o\right)$.

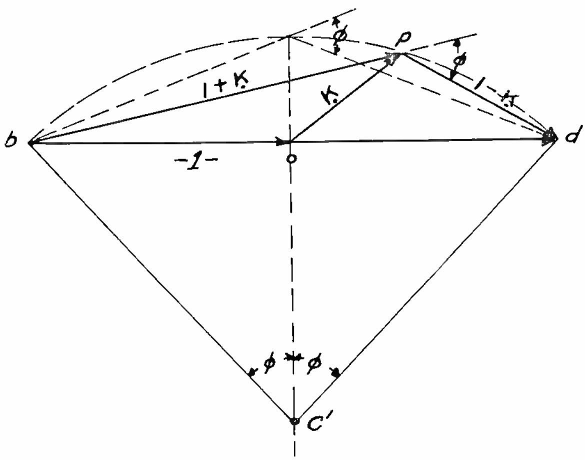


Fig. 6

Figure 6 shows the construction for the curve ϕ constant $|Z|$ variable. It is clear from the geometry that the radius of circle $= \dfrac{1}{\sin \phi}$ and that the distance to the center of the circle $oc' = \text{Radius} \times \cos \phi = \cot \phi$.

The mathematical relations between the load impedance $\underset{\cdot}{Z}_e$ and the coefficient of reflection $\underset{\cdot}{K}$ are:

$$\underset{\cdot}{K} = \frac{\underset{\cdot}{Z}_e - Z_o}{\underset{\cdot}{Z}_e + Z_o} \tag{1}$$

$$\underset{\cdot}{Z}_e = Z_o \frac{1 + \underset{\cdot}{K}}{1 - \underset{\cdot}{K}} \tag{2}$$

To express Z_e directly in terms of $|K|$ and ψ we may write (2) as

$$Z_e = Z_o \frac{1 + |K|\varepsilon^{j\psi}}{1 - |K|\varepsilon^{j\psi}} = Z_o \frac{(1 + |K|\varepsilon^{j\psi})\ (1 - |K|\varepsilon^{-j\psi})}{(1 - |K|\varepsilon^{j\psi})\ (1 - |K|\varepsilon^{-j\psi})}$$

$$= Z_o \frac{1 - |K|^2 + |K|\,(\varepsilon^{j\psi} - \varepsilon^{-j\psi})}{1 + |K|^2 - |K|\,(\varepsilon^{j\psi} + \varepsilon^{-j\psi})} \tag{3}$$

or

$$Z_e = Z_o \frac{1 - |K|^2 + j2|K| \sin \psi}{1 + |K|^2 - 2|K| \cos \psi} \text{ ohms.} \tag{4}$$

If we wish to express Z in terms of $Q = \dfrac{1 + |K|}{1 - |K|}$ and the distance d to a current minimum we obtain after substitution

$$Z_e = Z_o \frac{2Q + j(Q^2 - 1) \sin (720^\circ\ d/\lambda)}{Q^2 + 1 - (Q^2 - 1) \cos (720^\circ\ d/\lambda)} \text{ ohms.} \tag{5}$$

By a similar procedure we may obtain for the input impedance Z_i:

$$Z_i = Z_o \frac{1 - |K|^2 + j2|K| \sin (\psi - 2\theta)}{1 + |K|^2 - 2|K| \cos (\psi - 2\theta)} \tag{6}$$

MEASUREMENT OF EFFECTIVE HEIGHT OF AUTOMOBILE ANTENNAS

By

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Summary—*The requirements of a measurement method for obtaining the effective height of antennas are outlined, with particular reference to the problem of the automobile antenna. An instrument is described which was developed for making these measurements, and its method of calibration and use treated.*

INTRODUCTION

THE importance of the antenna on automobile radio reception has been realized to a greater extent recently than ever before. The use of steel tops in automobile construction has been one factor increasing the interest in antenna construction, since it forced abandonment of the roof antenna and caused new types to be devised. Since the effect of the antenna on satisfactory reception is so much greater than in home radio, some means of comparing automobile antennas quantitatively is much needed.

ANTENNA CHARACTERISTICS

There are three fundamental characteristics of antennas which determine their performance in radio reception. These are:

(a) Effective Height
(b) Reactance
(c) Effective Resistance

Of these, the effective height is most important as it determines the signal the antenna will collect and deliver to the receiver. This article deals primarily with measurement of the effective height.

The reactance of the antenna influences the suitability of the antenna input system of the receiver for use on a given antenna. In automobile antennas the reactance is predominantly capacitive, the inductance being negligible.

The effective antenna resistance affects the maximum signal current which can flow with a given induced voltage. It is of greatest importance when the input system of the receiver has lowest losses.

The intensity of a radiation field is measured in microvolts per meter. If a given antenna in a field of H microvolts per meter has induced in it a voltage of e microvolts, then it has an effective height h', which is equal to e/H meters. The actual physical height of the antenna is different from, and greater than, the effective height, but it is the effective height, as the term implies, which is the measure of the utility of an antenna as a collector of radio waves.

We have seen that an antenna of effective height h', when in a field of intensity H, will have induced in it a voltage e. We can then, in considering antenna effects, replace the antenna by a generator of voltage e. This equivalent generator will have an impedance the same as that of the antenna, which must also be taken into account.

In measuring antenna impedance and effective height we inevitably include the effect of any shunting capacity, such as shielding, etc. This will become apparent from considering Figure 1 and Figure 2.

In Figure 1 e is the antenna voltage
c_1 antenna capacity
c_2 lead-in capacity, etc.
e_2 voltage at receiver

It will be seen that c_1 and c_2 form a capacity voltage divider so that e_2 depends upon the relative magnitudes of c_1 and c_2 as well as upon e. Now by any measurement at e_2, that is, at the receiver, we cannot determine what c_1 and c_2 are individually, but only the total capacity c_1 plus c_2. The situation is then that illustrated by Figure 2. The voltage e' in Figure 2 can be shown to be equal to

$$e \frac{c_1}{c_1 + c_2}. \tag{1}$$

The voltage that we measure will then be e', which is proportional to the effective height of the antenna including capacity effects of lead-in and any other shunt capacities. It can be readily seen how desirable it is to keep the lead-in capacity small, so that as much of the voltage induced in the antenna as possible will be available at the receiver.

It is, in general, easier to design an efficient receiver input system for a high-capacity antenna than for a low-capacity antenna, so that if two antennas have the same effective height but different capacities. the one having the larger capacity will be preferable.

Measurement Methods

We have seen that if we subject an antenna, whose effective height we desire to determine, to a field of known intensity, the resultant voltage is a measure of the effective height. The method then consists in determining the field intensity and resultant voltage. We have available, in almost every locality, broadcast stations of sufficient strength and frequency range which can be used as the radiation field source. It will then only be necessary to measure the field intensity of such broadcast stations.

The usual field-intensity meter consists of a loop antenna of known characteristics, and a receiver calibrated to measure the voltage at the loop terminals. The loop antenna when carefully designed, calibrated,

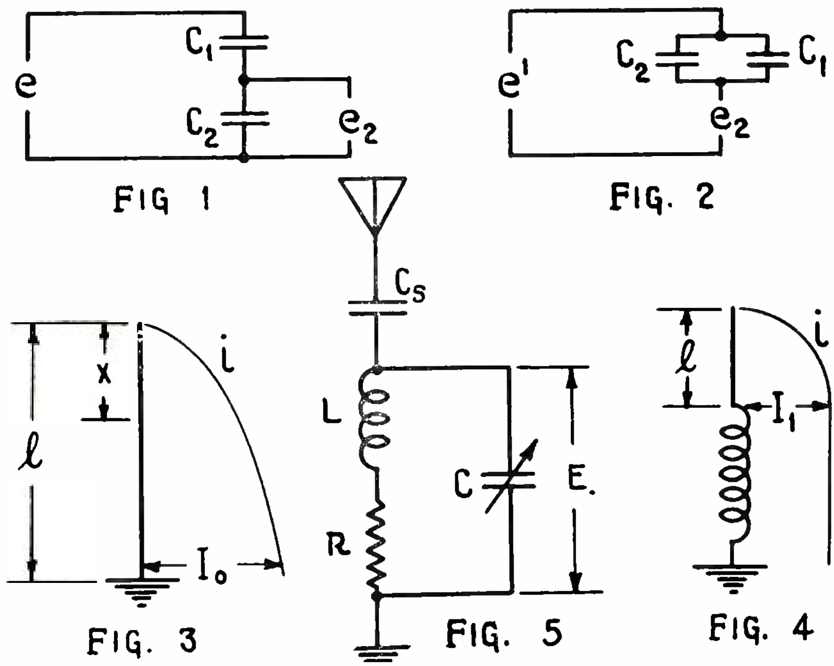


FIG 1 FIG. 2 FIG. 3 FIG. 5 FIG. 4

and used is the most accurate means of measuring field intensity. However the difficulties of constructing and calibrating a loop are such that the use of a rod antenna in place of a loop recommends itself. The rod has the advantages of easier construction and calibration, as well as greater convenience and rapidity of use.

The voltage developed in the rod antenna is measured by means of a radio receiver equipped with an indicating meter. This same receiver can be used to measure the voltage developed in the automobile antenna, so that a comparison is made between the known rod antenna and the car antenna. The absolute sensitivity or voltage calibration of the receiver need not be known, so long as the output meter indication is proportional to input voltage, and the sensitivity of the receiver need not stay constant except during the short period required to compare the output indication of the car antenna with the rod.

It is necessary, for reliable measurements, that they be made in a uniform field so that small changes in position of the car or measuring equipment do not cause differences in readings. Therefore it was felt that the measuring equipment should be entirely battery operated and self contained, so that it could be operated in the open country. Furthermore, inclusion of the batteries within the case of the instrument eliminates variable factors due to external wiring. It is possible to use the equipment without calibration of the rod, in which case it gives comparative results only. It is believed desirable, however, to calibrate the rod antenna so that the effective height of the antenna under measurement can be expressed in meters and results compared between different observers and measurements made at different times.

The effective height of a rod antenna may be calculated from measurement of its physical length. Maximum antenna current will flow, and consequently maximum output meter deflection will occur when the antenna is electrically one-quarter wave length long. Tuning to maximum meter deflection means then that the electrical length of the rod is made one-quarter of the wave length being received, regardless of the actual physical rod length.

A grounded, vertical antenna self-resonant to one-quarter wave length will appear as in Figure 3. The current, i, will have a maximum value, I_0, at the base and vary sinusoidally throughout the length. If the antenna is less than one-quarter wave length long, but is tuned to the quarter wave length condition by inserting inductance, L, at the base, the current across the coil is constant, and the current along the antenna varies sinusoidally as shown in Figure 4.

Let

i be current at any point
l be length of antenna
λ be wave length
x distance from top of antenna
I_0 current at base of quarter-wave long antenna
I_1 current at base of antenna less than quarter-wave long

Then

$$i = I_0 \sin x$$

If h' is effective height

$$h' I_1 \doteq \int_0^l i\,dl$$

$$I_1 = I_0 \sin \frac{2\pi l}{\lambda}$$

$$\text{then } h'I_1 = \int_0^l I_0 \sin \frac{2\pi l}{\lambda}\, dl$$

$$h'I_1 = I_0 \frac{\lambda}{2\pi} \left[1 - \cos \frac{2\pi l}{\lambda} \right]$$

$$h'I_0 \sin \frac{2\pi l}{\lambda} = I_0 \frac{\lambda}{2\pi} \left[1 - \cos \frac{2\pi l}{\lambda} \right]$$

$$h' = \frac{\lambda}{2\pi} \left[\frac{1 - \cos \frac{2\pi l}{\lambda}}{\sin \frac{2\pi l}{\lambda}} \right]$$

$$\text{but } \frac{1 - \cos a}{\sin a} = \tan \frac{a}{2}$$

$$\text{so } \quad h' = \frac{\lambda}{2\pi} \tan \frac{\pi l}{\lambda} \tag{2}$$

$$\text{if } \quad l < \frac{\lambda}{12} \qquad \tan \frac{\pi l}{\lambda} = \frac{\pi l}{\lambda} \text{ (nearly)}$$

$$\text{then } h' = \frac{l}{2} \tag{3}$$

to an accuracy of better than 2%. We thus see that the effective height of a rod antenna can be calculated from its physical dimensions.

The voltage developed in any antenna will not be measured directly since a selective circuit is necessary in order to separate a desired broadcast station from others also present. This means that it is the voltage across the input circuit of the instrument which is read and this is a function of the current produced by the antenna. The reactance of the antenna is tuned out (*i.e.* the input circuit as a whole becomes non-reactive) at any given measurement frequency, but it is also important that the effective resistance of the antenna be small compared to the effective resistance of the input circuit so that the resistance of the different antennas measured can be neglected and instrument readings be considered proportional to effective height.

Use of the input circuit of Figure 5, where the coupling capacity C_s is small (in the case of the instrument to be described approximately 3 $\mu\mu f$), will accomplish the desired result. Here L is the inductance which when tuned by capacity C covers the desired frequency range of the broadcast band, and R represents the radio-frequency resistance of L. The voltage E is that measured by the instrument and this voltage will be a maximum, at any given frequency when the effective inductance of the input circuit equals the capacity of C_s and the antenna in series. Under these conditions the effective resistance of the input circuit will be from 100 to 1000 times that of the antenna proper so the antenna resistance can be neglected and E becomes proportional to the voltage induced in the antenna and therefore proportional to the effective antenna height.

The variation of resistance of different antennas therefore need not be taken into account, but some tuning compensation for antenna capacity must be provided, since C will be ganged to the oscillator variable condenser.

Two methods were investigated for compensating variations in antenna capacity. The first considered was making C_s variable. The difficulties in this case were that stray capacities from the high potential side of C_s affected readings, since they were similar in effect to a capacity shunt as explained above in connection with Figure 1 and equation (1). It was also difficult to obtain exact resonance at low frequencies where the effective Q of the input circuit is very low. By shielding C_s and connecting the shield to the top of the tuned circuit, the first difficulty was overcome, but it was found extremely difficult to construct a shield of sufficiently low capacity to permit use of a wide range of antenna capacities, including the rod antenna. Furthermore the lack of precision of resonance at low frequencies still existed.

The second method considered, and the one finally adopted, was the use of a small trimmer in parallel with the tuning capacity and adjustable from the front panel. In this case C_s is fixed. The trimmer adjustment is much easier and more precise than with the series trimmer method. The capacity C_s is made equal to that of the rod antenna, but it is shorted out when the rod is used. There is a variation in effective Q of the input circuit under these circumstances when the change from the rod to antenna is made. If the capacity of the antenna is C_A the error is C_s/C_A and the indicated effective height will be too low by this amount. Since C_s is small however, the error is not appreciable except for small antennas. Most car antennas have a capacity of 100 $\mu\mu f$ or more, so the error will be less than 3 per cent and can be neglected. If measurements are made on smaller capacity antennas the indicated effective height should be multiplied by the factor

$$1 + \frac{C_S}{C_A} \tag{4}$$

to obtain the true effective height.

We have seen that the effective height of a rod antenna can be calculated readily from its dimensions, but the capacity of such an antenna is not so easily computed. Accordingly a curve is shown in

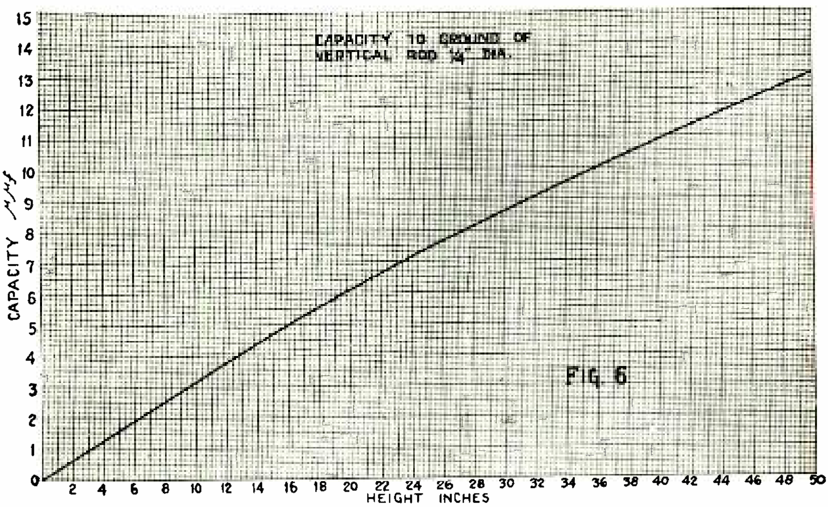


Figure 6 of the measured value of capacity for a vertical grounded rod. These measurements were made at 1000 kc with a large metal plate (approximately 3 feet × 6 feet) simulating the ground plane.

Measurement Apparatus

An instrument for measuring the effective height of automobile antennas was constructed in accordance with the requirements outlined above. A general view of the equipment is shown in Figure 7 and the schematic diagram thereof in Figure 8. It is essentially a battery-operated superheterodyne receiver with a rod antenna and is equipped with an output meter indicating carrier intensity.

The general specifications are:

Frequency range	540 — 1600 kc

Usable field intensity range 1000 — 1

185 $\mu v/m$ to 185 mv/m at 1500 kc

830 $\mu v/m$ to 830 mv/m at 600 kc

Antenna capacity range, above 100 $\mu\mu f$ without correction, below 100 $\mu\mu f$ by use of correction factor.

Tubes

1 — 1C6
1 — 34
1 — 1F6
1 — 1D1

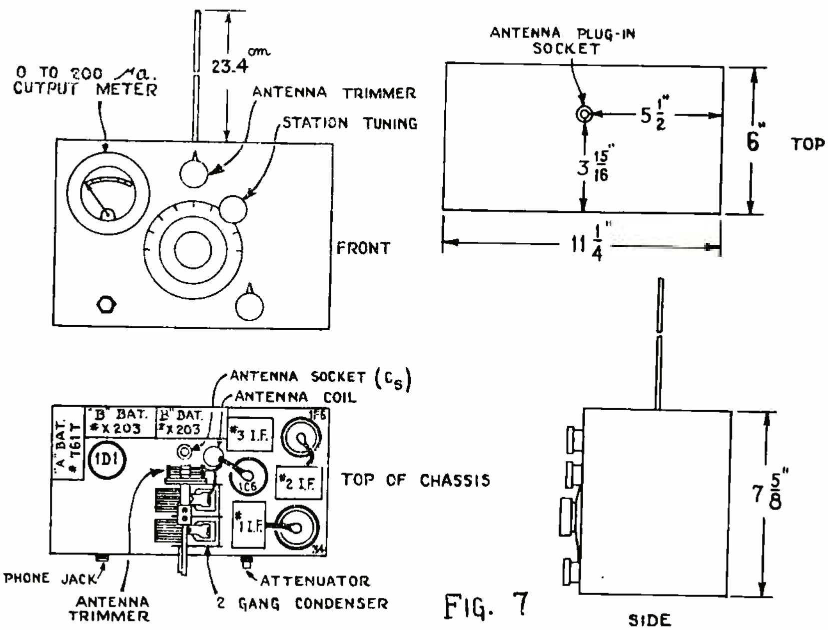


FIG. 7

Battery drain 9 ma, plate. 240 ma, filaments.

Battery voltages 90-volt B
3-volt A

Estimated battery life (size batteries shown in Figure 7) 8 hours of use.

The batteries used in the construction of the instrument were the smallest and lightest available in order to obtain a compact, easily portable instrument. Larger batteries would make the instrument more bulky, but would have the advantages of greater ease of obtaining replacements and increased hours of service. The type numbers of batteries shown in Figure 7 are those of the National Carbon Company (Eveready).

The antenna is coupled to the receiver by means of a 2.9 $\mu\mu f$ capacity, which is shorted out when the rod antenna is used. The details of construction of the coupling capacity, the rod antenna, and the adapter for connecting to the car antenna are shown in Figure 9. A coupler made up to these dimensions was used in the equipment and had a measured capacity of 2.9 $\mu\mu f$. The rod antenna should have the same capacity, so that changing from the rod to car antenna will not introduce detuning. From Figure 6 we see that this requires a length of 9.2 inches or 23.4 *cm*, which is the required height of rod above the case.

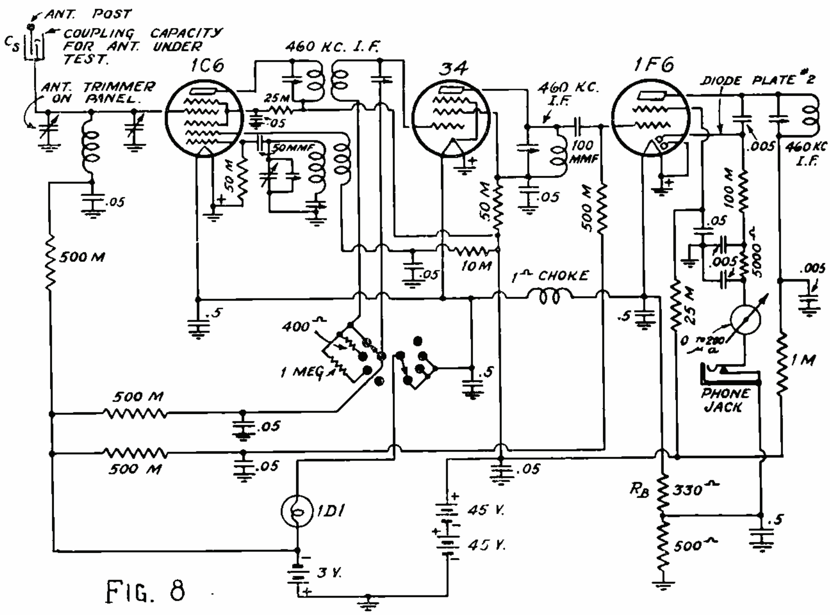


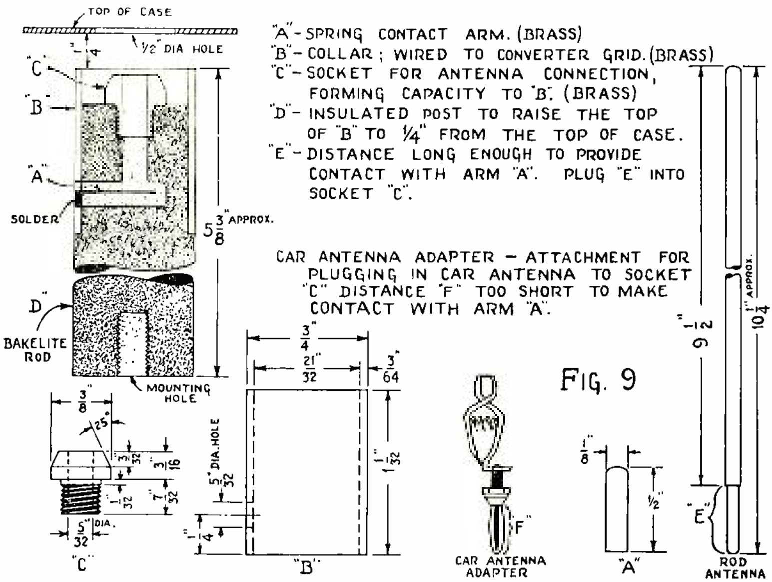
FIG. 8

The receiver has a 1D1 regulator tube to maintain the filament voltage constant.

A phone jack is provided for monitoring and to assist in identifying the station to which the receiver is tuned.

No volume or sensitivity control is provided since it was found to be unnecessary and greatly complicated the securing of a satisfactory attenuator. The attenuator consists of resistors in the secondary of the first i-f transformer for reducing the sensitivity by ten to one and one hundred to one respectively. This attenuator acts as a multiplier for the output meter, giving multiplying factors of 1, 10, and 100 for the meter indication so that a wide range of signal intensities and antenna heights can be accommodated. The resistance values for the attenuator shown in Figure 8 gave the desired multiplying factors for the i-f transformers used in the model. These values will require modification if different i-f transformers are used. Precautions should be taken to reduce regeneration to a low value, otherwise the attenuator will be inaccurate after the batteries drop in voltage or other causes operate to change the sensitivity.

The output meter is a 0-200 μa d-c meter and is connected in the No. 2 diode plate of the 1F6 so that it indicates carrier intensity values. Since effective height values are dependent on the indications of this meter, it is important to have it linear over the useful range. Since diode characteristics are not sufficiently linear, a bias from the filament circuit was added. The value of resistor R_B of Figure 8 may be



adjusted to secure proper bias. The value of R_B in the model was 330 ohms and the resultant output meter characteristic, as shown in Figure 10, is seen to be linear from 20 to 200 μa.

The last step of the attenuator switch serves to turn off the filaments.

Instrument Calibration

The attenuator multiplying factors and output meter linearity may be adjusted as described above and checked on a signal generator. If the instrument is to measure the absolute value of effective height of car antennas, the effective height of the rod must be known. We have seen that the effective height of a rod, short in comparison with the wave length, is one-half the physical height. The mathematical derivation of this relation assumes that the lower end of the rod is on an infinite ground plane. The size of the top of the instrument case is of finite extent so that the case acts to produce an increase in the effective height of the rod.

This increase of effective height may be determined and the appropriate correction made in the calculated effective height. The correction is found by temporarily increasing the area of the top of the case. If a signal is received which produces an output meter deflection A_1, as the size of the top is increased by placing thereon a large metal plate, the output meter deflection decreases. We have found that a plate 24 inches in diameter was large enough so that further diameter

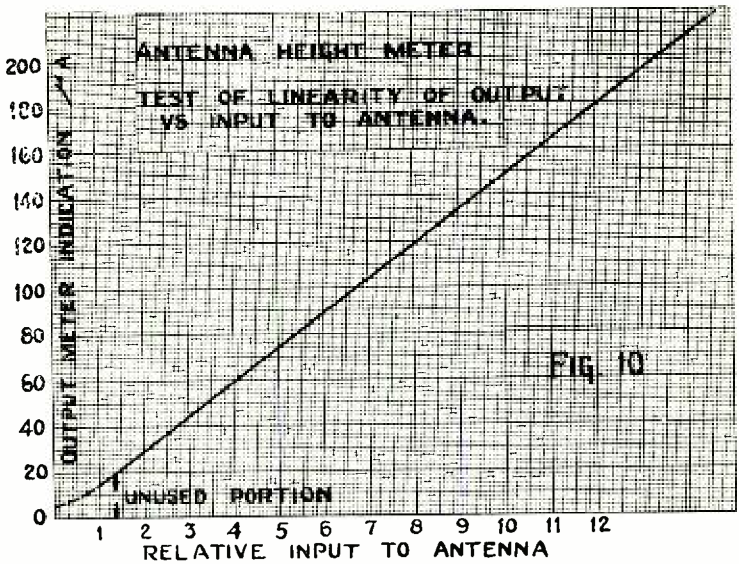


FIG. 10

increase produced practically no change of meter deflection. Let us call the deflection with the added plate A_2. The correction factor for calculated effective height is thus A_1/A_2.

With the specific apparatus dimensions shown in Figure 7 A_1/A_2 was found to be 1.2. Since the rod height above the case was 23.4 cm the effective height was

$$h' = \frac{23.4}{2} \times 1.2 = 14.05 \text{ cm}$$

Most automobile antennas have less effective height than this value.

Measurement Procedure

Measurements should be made in the open, free of buildings and irregularities of ground which might produce distortion of field intensity pattern. Set up the instrument at a distance of 20 feet or more from the car. The instrument should be placed on a wooden support three or four feet high. A metal support should not be used as it changes the effective rod height. Tune in a station and adjust the antenna trimmer to maximum meter deflection, setting the attenuator so that the meter reading is between 20 and 200 μa. The headphones should be removed from the phone jack and the observer should be four to six feet from the instrument when reading the output, to eliminate the effect of the observer's body on readings. The output meter reading multiplied by the attenuator setting then gives the rod antenna output.

The car may then be driven close to the same point that the meter occupied. Remove the rod antenna and plug in the car adapter clip and ground the lead-in shield to the car frame and to the case of the meter. Tune in the same station, adjusting the antenna trimmer and attenuator as before. Multiplication of the attenuator setting and output meter reading gives the car antenna output. Divide the car antenna output by the rod antenna output and multiply by the effective height of the rod to obtain the effective height of the car antenna.

The same measurements should be repeated on stations of different frequencies to determine variation of effective height of the car antenna with frequency. Measurements with the car pointing in four or more directions should also be made to determine the directive effect of the car antenna.

The effective resistance and capacity of the car antenna should be measured. If the capacity of the car antenna is less than 100 $\mu\mu f$ the correction factor for effective height given by equation (4) should be applied.

Conclusion

An instrument, such as described, was built and observations made thereon. It was found that the field intensity was greatest, in every case observed, when the rod was vertical. The height of the instrument above ground was found to have no effect, provided the distance to ground was at least two feet and the device was on an insulating support. It was, of course, also necessary that the observer be at least four feet from the instrument to avoid body effect. When the calibrating plate was temporarily placed on the instrument top it was found that the indication was the same whether the device was resting on the ground or several feet from the ground. Measurement of field intensities of several stations were made at a point where readings of these values had previously been obtained by a standard loop-type field-intensity meter. It was found that the indications of field intensities obtained by the meter described herein were, in no case, more than 20% different than those obtained on the standard field-intensity meter, and the average deviation was about 5%. To convert output indications to voltage equivalent for determining field intensities, the sensitivity of the instrument was measured on a standard signal generator. The signal generator output was applied through the 2.9 $\mu\mu f$ series capacity of the antenna socket. The sensitivity for 100 μa meter deflection was 600 kc, 570 μv; 1500 kc, 127 μv.

ERRATA

A Survey of Ultra-High Frequency Measurements

By L. S. Nergaard

RCA Review, October, 1938.

Mr. K. M. Soukaras of the Naval Research Laboratory has kindly called attention to an error in the formula for the phase constant β on Page 160. The last two signs should be reversed to give

$$\beta = \omega \sqrt{L_o C_o} \left\{ \tfrac{1}{2} \left[1 + \left(\frac{r_o}{\omega L_o} \right)^2 \right]^{\frac{1}{2}} \cdot \left[1 + \left(\frac{G_o}{\omega C_o} \right)^2 \right]^{\frac{1}{2}} + \tfrac{1}{2} \left[1 - \frac{G_o r_o}{\omega^2 L_o C_o} \right] \right\}^{\frac{1}{2}}$$

On Page 195, Reference 18, it is stated that Formula VIII-17 assumes that $B = \omega C$ or $B = -1/\omega L$. This statement is correct for $B = \omega C$, but in error for $B = -1/\omega L$.

When $B = -1/\omega L$, Formula VIII-17 becomes

$$G + G_L = \sqrt{\frac{C_o}{L_o}} \cdot \frac{2\pi l}{\lambda_o} \cdot \frac{\Delta\lambda_o}{\lambda_o} \left[1 - \frac{\sin \dfrac{4\pi l}{\lambda_o}}{\dfrac{4\pi l}{\lambda_o}} \right] \csc^2 \frac{2\pi l}{\lambda_o}$$

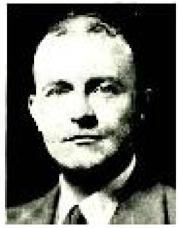
A New Converter Tube for All-Wave Receivers

By E. W. Herold, W. A. Harris, and T. J. Henry

RCA Review, July, 1938.

The date "1932" appearing in the fourth line of this article should be "1933".

OUR CONTRIBUTORS



IRVING F. BYRNES entered the General Electric Test Department in 1918 and later engaged in radio development in their Engineering Laboratory. From 1920 on he was occupied in the development of radio equipment for commercial and military vessels, submarines and aircraft. He participated in the design and tests of the early ship-to-shore duplex radio telephone equipment used on the *SS. America* in 1922. Mr. Byrnes joined the Engineering Department of RCA Manufacturing in 1930, later transferring to the Radiomarine Corporation of America in charge of engineering activities.

PHILIP S. CARTER received his A.B. degree in Mechanical Engineering from Stanford University in 1918. From May to December, 1918, he was in the Signal Corps of the United States Army, and from 1919 to 1921 was in the employ of the General Electric Company. Since 1921 he has been with the Radio Corporation of America and R.C.A. Communications, Inc., Engineering Department. Mr. Carter is a member, and on the Board of Editors, of the Institute of Radio Engineers.

JOHN B. COLEMAN was active in the Westinghouse Company's early development of broadcasting. For a short time in 1918 he served as a radio operator for the Marconi Company and then as an instructor in the Signal Corps Air Service School for Radio Mechanics at the Carnegie Institute of Technology. He graduated from Carnegie "Tech" in 1923, and joined the Radio Division of the Westinghouse Company, where he was engaged as Chief Engineer of WBZ from 1923 to 1925, and from 1925 to 1930 conducted development and design work on transmitters. From 1930 to date he has been with the Engineering Division of the RCA Manufacturing Company and is Manager of the Transmitter Department.

WILLIAM S. DUTTERA, after receiving his E.E. degree from Gettysburg College in 1929, took the two years of Masters work at Union College in 1930 and 1931. He started in transmitter test work for the General Electric Company in 1929 and in 1930 became a member of the High Power Group in the Transmitter Engineering Department. His assignments included development of the power amplifier and rectifier for the television transmitter in the Empire State Building. He transferred to the National Broadcasting Company in 1931 on television and in 1932 joined the Radio Facilities Department where he engaged in transmitter work, radio wave propagation studies and, to a major extent, in the development of improved directional and nondirectional antenna systems and associated equipment.

WILLIAM A. FITCH received his degree of B.S. in Electrical Engineering from Michigan State College in 1926. Following his graduation he entered the General Electric Test Department and later engaged in radio development in the Engineering Division. While with the General Electric Company he obtained an M.S. degree from Union College in 1931. Mr. Fitch joined the Engineering Department of the National Broadcasting Company in 1933, specializing in coverage surveys and propagation problems. He is a member of the Institute of Radio Engineers.

DUDLEY E. FOSTER received his E.E. degree at Cornell University in 1922. Prior to his college days he served during the war as a radio operator for the Marconi Company. Following graduation from Cornell he became associated with the Electrical Alloy Company and Driver-Harris Company. In 1925 he joined the Malone-Lemmon Products Company as Production Engineer, and the next year became Chief Engineer of the Case Electric Company. Two years later he went with the United States Radio and Television Company and soon after was promoted to Chief Engineer. In 1933 he became Chief Radio Engineer of the General Household Utilities Company and in 1934 took up his present duties as engineer in the RCA License Laboratory.

DEWITT R. GODDARD, a native of New York, received his B.S. Degree in Electrical Engineering from the Worcester Polytechnic Institute in 1929. In that same year he joined R.C.A. Communications, Inc., engaging in communication receiver research and development, in which capacity he still is employed. Mr. Goddard has been an associate member of the Institute of Radio Engineers since 1930.

ALBERT P. KAUZMANN received his degree of B.S. at the Massachusetts Institute of Technology in 1927. From 1927 to 1929 he served as radio-tube engineer for the Champion Radio Works at Danvers, Mass., following which he spent a year at the University of Munich. Since 1930 he has been engaged in advanced tube development in the Research and Engineering Department of the RCA Manufacturing Company at Harrison, N. J. Mr. Kauzmann is an associate member of the Institute of Radio Engineers.

CHARLES N. KIMBALL received a B.E.E. degree from Northeastern University in 1931. From Harvard University he received his M.S. degree in 1932 and his D. Sc. degree in 1934. He spent two years with the National Union Radio Corporation and since 1936 has been connected with the RCA License Laboratory. Dr. Kimball is an associate member of the Institute of Radio Engineers.

GARRARD MOUNTJOY, following his graduation with a B.S. degree from Washington University in 1929, joined the Sparks-Withington Company as a development engineer. For that company he spent some time in Europe designing radio sets for local conditions. He later became Chief Engineer of the Sparks-Withington Radio Development Department which position he held until 1935. Prior to entering college he spent a year with the Westinghouse Company. Since 1935, Mr. Mountjoy has been on the staff of the RCA License Laboratory.

DAVID SARNOFF, President of the Radio Corporation of America, has been continuously identified with radio since 1906. He received his early education in New York public schools and later was graduated from Pratt Institute, where he took the electrical engineering course. He is a fellow, Institute of Radio Engineers, and served as secretary and director of I.R.E. for three years. Mr. Sarnoff is a member, Council of New York University; member, Academy of Political Science and member, American Institute of Electrical Engineers. He holds the honorary degrees of Doctor of Science from St. Lawrence University, Doctor of Science from Marietta College, and Doctor of Literature from Norwich University. He is an honorary member of Beta Gamma Sigma and an honorary member of Tau Delta Phi. He is a colonel SC—Res., U. S. Army.

STUART W. SEELEY received his B.Sc. Degree in Electrical Engineering from Michigan State College in 1925. He was an amateur experimenter and commercial radio operator from 1915 to 1924. Following this he joined the experimental research department of the General Electric Company, and a year later became Chief Radio Engineer for the Sparks Withington Company. Since 1935 he has been an engineer in the RCA License Laboratory.

VIRGIL E. TROUANT received his B.S. in Electrical Engineering degree from the University of Maine in 1921, and the same year joined the Automotive Engineering Department of the Westinghouse Electric and Manufacturing Company. In 1924 he was transferred to the Radio Engineering Department, Transmitter Division, and was engineer-in-charge of the development group until 1933. Mr. Trouant then joined the RCA Manufacturing Company and since has been an engineer in the Transmitter Division.