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A NEW ERA IN TELEVISION

By

David Sarnoff

President, Radio Corporation of America

SINCE July 1, television has been authorized to operate commercially. That date, therefore, is recorded in the annals of television as a day of scientific recognition and economic encouragement. For research and engineering it is but another starting point packed with new challenges.

Through the years of its growth, television was officially designated as "experimental." Now, although it invites the support of sponsors who advertise, technically television has not relinquished its experimental status. Like the unlimited radio realm in which it has been nurtured, television, if it is to survive and thrive commercially, will forever be experimental. As long as it is a subject of experiment it will progress. The word "experimental" signifies that, as a science and an art, it is alive and seeking opportunities to advance in order to take every advantage of the unlimited possibilities for expansion. The many millions of dollars, the days and days of work devoted to its development are but the foundation of research that will go on and on, if television is to keep pace with the record established by all other branches of radio communication.

Cartoonists may picture the fledgling television finding its radio wings and hopping off from the edge of the nest or laboratory. For those who have watched it there can be no doubt that it will fly far to perform a new and great service in entertainment and information. But the expansion of that service and the efficiency of television as an ever-progressive medium of communication enjoyed by Americans everywhere, will depend upon the ties it maintains with its nest—the laboratory and the research men who have the power to extend and enliven its magic.

The triumph of radio in all of its branches of communication on land, sea and in the air belongs primarily to research and engineering. Radio has progressed because it never for one moment stopped research. It took advantage of every new thought that might make it bigger and better as a service to the public. Now, television, embarking upon its commercial channels, has but to look back on the achievements of radio to see what is ahead if it never breaks faith with research. It may be in the air as a commercial service, but its future always is in the laboratory. But, television need not go back to radio's brilliant pages of advancement to read a story of the promise of science. It has but to look at its own achievements so far.

The television scanning disc was invented in 1884. Almost fifty years passed before scientific research brought forth the Iconoscope and Kinescope, which provide the electronic "eyes" of television, and freed it from the shackles of mechanical operation.

Inspired by the wonders of radio and its rapid development in the field of broadcasting, experimenters in the early Twenties were confident that radio was destined to handle sight as well as sound. It was too magic a force to remain sightless. The fact that radio could talk around the world was evidence enough for those who had faith in science to know that radio also could see, and they set out intent to do just that—to see "unto the ends of the earth." That is the goal of research. It is the challenge of television. There can be little doubt that American ingenuity will make radio "eyes" as numerous in the homes and schools of America as are loudspeakers today.

The fact that television has reached a stage in its development where it merits commercial support is but a step to a threshold that leads afar, eventually to every nook and corner of the earth. Then man will enjoy radio vision just as he has the music of radio broadcasting, as well as the speed of commercial radio communications and news transmission.

Fortunate are those who are working in the television field, for its opportunities both scientific and artistic, are as plentiful and even more widespread than were those of radio in the decade of the Twenties. Television as sound and sight combined, creates and combines new opportunities that startle even the most imaginative minds. Radio through the ear makes men think; television stirs the brain through both ear and eye. It is a new and fascinating medium in which thoughts are and will be generated more and more, as the dreams and plans of research men become reality.

At this opportunity, when industrial radio and the public in greater numbers are expressing appreciation of a new industry and service, I am glad to add my congratulations to the indefatigable research experimenters and development engineers, who have contributed to the successful opening of this new era in radio science.

A RÉSUMÉ OF THE TECHNICAL ASPECTS OF RCA THEATRE-TELEVISION

BY

I. G. MALOFF AND W. A. TOLSON

Research Laboratories, RCA Manufacturing Co., Inc., Camden, N. J.

It is obvious that in a development of this magnitude many individuals contributed to the project. It is not the purpose of this brief article to cover completely this large project but rather to present a short résumé of the most pertinent factors of the development. This is deemed to be of interest because of the formal showing on, and since, May 9, 1941. It is hoped that more detailed discussions on various phases of the project, prepared by some of those identified with the particular work, may follow.

Summary—The problems of theatre television are stated. Television projection experiences of the past are reviewed. A description of the RCA Theatre-Television System recently demonstrated in New York City is given.

STATEMENT OF THE PROBLEM

THE problem of theatre television is essentially that of providing a bright picture on a viewing screen of normal theatre size; this picture having adequate resolution, contrast, and freedom from distortions. The question of how much light is needed on a theatre projection screen has been studied in the past by the Society of Motion Picture Engineers. A committe of the S.M.P.E. charged with the study of screen-brightness requirements recommended[1] a temporary screen-highlight-brightness standard of from 7 to 14 foot-lamberts. The S.M.P.E. proposed a screen-brightness standard for the purpose of making it possible to print all the release films to the same degree of contrast (gamma) and to avoid making prints of different contrast for theatres with different screen brightnesses. So far as visual satisfaction and avoidance of eye fatigue are concerned, the range of acceptable brightnesses appears to be much wider than the recommended standard. Values of screen illumination from about 1.5 to 20 foot-candles have been regarded as satisfactory at one time or the other[2]. With the wide-angle screens used in most of the theatres, this is nearly equivalent to 1.5 to 20 foot-lamberts in screen brightness. From information available on deluxe motion picture theatres it appears that the screen-highlight brightness varies between 5 and 22 foot-lamberts. In television, due to its flexibility in contrast and levels,[3] the motion-picture standards need not be adhered to, but it is reasonable to conclude that in theatre-television pictures the limiting highlight brightness should be at least of the order of the lowest value encountered in good motion-picture houses, a value which is about 5 foot-lamberts.

In a television-projection system the luminous image originates on the screen of a cathode-ray tube. This screen radiates light nearly as a perfectly diffusing (wide angle) surface. To project the image on the viewing screen some sort of an optical projection system is required. It has been shown[4] that in projecting the light from a perfectly diffusing surface on to a viewing screen by means of a conventional lens, much of the light is lost. In fact (for large magnifications) the following relation exists:

$$\frac{(\text{lumens on screen})}{(\text{lumens on tube})} \, 100 \text{ per cent} = K \frac{1}{4F^2} \times 100 \text{ per cent}$$

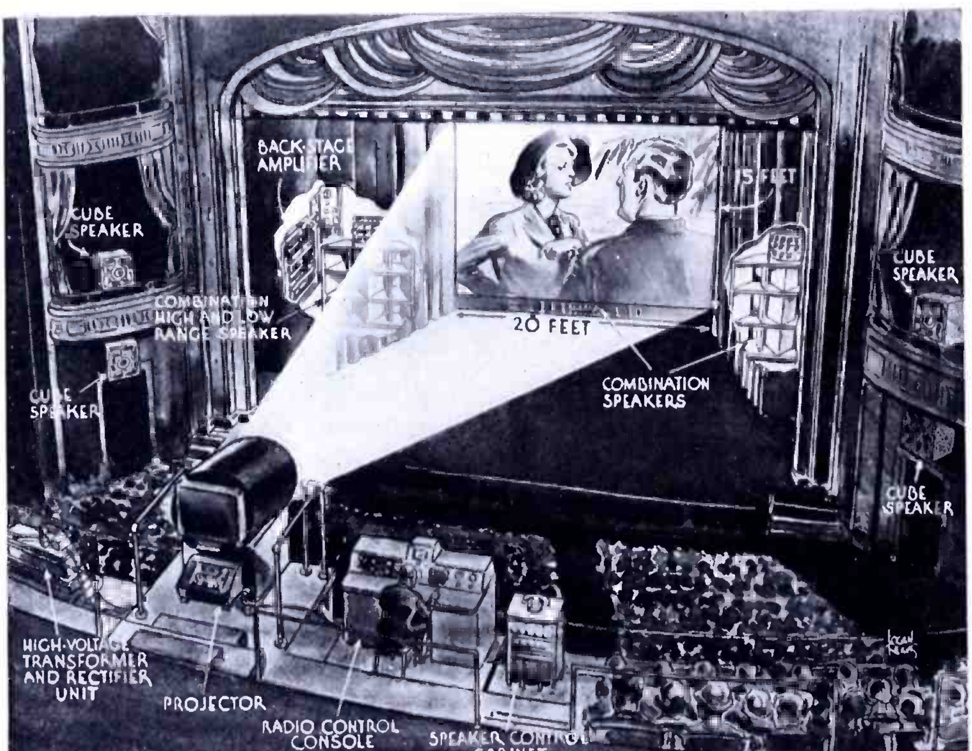

Fig. 1—Artist's sketch of the television installation at the New Yorker Theatre.

where K is the transmission of the lens and F is the "f" number of the lens. Good, commercially available, projection lenses, having a maximum numerical aperture of F:2 and transmission about 60 per cent of maximum, collect from the tube and deliver to the viewing screen only 3¾ per cent of the light generated. For a 15- by 20-foot wide-angle theatre screen (300 square feet) having 5 foot-lamberts maximum brightness, about 1500 lumens maximum of incident light is required. By wide-angle screen is meant a screen producing approximately 1 foot-lambert brightness for 1 foot-candle or 1 lumen per square foot of incident illumination. Narrow-angle directional screens produce as high as 5 foot-lamberts brightness for 1 foot-candle illumination. At 3¾ per cent efficiency this calls for the staggering figure of 40,000 lumens or 12,700 candlepower on the face of the cathode-ray tube.

On the basis of the discussion just given, the problems of theatre television may be resolved into the following:

(1) The problem of providing the most efficient optical system so as to utilize the largest possible percentage of the light generated.

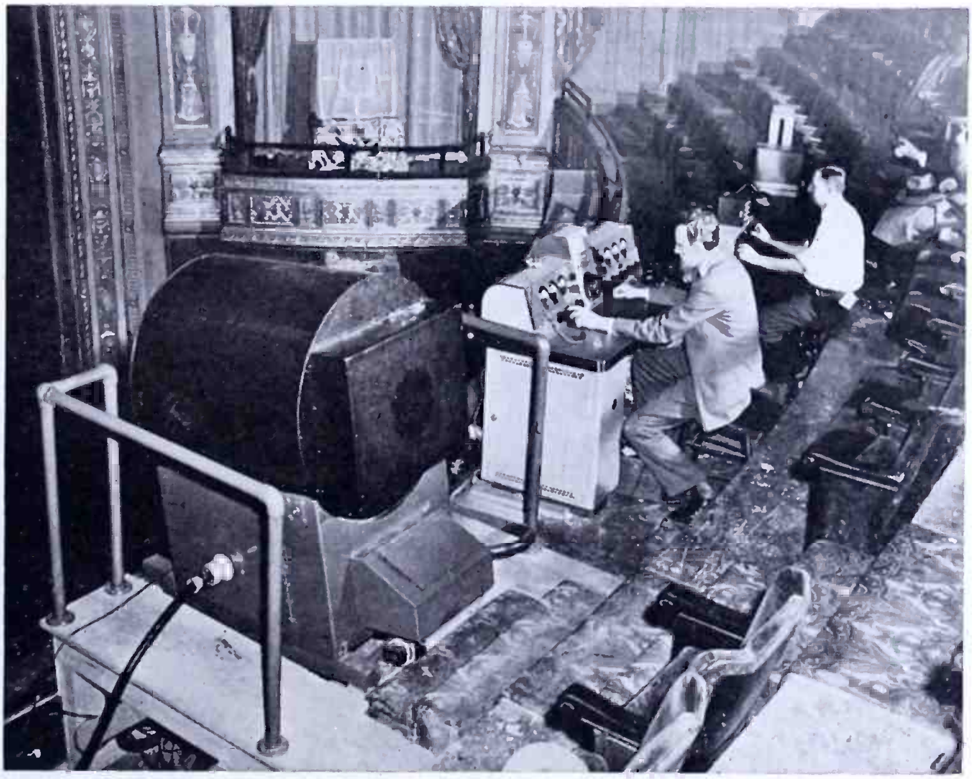

Fig. 2—Television-projection equipment installed in the balcony of the New Yorker Theatre.

(2) The problem of obtaining sufficient candlepower per unit area of the luminescent screen, by means of increased operating currents and voltages.

(3) The problem of providing a design of cathode-ray tube capable of operating at high currents and voltages.

(4) The problem of providing adequate accessories, such as deflecting circuits, video and power supplies, as well as providing adequate safety for viewers and the operating personnel from the high voltage and X-rays generated.

Television Projection Experiences of the Past

The basic aim of the RCA television-research program from the beginning has been twofold: (1) to develop apparatus for home-television service; and (2) to develop apparatus for theatre-television service. Even in the early stages of this program it was evident that while the first item could be accomplished with the aid of either the direct viewing or the projection system, the answer to the second item could be obtained only by a projection system. Therefore, the two systems—direct viewing and projection—have been carried along side-by-side, each benefiting from the other on the way. The results of the

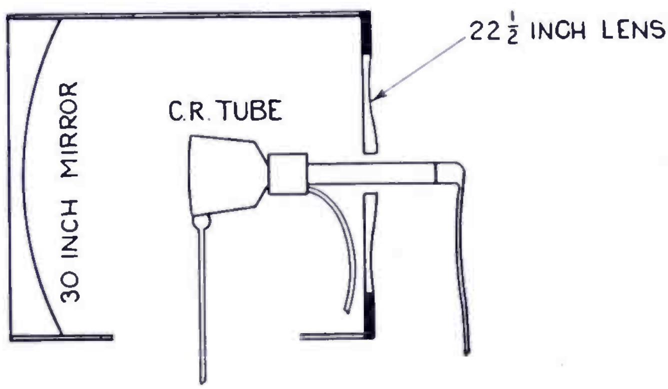

Fig. 3—Schematic diagram of reflective-projection optics showing location of cathode-ray tube.

earlier achievements in television projection have been published[5,6]. The papers cited indicate trends as they presented themselves at the time. These trends were very much in accord with the statement of the problem as it presents itself today.

The first public showing of a theatre-television system was made by the Radio Corporation of America when it demonstrated at its annual stockholders' meeting in New York City on May 7, 1940, a projected-television picture 4½ by 6 feet in size with brightness well above the 5 foot-lambert value. The demonstration was given before some 300 stockholders and press representatives. The same system was shown informally to members of the F.C.C. on February 5, 1940 in Camden, N. J.

RCA Theatre-Television System

The experience with the development, construction, and operation of the system giving a projected picture 4½ by 6 feet in size with adequate brightness, definition, and freedom from distortions indicated that the answers to the problems stated earlier in this paper had been found. The next obvious step was to build a system for a full-size theatre screen. This was done and on May 9, 1941 a demonstration of such a system, using a 441-line television signal and a projection screen 15 by 20 feet, was formally given before a large group of invited guests. The demonstration was held in the New Yorker Theatre, 254 West 54th Street, New York City. The program included dramatic sketches from the NBC Studios, Lowell Thomas, a singer, and was climaxed with a championship boxing bout. The general layout of the equipment

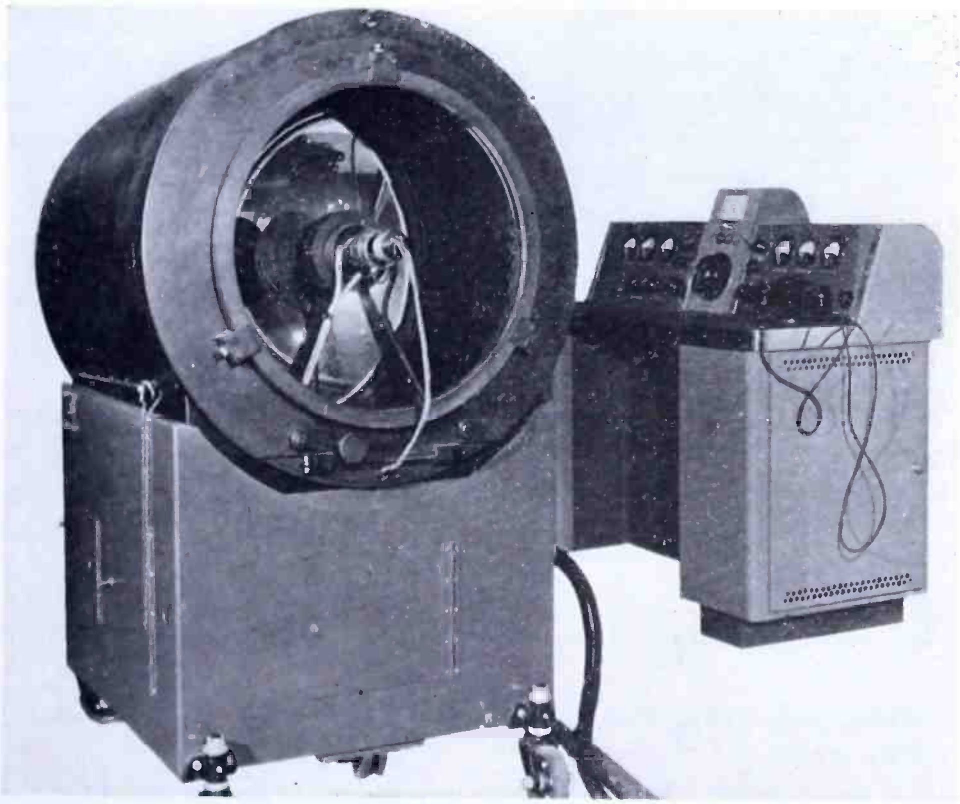

Fig. 4—Closeup of the projector and control console.

is shown in Figure 1. Previous to the formal May demonstration, one was held in the same theatre on January 24, 1941 for members of the F.C.C. and press.

All parts of the equipment used for the demonstration in the New Yorker Theatre were scaled up from the preceding system which gave the 4½ by 6-foot picture. In addition, a few improvements and refinements resulting from experiences gained in operating the smaller equipment were provided in the new unit. A photograph of the projector, control console, and sound-control cabinet in operation at the New Yorker Theatre is shown in Figure 2.

From the beginning of the development, the problem of providing an efficient optical system appeared to be the most formidable. A few per cent improvement was of no interest. Many-fold increase in the percentage of light delivered to the screen was sought. The answer was found in a reflective optical system consisting of a spherical mirror and an aspherical lens. The principle, that aspherical surfaces of various shapes may be combined into optical systems of high apertures and free of spherical aberration and coma, has been known for some time.[7,8,9] RCA opticians applied this principle to a television-projection system. In its final form the optical system is arranged as shown in Figure 3.

This system on actual tests showed 25 per cent optical efficiency; in other words, it delivered to the viewing screen 25 per cent of the light originating on the diffusing screen of the cathode-ray tube. The gain

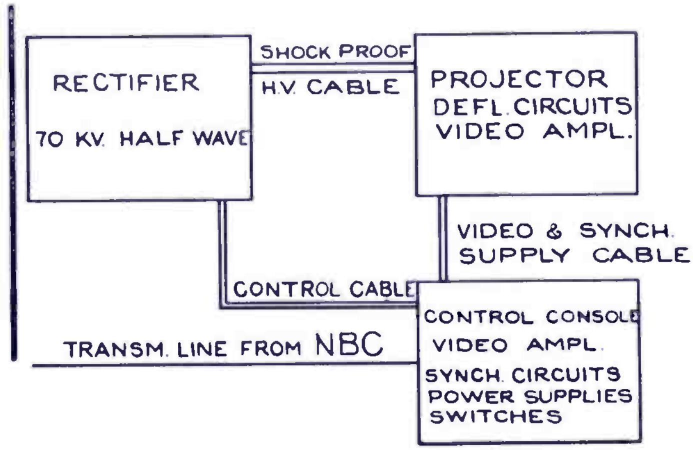

Fig. 5—Block diagram of the installation in the New Yorker Theatre.

over the conventional F:2 optical system is therefore *seven and one half to one.*

The problem of obtaining more candlepower by means of raising operating current and voltage has also been successfully solved. It was found that the thickness of the luminescent layer should increase with the operating voltages and optimum thickness was found for 60- to 70-kilovolt operation. Special provisions were worked out to avoid the so-called sticking effect.

The problem of designing a cathode-ray tube for reliable operation at 60 to 70 kilovolts was solved by introducing a new neck construction, which is now being identified as double-neck construction. The shape of electrodes had to be carefully selected and a number of refinements had to be introduced in the construction and processing of the tube. The general appearance of the tube mounted in the projector, and a closeup of the projector and control console, are shown in Figure 4.

The design of the video amplifier, deflecting and synchronizing circuits, and power supplies in a projection equipment in which the cathode-ray tube is operated at 70,000 volts maximum offered new problems; as did also the mechanical arrangement of the equipment. Some of these problems were solved by simply increasing the capacity of the units which had been used on the lower-voltage equipment. Other problems required radical changes in design and operating technique. A block diagram of the complete installation is shown in Figure 5.

Proper thicknesses of metal were chosen in the construction of the projector to insure complete safety from the X-rays generated by the high-voltage cathode rays. The installation was thoroughly checked under operating conditions to ascertain by actual measurements that the protection was adequate. Standard rules for protection from accidental contact with high voltage were followed.

The cathode-ray tube used in this installation is capable of delivering about 400 candlepower maximum of useful light. This is equivalent to about 1200 lumens. At 25 per cent optical efficiency this means 300 lumens delivered to the screen, producing 1 foot-candle illumination on the 15 by 20-foot screen. With a five-to-one directional screen, a highlight brightness of 5 foot-lamberts results. In actual demonstrations a compromise screen having directional gain of only two-to-one was used, giving a highlight brightness of slightly more than two foot-lamberts.

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[2] An Analysis of Theatre and Screen Illumination Data, by S. K. Wolf, *Journal S.M.P.E.*, Vol. 27, p. 139, (Aug. 1936).

[3] Gamma and Range in Television, by I. G. Maloff, *RCA Review*, Vol. 3, p. 416, (April 1939) and Tone Reproduction in Television (abstract), *Journal S.M.P.E.*, Vol. 34, p. 441, (April 1940).

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[5] Development of the Projection Kinescope, by V. K. Zworykin and W. H. Painter, *Proc. I.R.E.*, Vol. 25, p. 937, (Aug. 1937).

[6] High Current Electron Gun for Projection Kinescopes, by R. R. Law, *Proc. I.R.E.*, Vol. 25, p. 954, (Aug. 1937).

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[8] G. A. H. Kellner, *U. S. Patent* No. 969,785, granted Sept. 13, 1910.

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NBC'S INTERNATIONAL BROADCASTING SYSTEM

By

Raymond F. Guy

Radio Facilities Engineer, National Broadcasting Company, Inc., New York

Summary—International broadcasting as an instrument of national defense and a humanitarian service has assumed growing importance and stature. NBC stations WRCA and WNBI provide effective service in six different languages to the Portuguese and Spanish-speaking countries of South and Central America and to Europe. Recent increases in power to 50,000 watts and the development and construction of new directive transmitting antennas add to the scope and effectiveness of the NBC International Broadcasting activities. At times transmission to Latin America is accomplished with a total power of 100 kw on 9670 kc through the use of dual transmitters and antennas.

AN EXPANDING horizon of steel towers, tall wooden poles, and networks of elevated antenna wires attests to the growing international broadcasting activities of NBC stations WRCA and WNBI at Bound Brook, New Jersey. NBC has provided this service for many years.[1] International activities have been stimulated in recent years, first, in point of time, through the action of the Federal Communications Commission authorizing commercial operation, and second, by the more important contribution this service brings to National Defense.

A growing proportion of the program material consists of unbiased news broadcasts which are brought to the attention of foreign listeners with the aid of institutional sponsors and, especially to Europe, through the redoubled efforts of the National Broadcasting Company itself. Throughout Europe and the Americas there is a large and regular listening audience. In most, if not all of Europe, American news broadcasts constitute the best and often the only source of unbiased and complete news. Heavy penalties are meted out to those caught listening to such broadcasts in the oppressed countries, but there is nevertheless a substantial radio audience.

NBC has recently increased the power of the WRCA-WNBI transmitters from 25,000 to 50,000 watts and now transmits simultaneously through two separate 50,000-watt transmitters from 9:00 A.M. through the day until the early hours of the following morning and at times this service is continued around the clock 24 hours per day. A minimum of 16 transmitter hours per day are at present directed to Europe

[1] An International Broadcasting System, Raymond F. Guy, RCA Review, pp. 20-35, Vol. III—July, 1938.

during the most favorable listening periods in Europe. The balance of the program day consists of programs transmitted to Central America and to the Spanish and Portuguese areas of South America during their most favorable listening periods. The manner in which most favorable listening periods vary is indicated by the differences in time of various areas served by WRCA-WNBI.

Standard Time		*City*
Noon	in	New York corresponds to
1 P.M.	"	Buenos Aires
2 P.M.	"	Rio de Janeiro and Greenland
4 P.M.	"	Iceland
5 P.M.	"	London
6 P.M.	"	Berlin
7 P.M.	"	Athens
10 P.M.	"	Central Russia

In England the clock is advanced two hours during this Summer, contrasting with our own advance of one hour for daylight saving, thus increasing the difference to six hours.

The importance attached to international broadcasting by the European powers is evidenced by the fact that Italy directs toward the United States four program hours per day, the British send us about six and one-half hours, the Germans about ten hours and the Japanese about four and one-half hours. In addition to the programs directed to North America, these countries also provide service on from one to three frequencies simultaneously to about 20 other countries in many different languages. International personalities have been developed as a result of this service, of whom the best known is Germany's "Lord Haw Haw". The German program content covers a broad scope ranging from dance music to learned discourses on anthropology, and includes the reading of the names of new prisoners of war, request programs, question and answer periods, dramatizations of war developments, personalities, etc. etc. One offer by the German government to accept collect Radiograms relating to the German programs was accepted by several thousand listeners in the United States.

The Bound Brook plant contains NBC's international transmitting facilities plus the WJZ transmitting facilities. The original property was purchased for these combined facilities in 1925 when RCA built its first short-wave broadcast transmitter, the predecessor of the present busy stations. The original site consisted of 54 acres. Recently the expanding scope of NBC's international activities resulted in the purchase of an additional 16 acres, making a total of 70 acres, most of it devoted exclusively to international transmitting antennas. The majority of these antennas are horizontal broadside units and any future extension will probably include the replacement of other types with broadside units. They are economical of property requirements, less vulnerable than other types to damage from sleet and ice, and are very efficient. They have the added advantage that the systems can be

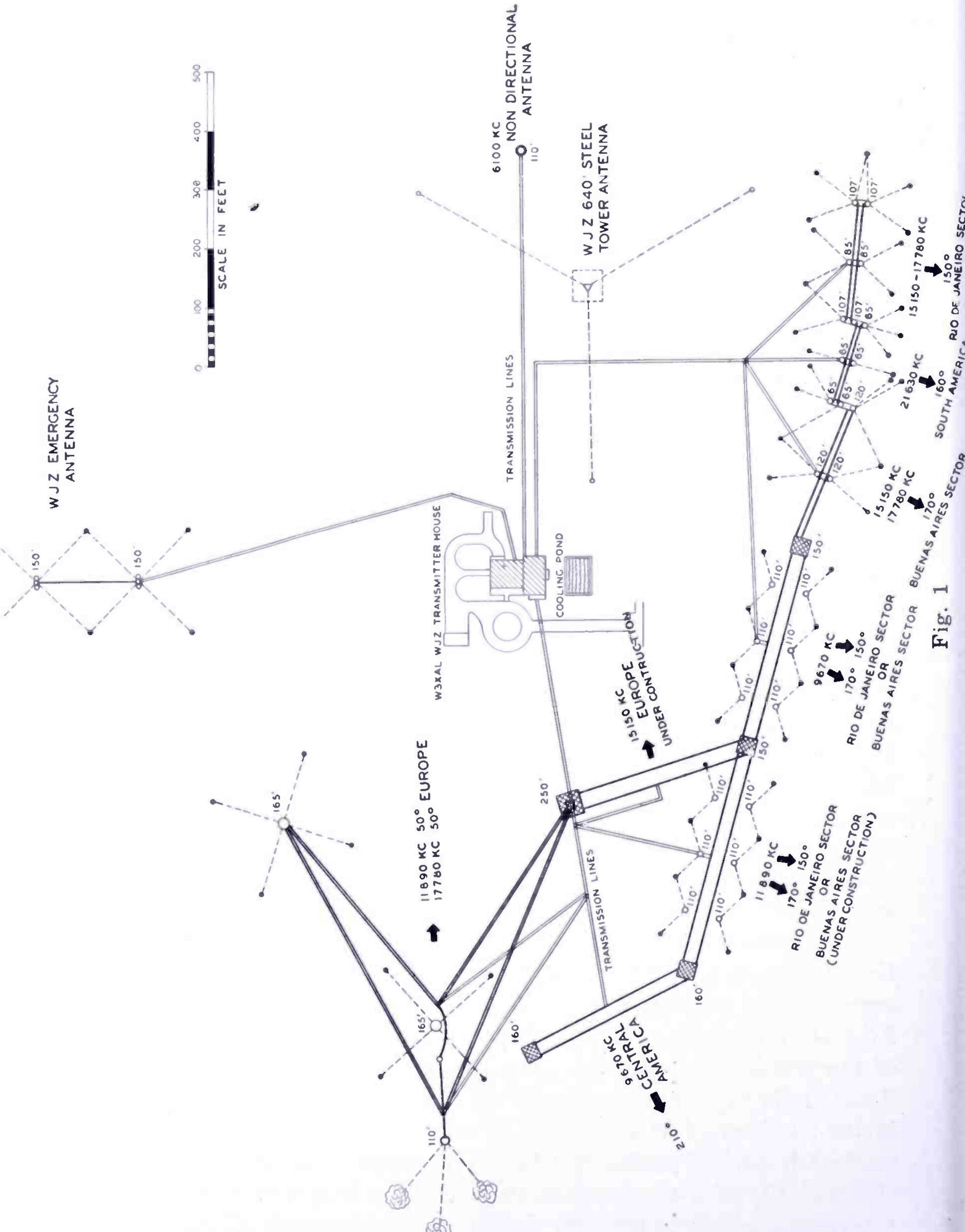

Fig. 1

designed to be steerable by phasing horizontally displaced groups of radiating elements, as described in the text which follows.

The Bound Brook site shown in Figure 1 is rectangular in shape with the long dimension facing in the general direction of the Latin American countries. This is important because a considerable number of antennas is necessary properly to serve the various sectors within the span of 100 degrees between Pernambuco and Mexico City. The short dimension of this site faces Europe. A smaller number of antennas will suffice to serve the European sector inasmuch as the span between Moscow and Madrid is only 30 degrees. To serve Latin America no fewer than three separate groups of antennas, divided approximately equally across the 100-degree Latin American sector, are required to give service to all areas. The problem presented in covering this great expanse is dealt with in more detail in other sections herein.

Antennas—Language Areas

Brazil is unique among Latin American countries in that it is the only one having Portuguese as its native language. Throughout the balance of Central and South America the native language is Spanish. Obviously listeners in Brazil, speaking Portuguese, are little interested in programs transmitted in Spanish, English or other languages. Similarly, listeners in the other portions of South and Central America, speaking Spanish, are little interested in programs in other languages. Therefore, to broadcast effectively to the two different language areas, individual antennas and separate language services are essential. As an example, nightly service is provided to Brazil on a 9670-kc beam during one part of the evening and to the Spanish speaking countries at another time on two other 9670-kc beams. To provide European service on this frequency a fourth 9670-kc beam would be required. It may be seen that to provide service to four basic areas with the most useful of six different assigned frequencies and six different languages, and do it with the optimum beam pattern, requires an impressive list of antennas.

In the design of a directive antenna the engineer must decide the proper compromise between power gain and beam width because one is, of course, obtained at the expense of the other. A beam broad enough to cover all of South America would have such low gain that it would not provide the signal intensity required to give satisfactory service. The design of antennas of very high power gain presents no special problem in itself. It would be a comparatively simple matter to increase the power gain to the point where a somewhat higher concentration of field were obtained in one area at the excessive sacrifice of adjoining areas of importance. To insure satisfactory field intensity to South America, at least two separate antennas must be used for each frequency, one directed toward the Portuguese areas and the other toward the Spanish areas. To cover Central America, still a third antenna is required. Service to Europe must of course be provided on a different array of antennas. It happens that the distribution of service areas in South America and Europe are such that the optimum

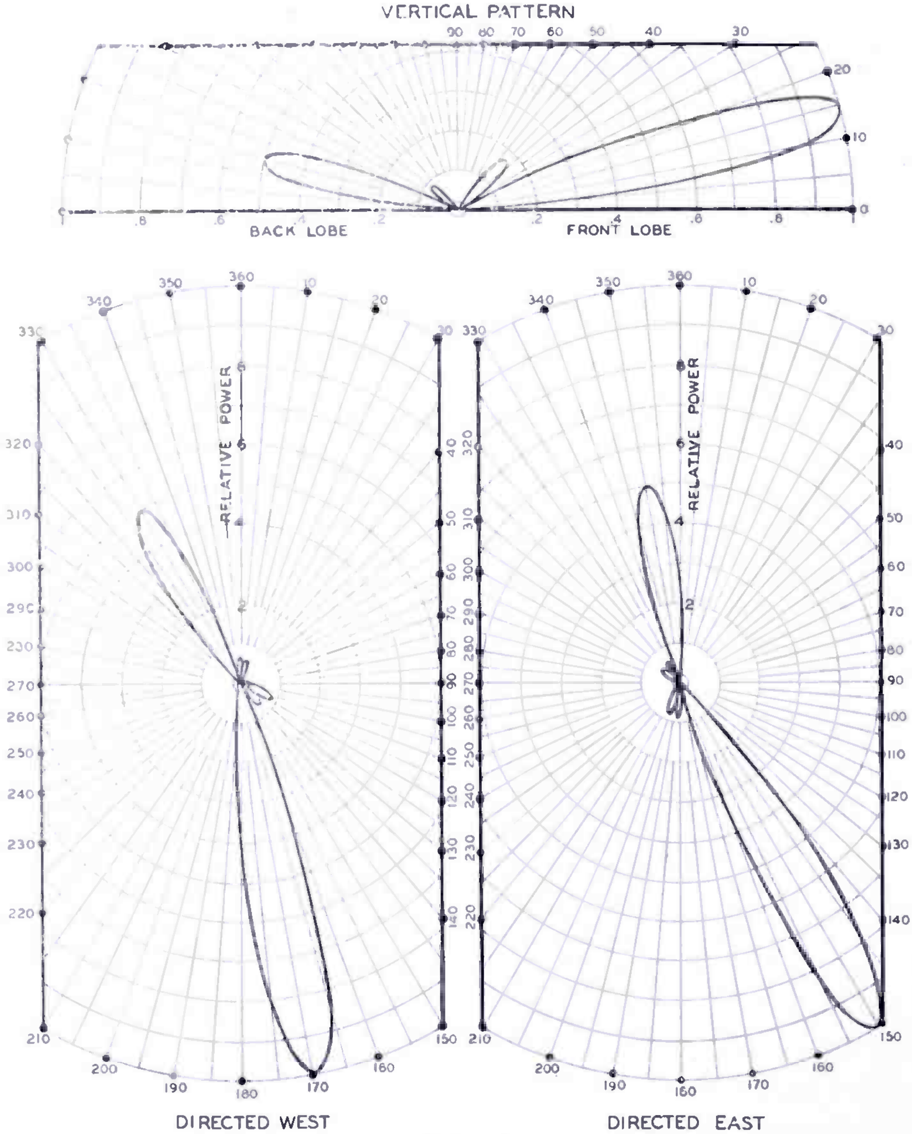

Fig. 2

beam width is in each case about the same. As a result NBC has practically standardized an antenna design which was evolved to give optimum performance to Europe and South America. It is the result of many years of applied experience and constant observation.

The angle 150° E of N centers on Rio de Janeiro, the middle of the concentration of population of Portuguese-speaking Brazil. It is also almost exactly in the center of the country as viewed over the great circle path from New York. It is natural that 150° should be the direction of a Portuguese language beam and it has been selected as optimum for that utility.

Buenos Aires is not quite so ideally centered geographically, but is the center of the greatest concentration of Spanish-speaking peoples on the South American continent. Considering population distribution, 170° was selected as the optimum angle for the Spanish language South American beam. On the theory that one picture is worth ten thousand words, the reader is referred to Figures 4, 5, 6, and 10 which show distributions of field intensity. In transmitting to Central America there is room for more latitude. The present 9670-kc antenna is directed 215° E, but in a pending move it is intended to shift the beam a few degrees to the south.

The European antennas are directed 50° E centering on the major European capitals which are quite closely grouped as viewed over the great circle path from New York.

The following tables show the number of radio receivers and in some cases the number of telephones per hundred persons in the United States of America and in the countries of Latin America.

Country	*Receivers*	*Telephones per hundred persons*
U. S. A.	....	15.37
Bahamas	1,880	
British Honduras	1,200	
Costa Rica	20,000	
Cuba	150,000	1.29
Dominican Republic	7,000	
Guatemala	21,700	
Haiti	3,000	
Honduras	16,000	
Mexico	300,000–350,000	.81
Nicaragua	4,000	
El Salvador	8,500–10,500	
Panama	24,000	
Argentina	1,050,000	3.13
Brazil	500,000	.59
Chile	160,000	1.69
Colombia	100,000	.44
Ecuador	6,500	.29
Paraguay	11,200	.35
Peru	68,000	.43
Uruguay	150,000	2.2
Venezuela	138,000	.67
Bolivia	....	.08
Central America	....	.40
Puerto Rico	....	.86

Over a period of many years there has never been an interval when construction of new NBC international radio facilities was not being carried forward at Bound Brook under special capital appropriations provided for the purpose. The increasingly diversified type of operations, the nature of cyclic changes in wave propagation, the expanding scope of the service, the growing importance to National Defense, new developments in circuits and circuit elements and increases in the number of assigned frequencies all contribute to the occasional need for new or improved radio facilities. The inventory of directive antennas in use or under construction is shown below. Of this inventory only two units, the third and sixth, are not yet ready for operation. The list of nondirective and comparison antennas is not included.

Frequency	*Language Area*	*Degrees*	
9670 kc	Brazil—Portuguese	150° E of N	Steerable
9670 "	Spanish	170° "	
6100–9670 "	Spanish—Central America	215° "	
11890 "	Brazil—Portuguese	150° "	Steerable
11890 "	Spanish	170° "	
11890 "	South America	160° "	To be dismantled
15150 "	Brazil—Portuguese	150° "	Dual Frequency Unit
17780 "	Brazil—Portuguese	150° "	
15150 "	Europe	50° "	
11890 "	Europe	50° "	
17780 "	Spanish	170° "	Dual Frequency Unit
15150 "	Spanish	170° "	
17780 "	Europe	50° "	
21630 "	South America	160° "	
17780 "	South America	160° "	To be dismantled

There are at least five methods of specifying antenna gain, all different. They all involve comparison of the "unknown" antenna field with the field of some simple "comparison" antenna. If the "unknown" antenna is compared with different types of "comparison" antennas, or similar "comparison" antennas in different positions, the apparent power gain of the "unknown" antenna is different in each case. Some of the methods of specifying antenna gain are described as follows:

1. Absolute directivity. This compares the maximum signal power of the "unknown" with the signal power of a vertically and horizontally non-directive comparison antenna, all assumed to be located in the center of a sphere.
2. Comparison with a Hertz doublet.
3. Comparison with a half-wave horizontal dipole at the same height above ground. This compares the maximum signal power of the "unknown" with the signal power of a half-wave dipole, the two ordinarily being located side by side.

4. Comparison with a vertical dipole with its center ¼ wavelength above the earth.
5. Comparison with a horizontal half-wave dipole one-half wavelength above the earth.

The indicated power gains are different for these methods over a range of more than 60 per cent. NBC has used, for comparison, a half-wave dipole horizontally polarized and at the same height above ground as the unknown. This method and one other, comparison with a half-wave dipole one-half wavelength above ground, are believed to be most used.

The "same height" method gives the smallest indicated power gain of any, but is the only method which expresses the actual gain of an array over an effective half-wave dipole such as would be used for broadcasting.

It may be seen on Figure 1 that the 70 acres are almost ideally laid out to provide service to the nations interested in the viewpoint of the United States. The transmitter building is located in the center of the property with transmission lines radiating in various directions to the transmitting antennas. The antennas are adequately isolated from each other and the frequencies are so staggered among these antennas that coupling between adjoining units is kept to a negligible amount.

Figure 2 shows the polar power pattern of the type of antenna which has been adopted by NBC as best for the service areas in South America and Europe. The figure shows the two beam patterns for the steerable type antenna to be described. This pattern was adopted years ago by NBC after consideration of population distribution, the size of the distant service area, the power gain and other factors.

This pattern is obtained with a broadside antenna having a horizontal width of six half-wave elements two tiers of such elements being stacked vertically, with a complete duplicate set of such elements to the rear, making a total of 24 radiating elements. The 12 rear elements forming the reflecting tiers are parasitically excited.

NBC Steerable and Dual-Frequency Antennas

The most important language areas of South America are centered around Rio de Janeiro and Buenos Aires. These areas are 20 degrees apart as viewed from New York. One of the earliest investigations conducted in connection with these antennas was to determine whether or not a single antenna could be built which could be steered to either of these two language areas. A cost study showed that for the lower-frequency antennas which involve massive supporting structures and long spans, an electrically steerable antenna entirely satisfactory in performance would be cheaper than the construction of separate antennas for the two areas. The 12 elements which are driven are segregated into three groups, each fed separately, because it was determined that the desired radiation pattern could be obtained by separately feeding three bays. The center bay is always kept at zero degrees for reference to the outside bays. It is desired to steer this beam ten degrees to one side or the other of the center line to serve either Rio de Janeiro or Buenos Aires. This can be accomplished when one outside bay is retarded 75 degrees and the other one is advanced 75 de-

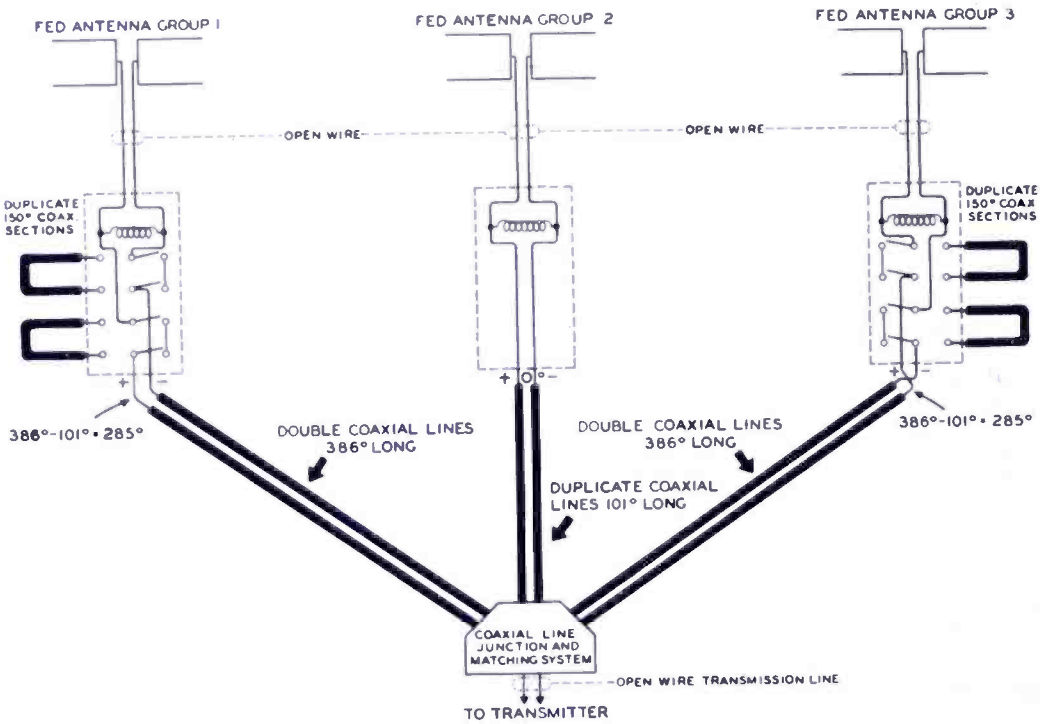

Fig. 3

grees with respect to the center bay. The manner in which this is accomplished is shown diagramically in Figure 3. The transmission line from the station building divides into three coaxial branches, terminating in phasing boxes directly beneath the three sets of down leads. The branch to the center down lead is 101 electrical degrees long. The branches to the outer phasing boxes are 386 degrees long. The net result is that the outside phasing boxes lag 285 degrees with respect to the center box. The advance condition which makes reversible steering possible requires that both the outside boxes be 75 degrees ahead of the center box. This was easily accomplished by proper choice of the branch line lengths as described above. In other words, a lagging phase of 285 degrees is equivalent to a leading phase of 75 degrees. Therefore, if all three of the down leads were connected directly through the phasing boxes to the transmission lines, the outside elements would each be advanced in phase 75 degrees with respect to the center box.

It was stated above that to steer the beam to the east or west, one outside box must be advanced 75 degrees and the other one retarded 75 degrees. Since, when connected directly through, both outside sections are advanced 75 degrees, it is possible to obtain the desired steer-

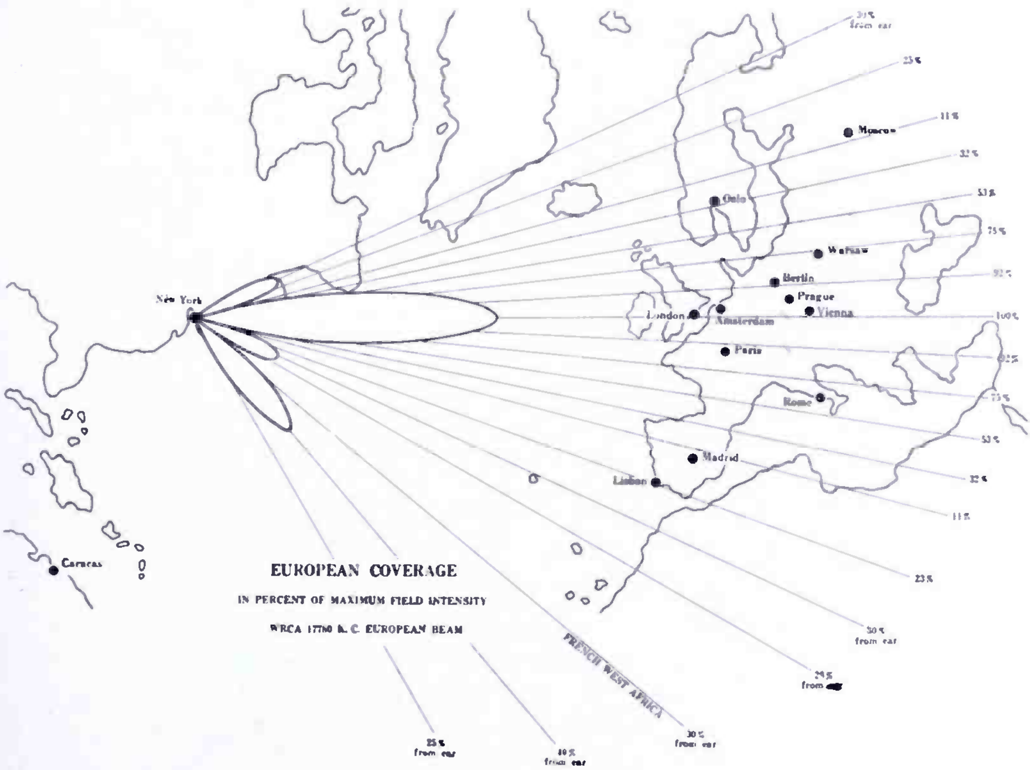

Fig. 4

ing condition by retarding the phase of one or the other outside elements 150 degrees. This is accomplished by inserting by means of specially built high-frequency contactors, 150-degree "building out" circuits which introduce a lag of 150 degrees into one side or the other. The contactors are operated from the transmitter control point in the transmitter building by remote control and are interlocked with the main rectifier in such a way that the rectifier may be shut down, the antenna directivity changed and the rectifier re-energized in one operation. Coaxial transmission lines are used for the 150-degree phasing-out sections and also to connect the phasing boxes to the common junction of the three branches, as shown in Figure 3.

The impedance of the down leads must be matched to the output of the phasing boxes and this is accomplished by circuit elements adjacent to the relays in the boxes.

The 9670-kc steerable antenna has been in daily service for some years and has been highly satisfactory. On the higher frequencies in which the supporting structures and spans are smaller, separate an-

Fig. 5

tennas are more economical to build. In this connection an investigation was made to see what advantage could be taken of operation of one antenna on two or more separate frequencies. It is possible to properly terminate and operate an antenna on several frequencies, but if they are too widely separated the antenna pattern departs too far from the optimum.

Two frequencies, not widely separated, may be used on one antenna without a significant change in the directive pattern. As a result of

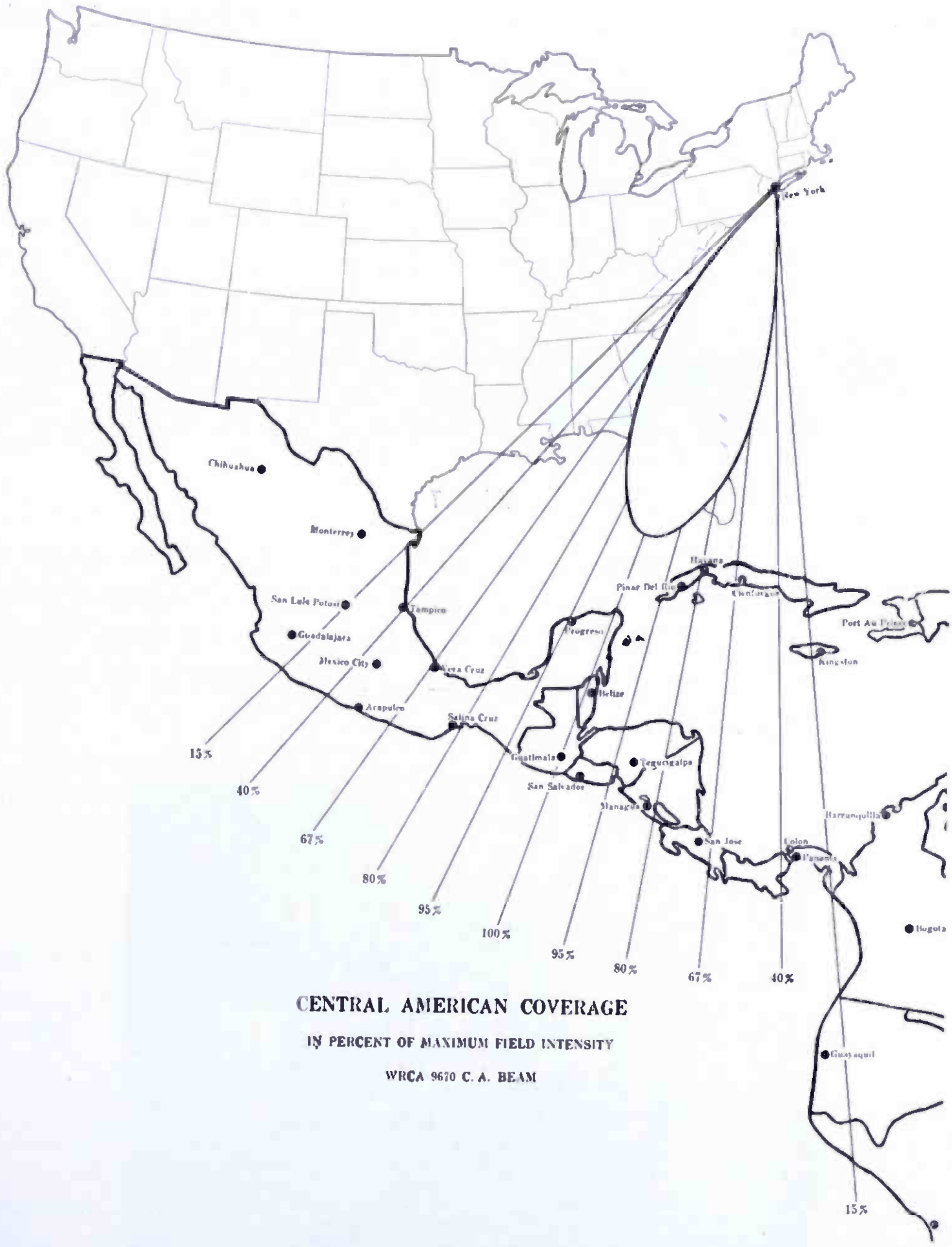

Fig. 6

this investigation two antennas, one directed to Buenos Aires and the other to Rio de Janeiro were designed and built so that each could operate on either 15150 kc or 17780 kc. The patterns are almost alike, the antennas can be terminated correctly for each frequency and the overall result is that two fixed antennas accomplish in this way substantially the same result that would be obtained with two steerable antennas.

Figure 4 shows the field intensity distribution, in per cent of maximum, throughout Europe, when the 17780-kc European antenna is transmitting. Figure 5 shows in the same manner the field intensity

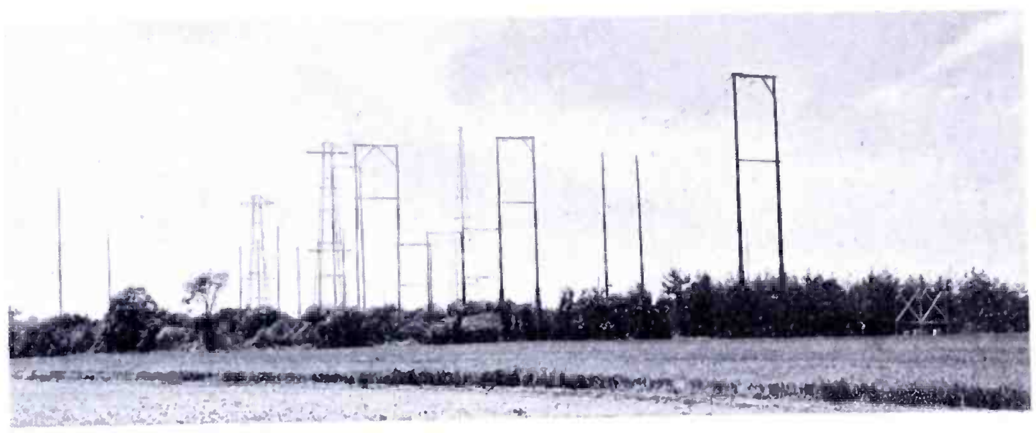

Fig. 7

distribution throughout Brazil for the east leg of the 9670-kc steerable antenna. Figure 6 shows the field intensity throughout Central America for the 9670-kc Central American antenna. Space does not permit showing patterns for all antennas nor is it important to do so when they are nearly all alike.

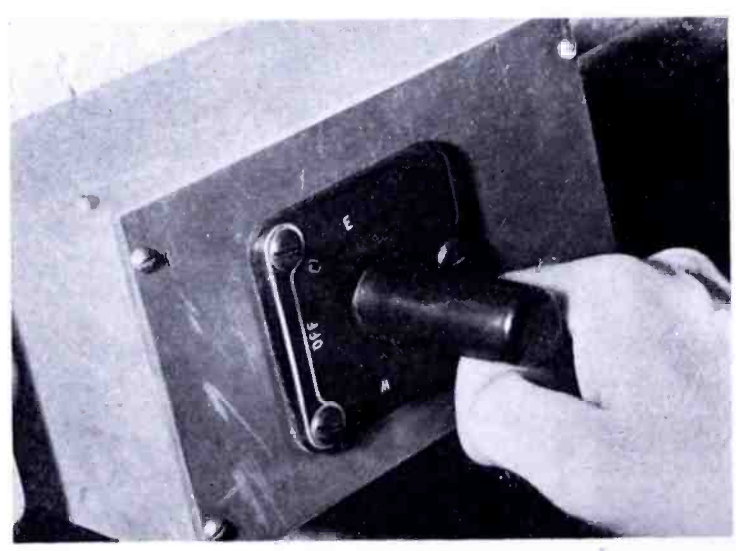

Fig. 8

The group of antennas used in transmitting to South America is illustrated in the photograph of Figure 7. Figure 8 shows the type of switch employed, and the only operation necessary, to direct the beam of the steerable type antenna developed by NBC.

In the design of beam antennas such as are used at WRCA and WNBI, maximum forward transmission is not accomplished with the same adjustment which would produce maximum suppression to the rear. Therefore, an appreciable amount of power is transmitted to the rear of such a system. When the steerable antenna is directed to the west leg an effective beam power of about 120 kw is transmitted in the direction of Alaska. A Fairbanks station has requested and received permission to rebroadcast the WRCA programs thus received. Simi-

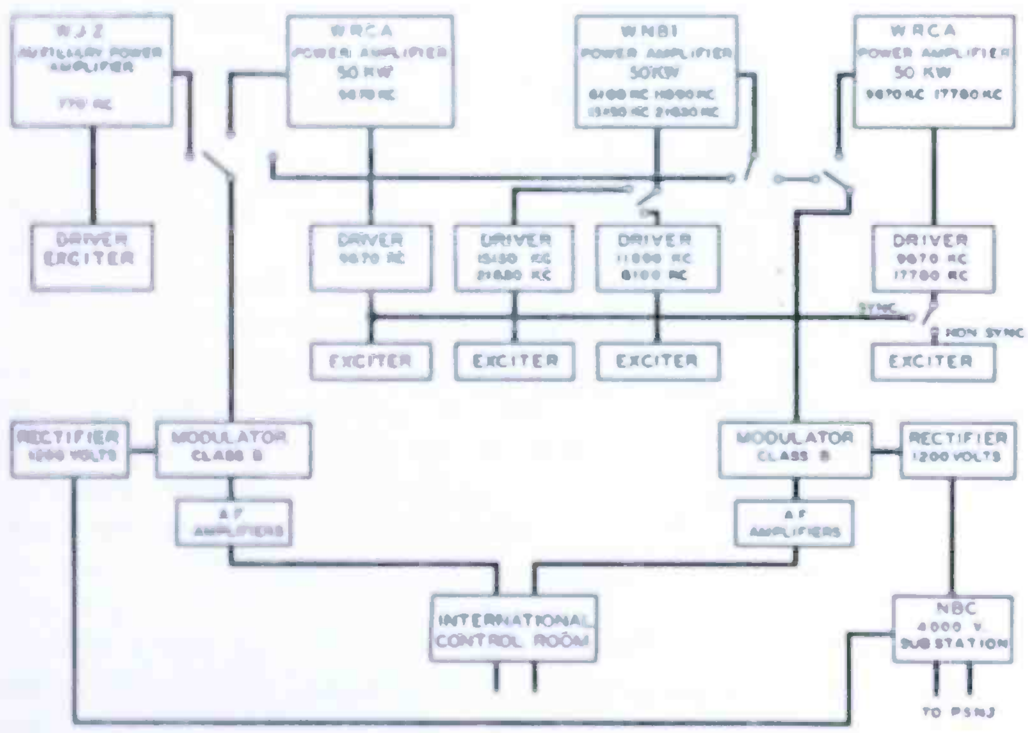

Fig. 9

larly, a station in Puerto Rico has requested and received permission to rebroadcast the programs of WRCA and WNBI.

WRCA-WNBI Transmitters

The transmitter facilities of WRCA and WNBI are shown in Figure 9. Facilities are provided for the operation of two transmitters simultaneously. The frequencies are divided between these two transmitters in such a way that the maximum number of useful adjacent frequency combinations can be used together. Six frequencies are assigned to WRCA and WNBI, 6100 kc, 9670 kc, 11890 kc, 15150 kc, 17780 kc, and 21630 kc. Dual 50-kw transmitters are available, the WNBI transmitter having one power amplifier and duplicate exciters identical except for the operating frequencies. WRCA is provided with similar duplicate exciters and also two duplicate power amplifiers differing only in operating frequency. The frequencies are licensed as follows:

WRCA	WNBI
9670 kc	6100 kc
17780 kc	11890 kc
	15150 kc
	21630 kc

Since adjoining frequencies are ordinarily most desirable for duplicate operation over a particular circuit, the above division was requested of the FCC, and granted. At the proper times, governed by time of day, path distance, direction, and 11-year sun spot cycle conditions, the following useful combinations are possible.

6100-9670
9670-11890
11890-17780
15150-17780
17780-21630
9670-15150
9670-9670 synchronized for 100 kw

These combinations permit combined operation on two separate frequencies to Europe, the Rio de Janeiro sector, the Buenos Aires sector, or Central America. They also permit operation separately to any two of the four basic service areas. In addition another interesting operating condition has been worked out and has been successfully used for some time as described below.

100-Kw Operation

During certain evening hours it is desirable to serve all of the Spanish speaking areas of Central and South America at the same time with the same program on the same frequency. This could be accomplished by the use of a very wide beam of low power gain, but it would, under such conditions, be necessary to increase the transmitter power to make up for the reduced gain of the wide-beam antenna. The alternate method would be the use of separate high-gain antennas serving adjoining areas, each antenna being driven by a separate transmitter. The latter method has been developed and is in daily use at WRCA.

The two WRCA 50-kw amplifiers are used simultaneously, one to excite the Central American antenna and the other to drive the west leg of the South American steerable antenna, both operating synchronized on 9670 kc. The beam-patterns overlap to provide service over the vast Spanish-speaking area extending from the Brazilian border to Central Mexico. Figure 9 shows the two r-f systems. Successful operation of twin transmitters requires, of course, that the frequencies be exactly synchronous. As may be seen, synchronization is simply and easily accomplished by cross connecting the separate drivers, eliminating one of the crystal units, and driving both transmitters from the other one alone. During the experimental period of twin frequency operation there was some concern that circuit irregularities might cause independent variations in two separate signals to result in increased distortion or violent fading in the overlap areas. The results were observed as closely as circumstances permitted, but no evidence has come to hand to indicate that twin frequency operation as used at WRCA is attended by any disadvantage. This system has been in nightly operation over an extended period and all reports are favorable.

Figure 10 shows the field intensity distribution throughout the Spanish-speaking areas of South and Central America when synchronized 100-kw operation is employed to cover them on 9670 kc.

The power gain of the steerable antenna is 24, in comparison with the gain of a half-wave horizontal dipole. Actual field intensity measurements made in Argentina during alternate transmissions on the steerable unit and the comparison dipole showed that this gain was obtained within a few per cent. Thus, the effective power with 50-kw transmitter output is 1,200,000 watts. The Central American antenna was built with a wide beam to cover the Central American arc of 60°. The requirements for serving this area are different than for South America or Europe. The power gain can be lower in this case to get the required wide pattern, while still maintaining high signal intensity, because Central America is comparatively close to New York. The power gain is 10 and, as a result, the effective power on this beam is 500,000 watts. Therefore, the combined effective powers on these two adjacent beams is 1,700,000 watts. These beams and effective powers can easily be made more concentrated, but to do so would somewhat defeat their purposes since it would be at the expense of the beam widths required to serve reasonably their respective areas.

Prior to 1941 the WRCA-WNBI transmitters were operated daily with a normal carrier power of 25 kw, but recent reconstruction has increased the power to 50 kw. Because of the importance of the service which had been built up over many years it would have been most undesirable to discontinue transmissions on any of the regularly used frequencies for the period of time necessary to reconstruct the exciter and power-amplifier units for 50-kw operation. Therefore, before reconstruction could start on the first transmitter, it was necessary to arrange the operating schedules of the other two transmitters and install satisfactory but temporary wave-changing facilities in them to permit two transmitters to cover the schedules of transmissions originally assigned to three. In addition, r-f transmission lines and switching facilities had to be transferred. Thus, a great deal of preliminary work was required before a transmitter could be removed from service for reconstruction. Then upon completion of the first 50-kw transmitter, temporary wave-changing facilities had to be built into it to facilitate its operation on the frequencies of the second transmitter so that no program interruption resulted. After this was accomplished, the

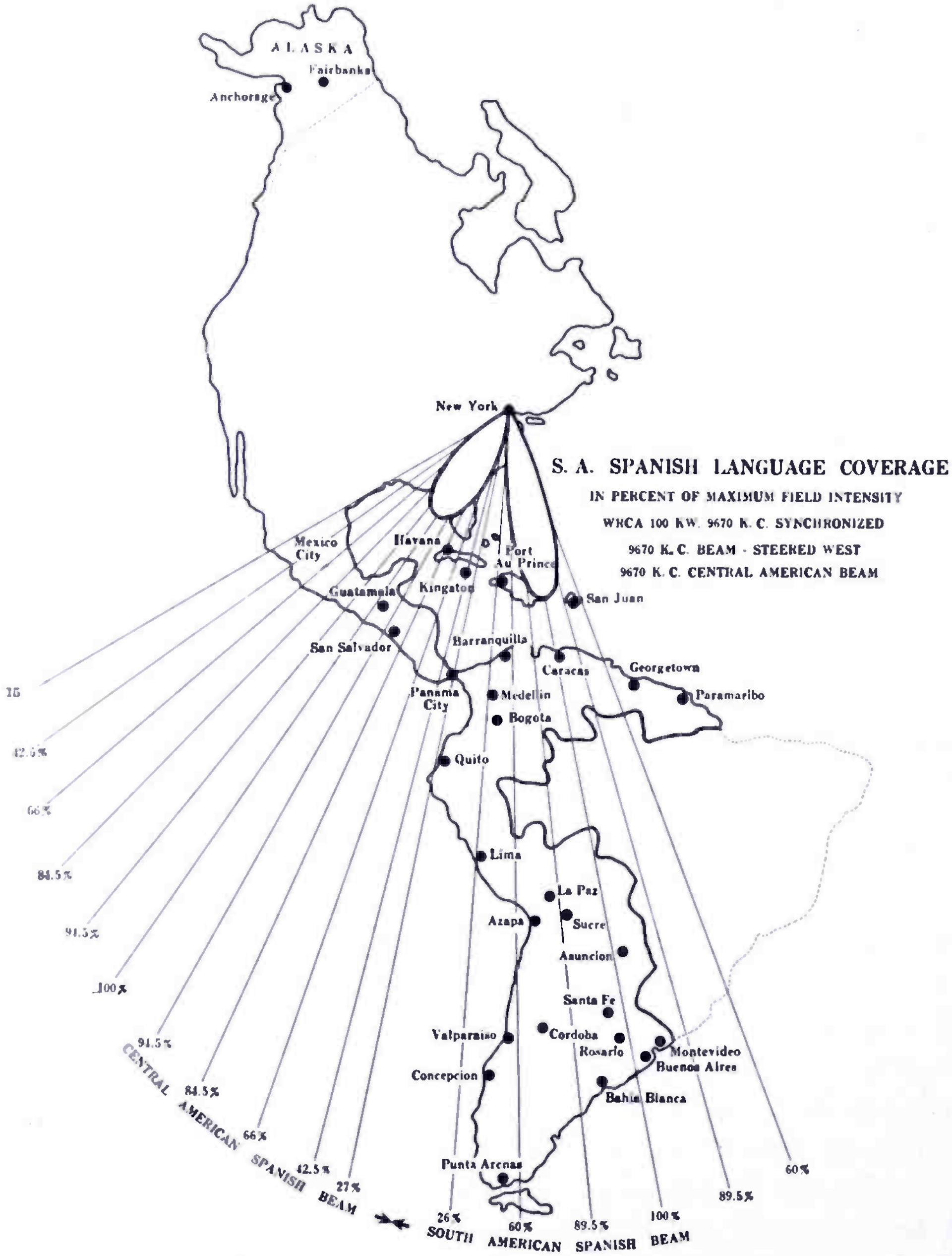

Fig. 10

second transmitter was made available for reconstruction for 50-kw operation and so on with the third transmitter.

An inordinate amount of extra work and planning were required under these conditions to rebuild completely all three transmitters. Most of it had to be done after the 1 A.M. sign-off period at night because only then could power be removed from the WRCA-WNBI equipment.

As a part of the project it was necessary to increase the power-handling capacity of the main 4000-volt substation and also expand the water-cooling facilities to withstand the added plate dissipation attendant upon the transmitter power increases.

Coinciding with the power increases of the transmitters, the power-handling capacity of r-f transmission lines and antenna systems had, also to be increased by installing larger transmission line and antenna cables and increasing the insulation. Most of this work also had to be performed at night after the end of the daily schedules.

Simplification of mechanical design has been accomplished by the use of removable shelf-type construction utilizing duralumin frames, panels, chasses, and shelves. For maintenance and repair of the low-power stages the shelves are removable on sliders from the front panels, like desk drawers. Other stages are made accessible through interlocked doors, or perforated screens.

As a result of many years of high-frequency broadcasting transmitter experience, the high level modulating system was retained as the most satisfactory. The RCA tube lineup in each transmitter starts with crystal controlled oscillator and frequency-multiplier units containing an 802 crystal oscillator and two 802 doubler stages. The outputs of the second doubler stages are amplified by 813 amplifier stages and push-pull 803 stages. These in turn are used to drive the intermediate power-amplifier stages utilizing four 833A tubes in push-pull parallel arrangement.

The outputs of the 833A stages of the exciter units are normally 3.5 to 4 kw and drive the grids of the push-pull 880 modulated power amplifier stages to saturation. All r-f stages are equipped for a-c filament heating from the 60-cycle power mains through suitable transformers.

The modulator units of these transmitters consist of cascade 211 and 845 audio amplifier stages and push-pull 891 driver stages. These driver stages have the desired excellent regulation for driving the 893 Class B modulators. The drivers and modulator stages operate at 10,500 volts of d-c power supplied by 3-phase full-wave main rectifier units utilizing six 857B hot-cathode mercury vapor rectifier tubes in each.

The power amplifier units of these transmitters are of improved design with a view toward greatest efficiency and operating stability. In each case two 880 power tubes, shown in Figures 12 and 13, amplify the transmitter carrier power to 50 kw. Modulation is accomplished with audio power provided by the RCA 893 Class B modulator unit.

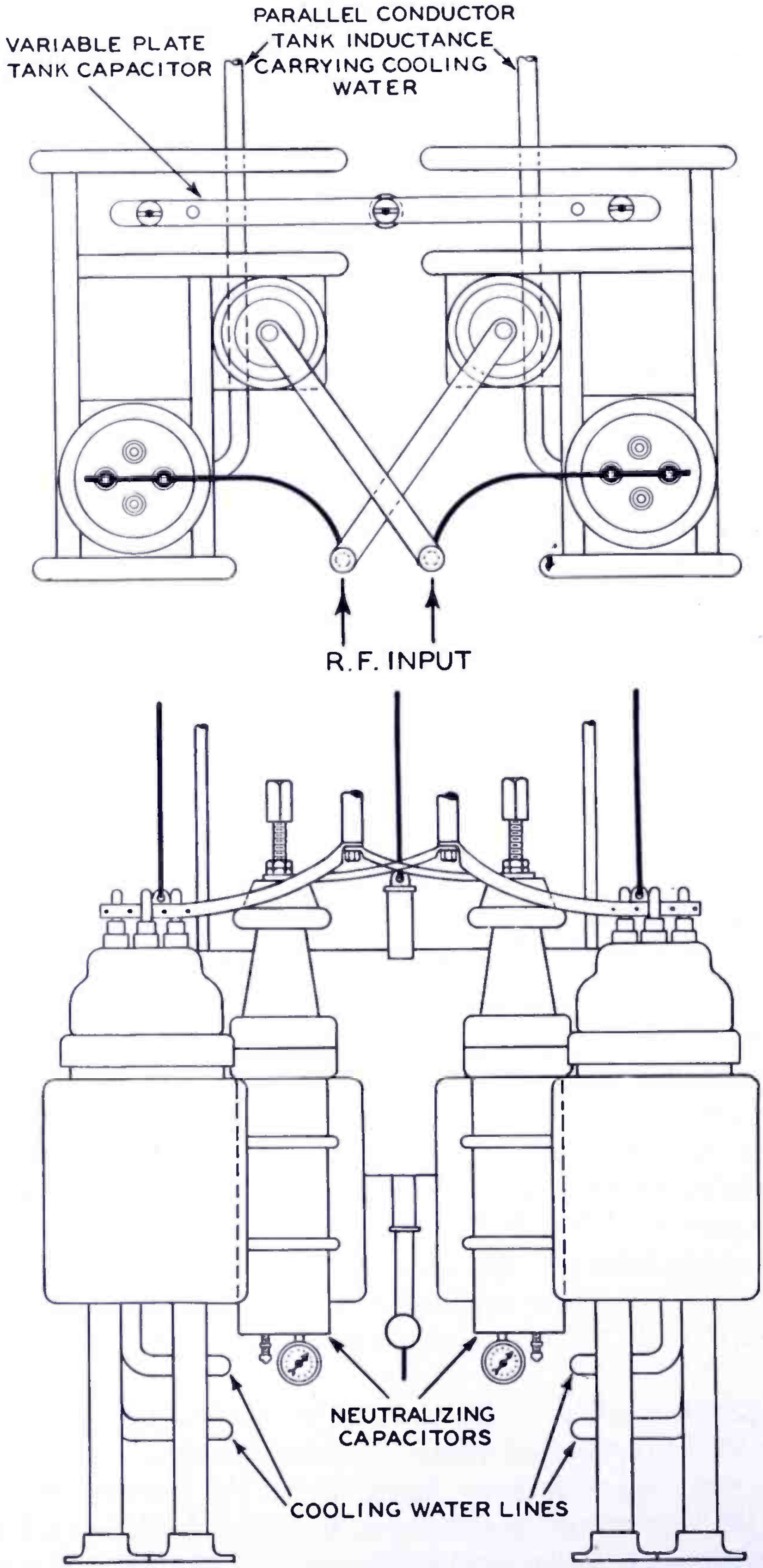
PARALLEL CONDUCTOR
TANK INDUCTANCE
CARRYING COOLING
WATER
VARIABLE PLATE
TANK CAPACITOR
R.F. INPUT
NEUTRALIZING
CAPACITORS
COOLING WATER LINES

Fig. 11

In Figure 11 there is shown an assembly drawing of the water-cooled RCA 880 tube mountings, the output tank circuit tuning capacitor and the neutralizing capacitor arrangement. The variable compressed-air neutralizing capacitors are used in a standard cross-neutralized circuit with excellent stability over the wide frequency range in which the transmitters are used. An air pressure of 150 pounds

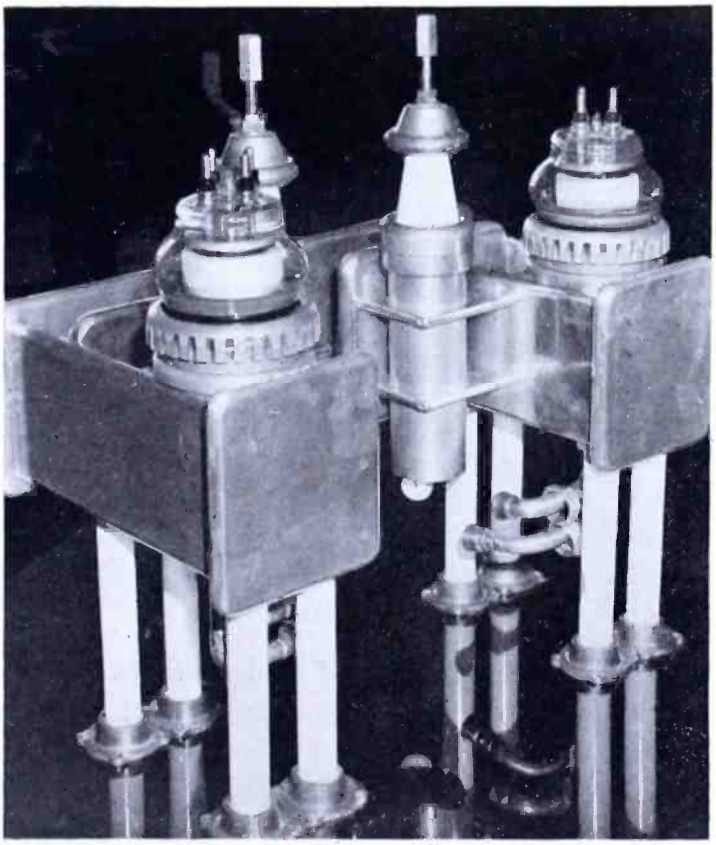

Fig. 12

per square inch has been found ample to provide air-dielectric strength suitable for this purpose. Two blower units are provided to supply sufficient air cooling for the grid and filament seals of each RCA 880 tube by means of three-inch bakelite ducts leading to the tops of the power amplifier tubes.

In the compact assembly shown in Figure 11, one-inch solid duralumin plates are used throughout. All corners and edges are carefully rounded and all surfaces highly polished. The variable, output, tank, tuning capacitor consists of a single movable grounded plate sliding vertically into the U-shaped chambers on each side of the assembly. This is illustrated in Figure 11. Figure 12 is a photograph of the assembly.

By this arrangement the capacitance from each side of the tank circuit to ground is perfectly balanced. Stray and minimum capacitance is reduced to a very low value. This is especially true with the movable plate raised to its extreme upward position where the capacity effects to it are negligible. A wide range of tuning of the parallel-conductor tank inductance output circuit is thus obtained by the sliding plate which is operated by means of gears from a knob on the front panel of the power amplifier unit. Counterweights are provided to reduce the effort required to vary the position of the movable plate.

The operation of high-powered, high-frequency amplifiers depends upon short direct r-f connections, lumped circuit elements, and simple construction if it is to be stable in operation. This is particularly true if the amplifier is to operate on two or more different frequencies. One of the difficulties in designing equipment of this kind is properly to locate the circuit elements with respect to each other, to minimize the length of the connections, and at the same time provide proper insulation for high-powered operation. One of the troublesome circuit elements is the neutralizing capacitor. In a push-pull balanced amplifier the neutralizing circuit must be insulated for d.c. plus r.f. but at the same time the capacitor itself should be as small and compact as possible. The use of compressed-air neutralizing capacitors offered an excellent solution. The capacitor is of a type developed and used by R.C.A. Communications, Inc. in high-powered telegraph transmitters and has been adapted to these broadcast units.

Figure 13 shows a photograph of one of these units. Figure 12 shows the manner in which it is mounted in the plate tank assembly. It may be seen that the RCA type 880 tube jacket is mounted adjacent to the compressed-air neutralizing capacitor, both being mechanically and electrically enclosed within the machined aluminum frame assembly.

These capacitors normally are kept at an air pressure of 150 pounds by means of a small motor-operated compressor. They have been tested under normal operating conditions on pressures as low as 60 pounds without any tendency to flash-over and have been tested, with normal pressure, at 90,000 volts without flashover. The capacitance is variable over a sufficiently wide range to insure optimum neutralization. The units are, of course, built to have the proper range of capacitance for the particular circuit in which they are to be used. The variation of capacitance is accomplished by turning the threaded end rod which varies the length of a copper bellows in the pressure chamber.

Distilled water for cooling the anodes of the RCA 880 tubes flows to and from each power amplifier tube through the parallel-conductor tank inductance, which consists of a one-inch square (outside dimension) bronze pipe and is illustrated in Figure 13. The tank inductance thus forms a part of the distilled-water system so that water is fed into the circuit through the porcelain water coils at the low r-f voltage point of the output tank with the result that r-f leakage to ground through the insulating water column is a minimum and at the same time the temperature of the tank inductance is maintained at a lower value.

Some of the problems encountered in building and operating dual high-power international broadcasting transmitters include those of eliminating harmonics and spurious frequency radiation. The harmonics usually can be eliminated by suitable filters located at the sending end of the transmission lines leading to the antenna. However, in addition to harmonic frequencies, there must be eliminated a wide variety of other spurious radiations which may be produced by a combination of two or more transmitters operating simultaneously on different frequencies, especially where open wire lines operate in the fields from the other antennas.

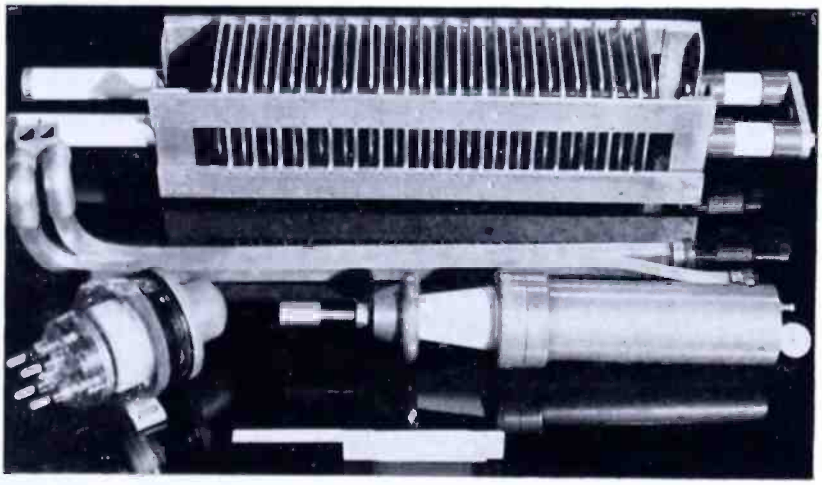

Fig. 13

Effectiveness of the Service

For many years NBC has conducted an effective International Broadcasting Service. With the outbreak and spread of the present world conflict and the continuous improvement in NBC radio facilities, the value of this service has multiplied in importance and scope. New thousands of letters and telegrams incessantly convey that fact with stirring impact. One cannot but be deeply impressed by those many communications from Europe which express despair and misery, nor can one read their pleas to continue and expand our efforts without more fully appreciating the high patriotic and humanitarian function which has come to be served by American short-wave licensees. International broadcasting as conducted by our country has become a powerful instrument which has earned the respect and confidence of foreign listeners through the truthfulness of its reporting and the character of its programs.

It has been the writer's privilege to be associated with radio for 26 years and with broadcasting for 20 years. Rarely has there been an opportunity in those eventful times for any form of communication to demonstrate such unique feats as are now becoming accepted commonplaces in international broadcasting. Barely 16 years ago the first rebroadcast from across the seas took place[1]. Scheduled rebroadcasts from the far corners of the earth have since become matters of but casual interest. But only in recent months has the bewildered victim of catastrophe, propaganda, and censorship so fully appreciated the modern miracle of radio which enables him to listen, perhaps secretly, to free stations thousands of miles overseas for frequent and authoritative reports of world events, at times taking place in his own country, frequently at points but a few blocks distant.

The following are excerpts from typical long letters from Europe:

"I am sending you this letter through a friend of mine to thank you for your excellent broadcasts of the French Hours. Whatever happens, keep on with them . . . At least may your radio stations remain with us, we beg of you. In our great sorrow, crushed as we are . . ."

"I follow your transmissions regularly. For months they have been helping me to bear up under the present situation and I cannot thank you enough for the moral support you give us".

"Living in a channel port you can't imagine the comfort my friends and I receive from hearing your transmissions . . . We shall be listening especially on April 13 unless a bomb shatters us."

"Literature you sent me before was destroyed during an air attack last June, from which, through a miracle, my family and I escaped although five bombs exploded surrounding our house."

The following is an example of the effectiveness of the WRCA-WNBI service in Central America. It is a telegram from Tegucigalpa, Honduras. "The broadcast was received in this capital perfectly. The speech has caused a sensation in our country as it did in the rest of Latin America. Rarely have I seen so many people gathered on the plazas where we have had our loudspeakers installed as I saw last night. We can guarantee that in our country there were many thousands of people who listened to the Spanish version of the presidential speech (Fireside Chat to the Americas) which we might add was very well read by your announcer."

NBC broadcast the President's speech direct to every one of the American republics and to Europe. Stations in the 20 Latin American republics and in Puerto Rico rebroadcast the Spanish and Portuguese versions which were given from Radio City simultaneously with the English broadcast from the White House. The speech was rebroadcast by the British Broadcasting Co. and Italian, German, and French translations were beamed to Europe on the regular times assigned to those languages.

The following is an example of the technical character of the 9670-kc signal received in Rio de Janeiro from the steerable antenna. It covers from 6.00 to 7.45 P.M. E.D.S.T. on May 15, 1941. Reception was on a common receiver with six feet of wire for an antenna.

	WRCA—9670 kc
Carrier strength	Very strong
Fading—Depth	Very slight to slight
Rate of fading	Very slow to slow
Interference from other stations	None
Static	Very weak
Background noise	Very low
Transmitted noise	None
Degree of modulation	High
Quality	Excellent
Overall rating	Excellent

ACKNOWLEDGMENTS

The combined cooperative efforts of many men are required in the planning and fulfillment of a long project. The writer wishes to gratefully acknowledge the unusual engineering ability and unselfish application of Mr. Carl G. Dietsch and Mr. William S. Duttera in planning and building the NBC radio facilities described herein. Mr. Dietsch has been assigned to the WRCA-WNBI project for many years and has directly superintended the construction of all of the transmitter and antenna facilities described. He has also invented and developed many improvements in methods and devices including the high-power tank-circuit assembly described. Mr. Duttera has from time to time collaborated in antenna development, design and adjustment, including particularly the steerable antenna system described.

In making major construction changes in equipment in daily operation countless vexing problems arise. Their solution has been simplified over a period of 17 years through the unfailing and sympathetic cooperation of Mr. D. N. Stair, Station Engineer at WJZ, WRCA and WNBI.

CALIBRATION OF MICROPHONES BY THE PRINCIPLES OF SIMILARITY AND RECIPROCITY

By

Harry F. Olson

RCA Manufacturing Company, Inc., Camden, New Jersey

Summary—For the application of the reciprocity principle to the calibration of a microphone, three transducers are used as follows: the microphone to be calibrated, a reversible microphone-loudspeaker, and a loudspeaker. The absolute sensitivity of a microphone may be obtained from two experiments involving the three transducers and electrical standards of voltage, current, and frequency. For the application of the principle of similarity, two velocity microphones are used having linear dimensions in the ratio 1:x. Under these conditions the configuration of the response characteristics will be the same, but shifted 1:x in frequency. The principles of similarity and reciprocity have been applied to the calibration of a commercial velocity microphone.

Introduction

THE field response[1] of a microphone at a particular frequency is given by the ratio e/p where e is the open-circuit voltage generated by the microphone and p is the sound pressure in dynes per square centimeter as measured in a plane-progressive sound wave prior to the introduction of the microphone. The pressure response is the same as the above except that p is the uniform sound pressure in dynes per square centimeter on the microphone diaphragm at a specified frequency.

There are four methods[1] in common use today to obtain primary microphone calibrations. These may be designated as: thermophone, electrostatic actuator, tourmaline crystal, and Rayleigh disc. The first two, in general, apply to pressure calibration and the last two to field calibration. The calibration of a microphone by any of the above methods requires considerable time and skill. Furthermore, in some cases the apparatus is quite delicate and costly. It is the purpose of this paper to describe the calibration of a microphone by means of the principles of similarity and reciprocity. By the use of these principles and relatively simple apparatus the absolute calibration of a microphone over a wide frequency range may be obtained.

[1] American Standards Association Sectional Committee Z-24, Report on Calibration of Microphones. *Jour. Acous. Soc. Amer.* Vol. 7, No. 4, p. 33, 1936. This report also lists the publications on calibration of microphones.

The Principle of Similarity

The free-field calibration of a microphone above 10,000 cycles presents some difficulties due to the relatively short wavelength. If a Rayleigh disc is used at 15,000 cycles the diameter of the disc should not exceed 3 millimeters, or 0.11 inch, to satisfy the condition that the diameter be less than one-eighth wavelength. The use of a disc of this small size is impractical. If a free-field calibration in terms of a pressure calibration is used, the dimensions of the microphone must be less than one-eighth wavelength to eliminate errors due to diffraction. The construction of a microphone of these dimensions with a usable output presents some difficulties. However, it is possible to obtain the frequency response over almost any range by employing the principle of similarity.

The principle of similarity[2] states that: if any system of connected particles and rigid bodies is given, it is possible to construct another system exactly similar to it, but on a different scale.

To find the relation between the various ratios involved let the linear dimensions of the model and pattern be in the ratio $x:1$, let the masses of the corresponding particles be in the ratio $y:1$, let the times between corresponding phases be in the ratio $z:1$, and let the forces be in the ratio $w:1$. Then the following required relation between the numbers w, x, y, and z may be written

$$w = xyz^2 \tag{1}$$

For the application of this principle it seems logical to employ the velocity microphone because of the simplicity of the vibrating system. At the high frequencies the vibrating system is a mass reactance save for the negligible radiation resistance. Further, in the case of the velocity microphone it is possible to design the structure so that the difference in pressure between the two sides of the ribbon is governed entirely by diffraction. This requirement is sufficient for the application of the principle of similarity in acoustics. Under these conditions the configuration of the response characteristics will be identical, save for a shift in frequency and sensitivity, if logarithmic scales are used for both the ordinates and the abscissas. If the ratio of the linear dimensions is $x:1$ the corresponding configurations of the two frequency characteristics will be displaced $1:x$ in frequency. The consideration of sensitivity may be neglected because the significant factor is the shape of the response-frequency characteristic.

To apply the principle of similarity, a velocity microphone was designed and built which exhibited very uniform response to beyond

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[2] Whittaker, "Analytical Dynamics," Cambridge University Press, 1927.

8000 cycles. This microphone is termed the pattern of Figure 1. Calibrations by means of the actuator, Rayleigh disc, pistonphone[1], and the method to be described are all fairly reliable and consistent in the low- and mid-frequency ranges. These calibrations were used to calibrate the pattern to 8000 cycles. Then a smaller microphone similar to the pattern, but with all the linear dimensions reduced by one-half was built. This microphone is termed the model of Figure 1. The constants of the two microphones were chosen so that the conditions

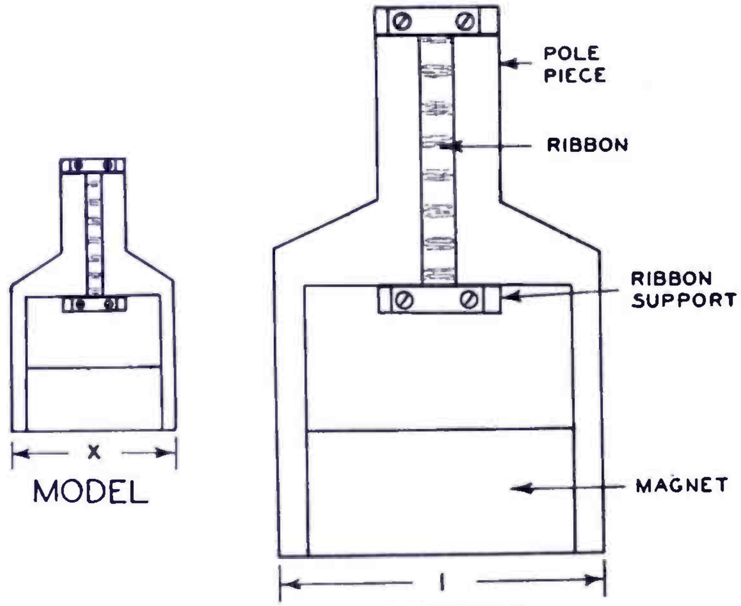

Fig. 1—Two similar velocity microphones with linear dimensions in the ratio of x:1 termed the model and pattern, respectively.

of Equation 1 were satisfied for the pattern and the model. The microphones were designed so that the actuating forces are governed by diffraction phenomena. Under these conditions the shape of the response-frequency characteristic of the small microphone is the same as that of the large microphone but advanced one octave in frequency.

PRINCIPLE OF RECIPROCITY

The acoustical reciprocity theorem was originally enunciated by Helmholtz and Rayleigh[3]. Ballantine[4] has established reciprocity theorems for mechano-acoustic, electro-mechano, and electro-mechano-acoustic systems. Ballantine has also carried out a generalized dis-

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[1] Loc. Cit.

[3] Rayleigh, "Theory of Sound," MacMillan Co. Vol. I, p. 145.

[4] Ballantine, S., *Proc. I.R.E.*, Vol. 17, No. 6, p. 929, 1929. Later, similar theoretical conclusions were reached by R. K. Cook, *Jour. of Research, Nat. Bur. of Standards*, Vol. 25, No. 5, p. 489, 1940 and by W. R. McLean, *Jour. Acous. Soc. Amer.*, Vol. 12, No. 1, p. 140, 1940. Cook applied these relations to the calibration of condenser and crystal microphones.

cussion to show that a microphone may be calibrated in terms of electrical standards by the use of the extended reciprocity relations. This method, applied to a specific problem, will now be outlined.

For the application of the reciprocity principle to the calibration of a microphone, three transducers are used as follows: the microphone, M, to be calibrated, a reversible microphone-loudspeaker S_1, and a loudspeaker S_2. For the reversible microphone-loudspeaker it is convenient to use a small back-enclosed loudspeaker.

The first experiment is shown schematically in Figure 2. An alternating current is fed to the loudspeaker S_2. A sound pressure p_1 is produced at a distance d. Let the open-circuit voltage, in abvolts, of

Fig. 2—Arrangement for obtaining the ratio of the open-circuit voltages e_M and e_S of the microphone M and the microphone-loudspeaker S_1 when actuated by the same sound pressure p_1 produced by the loudspeaker S_2.

S_1 used as a microphone be designated as e_S and the output of the microphone M be designated as e_M. Let K_S = abvolts per dyne per square centimeter of S_1, and, K_M = abvolts per dyne per square centimeter of M_1. Since the sound pressure p_1, in dynes per square centimeter, is the same for S_1 and M_1, it is evident that

$$p_1 = \frac{e_S}{K_S} = \frac{e_M}{K_M} \tag{2}$$

The voltage output, in abvolts, of the microphone-loud speaker S_1 used as a microphone is

$$e_S = Bl\dot{x}_1 \tag{3}$$

where B = flux density in the air gap, in gausses,
l = length of the conductor, in centimeters, and
$\dot{x}$ = velocity of the voice coil, in centimeters per second.

The velocity, in centimeters per second, of the vibrating system of S_1 as a microphone is

$$\dot{x}_1 = \frac{p_1 A}{Z_M} \tag{4}$$

where p = actuating sound pressure, in dynes per square centimeter,
A = area of the diaphragm, in square centimeters, and
Z_M = mechanical impedance of the vibrating system, in mechanical ohms.

From Equations 2, 3, and 4

$$\frac{e_S}{p_1} = \frac{BlA}{Z_M} = K_S \tag{5}$$

The second experiment is shown in Figure 3. The velocity, in centimeters per second, of the diaphragm and voice coil of S_1 for a current i, in abamperes, in the voice coil is

$$\dot{x} = \frac{Bli}{Z_M} \tag{6}$$

The pressure[5] p at M, in dynes per square centimeter, at a distance d, in centimeters, produced by S_1 in the range where the dimensions are small compared to the wavelength is

$$p = \frac{\rho c k A \dot{x}}{4\pi d} \tag{7}$$

where A = area of the diaphragm, in square centimeters,
$\dot{x}$ = velocity of the diaphragm, in centimeters per second
ρ = density of air, in grams per cubic centimeter and
$k = 2\pi/\lambda$
λ = wavelength, in centimeters, and
c = velocity of sound.

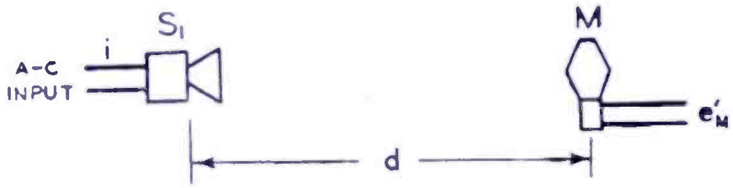

Fig. 3—Arrangement for obtaining the open-circuit voltage e'_M of the microphone when driven by a pressure p produced by the microphone-loudspeaker actuated by a current i.

From Equations 6 and 7,

$$p = \frac{\rho c k A B l i}{4\pi d\, Z_M} \tag{8}$$

From Equations 5 and 8,

$$p = \frac{\rho c k i K_S}{4\pi d} = \frac{r i K_S}{2 d \lambda} \tag{9}$$

where $r = \rho c$

[5] Olson, "Elements of Acoustical Engineering," D. Van Nostrand Co., 1940, p. 19.

The pressure p in dynes per square centimeter, at M_1 in terms of the constant K_M and the open-circuit voltage e'_M, in abvolts, is

$$p = \frac{e'_M}{K_M} \tag{10}$$

From Equations 9 and 10

$$\frac{e'_M}{K_M} = \frac{riK_S}{2d\lambda} \tag{11}$$

From Equation 2

$$K_M = \frac{e_M}{e_S} K_S \tag{12}$$

When K_S is eliminated from Equations 11 and 12, the response of the microphone M in abvolts per dyne per square centimeter is

$$K_M = \sqrt{\frac{2d\lambda e_M e'_M}{rie_S}}$$

where e_S and e_M are obtained from setup of Figure 2 and e'_M and i are obtained from setup of Figure 3. The units are as follows: Voltages

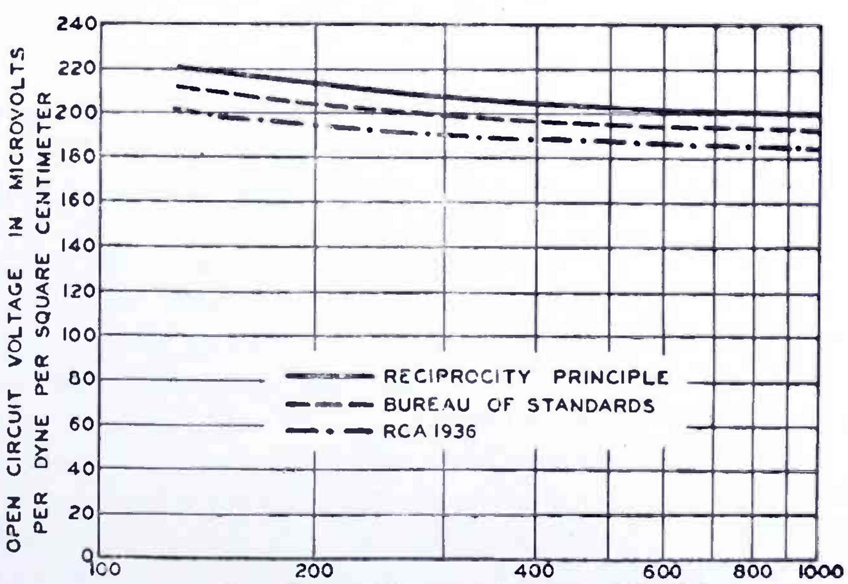

Fig. 4—The open-circuit voltage response vs. frequency characteristics at the 250-ohm output terminals of a commercial-type velocity microphone for a free field pressure of one dyne per square centimeter.

in abvolts, currents in abamperes, distances in centimeters, wavelengths in centimeters and $r = \rho c = 41.5$.

Experimental Results

A commercial-type velocity microphone was calibrated by the principle of reciprocity as outlined in the preceding section. The results are shown in Figure 4. The microphone was also calibrated by means of

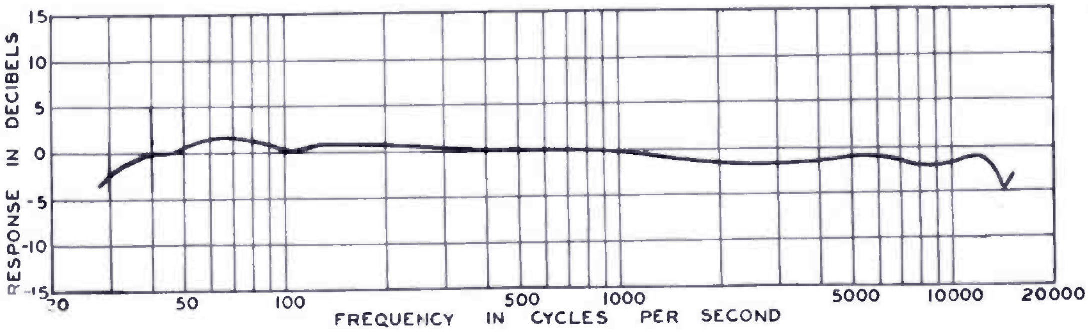

Fig. 5—The open-circuit voltage response vs. frequency characteristic at the 250-ohm output terminals of a commercial-type velocity microphone obtained by the principles of reciprocity and similarity. 0 db = 200 microvolts per dyne per square centimeter.

two secondary standards; one calibrated six years ago with a Rayleigh disc, pistonphone, and actuator by Mr. M. L. Douthit of the RCA Manufacturing Company and the other calibrated by the Bureau of Standards. The maximum deviation between the three methods is ⅔ decibel or 8 per cent.

Using the above data and the two microphones of Figure 1 the remainder of the range was obtained by the principle of similarity. The response frequency characteristic of the commercial-type microphone for the frequency range 30 to 15,000 cycles is shown in Figure 5.

RECENT DEVELOPMENTS IN PHOTOTUBES

By

R. B. Janes and A. M. Glover

RCA Manufacturing Co., Inc., Harrison, New Jersey

Summary—Developments of the last two years in photoemissive tubes are discussed. These include the introduction of a surface highly sensitive to blue-rich light and its incorporation in a phototube and an electrostatically focused photoelectric multiplier. The phototube has a sensitivity of about 45 microamperes per lumen to incandescent light of color temperature 2870°K even though the tube has no infra-red sensitivity. The photoelectric multiplier incorporates the same surface, but because of the use of nine stages of secondary emission multiplication has the tremendous sensitivity of 0.6 ampere per lumen to the same light source.

FOR many years after the discovery of the photoelectric effect by Hertz in 1887, its usefulness was appreciated chiefly by laboratory workers. Laboratory-made phototubes were used, in most cases, for measurements in optical and chemical research. However, the small size of the currents obtained required the use of delicate electrometers, thus making measurements tedious and exacting. With the development of reliable amplifying tubes in which the input impedance was relatively high, new fields of usefulness for phototubes were opened. By the use of amplifying tubes, output currents of the order of milliamperes or more could be delivered to rugged meters or other output devices. The first large-scale commercial application of phototubes was for sound motion pictures. In most sound motion pictures the sound track, as the record of the speech or music accompanying the picture is called, is photographed on the edge of the film. The light from a suitable source, generally an incandescent lamp, is projected upon this track and variations in the light transmitted through it are registered by a phototube. Since the radiation from an incandescent lamp, even when operated at a high temperature, is largely in the red and infra-red regions of the spectrum, efforts were made to develop a phototube surface with a high sensitivity in this region of the spectrum. In 1928, this work was successful. The spectral sensitivity curve for the new photosurface is shown in Figure 1. It is designated as photosurface S1. The energy distribution from a tungsten lamp operating at a color temperature of 2870°K is also shown in Figure 1. It is apparent that a phototube with an S1 surface is a very satisfactory device for use with an incandescent lamp. Manufacturing improvements over a period of several years permitted the introduction of a second surface designated as S2. This surface is similar to the S1, but has a higher infra-red response.

The reliable performance of phototubes in sound motion pictures has encouraged their use in many industrial applications. In some of these the output of the light source is largely in the green or blue. Also, there are many applications in which the phototube is used to replace the eye for color matching. Hence, a spectral sensitivity having high green and blue response and more nearly approaching that of the eye is needed. The desired spectral-response characteristic can be obtained from an S2 surface by proper filtering, but the resultant sensitivity is very small. The S3 surface, for which the spectral sensitivity is shown in Figure 2, was developed to meet these applications. Although its overall sensitivity to an incandescent lamp is only 5 or 10 microamperes per lumen as compared to 20 to 30 microamperes per lumen for the S2 surface, its sensitivity to blue and green light is several times larger than that of the S2 surface. The problem of filtering to match the response of the eye is, therefore, considerably simplified. The sensitivity curve of the eye is also shown in Figure 2 for comparison.

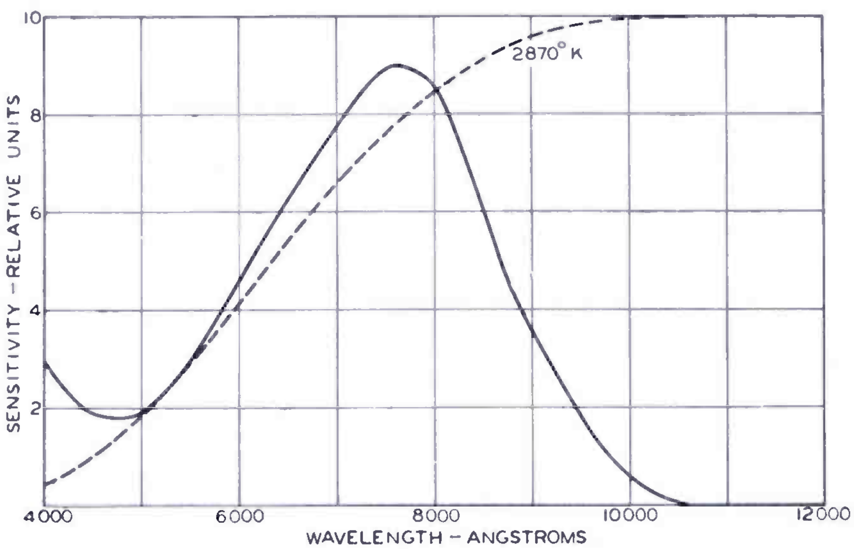

Fig. 1—Spectral distribution of the S1 surface. Spectral energy distribution of an incandescent lamp at 2870°K.

In the last year, two phototubes employing a new surface called S4 have been introduced.[1] These are the types RCA-929 and RCA-931. The 929 is a vacuum phototube while the 931 is a multiplier phototube in which both the photocathode and the secondary-emissive stages employ the S4 surface. The 929 will be considered first.

The 929 phototube is different in many respects from tubes with the S2 and S3 surfaces. The spectral response of the three surfaces are shown together in Figure 3. It will be observed that the spectral response of the 929 has no infra-red sensitivity, but is confined to the

[1] A. M. Glover and R. B. Janes, *Electronics* 13, 26, August 1940. "A New High Sensitivity Photosurface".

visible and ultra-violet region of the spectrum. The long wavelength limit is at about 6300 angstroms. This characteristic is especially useful for applications such as flame-control measurements, in which it is desirable for the tube to respond to the visible part of the flame spectrum only and not to the background illumination. In the ultra-violet end of the spectrum the sensitivity of the 929 drops to very low values near 3000 angstroms, because of the high optical absorption of the soft-glass envelope. The sensitivity of the surface itself extends much farther into the ultra-violet and is high to at least 2537 angstroms. The response in the ultra-violet is not directly proportional to

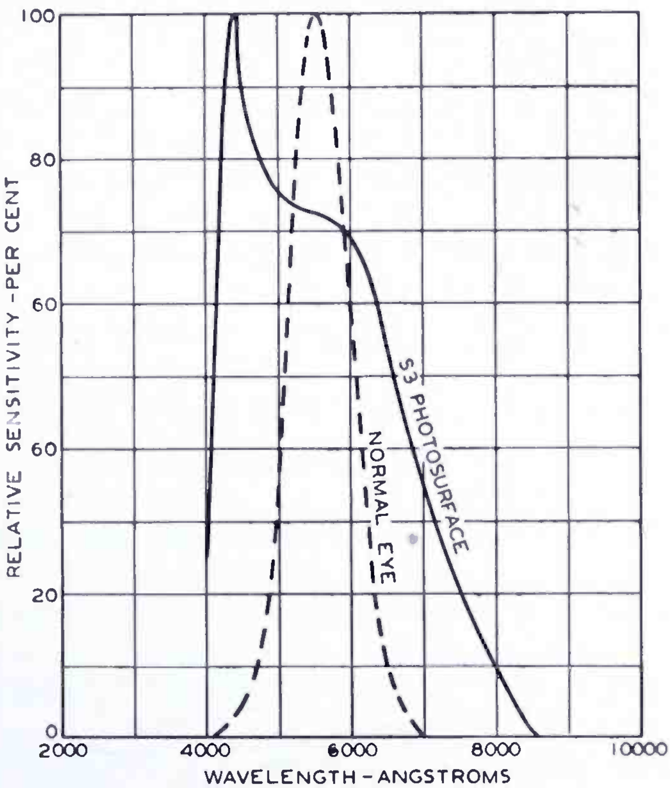

Fig. 2—Spectral distribution of the S3 surface. Spectral sensitivity of the human eye.

the sensitivity in the visible spectrum, since a small variation in the shape of the spectral-sensitivity curve in the neighborhood of the threshold will, of course, affect the visible sensitivity without changing the ultra-violet sensitivity.

The high sensitivity of the 929 in the ultra-violet and the blue region of the visible spectrum makes the tube very useful for measuring the light output of mercury arcs, fluorescent materials, or the sun, since such sources radiate chiefly in this region. For these and similar light sources, the 929 has a response several times that of other surfaces.

Although the response of the 929 is greatest in the blue and ultra-violet, its sensitivity is also high for light from tungsten filaments. In fact, while the S2 surface has a sensitivity to incandescent illumination of color temperature 2870°K of 20 to 30 microamperes per lumen, the S4 surface has a sensitivity of 30 to 60 microamperes per lumen. Thus, the 929 can be used to replace phototubes with S1 or S2 surfaces where they are used with an incandescent light source such as in sound reproduction equipment. However, if the temperature of the incandescent source is lowered, the response of the 929 falls more rapidly than that for a tube with an S1 or S2 surface. The customary practice of reduc-

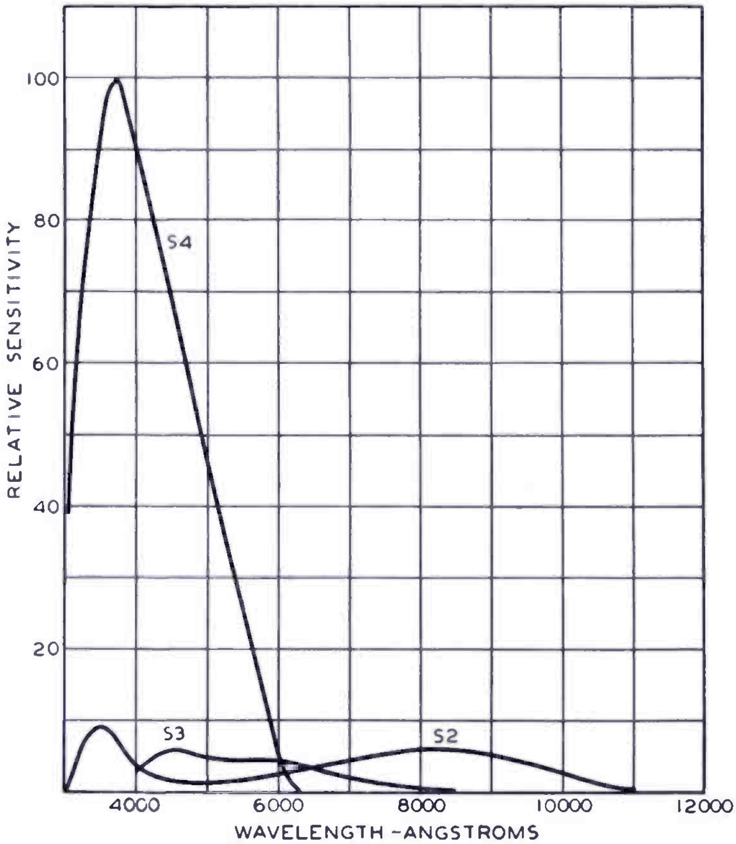

Fig. 3—Spectral distribution of the S2, S3, S4 surfaces.

ing the lamp voltage to conserve lamp life can not be utilized to the same extent with the 929 as with tubes having S1, S2, or S3 surfaces without the loss of considerable sensitivity. However, when the life of the lamp is not an important factor a large increase in sensitivity can be attained with the 929 by increasing the color temperature of the lamp.

A consideration of the relative sensitivity of the 929 and of tubes with S2 and S3 surfaces discloses the following facts. For incandescent light at a color temperature of 2870°K the 929 has nearly twice the sensitivity of the best S1 and S2 surfaces and about seven or eight times that of the S3. For daylight illumination, the sensitivity of the

929 is at least 120 microamperes per lumen, a value which is many times that of the other surfaces. For a high-pressure mercury arc, the response is fifty times as great as for the S2 surface. It is interesting to calculate for these surfaces the number of electrons emitted per incident photon, i.e. the quantum efficiency. For the S2 surface with a peak at about 8000 angstroms, the number is about one for each 100 incident photons or an efficiency of 1 per cent. For the S4 surface at the peak at 3750 angstroms, the efficiency for an average tube is about 12 per cent, a range in the efficiency from 8 to 20 per cent being commonly obtained. Experiments have shown that for the S1 and S2 surfaces the sensitivity resides largely in a very thin layer on the surface. The quantum efficiency of this layer is very high, of the order of 10 or 20 per cent, but the amount of light absorbed in the sensitive layer is a small part of the total light. The reason for the high sensitivity of the S4 surface is that a great deal more light is absorbed in the sensitive layer than is the case for the S2 surface.

The high sensitivity of the 929 to nearly all sources of light is not its only advantageous characteristic. Its dark resistance (that is, its resistance when no light is permitted to fall on the cathode) is also very high. The high value of dark resistance is equivalent to an increase in sensitivity since it permits the use of a large load impedance with a consequent increased voltage sensitivity. The dark resistance is determined by several factors. The most important is the leakage across the glass press and across the base of the tube. This leakage is low in the 929 because the characteristic of the surface is such that no excess of alkali metal is left beyond that which is required to sensitize the surface. In the average 929 the dark resistance is of the order of 50,000 to 500,000 megohms and is limited by leakage in the base. A second factor, which adds to the dark current, is thermionic emission, which is the release of electrons from the photocathode as a result of its operating temperature. In tubes with an S2 surface this current may be as large as 10^{-10} to 10^{-11} amperes per square centimeter of cathode area at room temperature. Tubes employing this surface are frequently immersed in liquid air to reduce the thermionic emission. In the S4 surface the thermionic emission is of the order of 10^{-13} to 10^{-14} amperes per square centimeter. The reduction is very considerable and is especially valuable when very small amounts of light are being measured. A third factor is photosensitivity to unwanted radiation either visible or invisible. Ordinarily, the most difficult stray radiation to eliminate is infra-red radiation which may come from the walls of the phototube shield. The absence of sensitivity of the S4 surface to such radiation is, therefore, an asset for measurement work.

The 929 is of the high-vacuum type. Its photosensitivity to tungsten light is higher than that of high-vacuum tubes with S2 surfaces although its output for such light is lower than for a gas-filled tube with an S2 surface such as the 918. Gas filling of the 929 to date has proved undesirable because of a loss in sensitivity with time which takes place when the tube is illuminated with sufficient light to cause a drain of more than a microampere or so from the cathode surface. However, its high sensitivity makes gas filling of minor importance. Because the 929 has such a high intrinsic photosensitivity, its signal-to-noise ratio is better than that of the S2 surface. As a result, the subsequent amplification can be raised to offset partially the greater output of a gas-filled tube such as the 918. Measurements under actual operating conditions verify this conclusion. Because the 929 is a high-vacuum type, an improvement in frequency response also results from the absence of the time delay which would be caused by the transit time of the ions and metastable atoms of the gas. These in a gas-filled type drift relatively slowly to the cathode and, thus, cause a decrease in sensitivity as the frequency of an audio signal is increased. The absence of the gas also provides a more linear output characteristic as a function of the light intensity with resultant decrease in amplitude distortion.

Another valuable characteristic of the 929 is its extremely steady output under constant illumination. Readings taken over periods of many hundreds of hours show that the variations in output are of the order of a few per cent. This stability makes the 929 especially useful for the measurement of small changes in light. Fairly large amounts of current can be drawn from the tube. Over long periods of time the current is limited to about ten microamperes per square centimeter of cathode surface but for short periods of operation this value can be exceeded. However, this type phototube should not be operated at as high an ambient temperature as other types. Its ambient temperature must not exceed 50°C. Figure 4 shows curves of the anode current versus anode voltage of the 929 for several different values of light flux. It is apparent that the output of the tube is essentially independent of anode voltage over a very wide range, and is, therefore, independent of line-voltage fluctuations. This fact, together with the high internal resistance of the tube, permits the use of a load having high resistance. It should be noted that a high-vacuum phototube is especially suitable for use directly from an alternating-current supply, since the tube may be operated at the high peak voltage of the supply. The tube serves as its own rectifier to give a unidirectional output which may be filtered if desired.

The spectral response characteristic and the high sensitivity of the tube open up many fields for its application. Because of its high sensitivity to daylight, it is possible to design around it a sensitive photographic exposure meter for use at light levels at which present instruments are insensitive. A rugged inexpensive output meter can be used because the high impedance of the tube permits vacuum-tube amplification. The 929, because of its inherent stability and high sensitivity in the blue, is also especially adaptable for photographic densitometer equipment. The lack of infra-red sensitivity and the rapid rise of response in the yellow makes it applicable to the shutting-off of incandescent devices after they have reached a certain color temperature. Since the response of the tube changes very rapidly with color temperature, a small change in color temperature may be detected. This type

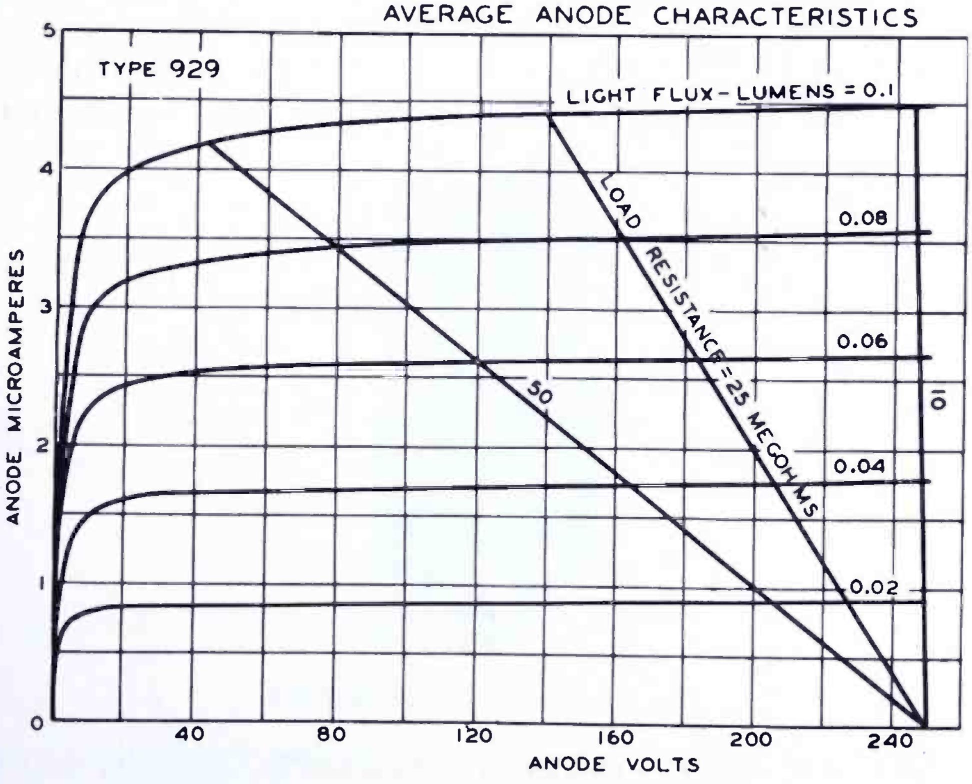

Fig. 4—Plate family of RCA-929.

of response also makes it possible to obtain a good sorting of red from yellow and from blue objects. In three-color printing processes, the response may be used to advantage either with a 929 phototube alone or in conjunction with another phototube with a red-sensitive S2 surface. Also, just as phototubes with S2 surfaces are used with infra-red radiation to produce a concealed automatic burglar alarm, the S4 surface can be so used with ultra-violet radiation.

There are many applications for the 929 in scientific fields. For astronomical purposes the S4 surface is exceptional. The very high sensitivity in the blue and near ultra-violet closely matches the maximum of the spectral output of the stars. Also, the very small dark current gives a very favorable signal-to-noise ratio. An increase in sensitivity of the phototube is equivalent to an increase in the light-gathering power of an astronomical telescope, or, specifically, to an expensive increase in the diameter of the light-collecting system. There are also many applications in colorimetry, organic analysis, and biochemical analysis where this new tube will prove useful. One other special use is for the measurement of X-ray intensities. In this application, the tube is exposed to the light from a calcium-tungstate screen which fluoresces under X-ray bombardment. The spectral response of the fluorescence of calcium tungstate is a maximum at a wavelength at which the 929 is most sensitive. The combination of a calcium-tungstate screen and a 929 tube, therefore, makes a very sensitive detector of X-rays.

Figure 5 is a photograph of the 929. An intermediate five-pin octal

Fig. 5—Photograph of the 929.

base is used to provide a short overall tube length. Recently, the type RCA-930 has been introduced. This tube utilizes the S2 photosurface; its construction and basing are identical with that of the 929. Hence, these phototubes with S2 and S4 surfaces are readily interchangeable.

The type RCA-931 is a photomultiplier tube in which the photocurrent produced at a light-sensitive cathode is multiplied many times by secondary emission occurring at successive dynodes within the tube. It employs complete electrostatic focusing as used in the development tube described by Rajchman and Snyder.[2] A photograph of the tube and its internal structure are shown in Figures 6 and 7. As can be seen, the design is extremely compact with all leads brought out to the base pins. In fact, the tube resembles very closely in appearance the 929 and 930 phototubes and is nearly as small.

[2] J. A. Rajchman and R. L. Snyder, *Electronics* 13, 20, December 1940. "An Electrically Focused Multiplier Phototube".

Although the system of focusing is that of Rajchman and Snyder, the 931 differs in that the S4 surface has been employed for both the photoemissive and secondary-emissive surfaces. The photosurface has the high blue sensitivity and the very small dark current which is associated with this surface. The factors of high sensitivity and low dark current are particularly necessary in a multiplier in order to take advantage of its high signal-to-noise ratio at very low light levels. The blue sensitivity also is very desirable in many applications as already mentioned, particularly for spectroscopy and astronomical measurements, where the multiplier phototubes will find many uses.

In the secondary-emissive stages the use of the S4 surface is also advantageous. An important advantage is the very high stable second-

Fig. 6—Photograph of the 931.

ary emission which the surface gives at low voltages. At 100 volts per stage, the secondary emission is about 3 or 4 to 1 while at 150 volts per stage, it is 4 or 5 to 1. The use of this surface also considerably simplifies the processing of the tube. In tubes using other surfaces, particularly the S2, for both the photo and secondary-emissive surfaces, considerable difficulty has been experienced in obtaining good secondary emission and good photoemission in the same tube. With the new S4 surface, this difficulty has been overcome.

In addition to the use of the S4 surface, the 931 represents much effort in attaining a design adaptable to economical manufacturing methods. Both the bulb and the stem press are of soft glass. The bulb is a standard T9 bulb such as is used in receiving tubes and other phototubes while the stem which contains eleven leads is of the button construction used in many cathode-ray tubes. Figure 6 is an illustration of the tube. As Figure 7 shows, the 931 consists of a photocathode (element Number 0), nine stages of secondary-emission multiplication (electrodes Numbered 1 to 9) and a collector (element Number 10). Such a large number of stages is necessary to obtain a signal-to-noise ratio that will be well above that of a regular phototube and amplifier.

The sensitivity obtainable with the 931 multiplier depends on the photosensitivity of the photocathode and the secondary emission of the multiplier stages. In average tubes a photoemission of 15 microamperes per lumen is obtained, a value about one-third of that of an average 929. The average gain per stage is about 3.5 at 100 volts per stage and about 4.0 at 125 volts per stage. Thus, at 100 volts per stage a gain of 60,000 can be expected while at 125 volts per stage the gain will be about 230,000. Because the overall gain varies so rapidly with

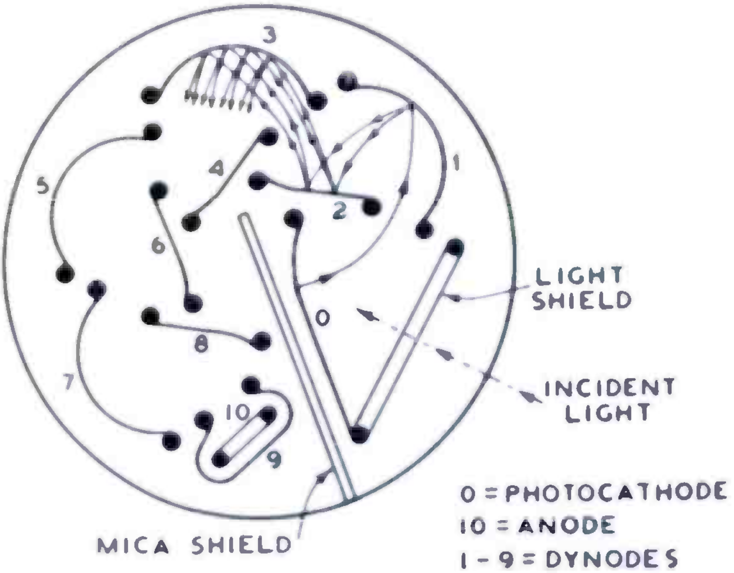

Fig. 7—Internal structure of the 931.

a small change in secondary emission of the multiplier surfaces, variations in gain must be expected from tube to tube. However, the change in sensitivity with supply voltage may be used as a gain control to compensate for this variation.

Because of the large number of stages, a high-voltage supply of 1000 to 1250 volts is necessary for the operation of the multiplier. An increase in the voltage from about 1000 to 1250 volts gives a four-fold increase in output. The gain of the tube can, therefore, be readily controlled by changing the supply voltage, but this rapid change makes necessary good regulation and constancy of the supply voltage. If the output of the tube is to be held constant to 1 or 2 per cent, the voltage supply must be constant to about 0.1 per cent.

An average sensitivity of 0.6 ampere per lumen to a light source at 2870°K is obtained for a 1000-volt supply. A background current of less than 0.25 microampere at this voltage may be expected. The background current increases faster than linearly with the voltage due to the contribution of field emission, and bombardment of the photocathode by positive ions of the residual gas which have evaded the trapping screen placed between the collector and the photocathode. At the recommended maximum output of 2.5 milliamperes, several thousand hours of life may be expected.

The response of the tube is linear to illumination over a range from zero illumination up to values of illumination which cause the output stage to saturate due to space-charge limitations. This value is above

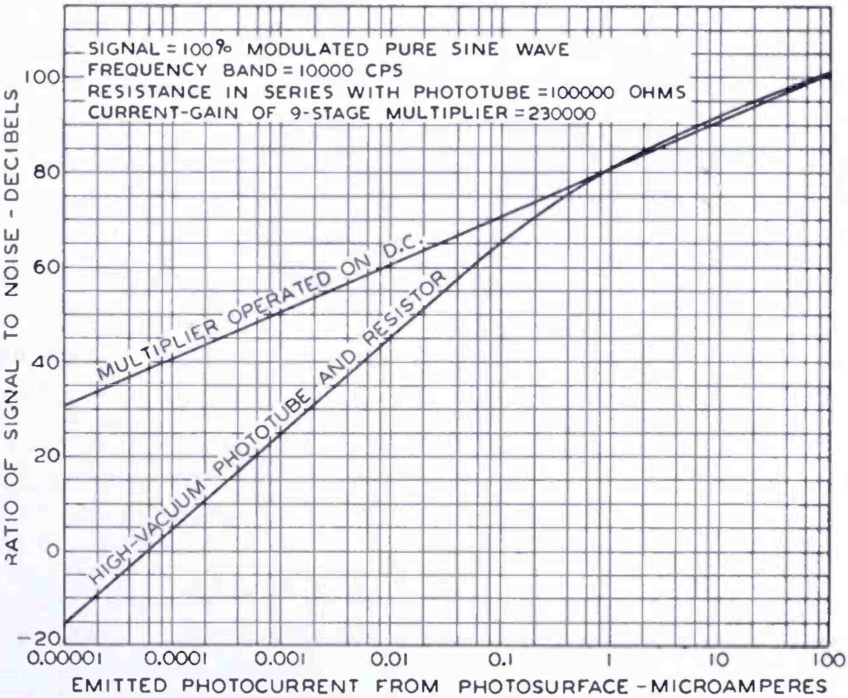

Fig. 8—Signal-to-noise ratio of a multiplier phototube.

the recommended maximum operating current. Also the output characteristic of the tube is flat, a feature which makes it possible to use a high load resistor giving an output voltage swing which is a large fraction of the total available voltage, yet with linear response to a variable input illumination.

Probably the most important characteristic of a phototube amplifier is its signal-to-noise ratio. In a phototube-amplifier circuit, noise arises from the shot noise of the photoemission and from the thermal noise of the amplifier coupling resistor. In the 931 multiplier with its very high gain, the shot noise alone is significant. In the 929 vacuum-type phototube with an amplifier both must be taken into consideration. In Figure 8 are reproduced curves taken from Rajchman and Snyder's paper for the signal-to-noise ratio of a multiplier and a vacuum-type phototube. The ratio of signal-to-noise in decibels is plotted against the emitted photocurrent in microamperes. For the same emitted photocurrent the curves show that the multiplier at low currents has a much better ratio than a vacuum phototube and amplifier. However, it should be remembered that the multiplier has only about one-third the cathode photoemission of a 929. Therefore, for a given light intensity the photocurrent from a 929 will be three times more than the photocurrent of the 931. Taking this ratio into consideration, we find that a light level which will produce 0.00001 microampere photocurrent in the multiplier (or about 2.3 microampere output if the gain is 230,000) the multiplier will have a signal-to-noise ratio about 35 decibels above that of the 929. For a light level that will give about 0.0001 microampere photocurrent in the multiplier (or 23 microamperes output if the gain is 230,000) the multiplier ratio is about 25 decibels above that of the phototube. Only when the light is large enough to produce a photocurrent of about 0.1 microampere in the multiplier are the ratios about equal.

The dark current of the multiplier determines how small a light level can be measured. As mentioned before, the use of the S4 surface keeps the dark current to a very small value. In the 931 the upper limit of dark current is 0.25 microampere measured at the output with 100 volts per stage. The variations in the dark current are, of course, smaller than this figure. Consequently, for a tube with a gain of 50,000, an amount of incandescent light of the order of 10^{-6} to 10^{-7} lumens can be detected.

The multiplier may be operated with a very simple power supply by the use of an alternating instead of direct supply. The frequency of the supply should be at least ten times as high as the frequency of the variations of light intensity. If measurements of a slowly varying light source are to be made, a 60-cycle supply is satisfactory. Although the signal-to-noise ratio for a-c operation is not as high as that for d-c operation, the simplicity of the a-c circuit makes it very useful for operating a relay directly.

The ability of the multiplier to detect and measure extremely small amounts of light is, of course, its greatest advantage. Moreover, its small size and economical construction will permit its installation in many places where a phototube is used with an amplifier. This includes sound-movie equipment and the whole field of industrial control.

AN OMNIDIRECTIONAL RADIO-RANGE SYSTEM

By

David G. C. Luck

Research Laboratories, RCA Manufacturing Co., Inc., Camden, New Jersey

PART I — PRINCIPLES OF OPERATION

Summary—*Radio navigation may be done with direction-finding receivers on mobile craft, with fixed direction finders on the ground or with directional beacon transmitters on the ground. Each method has its unique merits and faults, but the last seems especially suited for aircraft guidance in the United States and has, in the form of four-course radio "range" beacons, rendered outstanding service. The disadvantages of limited choice of courses and of difficulty in definitely determining on which course a craft may be, inherent in the present four-course ranges, may be avoided by rotating a transmitted radio beam and timing its passage over the receiving craft, to determine uniquely the bearing of that craft from the known location of the beacon transmitter.*

A rotating beam, of figure-eight shape, may be produced without mechanical motion by setting up two fixed antenna systems, having figure-eight directivity, at right angles and feeding them with radio-frequency signals modulated at the desired rotation frequency, the modulation of the separate supplies to the two crossed antennas being in phase quadrature. Unmodulated carrier to resolve the ambiguity of the figure-eight beam, by changing its shape to a limaçon, is radiated from a non-directive antenna, and a timing reference is provided by interrupting all transmission momentarily just as the beam points north.

The audio output from a receiver tuned to this beacon comprises a sine wave produced by the sweep of the beam and a train of impulses produced by the reference keying. The sine component is filtered, split in phase and used to drive a cathode-ray beam in a circle, in step with the rotation of the transmitted beam. The impulses are used to slow up the beam electrons momentarily, marking the swept circle with an outward jog and so indicating receiver bearing directly. The impulses also actuate a zero-center meter, while the sine wave renders this meter insensitive at a certain moment of the cycle and oppositely sensitive just before and just after that moment. By adjusting the sine wave phase, the meter may be centered when the receiver is on any desired bearing, and thereafter will indicate any departure from that bearing. A special broadcast transmission may be used to check adjustments of receiving indicators.

Certain conditions as to modulation phases and amplitudes, antenna-current phases and amplitudes, antenna geometry and cathode-ray indicator voltage phases, amplitudes and tube geometry must be fulfilled if accurate bearings are to be obtained. Study of these conditions shows all adjustment tolerances to be of reasonable magnitude, though considerable care in antenna construction is necessary to insure adequate symmetry of antenna-current phase.

1—Introduction

A—*Navigation*

EVERY form of transportation is faced with the problems of navigation, which are to keep track of and to plan the travel of mobile units. Navigation becomes increasingly difficult and important as freedom of mobility increases. Aircraft, therefore, provide the ultimate in navigational problems to date. The difficulty of solution is enhanced by the high speeds attained and by the preoccupation of a necessarily minimal operating personnel. The last factor makes convenience in use a paramount virtue in any proposed solution of problems of aircraft navigation.

Three methods of navigation have come into use. When possible, direct visual observation of the earth's surface with its landmarks is a very convenient and valuable method, direct celestial observation following as a less convenient, but sometimes even more valuable variant. When visual observation is impossible, blind calculation or "dead reckoning" from all available position, speed, and drift data furnishes a second method. This is often convenient, but can permit the accumulation of serious amounts of unsuspected error between checks by direct observation. The third method of navigation, radio direction-finding, is unique, both in its ability to supply direct navigational data when there is no visibility at all and in the convenience of use of its recent refined embodiments.

B—*Radio Direction-Finding*

Radio direction-finding also has come into use in three main forms, each with its peculiar advantages and disadvantages.[1] In all cases, radio direction-finding equipment consists primarily of a precision directional antenna system. The remaining apparatus is subsidiary; its specific form determines the convenience and completeness with which the properties of the antenna system may be utilized. In particular, the accuracy of direction-finding is always limited by the degree to which the ideal properties of the antenna system satisfy certain requirements, such as insensitivity to vertically downcoming signals, and by the accuracy with which the actual antenna characteristics agree with the ideal ones. Further limitations are imposed under certain conditions by vagaries inherent in radio-wave travel.

One main form, the aircraft direction-finder, or radio compass, requires a directive receiving antenna and its auxiliary apparatus to be carried on board, and gets directional information from signals from ordinary radio transmitters on the ground. It gives the flight personnel full control of and responsibility for radio navigation and requires a proportionate amount of attention from them. In effect, it permits them to "see" ground transmitters when there is no visibility. It is very convenient for "homing," or flight headed directly toward the transmitter, but requires correction for the ship's heading, as

[1] R. Keen, "Wireless Direction-Finding," 3rd ed., *Iliffe*, London, 1938.

determined by magnetic compass, to be useful for navigation. In its newest forms, in which the indicating means automatically point directly at any station tuned in,[2,3] the effort required to obtain the information which the radio compass can supply is reduced to a very low level. Requirements of compactness and mechanical simplicity have always made necessary the exclusive use of loop antennas for aircraft direction-finders. Loop direction-finders are fundamentally subject to errors when receiving waves that arrive from other than horizontal directions with other than vertical polarization,[4] conditions often found at night. Thus, the reliability of the radio compass is sometimes limited by conditions beyond control. Also, no really satisfactory radio compass has yet been developed for use at the ultra-high radio frequencies, where natural static normally causes only negligible interference.

In another form, the ground direction-finder, the directional receiving antenna and its auxiliaries are on the ground. Thus, cumbersome, but precise Adcock (spaced vertical)[5] or spaced loop[6] antennas may be used and errors due to oblique wave incidence and polarization may be avoided. Use of ground direction-finders keeps ground personnel, both authorized and unauthorized, directly informed of aircraft positions, and in fact transfers to them the work of position finding. In effect, they enable ground staff to "see" transmitting aircraft, despite poor visibility. Aircraft personnel is relieved of much of the work of navigation, relying on the observations and calculations made on the ground. No extra aircraft equipment is required, but much extra transmission from aircraft is involved in the intensive use of ground direction-finders. This creates a serious interference problem, particularly since only one aircraft can be served at a time and each one requires especially frequent service when conditions are difficult. Recently, automatically indicating ground direction finders[7] have become available, so that these, like modern automatic radio compasses, are very convenient to use. Ultra-high-frequency forms are not yet

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[2] "An Automatic Direction-Finder," *Communications*, Vol. 18, Oct. 1938, p. 10.

[3] H. Busignies, "The Automatic Radio Compass and its Application to Aerial Navigation," *Elect. Communication*, Vol. 15, 1937, p. 157.

[4] T. L. Eckersley, "The Effect of the Heaviside Layer on the Apparent Direction of Electromagnetic Waves," *Radio Review*, Vol. 2, Feb. 1921, No. 2, p. 60 and May 1921, No. 5, p. 231.

[5] F. Adcock, "Improvements in Means for Determining the Direction of a distant Source of Electromagnetic Radiation," *Brit. Pat.* No. 13,0490, 1919.

[6] R. H. Barfield and W. Ross, "The Measurement of the Lateral Deviation of Radio Waves by Means of a Spaced Loop Direction-Finder," *Journ. Inst. Elec. Eng.*, Vol. 83, July 1938, p. 98.

[7] Nat. Phys. Lab. Staff, "A Short-Wave Cathode-Ray Direction-Finding Receiver," *Wireless Engineer*, Vol. 15, Aug. 1938, No. 179, p. 432.

fully developed,[8,9] but here no fundamental difficulty appears to bar the way as in the case of the radio compass.

Finally, the directional antenna system has been used at ground transmitters to give a third main radio navigational method. These directional radio beacons or, to borrow a nautical term, radio "range" beacons, have appeared in a variety of forms, which happen to be classifiable into four main groups. In general the directional antenna, being on the ground, can be precisely executed and of a type free from night effects. Responsibility for providing correct information rests with the ground personnel and responsibility for obtaining and correctly using it, with flight personnel: this seems to be an essentially satisfactory division of effort. Like the ground direction-finder, but in contrast with the radio compass, the radio range beacon provides direct information as to the positions of aircraft over the ground and is not at all affected by their headings. Like the radio compass and in contrast with the ground direction-finder, as many ships as wish to do so may obtain bearings from a single ground station, simultaneously and continuously. A number of radio ranges of various types have been made to operate satisfactorily at ultra-high frequencies, where the main interference problem is the essentially curable one of noise generated on the receiving craft. But the range to be described in this paper is the only one yet available that offers the flexibility and convenience in use of the modern ground and aircraft direction-finding receivers.

C—*Radio Range Beacons*

The only form of directional beacon which has come into wide use so far is the aural equisignal type of radio range[10,11] that has played so important a part in the development of American commercial air transport to its present outstanding position. In this device, two differently directed figure-eight patterns of tone-modulated radio-frequency energy are radiated alternately. Energy is transferred between the two patterns in the rhythm of interlocking A and N telegraphic characters. The pilot of an aircraft located along the zero line of one pattern hears only the character transmitted on the other pattern; when anywhere else, he hears an admixture of both characters. In

[8] R. L. Smith, Rose and H. G. Hopkins, "Radio Direction-Finding on Wavelengths Between 6 and 10 Meters," *Journ. Inst. Elec. Eng.*, Vol. 83, 1938, p. 87.

[9] R. L. Smith, Rose and H. G. Hopkins, "Radio Direction-Finding on Wavelengths Between 2 and 3 Meters," *Journ. Inst. Elect. Eng.*, Vol. 87, Aug. 1940, No. 524, p. 154.

[10] J. H. Dellinger and H. Pratt, "Development of Radio Aids to Air Navigation," *Proc. I.R.E.*, Vol. 16, 1928, p. 890.

[11] W. E. Jackson and D. M. Stuart, "Simultaneous Radio Range and Telephone Transmission," *Proc. I.R.E.*, Vol. 25, 1937, p. 314.

particular, along a bisector of the angle between the patterns both signals are of equal strength and a pilot there hears a steady tone, which signifies that he is on a "course," or "range," of the beacon. If he drifts off course to one side, one pattern predominates and he hears an A above the steady tone, while if he deviates to the other side he hears an N. The range beacon thus lays out straight radial tracks over the ground, in effect painting the air to the right of a track green and the air to the left red.

The equisignal range is thoroughly satisfactory if a course happens to lie along the route one wishes to follow and if, once on course, one never loses the signal nor departs seriously off course. But the range is of little use if its courses do not fit one's plans. And the four courses are distinguishable only by maneuvering, so that a pilot must exercise care to be sure of starting out on the right one. If he goes far off course for any reason, he must carry out in detail a carefully planned series of maneuvers, an "orientation procedure," to find and identify the course to which he later returns. Range beacons with distinguishable courses,[12] with visual rather than aural indication of deviations from course,[13] and with as many as twelve courses[14] have been developed, but for various reasons none of these improvements has come into use. A development which is universally used, however, enables all four courses to be oriented as desired by deforming the radiated field patterns.[15] Visual indicating equisignal ranges are in experimental use as runway localizers for instrument landing and their use is proposed for ultra-high-frequency navigational aids.

A second type of range also involves the comparison of distinct signals from two or more differently oriented antenna arrays. This may be done in terms of amplitudes of distinguishable modulations,[16] a generalization of the equisignal fixed-course range. Another ingenious proposal of this[17] sort involves comparison of signal-travel times from three spaced antennas, by transmitting frequency-modulated waves from two of these antennas at a time and measuring the resulting beat frequencies at the receiver.

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[12] F. W. Dunmore, "A Method of Providing Course and Quadrant Identification With the Radio-Range Beacon System," *Bur. Stand. Jour. Res.*, Vol. 11, 1933, p. 309.

[13] J. H. Dellinger, H. Diamond and F. W. Dunmore, "Development of the Visual Type Airway Radio-Beacon System," *Proc. I.R.E.*, Vol. 18, 1930, p. 796.

[14] H. Diamond and F. G. Kear, "A 12-Course Radio Range for Guiding Aircraft With Tuned-Reed Visual Indication," *Proc. I.R.E.*, Vol. 18, 1930, p. 939.

[15] F. G. Kear and W. E. Jackson, "Applying the Radio Range to the Airways," *Proc. I.R.E.*, Vol. 17, No. 12, 1929, p. 2268.

[16] F. W. Dunmore, *U. S. Patent*, No. 2,128,928; 1938.

[17] E. N. Dingley, Jr., "A True Omnidirectional Radio Beacon," *Communications*, Vol. 20, No. 1, Jan. 1940, p. 5.

Still a different approach to the directional beacon was proposed long ago.[18] In this third form, a radio beam is transmitted and rotated about the transmitter. From sector to sector of the rotation, the character of the radiation is changed, as by telegraphic keying with different characters.[19,20] A pilot listening to such a beacon need only note which character is heard the loudest in order to determine in which sector about the beacon his craft is. In a more modern version,[21] direct, automatic indication of bearing all around the beacon is provided by varying modulation frequency linearly as the beam is rotated and observing, on a series of tuned reeds, the modulation frequency of the loudest signal received.

The fourth type of directional beacon also employs a rotating radio beam and, like the second and third types, merits the name of "omnidirectional radio range," since it distinguishably marks out straight radial courses in all directions about the beacon. In this form, also proposed and even developed to some extent many years ago,[22] the instant at which the rotating beam points in some chosen direction, for instance due north, from the transmitter is marked by a signal transmitted in all directions. Thus, the proportion of the interval, between successive marking signals, which elapses between a marking signal and the next following beam signal maximum is directly that proportion of the full 360-degree sweep of the beam which represents the azimuth or bearing of the receiver from the transmitter.

A practical form of this device was developed in England, using first a mechanically rotated loop antenna[23] and later avoiding antenna motion.[24] This system involved the manual timing of intervals at the receiver with a stop watch, and therefore, was too slow and inconvenient in use to be satisfactory for aircraft guidance. Englund,[25] and later several others,[26,27,28] realized that limitations on speed of operation could be removed by rotating the beam by purely electrical means and that indication could be made automatic. Two channels, one for the rotating beam and one for the marking signal, are required,

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[18] Lee deForest, *U. S. Patent.* No. 833,034; 1906.

[19] C. S. Franklin, "Short-Wave Directional Wireless," *Journ. Inst. Elec. Eng.*, Vol. 60, 1922, p. 933.

[20] J. Robinson, *Brit. Pat.* No. 327,112; 1928.

[21] C. W. Hansell, *U. S. Pat.* No. 2,014,732; 1935.

[22] J. Zenneck (Trans. Seelig) "*Wireless Telegraphy,*" McGraw-Hill, New York, 1915, p. 368.

[23] T. H. Gill and N. F. S. Hecht, "Rotating Loop Radio Transmitters and their Application to Direction-Finding and Navigation," *Jour. Inst. Elec. Eng.*, Vol. 66, 1928, p. 256.

[24] H. A. Thomas, "A Method of Exciting the Aerial System of a Rotating Radio Beacon," *Jour. Inst. Elec. Eng.*, Vol. 77, 1935, p. 285.

[25] C. R. Englund, *U. S. Pat.* No. 1,815,246; 1931.

[26] P. H. Evans and J. W. Grieg, *U. S. Pat.* No. 1,933,248; 1933.

[27] J. W. Grieg, *U. S. Pat.* No. 1,988,006; 1935.

[28] G. H. Brown and D. G. C. Luck, *U. S. Pat.* No. 2,112,824; 1938.

and the various proposed schemes differ mainly in the methods suggested for multiplexing these two channels onto a single radio carrier wave.

Only two of the proposed methods seem particularly well suited to practical development, however. One of these employs two very different waveforms[29] and the other, double modulation and a subcarrier[30]. The purpose of this paper is to describe the principles of an experimental system developed on the basis of modulation waveform multiplexing. This system, giving direct, immediate, and automatic indication of the bearing of a mobile receiver from the beacon transmitter to which it is tuned, is believed to be the first omnidirectional radio range to reach a practically useful stage of development. It not only lays down straight, radial tracks over the ground, wherever they may be wanted, but also stamps each one with a unique distinguishing number. With a number of beacons of this type in service, full navigational information would be available at all times to all craft within reach of the signals of two or more beacons, in a directly useful form and with expenditure of a minimum of effort on the part of flight personnel.

2—Principles of Omnidirectional Range Oepration

A—*Beacon*

A short, free-ended rod carrying radio-frequency current acts, as is well known, as an oscillating electric dipole coaxial with the rod. A small plane loop similarly acts as a magnetic dipole coaxial with the loop, with electric field in the loop plane. Either type of dipole radiates uniformly in any cone of directions making a constant angle with its axis, giving no radiation along the axis and maximum radiation in the equatorial plane perpendicular thereto. A parallel pair of dipole antennas, of either type, may be placed close together in a common equatorial plane and fed with currents of accurately equal amplitude and opposite phase. This array is called an Adcock antenna system.

In the plane perpendicular to and bisecting the line of separation of the antennas, the fields produced by the two oppositely fed antennas just cancel, so no signal is received anywhere in that plane. Any point elsewhere in space is not equidistant from the two antennas, so waves from each of the latter must travel for different lengths of time to reach such a point. Consequently, the two waves are not in exactly opposite phase upon arrival and fail to cancel exactly. The path difference is greatest for points along the extended line of antennas, so

[29] D. G. C. Luck, *U. S. Pat.* No. 2,208,376; 1940.
[30] J. W. Grieg, *U. S. Pat.* No. 2,129,004; 1938.

points there receive the strongest signals. The instantaneous polarity of the signal at any particular place depends on which of the oppositely fed antennas is the nearer to that place.

A polar plot of the signal strength in the common equatorial plane, at a constant distance from the center of the antenna system, shows the well-known "figure eight" form, the two lobes of the figure being circular in shape and representing signals of opposite polarities. Analytically, the radio-frequency field e for points at a constant distance from the antennas is described by

$$e = e_0 \cos \theta \tag{1}$$

where e_0 is the field on the line of antennas and θ is the angular departure of the point of measurement from that line. The field from a closely spaced, oppositely fed pair of antennas has the peculiarity of being, everywhere except in the immediate neighborhood of the antennas, just 90 degrees different in phase from that produced by a single antenna located at the center of the pair and fed with current in phase with that in an antenna of the pair.

Feeding the antenna pair through a balanced modulator, excited by unmodulated carrier and alternating modulation voltage the radiated pattern will pass successively through alternate maxima E_1, of opposite polarities, falling to zero between maxima as the modulating voltage passes through zero in its alternation. That is, for sinusoidal modulation,

$$e_0 = E_1 \cos at \tag{2}$$

and

$$e = E_1 \cos \theta \cos at \tag{3}$$

where $a/2\pi$ is modulation frequency in cycles per second and t is time in seconds.

Suppose that we erect a pair of vertical dipole antennas spaced along a north-south line and another pair, symmetrical with respect to the first, spaced along an east-west line. We feed both pairs with symmetrically modulated radio-frequency carrier energy, in phase as regards radio frequency, but with the east-west modulation lagging behind the north-south modulation by a quarter period of the modulation wave. If θ is now azimuth measured clockwise from the northward geographic meridian through the antenna system, or true bearing from the antennas,

$$e_{NS} = E_1 \cos \theta \cos at \tag{4}$$

is the amplitude of the field radiated by the north-south pair.

$$\left.\begin{aligned} e_{EW} &= E_1 \cos(\theta + 90^\circ)\cos(at - 90^\circ), \text{ or} \\ e_{EW} &= -E_1 \sin \theta \sin at, \end{aligned}\right\} \tag{5}$$

is that of the field radiated by the east-west pair. Since these fields are in phase at radio frequency, they add algebraically to give

$$\left.\begin{aligned} e_D &= e_{NS} + e_{EW} = E_1(\cos\theta\cos at - \sin\theta\sin at), \text{ or} \\ e_D &= E_1\cos(at+\theta) \end{aligned}\right\} \quad (6)$$

as the overall directive antenna field.

A single antenna fed with unmodulated carrier energy producing a steady non-directive, or broadcast, field is placed symmetrically with respect to the four antennas of the directive array. It is fed with radio-frequency current in phase quadrature with those fed to the directive antennas. The broadcast and directive fields will then be in phase, at a distance from the antennas, in all directions, and so may be added algebraically. Representing the steady broadcast field amplitude by E_o, the resultant field amplitude is

$$e_R = e_{BC} + e_{NS} + e_{EW} = E_o + E_1\cos(at+\theta) \quad (7)$$

which is the equation of a limaçon. The pattern would have the special one-cusped form called a cardioid if E_1 were equal to E_o, but we require E_o to exceed E_1. The phase of the modulation of the radiated field amplitude, at constant distance, is always and everywhere $at+\theta$. So we see that each stage of the modulation cycle, characterized by its own particular phase value, occurs for uniformly smaller and smaller values of t, that is, earlier and earlier, for greater and greater values of θ, that is, for successively more clockwise directions. In other words, Equation (7) represents a limaçon rotating uniformly counterclockwise.

We now have a channel, transmitting a signal to which things happen at different times in different directions, which can obviously be the beginning of a direction-finding system. But, to use it, we must make sure that time measurements correspond in all directions. That is, we must transmit a suitable clock-setting, or reference phase, signal in all directions, over a distinguishably separate channel. We do this by switching all radiation completely off, very briefly, while the maximum of the rotating limaçon pattern points within the small angle θ_k (about ½ degree) of due North. An analytic expression for the complete transmitted signal amplitude may be included for completeness:

$$e = E_o\left[1 + \frac{E_1}{E_o}\cos(at+\theta)\right]\cdot\left(1 - \sum_{n=1}^{n=\infty} K_n\cos nat\right) \quad (8)$$

The coefficients K_n depend on the exact waveform of the keying impulses. For instantaneous transitions from "on" to "off" when at is $-\theta_k$ degrees and back to "on" when at is θ_k degrees, they would be

$$K_n = \frac{2\,\theta_k}{180 - \theta_k} \cdot \frac{\sin n\pi \dfrac{\theta_k}{180}}{n\pi \dfrac{\theta_k}{180}} \tag{9}$$

These coefficients decrease only slowly with increasing harmonic order n.

While the basic principle of the beacon is thus very simple, the fact that the variation of signal amplitude with both azimuth and time is essentially a three-dimensional affair tends to make it difficult to visualize all details of operation in their proper sequence. The elementary graphic presentation of Figures 1 and 2 may help to clarify the situation, each picture being only two-dimensional in its significance. Figure 1 is a sequence of instantaneous polar graphs, snapshots

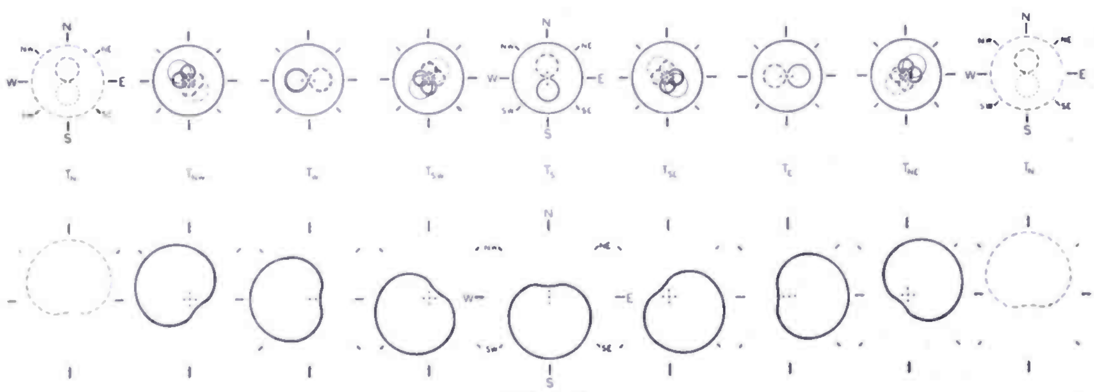


Fig. 1

of the azimuthal variation of signal strength at chosen instants of time. Figure 2 is a set of time graphs, oscillograms of signal-amplitude variation at fixed points, in chosen azimuths from the transmitter. To interconnect the figures, the instants of the snapshots of Figure 1 are noted on the time scales of the oscillograms of Figure 2, while the azimuths yielding the oscillograms of Figure 2 are marked off on the snapshots of Figure 1.

In Figure 1, the upper row of polar graphs shows separately the three component directive patterns, the broadcast unmodulated carrier pattern and the north-south and east-west modulated figure-eight patterns. Figure-eight lobes in phase with the broadcast signal are shown in full lines, while the contraphased lobes are shown dashed. The resultant of the two figure-eights, itself a figure-eight of constant size, is shown in light lines on each graph. Round dots mark the positions of the antennas energized at the instant of each snapshot. It will be noted that none are energized at the instants T_N: these pictures should properly be blank, the signal then being off entirely, but that would be too uninformative, so ghost patterns are dashed in lightly to show what would be there if it had not just been turned off. The lower row of graphs in Figure 1 shows the limacoid patterns resulting

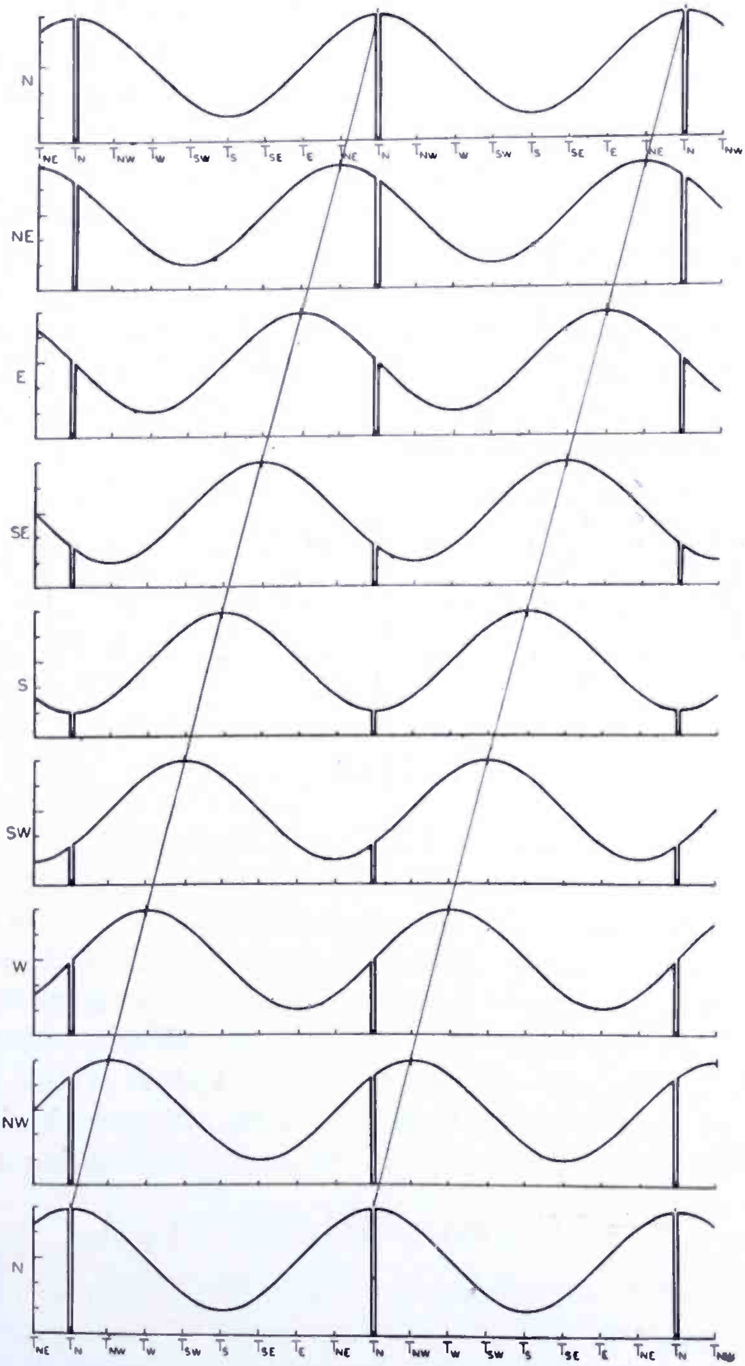


Fig. 2

from the addition of the component patterns above them. Uniform rotation of the resultant pattern is obtained.

In Figure 2, the change with azimuth of the time relation between the component sinusoidal and impulsive amplitude-modulation waves shows immediately. Comparing, for example, the modulation found southeast of the beacon, in Figure 2, with the signal in that direction shown by the space graphs of Figure 1, the minimum is seen either way to occur at time T_{NW}, the maximum half a cycle later at T_{SE} and the reference impulse three-eighths of a cycle later still, at T_N. The slant lines of Figure 2 indicate how the delay between each sine wave maximum and the next following impulse is linearly related to the azimuth of transmission. In fact, the electrical phase angle representing this delay is always directly that azimuth.

The reason for avoiding a true cardioid is evident from the figures. Such a pattern, with zero signal at its one minimum, would not allow any impulse to be transmitted due southward, as may be seen from the snapshot at T_N and the oscillogram for southward transmission. It is, of course, a fundamental of successful multiplexing that no one channel may completely suppress the total signal just when another channel must carry information.

Fig. 3

For checking adjustment of indicating equipment, it is necessary to know to what azimuth some particular received field variation really belongs. To facilitate this, it is convenient to be able to transmit, when desired, a signal independent of azimuth and everywhere identical with that corresponding normally to one definite azimuth. This is accomplished by switching off all power to the directive antennas and sinusoidally modulating the carrier current fed to the broadcast antenna, with such modulation phase that the reference keying occurs just at the instant of peak output. This replaces Equation (8) by

$$e_o = E_o(1 + m\cos at)(1 - \Sigma K_n \cos nat) \tag{10}$$

where the modulation depth m should have the same value, E_1/E_o, as in normal operation. Figure 3 shows a series of polar graphs of the calibrating signal, at instants corresponding to some of those for the graphs of Figure 1 for the normal signal. Figure 4 is an oscillogram of amplitude variation with time of the calibrating signal. This is, of course, the same for all directions. Comparison of Figure 4 and Figure 2 shows that the calibrating signal, the same everywhere, is identical with the signal normally transmitted due North only.

It may be as well to conclude this explanation of the operation of the beacon by pointing out that the concept of a signal of single, carrier frequency, but of varying amplitude, has been used throughout because it seems to give directly an especially clear picture of what takes place. The mathematically equivalent concept of a carrier and side bands, all of constant amplitude, might of course have been used, throughout, instead, but seems less clear in the present case. Only if both concepts are intermingled is serious confusion almost certain to arise. It should also be noted that nowhere do we deal with a true rotating field vector, such as occurs in the somewhat analogous case of a synchronous motor. What rotates is the distribution-in-azimuth of the otherwise unvarying magnitude of a radiated carrier field of constant direction.

B—*Azimuth Indicator*

An ordinary radio receiver serves to convert the amplitude variations of the radio signal transmitted by the beacon into similar varia-

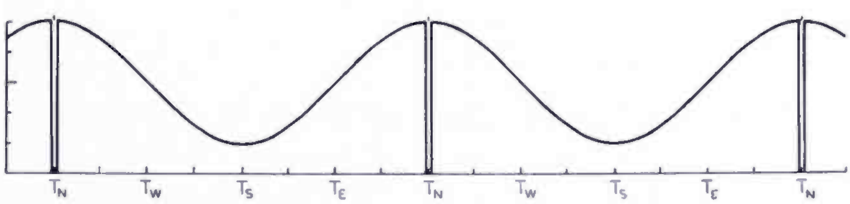


Fig. 4

tions of direct output voltage. The receiver should have good automatic gain control to hold the average of this output voltage constant despite wide variations of field strength. Receiving antennas should, of course, be nondirectional. The receiver-output voltage variation, like the signal amplitude variation of Figure 2, comprises a sinusoidal component produced by the pattern rotation and a train of impulses corresponding to the reference keying, but contains as well assorted additional components representing received noise of all kinds. The task of the indicators is measurement of the phase angle between the two differently shaped waves, which is also the true bearing of the receiver from the transmitter, without serious disturbance by noise.

A narrow band-pass filter selects the sinusoidal pattern-rotation component from the receiver output. Filter output is applied directly to the vertical deflecting circuit of a cathode-ray oscillograph. Assuming no delay in wave travel, receiver or filter, the filter output voltage is

$$v_f = V_f \cos(at + \theta) \qquad (11)$$

Changing to a new time variable, we obtain

$$t' = t + \frac{\theta}{a} \tag{12}$$

This becomes

$$v_f = V_f \cos a\, t' \tag{12a}$$

for any bearing θ. The vertical deflection of the oscillograph spot is

$$Y = S_v \frac{v_f}{V_a} \tag{13}$$

where V_a is the cathode-ray beam accelerating voltage and S_v is a vertical-deflection sensitivity.

Delaying the filter output a quarter cycle by means of a phase shifter produces a quadrature voltage

$$v_q = V_q \sin a\, t' \tag{14}$$

Applying this to the horizontal deflecting circuit of the oscillograph gives a horizontal deflection

$$X = S_H \frac{v_q}{V_a} \tag{15}$$

The overall result is travel of the oscillograph spot in the locus

$$X^2 + Y^2 = \left(\frac{S_h V_q}{V_a} \right)^2 \sin^2 a\, t' + \left(\frac{S_v V_f}{V_a} \right)^2 \cos^2 a\, t' \tag{16}$$

If $S_v V_f$ and $S_h V_q$ are equal, this becomes

$$X^2 + Y^2 = \left(\frac{SV}{V_a} \right)^2 = \text{constant}, \tag{17}$$

a circle centered at the undeflected position of the spot. At the instants when t' is an integral multiple of $2\pi/a$, the spot is just at the top of its path. It travels, clockwise, along that path at constant speed, the arc traversed subtending an angle at the center equal to the value of $a\, t'$.

Impulses may be segregated from the receiver output by a high-pass filter and used to decrease momentarily the cathode-ray accelerating voltage V_a. As transmitted, the impulses occur when t is an integral multiple of $2\pi/a$, but—in terms of the new time variable—this is when $a\, t'$ exceeds an integral multiple of 2π by the azimuth θ (Equation 12). A decrease of V_a increases both spot deflections proportionally (Equations 13, 15), so that each impulse momentarily increases the radius of the circular spot path. As the impulses occur θ degrees after the spot passes the top of its path, the outward jog on the uniformly swept circle marks off this angle, the true bearing of receiver from transmitter, directly. The angle may be read against a circular scale, having its zero at the top adjacent to the swept circle. It is now clear that the assumption made above, of no delay in sinusoid transmission, merely means equal delay of sinusoid and impulses.

Figure 5 shows graphically all essentials of the operation of the cathode-ray azimuth indicator. At the left center is shown an oscillogram of the filtered receiver output, used for vertical deflection, and

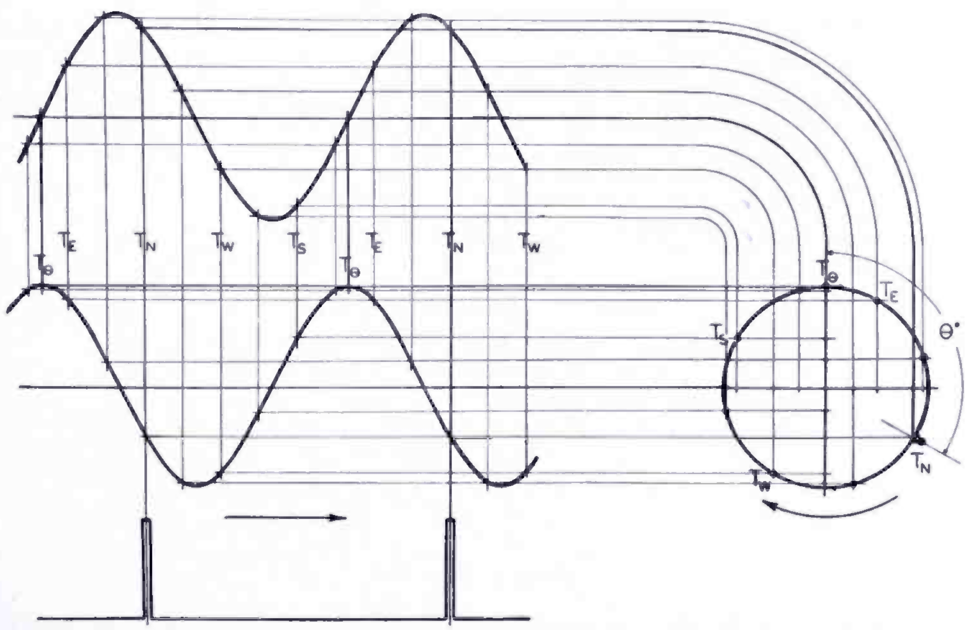


Fig. 5

above this appears an oscillogram of the quarter-cycle-delayed filtered wave used for horizontal deflection. The instant at which the limaçon points at the receiving airplane in azimuth θ, as in the full-line graph of Figure 6, is marked on the oscillograms of Figure 5 by T_θ. Other special instants, at which the cardioid points in the cardinal directions, are marked in Figure 5 as in Figure 2. Keying impulses are shown at the instants of their occurrence, T_N, at the lower left. Projection of the marked values of deflecting voltage, along the light lines, onto the perpendicular axes of oscillograph deflection gives the instantaneous deflection components. Instantaneous resultant deflections are given by the intersection of corresponding projection lines and are marked by dots and labeled according to the time of their occurrence. The locus of resultant deflections—the path of the spot—is shown by the heavy circle connecting the resultant points. At time T_N, the impulse kicks the spot outward, marking the jog seen in the figure. From T_θ to T_N, the limaçon maximum of Figure 6 rotates counter-clockwise through the azimuth angle θ. Since the indicator spot is at the north point of its scale at T_θ and thereafter moves clockwise in step with the limaçon, at T_V, it has also swept through the azimuth angle θ and the jog then marked by the impulse indicates directly the value of θ.

Correct indications appear only if the indicator spot is at the zero of its scale exactly when the transmitted beam points directly at the receiver. This could be checked by reading the indication when at a known bearing from the transmitter, but such a known bearing is not always likely to be available when needed. It is to permit checking at any time that the calibrating signal of Figures 3 and 4 can be transmitted upon request. This special signal, everywhere just the same as that sent due north in normal operation, will give a bearing indication of zero degrees on any correctly adjusted indicator. Deviation from this value can be corrected, if present, by rotating the scale, as is done

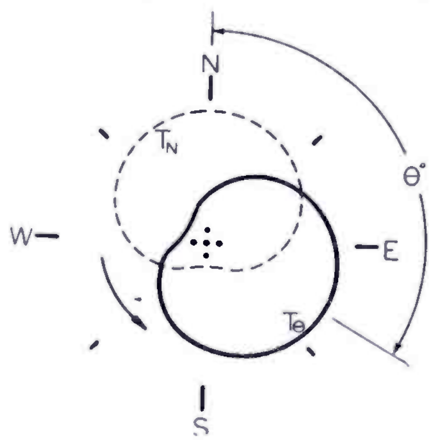


Fig. 6

in setting other aircraft instruments, or by providing an adjustable electrical delay between the receiver output and the band-pass filter.

C—*Deviation Indicator*

It is a great convenience, in radial flight directly toward or away from the range beacon, to avoid the necessity of repeatedly reading and interpreting the quantitative azimuth indication of the cathode-ray instrument.[31] This reading may be anywhere around the scale and, in radial flight, must merely be held constant.

From the filter separating out the pattern-rotation component of the receiver output, two outputs varying in push-pull fashion may be derived, one increasing as the other decreases. Suppose the push-pull filter output to be connected to the second control grids of a pair of multigrid vacuum tubes, with the tube plates connected to a zero-center meter so that the latter reads the average value of the difference of the two plate currents. If the first control grids of both tubes are normally biased beyond plate-current cutoff, but are brought momentarily to

[31] D. G. C. Luck, *U. S. Pat.* No. 2,208,377; 1940.

zero bias by the reference keying, plate current will flow only during the keying impulses. The relative average current values in the two plate circuits will be determined by the relation of the push-pull filter outputs at the keying instants.

Due east of the transmitter, the reference impulse comes just as the strength of the rotating pattern has fallen halfway from its maximum to its minimum value, as may be seen from Figures 1 and 2. In that direction, the oppositely varying currents will be equal at the instant when a reference impulse occurs, as shown at *A* in Figure 7. In this figure, the ordinates of the light full and dashed curves represent, respectively, the values that the two oppositely acting plate currents would have if turned on. Impulses are outlined in light lines, three different timing conditions being shown. The short heavy lines

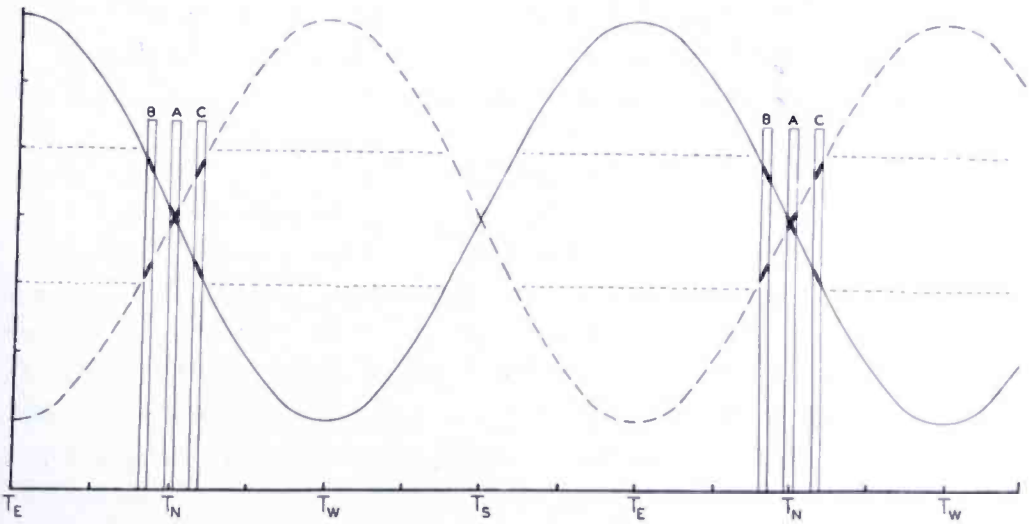


Fig. 7

represent actual plate currents during impulses. For a receiver a little north of east, the limaçon maximum will sweep by a little later and the keying will occur just before the receiver output has fallen to its halfway value, as shown at *B* in Figure 7. One tube plate current will predominate and the meter will deflect one way. South of east, maximum signal will be received earlier and the keying will come more than halfway down, as at *C* Figure 7. The other current will predominate and the meter will deflect the other way. Thus, the zero-center meter is made, by a differential vacuum-tube circuit, to indicate sense and magnitude of small deviations of the receiver position from the radial course due eastward of the beacon.

Signals in various azimuths from the beacon differ only in the relative timing of reference keying and pattern-rotation modulation. Therefore, the signal in any chosen direction may be changed to simulate the eastward signal merely by delaying the effect of pattern rotation. Using a fully variable phase shifter between the narrow band-filter and the vacuum-tube grids, any desired delay may be introduced, so the condition at *A* of Figure 7 may be produced for any chosen course and the meter will then indicate deviations from that course. Examination of Figure 7 will show that there are two instants during the cycle at which an impulse would produce equal currents. Similar deviations from these two balance conditions will produce opposite indications. To avoid this ambiguity, a rough bearing scale may be applied to the phase shifter and care be taken, in setting the meter to zero when on a course to be held, that the reading of this scale shall agree approximately with that of the cathode-ray azimuth indicator.

Since the deviation indicator is only useful when near a chosen course, it should reach full deflection for quite small deviations. But large deviations must not injure the instrument. These conditions can be met by limiting the push-pull signal applied to the switching tubes, as indicated in Figure 7 by the dotted straight lines cutting off the signal peaks.

3—Sources of Error

A—*General*

Obstacles, ground contours, and ground-conductivity variations are able to affect markedly the travel of radio waves. In this way there arise "terrain errors" in all radio direction-finding. These are, in principle, neither better nor worse with an omnidirectional range than with any other radio system. However, the continuous indication given by the former makes the course of such vagaries unusually clear and so minimizes the difficulty they can cause. Terrain errors need not be further mentioned here, except to point out that when caused by obstacles near the directive radiator they are called "site errors" and may be especially bad. In consequence, it is important to select as large a clear site as can be found for erection of the beacon antennas.

Many different sorts of imperfection in operation of the equipment itself can also produce errors of bearing indication, however, and these it is worthwhile to consider in some detail. The procedure followed in each case below is to consider a single possible source of error to act alone, choosing suitable parameters to characterize that error source quantitatively. In the presence of an error source, the variation of azimuth indication with true bearing is determined and the difference between indicated and true bearings computed. This gives the error produced, as a function of true bearing, in the form of a correction to be added to true bearing to give indicated bearing. As the method is always the same only one or two cases will be worked through as examples, the results alone being given for others. Errors in a properly operating system will be small, so that first approximations are adequate and the relations of the individual errors to their causes become very simple.

B—*Conditions for Ideal Operation*

Consideration of the various errors of a real system may well be prefaced by an enumeration of the conditions which would have to be fulfilled by an ideal system. When all modulation phases are referred to that of the north-marking impulse train, the modulation phase conditions for perfect operation become: calibrating and north-south directive modulation must be in phase, and east-west directive modulation in phase quadrature, with the reference impulses. Pattern modulation-amplitude relations for perfect operation are: north-south and east-west directive-modulation amplitudes must be equal to each other and to the calibrating-modulation amplitude, no rotation-frequency modulation of the broadcast carrier (hum) may be present during directional operation, and the amplitude of the fundamental-frequency component of the impulse modulation must be negligible compared with that of the desired directive modulation.

There are also certain radio-frequency relationships to be fulfilled for perfect operation, as radio-frequency current variations in all four antennas of the directive array must at all times be in phase quadrature with that in the broadcast antenna, while currents in the two antennas of each directive pair must be accurately equal in magnitude and opposite in phase. It is a great operating convenience, though not fundamentally necessary, to have no coupling between any two antennas save those of a single directive pair.

Geometrically, the antennas must be accurately parallel and vertical, over an accurately horizontal, symmetrical ground plane free of undulations or obstacles for a large distance in all directions. The antennas must be located accurately at the corners and center of a square having its diagonals aligned accurately north-south and east-west and must be close enough together to ensure the true cosine form of Equation (1) (p. 62) for the figure-eight patterns.

At the receiver, any time lag between pattern-rotation variation of the received field and the resulting variation of vertical deflecting voltage at the cathode-ray tube must be the same as the time lag occurring between impulsive field variation and impulsive velocity-modulating voltage applied to the cathode-ray tube. The receiver must not have its output affected by slight variation of the radio frequency of the received field, and it is a great convenience to have the amplitude of the voltage variations applied to the cathode-ray tube accurately independent of received signal level. No spurious receiver output voltage variations due to local signal sources, obstacle motions, or receiver variations are, ideally, permissible. The vertical and horizontal cathode-ray deflecting voltages must be accurately in phase quadrature and must be related in amplitude so as to give equal amplitudes of cathode-ray spot deflection. Electromechanically, the cathode-ray-tube deflection components must be accurately linear, symmetrical functions of deflecting voltages and deflection axes must be rectilinear and mutually perpendicular. Mechanically, the cardinal points of the uniform indicator scale must be accurately aligned with the deflection axes of the cathode-ray tube. Rotation of the scale as a whole after alignment is, however, permissible if phase compensation is available.

Manifestly, ideal operation is not practically realizable. It is, however, rather amazing how nearly the ideal may be approached by careful attention to the control of each error source, especially if some error sources can be adjusted to cancel the effects of other, less tractable ones. While not all of the above enumerated conditions for perfection are explicitly present in the case of some other versions of the omnidirectional range principle, close scrutiny will usually reveal an equivalent for each of them in any given case. Often, additional conditions must also be satisfied. For example, many types require overall amplitude linearity of the entire electrical system. This requirement is largely relaxed in our case because of the smallness of the fundamental rotation-frequency component of the impulsive timing wave, a feature which minimizes cross-talk between the directive and reference modulation channels.

C—*Transmitter Modulation Conditions*

To return to the subject of errors produced by individual causes: if both north-south and east-west modulation are equally displaced in phase from their ideal relations to the reference wave, an error equal in magnitude to the common phase displacement and independent of azimuth is produced. For a single range this is of no consequence, as it may be compensated by receiving indicator zero setting. But in a network of range beacons such displacements, peculiar to individual beacons, require realignment of indicators whenever receivers are tuned from one beacon to another to ensure concordant bearings. For this reason, such errors must be avoided for each beacon, as they will be if both directive-modulation channels are individually correctly phased. If the calibrating modulation is incorrectly phased with respect to the impulsive reference wave, an azimuth-independent error of like amount will be produced in a less direct way. This error will become effective by being transferred to receiving indicator zero settings, so that any indicator set by a beacon with such an error will read incorrectly on any beacon thereafter received.

In the event of north-south directive modulation only being out of phase with the reference wave, by an amount P_1, Equation (4) for the north-south field becomes

$$e_{NS} = E_1 \cos\theta \cos(a\,t + P_1) \tag{18}$$

which combines with Equation (5) for the east-west pattern and with the broadcast pattern to give

$$e = E_0 + E_1 [\cos\theta \cos P_1 \cos at - (\cos\theta \sin P_1 + \sin\theta) \sin at] \tag{19}$$

for the complete radiated field amplitude. This is trigonometrically equivalent to

$$e = E_0 + E_1 \sqrt{1 + \sin 2\theta \sin P_1}\ \cos(at + \phi_1) \tag{20}$$

where ϕ_1 is the angle which will now be indicated as the bearing of a receiver and is given by

$$\tan\phi_1 = \frac{\cos\theta \sin P_1 + \sin\theta}{\cos\theta \cos P_1} \tag{21}$$

Calling $\phi_1 - \theta$, the error to be added to true bearing θ to give indicated bearing ϕ_1, ρ_1, use of Equation (21) for ϕ_1 in the usual trigonometric formula for the tangent of the difference of two angles gives

$$\tan\rho_1 = \frac{\sin\theta\cos\theta + \cos^2\theta \sin P_1 - \sin\theta\cos\theta\cos P_1}{\sin^2\theta + \cos^2\theta\cos P_1 + \sin\theta\cos\theta\sin P_1} \tag{22}$$

This is in turn trigonometrically equivalent to

$$\tan\rho_1 = \sin\frac{P_1}{2}\cdot\frac{\cos(2\theta - P_1/2) + \cos P_1/2}{1 + \sin\dfrac{P_1}{2}\left[\sin\left(2\theta - \dfrac{P_1}{2}\right) - \sin P_1/2\right]} \tag{23}$$

Since P_1 and, in consequence, ρ_1, is small in proper operation, a first approximation to Equation (23), obtained by letting $\tan\rho_1 = \rho_1$, $\sin\dfrac{P_1}{2} = \dfrac{P_1}{2}$, and $\cos\dfrac{P_1}{2} = 1$ is satisfactory: this yields

$$\rho_1 = P_1 \cos\theta \cos(\theta - P_1/2). \tag{24}$$

This error, as was to be expected, is therefore zero east and west of the range, where the misadjusted north-south pattern falls to zero, and grows to a maximum directly equal to the pattern misphasing in directions due north and south of the transmitter. In case the east-west modulation only is misphased, by P_2 degrees, a similar process gives

$$\rho_2 = P_2 \sin \theta \sin (\theta - P_2/2) \tag{25}$$

as the relation valid for small errors. This error is zero north and south of the range and rises to equal the phase displacement east and west of the range.

If the figure-eight pattern modulation amplitudes are unequal, the north-south one being greater than the average of the two and the east-west one less than the average by the fraction A of the average, the resultant field is

$$e = E_o + E_1 [(1 + A) \cos at \cos \theta - (1 - A) \sin at \sin \theta] \tag{26}$$

which is the same as

$$e = E_o + E_1 \sqrt{1 + A^2 + 2A \cos 2\theta} \cos (at + \phi_2) \tag{27}$$

where

$$\tan \phi_2 = \frac{1 - A}{1 + A} \tan \theta \tag{28}$$

Calling $\phi_2 - \theta$, this time, ϵ, we have

$$\tan \epsilon = - \frac{A \sin 2\theta}{1 + A \cos 2\theta} \tag{29}$$

which for small inequality is

$$\epsilon = - 0.57\, A \sin 2\theta \tag{30}$$

with A in per cent and ϵ in degrees. This is a quadrantal error, the factor $\sin 2\theta$ indicating that ϵ has maximum absolute values for four azimuths, 45, 135, 225, and 315 degrees; ϵ max. is one degree for $A = 1\frac{3}{4}$ per cent, or a total difference between pattern amplitudes of $3\frac{1}{2}$ per cent.

Modulation of the rotation frequency present on the carrier during directional transmission—that is, hum—also produces an error of indication. Hum of modulation depth H per cent and phase G degrees results in a pattern

$$e = E_o \left[1 + \frac{H}{100} \cos (at - G) \right] + E_1 \cos (at - \theta) \tag{31}$$

For low hum levels, this gives an error, in degrees of

$$\eta = - 0.57 \frac{HE_o}{E_1} \sin (\theta + G) \tag{32}$$

This error has maximum magnitude in two directions, bearing $90 + G$ and $270 + G$ degrees, and may be called a duantal error; for a normal directional modulation depth $E_1/E_o = 0.40$, the maximum error is 1.42 H degrees.

Reference keying has a fundamental-frequency component which acts as hum on the carrier. Since the keying comes at maximum north-south directional output, but just at zero east-west output, it acts also to unbalance the two pattern amplitudes and so to produce quadrantal errors of the amplitude inequality type discussed above. For narrow rectangular impulses keeping the signal off for K per cent of the time, the errors are

$$H = 1.14 \frac{KE_o}{E_1} \sin\theta - 0.57\, K \sin 2\,\theta \qquad (33)$$

which varies up to maximum values of 1 degree if K is 1.1 per cent, with $E_1/E_o = 0.40$. Obviously, these two error components may be compensated by injecting broadcast hum and intentionally unbalancing the unkeyed directive patterns by the proper amounts.

Should a balanced modulator have tubes of unequal emission, or for other reasons show an amplitude unbalance, the only effect is to introduce harmonics into the modulation waveform. No error is produced in this case.

D—*Radio-Frequency Conditions*

Radio-frequency effects are calculated much like the modulation-frequency conditions just discussed. However, the complete wave, of which we have only written down the amplitude heretofore, must of course be used explicitly in the radio-frequency analyses. Sometimes intractable forms are encountered and approximate handling is necessary. If the broadcast antenna carrier current is misphased equally with respect to the modulated carrier currents in all four directive-array component antennas, no direct error is produced. But frequency modulation of the radiated energy then takes place and may give rise in a poorly aligned receiver to spurious amplitude modulation and consequent errors. Leakage of unmodulated carrier through a balanced modulator, in phase with the modulated output, causes no error. Quadrature-phase carrier leakage causes a little azimuth-dependent frequency modulation, but no direct errors.

If feed to one directive pair only is misphased with respect to that to the broadcast antenna, by B degrees, frequency modulation is present and a combined quadrantal and octantal error is produced as well, as given when B is small by

$$\beta = \left(\frac{B}{\cos\psi/2}\right)^2 \frac{1}{229 \sin\psi} \cdot \sin 2\,\theta + \left(\frac{B}{\cos\psi/2}\right)^2 \frac{\tan\psi/2}{916} \cdot \sin 4\,\theta \qquad (34)$$

Both B and β are in degrees and $\sin\psi = E_1/E_o$. For $E_1/E_o = 0.40$, a maximum error of 1 degree occurs for a misphasing of 9½ degrees. High-frequency balanced modulators tend to give different output phase from each tube: if one tube of a modulator gives misphasing $+B$ and the other tube $-B$ degrees, the above result for the error produced remains unaltered.

The result of Equation (34) above was obtained by writing down the two directive and the one broadcast wave components of the total radio-frequency field, taking account of radio-frequency phase, and reducing the resultant of the three components to the form of a single-frequency and amplitude-modulated wave. Assuming B small, the radical representing the modulated amplitude was expanded by the binominal theorem and only the first term of the expansion was retained. This rather complicated first term was expressed as a Fourier series, with coefficients calculated by direct integration. The phase of the fundamental-frequency term of the Fourier Series, as a function of θ, was then compared with θ to give the final result of Equation (34). This procedure, while more complicated, is essentially similar to that followed in determining the errors caused by modulation-frequency misadjustments.

The signal from a pair of spaced, non-directive antennas fed with equal and opposite currents results from failure of the two antenna fields to cancel exactly. This failure to cancel is normally caused only by the different time delays for wave travel from transmitting to receiving antenna, which result from one transmitting antenna being nearer to the receiving antenna than the other. But if the contraphased currents in the antennas of a pair are not of exactly equal amplitude, an additional departure from cancellation occurs, and in such a way that the resulting extra radiated field component is in radio-frequency phase quadrature to the desired field. This produces only a negligible direct error for moderate degrees of current inequality, but does give rise to frequency modulation.

The forms of misadjustment causing frequency-modulated radiation to occur give, individually, only small direct errors. The frequency modulation itself is also harmless, unless misalignment or mistuning of the receiver enables it to be converted there to amplitude modulation. But the combined presence of two or more such misadjustments, by permitting out-of-phase field components to combine, may give rise to appreciable direct errors. Therefore, the limits on broadcast carrier misphasing, modulator misphasing or unbalance, B, and antenna pair current inequality are more severe than consideration of each fault separately would imply. If the carrier pattern is misphased with respect to both directive patterns by C degrees while north and south antenna currents differ by M per cent, the resultant error in degrees is

$$\zeta = -0.0045 \frac{\lambda}{d} MC \frac{\tan \psi/2}{\sin \psi} \cos(\theta - 45°) \qquad (35)$$

where $\sin \psi$ is again E_1/E_o. With $E_1/E_o = 0.40$ and antenna spacing, d, of 1/10 wavelength, the error reaches 1 degree when MC is 43.7. This restricts M to 6 per cent when C is 7 degrees, for 1-degree peak error, so is still not very stringent.

Failure of the currents, of equal amplitude, in the antennas of a directive pair to be exactly opposite in radio-frequency phase also modifies the manner of cancellation of their fields. Specifically, if the antennas are closely spaced, the figure-eight pattern amplitude is decreased from its value with correct phasing and a modulated broadcast pattern is also radiated by the pair. If the phase difference between north and south antenna currents is less than 180 degrees by the small angle D degrees, the resulting error, in degrees, is

$$\delta = 0.0044\, D^2 \sin 2\theta + 0.159 \frac{\lambda}{d} D \sin \theta \qquad (36)$$

If $d = 1/10\lambda$, this gives a maximum error of 1 degree for a misphasing of only ⅝ degree. This is the most critical condition encountered, and is of course matched by similar behavior of the other antenna pair. It should be noted, however, that if the error caused by antenna misphasing can be determined, it can be compensated by introducing the proper modulation of the broadcast pattern.

E—*Antenna Geometry*

Directive patterns rigorously following a cosine law are only produced if the push-pull antennas of a pair are infinitesimally separated. The modification of the pattern shape for small finite separations is such as to give rise along the horizon to an error

$$\chi = 23.6 \left(\frac{d}{\lambda} \right)^2 \sin 4\theta \qquad (37)$$

which decreases with increasing angular altitude. This is the well-known octantal error of widely spaced Adcock antennas and reaches 1 degree, at its maxima, for spacing $d = 1/5\,\lambda$.

Unequal spacing of the two pairs merely requires different antenna currents to give equal pattern amplitudes and, so long as those amplitudes are equal, produces no error. Improper orientation of a pair has the same effect as phase displacement, of the same amount, of the modulation of the supply to that pair. Stray coupling between pairs is equivalent to misorienting each pair, while stray coupling between pair and broadcast antennas may be equivalent to broadcast modulation.

An antenna system tilted out of the vertical, but remaining perpendicular to a locally tilted ground plane introduces an error on the horizon

$$\sigma = -0.0044\, S^2 \left[\sin 2Q + \sin 2\,(\theta - Q)\right] \tag{38}$$

for a tilt of S degrees from the vertical in azimuth Q. This reaches maxima of only 1 degree for a tilt of $10\frac{1}{2}$ degrees.

F—*Receiving Indicator Conditions*

At the receiver, incorrect phasing of deflecting voltages applied to the cathode-ray tube produces the same errors as corresponding misphasing of modulating voltages at the transmitter, errors given by Equations (24) and (25). Decentering of the circular cathode-ray pattern with respect to the circular azimuth scale of the indicator displaces the deflection axes of the tube out of line with the cardinal axes of the scale. A simple geometrical construction permits analytical determination of the decentering error which, for small decentering of r per cent of the pattern radius, toward azimuth A of the scale, is

$$\alpha = -0.57\, r \sin\,(\theta + A) \tag{39}$$

degrees. This error resembles that caused by transmitter hum, and has maximum absolute values of 1 degree, at right angles to the direction of decentering, for $1\frac{3}{4}$ per cent decentering. Deflection amplitude inequality produces errors in the indicator similar to those produced by directive modulation amplitude inequality in the transmitted field (see Equation 30).

Cathode-ray tubes of usual commercial construction are not always accurately linear in deflection sensitivity or input impedance. However, if pattern centering, deflection phases, and relative amplitude are given optimum values, all cardinal points will be indicated correctly, primary intercardinal points will be correct on the average and the residual errors at intermediate points will be small. To permit setting all five variables easily and accurately, it is convenient to be able to transmit, when required, a calibrating signal keyed off eight times at equal intervals, rather than once, per cycle. With this signal, one may adjust so that all cardinal and primary intercardinal points are correctly indicated at once. This adds another condition at the transmitter: that the intervals between adjacent keying impulses, during such special transmission, shall each be accurately $\frac{1}{8}$ cycle of the modulation. It also provides another transmitter function, since the seven extra marks can be switched on and off for telegraphic communication, which may be read either visually by the extra marks on the indicator or aurally by the change in keying tone, without disturbing bearing indications.

4—Conclusions

From the foregoing general discussion of the place of radio in navigation, and the description of principles and conditions of operation of one particular radio navigational device, some conclusions may be drawn. Radio-range beacons are inherently accurate and are adaptable to ultra-high-frequency operation. They create no extreme interference problems and divide responsibility equitably between ground and flight personnel. Therefore, the decision of our government a number of years ago, to develop radio ranges as the major radio aid to air navigation in this country, was a very wise one. Recent developments in the ground and airplane direction-finder art have opened new vistas of convenience and flexibility in navigational use of radio. Comparable refinement of radio ranges is needed to enable them to retain their outstanding position in the face of these new and attractive alternatives. The impending change to ultra-high frequencies permits a free choice now of the course which development shall take. Indistinguishability and restricted number of courses with presently used ranges point clearly to possible lines of development.

The rotating beacon principle is well suited to the provision of a type of radio range more flexible and convenient than those now in use. By employing basically simple electrical means to eliminate mechanical motion altogether, limitations on speed of rotation are removed. This permits automatic, direct indication of the position of a receiver anywhere around the range beacon, or of deviation of the receiver from a chosen line through the range. In effect, a radio range course is available wherever wanted, with no ambiguity. With such an "omnidirectional radio range" the indicating means, like the beacon, are basically simple.

Many conditions must be fulfilled for ideal, error-free omnidirectional range operation. Analysis of imperfect fulfillment of these conditions shows errors resulting from small imperfections to be moderate in amount. Tolerances on adjustments are therefore reasonable. Great flexibility of strictly electrical beacons and indicators permits provision of convenient adjustments for fulfilling all necessary conditions for all-around accuracy.

A system which combines the advantages of the range beacon with the convenience of its new alternatives emerges from these considerations. An experimental model of such a system and the results obtained with it, as well as ways of using it in flight, will be described in a further paper.

NBC SHORT-WAVE LISTENING POST

By

G. O. Milne

Eastern Division Engineer, National Broadcasting Company, Inc.

Summary—The reasons for establishing a central point for the monitoring of international broadcasting stations are given. The physical arrangement of the equipment used is described, and because of the peculiar demands placed on men working in this service the problems encountered in obtaining necessary personnel are discussed.

THE National Broadcasting Company's short-wave listening posts are located in the NBC Newsroom in RCA Building, at 30 Rockefeller Plaza in New York City, and at Bellmore, L. I. The listening post was first established as an experiment on April 1st, 1940, operating two receivers during the late afternoon and early evening hours. Its importance has grown such that the service has been expanded to include eight receivers and twenty-four hour coverage. The Staff, under the direction of Mr. A. A. Schechter, Director of the News and Special Events Division, is headed by Mr. J. Van Item, as Supervisor, and includes five interpreters and four junior engineers. The men work in pairs, composed of one linguist and one engineer per eight-hour shift.

As the European War began to spread, it became necessary to obtain up-to-the-minute news from Europe for use on the NBC News broadcasts. As a trial, two receivers were installed and monitoring of the news periods in English from foreign short-wave broadcasting stations was observed. Several instances during the invasion of Holland and Belgium proved the value of this service, but indicated that it would have to be extended to cover more languages than English alone. We, thereupon, started on the two problems that had to be solved to enable us to provide a complete service. These two were first—as efficient a technical plan as could be had at the location available and second, the employment of linguists capable of using these facilities to best advantage.

It was early indicated that facilities should be as near as possible to the NBC news desk. This was necessary in order to pass along with an absolute minimum of delay, any information received concerning the rapidly changing events in Europe. Usable information received

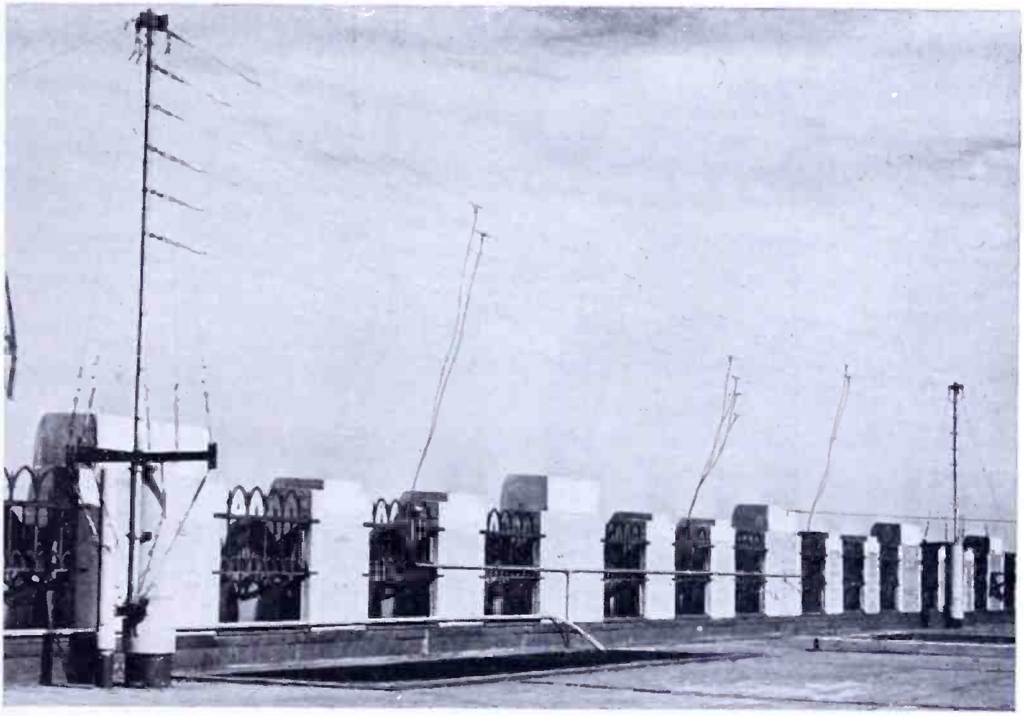


Fig. 1

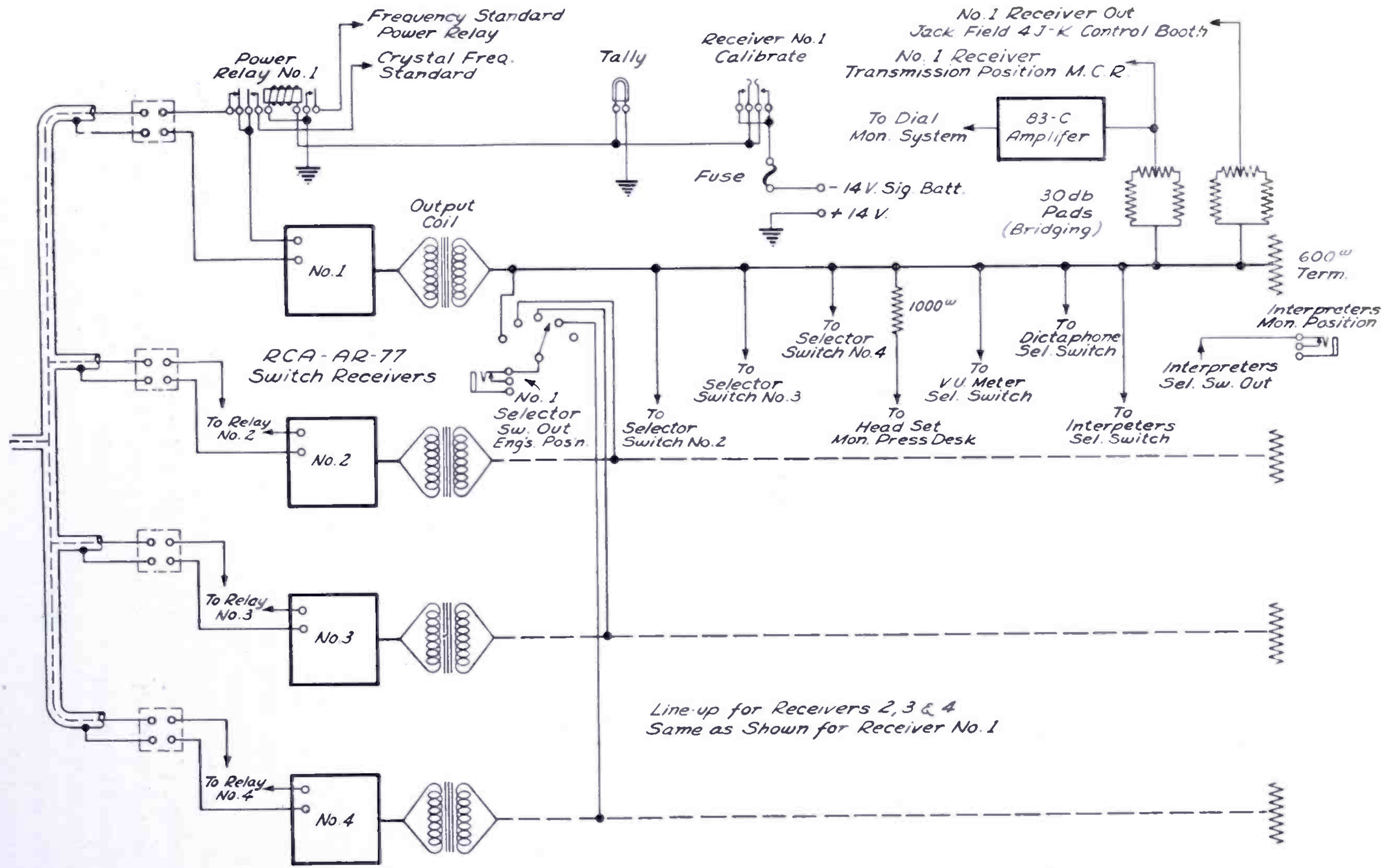
was broadcast instantly through the NBC networks as well as placed on the teletypewriters to the major news associations.

The Rockefeller Center Organization had installed on the roof of the International Building a standard RCA multi-wave antennaplex system, Figure 1. This system gives very efficient coverage on the following bands:

15.4 - 18.1 Mc
12.7 - 15.4 Mc
10.0 - 12.7 Mc
7.3 - 10.0 Mc
4.6 - 7.3 Mc
2.0 - 4.6 Mc
0.53 - 2.0 Mc

Separate antennas and wide-band amplifiers are used for each of the frequency bands. The antennas are half-wave doublets arranged for use with low-impedance two-wire transmission lines. Reference to the antenna photograph will show that the various horizontal doublets are disposed vertically. It should be noted that this arrangement allows for convenient maintenance of the system. A two-wire transmission line of about 100 feet in length is brought from each doublet to the RCA multi-wave antennaplex amplifier which is located in a fan room on one of the upper tiers.

The length of the coaxial cable which starts at the antennaplex unit



SHORT WAVE MONITORING FACILITIES
NBC LISTENING POST

Fig. 2

near the top of the International Building and runs to the NBC newsroom in the RCA Building is about 1500 feet. The antennaplex unit, as it was installed, did not have enough gain on the high-frequency bands to make up for the loss in the coaxial cable. An auxiliary amplifier was added which contained two more amplifier stages for the 9, 12, and 16-megacycle bands.

Four RCA AR-77 receivers mounted in a special table are used for listening. An input switching system is provided so that the receivers can be fed with a series of calibrating radio frequencies.

Since only two men are on duty at one time, it was deemed advisable to provide an audio-switching system whereby either or both men could monitor the output of any one of the receivers without moving from his normal position, Figure 2. Also, because at times there might be as many as three simultaneous news transmissions to be observed, special provisions were installed enabling both men to monitor one receiver each while the third was fed into a recording machine. Selector switches were provided at each of the receiver positions for the engineer. The interpreter's normal position, as can be seen in the photograph of the complete listening post, Figure 3, is at the left end of the table. Here he has one selector switch connecting him to the output of any receiver. He also has a selector switch by means of which he can feed the output of any receiver into the recording machine. Facilities are also provided so that the output of any receiver may be fed directly into the NBC networks through the adjoining news studios, one of which can be seen in the photograph. This has been done on a few occasions when the monitors, tuning over the various bands between scheduled news periods, came upon an event of great enough importance to be put on the air immediately. Ordinarily, the news broadcasts from foreign countries retransmitted over the networks, are received in the United States by regular communication companies. The material received by the monitoring post is used if warranted in domestic news broadcast periods.

The outputs of all receivers are also permanently connected to the dial monitoring system in all offices and studios of the National Broadcasting Company. It is thereby possible for any point equipped with the dial monitoring system and a loudspeaker to listen to any of the channels received by the listening post. Since the input to the recording machine, the input to the broadcasting circuits, and the input to the house monitoring system must be kept at a constant level, a VU meter is available on a selector switch at the engineer's position. Thus, maximum flexibility of control of the entire short-wave monitoring facilities has been provided. The receivers are in use 24 hours a day and have given excellent service.

The duty of the engineers assigned to this service is to have the receivers tuned to the proper stations at the required time and to assist in the copying of reports. They are also constantly tuning throughout the entire radio-frequency spectrum looking for new international broadcast stations. At the outset, engineers were assigned to this service when they were free from regular broadcasting assignments. At present, however, there are four men assigned exclusively to this service as it was found that by monitoring regularly a kind of sixth sense is developed that enables a man to understand weak or noisy signals which one assigned only occasionally could not decipher. International broadcasting stations today are operating as close as five kilocycles and in some cases only three kilocycles apart. It is no longer

Fig. 3

possible to be certain that one knows the nationality of the station by the language in use, as so many of these stations use many languages, the British at present transmitting in twenty-seven different tongues. Only an experienced operator can be reasonably sure that he has the proper station tuned in at the required time.

Difficult as was the technical problem, that of securing the services of qualified monitors was even greater. Since the profession of news monitoring was comparatively new, very little was known concerning the qualifications of a monitor, with the result that a large number of applicants presented themselves merely on the strength of their knowledge of languages. The chief difficulty in finding duly qualified monitors lay in an apparent contradiction. The monitor must be fairly young in order to stand up under the incessant pounding he receives by listening through earphones day in and day out, and at the same time, his knowledge of languages must be such as is usually encountered only in persons of advanced age.

The following qualifications were finally determined:

Must be American-born or a naturalized citizen of long standing, a college graduate or equivalent. Must know thoroughly at least four languages: English, French, German, Spanish (or Italian). Some newspaper or writing experience essential. Must have European educational background, or have lived in Europe for a great number of years, in order to speak at least German and French like a native, so that he understands dialects, patois, colloquialisms, idioms, etc. Must be thoroughly familiar with geography, international affairs, political situations, foreign prejudices, customs and opinions, minority problems and similar factors.

While these requirements may seem rigid, such a background was necessary in order to properly evaluate news items, talks, special features, etc., as distinguished from the wealth of pure propaganda which was transmitted by all foreign short-wave broadcast stations. After several months of trial and error, the present staff was assembled.

Between 650 and 700 individual reports are written by the monitors each month. Quite frequently the reports of the English, German, and Italian broadcasts of a single important incident such as the bombing of a city or sinking of a battleship are recorded at different periods and then written into one report in an effort to obtain from them all, as nearly as possible, the true story. Recently, transmitting stations have been opened in various parts of the world, such as Japan, Indo-China, Belgian Congo, Greece, and Eire of sufficient power to be received regularly. To these can be added Australia, South Africa, Turkey, French Equatorial Africa, Persia, China, all of South America as well as all the major countries of Europe. Thus, through the short-wave listening post the National Broadcasting Company is in constant touch with the news of the world. The material received is used daily in the making of the news bulletins broadcast on the networks. The success of the plan is also attested by the fact that a monthly average of 150 of the reports written are used by the news associations in their releases.

Of the several news "scoops" received by this group, the outstanding one was the speech of Marshal Petain, announcing France's capitulation. Regular communication channels with France were jammed and delayed and radio reception was very poor. After playing the recording of the speech over twelve times, a complete transcript was obtained and broadcast and sent to the newspapers long before its receipt by regular communication channels. With the present war situation expanding throughout the world, the listening post takes on added importance in the proper transmission of the important news of the world to our American listening audience.

DUPLEX TRANSMISSION OF FREQUENCY-MODULATED SOUND AND FACSIMILE

By

MAURICE ARTZT

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

AND DUDLEY E. FOSTER

RCA License Laboratory, New York

Summary—*Laboratory and field tests of multiplexing aural and facsimile programs on a frequency-modulated wave have been made. The results obtained show that while technically possible, difficulties are encountered that will probably make such operation undesirable. Enough of the theory of multitone modulation is discussed to show that circuit linearity requirements are severe, and that the side bands can easily exceed the 200-kilocycle channel width if the aural deviation is not reduced. Quantitative data are given on the types and percentages of cross-talk obtained.*

1. INTRODUCTION

THE F.C.C. rule defining requirements of high-frequency, aural and facsimile broadcasting stations is as follows:

"S 3.228 Facsimile Broadcasting and Multiplex Transmission. The Commission may grant authority to a high-frequency broadcast station for the multiplex transmission of facsimile and aural broadcast programs provided the facsimile transmission is incidental to the aural broadcast and does not reduce the quality of or the frequency swing required for the transmission of the aural program. The frequency swing for the modulation of the aural program should be maintained at 75 kilocycles and the facsimile signal added thereto. No transmission outside the authorized band of 200 kilocycles shall result from such multiplex operation nor shall interference be caused to other stations operating on adjacent channels."

Little practical experience has yet been obtained with multiplex operation of the kind called for in this regulation. For this reason it has seemed worthwhile to report the results of some laboratory and field tests carried out by RCA in Camden and New York. These tests were made with the object of determining first, if any limitations and practical difficulties were likely to be encountered, and second, to what extent the introduction of multiplex facsimile signals might effect the design of receivers for aural service only.

It is generally conceded that in such a multiplex system the signal-to-noise level in the aural program should be favored as much as possible. In order to maintain and protect adequately the high signal-to-noise ratio of the aural service, the facsimile signal must use a much lower deviation. The ability of the facsimile equipment to operate on a much less favorable signal-to-noise ratio then becomes increasingly important. The facsimile equipment used in these tests was designed to operate under conditions heretofore considered too severe for good reproduction, and the final results obtained in the multiplex operation should therefore be considered on the basis that all possible favoring has been given to the aural channel. When used with a facsimile system requiring a higher signal-to-noise ratio, the facsimile would in general become less reliable and reproduction poorer.

Practical difficulties were encountered in the field tests described later that limited the frequency deviations used for the aural portion of the multiplexing to 60 kilocycles. The facsimile service was added at 15 kilocycles deviation, thus sharing the maximum deviation of 75 kilocycles between the two services. Many things contributed to this limitation of aural deviation and each will be described in turn.

Before going to the tests and their results a description of the facsimile signals used in these tests will be given. A general discussion of the frequency-modulation theory for multitone use is also in order if the proper interpretation of the results is to be made.

2. The Facsimile Signals

In the usual broadcast facsimile services contemplated or now in use, copy areas of from four to twenty square inches per minute are transmitted. A good average at present would be 10 square inches per minute, the figure chosen for these tests. This was transmitted at 120 scanning lines per inch, using a picture signal frequency range of from zero to 1200 cycles. This signal can be transmitted in a variety of ways, the usual one being to carry the signals as an amplitude modulation of an audio tone. However, previous facsimile work has shown that the use of a frequency-modulated sub-carrier for the facsimile signals has many advantages over the amplitude-modulated sub-carrier.[1] The sub-carrier chosen in this case was 18 to 24 kilocycles, the picture signals occupying the band of 20 to 24 kilocycles and the frequency being shifted down to 18 kilocycles for a short time once per scanning line to indicate phasing. The spectrum then is as follows:

18 kilocycles—phase signal
20 kilocycles—white picture signal
22 kilocycles—50/50 gray picture signal
24 kilocycles—black picture signal

The use of a phase signal to hold synchronism of the recorder is usually insufficient if very fine detail of recording is expected, and a

[1] Mathes and Whitaker, "Radio Facsimile by Sub-Carrier Frequency Modulation," RCA Review, Oct. 1939.

secondary synchronizing signal to control motor speed between phase pulses is desirable. A simple method of accomplishing this is to transmit a control tone (60 cycles in this case) as an amplitude modulation super-imposed on the already frequency-modulated sub-carrier. An amplitude-modulation of 30 per cent was used in this case, and the 60-cycle wave is thus transmitted without requiring a third multiplex channel and attendant filtering problems.

This complex wave of 18- to 24-kilocycle frequency variation and 30 per cent amplitude modulation then carries the picture, phasing, and synchronizing signals. When used in a simplex service and with the proper limiters and filters, these signals have been found to be practically immune to any normal signal fluctuations, and to reject all

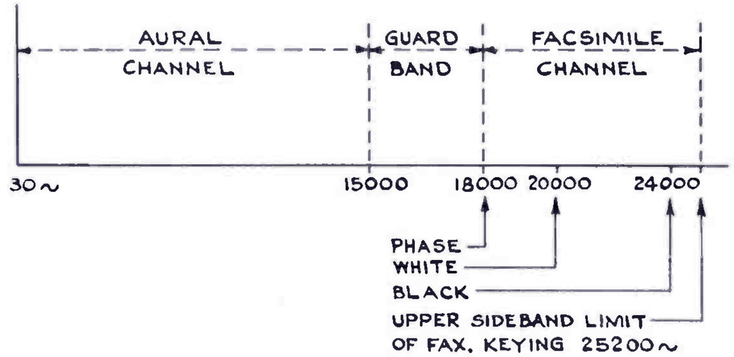


Fig. 1—Audio spectrum for multiplexed aural and facsimile.

noise 6 decibels or more below the signal itself. Occasional interferences equal to or even greater than the signal usually cause little marring of the copy.

This composite signal was then used to deviate the r-f carrier 15 kilocycles, the aural channel being added at 60 cycles deviation. The overall audio spectrum transmitted is then represented in Figure 1.

3. Multiplex Theory

In tone multiplexing on an amplitude modulated system, the addition of the second tone produces an additional pair of side bands displaced from the central carrier by the tone frequency itself. All side bands are proportional to the carrier amplitude and depth of modulation. In multitone frequency modulation the relationship of side bands is not as simple as this, for each tone in itself produces an infinite series of side frequencies. The case for two tones involves two such series of side frequencies, and their cross beat products. A derivation of the side bands produced in such a two-tone system is given by Crosby[2] and can be readily applied to this type of multiplexing.

[2] Murray G. Crosby, "Carrier and Side Frequency Relations with Multitone Frequency or Phase Modulation," RCA Review, July 1938.

Applying the Bessel function expression to the facsimile signal only, with no aural signal, gives the plot in Figure 2 of the relative amplitudes of carrier and side bands to transmit this signal with deviations up to 25 kilocycles. It can be seen that for deviations below 15 kilocycles the second-harmonic side band (J_2) is small and can probably be neglected without seriously affecting the facsimile signal itself.

The radio-frequency spectrum for the facsimile signal when operating simplex is shown in Figure 3. Suppose now that an aural tone is introduced and with such a low-frequency period that any small section of the wave may be considered as constant in frequency com-

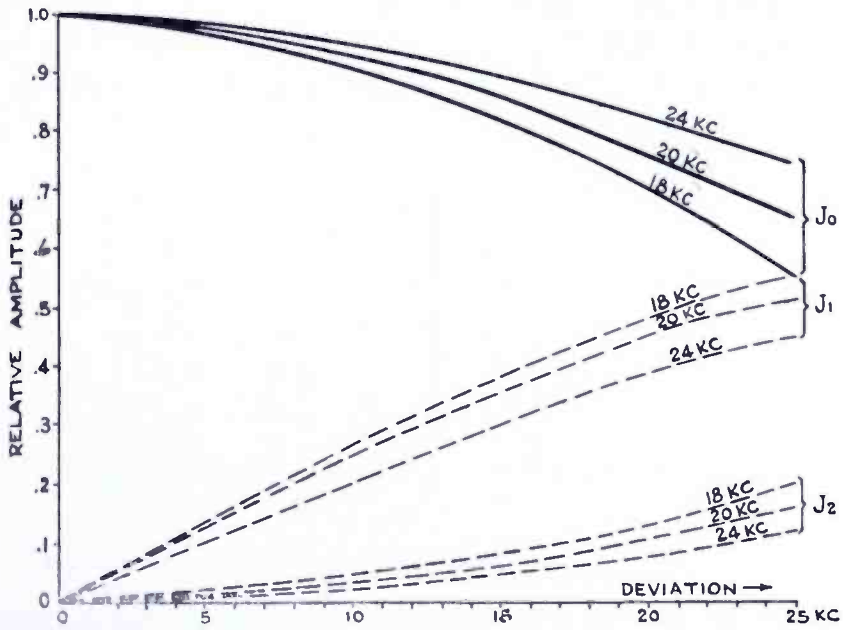


Fig. 2—Facsimile sideband amplitudes (No aural).

pared to the rapid deviation caused by the facsimile signal. Figure 4 will then represent the wave. This is strictly true only when the aural tone introduced has zero frequency, for frequency modulations cannot be added directly; but for frequencies in the aural band up to several hundred cycles there is very little error. If only the first-harmonic side bands of the facsimile signal (± 20 kilocycles) are considered, the band width under this condition will be ± 60 ± 20 or ± 80 kilocycles. If the second-harmonic side bands of ± 40 cycles are included, the width is ± 100 kilocycles.

With an audio frequency of 3125 cycles having a frequency deviation of 75 kilocycles applied alone to the frequency modulator the spectrum appears as in Figure 5 for frequencies on one side of the carrier, the side bands for frequencies on the opposite side of the carrier being similar.

At an audio frequency of 2500 cycles the resultant argument (Deviation divided by modulating frequency) comes within the range of the usual Bessel tabulations, and the sidebands of the multiplex modulation can be determined. Thus with a 2500-cycle tone at 60 kilocycles deviation and the 20,000-cycle tone at 15 kilocycles deviation the amplitudes can be plotted as in Figure 6. The first chart gives the side bands of the audio tone only, and the second the side bands for the combination of both tones. In both cases all side bands and side frequencies are plotted as positive quantities, but due account of sign of the coefficients has been taken where several terms of the expansion are involved for a particular frequency. In these charts any coefficient greater than 0.1 per cent was included.

It can be seen that the bandwidth in Figure 6 is somewhat greater than in Figure 5, but that the side bands outside the 200-kilocycle channel have less than 1 per cent amplitude.

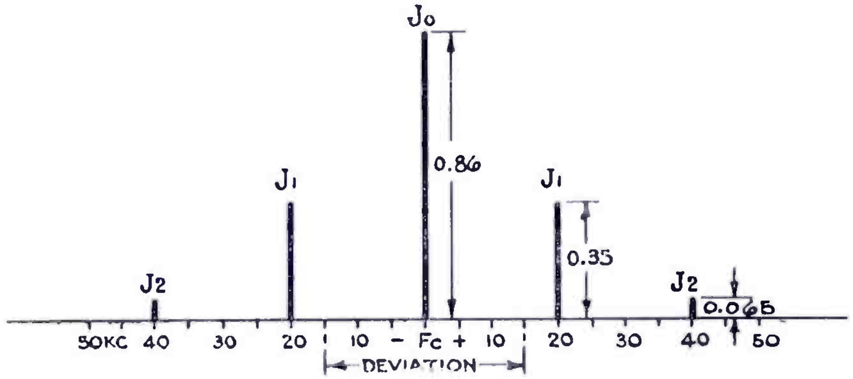


Fig. 3—Sideband chart of 20-kc signal at 15-kc deviation.

By comparing Figure 4 with Figures 5 and 6, it may be seen that the low aural frequencies cause a greater side band extension than the higher aural frequencies. For the higher frequencies, the product of the Bessel coefficients reduces the amplitude of the higher order side bands to a degree where the bandwidth of consequence becomes less than for a single modulating tone.

The 60-kilocycle deviation for aural signals was used in the multiplex calculations rather than 75 kilocycles, in order to keep the signal within the 200-kilocycle band width. As can be readily seen from Figure 5, maintaining the aural deviation at 75 kilocycles and using the 15-kilocycle facsimile deviation would produce side bands outside the 200-kilocycle channel. Reducing the facsimile deviation to 10 kilocycles would lower the amplitude of the out-of-channel side bands to 64 per cent of their former value. This reduction would be insufficient to clear up the out-of-channel radiation, and furthermore would handicap the facsimile in signal-to-noise ratio. Thus Figure 5 shows the first practical reason for reducing the aural deviation to 60 kilocycles when multiplexing with facsimile.

Before going to the experimental results it would be in order to point out the types of cross-talk that can occur if the equipment fails to pass this entire band as represented in Figure 6. If the bandwidth of the single-tone transmission is restricted at the extreme ends, only harmonic distortion of the resulting detected tone will be present, and a few per cent of distortion of this type will be little noticed. This same amount of restriction on the two-tone transmission will result quite

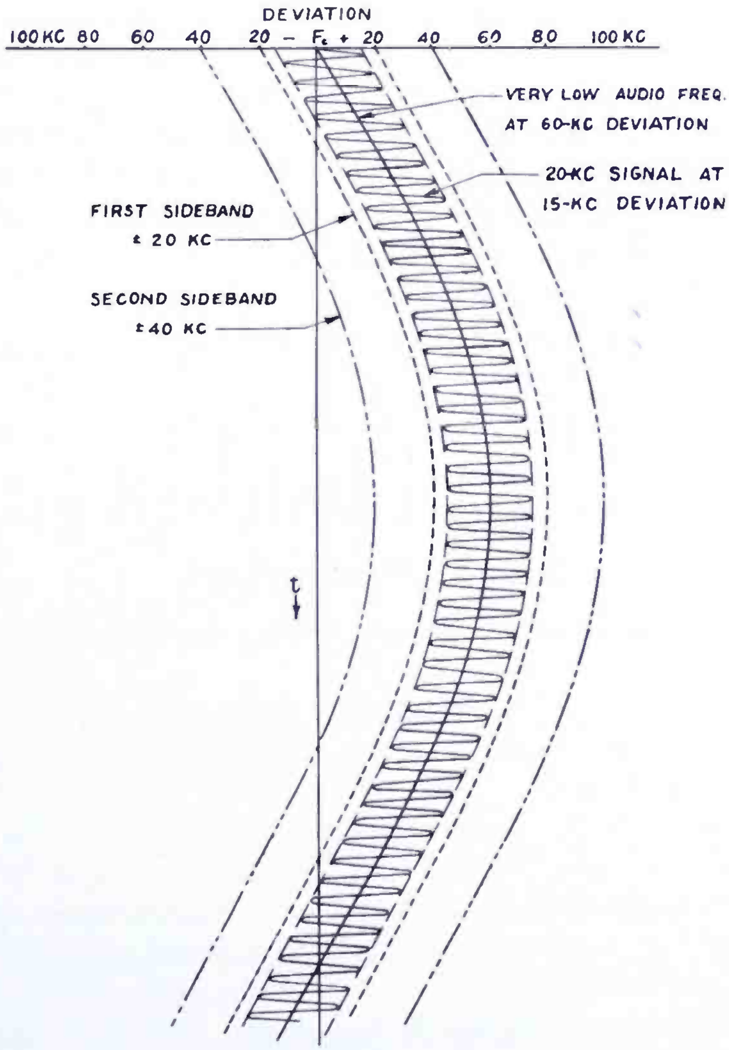


Fig. 4—Facsimile signal of Figure 3 added to very low frequency signal at 60-kc deviation.

differently, for such band limiting will act as a preliminary frequency discrimination and change various side-band frequency modulations to amplitude modulations. This will introduce cross-talk in a manner similar to the action of a non-linear tube in amplitude multiplexing. It should be pointed out that a small percentage of cross-talk is far more objectionable than the same percentage of harmonic distortion.

The process of detecting a frequency-modulated wave requires differentiating[3] the wave to change the frequency variations into an amplitude-modulated wave. Amplitude detection follows this differentiation (or discrimination) to obtain the amplitude envelope. If the waves represented by Figures 4 and 6 are differentiated, the original modulation frequencies are obtained intact. However, if one or more of the terms in these side frequencies are eliminated before differentiating, the differentiated (or discriminated wave) tone will contain terms proportional to the derivative of these missing terms. Any two of these waves can then beat together in the amplitude detector that follows this discriminator to give cross talk.

Bandwidth limitation could occur in the transmitter itself, but the more likely place will be in the receiver intermediate-frequency amplifier or discriminator circuits. This is especially true in receivers hav-

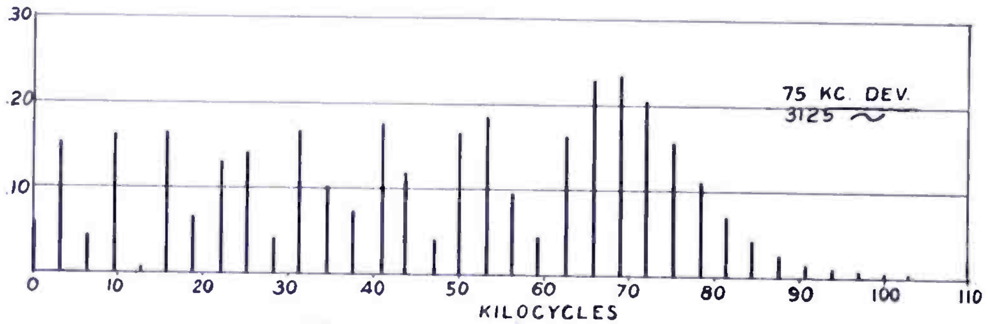


Fig. 5—Sideband amplitude.

ing good adjacent channel selectivity and consequent narrow band-pass action in the i-f transformers.

After detecting, the requirements of linearity in any tube that both aural and facsimile signals pass through together are the same as for amplitude multiplexing.

In the case of two modulating frequencies which differ in character as greatly as do those of sound and facsimile programs the linearity requirements imposed on the system are much more severe than in the case of single program modulation. With a single program, any non-linearity of the modulator, the detector, or the audio-frequency portion of the receiver causes harmonic distortion. Distortion may also be caused by too narrow a receiver passband. This latter distortion will be decreased appreciably, but not entirely eliminated by a limiter. All of these causes of distortion may be classed as non-linearity.

The distortion which results from non-linearity in the case of a single program, if not too great, is not readily distinguished by the ear. For example, there are many radio receivers having 3 per cent or even 5 per cent distortion which are not objectionable to the major-

[3] Carson and Fry, "Variable Frequency Electric Circuit Theory With Application to the Theory of Frequency Modulation," *Bell System Tech. Jour.*, Oct. 1937.

ity of listeners. In the case of modulation by two different types of program, however, non-linearity anywhere in the system results in combination tones or cross modulation. The cross modulation of one program by the other has the characteristics of interference which is readily noticed by the listener. An interfering program 30 db below the desired program is easily apparent, whereas distortion of a single program 30 db below (approximately 3 per cent of) the fundamental, is seldom noticed.

Consequently, duplex frequency-modulation transmission of sound and facsimile requires a somewhat greater bandwidth than a sound

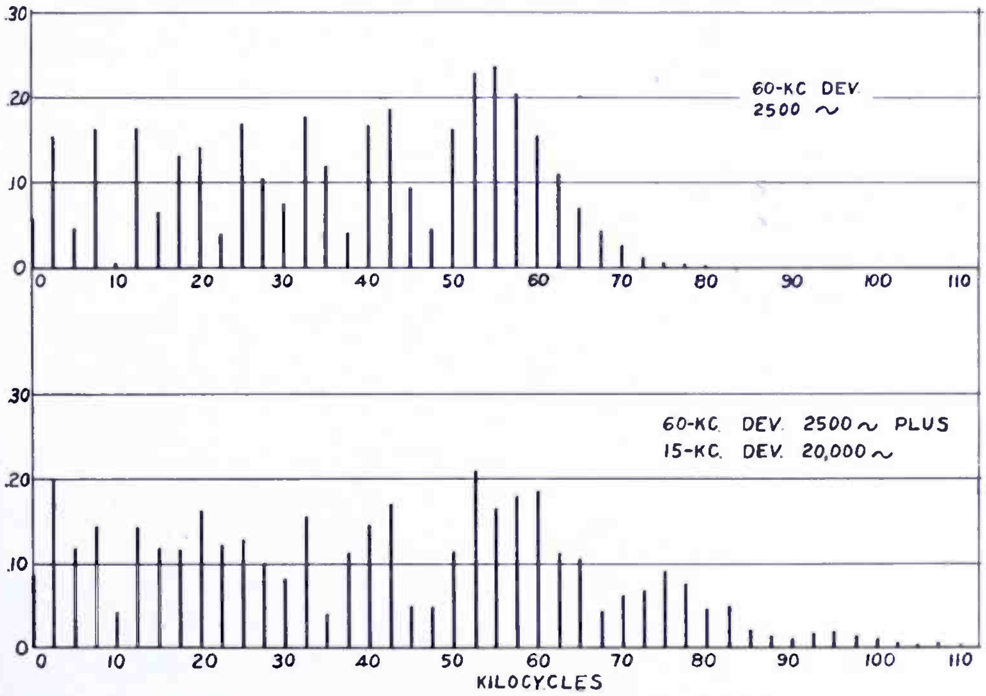


Fig. 6—Sideband amplitude.

program alone of comparable deviation, particularly for low modulating frequencies, and imposes much more stringent requirements on system linearity.

4. Duplex Receiver

The receiver for duplex transmission of sound and facsimile, in addition to the requirement of exceedingly linear circuits, must be provided with filters for separating the two programs following the detector.

A block diagram of a receiver constructed for tests of duplex transmission is shown in Figure 7. Two selective circuits were used ahead of the converter and an i.f. of 8.25 Mc was employed to minimize likelihood of spurious responses. The i-f passband was made somewhat wider than normal to guard against non-linearity from that source.

It had a response characteristic down 14 db at 100 kc each side of the center and down 37 db at 200 kc each side of the center.

The detector characteristic was more linear than is usually the case, having very small departure from a straight line up to 100 kc on each side of the center.

It was found necessary to separate the sound and facsimile programs immediately following the detector, as passing both through a single common amplifier stage, even a fully degenerated stage, resulted in excessive cross modulation. The de-emphasis filter in the output of the detector attenuated the facsimile frequencies sufficiently so that cross-modulation did not occur in the succeeding stage. An additional low-pass filter was used following the first audio-frequency

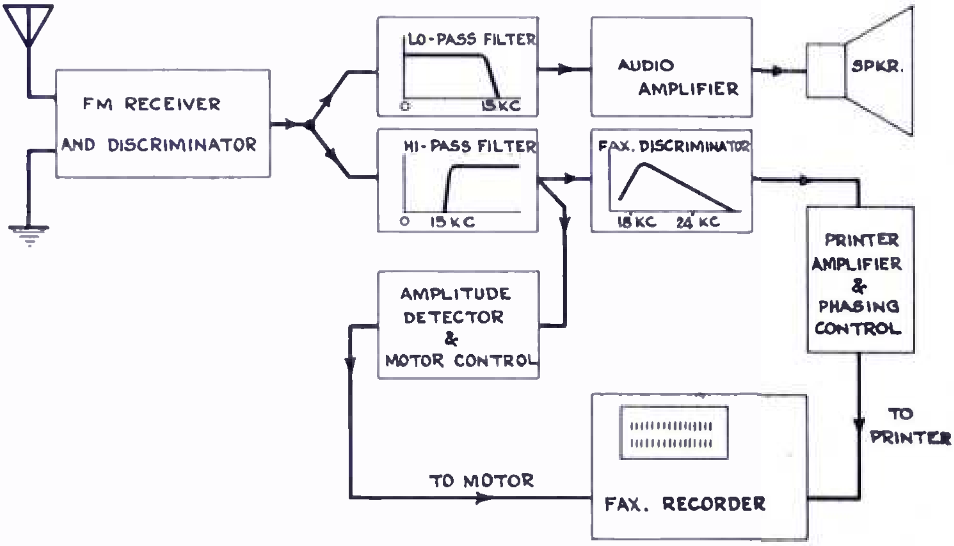


Fig. 7—Receiver block diagram for multiplex operation of aural and facsimile.

amplifier since the impedance at that point could be made low by use of a cathode load and a better filter designed for the resultant low impedance than for the high impedance at the output of the detector. The audio program was thus subjected to proper compensation for the high-frequency pre-emphasis at the transmitter and also had an additional filter to attenuate further any facsimile frequencies in the sound channel. The resultant attenuation for facsimile frequencies in the sound channel was greater than 45 db for even the lowest (18 kc) facsimile frequency.

High-pass filters to attenuate the sound program in the facsimile channel were provided in the facsimile equipment.

A high-quality, Type 64B speaker was used with the receiver and was fed by a pair of 6L6G tubes in push-pull so that the reproduction of the sound program was of high fidelity.

5. Laboratory Tests of Duplex Transmission

The first laboratory tests were of a preliminary nature and used a special receiver (with audio circuits similar to those in Figure 7) and a signal generator set at 43.5 megacycles. The facsimile signal was as previously described, and the aural channel was fed by a beat oscillator or from a phonograph pickup. In these tests the receiver bandwidth was slightly higher than 200 kilocycles and good results were expected with the full aural deviation of 75 kilocycles and a facsimile deviation of 15 kilocycles. A tabulation of the results obtained follows:

Audio Freq.	*Audio Deviation*	*Facsimile Deviation*	*Interference*	
			On Aural	*On Facsimile*
10 kc	75 kc	15 kc	— 25 db	Prints Interference Synch. Erratic
5 kc	75 kc	15 kc	— 69 db	Prints Interference Synch. Erratic
400 cy	75 kc	15 kc	— 70 db	None Noticeable
*400 cy	75 kc	15 kc	— 58 db	Prints Interference Synch. Erratic

* Receiver detuned 18 kc.

These results show that only at the low frequencies of aural tone (400 cycles in this case) were the cross-beat products low enough substantially to eliminate all cross-talk. When high aural frequencies were used (5 and 10 kilocycles), the combination frequencies increased enough to cause deterioration of the signal-to-noise ratio in both channels. It was to be expected that the facsimile channel would become unusable first, for the cross-beat interference tones were of a high frequency and would come through the high-pass filter network feeding the facsimile amplifiers.

Two general observations can be made from these tests: First, the use of pre-emphasis of higher aural frequencies will occasionally bring enough interference into the facsimile channel to mar the recorded copy if the full 75-kilocycle deviation is used for aural programs. Second, only a very wide band receiver could receive such a signal combination without cross-talk. Any reduction in bandwidth in the receiver will aggravate the interference in either of the above cases and a reduced bandwidth would probably be used in the average receiver to obtain a fair degree of adjacent channel selectivity.

The most apparent cure for these difficulties was to reduce the aural deviation to 60 kilocycles, and this was done in all subsequent tests.

A further test was made in the laboratory to simulate duplex transmission of sound and facsimile. In this case a 4000-cycle tone was used for the audio frequency and a 17,000-cycle tone for the facsimile frequency.

Simultaneous application of two tones to a system having any non-linearity results in combination tones that are the sum and difference of multiples of the individual tones. For example, if two tones having frequencies a and b are applied to a non-linear element, the result will be frequencies $2a$, $2b$, $3a$, $3b$, $2a - b$, $2a + b$, $3a - b$, $3a + b$, $2b - a$, $2b + a$, $3b - a$, $3b + a$, etc. In general, in a non-linear system the coefficients of the higher-order terms are much smaller than those of the lower-order terms so that the greatest amplitude of resultant combination tones is likely to occur with a high audio frequency and a low facsimile frequency. However, audio frequencies much above 4000 or 5000 cycles seldom have sufficient energy, even when pre-emphasis is used, to modulate the transmitter fully. Consequently interference between sound and facsimile is most likely for audio frequencies in the 2000 to 5000-cycle range.

The test was performed by modulating a laboratory frequency-modulation signal generator by 4000 cycles and 17,000 cycles simultaneously and measuring the resultant frequencies in a dummy load replacing the voice coil in the output of the receiver by means of a wave analyzer. Throughout the series of tests the input to the receiver was 50 microvolts and the output was held constant by means of the volume control at 0.5 watt. In all cases, except where otherwise noted, tuning was performed by applying approximately 150-kc deviation and the receiver tuned to symmetrical distortion of the top and bottom of the wave as observed on an output oscilloscope.

As an initial test, the 4000-cycle tone alone was applied and the harmonics read as the deviation was varied.

TABLE I

4000-cycle modulation

Deviation	*2nd*	*3rd*	*4th*
75 kc	7.5%	3.1%	0.25%
60	3.6	1.2	0.14
40	1.98	0.62	0.13
20	1.85	1.1	0.10

With the exception of the second harmonic with 75-kc deviation which was believed due to the signal generator curvature, the harmonic content was low. In any event the second harmonic would be due to a second-order term and with 4000- and 17,000-cycle tones, the second-order term would give a 13,000-cycle resultant, and the ear is much less responsive to that frequency than to frequencies of the order of 1000 or 2000 cycles. If we designate the 4000-cycle tone as a and the 17,000-cycle tone as b, the resultant tones up to the fifth order, would be $b - a$, $b - 2a$, $b - 3a$, $b - 4a$, or 13,000 cycles, 9000 cycles, 5000 cycles and 1000 cycles respectively. The amplitude of these resultant frequencies was measured in the output of the receiver with the two tones modulating the signal generator.

In the second test, shown in Table II, the 4000-cycle tone had a constant deviation of 60 kc and the deviation of the 17,000-cycle tone was varied from 5 to 15 kc. The amplitude of 4000-cycle output was taken as 100 per cent.

TABLE II

4000 cycles; constant deviation, 60 kc

17,000-cycle Deviation	*4000*	*1000*	*5000*	*9000*	*13,000*
15 kc	100%	4.0%	5.1%	18.5%	1.0%
10	100	2.1	5.0	15.0	1.0
5	100	3.1	5.1	13.7	1.5

Variation of deviation of the 17,000-cycle component is seen to have little effect on most resultant frequencies.

The results when the 17,000-cycle tone deviation was kept constant at 15 kc and the deviation of the 4000-cycle tone varied are shown in Table III.

TABLE III

17,000 cycles; constant deviation, 15 kc

4000-cycle Deviation	*4000*	*Output frequencies, cycles*			
		1000	*5000*	*9000*	*13,000*
60 kc	100%	3.1%	5.1%	13.7%	1.5%
40	100	1.9	4.6	12.2	1.1
20	100	1.2	1.2	8.3	0.46

Decreasing the deviation of the 4000-cycle component is seen to decrease materially the magnitude of the resultant components.

With 60-kc deviation of 4000 cycles and 15-kc deviation of 17,000 cycles applied, the receiver was then tuned to minimize the audible resultant tones by listening. This tuning was found to be very critical and to depart by about 20 kc from the tune position determined with a single modulating tone. The resultant outputs were then:

4000 cycles	*1000*	*5000*	*9000*	*13,000*
100%	0.62%	3.0%	9.7%	4.2%

This method of tuning the receiver is seen to reduce greatly the resultant components, especially 1000 cycles, which was the most prominent in listening. Even with the receiver so tuned, the output to the ear was readily evident as being different from pure 4000 cycles.

The results of these laboratory tests confirm the theoretical considerations that two tones applied to a system having any non-linearity whatsoever will result in combination frequencies exceeding in amplitude the harmonics which occur with a single modulating tone. When these two tones differ widely in frequency, the low modulating frequencies, while producing greatest side-band extent, produce combination tones outside the audible range, hence are less disturbing to the sound program. The higher-frequency audio tones, combining with the facsimile frequencies, however, do produce serious interference in the sound channel.

6. Field Tests

A series of tests was run on the system of duplex sound and facsimile on W2XWG (W51NY) frequency-modulation transmitter located on the Empire State Building in New York City during January 1941. The receiver used in the tests was the one described above and was tried at three different locations in Radio City, about one mile from the transmitter.

The facsimile and sound programs were applied in parallel to the grid of the reactance tube modulator of the f-m transmitter and were independently monitored as to deviation. The response of the facsimile system was flat from scanner to transmitter within 0.5 db up to 24 kc.

With careful tuning of the receiver and limitation of the deviation to 60 kc for the sound and 15 kc for the facsimile, substantial freedom from interference between the two was possible.

Even with these conditions, a sound program with prominent sibilants produced a short skip (white space) in the facsimile copy for each "s" sound, and during pauses in sound programs the facsimile could be heard in the background.

If the receiver was not precisely tuned, little increase in interference with facsimile was apparent, but the facsimile background became markedly evident in the sound reproduction.

During the progress of the test one practical operating factor was noticed that illustrated the interdependence of sound and facsimile in duplex transmission. There was a tendency for unwarned individuals to tune to a different sound program for check purposes while receiving facsimile. This caused the facsimile to drop out of synchronism which had to be regained upon returning to W2XWG, so that about ¼ inch of copy was lost for even momentary tuning to a different station.

7. Conclusions

The laboratory and field tests indicated that the frequency-shift facsimile method was capable of producing good copy and could be synchronized by a simultaneous amplitude modulation of the picture frequency modulation. This resultant signal could also be used to frequency-modulate a transmitter at the same time as a sound program to give duplex transmission. The required spectrum space for duplex operation is greatest for very low modulating frequencies, and for such frequencies, is likely to exceed the channel width.

The test indicated several practical difficulties which may be expected in duplex sound and facsimile operation.

1. Necessity for adequate filtering to separate signals before passing through any audio amplifier. This could be done on receivers designed for duplex operation, but would be necessary on all f-m receivers whether designed for facsimile reproduction or not, if cross modulation of the sound by the facsimile is to be avoided. Thus, receivers designed for sound reproduction only would require design provisions not now needed.

2. Necessity for continuous uninterrupted reception of an undesired sound program to obtain a desired facsimile program and vice versa.

3. Necessity for tuning accuracy beyond the capabilities of the average user for acceptable freedom from cross-talk.

4. Necessity for unusually high degree of oscillator frequency stability to maintain this tuning accuracy.

5. Necessity for unusually high degree of linearity in any circuit carrying both sound and facsimile signals.

These considerations result in the opinion that duplex operation of f-m sound and f-m facsimile is technically possible, but has certain difficulties both technical and commercial which would appear to make it undesirable.

The two services have too many conflicting requirements in receiver design and sale, in program design and in user's desire to change one, but not the other, to be tied together thus intimately. It would appear more desirable to operate with independent modulation of sound and facsimile on a time division basis, the facsimile being transmitted at a period when the station is not in use for sound transmission.

The authors are indebted to W. A. R. Brown and T. J. Buzalski of the National Broadcasting Company for furnishing and operating the transmitter facilities for the field tests, and to W. L. Carlson, V. D. Landon, and H. Kihn of the RCA Manufacturing Company for a part of the laboratory tests described. Credit is also due J. A. Rankin of the RCA License Laboratories for able assistance in the field and laboratory tests, and to M. G. Crosby of R.C.A. Communications for his helpful discussions of the theoretical aspects of the problem.

THE EQUIVALENT CHARACTERISTICS OF VACUUM TUBES OPERATING IN FEEDBACK CIRCUITS

By

JOHN H. PRATT

Engineering Products Division, RCA Victor Company Limited, Montreal, Canada

Summary—A graphical method of determining the effects of feedback upon the characteristics of a vacuum-tube amplifier is described. The method depends upon the fact that the characteristics of vacuum tubes may be considered to be changed by feedback.

The method of obtaining vacuum-tube characteristics with feedback from the normal characteristics is described, and a number of examples of "feedback characteristics" for various percentages of positive and negative feedback of both the voltage and current type are illustrated.

THE effects of inverse feedback in reducing distortion, improving frequency response, and changing apparent source impedance in vacuum-tube amplifiers is very well known and since the original paper by Black[1] the subject has been treated mathematically by a number of authors.

It is the purpose of the present article to describe a simple graphical method of calculating the effects of feedback by assuming that the characteristics of the vacuum tube are changed by the presence of the voltage fed back from the output to the input circuit. O. Schade[2] has shown that the plate characteristics of a vacuum tube may be said to be changed by feedback and has shown how equivalent characteristics may be obtained.

The method to be described is most generally applicable when feedback is over one stage only and when the phase of the fraction of the output voltage applied to the input is assumed to be exactly 180° or 0°. However, the method can be extended to cover the case of feedback over several stages and also to take into account a phase angle associated with the feedback factor, although the graphics become somewhat complicated and tedious in the latter cases.

Figure 1 represents the circuit of a common type of inverse-feedback amplifier simplified to show a-c components only. The feedback factor in this case is

$$\beta = -\frac{R_2}{R_1 + R_2} \qquad (1)$$

[1] H. S. Black, Stabilized Feedback Amplifiers, *Elec. Eng.*, Vol. 53, p. 114, January 1934.

[2] O. H. Schade, Beam Power Tubes, *Proc. I.R.E.*, Vol. 26, pp. 176-181, Feb. 1938.

For any instantaneous value of plate voltage e_b the voltage fed back to the grid circuit is

$$e_\beta = \beta e_b \tag{2}$$

and the grid-cathode voltage is

$$e_c = e_g - \beta e_b \tag{3}$$

The above equation gives the instantaneous grid-cathode voltage for any particular combination of instantaneous signal voltage and instantaneous plate voltage. By taking various values of e_g as parameters and allowing e_b to vary, an equivalent family of plate characteristics may be obtained in which grid-cathode voltage is replaced by instantaneous signal voltage.

For example, take the Type 6L6 with $\beta = -0.1$ and assume $e_g = -40$ volts. When $e_b = 100$ volts, $e_c = -40 - (-0.1 \times 100) = -40 + 10 = -30$ volts.

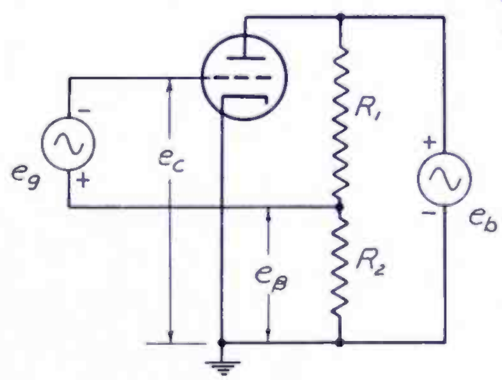


Fig. 1—Negative voltage feedback

Referring to the plate characteristic curves for this tube it will be seen that the plate current for $e_c = -30$ volts, and $e_b = 100$ volts is 5 milliamperes. (Screen grid voltage = 250 volts.)

Let e_b increase to 150 volts.

Then $e_c = -40 + 15 = -25$ volts and

$i_b = 18$ milliamperes

It will be noted that the grid voltage increases by intervals equal to $-\beta$ times the plate voltage, so if the plate-voltage intervals are chosen equal to $-1/\beta$ times the published grid voltage intervals it is a simple matter to plot the equivalent characteristics with feedback by simply moving one grid voltage curve for each plate voltage increment. A curve of this type was commenced above since the published E_{C1} curves are at intervals of 5 volts making the necessary e_b intervals equal to 50 volts.

Continuing: When $e_b = 200$ volts we move up to the curve for $E_{C1} = -20$ and obtain 40 milliamperes. When $e_b = 250$ volts, we obtain $i_b = 67$ milliamperes at the $E_{C1} = -15$ curve. Continuing thus the feedback characteristics may be plotted without any further calculations once the plate-voltage interval has been obtained for the value of β desired, and the proper starting point has been found.

In Figure 2 the complete equivalent plate characteristics for the Type 6L6 with 10 per cent voltage feedback ($\beta = -0.1$) are plotted and Figure 3 shows the effect of a feedback factor of -0.2. The reduction in plate resistance is immediately apparent and the reduction in distortion may be easily seen by plotting a load line on the equivalent characteristics in the usual manner and obtaining a dynamic characteristic curve (e_g vs. i_b). In order to obtain the d-c operating point from the equivalent characteristics, Equation (3) above may be used to find the grid bias corresponding to particular values of e_g and e_b, or the desired plate-voltage vs. plate-current point may be transferred to

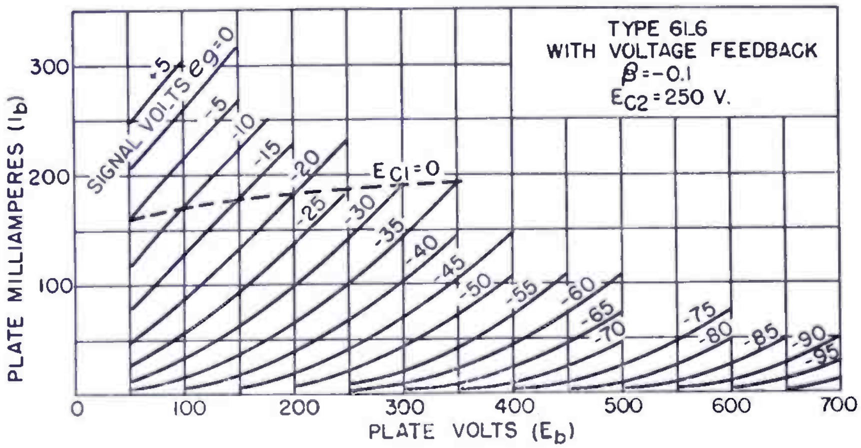


Fig. 2

the normal characteristics and the corresponding grid-bias voltage obtained from them. The latter is probably the simpler method.

Improvement in frequency response may be considered to be due to the reduction of plate resistance and the actual improvement with feedback may be calculated by comparing the reduction in output voltage which occurs as the load impedance is reduced on the published and the equivalent characteristics. Estimation of frequency response in this manner will not be accurate because the reduction in load impedance which causes a drop in frequency response is almost always accompanied by a change from a pure resistive to a reactive load. Thus, the feedback factor β will have a phase angle associated with it, and in order to obtain an accurate equivalent plate characteristic, this phase angle must be used in Equation (3) which then becomes

$$e_c = e_g - e_b|\beta|\angle\theta \tag{4}$$

Equivalent plate characteristics may be plotted for various phase angles and the load line for the reactive load may be plotted upon them, but the amount of labor involved would hardly be justified by the value of the results.[3]

Figures 4 and 5 are the equivalent plate characteristics with a feedback factor of −1, of the Type 6L6 and 2A3 respectively. These might be termed the "cathode characteristics" of these tubes since they represent the equivalent characteristics when the load is placed in the cathode circuit. Note that for Figure 4 to represent the true cathode characteristics of the Type 6L6 the screen grid must be fed through a high impedance and bypassed to the cathode. The cathode-loaded amplifier has been used as a driver for a class B audio stage where the load impedance varies widely over the cycle, and in television circuits where a low source impedance is required and a transformer is not practical because of the wide frequency range.

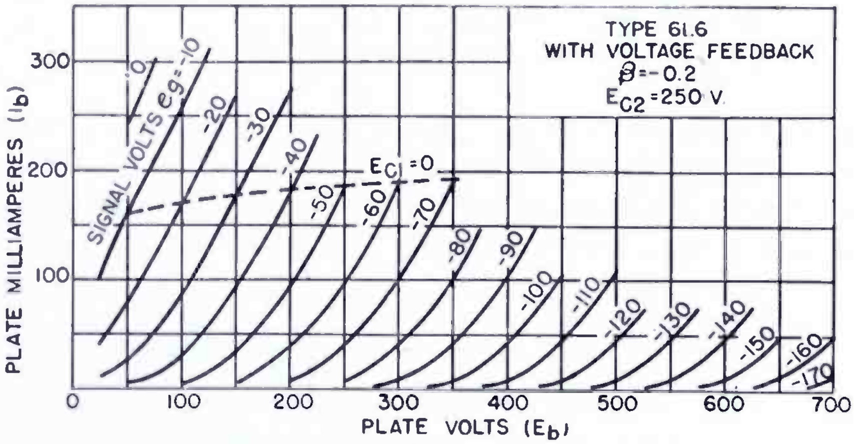


Fig. 3

Examination of Figures 3 to 5 brings out clearly a point which is generally neglected in a mathematical treatment of inverse feedback. The dashed lines on the four figures are the normal characteristic curves for $E_{C1} = 0$. The $E_{C1} = 0$ curve is in the same position no matter what the value of β, and if the load resistance and signal voltage are such as to enter the region above this line, grid current will flow and grid-circuit distortion will normally occur.

These equivalent plate characteristics show clearly the fact that although the apparent plate resistance of a vacuum tube may be lowered by the use of inverse feedback, the load impedance must be kept high throughout the frequency range over which full power output is required. It may be shown that the low-frequency response of a transformer-coupled Class A amplifier without feedback is given approximately by the equation

$$n = \frac{1}{\sqrt{1 + \left(\frac{R}{\omega_L L}\right)^2}} \tag{5}$$

[3] Albert Priesman, Graphics of Non-Linear Circuits, *RCA Review*, Vol. 2, No. 1, p. 124, July 1937.

where n is the ratio of the output voltage at some low frequency $\omega_L/2\pi$ to the output voltage in the middle range of frequencies, R is the resistance of the plate resistance and reflected load resistance in

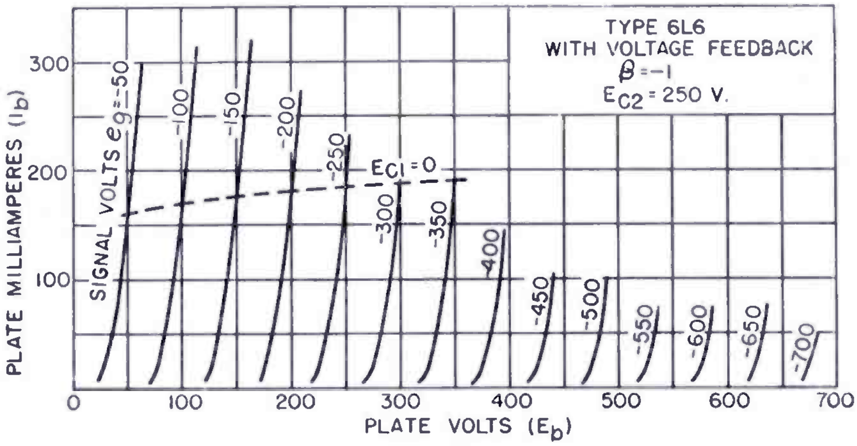


Fig. 4

parallel, and L is the primary inductance of the transformer. This equation indicates that for a particular value of n, a transformer with a lower primary inductance may be used with a tube having a low plate resistance than with a tube having a high plate resistance. Since the

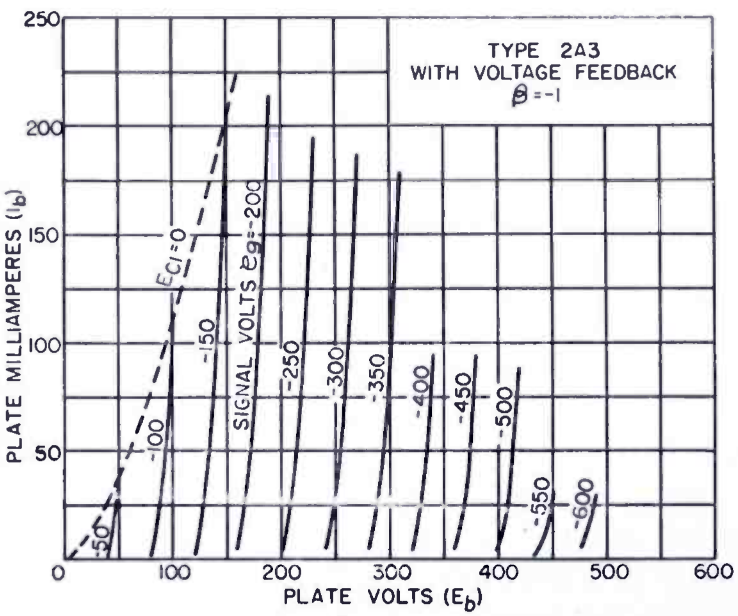


Fig. 5

plate resistance of a vacuum tube is apparently lowered by inverse feedback it would be expected that for a certain low-frequency response, the primary inductance of the output transformer used in a feedback amplifier could be lower than if feedback were not employed. This is true, but only if full power output is not required at low frequencies, for if the grid is driven to zero volts with respect to the cathode at middle frequencies the same input voltage will drive the grid positive when the load impedance begins to drop. The primary inductance of the output transformer used in a feedback amplifier must be great enough to keep the load impedance so high that the grid is not driven positive at the maximum amplitude of the lowest frequency applied to the grid circuit if grid-circuit distortion is to be avoided. The same, of course, also applies to the load impedance at high frequencies, although in speech and music the amplitude of frequencies in the upper range, where load impedance begins to drop, is usually so much less than the amplitude of mid-range frequencies that grid-circuit distortion is not likely to occur.

The limitation imposed on peak-signal voltage by the grid-current point makes a cathode-loaded triode a better driver for a Class B stage

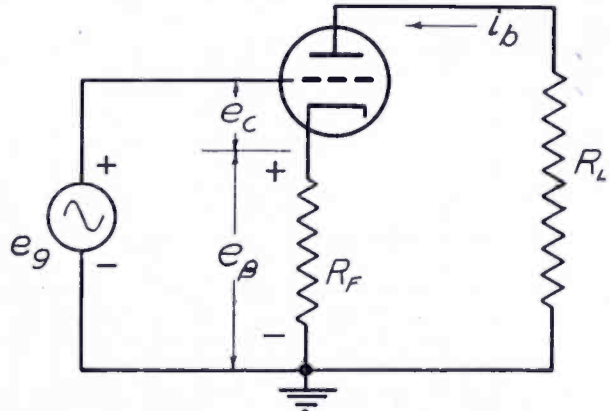


Fig. 6—Negative current feedback

than a cathode-loaded pentode or tetrode since, for a limited peak-signal voltage, the load resistance may vary over a much wider range in the case of the triode than in the case of the pentode or tetrode, even though the latter may have almost as low a "cathode" resistance as the former.

Current Feedback

In all of the preceding paragraphs it has been assumed that a portion of the load voltage was being fed back to the grid circuit. Another form of inverse feedback is commonly employed in which the feedback voltage is proportional to the load current instead of the load voltage. A practical way of obtaining this type of feedback is by the use of an unbypassed cathode resistor. See Figure 6.

With this type of feedback β is not independent of the load resistance as it is with voltage feedback. For any particular value of load resistance R_L

$$\beta = -\frac{R_F}{R_L} \tag{6}$$

From Figure 6 it may be seen that the grid-cathode voltage is equal to

$$e_c = e_g - e_\beta \tag{7}$$

$$\text{or } e_c = e_g - i_b R_F \tag{8}$$

Thus, it is possible (for any particular value of R_F) to take e_g as a parameter, find the values of e_c for various values of i_b and determine the corresponding values of e_b from the published characteristics. For tubes with a high plate resistance this procedure is not very practical

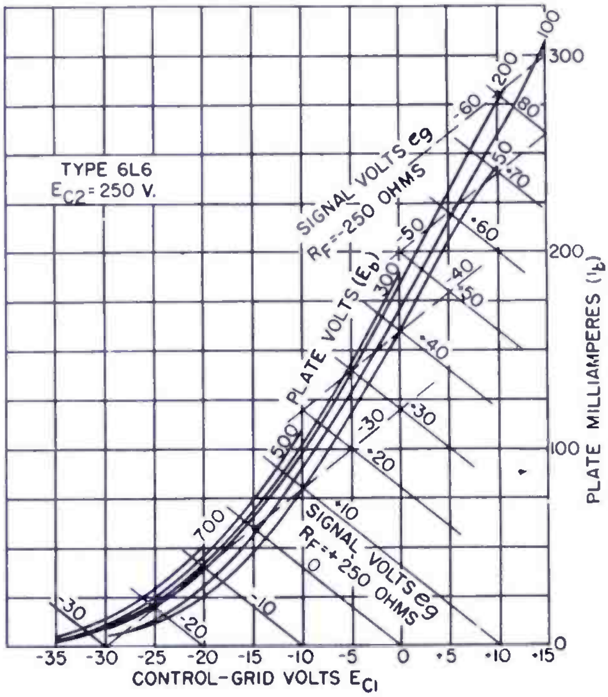


Fig. 7—Transfer characteristics of Type 6L6 with construction for obtaining current feedback characteristics

because if the i_b R_F increments are chosen so as to make e_c vary in the steps shown in the published characteristics, only one or two e_b points can be obtained for each value of e_g. A more practical method is to use the grid-voltage vs. plate-current characteristics. See Figure 7. In Figure 7 are shown the transfer characteristics of the 6L6 and across these are a number of parallel lines drawn at a slope equal to —250 ohms through 10-volt steps of E_{C1} at the zero axis for I_b. The junction of any of these lines with a particular value of I_b extended down to the zero axis represents the grid-cathode e_c for the value of e_g on the line. In other words, the process indicated by (8) is done graphically. Thus, the points where the e_g lines cross the plate-voltage curves may be transferred directly to a set of equivalent I_b vs. E_b characteristics. This is done in Figure 8. Figure 9 is obtained in the same manner, but using a value of R_F equal to 500 ohms.

It should be noted that this graphical construction is only valid for triodes, when current feedback is obtained by means of an unbypassed cathode resistor. To hold true for tetrodes and pentodes the feedback resistor must be placed in the plate circuit so that only plate current flows through it. Both plate and screen current flow in the cathode circuit and since screen current varies with grid voltage somewhat as though the screen were the plate of a triode, the voltage developed

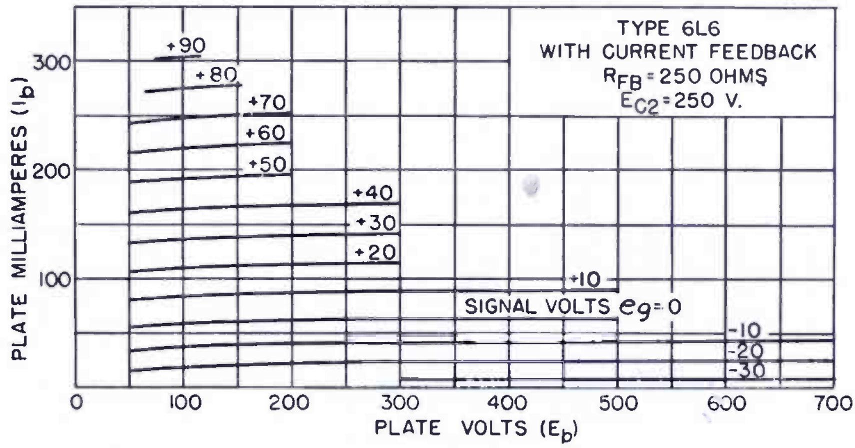


Fig. 8

across the cathode resistor will not be given by equation (8). The current-feedback circuit for which this construction is valid might be similar to Figure 1, except that R_1 would be the load resistance, R_2 the feedback resistor and the output voltage would be developed across R_1 only.

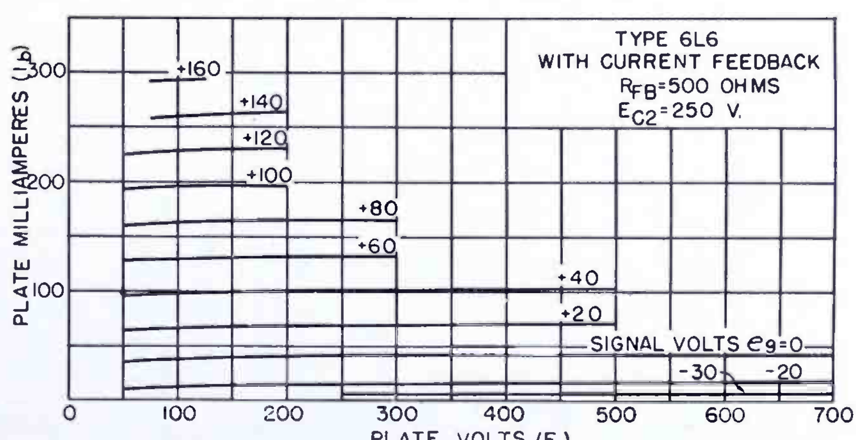


Fig. 9

The effect of current feedback in increasing plate resistance is easily seen in Figures 8 and 9 although it would be even more striking if a tube with a lower plate resistance had been chosen for the illustration. The fact that the e_g curves are more evenly spaced than the E_{C1} curves of the regular characteristics shows that distortion is reduced by current feedback. Figures 8 and 9 represent values of β of -0.1 and -0.2 respectively for a load resistance of 2500 ohms. If this load line is plotted on these equivalent characteristics and also on those of Figures 2 and 3, (using the same operating point), it will be found that the resulting dynamic characteristic from Figure 2 will be identical with that from Figure 8 and the dynamic characteristic from Figure 3 will be identical with that from Figure 9. Since current feedback

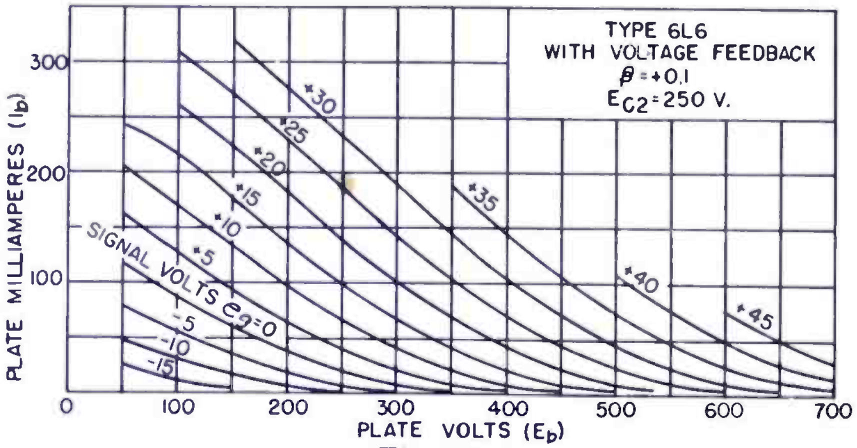


Fig. 10

increases the apparent plate resistance of a vacuum tube, it is immediately apparent that frequency response will not be improved by current feedback. In fact it is actually made poorer. It can be seen from Figures 8 and 9 that as the load resistance is decreased the output will drop off rapidly.

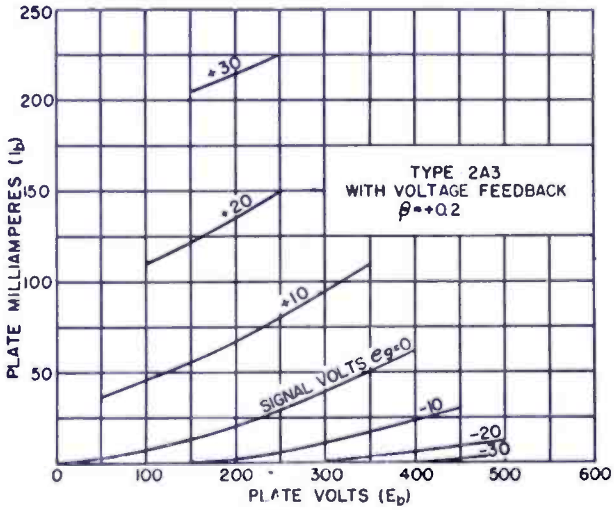


Fig. 11

Regeneration

The methods described for obtaining equivalent characteristics of vacuum tubes with feedback are not limited to the case of negative, or inverse, feedback. They may be employed equally well to illustrate the effects of positive, or regenerative, feedback. Figure 10 shows the equivalent characteristics of the 6L6 with 10 per cent positive voltage feedback. The plate resistance of the tube has been made approximately 1400 ohms negative and for any load resistance greater than 1400 ohms the tube will oscillate. Figure 11 illustrates 20 per cent positive feedback applied to the Type 2A3. The plate resistance has been greatly increased, but is not negative at any point and thus, the tube could not be made to oscillate with this value of β. The increase

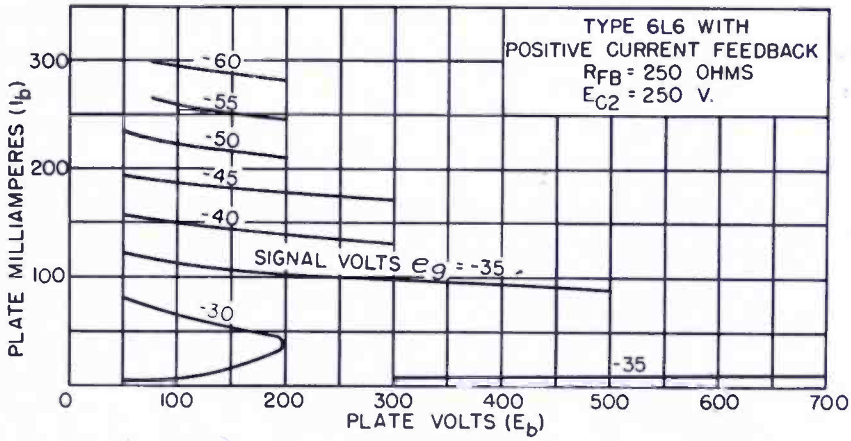


Fig. 12

in non-linearity of the signal-voltage vs. plate-current characteristics is very obvious.

To complete the picture, Figure 12 shows the equivalent plate characteristics of the 6L6 with positive current feedback, the feedback voltage being developed across a resistance of 250 ohms. These equivalent characteristics were obtained in the same way as those for negative current feedback except that the slope of the feedback resistance was drawn opposite. These are the dashed lines of Figure 7. The reversal in the characteristics at the higher signal voltages is caused

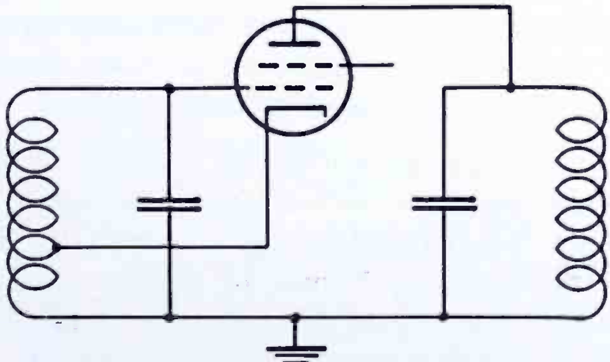


Fig. 13—Oscillator circuit employing positive current feedback

by the line representing the feedback resistance crossing the plate-current vs. grid-voltage curves at two points, as for example does the line representing $e_g = -30$ volts.

The so-called electron-coupled oscillator in which feedback is obtained from the cathode circuit and output is obtained from the plate circuit employs positive-current feedback. See Figure 13. The feedback resistance in this case (at the resonant frequency of the grid circuit) is the resistance seen between the cathode tap and ground, multiplied by the ratio of the voltage between cathode and ground to that between grid and cathode.

It is interesting to note that negative voltage feedback and positive current feedback tends to rotate the plate characteristics in a counter-clockwise direction, while positive voltage feedback and negative current feedback rotate them in a clockwise direction. Negative feedback cannot produce either zero or infinite plate resistance, but zero and infinite plate resistance may be produced by the proper values of positive current and positive voltage feedback, respectively. H. F. Mayer[4] has noted these effects and has shown some practical applica-

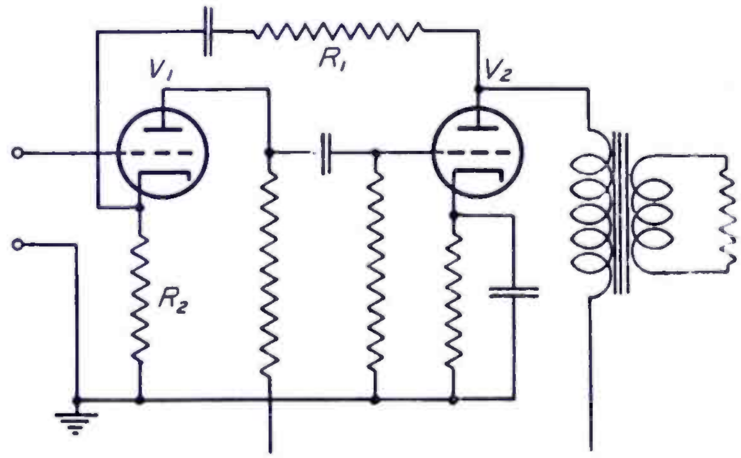


Fig. 14—Feedback over two stages

tions of the use of combinations of the various types of feedback in controlling the effective internal impedance of amplifiers.

Feedback Over Several Stages

To apply the principle of equivalent characteristics to feedback over more than one stage it is necessary to obtain first a set of equivalent characteristics representing all the stages in the feedback loop. For example, in the circuit of Figure 14, first the equivalent characteristics of V_1 due to the current feedback of R_2 would be plotted. On these characteristics the load line would be plotted in the usual manner and a dynamic characteristic obtained. From this dynamic characteristic the grid voltage applied to V_2 for various input voltages could be found and finally the plate current of V_2 could be obtained from its grid-voltage vs. plate-current characteristics. Thus, a set of equivalent plate characteristics of the two stages (without feedback from the

[4] H. F. Mayer, Control of the Effective Internal Impedance of Amplifiers by Means of Feedback, *Proc. I.R.E.*, Vol. 27, No. 3, p. 213, March 1939.

second tube) may be obtained. By the use of Equation (3) equivalent plate characteristics with feedback from V_2 may be obtained in which,

$$\beta = \frac{R_F}{R_1 + R_F} \tag{9}$$

where R_F is equal to R_2 in parallel with the cathode resistance of V_1.

The cathode resistance of V_1 may be shown to be equal to

$$R_K = \frac{Rp + R_L}{1 + \mu} \tag{10}$$

where Rp and μ are the plate resistance and voltage amplification factor of the tube at the operating point and R_L is the resistance formed by the load resistor and grid leak resistor in parallel.

In this example the grid-cathode voltage

$$e_c = e_g - e_\beta \tag{11}$$

Although this is the same as the equation for current feedback (6), the plate resistance will not be increased because on the equivalent characteristics for the two stages the plate current increases as the grid voltage becomes increasingly negative due to the phase reversal through the first tube.

The above procedure is rather involved and entails considerable tedious graphical construction, but by its use the effects of feedback over two stages may be predicted quite accurately.

Conclusion

The methods of graphically determining the effect of feedback are perfectly feasible, and not too complicated to use in general engineering practice. Mathematical methods of treating feedback are quite simple and are probably more convenient to use when it is desired to calculate the performance of a feedback amplifier under a single set of operating conditions. However, when it is desired to try several sets of operating conditions in order to arrive at a desired result, the "feedback characteristics" are very useful for with them it is possible to judge distortion by the shape of the dynamic characteristic without going to the trouble of expressing it as a number and then applying it in an equation for the reduction due to feedback. Gain is also obtained directly from the feedback characteristics without calculation.

FLUCTUATIONS IN SPACE-CHARGE-LIMITED CURRENTS AT MODERATELY HIGH FREQUENCIES

By

B. J. Thompson, D. O. North, and W. A. Harris

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

PART V—FLUCTUATIONS IN VACUUM-TUBE AMPLIFIERS AND INPUT SYSTEMS

(Continued from April RCA Review)

Television Camera Pick-up Tube

THE output signals from an Iconoscope or from other television camera pick-up tubes have certain characteristics in common with those of phototube-output signals. The output current is proportional to the light input; the signal which we wish to amplify appears as modulation of this current; and fluctuation current is produced in the pick-up tube in accordance with the shot-effect law. The major difference is in the effective band width, which is of the order of 5 megacycles in the television case.

When the capacitance of the pick-up tube and the first amplifier tube total 20 $\mu\mu$f. a resistor of 1600 ohms could be used to obtain approximately uniform response over a 5-megacycle band. It would be practical to use an amplifier tube with less than 1600 ohms for its noise-equivalent resistance, thus insuring that the amplifier tube does not contribute seriously to the noise level. However, the thermal-agitation noise of the resistor would exceed the pick-up tube noise for all pick-up tube currents less than 30 microamperes, and would seriously limit the sensitivity of the system.

This thermal-agitation noise can be greatly reduced by the substitution of a resistor of considerably greater value. The signal voltage developed in the input circuit then becomes greater at low frequencies than at high frequencies, but this effect can be compensated for by using an inductance as the plate load for the first amplifier tube. A resistance in series with this inductance balances the effect of the resistance in parallel with the input-circuit capacitance. This circuit arrangement is indicated in Figure 4-a; additional refinements, such as circuit elements to provide the right amount of damping at the plate-circuit resonance frequency, are not considered.

Figure 4-b is an equivalent circuit showing the pick-up-tube noise and the resistor noise introduced by constant-current noise generators

with noise-equivalent conductance values of $20I$ and $1/R$, respectively. The tube noise is shown as introduced by a constant-voltage noise generator R_{eq}. The diagram gives a correct indication of the noise-frequency distribution; noise current from g_a and g_b is amplified in a uniform manner over the amplifier-frequency band since it enters the system in the same manner as the signal, while noise from R_{eq} is amplified in the manner indicated by the plate load of the first tube.

The noise voltage produced by R_{eq} could also be produced by a constant-current noise generator represented by a noise-equivalent

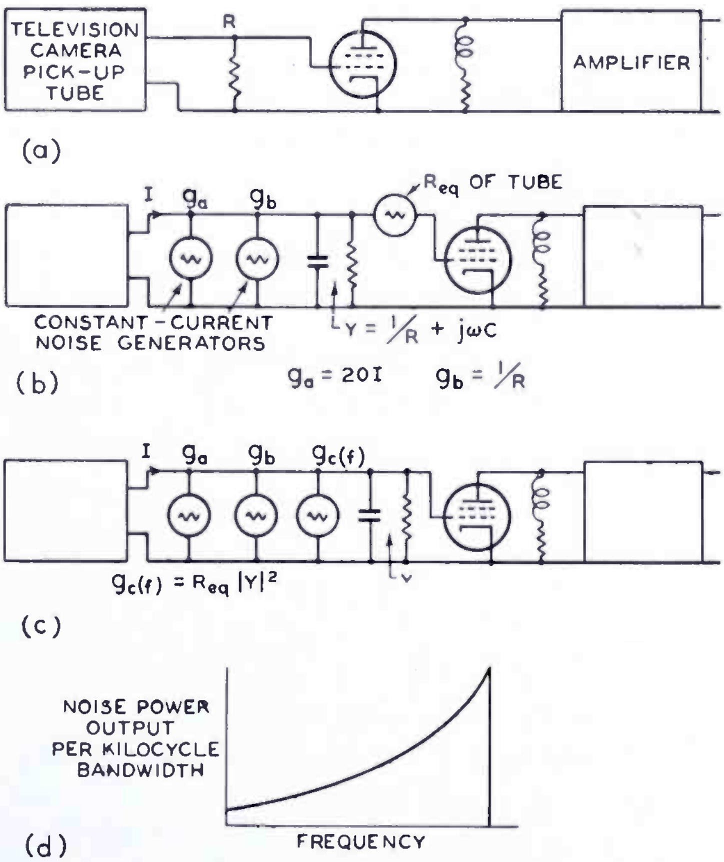


Fig. 4

conductance varying in the proper manner with frequency. This "generator" is represented as $g_c(f)$ in Figure 4-c. The mean-square voltage delivered by R_{eq} in a small frequency band df is

$$d\,\overline{e^2} = 4kT\,R_{eq}\,df$$

The square of the admittance of the input circuit is

$$|Y|^2 = \left(\frac{1}{R}\right)^2 + \omega^2C^2 = g^2 + \omega^2C^2$$

The mean-square current in Y which would produce the mean-square voltage $d\,\overline{e^2}$ is

$$d\,\overline{i^2} = d\,\overline{e^2}\,|Y|^2 = 4kT\,R_{eq}\,|Y|^2\,df$$

The mean-square current produced by a conductance g_c is

$$d\,\overline{i^2} = 4kT\, g_c\, df$$

Consequently, the required noise-equivalent conductance to produce the same noise-equivalent mean-square voltage as that from R_{eq}, in a specified narrow frequency band, is

$$g_c = R_{eq}\,(g^2 + \omega^2 C^2)$$

We can symbolize this by the equation

$$g_{eq(f)} = R_{eq}\, |Y|^2$$

The total noise-equivalent conductance represented in Figure 4-c is

$$g_{eq(f)} = g_a + g_b = g_{c(f)}$$

or, substituting values,

$$g_{eq(f)} = 20\,I + \frac{1}{R} + \frac{R_{eq}}{R^2} + R_{eq}\,\omega^2 C^2$$

The noise power output per unit bandwidth for the system is proportional to $g_{eq(f)}$ so it can be represented by the curve of Figure 4-d. An average noise-equivalent conductance value can be obtained readily from this equation. It is

$$\overline{g_{eq}} = 20I + \frac{1}{R} + \frac{R_{eq}}{R^2} + \frac{1}{3} R_{eq}\,\omega^2_{max}\, C^2$$

The total noise-power output will be proportional to this value.

$$\begin{aligned} \text{Let } R &= 50{,}000 \text{ ohms} \\ C &= 20\ \mu\mu\text{f.} \\ R_{eq} &= 720 \text{ ohms} \\ f_{max} &= 5 \times 10^6 \text{ cycles per second} \end{aligned}$$

Then, when I is given in microamperes,

$$\begin{aligned} \overline{g_{eq}} &= (20I + 20 + 0.29 + 96) \times 10^{-6} \\ &= (20I + 116) \times 10^{-6} \text{ mho} \end{aligned}$$

The corresponding equation for the circuit using the 1600-ohm resistor is

$$\begin{aligned} g_{eq} &= 20I + \frac{1}{R} + \frac{R_{eq}}{R^2} \\ &= (20I + 625 + 281) \times 10^{-6} \\ &= (20I + 906) \times 10^{-6} \text{ mho} \end{aligned}$$

At low current inputs an advantage of about 9 db in signal-to-noise ratio is indicated for the circuit of Figure 4-a. The current input required to produce noise equal to the fixed noise under the assumed conditions in the circuit of Figure 4-a is 5.8 microamperes and the rms fluctuation current in a 5-Mc band with this current flowing is 0.0043 microamperes; the rms fluctuation current with no signal input ($I = 0$) is 0.003 microampere. Since noise is evidently not at a prohibitive level with the 5.8-microampere signal, operation at lower levels may be assumed. At low-input levels the first amplifier tube is the principal noise source.

The frequency-distribution curve for the circuit, shown in Figure 4-d, will quite evidently produce a different effect on the television screen than would a uniform distribution. Fluctuation "noise" on a television screen gives somewhat the appearance of a snow storm. When the noise is concentrated in the high-frequency portion of the band to the extent indicated in Figure 4-d the result is finer "flakes." Tests have shown that greater noise power outputs can be tolerated under this condition than under the condition of uniform frequency distribution. Therefore, the effective improvement obtained from the circuit of Figure 4-a (under the assumed conditions) over the circuit using the 1600-ohm load resistor would be even greater than the 9-db power-output-ratio improvement.

Receiver Input Circuits

The noise from the input circuit of a radio receiver can be divided into that produced by thermal agitation in the circuit itself and that picked up by the antenna. The first part is readily determined. An equivalent circuit such as the circuit of Figure 1-d* can be used to represent the system, with a noise-equivalent resistance value equal to the resonant impedance of the circuit assigned to the noise generator R_b. When the input circuit does not have much effect on the overall selectivity of the receiver the "filter" of Figure 1-d can be eliminated and the noise-equivalent resistance for the amplifier can be added directly to that for the circuit.

When the first circuit is selective enough to have an effect on the overall selectivity, the noise-equivalent resistance for the amplifier should be multiplied by the ratio of the effective bandwidth determined for the amplifier only, to that for the complete system. The noise-equivalent resistance value thus obtained corresponds to the value for the noise generator R'_a of Figure 1-e.* It can be added to the noise-equivalent-resistance value for the input circuit as before and the

* See page 511, RCA Review, April, 1941.

noise-equivalent sideband input for the receiver can be determined by using the total noise-equivalent resistance and the overall effective bandwidth.

Example: Let the first tube be an r-f amplifier with a noise-equivalent resistance of 10,000 ohms. Assume an effective bandwidth of 6000 cycles with the first circuit disconnected and 4000 cycles with the first circuit in use; an input circuit operating at 600 kilocycles with a "Q" of 100 could produce this result. Let the resonant impedance of the input circuit be 50,000 ohms. The resulting total noise-equivalent resistance is 65,000 ohms. "Ensi" for this receiver, referred to the grid of the first tube, would be 2.1 microvolts.

Of course, we could obtain the same result by using the effective bandwidth and the noise-equivalent resistance of the amplifier as reference values and assigning a correspondingly reduced noise-equivalent resistance to the circuit. The total noise-equivalent resistance would then be 43,000 ohms and the effective bandwidth 6000 cycles; "ensi" would still be 2.1 microvolts.

An extension of the same idea can be applied in the analysis of a receiver using an "untuned r-f" stage ahead of a frequency converter. A circuit for such an arrangement is shown in Figure 5-a, and a representation of a curve of power output against frequency for a constant-current "sideband" signal introduced at point "A" is shown in Figure 5-b. This curve applies when the receiver is adjusted for a signal of 1000 kc; the assumed intermediate frequency is 455 kc and the corresponding oscillator frequency is 1455 kc. The dotted curve (Figure 5-b) represents the response of the interstage coupling circuit, and the relations between the heights of the rectangles and the dotted curve represent the relative gains (squared) of the converter itself at the different frequencies indicated. The squares of the relative overall gains determined for one such system were as follows:

	I-f	*Signal*	*Image*	*2d-harmonic Images*	
Frequency	455	1000	1910	2455	3365
(Relative gain)^2	1.72	1.00	0.75	0.22	0.15

Small responses could be observed at higher frequencies but their importance with respect to the total noise output was small.

Additional data for this system were:

Converter noise-equivalent resistance	240,000 ohms
R-f gain	5
R-f tube noise-equivalent resistance	10,800 ohms
Circuit resistance (1000 kc)	100,000 ohms

Converter noise-equivalent resistance values are referred to the signal frequency only; the value of 240,000 ohms referred to the r-f grid (divide by 25) gives 9600 ohms.

The r-f tube produces noise at all the response frequencies, but comparison is to be made with a circuit operating at only one of those frequencies—1000 kc. Consequently the noise-equivalent-resistance value to be added on account of the r-f tube is obtained by adding the

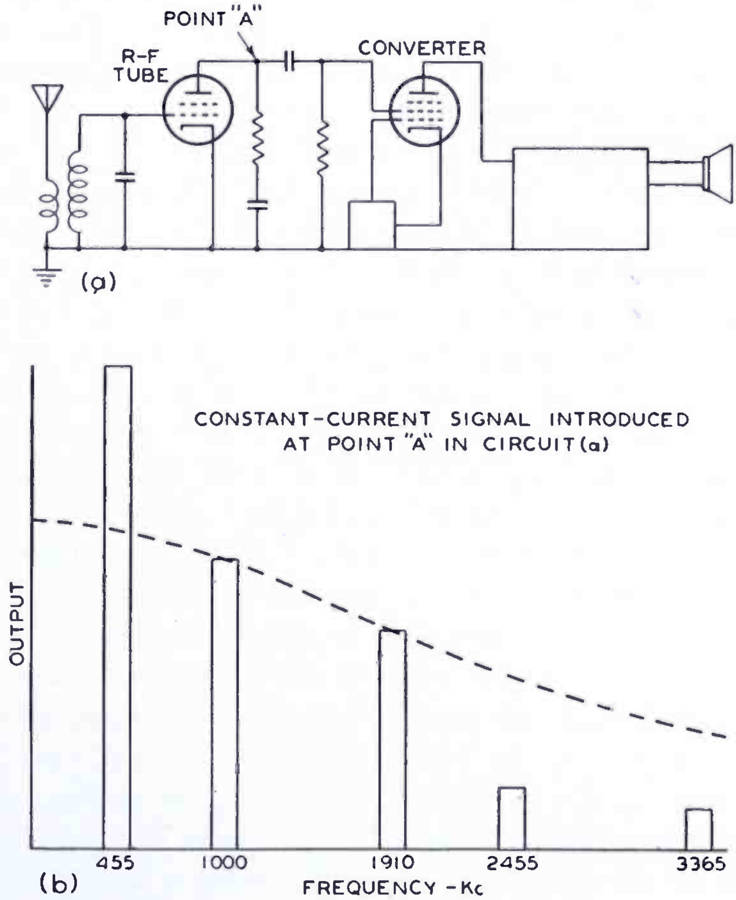


Fig. 5

(relative gain)² factors and multiplying their sum by the noise-equivalent resistance for the r-f tube; the resultant product is 41,500 ohms. The total noise-equivalent resistance for the system is consequently 151,000 ohms. A "broad" input circuit is assumed.

Antenna Noise

An antenna can be regarded as a transformer coupling the input circuits of a receiver to "space". In this sense, the radiation resistance of an antenna is the reflected resistance of "space" measured at the antenna terminals, and the efficiency is the ratio of the radiation resistance to the total resistance of the antenna. There will be a thermal-agitation voltage at the terminals of an antenna of the magnitude determined by the antenna temperature and the ohmic or dissipative resistance of the antenna, but noise associated with the "radiation resistance" component does not originate in the antenna; it is properly classified as "received" noise.

There is one situation in which this received noise can be calculated. If we imagine an antenna in a large enclosure, with the boundaries and the contents of the enclosure maintained at a constant temperature, we know that the antenna must then be in thermal equilibrium with any resistor connected across its terminals. This condition means that the thermal-agitation law must apply to the radiation resistance of the antenna as well as to the dissipative resistance.

When a resistor is connected to the antenna, part of the thermal-agitation energy is radiated into the enclosure. Maintenance of thermal equilibrium requires that an equal amount of energy be received by the antenna; otherwise the resistor would be cooled by the loss of thermal energy to the surrounding space. The thermal-agitation voltage associated with the radiation resistance would be a measure of this received energy. If the boundaries of the enclosure were cooled to a temperature below that of the resistor, the resistor could actually lose heat by radiation, no thermodynamic principle would be violated and the thermal-agitation voltage association with the radiation resistance would be reduced. We could take this into account by using a noise-equivalent resistance for the antenna equal to the sum of the dissipative resistance and the proper fraction of the radiation resistance; or if the temperature of the enclosure was higher than that of the antenna and the resistor, we could use a suitable multiple of the radiation resistance.

Unfortunately, since we cannot locate any boundaries for the space enclosing a real antenna, this method of noise calculation cannot be used. An experimental approach would require a determination of the amount of noise picked up by antennas of known characteristics at various frequencies. Perhaps analysis of the results of such an investigation would reveal a steady component of noise which could be classified as thermal agitation. Our real interest, however, would be in the determination of that minimum noise level which could be expected to persist for time intervals long enough to permit the reception of useful signals. We need not be greatly concerned as to whether or not the name "thermal agitation" is properly applicable to that minimum noise.

The "transformer" analogy indicates that the received noise power will be proportional to the radiation resistance when other factors (such as directivity) remain constant. If we assume that the relation between radiation resistance and noise input is known for a particular system we can represent this component of noise by a noise-equivalent resistance such as R_c in the circuit of Figure 6-b; R_c will be a multiple (or fraction) of that component of the circuit resistance measured at the grid, which is due to the antenna radiation resistance. This component is indicated as R_2 in Figure 6-b. Its magnitude can be calculated by transformer theory and the ratio of its magnitude to the total resistance $(R_2/(R_1 + R_2)$, Figure 6-b) determines the efficiency of the system as a whole. The ratio of "R_c" to the total R_{eq} determines the efficiency of the system with respect to signal-to-noise ratio, for if R_c is much greater than $(R_a + R_b)$ the signal-to-noise ratio in the receiver approaches equality with the signal-to-noise ratio existing in "space." It is evident that this desirable condition can be realized most readily

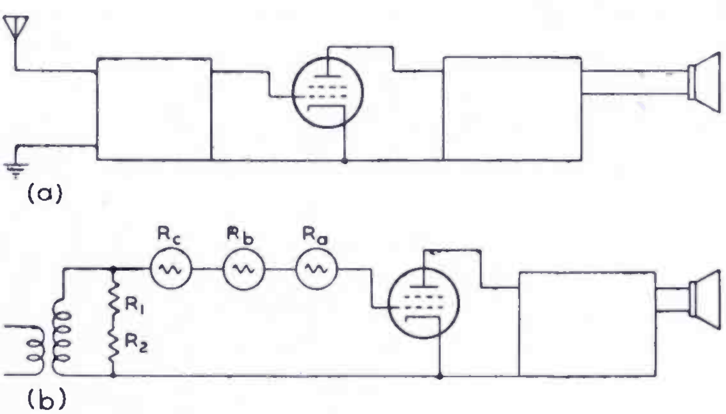


Fig. 6

if the ratio of the noise-equivalent-resistance value corresponding to the radiation resistance, to the radiation-resistance value happens to be large; but this is no more significant than a statement that a quiet receiver is of no particular value in a noisy location. When it is found that the ratio is small in a given situation, we are justified in doing everything possible to improve the coupling between the antenna and the grid of the first tube and to reduce circuit losses to a minimum.

In most home receivers operating at low frequencies the component of circuit impedance derived from the radiation resistance is so small that noise picked up by the antenna under quiet conditions is negligible in comparison with the internally-generated fluctuation noise. At higher frequencies, however, it is practical to build efficient antennas of reasonable dimensions. Consequently the determination of minimum values for the ratio of radiation resistance to noise-equivalent resistance at the higher frequencies may become a matter of considerable importance.

Feedback

A simple circuit involving feedback is shown in Figure 7. A signal current i is introduced from a high-impedance source. A voltage e_g is developed between the grid and the cathode of the tube. The plate current i_b flowing through the "tickler coil" induces a voltage in the grid circuit. This induced voltage has the same effect as a second current

$$i_1 = \frac{M}{L} i_b = \frac{M}{L} g_m e_g$$

introduced into the circuit in the same manner as the signal current, and the value of e_g is the resultant of the effect of these two currents in the grid circuits. Represent the impedance of the grid circuit as Z. Then,

$$e_g = \left(i + \frac{M}{L} g_m e_g \right) Z$$

$$\text{or } e_g = \frac{iZ}{1 - \frac{M}{L} g_m Z}$$

$$\text{and } i_b = \frac{i g_m Z}{1 - \frac{M}{L} g_m Z}$$

A spurious current component originating in the plate circuit will be represented by i'_b and a spurious component originating in the input circuit will be represented as i'. Inclusion of these components gives the total current into the grid circuit as

$$i_{\text{total}} = i + i' + \frac{M}{L} i'_b + \frac{M}{L} g_m e_g$$

$$\text{consequently } e_g = Z \left(i + i' + \frac{M}{L} i'_b \right) + \frac{M}{L} g_m Z e_g$$

$$\text{or } e_g = \frac{\left(i + i' + \frac{M}{L} i'_b \right) Z}{1 - \frac{M}{L} g_m Z}$$

$$\text{and } i_b = g_m e_g + i'_b$$

$$= \frac{(i + i')\, g_m Z}{1 - \dfrac{M}{L} g_m Z} + i'_b \left[1 + \frac{\dfrac{M}{L} g_m Z}{1 - \dfrac{M}{L} g_m Z} \right]$$

$$= \frac{(i + i')\, g_m Z + i'_b}{1 - \dfrac{M}{L} g_m Z}$$

The numerator represents just the plate-current components which would be observed in the absence of feedback, so it is evident that the signal component and both spurious components are affected to the same degree. The result should not be surprising; it is difficult to see how feeding part of the output back into the input could affect the relation between the desired signal and spurious signals introduced

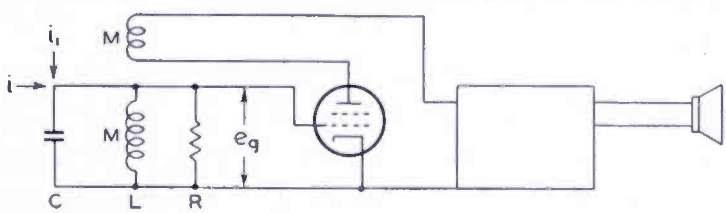


Fig. 7

with the desired signal or generated between the "input" and "output" points.

It is conceded that situations of greater complexity, to which the above statement does not apply, can be developed in vacuum-tube amplifier circuits. However, the circuit of Figure 7 can be used to illustrate some of the effects of feedback on signal-to-noise ratio and particularly on the measurement of noise.

(1) When a second amplifier stage following the circuit of Figure 7 produces a significant amount of fluctuation noise, the amount of feedback affects the signal-to-noise ratio by its effect on the first-stage gain.

(2) The feedback in the circuit of Figure 7 changes the effective bandwidth of the circuit and, consequently, it may change the effective bandwidth of the whole system. The signal-to-noise ratio in the plate circuit of the tube would be unchanged if the signal power were distributed uniformly over the frequency band passed by the system, or if the frequency band passed by the following amplifier was so narrow that Z could be replaced by R in the equations for Figure 7 for that

band; otherwise, the total signal-to-noise ratio in the output of the amplifier of Figure 7 would change with the amount of feedback.

(3) The frequency distribution of the output noise changes when feedback changes the effective bandwidth by a significant amount.

In the measurement of noise it is common practice to make a measurement of noise-power output with the input circuit of the first tube effectively shorted (a "carrier" signal is introduced when necessary) and to compare the result of this measurement with the noise-power output observed with the input circuit in use. With the circuit of Figure 7 the feedback is eliminated when the input circuit is shorted. Shorting the input circuit might result in a considerable change in the noise-power-output component due to tube noise, especially when the feedback is regenerative. It is evident, therefore, that the ratio of the noise-power output measured with the input circuit shorted to that measured with the input circuit operative does not always represent the ratio of tube noise to total noise. When there is regenerative feedback into the input circuit there will be a decrease in noise-power output when the input circuit is shorted, even though the circuit noise is entirely negligible.

The principal limitation to the scope of this series is implied by the title phrase "At Moderately High Frequencies." There is a source of current fluctuations associated with the component of tube-input conductance produced by electron transit-time effects, a source which becomes important when the input frequency is high enough to make the transit-time input conductance relatively large.[12] There are also complications introduced at high frequencies by electron-transit angles, which make it more difficult to take feedback into account in noise determinations. The lower limit of frequencies at which these effects become important depends on the high-frequency operating characteristics of the first amplifier tube; the limit may range from about 30 to 200 megacycles. The theory and application methods presented in this series are believed adequate to permit the complete determination of the more important manifestations of fluctuation phenomena at lower frequencies.

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[12] D. O. North and W. R. Ferris, "Fluctuations Induced in Vacuum Tube Grids at High Frequencies." *Proc. I.R.E.*, Vol. 29, pp. 49-50, February (1941).

OUR CONTRIBUTORS

MAURICE ARTZT attended the University of Texas, where he obtained a degree of B.S. in Electrical Engineering in 1925. He joined the General Electric Company in 1925, starting in the test department and later transferring to the radio engineering department. While in Schenectady he attended Union College and received a degree of M.S. in Electrical Engineering in 1929. Since 1930 he has been employed in the laboratories of the RCA Manufacturing Company in Camden. Since 1928 his work has been in the field of facsimile development. He is an associate member of the Institute of Radio Engineers and a member of Tau Beta Pi.

DUDLEY E. FOSTER received his E.E. degree at Cornell University in 1922. Prior to entering college he served during the war as a radio operator for the American 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 that company was merged with the United States Radio and Television Company and soon thereafter Mr. Foster 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. He also is lecturer in television engineering at Stevens Institute of Technology and is a member of the Institute of Radio Engineers.

ALAN M. GLOVER was born in Rochester, New York on September 26, 1909. He received the B.A. degree in 1930 from the University of Rochester; the M.A. in 1932, and the Ph.D. degree in Physics in 1935, also at the University of Rochester. In 1935 and 1936, he was on the staff of the Institute of Paper Chemistry, Laurence College, Appleton, Wisconsin. Since 1936, he has been with the Research and Engineering Department of the RCA Manufacturing Company, engaged in the development of phototubes and other special tube activities of the company.

RAYMOND F. GUY has had a long and rich experience in broadcasting. He entered the Marconi marine service in 1916, resigning in 1918 to enlist in the regular army. After a year overseas he entered Pratt Institute, graduating in Electrical Engineering in 1921. Following short periods as Inspector for the Shipowners Radio Service and the Independent Wireless Telegraph Company he was retained by Westinghouse in 1921 for the first WJZ staff. In the ensuing years he was prominent in the development of engineering and operating techniques which have been followed ever since in the broadcasting industry. In 1924 he was drafted by the RCA Research Laboratory to head the Broadcast Engineering Section which engineered the RCA stations, did consulting work for clients and supervised the development of all RCA broadcast transmitting apparatus used or sold. During this period he collaborated in the first Transatlantic rebroadcast and its ensuing development, initiated short-wave broadcast relaying, collaborated in the development of a network of RCA stations using telegraph wire facilities and the construction and design of the first 50-kw broadcasting and short-wave stations. In 1929 he accepted Mr. Hanson's invitation to become NBC's Radio Facilities Engineer and has since been responsible to him for design, construction and engineering of all NBC broadcasting, television, short-wave and u-h-f radio facilities, frequency allocation work, power tube engineering, etc. He is licensed to practice as a professional engineer in New York and New Jersey, is a Fellow of the Institute of Radio Engineers, a Fellow of the Radio Club of America, a Member of the New Jersey State Society of Professional Engineers and a Member of the New York Electrical Society.

WILLIAM A. HARRIS is a native of Indiana. He received his B.S. degree in Electrical Engineering from the Rose Polytechnic Institute in 1927. He was in the radio department of the General Electric Company, 1927-1928, and in receiver development work for the same company, 1928-1929. Since 1930 he has been engaged in research and engineering work for the RCA Victor Company and the RCA Manufacturing Company at Harrison, N. J. Mr. Harris is an associate member of the Institute of Radio Engineers.

R. B. JANES was born February 2, 1909 at Burlington, Mass.; Kenyon College, B.S., 1928; Graduate Student, Harvard University, 1928-1929; University of Wisconsin, Ph.D., 1935; Instructor in Physics, Colgate Univ., 1929-1931; Research Assistant in Physics, Univ. of Wisconsin, 1931 to 1935; Since 1935 in the Research and Engineering Department of RCA Mfg. Co., Harrison, N. J. engaged in work on television tubes and phototubes. Member of Institute of Radio Engineers.

DAVID G. C. LUCK was born July 1906 in Whittier, California. Bachelor of Science Mass. Inst. of Tech. 1927; Ph.D. 1932. Swope Fellow in Physics 1927-28, Malcolm Cotton Brown Fellow 1928-29. Assistant, Department of Physics, M.I.T., 1929-32. Research Division, RCA-Victor Co. 1932-35, Victor Division of RCA Manufacturing Co. 1935. Member, American Physical Society.

IOURY G. MALOFF, a former lieutenant of the Russian Navy, arrived in this country in 1923 and joined the General Electric Company. In the latter part of 1926 he severed his connection with that company to engage in independent consulting work, and in 1929 became Chief Engineer of the Valley Appliance Corporation at Rochester, N. Y. Later in that year he became Chief Engineer of the Colonial Radio Corporation. In 1931 he joined the RCA Research Staff at Camden where he is now Engineer in Charge of the Television Tubes and Tube Applications Section of RCA Manufacturing Company. Mr. Maloff is a Fellow, Institute of Radio Engineers.

GEORGE O. MILNE, Eastern Division Engineer for the National Broadcasting Company, attended the Western Electric Company Installation and Machine Switching School in 1922 and assisted in the installation of the first full mechanical telephone exchange in the East at Paterson, N. J. In 1924 he was transferred to the radio broadcasting department of the American Telephone and Telegraph Company and worked in the studio, field and transmitter groups. Two years after the formation of NBC in 1926 he was appointed Operation Supervisor. He was placed in charge of NBC's Eastern Engineering Division when it was organized in 1930.

HARRY F. OLSON received his B.E. degree in 1924, M.S. in 1925, Ph.D. in 1928, and E.E. in 1932, from the University of Iowa. Eight years ago he became an RCA engineer and in that time spent two years in Photophone development work. He is now Research Engineer in the Victor Division of the RCA Manufacturing Company, Inc. Dr. Olson is a member of Sigma Xi, and of the American Physical Society, and is a Fellow of the Acoustical Society of America.

JOHN H. PRATT, born in Toronto, Ontario, graduated from the General Engineering Course at the New York School of RCA Institutes in 1938. In 1939 he joined the RCA Victor Company, Montreal, Canada in the Engineering and Development Section of the Engineering Products Division where he has since remained.

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, Marietta College, and Suffolk University; Doctor of Literature from Norwich University; and Doctor of Commercial Science from Oglethorpe 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.

WILLIAM ARTHUR TOLSON is a native of Texas. He received his B.S. degree in Electrical Engineering from Texas A. and M. College in 1923. He was in the testing department of General Electric Company at Schenectady, N. Y., 1923-1924. From 1924 to 1928 he was in charge of carrier current communication development at the General Electric Company, Schenectady. From 1928 to 1930 he was in charge of television development for the same firm. From 1930-1931 he was in charge of television development for RCA Victor Corporation, Camden, N. J. From 1931 to date he has been a member of the Research Division, RCA Manufacturing Co., Inc., Camden, N. J.

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