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Microwave Repeater Research

By H. T. FRIIS

Introduction

IT WAS some 80 years ago that Maxwell and Hertz demonstrated that free space is a good transmission medium for electromagnetic waves. Since this fundamental contribution, the radio art has advanced tremendously and a decade ago it had progressed to the point where it was possible to construct equipment suitable for quantitative propagation studies of microwaves. Such studies were made and they indicated that normal propagation over "line-of-sight" paths of signals of 10 to 20 centimeters wavelength was characterized by free space attenuation and freedom from atmospheric interference. These results, together with the facts that in this wavelength range wide bands of frequencies are available and it is possible to design small antennas having high directivity, encouraged us to start more comprehensive research work on microwave repeater circuits. This paper gives the present status of the work which was interrupted by our war efforts and resumed at the end of the war with the construction of an experimental New York-Boston system as an initial objective.

The first section will describe our propagation studies. It will be followed by sections on repeater circuit planning, antennas, radio frequency channel filters, the construction and testing of the repeater amplifier, and a concluding section on the whole repeater.

I. Propagation Studies*

That portion of the radio frequency spectrum represented by wavelengths shorter than about five meters has long been considered as the proper domain for point-to-point communication links, local broadcasting, and mobile radio communication. Since these ultra-short waves are not reflected by the ionosphere, their effective range is not much greater than the horizon distance and it therefore becomes possible for a number of stations, properly separated, to operate in the same frequency band; for the same reason, atmospheric interference is not an important factor in this wave-

* This section was prepared by A. B. Crawford who, with W. M. Sharpless, is at present engaged in microwave propagation studies.

length range. Also, as the wavelength decreases, it becomes possible to construct antennas large in comparison with the wavelength so that high antenna gains are obtained and the corresponding directivity further reduces the interference areas.

Since about 1930, with the exception of the war years, we have conducted fundamental studies in radio propagation, taking advantage of advances in the art to extend the wavelength range from about four meters (ultra-short wave region) in the beginning to 1.25 centimeters (microwave region) at the present time. A considerable portion of the effort of those engaged in propagation studies has, of necessity, been devoted to the development of measurement techniques and reliable measuring apparatus. The present discussion, however, will be concerned with the results of experiments rather than with a description of the apparatus and methods. Most of these results have been described in the literature; the following is a review intended to show the development of the background leading to the present field trial of a microwave repeater circuit.

The object in making propagation studies has been to evaluate and to understand the effects of the terrain and of the lower atmosphere upon the transmission of ultra-short-wave and microwave signals. The evaluation is usually obtained by amassing sufficient data on a particular transmission experiment so that a statistical analysis can be made. Efforts to understand the transmission phenomena usually take the form of experiments involving specially designed apparatus. These experiments are varied from time to time as information is obtained or as it becomes desirable to check the validity of such theories as may be devised. The hope is always present that an understanding of the phenomena may suggest a means for reducing the transmission difficulties.

The absence of ionospheric reflections at these frequencies suggested at the start that propagation studies would probably be concerned mainly with phenomena familiar in optics, namely: reflection, refraction and diffraction. Two of the early papers[1, 2] treated ultra-short-wave propagation from this viewpoint. It was soon observed that diffracted signals tended to be unstable in the shadow region; furthermore, as the wavelength is decreased the shadows cast by obstacles such as hills or the bulge of the earth itself become more sharply defined. For these reasons, a considerable part of our experimental work has been done on paths for which a line-of-sight exists between transmitter and receiver. The chief interest, therefore, has been in ground reflections and the effect of the atmosphere.

[1] J. C. Schelleng, C. R. Burrows and E. B. Ferrell, "Ultra-Short-Wave Propagation", *Proc. I. R. E.*, vol. 21, pp 427–463; March 1933.
[2] C. R. Englund, A. B. Crawford and W. W. Mumford, "Some Results of a Study of Ultra-Short Wave Transmission Phenomena", *Proc. I. R. E.*, vol. 21, pp 464–492; March 1933.

GROUND REFLECTIONS

Some of our first experiments with ultra-short-waves showed that regular reflections were obtained locally from open, relatively flat fields. The reflection coefficients were in good agreement with theory. Later, measurements of propagation between a transmitter located on a hill top and a receiver carried in an airplane[2] showed that for near-grazing angles of incidence, the irregular and wooded terrain, typical of the New Jersey countryside, could give rise to regular reflections at wavelengths as short as four meters. The depth of the minima in received signal strength, caused by wave interference between the direct and ground reflected components, corresponded to a reflection coefficient of about 0.9. In 1939, unpublished results obtained over the 39-mile Beer's Hill-Lebanon optical path (See map of Fig. I-1) indicated that for a wavelength of 30 centimeters the reflection coefficient was still large, about 0.8.

More recently, microwave propagation studies have been made over the same type of terrain at wavelengths of 3.25 centimeters and 1.25 centimeters and the situation in regard to ground reflections seems to have changed somewhat. Experiments were conducted over the 12.6 mile Beer's Hill-Deal path in which the height of the transmitting terminal was varied and which also made use of narrow-beam scanning antennas to separate the direct wave from a possible ground reflected component. The results showed the apparent reflection coefficient to be of the order of 0.2 at 3.25 centimeters and to be even less at 1.25 centimeters. Figure I-2 shows typical curves of signal level versus transmitter heights for wavelengths of 3.25 and 1.25 centimeters. Actually, the shapes of the curves can be accounted for better by diffraction, for which the hill about two miles from Deal is considered to be a straight edge, than by reflection from an assumed average ground plane. The true picture is probably a combination of reflection and diffraction effects.

In an effort to minimize ground reflection, over-water paths were avoided in the layout of the New York-Boston microwave repeater circuit and as a final check a number of variable antenna-height tests* were made in the preliminary survey of all sites. A few curves obtained at a wavelength of 7 centimeters are reproduced in Fig. I-3. Similar results were observed during a survey of sites between Chicago and Milwaukee.

It is concluded, therefore, that although in the wavelength range down to 30 centimeters, at least, the effects of ground reflection must be taken into account in the choice of sites for an optical path radio circuit, in the lower microwave range, below say 10 centimeters, scattering and absorption of the reflected wave by rough terrain and vegetation usually results in substantially free-space propagation under normal conditions when the line of

* F. F. Merriam was in charge of this work.

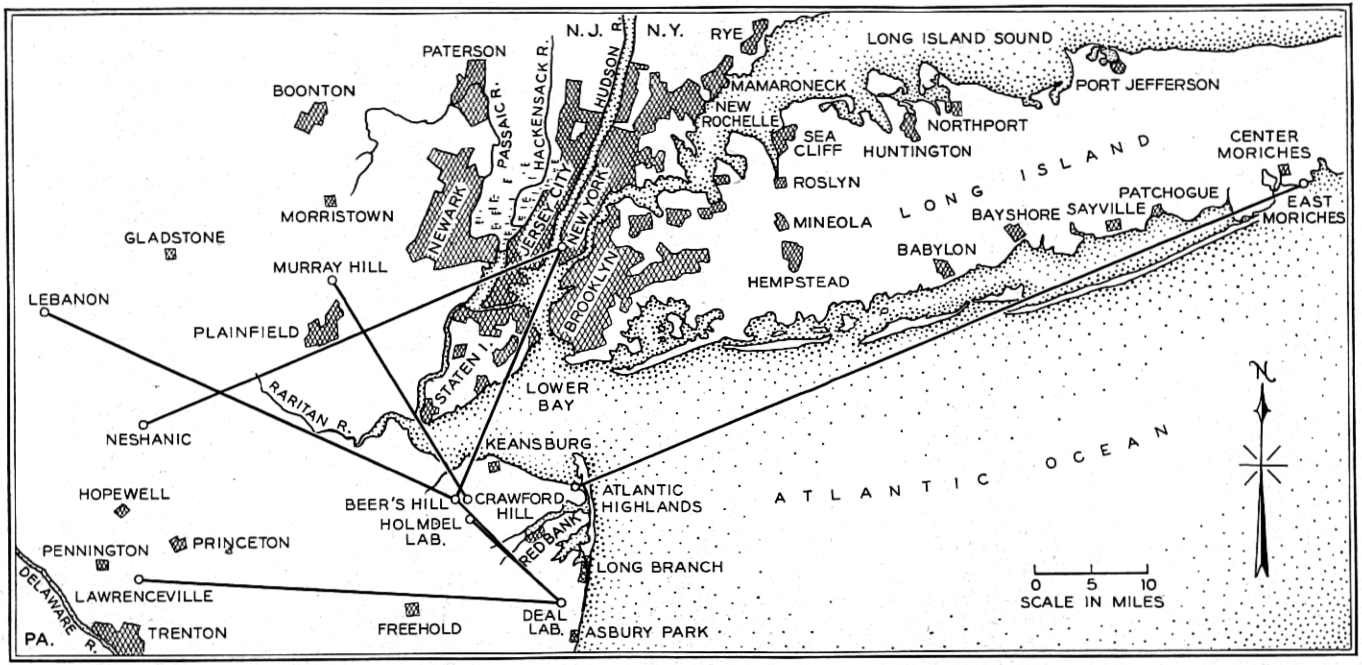


Fig. I-1.—Map showing principal Propagation Paths.

sight is well clear of intervening obstructions. In order to have a rule-of-thumb as to the amount of path clearance desirable, we have suggested that the first Fresnel region should be clear of all obstacles. The first Fresnel region for a given transmitter and receiver is bounded by points for which the length of the path, transmitter to point to receiver, is greater by one-half

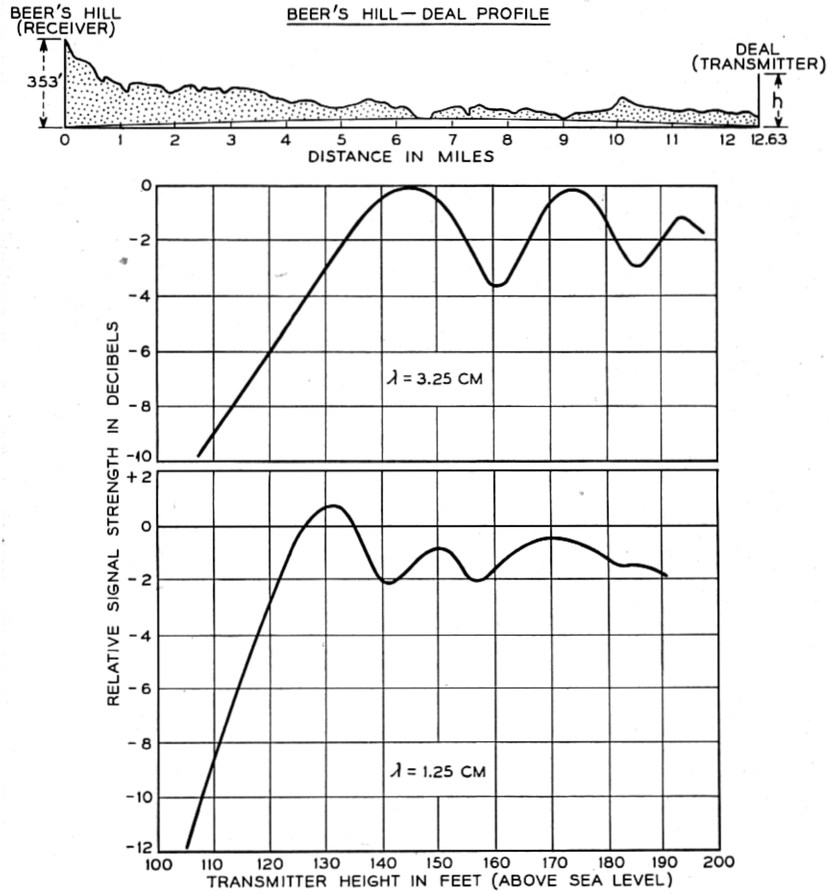


Fig. I-2.—Variable antenna-height tests on Beer's Hill-Deal Path.

wavelength than the direct path from transmitter to receiver; its cross-section by any plane perpendicular to the direct path is the first Fresnel zone in the sense used in optics. A wave can be transmitted with practically no loss through an opening whose area is of the order of the first Fresnel zone. Also, in the case of a smooth reflecting surface between transmitter and receiver, the first Fresnel zone clearance provides a maximum in received field strength since the half wavelength path difference plus the 180-degree phase change at reflection causes the direct wave and the reflected wave to arrive in phase at the receiver. In Fig. I-4, the first Fresnel region is sketched on the profile map of a typical microwave link for wavelengths of 3 meters and 3 centimeters.

It should be emphasized that the above remarks on ground reflections apply only for rough terrain and for the case of reflection at a distance from

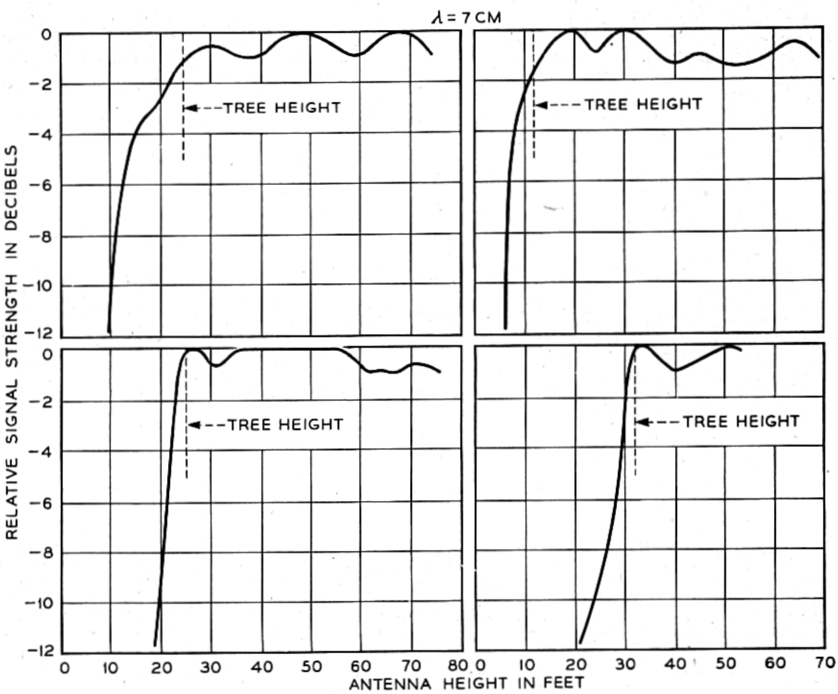


Fig. I-3.—Variable antenna-height tests on several of the New York-Boston repeater circuits. $\lambda = 7$ cm.

the terminals. Variable height experiments involving short distances over open flat fields reveal the presence of almost perfect ground reflections at wavelengths as short as 1.25 centimeters. For transmission paths over water, strongly reflected components are often observed. Reports[3] of experiments in the Arizona desert indicate a strong ground reflection at wavelength of 3 centimeters. In such locations, and most likely in the plains regions, the presence of substantial ground reflected components may prove to be troublesome.

[3] Report No. 6. Electrical Engineering Research Laboratory, The University of Texas, February 1, 1947.

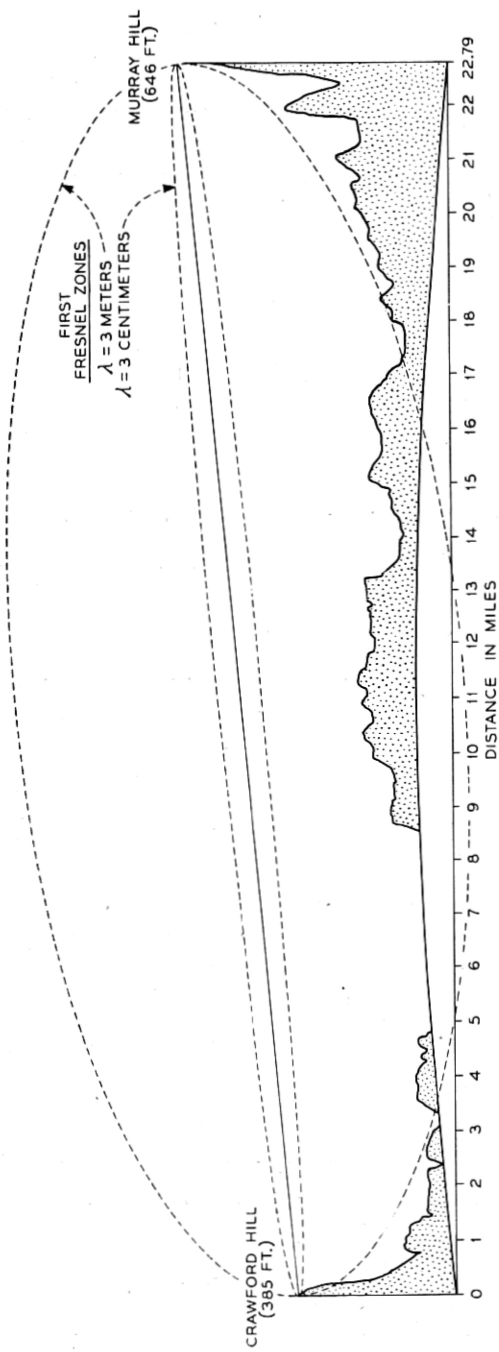


Fig. I-4.—Profile map of Murray Hill-Crawford Hill path showing first Fresnel regions for wavelengths of 3 meters and 3 cm.

Atmospheric Refraction

As in the case of ground reflection, the refractive effect of the atmosphere has been found to play a somewhat varying role, depending upon the wavelength employed. In some of the early work on ultra-short-wave propagation,[1, 2] the concept of average atmospheric refraction was found to bring about better agreement between observed and calculated results. Due to the variation of temperature and water vapor content of the atmosphere with height above ground, the dielectric constant of the atmosphere normally decreases with height. The effect of this negative dielectric constant gradient is to cause the path of a radio wave to be bent slightly downward toward the earth, thus effectively increasing the horizon distance. It has been suggested that a good approximation for average refraction was to assume the radius of curvature of the ray to be four times that of the earth.[1] This condition is used at the present time to describe a "standard atmosphere."

It was soon found, however, that atmospheric refraction could vary between rather wide limits depending chiefly on the gradient of water vapor with height.[4] Refraction effects were found to be greater in summer than in winter since the air contains a higher percentage of water vapor in the summertime. A diurnal variation in refraction was also observed on overland transmission paths. During the day, rising convection currents and surface winds, caused by surface heating of the earth, usually produce a well mixed atmosphere near the earth so that "standard" atmospheric conditions prevail. On clear nights, however, particularly if the wind velocity is low, radiation cooling of the earth may cause a temperature inversion in the lower atmosphere; if, also, the water vapor decreases with height, the combined temperature and water vapor effects may add to produce a steep negative gradient in the dielectric constant. Stormy weather and overcast skies usually result in standard atmospheric conditions.

Most of the signal variations observed during a two-year study of propagation of two and four-meter waves over the 39-mile over-land optical path between Beer's Hill, N. J. and Lebanon, N. J.[5] could be explained satisfactorily on the basis of wave interference between direct and ground-reflected radiations; the relative path lengths, and hence the phases, of these two components of the received field varied with the refractivity of the atmosphere. The fading on the two wavelengths was usually similar in major detail as might be expected from the geometry of the path. On the

[1, 2] Loc. cit.

[4] Englund, Crawford and Mumford, "Further Studies of Ultra-Short-Wave Transmission Phenomena", *B. S. T. J.*, vol. 14, pp 369–387; July 1935.

[5] Englund, Crawford and Mumford, "Ultra-Short-Wave Transmission over a 39 mile 'Optical' Path", *Proc, I. R. E.*, vol. 28, pp. 360–369; August 1940.

few occasions when the fading could not be accounted for in this simple fashion, it was assumed that signal components were arriving from above by virtue of reflections from small, relatively abrupt changes in the dielectric constant of the atmosphere. The existence of such reflections was demonstrated by a frequency-sweep method during propagation studies on the 70 mile over-water path between Highlands, N. J. and East Moriches, Long Island.[6]

With microwaves, where, as stated previously, ground reflections are usually small or absent, one might surmise that changes in atmospheric refraction would have a smaller effect on transmission than at ultra-short-wavelengths where strong ground reflections are present, and that fading should, therefore, be less. Actually the opposite is observed. Fading is found to be more frequent, faster, and deeper as the wavelength is decreased. This frequency effect may be explained in a qualitative fashion by a consideration of the relative sizes of Fresnel zones at ultra-short waves and at microwaves. It is known that the dielectric constant of the atmosphere usually does not vary with height in a smooth linear manner; on calm nights, particularly, very steep gradients in the dielectric constant may exist over small vertical ranges measuring only tens of feet. The effectiveness of these steep gradients would be expected to depend on their extent relative to the size of a Fresnel zone. Thus on a path such as that in Fig. I-4, a steep gradient extending over only a hundred feet would include practically the whole first Fresnel zone at 3 centimeters while it would cover only a small part of a zone at 3 meters wavelength; the effective gradient, therefore, would be considerably less at 3 meters than at 3 centimeters. Analyses based on wave theory show that atmospheric layers, in which the dielectric constant has a steep negative gradient, tend to confine or guide the radiation in much the same way as a waveguide, and that this "trapping" phenomenon, for a given layer thickness, becomes more pronounced as the wavelength is decreased.[7]

The mechanism of microwave propagation is certainly a complicated one, and a considerable amount of experimental work in the fields of radio and meteorology will be required to unravel it. However, it is very difficult to interpret the radio measurements in terms of meteorological data. The chief difficulty is that meteorological measurements often do not give an accurate picture of the atmosphere, particularly at those times when microwave fading indicates that rapid changes of some sort are occurring in the

[6] Englund, Crawford and Mumford, "Ultra-Short-Wave Transmission and Atmospheric Irregularities", *B. S. T. J.*, vol. 17, pp. 489–519; October 1938.

[7] H. G. Booker, in England, was the first to call attention to this phenomenon. For more recent work see: C. L. Pekeris, "Wave Theoretical Interpretation of Propagation in Low-Level Ocean Ducts, *Proc. I. R. E.*, vol. 35, pp. 453–462; May, 1947. This paper gives references to other work in this field.

atmosphere. The instruments used to measure temperature and humidity require a few seconds to reach equilibrium—a length of time comparable at times with the period of fading. To measure the variation of dielectric constant with height, the measuring instruments are usually carried aloft by means of captive balloons. A half hour may be required to measure to heights of six or seven hundred feet with the result that the final curve represents an unknown combination of variations of dielectric constant with height and with time. It is also extremely doubtful that the atmosphere is uniform in the horizontal plane—an assumption which is usually made in the theoretical treatment of microwave propagation. It seems likely that the lower atmosphere is far from being a homogeneous fluid but rather may contain small air masses or "boulders" with properties which differ considerably from those of the surrounding air. Reflections from these boulders may be the cause of radar echoes received from the lower atmosphere.[8] Scintillation fading of microwaves is another evidence of these inhomogeneties in the atmosphere. Scintillation fading, a rapid fluctuation in signal level about a more or less steady average value, increases as the wavelength becomes less and as the path length is increased.

In order to evaluate, on a statistical basis, the effect of atmospheric changes on a typical microwave circuit, extensive measurements of transmission were made over a 40-mile overland path between New York City and Neshanic, New Jersey. The tests covered a period of about two years. Most of the data were obtained at wavelengths of 10, 6.5, and 3.2 centimeters although some data were taken at wavelengths of 42 centimeters and 1.25 centimeters. The results are described in a recent paper.[9] In many respects, observations were in agreement with those made earlier on the 39-mile Beer's Hill-Lebanon path at wavelengths of 4 and 2 meters and on the 38-mile non-optical path between Deal, N. J. and Lawrenceville, N. J. at a wavelength of 2 meters.[10] The same seasonal and diurnal trends in fading were found; transmission was generally steady during the midday hours and during periods of windy or rainy weather; fading was the same on vertical and horizontal polarizations. However, the character of the fading was different; the fading at microwaves was much faster and deeper than that observed on the ultra-short-wave path. The average daily fading range for July on the New Yoık-Neshanic path was 20 db at 6.5 centimeters compared with a median daily fading range of 8.5 db for 2.0 meters observed in July on the Lebanon-Beer's Hill path.

[8] H. T. Friis, "Radar Reflections from the Lower Atmosphere", *Proc. I. R. E.*, vol. 35, pp. 494–495; May 1947 (Correspondence Section).

[9] A. L. Durkee, "Results of Microwave Propagation Tests on a 40-mile Overland Path", *Proc. I. R. E.*, vol. 36, No. 2, pp. 197–205, Feb. 1948.

[10] C. R. Burrows, A. Decino and L. E. Hunt, "Stability of Two-Meter Waves", *Proc. I. R. E.*, vol. 26, pp. 516–528; May 1938.

Other observations on the New York-Neshanic microwave path may be summarized as follows: While all the wavelengths were affected at times of anomalous propagation, the shorter wavelengths faded more severely and the character of the fading was different from that observed at the 42 centimeters wavelength; apparently the 3–10 centimeter range was more sensitive to the fine structure of the atmosphere, as pointed out previously. During non-fading periods, signal levels were very close to the free-space values with the exception of the 1.25 centimeter signal which was usually 15 db or more below the free space value because of atmospheric absorption effects. Some special tests showed that fading was considerably more severe when one of the terminals was lowered so that the transmission path was grazing slightly below line-of-sight. It was also found that fading was about twice as great, in decibels, on the whole path as on either half-section. A statistical analysis, on an hourly basis, of all the data on 6.5 centimeters showed that only one-half of one percent of the total hours had signal minima deeper than 20 db below the free space field. Also during August 1, the day of the most severe fading, the signal was more than 20 db below free space for about one-half of one percent of the time. It was also found that signals of the order of 10 db above free space were equally probable. From a consideration of these statistics, it was decided to engineer the New York-Boston repeater circuit with -20 to $+$ 10 db allowance for fading on each link.

Specialized Experiments

Much of our more recent work on microwave propagation has been of a specialized nature in which apparatus and experiments have been designed more for the purpose of studying the mechanism of anomalous propagation than for making a statistical analysis of the transmission. Perhaps the most informative experiments have been those in which narrow beam scanning antennas were used to explore the incident wave fronts.

The first of these antennas had an aperture of 20 feet and a beam width between half-power points of $\frac{1}{3}$ degree at the design wavelength of 3.25 centimeters. It was built for the purpose of establishing a practical limit to the size, and hence the directivity, of microwave repeater antennas from the standpoint of variations in the angle of arrival of the received wave. It had been realized, of course, that variations in the refractivity of the atmosphere would cause some deviations in the path of the wave. While these deviations should be negligible in comparison with the beam width of antennas normally used in the ultra-short-wave region, they might be comparable with the beam widths readily obtainable in the microwave region.

Using this antenna for measurements in the vertical plane and another identical antenna for measurements in the horizontal plane, angle-of-arrival data were obtained during the summer of 1944 over a twenty-four mile, partly over-water, path between Beer's Hill, N. J. and New York and over a thirteen mile over-land path between Beer's Hill, N. J. and Deal, N. J.[11] In the horizontal plane, deviations in the angle of arrival were rather uncommon and were not greater than $\pm$ 0.1 degree from the true bearing of the transmitter. In the vertical plane, angles of arrival above the true elevation of the transmitter were observed to be as much as 0.5 degree on the New York path and 0.3 degree on the Deal path during times of anomalous propagation. From these measurements it was concluded that microwave repeater antennas could be made highly directive in the horizontal plane but should have beam widths somewhat greater than $\frac{1}{2}$ degree in the vertical plane unless means for steering the beams are provided.

Although the $\frac{1}{3}$ degree beam width of these scanning antennas was sharp enough to permit the separation of the direct and the water-reflected components on the New York path, and to demonstrate the anomalous behavior of each, there was evidence that occasionally there were signal components so close together in angle that a sharper antenna would be required to resolve them. Consequently a scanning antenna of the metal-lens type was constructed for operation at 1.25 centimeters. The aperture of this antenna was 20 feet in the long dimension; the beam width was 0.12 degrees. Using this antenna and also the 3.25 centimeter scanning antennas, angle-of-arrival measurements were made in the summer of 1945 on the Deal-Beer's Hill path.[12] The most noteworthy result of these observations was the demonstration of multiple-path transmission. Two, three and, at times, four distinct signal components were observed simultaneously during one night when the transmission was extremely disturbed. These transmission paths generally were above the true direction of the transmitter; at one time, a weak signal was arriving at an angle of 0.75 degree relative to the line of sight. These components varied in angle of arrival and in signal amplitude. Wave interference among them caused severe fading on broad beam antennas that would accept all the wave paths.

Another significant result of these angle-of-arrival measurements was evidence that the transmission mechanism was very similar for wavelengths of 3.25 and 1.25 centimeters. Angles of arrival, measured simultaneously at the two wavelengths, agreed very well for times of single-path transmission; multiple-path transmission was observed on both wavelengths although the 3.25 centimeter antenna was too broad to resolve the com-

[11] W. M. Sharpless, "Measurement of the Angle of Arrival of Microwaves", *Proc. I. R. E.*, vol. 34, pp. 837–845; November 1946.

[12] A. B. Crawford and W. M. Sharpless, "Further Observations of the Angle of Arrival of Microwaves", *Proc. I. R. E.*, vol. 34, pp. 845–848; November, 1946.

ponents completely. This result suggested that the 1.25-centimeter scanning antenna might be a very useful tool for investigating the fading mechanism at 7 centimeters wavelength.

The 22.8-mile path between Crawford Hill and a hill on the Murray Hill Laboratory property was chosen for study as a representative link in a repeater circuit. (See Profile Map, Fig. I-4.) Transmitters for the 1.25-centimeter and 7 centimeter wavelengths were installed in the 100-foot tower at Murray Hill. At the Crawford Hill receiving site were the narrow-beam scanning antenna and a broad beam antenna for 1.25-centimeter operation; also two broad beam antennas, spaced vertically 15 feet, for 7-centimeter operation. In addition, a 1.25-centimeter radar could be operated with the scanning antenna. A corner reflector target, $5\frac{1}{2}$ feet on a side, was located at the Murray Hill tower. The signal reflected by this target was about 10 db stronger than the spurious reflections from other objects at the same range. By making use of this target and ground echoes at intermediate distances, the radar technique provided a considerable amount of useful information concerning the transmission phenomena.

Measurements were made on this path during the summer of 1946. As had been hoped, the observations showed that transmission on 1.25 centimeters and 7 centimeters was often affected by the same conditions except, of course, for atmospheric absorption effects at the 1.25-centimeter wavelength. While it was not possible to arrive at explanations for all the fading observed, the deep minima in the 7-centimeter signal, i.e., fades to levels of 15 to 20 db or more below the free space field, usually were the result of one of three types of propagation*:—

Type 1. The 7-centimeter fading was of the rapid, large amplitude type characteristic of wave interference. The 1.25-centimeter scanning records showed the presence of multiple-path transmission in which two or more readily separable wave paths were observed. While the signals on both of the vertically-spaced 7-centimeter antennas faded about the same in amplitude, their signal minima did not occur simultaneously. A space diversity system would be successful in reducing the effects of this type of fading.

Type 2. The 7-centimeter fading was somewhat slower than in Type 1, but still had the appearance of wave interference. The 1.25-centimeter scanning records appeared to be of the single path variety. However, close inspection showed that, in all probability, more than one transmission path was involved but the 0.12 degree beam of the antenna was not sharp enough to resolve them. The signals received on the vertically spaced 7-centimeter antennas faded together so that space diversity would not be expected to be successful unless an extremely large spacing of antennas were used.

* Recently, on a different overland path having barely one Fresnel zone clearance, an important fourth type has been observed when atmospheric refraction gives the ray path a curvature opposite to that of the earth, thus effectively reducing the path clearance.

Type 3. The 7-centimeter signal would fade to a low level and remain there for a considerable period of time; sometimes for an hour or so. The character of the fading was unlike that caused by wave interference. The 1.25-centimeter signal was simultaneously at a low level and the scanning records showed that only one path was involved. Reception was almost

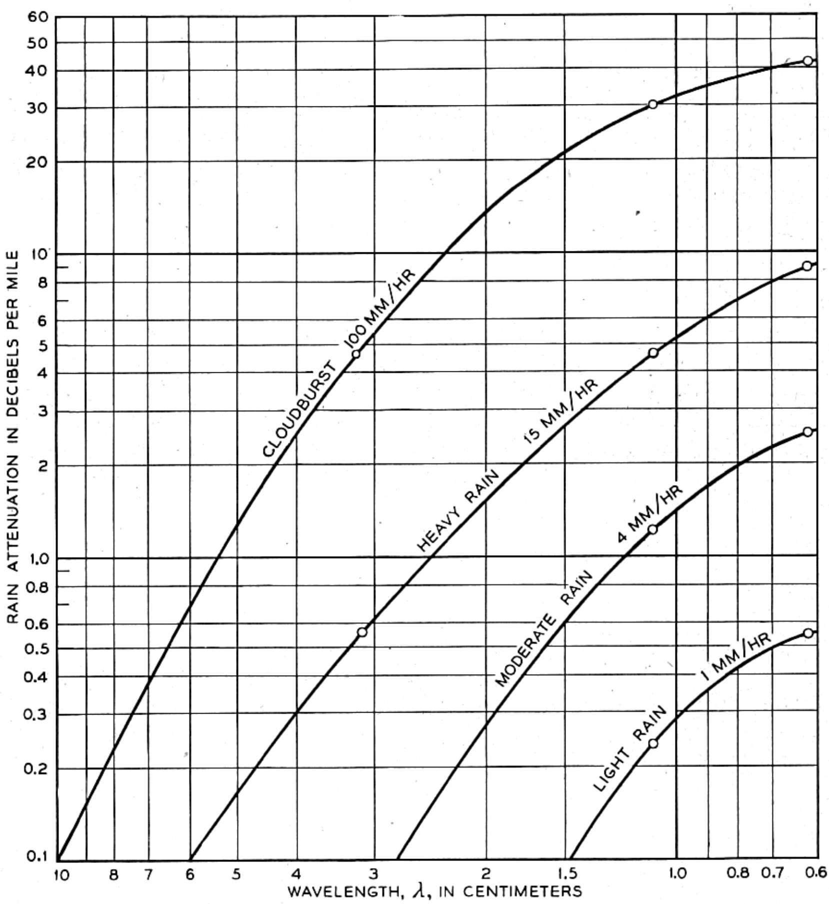


Fig. I-5.—Rain attenuation vs. wavelength.

identical on the two vertically-spaced 7-centimeter antennas. Radar observations suggested that this type of fading was due to attenuation by a reflecting layer in the atmosphere at a height intermediate to the heights of the transmitters and receivers. It was observed, for example, that while the echo from the Murray Hill corner reflector was absent, strong echoes were received from the hill directly in front of Murray Hill and some 250 feet lower in height; sometimes multiple paths were observed with this echo. Space diversity would fail to improve transmission under these propagation conditions, and no other means of improvement is apparent except, perhaps, an alternate path. Fortunately, this type of fading was the least frequent of the three types which were characterized by low signal levels.

Rain Attenuation and Atmospheric Absorption

Attenuation effects due to rainfall and absorption by atmospheric gases become increasingly important at the short-wave end of the microwave region. Measurements of rain attenuation have been made at the Holmdel Laboratory[13, 14]; the results are summarized in Fig. I-5. These curves show that for wavelengths above about 5 centimeters, rain attenuation is not very serious except for rains of cloudburst proportions. However, at wavelengths of one centimeter and less, even moderate rainfall will cause large attenuations on paths of the order of 10–20 miles in length.

Absorption by atmospheric gases, principally water vapor and oxygen, becomes important at wavelengths below about 1.5 centimeters. According to the theoretical work of Dr. J. H. Van Vleck, Harvard University, water vapor has an absorption band at 1.33 centimeters and oxygen has bands at 0.5 and 0.25 centimeters. Measurements made on the Deal-Holmdel path at 1.25 centimeters were in fair agreement with Van Vleck's results and indicated that a typical value of atmospheric absorption for this locality in summertime is about 0.4 db per mile.[14]

Summary

In the ultra-short-wave region, transmission has been found to be affected mainly by ground reflections and variable atmospheric refraction; only occasionally are atmospheric reflecting layers and trapping phenomena involved. These wavelengths ordinarily are not transmitted to great distances along the surface of the earth, but are diffracted around obstacles. They are used for local broadcasting and mobile radio communication.

Microwaves are attractive for radio repeater circuits since they permit the use of wide transmission bands. Ground reflections are apparently of small importance with terrain such as that of the Eastern seaboard and substantially free-space propagation is obtained during non-fading periods over optical paths which have approximately "first Fresnel region" clearance. Atmospheric reflecting layers and trapping phenomena are frequently observed and signal variations are considerably greater than in the ultra-

[13] Sloan D. Robertson and Archie P. King, "The Effect of Rain upon the Propagation of Waves in the 1- and 3-Centimeter Regions", *Proc. I. R. E.*, vol. 34, pp. 178P–180P; April 1946.

[14] G. E. Mueller, "Propagation of 6-Millimeter Waves", *Proc. I. R. E.*, vol. 34, pp. 181P–183P; April 1946.

short-wave region. Although fading becomes worse as the wavelength is decreased, the advantages of increased antenna gain and directivity possible at the shorter wavelengths suggest the use of a wavelength just above the region where rain attenuation becomes objectionable; i.e. above about 5 centimeters.

The use of two antennas, operated in space diversity, should reduce the effects of fading caused by multiple-path transmission. The use of space diversity may be essential in those localities where strong ground reflections are present. On the basis of the comparatively weak ground reflections measured on the New York-Boston path it was decided to avoid the complications that would result from the use of space diversity in this experimental system.

II. Repeater Circuit Planning

The diagram in Fig. II-1 shows schematically a repeater circuit. At the input terminal toward the left the signal, S, is fed to the terminal's transmitting antenna. The radiated signal is propagated as discussed in Section I and produces the signal power s_1 in the output of the receiving antenna of repeater 1. The signal is then amplified G_1 times and radiated toward repeater 2 and this process is repeated until the signal finally appears in the output terminal towards the right. In each repeater the signal is gain-controlled automatically for the same level of output powers, i.e., $S = S_1 = S_2 = \cdots = S_n$. It is assumed that the signals are amplitude or frequency-modulated C.W. carriers of substantially the same frequency in each link and that the repeaters have linear amplifiers. The diagram shows a West-to-East circuit only. A circuit for the opposite direction requires duplication of all the equipment with the exception of the antenna supporting towers.

Some simple formulas for the repeater gain and the signal-to-noise ratio at the terminal will be given in this section, without going into any details on propagation phenomena, antennas, amplifiers, etc. The formulas will orient the reader in regard to the importance of quantities such as:

$$\left.\begin{aligned} d &= \text{Repeater separation} \\ \lambda &= \text{Wavelength of signal} \\ A &= \text{Effective area of each antenna} \end{aligned}\right\}\text{same units of length}$$

F = Noise figure of each repeater amplifier
B = Bandwidth of circuit in cycles/sec.

The free space attenuation (S_{x-1}/s_x) of link "x" is[15]

$$L_x = \frac{d_x^2 \lambda^2}{A^2}$$

[15] H. T. Friis, "A Note on a Simple Transmission Formula", *Proc. I. R. E.*, Vol. 34, No. 5, pp. 254–256; May 1946.

Allowing the signal power to fade by a factor M below this value, the maximum gain of repeater "x" must be

$$G_x = \frac{d_x^2 \lambda^2}{A^2} M_x \tag{II-1}$$

The total maximum gain in the circuit is $G_T = G_1 \times G_2 \times \cdots G_n$ which for the same repeater spacings and fading allowances becomes

$$G_T = G^n = \left(\frac{d^2 \lambda^2 M}{A^2}\right)^n = \left(\frac{d^2 \lambda^2 M}{A^2}\right)^{D/d} \tag{II-2}$$

where D is the total length of the circuit. Because of distortion, original costs, and maintenance costs, this total gain should be made as small as possible.

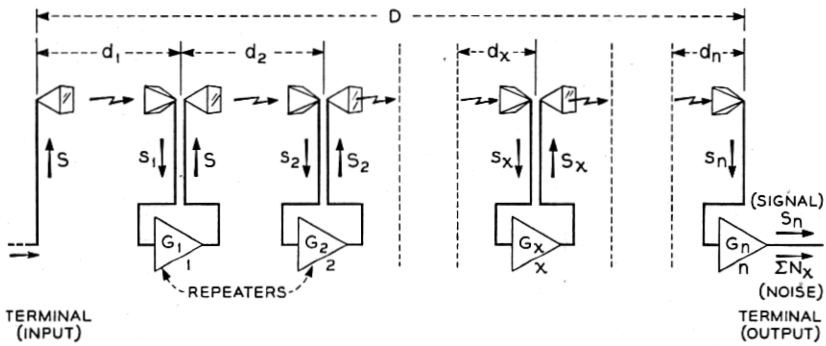


Fig. II-1.—Repeater circuit with n links.

The following example illustrates the maximum gain required of the amplifier in a repeater:

For $d = 4 \times 10^4$ meters (25 miles), $\lambda = 0.075$ meters ($f = 4000$ megacycles), $A = 4.6$ meter² (50 sq. ft.) and $M = 100$ (20 db), we have

$$G = 4.3 \times 10^7 \text{ (76 db)}$$

The noise output of a repeater due to noise sources in the repeater itself is approximately[16]

$$N_x = 4 \times 10^{-21} \, F \, B \, G_x \text{ Watts}$$

or from (II-1)

$$N_x = 4 \times 10^{-21} \, FB \frac{d_x^2 \lambda^2}{A^2} M_x$$

[16] H. T. Friis, "Noise Figures of Radio Receivers", *Proc. I. R. E.*, vol. 32, pp. 419–22; July 1944.

This noise power is transmitted without gain or loss to the output terminals of the repeater circuit. The total noise power at the output terminals is therefore

$$\Sigma N_x = 4 \times 10^{-21} FB \frac{\lambda^2}{A^2} \Sigma(d_x^2 M_x)$$

Assuming the same repeater spacings and fading factors M in each link and substituting D for nd,

$$\Sigma N = 4 \times 10^{-21} FBD \frac{d\lambda^2}{A^2} M \text{ Watts} \tag{II-3}$$

The signal-to-noise ratio at the output terminals is $S/\Sigma N$. The circuit should be designed for a signal power, S, as low as possible. Therefore, it is very important to choose values of the several factors in (II-3) which give a low level of output noise.

Assuming a noise figure $F = 20$ (13 db) and bandwidth $B = 10$ megacycles, eight repeaters of the type described in the above example will have,

$$\Sigma N = 2.8 \times 10^{-4} \text{ Watts}$$

or, assuming a required minimum output signal to noise ratio of 30 db, the output power must be $S \geqq 0.28$ Watts.

In this example it has been assumed that the signals in all links have faded simultaneously 20 db below the free space value, which may only happen a fraction of a percent of the time. Most of the time the signal-to-noise ratio will be higher than the assumed 30 db and under normal transmission conditions it will be 50 db (fading allowance factor $M = 1$).

Assuming the same repeater spacings and fading allowances in each link, equation (II-1) and (II-3) give the following formula for the ratio of the output power to noise figure of the repeater amplifier

$$S/F = 4 \times 10^{-21} G\, B\, (S/\Sigma N) n \tag{II-4}$$

Equations (II-1) and (II-4) are the important design equations for the repeater amplifier.

The factors in (II-2) and (II-3) will now be discussed briefly. (II-3) shows that the noise figure F should be as small as possible. If by improving the equipment the noise figure could be halved, then the signal power S could also be halved (unless interference from other microwave circuits predominate). Later on the noise figure will be discussed further.

The bandwidth B is determined by the characteristics of the signal it is desired to transmit and by the method of transmission. Our aim has been to provide 10-megacycle bands which are sufficient for transmission of standard television signals by AM or low index FM.

An increase in the effective area, A, of the antennas reduces both the total gain required of the amplifier and the output noise power. Crosstalk between the several antennas in a repeater station and interference from outside sources also decrease as the antennas are increased in size because of the increased directivity. Therefore, the antennas should be as large as maintenance and initial costs will permit. Antennas will be discussed in detail in Section III.

Equations (II-2) and (II-3) show that the wavelength λ should be small. Also more frequency space or signal channels may be had at shorter wavelengths. On the other hand, the fading factor M increases somewhat as the wavelength is decreased and, besides, attenuation due to rain sets a lower limit for λ in the region of 5 centimeters. The status of the apparatus art has also been an important factor, but it now permits utilization of the wavelength range extending upward from 3 centimeters. Since the war, our work has been concentrated on a 10% band around 4000 megacycles or $7\frac{1}{2}$ centimeter wavelength. The manner in which this 4000-megacycle band may be divided up into separate channels is explained in Section IV.

The effects of varying the repeater separation d will now be discussed. d appears in the denominator of the exponent of (II-2) which indicates that large separations are favorable, while (II-3) shows that a decrease in separation cuts down the noise. Small separations are very costly, the cost being almost inversely proportional to the separation. We have concluded from propagation studies and site surveys that in the eastern part of the United States it is desirable to use separations of about 30 miles, which generally provide line-of-sight paths with reasonable tower heights. It should be mentioned that the fading allowance factor M is not independent of d; an increased d requires a larger fading factor.

III. Antenna Research*

There are three electrical characteristics which repeater antennas should possess. The first is high gain (large effective area), as this will reduce the path loss and accordingly the requirements on transmitter power. The second is good directional qualities so as to minimize interference from outside sources and also interference between adjacent antennas. The third is a good impedance match so that reflections between the antenna and the repeater equipment will not distort the transmitted signals. These characteristics should preferably be attainable without the imposition of severe mechanical or constructional requirements.

It was felt that a 10-foot round or square antenna would be the largest that maintenance and initial cost would permit. Propagation studies also

* This section prepared by W. E. Kock who performed the major part of the work on antennas.

showed that the variations in angle of arrival of the distant signals would be small compared to the beam width of 10-foot antennas. First experiments were therefore made with 10-foot diameter parabolic "dish" type antennas. The experimental models were made of wood with a metallized reflecting surface consisting of silver conducting paint. Fairly satisfactory tolerances were met in these first models, but it was anticipated that trouble would be experienced in constructing a permanent metal paraboloid of that size to the required tolerances without the use of a heavy and costly supporting means for the parabolic sheet. It was also found that an ice coating a quarter wavelength thick on the reflecting surface, when wet, acted as an effective absorber of power,† since the sheet of water is resistive and is backed up by the reflector. Such a condition could produce an intolerable drop in received signal and would have to be prevented by providing the dish with a plastic cover. As this cover should preferably house the feed also, it would have presented a difficult supporting problem.

Two electrical shortcomings of the paraboloid antenna also presented themselves. First, it was found extremely difficult to obtain a satisfactory impedance match between the antenna and the feed line. This was true partly because of energy reflected from the dish re-entering the feed horn, (this produced a constant 0.6 db standing wave ratio in the feed line), and partly because of the problem of matching the feed horn itself over the desired 400 megacycle band. Secondly, the mutual interference or "crosstalk" between two paraboloids was found to be only 50 to 60 db down when placed back to back** (Fig. III-1).

A type of reflector antenna was later investigated, which, although larger physically than a dish having the same aperture area, overcomes the above two objections.[17] It is shown sketched in Fig. III-2. The photograph of Fig. III-3 shows the antenna lying on its side. It can be seen that the feed is effectively "offset" and reflection back toward the feed is eliminated; the experimental model of Fig. III-3 showed only 0.1 db standing wave ratio in the feed line over a 10% band of frequencies. Furthermore, a horn or "shielded" type feed is used which confines the energy and minimizes stray radiation, and measurements indicate that the back-to-back crosstalk suppression of two such antennas will be high. This long horn is also partly responsible for the excellent impedance match. A horn having a large aperture "matches" free space quite well and the slight mismatch at the throat can be tuned out over a wide band of frequencies. This is not true

† A waveguide termination in common use today employs a resistive sheet placed one-quarter wavelength in front of a conducting plate; this device absorbs practically all the power falling on it.

** Back to back crosstalk suppression in the order of 125 db would be desirable for repeaters receiving and transmitting on the same frequency.

[17] *U. S. Patent* #2,236,393, H. T. Friis and A. C. Beck.

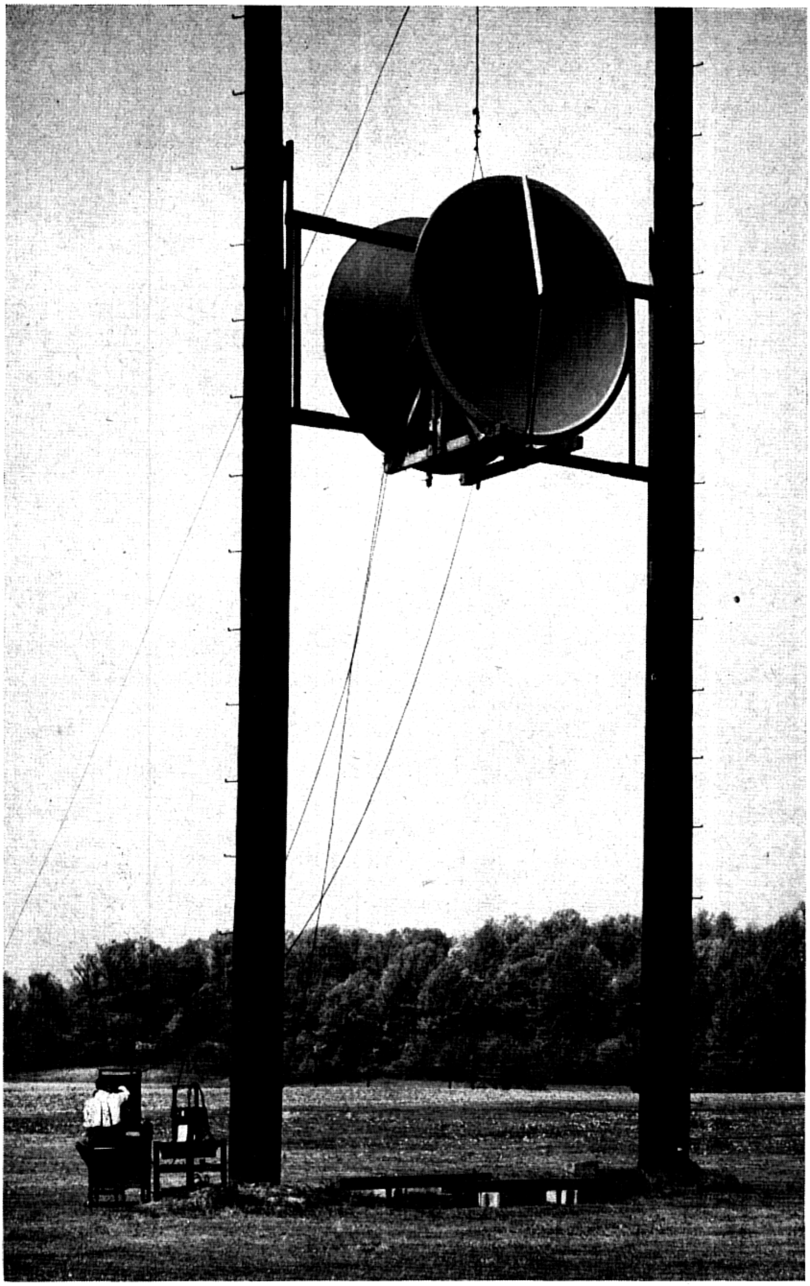


Fig. III-1.—Measuring the back to back crosstalk of two 10 ft. paraboloid antennas.

of the short, small aperture horn used in feeding the dish antenna. The antenna displayed an effective area which was 66% of its actual area which is only 0.9 db below the theoretical maximum.

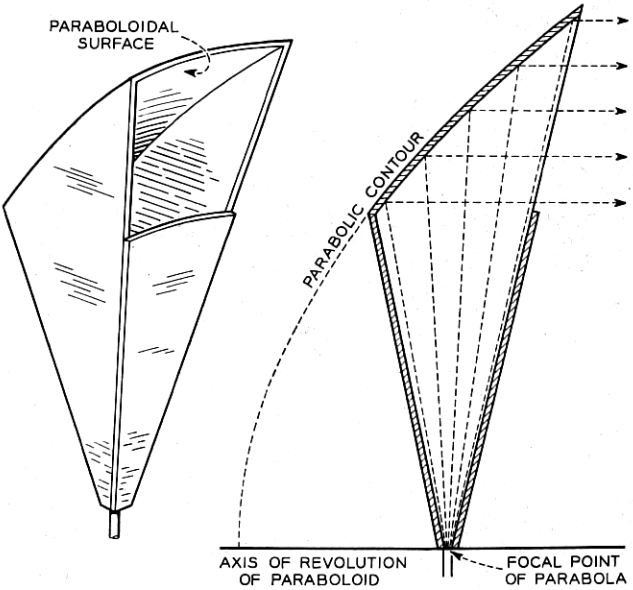


Fig. III-2.—Schematic of horn-reflector antenna.

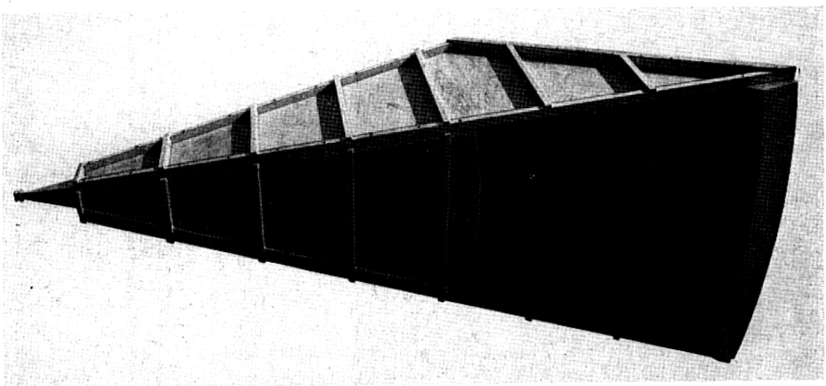


Fig. III-3.—Experimental model of horn reflector antenna.

The expected gain and directional characteristics of an antenna can be realized only if the emerging wave fronts are truly plane. Since deviations greater than $\pm \frac{1}{16}$ wavelength can materially impair the antenna performance,[18] reflector antennas have difficult tolerance requirements imposed upon them. For example, at 4000 megacycles, the 10-foot reflector must conform to parabolic shape to within $\pm \frac{1}{8}$ inches, and any twist or warp

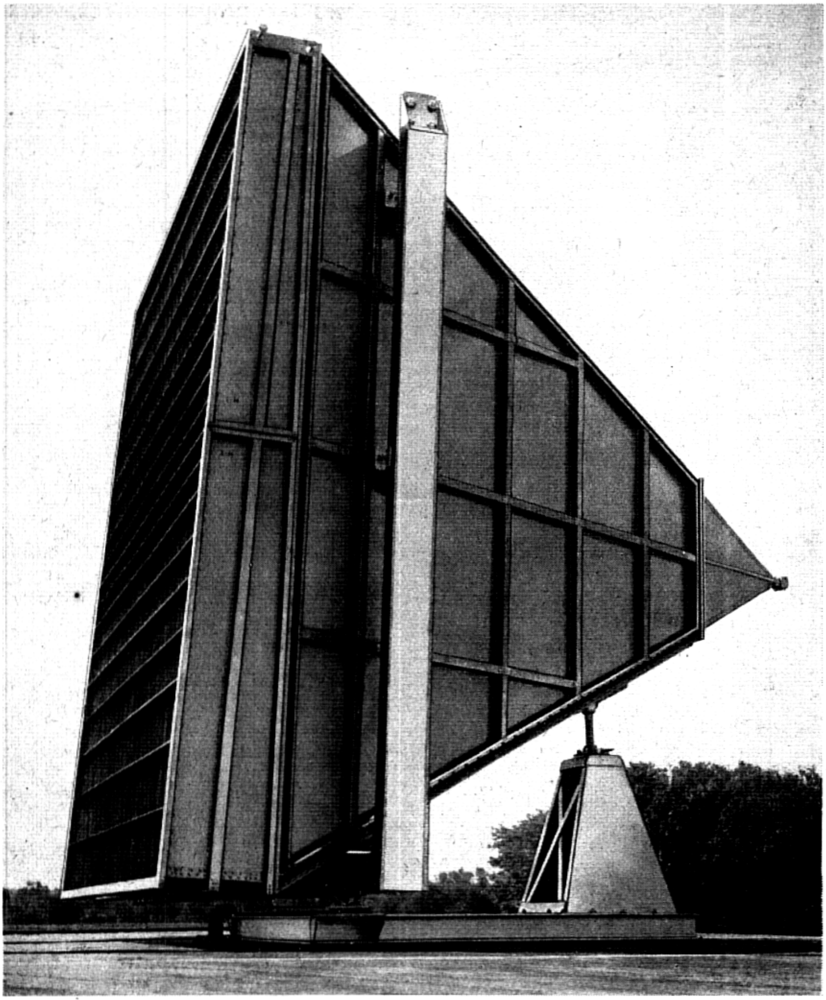


Fig. III-4.—Shielded metallic lens antenna.

of the reflector greater than this would be objectionable. Lenses, however, possess the property that a twist or warp in them does not impair their

[18] See, for example, "Radar Antennas", H. T. Friis and W. D. Lewis, *B. S. T. J.*, vol. 26, p. 219, April 1947, Figs. 17 and 28.

beam-forming properties, and, with the development of metal lenses for microwaves,[19] this type of antenna appeared to lend itself very well to repeater applications. The "shielded" type lens, which is a lens in the mouth of a short horn, is shown in Fig. III-4. This antenna, which was developed for the New York-Boston circuit,* possesses the property of excellent

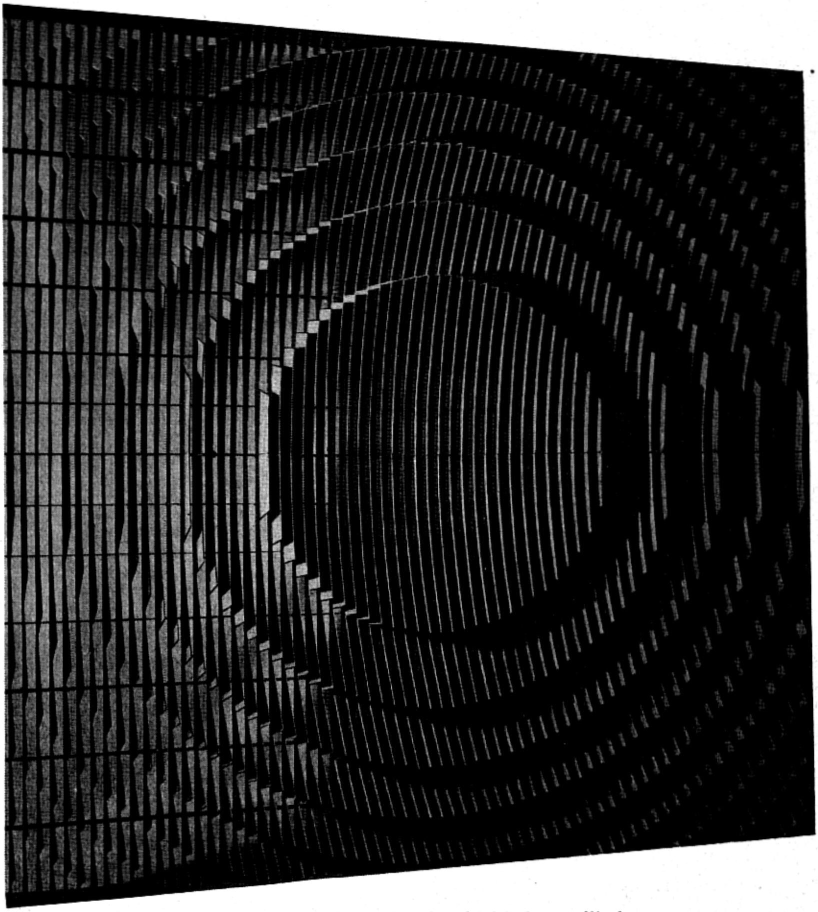


Fig. III-5.—Internal view of lens for the shielded metallic lens antenna.

crosstalk suppression both side to side (85 db) and back to back (125 db). Within the horn, the small amount of energy reflected back from the lens is directed away from the feed by tilting the lens, a procedure which does not noticeably affect the radiation characteristics, but which results in a fairly

[19] W. E. Kock, Metal Lens Antenna, *Proc. I. R. E.*, vol. 34, p. 828, November 1946.
* Developed by W. E. Kock and R. W. Friis.

good impedance match (under 0.8 db standing wave ratio) over the desired 400-megacycle band of frequencies. The lens in the mouth of the horn also provies a convenient support for a plastic impregnated Fiberglass sheet which acts as a weatherproof cover and protects the lens against ice forming between the plates.

The lens itself, Fig. III-5, is based on waveguide principles and causes the wave to be refracted by virtue of the fact that waves confined between plates parallel to the electric vector acquire a phase velocity higher than their free space velocity in accordance with the equation:

$$v_{\text{lens}} = v_{\text{free space}} \Bigg/ \sqrt{1 - \left(\frac{\lambda}{2a}\right)^2}, \qquad \text{(III-1)}$$

where λ is the wavelength and, a, the distance between the plates. The index of refraction is thus less than one, and a converging lens must be made concave.

As seen in Fig. III-5, the lens is stepped to reduce its thickness. As a consequence of this stepping, the efficiency at midband of the antenna (50%), is a good deal less than the theoretical value of 81%. Furthermore, the index of refraction varies with wavelength, as seen from equation III-1, and this results in a defocussing of the lens, with a consequent drop in gain, at wavelengths different from the design wavelength. This amounts to a drop in gain of 1.5 db at the edges of a 400 megacycle band; however, its other characteristics of impedance, side lobe suppression (70 db in the two rear quadrants), crosstalk, and ease of construction, help to make up for the gain deficiency.

Measurements taken on the antenna when a thick coating of ice had formed on the plastic cover indicated that the impedance match was impaired (the maximum standing wave ratio increased from .8 db to 1.6 db), but that the gain was not appreciably affected (less than 1 db). Since propagation experiments indicate that severe atmospheric fades are not likely to occur during the winter months, some of the fading allowance can be applied against ice loss.

There was some doubt that the crosstalk figures quoted above could be relied upon during heavy rainfalls, as there was indication that the signal transmitted from one antenna might be reflected from the rain and thus caused to enter an adjacent side-by-side antenna. Measurements during a moderately heavy rain proved that this effect was small, but large enough so that the 85 db side-to-side figure was approaching a limit for the 4000 megacycle band.

The measurement of antenna characteristics involves microwave techniques whose development is an important part of a research program.

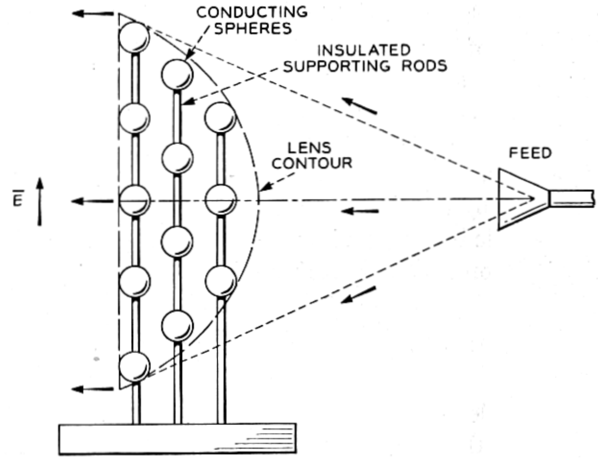


Fig. III-6.—Schematic of metallic delay lens using metal balls as the delay elements.

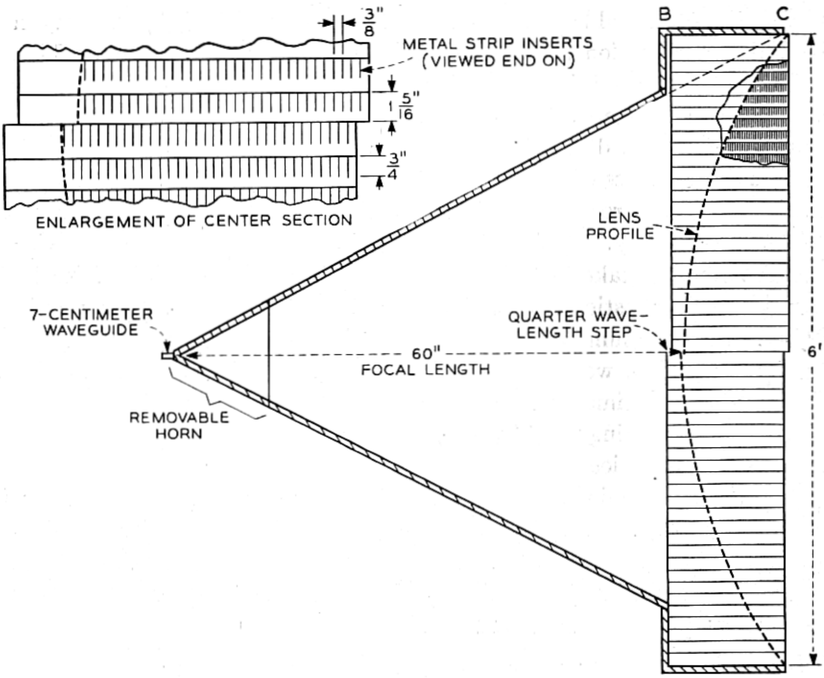


Fig. III-7.—Schematic of metallic delay lens for repeater applications using metal strips as the delay elements.

The antenna measuring methods which were employed in our repeater research follow along the lines of those described in a recent paper.[20] The very large signal ratios of 120 db or more necessitated double detection receivers and low noise figure crystal converters. Pattern and gain measurements of the antennas required measuring sites having large unob-

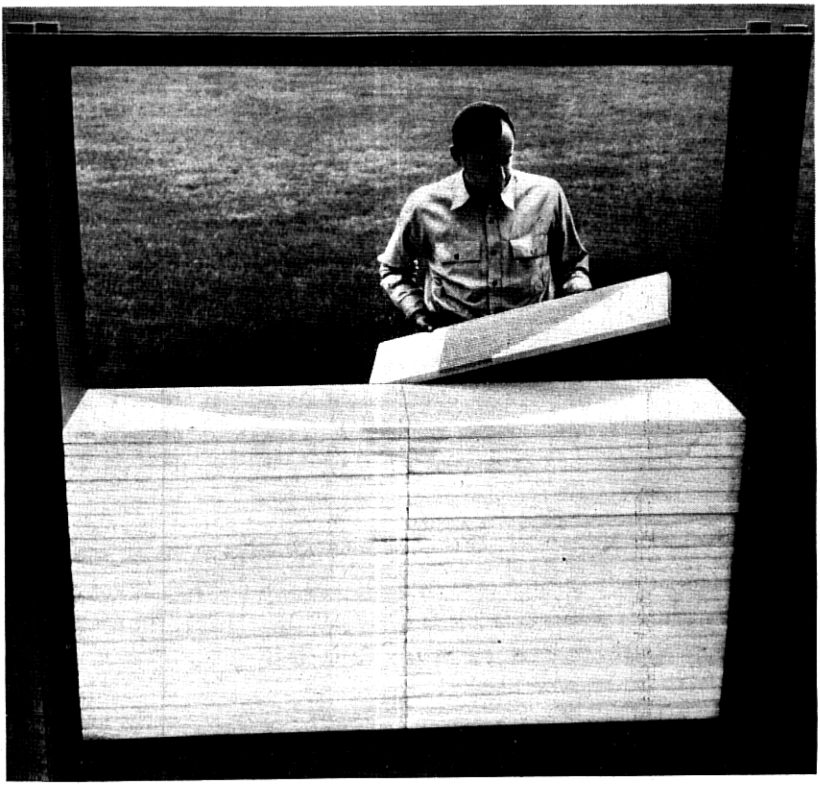


Fig. III-8.—A view of the partly assembled lens of Fig. III-7.

structed areas; these measurements were conveniently taken at the Holmdel Radio Research Laboratory. Impedance measurements involved the usual microwave equipment such as standing wave detectors in waveguide form, signal generators and calibrated receivers.

Research is now underway on an improved metal lens with gain and bandwidth properties which are superior to the lens of Fig. III-5. These lenses[21]

[20] C. C. Cutler, A. P. King, W. E. Kock, "Microwave Antenna Measurements", *Proc. I. R. E.*, vol. 35, No. 12, pp. 1462–1471, December 1947.

[21] Winston E. Kock, "Metallic Delay Lenses", *B. S. T. J.*, vol. 27, pp. 58–82, January 1948.

employ an artificial dielectric material which duplicates, on a much larger scale, processes occurring in a true dielectric. This involves arranging conducting elements in a three dimensional array or lattice structure to simulate the crystalline lattice of the true dielectric. Such an array responds to radio waves just as a molecular lattice responds to light waves, and if the spacing and size of the elements is small compared to the wavelength, the index of refraction is substantially constant, so that lenses made of this material are effective over large wavelength bands.

A lens employing conducting spheres as the lattice elements is sketched in Fig. III-6. A more convenient structure for large lenses is shown in Fig. III-7 and III-8; it uses thin metallic strips, with the width dimension parallel to the electric vector. Slotted polystyrene foam sheets support the strips and they are stacked up to form the lens. A quarter wavelength step in the lens causes the reflections from the lens surfaces to cancel at the feed point, which, in the drawing, is the apex of the horn shield.

Over a 10% wavelength band, a 6 foot square shielded lens antenna of this type exhibited an efficiency of better than 60% and the impedance mismatch due to the lens produces only a 0.2 db standing wave ratio in the feed line. This antenna thus retains the dimensional tolerance, weight, size and crosstalk advantages of the shielded lens over the shielded reflector, and has the advantage of higher gain and broader band performance over the shielded metal plate lens.

IV. Filter Research*

Frequency space for common carrier radio relay systems is available in blocks several hundred megacycles wide. Where heavy traffic is to be carried such bands must be efficiently exploited. This may in time be accomplished by using extremely wide band amplifiers, for example traveling wave tubes; however, more immediate success is offered by the possibility of operating a number of narrower band circuits of different frequencies. This could be done by using a separate transmitting and receiving antenna for each circuit. But each antenna, for sound technical reasons, must be large and expensive and in addition requires adequate tower support. Consequently there is a need for filters which can connect a number of individual radio channels to a common antenna.

The design[22] of these radio frequency branching filters must be coordinated with the design of the relay system as a whole. At lower frequencies where little or no antenna crosstalk protection can be counted on it is natural

* This section prepared by W. D. Lewis and L. C. Tillotson who were responsible for a large part of the research on filters.

[22] For more detailed discussion see, W. D. Lewis and L. C. Tillotson "A Constant Resistance Branching Filter for Microwaves," *B. S. T. J.*, vol. 27, pp. 83–95, Jan. 1948.

to lump transmitting frequencies in one group and receiving frequencies in another. When separate microwave shielded lens antennas are employed for transmitting and receiving in each direction it becomes practical to use a frequency plan in which transmitting and receiving frequencies are interleaved. Such a plan eases filter requirements considerably and has

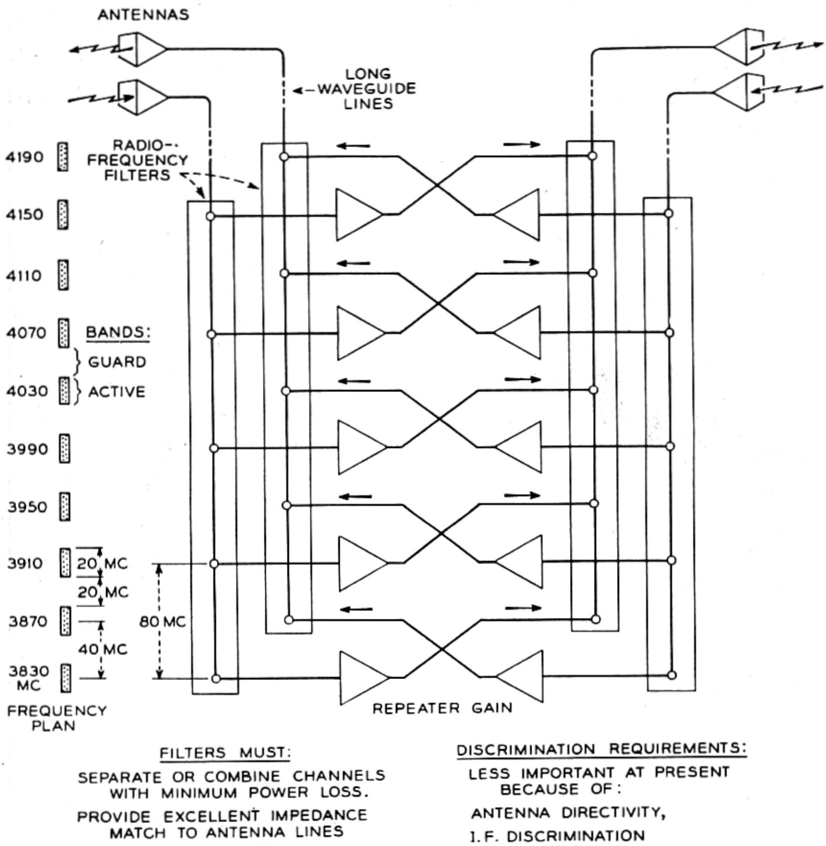


Fig. IV-1.—Schematic diagram of a possible five channel radio repeater station.

certain other advantages to be discussed in Section V. A possible repeater employing such a frequency scheme is illustrated schematically in Fig. IV-1.

If a radio frequency branching filter is to fit properly into a repeater it must separate or combine channels without excessive loss of signal. In addition it must provide an excellent match to the long line which leads to the antenna, otherwise troublesome echoes in this line may be caused.

Because of IF amplifier band-pass characteristics, suppression requirements on the filter, except possibly in the vicinity of receiver image bands, are not large.

Microwave band-pass filters consisting of one or more cavities arranged in sequence along a waveguide have been known for some time. The frequency, bandwidth and discrimination characteristics of such filters can all be chosen within wide limits by appropriate design of the cavities and the means for coupling to them. These filters are analogous to lumped-circuit channel-passing filters and can in principle, like them, be connected in groups to provide a branching network.

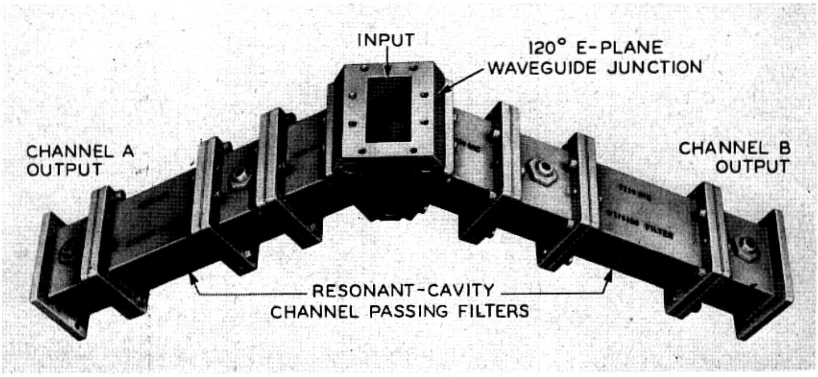


Fig. IV-2.—Photograph of a branching filter for an experimental radio relay system.

Several successful two-branch networks have been designed and constructed in this manner. One of these, developed for the New York-Boston circuit*, is illustrated in Fig. IV-2. Here two three-cavity filters are connected to an E plane Y junction, the waveguide analogue of a series connection. The two filters are tuned to different bands and each is connected to the junction through a line of length such that it causes no disturbance in the channel of the other. The electrical characteristics of this filter are plotted in Fig. IV-3.

Problems connected with the design of suitable microwave branching filters with more than two branches evidently differ considerably from previous filter problems. Channel passing networks which can be connected in series or parallel to form a channel branching filter can be designed at lower frequencies on the basis of lumped circuit theory and built of coils, condensers and resistances, but in the microwave region simple elements

* Developed by the group concerned with high-frequency filter design headed by A. R. D'heedene. A large part of the research underlying the design of these filters was performed by W. W. Mumford. Prior to the war a considerable amount of research on the band-pass type of waveguide filter had already been done by A. G. Fox.

and connections do not exist. Where more than two waveguide channel passing filters are to be connected to a common junction the design becomes complex, since in every channel the sum of the interactions of all the inactive filters on transmission through the active filter must be zero. It is evidently not easy to satisfy this condition, particularly since in doing so one must take account of the change with frequency of the effective length of all waveguide connecting lines. And even if such a solution is found it will be valid for only one set of channels, so that the problem must be solved all over again for every change in channel arrangement.

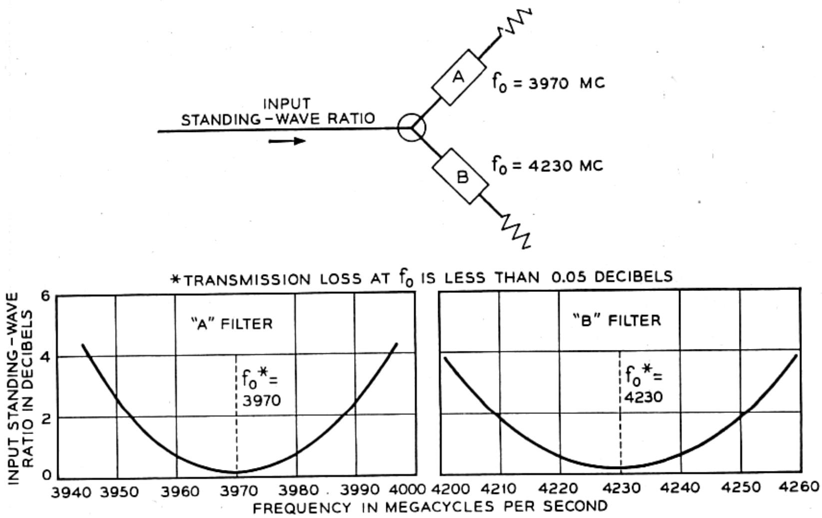


Fig. IV-3.—Input standing wave ratios of filter of Fig. IV-2.

As a result of these difficulties and after a few attempts to overcome them it became evident that a more flexible microwave branching filter technique should be found. Accordingly a solution on an iterative basis was developed. A channel dropping circuit was devised which, when inserted in a line, could extract or insert one channel while allowing others to pass through without disturbance. This circuit is of the constant resistance type; in other words it operates by diverting energy selectively but not by reflecting it back to the input. Consequently N such circuits placed in sequence do not interact reflectively; they, thus, form an N channel branching filter which is also of the constant resistance type.

An individual constant resistance channel dropping circuit is illustrated schematically in Fig. IV-4. It is made up of two hybrid[23] circuits, two

[23] For a general discussion of hybrid circuits see W. A. Tyrrell, "Hybrid Circuits for Microwaves", *Proc. I. R. E.*, vol. 35, No. 11, pp. 1294–1306, November 1947.

identical band reflection filters, and two quarter wavelengths of line. Each of the hybrids is analogous to a low-frequency hybrid coil and operates as follows. A wave in line C (See Fig. IV-4) incident on the hybrid is divided equally and with equal phase into A and B but does not appear in D or reappear in C. If waves in A and B are incident on the hybrid a wave proportional to their vector sum will appear in C, a wave proportional to their vector difference will appear in D but nothing will reappear in A or B. A wave in the input line incident on the channel dropping circuit will thus be divided by the input line into the lines leading to the two band reflection filters. These filters are designed to reflect frequencies lying within their band and pass all other frequencies. If the frequency is outside of the reflected band the two waves will travel on to connections A and B of the

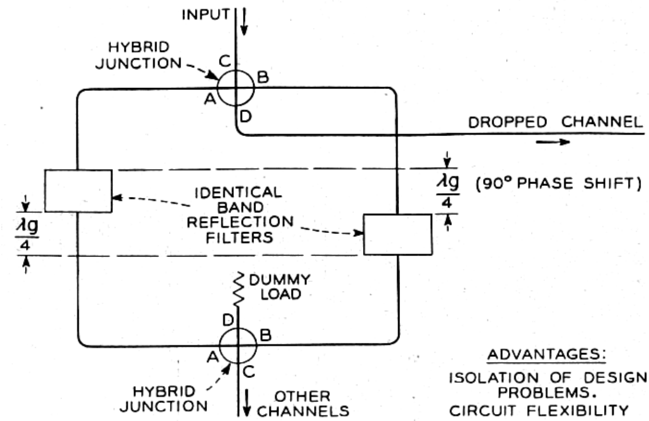


Fig. IV-4.—Schematic diagram of a constant impedance channel dropping filter using hybrid junctions and band reflection filters.

output hybrid. Here they will have equal phase and amplitude, their vector difference will be zero and the wave appearing in C of the output hybrid and consequently in the output line will contain all the power. If the frequency lies within the band of the reflection filters the two waves will be reflected by them and will travel back to the connections A and B of the input hybrid. The two waves strike these connections with opposite phase since one of them has traveled twice over an extra quarter wavelength of line. Their vector sum will consequently be zero and the wave which appears in terminal D of the input hybrid and consequently in the dropped channel line will contain all the power. The circuit of Fig. IV-4 is therefore a constant resistance channel dropping network which diverts energy lying within the band of the reflection filters but allows all other energy to pass through without disturbance. Conversely, by the law of reciprocity, this circuit can insert energy lying within the band of the reflection filters into the main line without disturbing any energy passing through it at other frequencies.

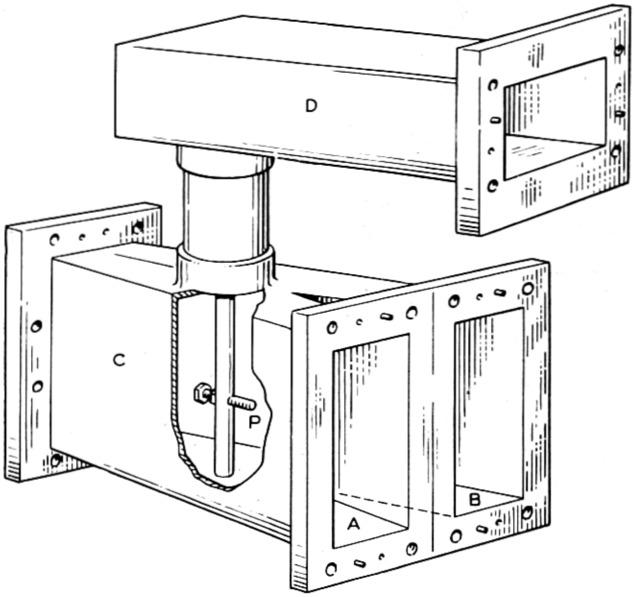


Fig. IV-5.—Hybrid junction used in the filter of Fig. IV-4.

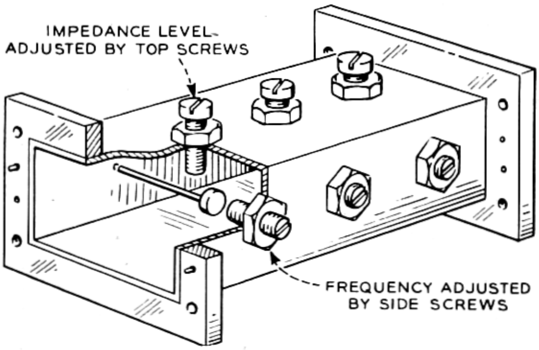


Fig. IV-.6—Waveguide band reflection filter used in the filter of Fig. IV-4.

An embodiment of the circuit of Fig. IV-4 suitable for use in the repeater arrangement of Fig. IV-1 has been constructed and tested. Figure IV-5 illustrates the waveguide hybrid employed. Here the waveguide opening

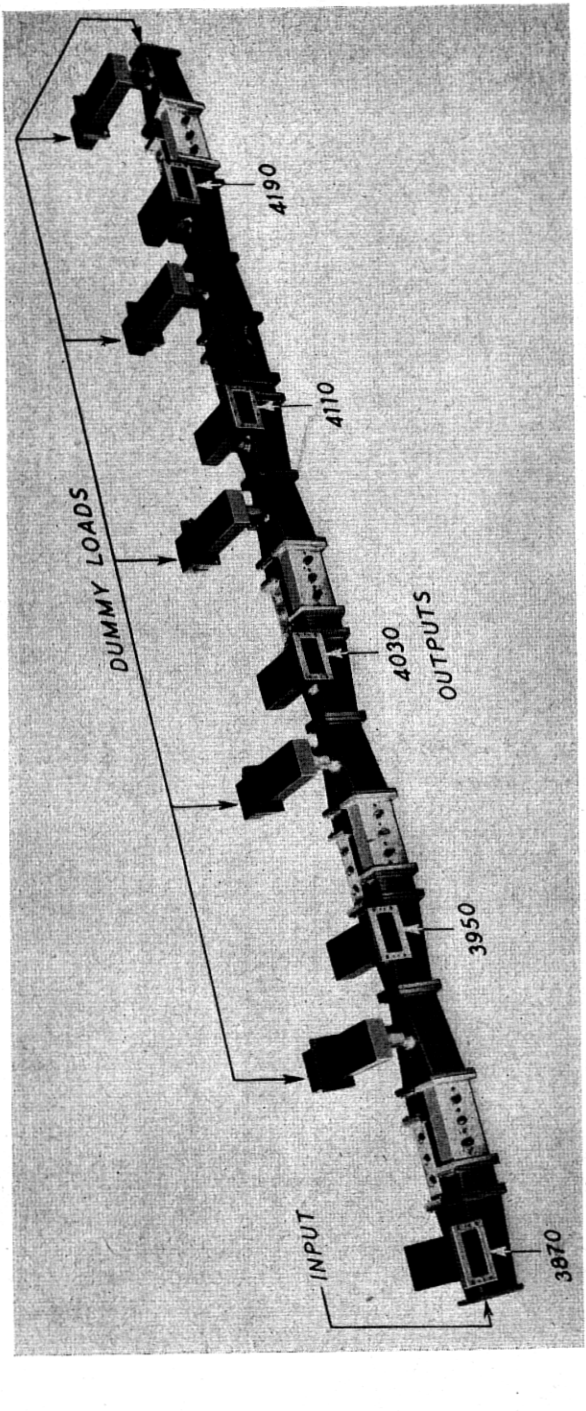


Fig. IV-7.—Photograph of a five channel branching filter.

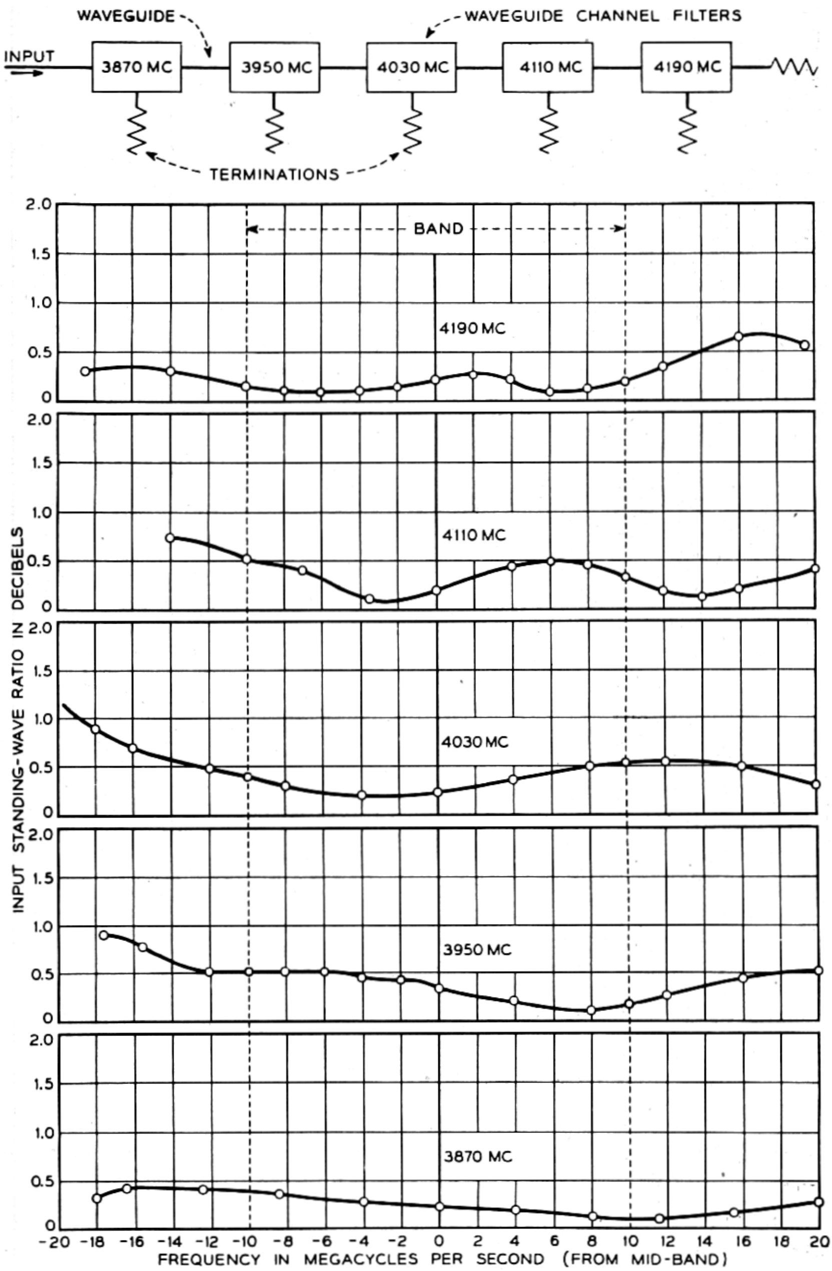


Fig. IV-8.—Input standing wave ratios for the filter of Fig. IV-7.

for C is physically parallel to those for A and B and is connected to them through a broad-band waveguide taper. Connection D is made through a relatively narrow band coaxial and probe arrangement. Figure IV-6 illustrates one of the waveguide band reflection filters. In this filter reflection occurs at three resonant rods, each tuned by an adjustable capacita-

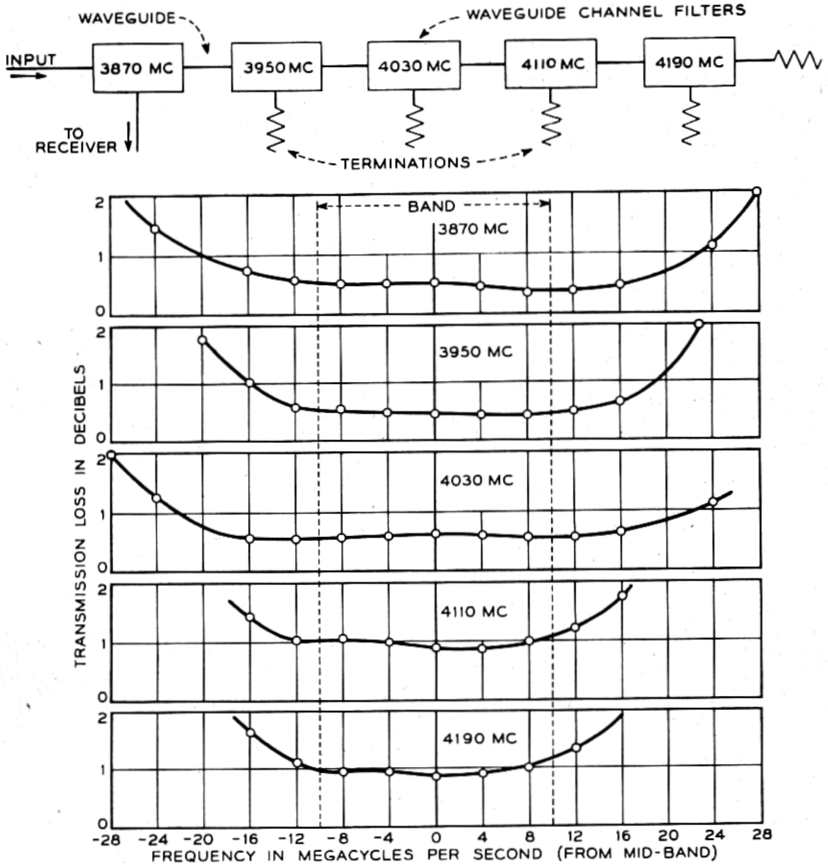


Fig. IV-9.—Transmission loss for the filter of Fig. IV-7.

tive plug at one end. These rods are placed perpendicular to the electric vector of the guide and are coupled to it by means of adjustable screws placed over them in the wide wall of the guide.

Channel dropping units for five channels in the 4,000 megacycle region were made up of these components and suitable quarter wavelengths of guide. These units were connected in sequence as shown in Fig. IV-7 and adjusted systematically. The electrical performance of the resulting five-channel branching filter was measured with a double detection measuring set and is plotted in Figs. IV-8 and IV-9.

These measured electrical characteristics serve as a check on the general theoretical ideas concerning constant resistance hybrid branching networks. They also indicate that when these ideas are embodied in the form illustrated in Fig. IV-7 the result is a branching filter which can be used in currently planned radio relay circuits.

The circuit of Fig. IV-7 provides a satisfactory channel splitting network. It does not, however, provide consistent high off-frequency discrimination between one of the channel output terminals and the other terminals of the filter. When systems requirements* are such that extra discrimination or special impedance behavior is required, this can be supplied by inserting suitably designed reflecting filters in the branch lines. These added filters will not interact reflectively with the branching filter.

V. The Repeater Amplifier†

In a microwave repeater circuit, Fig. II-1, the signal is amplified at each repeater to compensate for the transmission loss in the preceding radio path. Since we cannot build perfect amplifiers, the signal will not appear at the output terminals as a true replica of the input signal; the circuit will distort the shape of the signal and it will also add noise. Therefore, the main objectives in amplifier work have been to keep the signal distortion and the added noise within certain requirements.

To be more specific, the repeater amplifier in a relay system must be capable of supplying a maximum gain, G, as given by the equation II-1; it must have a ratio of output power capacity to noise figure which will meet the signal to noise ratio requirements of the system as given by equation II-4; since distortionless transmission is desired, it must have an amplitude characteristic as flat as possible and a phase characteristic as linear as possible over the essential range of frequencies of the signal it is desired to transmit;[24] and it must be equipped with an automatic gain regulating circuit to hold the output power constant over the expected range of input levels.

The simplest relay amplifier would be one which amplifies the signal and sends it on without a change in frequency. However, two major considerations indicated that early repeater amplifiers could not be so simple. No

* E.g., the converter may require a reflection in the input line at the image frequency, See Section V and Fig. V-4.

† Those parts of this section dealing with the general layout, the requirements, and the over-all testing of the repeater amplifier were prepared by D. H. Ring, who together with A. C. Beck did the work on this phase of the problem.

[24] Sallie Pero Mead, "Phase Distortion and Phase Distortion Correction", *B. S. T. J.*, vol. VII, No. 2, pp. 195–224, April 1928.

microwave amplifiers were known which gave promise of yielding an adequate ratio of output power capacity to noise figure, and there was considerable doubt of our ability to reduce, sufficiently, the feedback from the transmitting antenna to the receiving antenna.

These difficulties with straight-through amplification can be avoided by a repeater amplifier such as is shown schematically in Fig. V-1. The incoming signal is converted to an intermediate frequency, IF, where better amplifiers are available and where the major part of the required gain is supplied. The amplified IF signal is then converted back to the microwave frequency $f + \Delta f$, where Δf is relatively small. The difference Δf between the incoming and outgoing frequencies permits the use of circuit selectivity to counteract feedback troubles, and the radio frequency amplifier following the transmitting converter can have a relatively large noise figure.

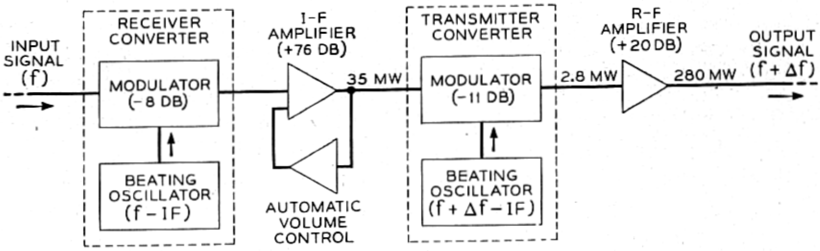


Fig. V-1.—Schematic of a repeater amplifier.

Our initial research on microwave repeaters was directed toward solving the problems associated with an amplifier of the type shown in Fig. V-1. The gain and level figures shown in this figure apply to the example of an eight-link relay system given in section II. They indicate approximate minimum objectives for the various components of the repeater amplifier to be discussed later.

Choice of I.F. Frequency—When selecting the intermediate frequency for a multichannel repeater circuit utilizing the interleaved radio frequency plan of Fig. IV-1 and the intermediate frequency type repeater amplifiers of Fig. V-1, the relative position of the various discrete frequencies and frequency bands shown in Fig. V-2 must be considered. In order to minimize the possibility of crosstalk from the image bands and interference from the beating oscillators caused by insufficient shielding, it is desirable to choose the intermediate frequency in such a way that the image bands fall midway between the active bands, and the oscillator frequencies fall midway between the image and active bands. These conditions are realized, as shown in Fig. V-2, if the intermediate frequency satisfies the relation

$$2\ \mathrm{IF} = n\Delta f + \frac{\Delta f}{2}$$

or

$$\mathrm{IF} = \frac{\Delta f}{4}(2n + 1)$$

where Δf is the frequency spacing and n is any integer greater than zero.

In general, better noise figures and circuit stability are obtained with low intermediate frequencies, while high intermediate frequencies lead to

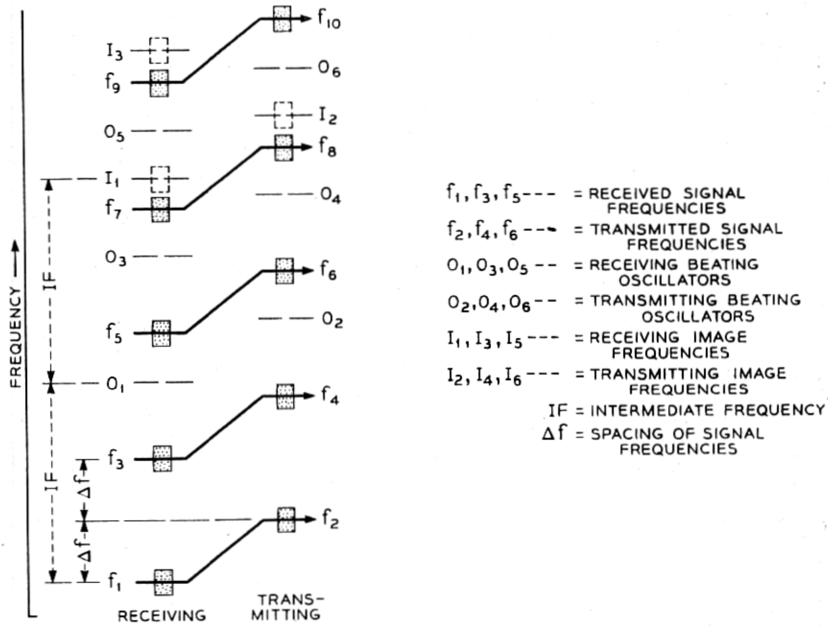


Fig. V-2.—Frequency spectrum of a multichannel microwave repeater circuit.

more symmetrical amplitude and phase characteristics. The research on the intermediate frequency components described below was conducted in the 60 to 70-megacycle range.

Frequency Stability—If all receiving and transmitting beating oscillators are independently controlled in a multichannel relay circuit of this kind, there is a possibility that small variations will add to produce large variations at the distant end of the system. This difficulty can be overcome by the frequency control system shown schematically in Fig. V-3. The transmitting and receiving beating frequencies are both derived from an oscillator operating at a frequency suitable for receiving the incoming signal. The frequency of this oscillator is controlled by a high Q cavity and a servo mechanism to 0.2 megacycles or better.[25] One portion of the output of the oscillator is used as the beat frequency in the receiving converter. A second portion is combined with the output from a crystal oscillator operating at a frequency equal to the difference, Δf, between the incoming and outgoing frequencies. In this way a beat frequency for the transmitting converter is obtained which has the same variations as that for the receiving converter except for negligibly small variations that may occur in the crystal oscillator. As a result of this method of deriving the beat frequencies the outgoing frequency always differs from the incoming frequency by an amount equal to the crystal oscillator frequency and is not influenced by variations in the high-frequency local oscillator. The result is that, except for the small

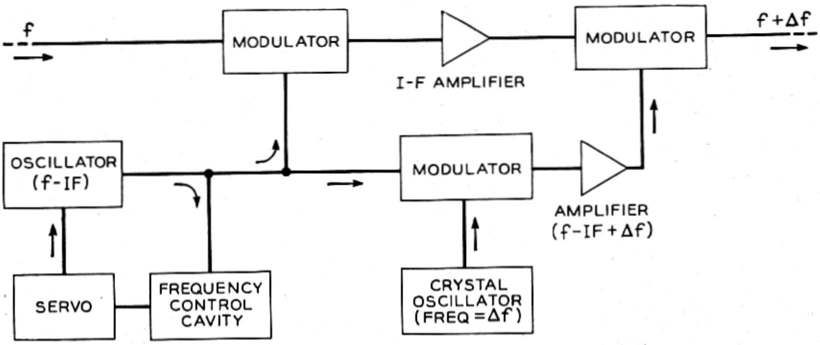


Fig. V-3.—Frequency control system for a microwave repeater.

crystal oscillator variations, all radiated frequencies in a long circuit carry only the variations in the transmitting oscillator of the originating terminal, while the intermediate frequency of each repeater may vary by an amount equal to the sum of the variation of its own local oscillator and that of the terminal transmitter frequency.

Automatic Gain Regulation—The function of the automatic gain control circuits is to hold the repeater output constant over the expected fading range. As already stated an allowance for fades 20 db down and 10 db up from free space have been made for 30-mile paths at a wavelength of about 7 centimeters. In addition to knowledge of the fading range to be compensated, it is necessary in the design of suitable circuits to know what the maximum fading rate is likely to be. Analysis of the records of a number of disturbed periods on the New York-Neshanic path indicate a maximum rate of 5 db per second.

[25] V. C. Rideout, "Automatic Frequency Control of Microwave Oscillators", *Proc. I. R. E.*, vol. 35, pp. 767–771, August 1947.

A conventional delayed automatic gain control circuit has been used in which a d.c. voltage, supplied by a peak rectifier in the output of the last IF stage, is amplified and fed back to bias several of the IF stages. Analysis of the transient response of the last repeater in response to a step function disturbance in the input to the first repeater, where the number of repeaters is of the order of 5 or more, shows that great care must be taken in shaping the frequency characteristic of the feedback circuit.

General Requirements of Components—The important factors bearing on the basic layout of the amplifier components have been discussed. Others, affecting the design of these various components will now be considered, after which a brief description of the research work on each component will be given.

Different repeater circuits will, in general, have different numbers of repeaters; also it may be necessary to feed signals into and extract signals from them at any point. Under these conditions it is impractical to specify only the over-all characteristics for a given number of repeaters. Each repeater itself should be individually good and should not depend upon any systematic compensation or equalization at any other point in the system. The repeater was designed in accordance with this philosophy and, in the interest of flexibility and ease of testing, the same line of reasoning was extended to cover the design of the components of the repeater as well.

In accordance with the above considerations major emphasis was placed on obtaining a minimum of amplitude variation and phase distortion over a 10-megacycle band for each repeater component. However, it was appreciated that even with the simplest circuits the inherent phase distortion would be excessive in long relay systems. Phase equalizers can be designed to equalize this distortion, but the difficulties of design and alignment increase with the magnitude of the distortion to be equalized. Accordingly, simple circuits were used wherever gain requirements would permit and at the same time parallel research was carried out on the problems of designing and testing appropriate phase equalizers.

While our aim was to provide a repeater 10 megacycles wide suitable for any type of modulation, it soon became apparent that it would be uneconomical to attempt to provide the extreme degree of linearity that would be required; for example, for a long relay circuit carrying an amplitude-modulated television signal. However, early tests indicated that very satisfactory transmission could be obtained with low-index frequency-modulated signals, for which reason the later stages of the work were aimed at providing an amplifier to be used for such signals. Nevertheless an attempt was made to limit the compression in each unit except the R.F. amplifier to 0.1 db at maximum rated load.

A further important requirement placed on all components was that their input and output impedances should match the corresponding impedances of the components to which they were to be connected with a minimum of reflected power over the 10-megacycle band. This requirement was imposed on the separate units to provide for flexibility in testing and to permit easy patching in of spare units in case of failure.

RECEIVING CONVERTER*

The receiving converter, together with the input to the first stage of the intermediate-frequency amplifier, occupies a unique position in the repeater amplifier in that it is located at that point in the circuit where the signal level is lowest. As a consequence, research on receiving converters has been directed toward insuring that the least possible noise be added to the signal to be amplified. An extensive background of microwave converter design information was available from the work done on converters during the war[26,27] which led to the selection of a balanced converter using a waveguide hybrid junction and 1N23-B silicon point contact rectifiers. The information already available would probably have been adequate except for the two additional requirements imposed by the repeater amplifier: first, that the standing wave ratio at the input have a low value and, second, that uniform conversion efficiency be maintained over a band of at least 10 megacycles.

A receiving converter with its associated input and output circuits is shown in Fig. V-4. Frequency conversion is accomplished in a device of this kind by virtue of the fact that when two sinusoidal voltages (in this case the signal and the beating oscillator) are applied to a non-linear impedance such as a rectifier, new frequencies given by the sums and differences of the applied frequencies and their harmonics are generated. The difference frequency may thus be selected as the desired output frequency and passed through the IF transformers to the IF amplifier. The performance of a converter is, however, influenced by the impedances encountered by some of the other frequencies generated. For this reason, the separation S between the input filter, more properly a component of the channel selecting network, but here shown as part of the converter, and the converter must be given consideration, since it determines the phase of the reflection from the filter at the image frequency (the difference between the beating

* This section prepared by C. F. Edwards who was responsible for the research on the receiving converters.

[26] "Developments of Silicon Crystal Rectifiers for Microwave Radar Receivers", J. H. Scaff and R. S. Ohl, *B. S. T. J.*, vol. 26, pp. 1–30, January 1947.

[27] Descriptions of several converters developed prior to and during the war as well as a more complete description of the present converter are given in C. F. Edwards' paper, "Microwave Converters", *Proc. I. R. E.*, vol. 35, No. 11, pp. 1181–1191, November 1947.

oscillator second harmonic and the signal) which in turn affects the converter output impedance and hence the match between the rectifiers and the IF transformers. Failure to obtain a proper match results in non-uniform transmission over the channel band.

In addition, to maintain uniform conversion efficiency over a wide band, it is necessary to control the impedance encountered by frequencies near the beating oscillator second harmonic. This is done by means of the harmonic

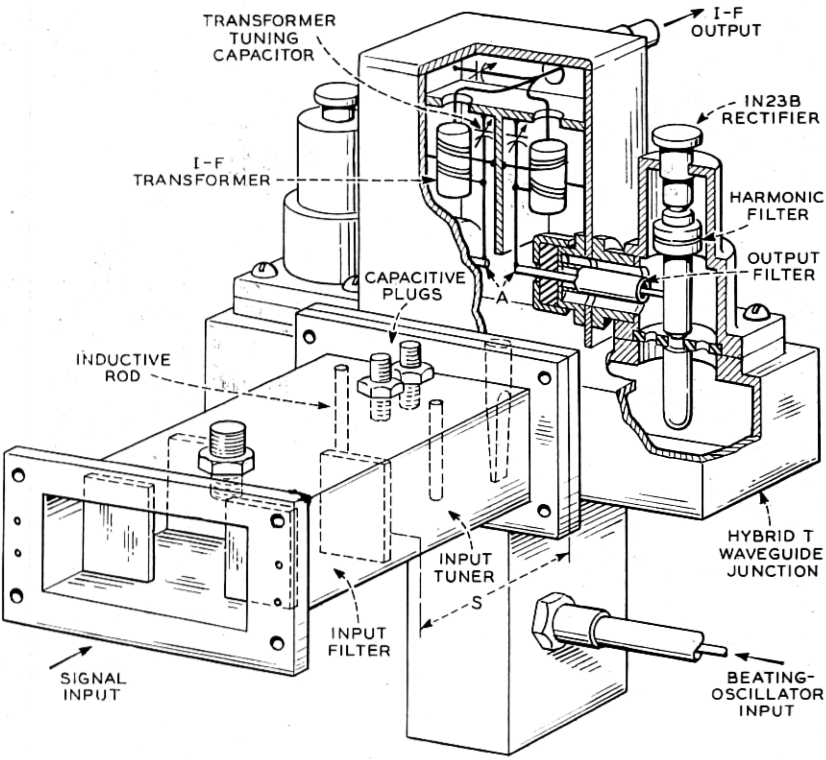


Fig. V-4.—Receiving converter with input filter.

filters shown which reflect these frequencies from a point close to the rectifier. Since the point of reflection is close, there is little opportunity for the phase of the reflection, and hence the conversion loss, to vary over the band of frequencies to be converted.

The converter in Fig. V-4 was designed to provide as good a termination as possible to the incoming waveguide over the band of channel frequencies so that a minimum of additional adjustment would be required to match the guide accurately at any particular channel frequency. This additional adjustment is provided by the input tuner shown. With it, the reflection coefficient may be reduced to zero at the desired channel band center, and when this is done the standing wave ratio at the input is less than 2 db over a 20-megacycle band.

The bandwidth of the converter is largely determined by the IF transformers shown which transform the balanced output impedance to the unbalanced 75-ohm coaxial line connecting to the IF amplifier. When transformers having a coupling coefficient of about 0.5 are used, transmission variations of less than 0.1 db over a 20-megacycle band are obtained.

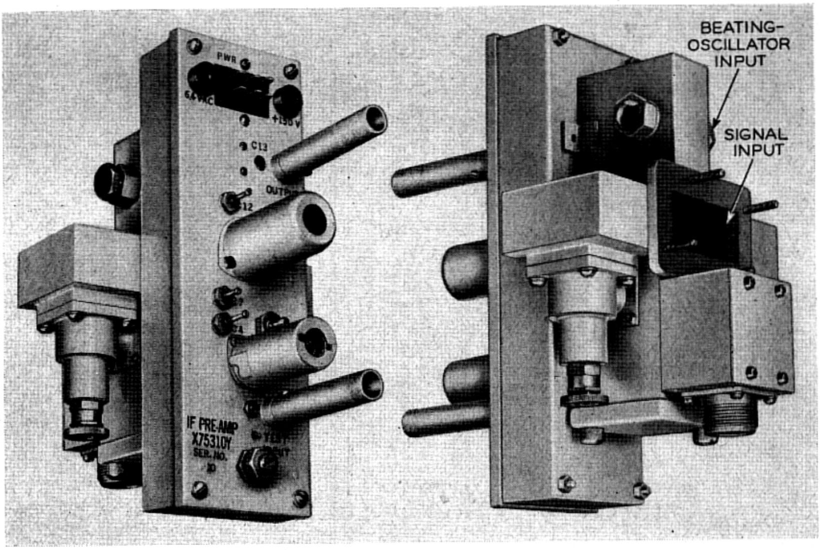


Fig. V-5.—Receiving converter with low noise IF preamplifier.

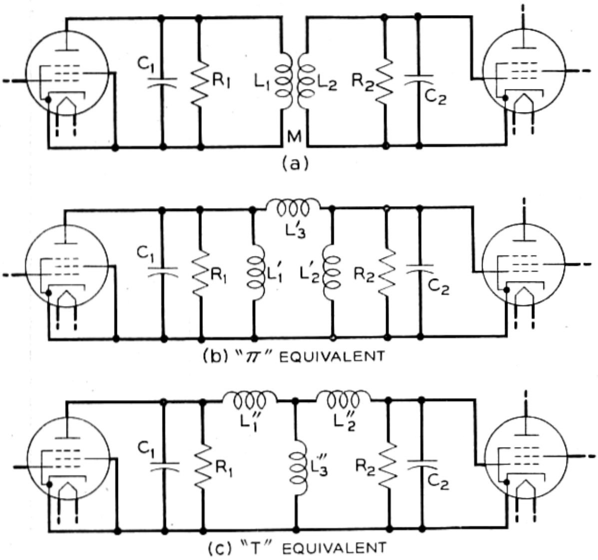
The noise figure of the repeater amplifier is determined by the conversion loss and noise figure of the converter and the noise figure of the IF amplifier.[16] The converter designed for the New York-Boston circuit has a conversion loss of about $5\frac{3}{4}$ db and its noise figure is 10 db. Thus when it is used with an IF amplifier having a noise figure of 7 db, a figure of 14 db is obtained for the over-all noise figure of the repeater amplifier. Figure V-5 shows this converter with a low noise preamplifier attached.*

I.F. Amplifier†

Most of the gain of the IF amplifier is obtained with stages using double tuned, symmetrically loaded (or 'matched") interstage transformers, de-

[16] Loc. cit.
* Developed by H. C. Foreman and B. C. Bellows, Jr.
† This section prepared by Karl G. Jansky who, together with V. C. Rideout, did the work on IF amplifiers.

signed in accordance with the formulas given by a former member of this laboratory.[28] In Fig. V-6, "a" shows a schematic diagram of such a transformer and "b" shows the equivalent π network generally used. The equivalent T network shown at "c" has also been used. In Fig. V-7 "a" shows the theoretical band-pass characteristic for this type of transformer with a coupling coefficient of 0.5 which was the value used in most stages. The circles indicate points measured on a typical transformer. This matched network is relatively insensitive to small changes in capacitance so that wherever it is used it is possible to change tubes without having to realign the amplifier. The delay distortion for this type of network is, as shown by



$$\omega_0 = \frac{1}{\sqrt{L_1C_1}} = \frac{1}{\sqrt{L_2C_2}}$$

$$\frac{\sqrt{1-K^2}}{K} = \omega_0 R_1 C_1 = \omega_0 R_2 C_2$$

$$M = K\sqrt{L_1L_2}$$

$$L_1' = \frac{\sqrt{L_1L_2}(1-K^2)}{\sqrt{\frac{L_2}{L_1}}-K}$$

$$L_2' = \frac{\sqrt{L_1L_2}(1-K^2)}{\sqrt{\frac{L_1}{L_2}}-K}$$

$$L_3' = \frac{\sqrt{L_1L_2}(1-K^2)}{K}$$

$$L_1'' = L_1 - K\sqrt{L_1L_2}$$

$$L_2'' = L_2 - K\sqrt{L_1L_2}$$

$$L_3'' = K\sqrt{L_1L_2}$$

Fig. V-6.—The double tuned IF transformer.

"b" in this figure, also relatively small. The gain per stage for a coupling coefficient of 0.5 is approximately 6 db for 6AK5 vacuum tubes and about 12 db for the recently developed WE 404-A tubes. As shown by W. J. Albersheim[28] it is possible to design circuits which will give more gain than the "matched" transformer, but only at the cost of increased sensitivity to capacitance changes. For convenience the IF amplifier is usually divided into a preamplifier closely associated with the receiving converter and a main IF amplifier.

Pre-amplifier—At the time work was begun, noise figures of the order of

[28] V. C. Rideout, "Design of Parallel Tuned Transformers", *B. S. T. J.*, vol. 27, pp. 96–108, January 1948.

12 to 14 db were obtainable for 65 mc amplifiers of the required bandwidth with 6AK5 tubes and matched input transformers. This was much worse than was desired. By using a 6J4 close spaced grounded grid triode in the input stage, much lower noise figures were obtained, but the gain with matched input and output circuits was so low (approximately 3 db for a coupling coefficient of 0.5) that the noise from the following stage contributed considerably to the overall noise figure.

By removing the loading resistance on the output side of the first interstage transformer and reducing the coupling coefficient to 0.3, the gain was

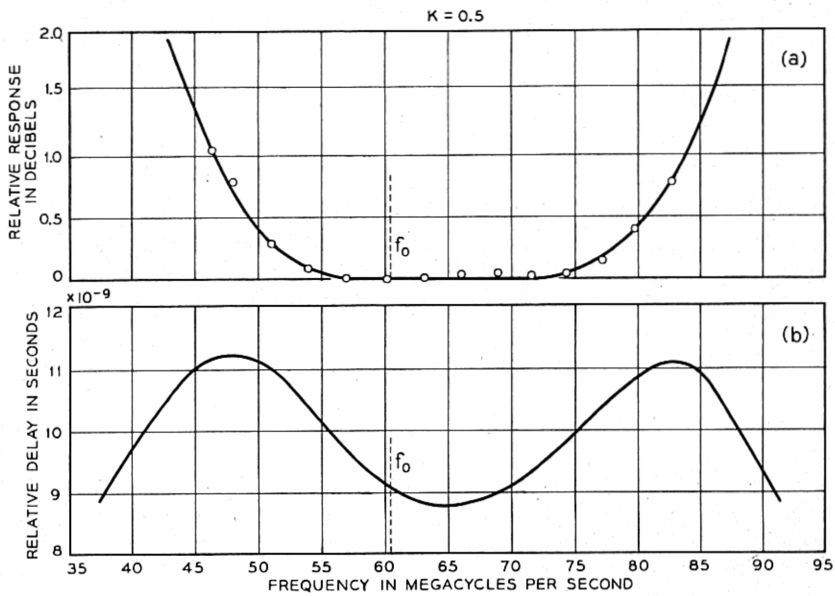


Fig. V-7.—Band pass and delay characteristics of the IF transformer.

raised sufficiently to give an overall noise figure of the order of 7 db. Figure V-5 shows the preamplifier developed for the New York-Boston circuit. This amplifier employs a 6J4 and a WE 404-A and provides a gain of 23 db.

The 6J4 tube, when used in a grounded grid circuit, has an input impedance of approximately 85 ohms which is close to the desired 75 ohm impedance. When a good match is required, it is necessary to use an input transformer, but it should be noted that an improvement in the noise figure can be obtained by deliberately producing a mismatch in the right direction at the input. Noise figures as low as 4 db have been obtained in this manner with recent experimental tubes.

Figure V-8 is a schematic diagram of an amplifier with a very low noise figure consisting of two of these experimental grounded grid tubes in tandem. The circuits were designed so that there is a mismatch at the input of each tube. The overall noise figure is 4.0 db with the amplifier connected between a 75-ohm generator and a 75-ohm load; the gain is $17\frac{3}{4}$ db; and the bandwidth at the 0.1 db down points is about 13.5 megacycles.

Main IF Amplifier—In order to obtain the required output power over the desired bandwidth it was necessary, at the beginning of this work, to use the WE 367-A tube in the output stage. Since the gain of a stage using this tube is very low, it must be driven by a tube similar to the 6AG7 and to prevent compression in this latter tube a special high gain, triple-tuned interstage transformer was designed, with a relatively low coupling coefficient.

Methods of paralleling tubes to get more power output or the same power output with a much wider bandwidth have been worked out, but recent

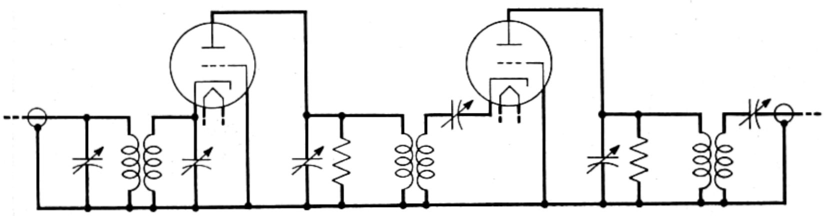


Fig. V-8.—Schematic of a "low noise figure" preamplifier.

developments in power output tubes and in transmitting converters have made them unnecessary, except for applications requiring very wide bandwidths.

Automatic gain control can be applied to the IF amplifiers by controlling the grid bias of the various stages. However, it is not advisable to apply the gain control bias to either of those stages which are preceded by the special high-gain transformers. In these stages the slight changes in input impedance of the tubes caused by variations in the grid bias would be sufficient to alter significantly the band-pass characteristics of the transformer.

Figure V-9 shows an amplifier with about 55 db gain developed for the New York-Boston circuit.* Automatic gain control bias is applied to the grids of the first three stages which employ wide band (K = 0.7) matched interstage transformers.

It will be noted that the first interstage transformer of the preamplifier and the last interstage transformer of the main IF amplifier require low coupling coefficients to obtain the gain desired at these points. For this

* Developed by A. L. Hopper and B. C. Bellows, Jr.

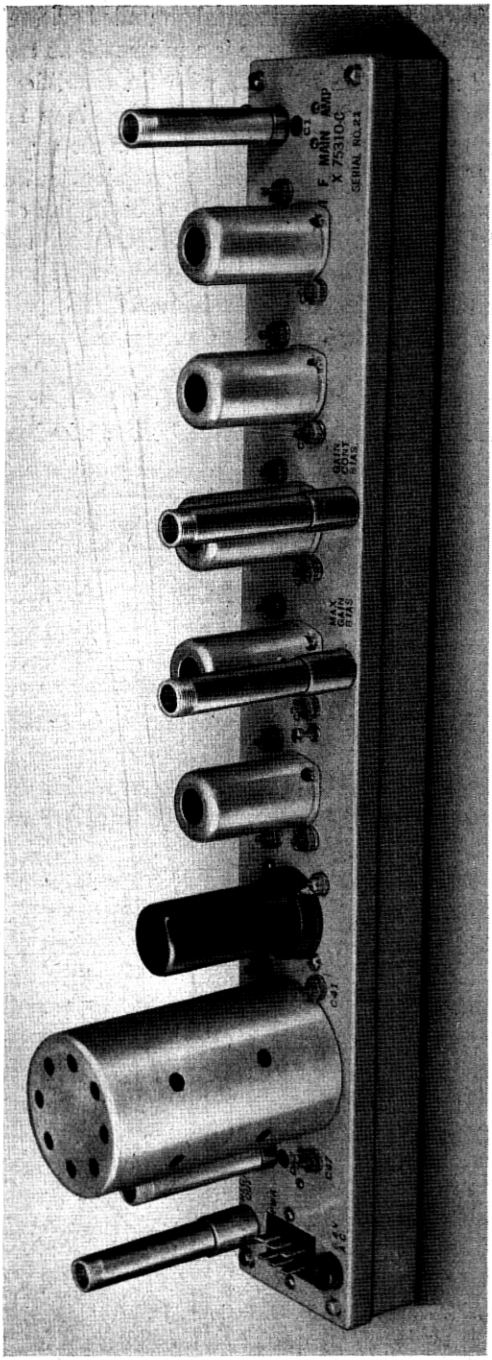


Fig. V-9.—The main If amplifier.

reason, these two transformers largely determine the band pass and delay distortion of the whole amplifier. Typical overall characteristics for a complete IF amplifier are shown in Fig. V-10.

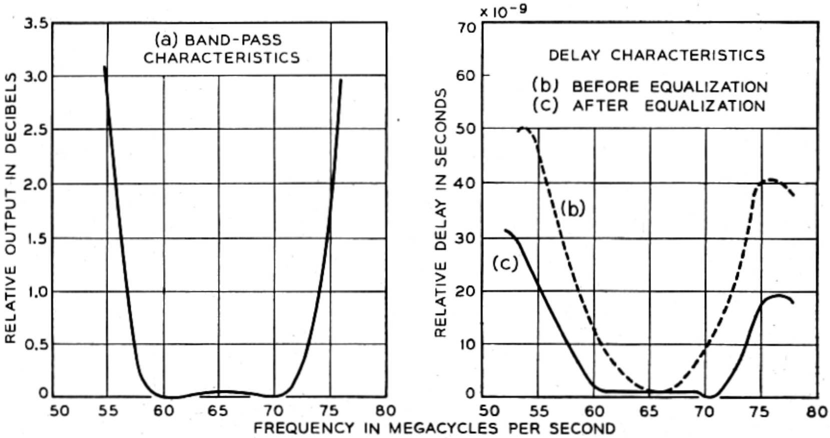


Fig. V-10.—Band pass and delay characteristics of complete IF amplifier.

TRANSMITTING CONVERTER, OR MODULATOR*

The transmitting converter problem differed from the first converter problem in that high output power was the major consideration, rather than low conversion loss and noise figure.

There will normally be three frequencies present in the output of a converter. These are the desired output sideband frequency, f', the beating oscillator frequency, $f' + \text{IF}$ or $f' - \text{IF}$, and the unwanted sideband frequency, $f' + 2\ \text{IF}$ or $f' - 2\ \text{IF}$. The strongest of these is the beating oscillator frequency, but fortunately this can be suppressed by 20 to 30 db by balance in a balanced converter. Investigation showed that the input impedance at the IF terminals of the modulator was affected by the load impedance of the modulator at both the wanted and the unwanted sideband frequencies. The load impedance consists of the input impedance of the RF amplifier as seen through the length of transmission line which connects the two components. At the wanted sideband frequency, the RF amplifier presents a good termination and the load impedance is independent of the length of the connecting line. At the unwanted sideband frequency, however, the RF amplifier presents a short circuit and hence the load impedance is a function of the length of the connecting line. A waveguide filter was incorporated in the output circuit of the modulator so as to reflect the un-

* This section prepared by W. W. Mumford who did the work on the transmitting converters.

wanted sideband back into the modulator in proper phase, thereby making the length of the line going to the RF amplifier much less critical. The proper phase of reflection was obtained by adjusting the length of the waveguide between the modulator and the reflecting filter. The IF impedance of the crystals could thus be varied so as to obtain an impedance match between the IF amplifier and the crystals.

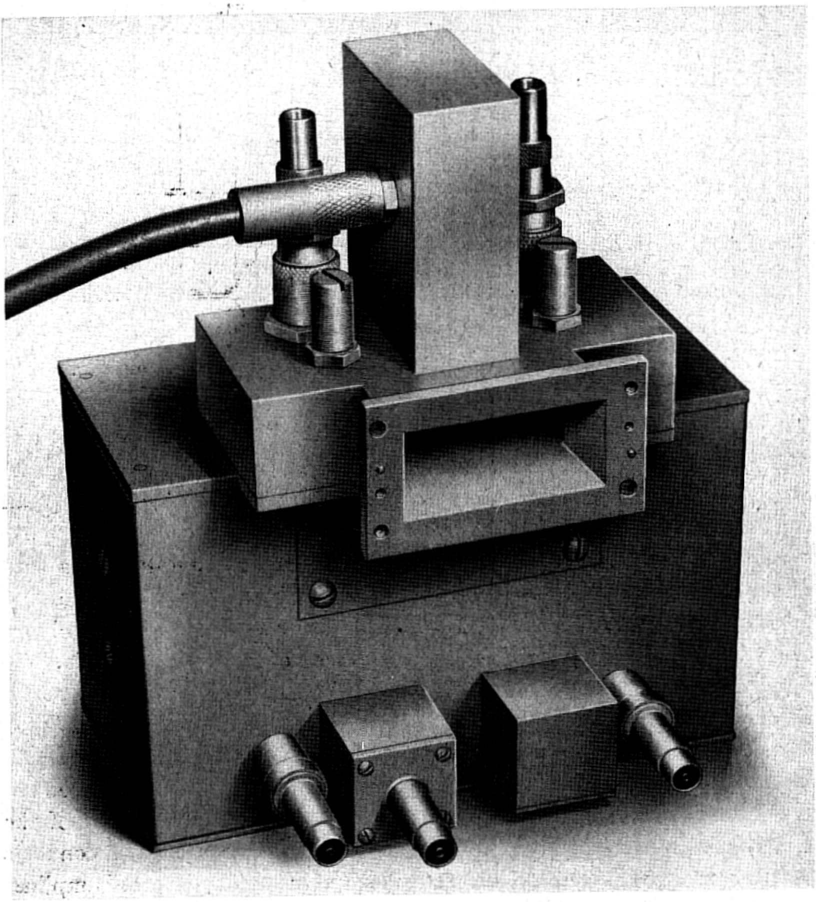


Fig. V-11.—Experimental model of transmitting converter.

An investigation of the power capacity of various types of standard and special silicon crystals indicated that one similar to the 1N28 crystal was the best available for this service, and life tests were made to check the stability of the crystal under high level conditions. Special IF input and crystal-to-waveguide matching circuits were developed to meet the stringent match requirements. The research on these major factors and many subsidiary details resulted in a converter which had an output of 6 milliwatts with less than 0.1 db compression, good input and output match, a flat amplitude response over more than 10 megacycles, and a conversion loss of about 11 db.

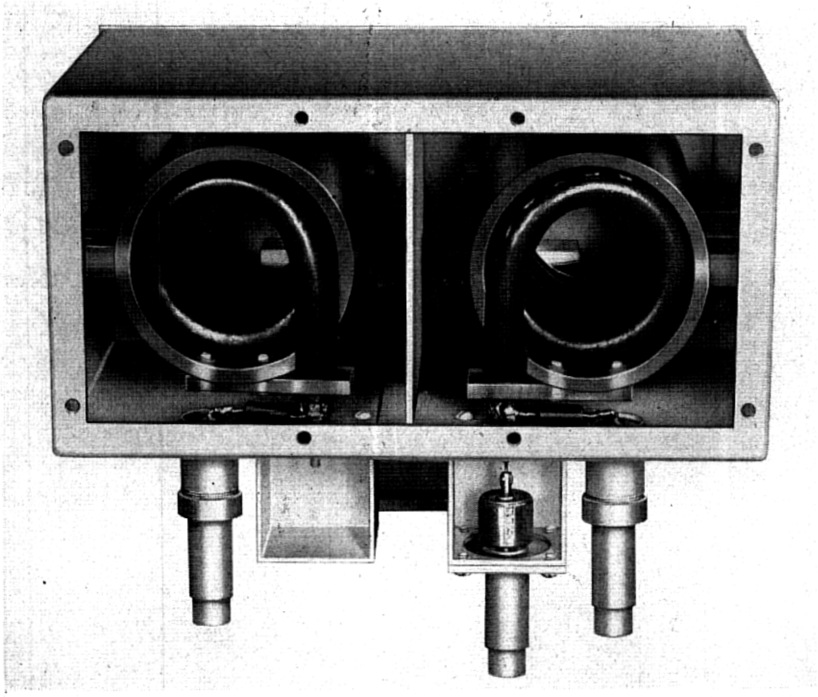


Fig. V-12.—Unbalanced to balanced coaxial transformer for feeding the 65 mc signal to the transmitting converter.

This converter, or transmitting modulator, had the 1N28 crystals mounted in the conjugate branches of a waveguide hybrid junction, as shown in Fig. V-11. Adjustable coaxial sleeves surrounded the crystal cartridges in order to effect an impedance match to the waveguide, and tuning studs were provided for trimming adjustments. The beating oscillator was injected into the hybrid junction through a broad-band coaxial-to-waveguide transducer in the upper branch of the hybrid junction, and the sidebands appeared in the conjugate waveguide branch. The 65-megacycle signal was fed onto the crystals in push-pull through the unbalanced to balanced coaxial transformer shown in Fig. V-12. Blocking condensers enabled the crystal currents to be monitored separately and RF filters kept the 4000-megacycle

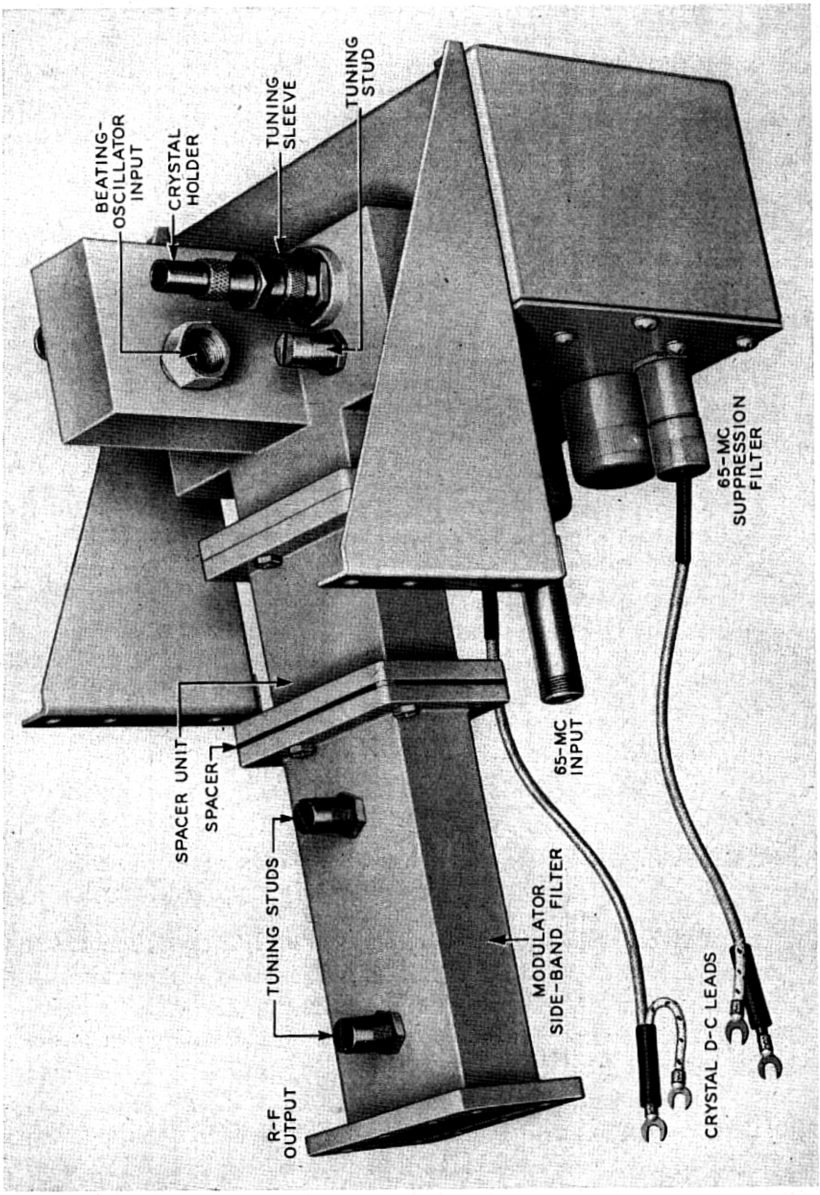


Fig. V-13.—Transmitting converter complete with output filter.

energy from entering the 65-megacycle transformer compartment. The converter developed for the New York-Boston circuit, complete with sideband filter, spacer, mounting brackets and d-c. leads, is shown in Fig. V-13.*

R. F. AMPLIFIER**

Prior to the war, a considerable amount of research had been applied in the Bell Telephone Laboratories to electron tubes operating on the velocity modulation principle, with the expectation that such tubes would find applications in radio-relay systems. Although this work was interrupted by the war, enough information had been obtained to make it apparent, upon resumption of the radio relay work, that such tubes were the only ones then known which showed promise of meeting the stated requirements.

Velocity modulation tubes have been described by several authors,[29, 30, 31] and the theory of their operation has been discussed adequately in several places.[32, 33, 34, 35] However, a review in 1939 of the structures then known had led to the decision that a new type of construction would be necessary to obtain a satisfactory amplifier tube for radio relay purposes. To keep the tube voltages within reasonable limits it is desirable to make the input and output gaps as small as possible. Resonant circuits external to the evacuated envelope are desirable to enable coverage of as large a frequency range as possible with a single tube, and to facilitate addition of broad-banding circuits. Grids on the input and output gaps are undesirable because of the large interception of current and the difficulty and expense of construction.

With the above considerations in mind, decision was then made to explore the possibilities of focussed beams, and of gaps comprised of copper discs sealed through a cylindrical glass vacuum envelope. The latter technique had been developed at the Bell Telephone Laboratories in connection with

* Developed by H. C. Foreman and W. W. Halbrook.

** This section prepared by A. G. Fox and A. E. Bowen. Messrs. Fox and Bowen, in collaboration with A. L. Samuel, A. E. Anderson, and J. W. Clark of these Laboratories, did the major part of the research which resulted in this amplifier.

[29] Varian, R. H. and Varian, S. F., "A High Frequency Oscillator and Amplifier", *Jour. Applied Physics*, vol. 10, No. 5, pp. 321–327, May 1939.

[30] Hahn, W. C. and Metcalf, G. F., "Velocity Modulated Tubes", *Proc. I. R. E.*, vol. 27, No. 2, pp. 106–116, Feb. 1939.

[31] Harrison, A. E., "Klystron Tubes", McGraw-Hill Book Co., New York, 1947. (Book)

[32] Hansen, W. W. and Richtmyer, R. D., "On Resonators Suitable for Klystron Oscillators", *Jour. Applied Physics*, vol. 10, No. 3, pp. 189–199, March 1939.

[33] Hahn, W. C., "Small Signal Theory of Velocity Modulated Electron Beams", *G. E. Review*, vol. 42, No. 6, pp. 258–270, June 1939.

[34] Hahn, W. C., "Wave Energy and Transconductance of Velocity-Modulated Electron Beams", *G. E. Review*, vol. 42, No. 11, pp. 497–502, Nov. 1939.

[35] Ramo, S., "The Electronic-Wave Theory of Velocity Modulation Tubes", *Proc. I. R. E.*, vol. 27, No. 12, pp. 757–763, Dec. 1939.

the development of water cooled tubes, and showed promise of providing a very convenient means for construction of resonant circuits in which a part of the circuit was external to the vacuum envelope and part was internal.

Numerous forms of focussed beam-disc seal velocity modulation tubes were constructed for examination of the various factors affecting the performance of such tubes. Sizes and shapes of gaps, length of drift space, voltage and beam current, and other factors were adjusted to optimize the performance. Both magnetic and electrostatic focussing of the beam were explored. Triple gap tubes, from which gains of more than 30 db were obtained, were experimented with.

The end result of this work was the development of a four-stage amplifier employing velocity modulation tubes of the disc-seal type (Fig. V-14) designed especially for this application.* The electron gun is at the left, and the collector is at the right. These tubes employ external circuits in the form of resonant cavities assembled around the tube. The electron

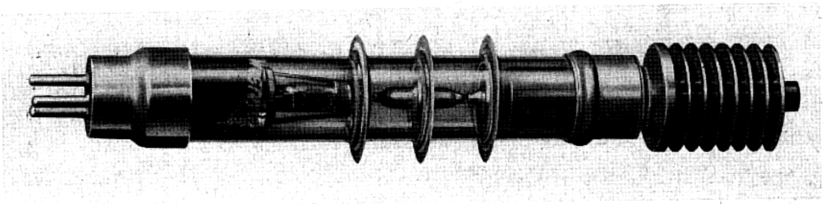


Fig. V-14.—Disc seal velocity modulation tube.

beams are accelerated by 1500 volts and focussed magnetically by means of small alnico permanent magnets placed around the cavity structure as shown in Fig. V-15. This shows a single stage of coaxial-coupled amplifier with half of each cavity removed to show internal details. The cavities are tunable over approximately a 250-megacycle range by means of metal screw plugs located around their periphery. Two different sets of cavities are sufficient to cover the entire frequency range of the radio repeater.

A single amplifier stage of the type shown in Fig. V-15 will exhibit a single-tuned type of gain characteristic which is only 5 megacycles wide at the 3 db points when operated under matched input and output impedance conditions. It was therefore necessary to widen the band by using double-tuned circuits throughout. In fact such a design would be difficult to avoid because each tube has its own resonant cavities. By running a short length of waveguide or coaxial transmission line from the output cavity of one stage to the input cavity of the next, a double-tuned structure results automatically. The coupling from cavity to transmission line is made by means

* These tubes were developed by A. L. Samuel and J. W. Clark.

of a window in the side of the cavity when a waveguide line is used, and by means of a wire loop projecting within the cavity when a coaxial line is used. Amplifiers employing both types of coupling were designed and tested. The coaxial coupling type with links approximately one-half wavelength long

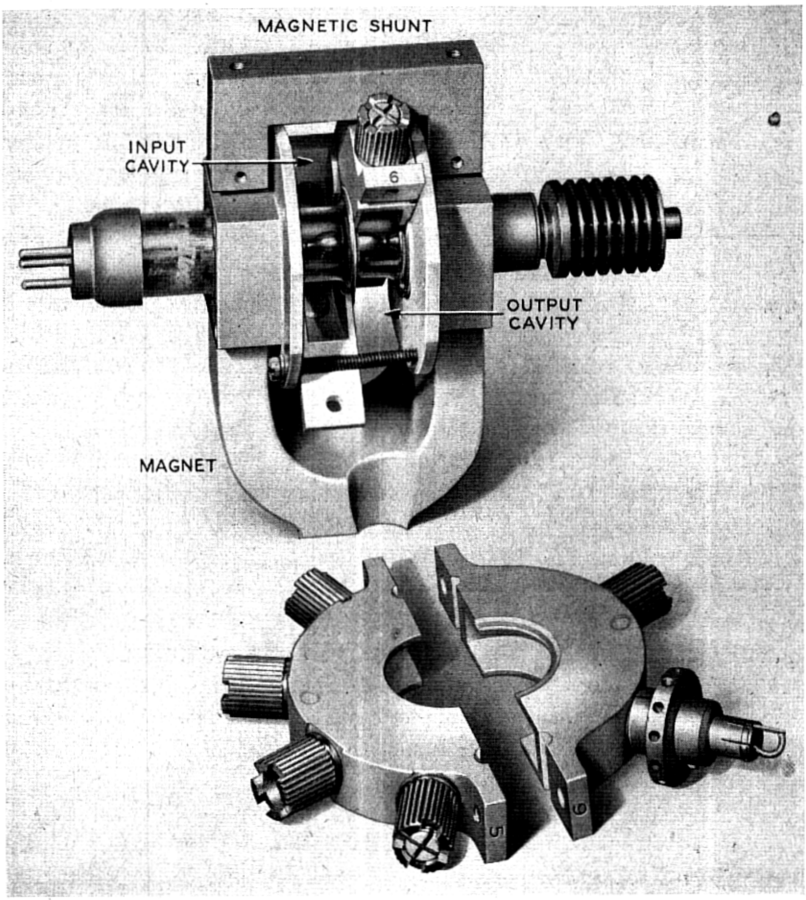


Fig. V-15.—A single stage of a coaxial-coupled amplifier using the velocity modulated tube of Fig. V-14.

proved the easier to adjust. Variation of the coupling from zero to a maximum is accomplished by rotating the plane of the coupling loop through 90°. Each cavity is provided with a small plate of resistance film projecting into the field in such a way that rotation of the plate about an axis lying in the plane of the plate will vary the resistive loading present in the cavity from zero to a maximum. Finally, the input and output cavities of the first and last stages respectively are coupled to separate tuned cavities in order to provide double-tuned terminals for the amplifier. This results in the overall structure shown schematically in Fig. V-16.

The required bandwidth is obtained by a process of stagger-coupling the circuits so that the individual responses are as indicated schematically at the top of Fig. V-16. Because there are available continuously variable adjustments of tuning, loading, and coupling for each circuit, it is possible to obtain a very smooth and symmetrical overall response characteristic as shown in Fig. V-17. The corresponding measured delay distortion characteristic is shown in Fig. V-18. For this type of tuning the output power is about .7 watt at $\frac{1}{2}$ db of compression. More power is, however, obtainable if more compression is tolerable; and when used in the repeater for the transmission of FM signals, the amplifier is driven to an output of 1 watt.

Since the circuits are of fairly high Q, the frequency characteritics of the amplifier are markedly affected by changes in temperature of the cavities. In order to minimize such detuning effects, the amplifier has been placed in a temperature controlled compartment. Since about 45 watts are dissipated at the collector of each tube by the high-voltage electron beam, the collectors must be cooled by a forced air blast. In order to keep the cooling system separate from the temperature control system, the collector ends of the tubes project through a wall of the temperature controlled compartment into an external air duct. Figure V-19 shows a complete r.f. amplifier with the cover of the temperature control box removed. This is the amplifier developed for the New York-Boston circuit.* The electron-gun ends of the tubes face the reader. Short lengths of flexible coaxial cable couple the input and output of the amplifier to the associated equipment.

Testing of Components and Repeater Amplifier—Many special measuring and testing techniques had to be devised for the research work on each component. In addition, standard production tests had to be worked out and measuring equipment constructed.

Impedances were measured at RF with standing wave detectors, and at IF with three fixed taps on a coaxial line. It was found that input and output SW ratios could be held to 1.7 db or less over the 10 megacycle band. Both point by point and swept oscillator methods were used at RF and IF for measuring amplitude characteristics. Particularly as a result of the development of swept oscillator measuring techniques it was found practicable to adjust each unit to ± 0.1 db amplitude variation over the 10 megacycle band. Noise figures[16] of IF equipment were measured by the noise diode method.[36]

* Developed by F. E. Radcliffe and R. C. Carlton.
[16] Loc. cit.
[36] H. Johnson, "A Coaxial Line Diode Noise Source for U.H.F.", *R. C. A. Review*, vol. VIII, No. 1, pp. 169–185, March 1947.

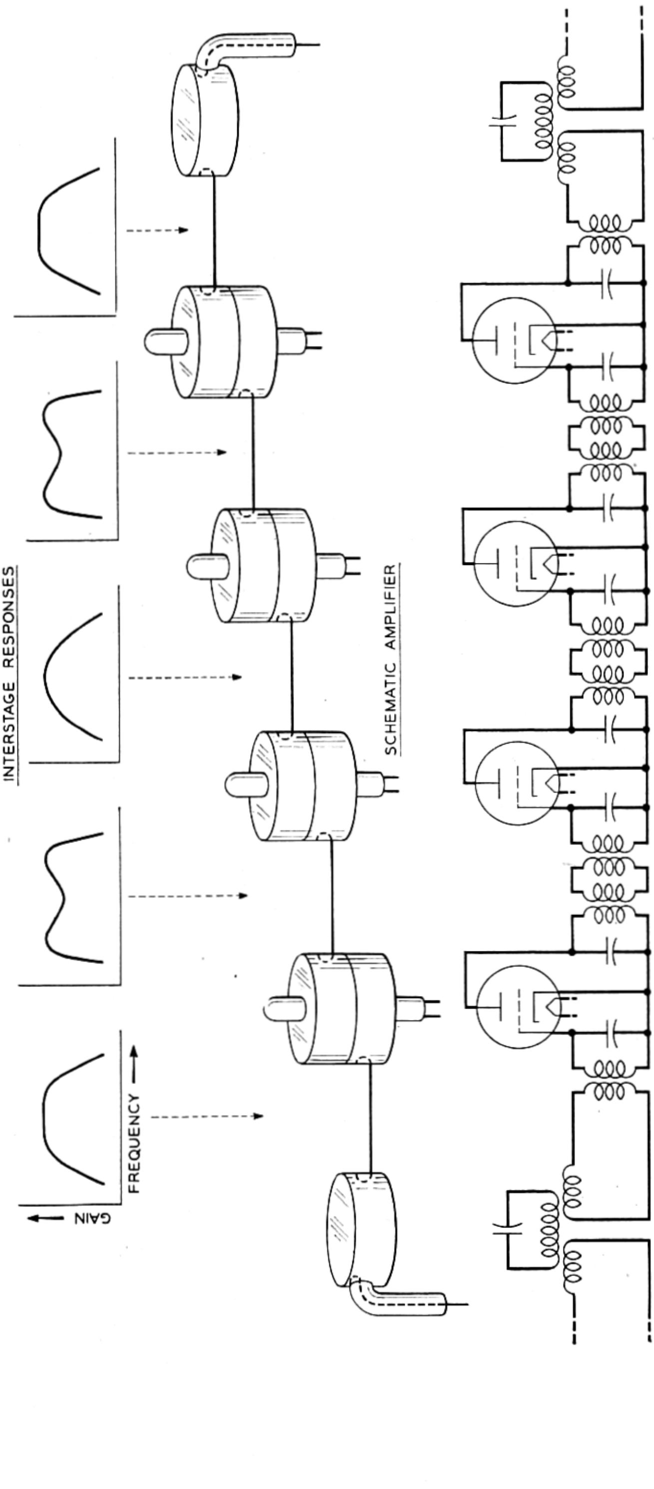


Fig. V-16.—Schematic diagram of four stage coaxial coupled amplifier using velocity modulated tubes.

A method[37] of measuring the phase and group delay characteristics of the repeater was worked out. This is a rather difficult measurement to make because the tolerable phase and delay distortion are very small due to the wide band of the system. Relative group delay through the band was measured with an accuracy of about ±0.001 microsecond. This corresponds to an accuracy in relative phase shift of about ±0.35°. The measured distortion agreed reasonably well with the distortion calculated from the constants of the various circuits. These measurements and calculations showed that while the characteristics of an 8-link repeater

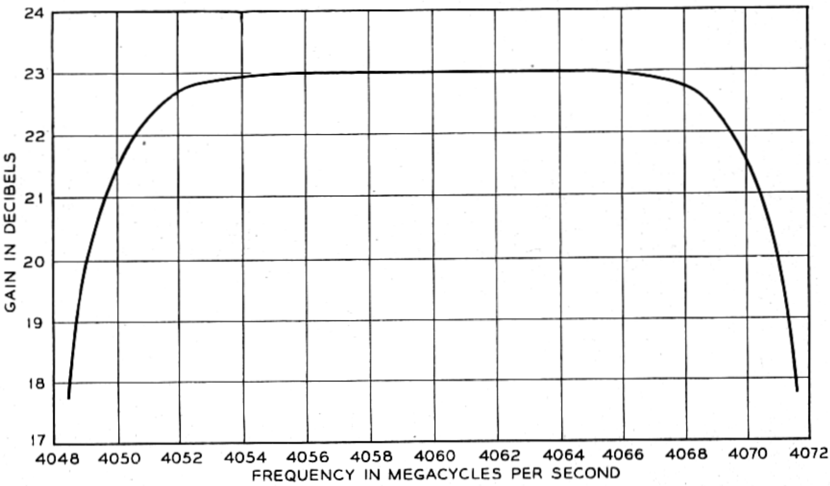


Fig. V-17.—Frequency response of RF amplifier.

circuit will give acceptable television pictures, phase equalization will provide a definite improvement. Longer circuits will therefore require phase equalization. Equalizers were devised for both the IF and RF circuits which made it possible to equalize the group delay of each component of the repeater amplifier to ±0.001 microsecond over the 10-mc band.

An experimental repeater amplifier was set up so that the output could be fed through an attenuator to the input. It was then possible to break the loop and make overall tests of delay, amplitude linearity, and transient response with IF measuring equipment. Transient response was measured by including a 2000-foot waveguide line between the input and output terminals and applying the circulated pulse testing technique.[38] This

[37] D. H. Ring, "The Measurement of Delay Distortion in Microwave Repeaters", elsewhere in this issue of the *B. S. T. J.*

[38] A. C. Beck and D. H. Ring, "Testing Repeaters with Circulated Pulses", *Proc. I. R. E.*, vol. 35, No. 11, pp. 1226–1230, November 1947.

testing method permits a study of the shapes of rectangular pulses which have passed through a repeater amplifier many times, and thus simulates transmission through a relay system with many repeaters. Figure V-20 shows an example of the results obtained from such tests made on the IF amplifier alone. The top row shows a 1.25-microsecond test pulse after 1,

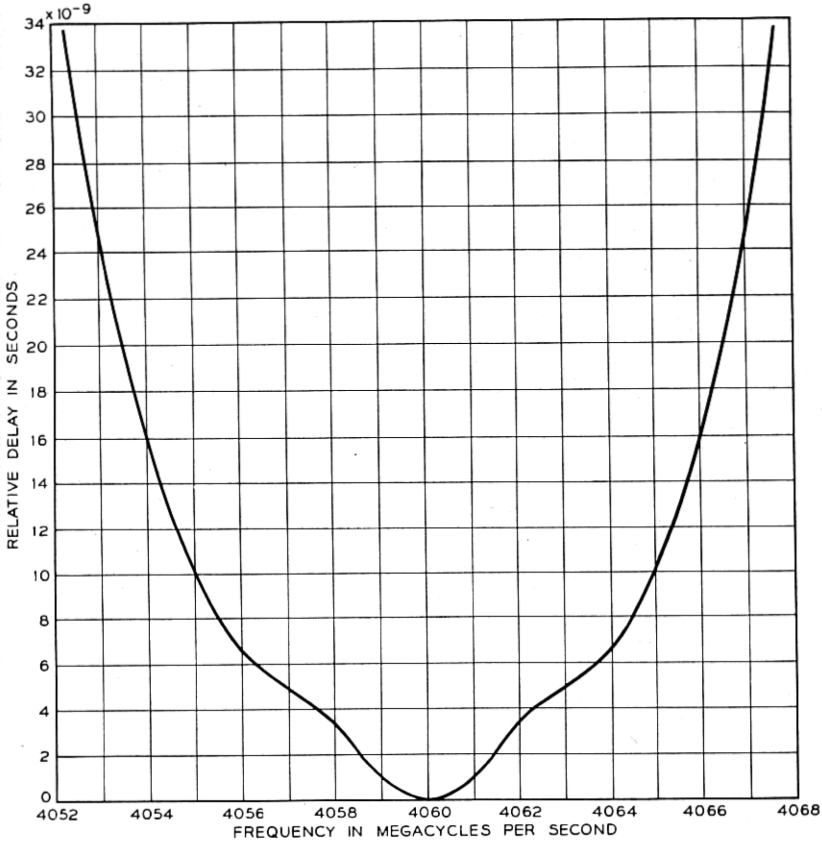


Fig. V-18.—Delay characteristic of RF amplifier.

10, and 30 trips through the IF amplifier. Part of the distortion appearing after one trip through the amplifier was in the viewing oscilloscope. It was difficult to detect distortion from a single trip through the amplifier, but Fig. V-20 demonstrates how it increases with successive trips. The second row of pulse pictures shows the improvement achieved by adding phase equalizers to the circuit. It was particularly noticeable in this test that without equalizers the details in the shape of the pulse after 10 or 30 trips were very sensitive to the exact value of the intermediate frequency.

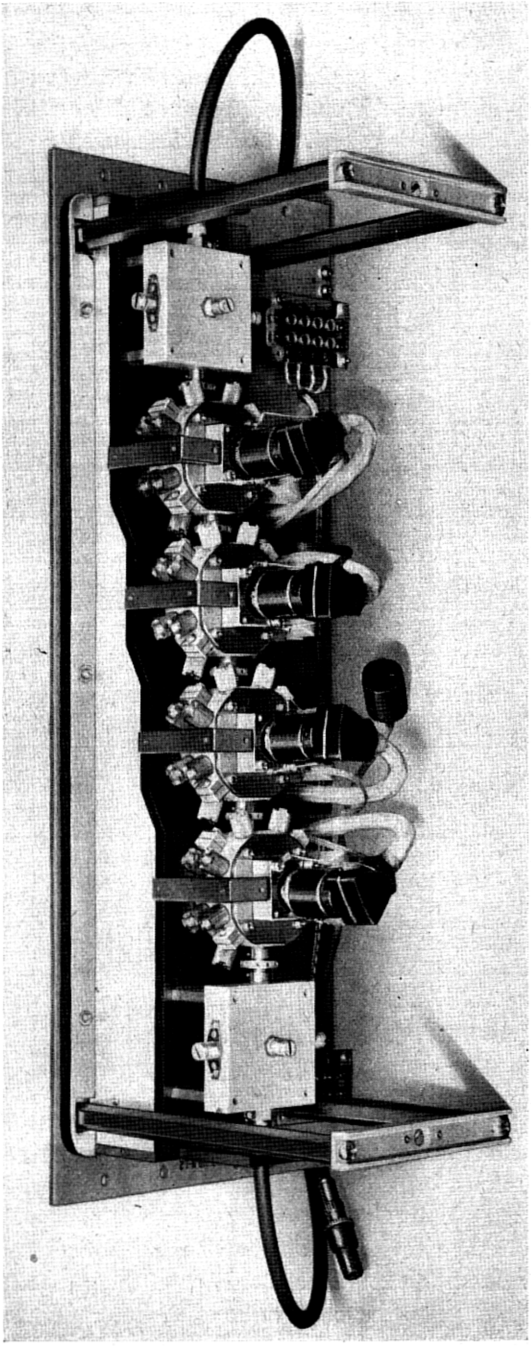


Fig. V-19.—The complete RF amplifier with cover removed.

When the equalizers were added, the pulse shapes shown in the second row were not greatly changed for variations of ±3 megacycles or more in the intermediate frequency.

In concluding this section it should be noted that when the various components were connected together to form an IF-type repeater amplifier, it was found that the amplitude and phase characteristics added as expected and resulted in a satisfactory amplifier with only a few millimicroseconds variation in delay and only a few tenths of a db variation in amplitude over a 10-megacycle band. Nevertheless, the equipment is very complicated.

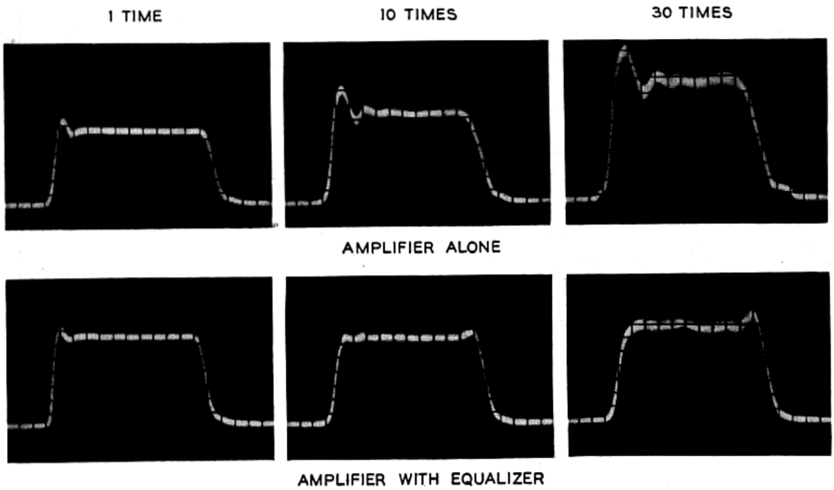


Fig. V-20.—Results of circulated pulse tests on IF amplifier.

A straight-through radio frequency amplifier repeater is still to be desired and, no doubt, further research will eventually produce such an amplifier.

VI. The Complete Repeater*

In the preceding section it has been pointed out how the various components that make up the repeater amplifier were added together without the introduction of additional distortion. The antennas, filters and amplifiers which go to make up one of the complete repeaters shown in Fig. II-1 were designed with the same ease of interconnection in mind. However, as will be discussed below, the length of the waveguide lines used for these connections has an important bearing on the distortion introduced.

The large amount of equipment in the repeater amplifier makes it desirable, from the maintenance standpoint, to locate the amplifier near the ground and to provide towers for the antenna where antenna elevation

* Prepared by D. H. Ring.

is necessary. This means relatively long transmission lines between the antenna and the rest of the repeater. If exact termination of the line by the antenna impedance or by the filter input impedance were possible no distortion would be produced, but in general there will be a slight mismatch and a corresponding reflection of energy at each of these junctions which

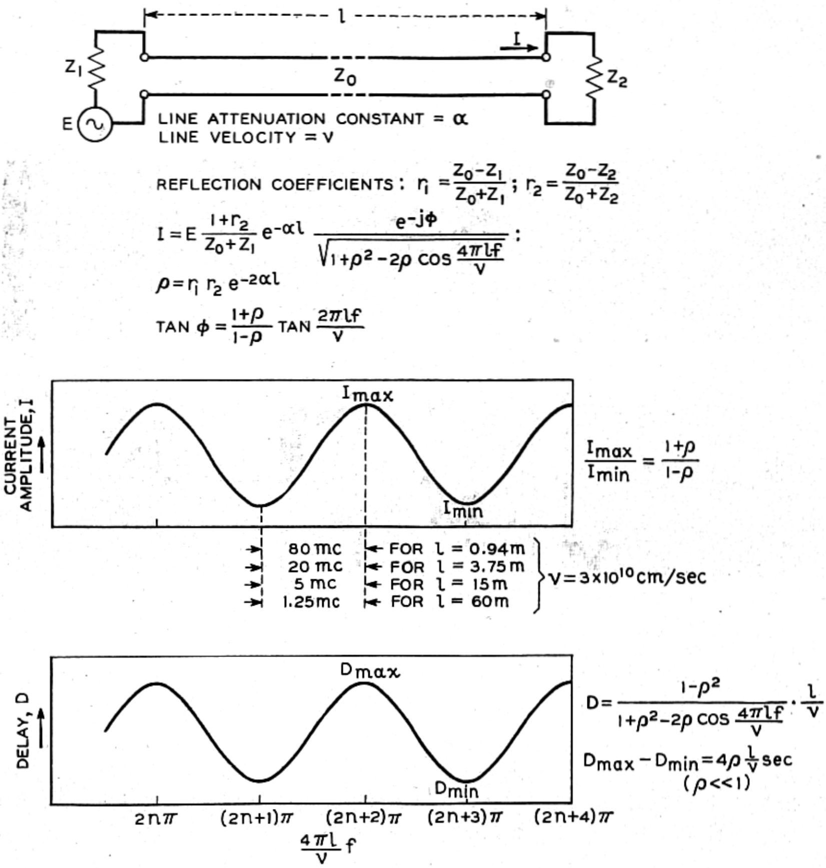


Fig. VI-1.—Effect of long lines on the amplitude and delay distortion of a repeater.

will produce variations in the amplitude and delay of the signal throughout the desired band. These variations may be greater than those produced anywhere else in the repeater.

The type of distortion originating in this way is illustrated by Fig. VI-1. In this figure there is represented an antenna of impedance Z_1 connected by a line of length ℓ and characteristic impedance Z_0, to a load, Z_2. In actual practice this load is the impedance presented by the filter to the line from the antenna. The variations produced in both the amplitude and delay of the signal currents are shown by the curves in the figure where relative change in the characteristics in question are shown as ordinates and values of the quantity $\frac{4\pi \ell f}{v}$ are shown as abscissas.

It will be seen from these curves that the amount of variation over a given band is a function of the line length. With short lines the amount of variation over a band 10 mc wide can be kept very small, but with lines about 150 feet in length there may be three or four full cycles of variations in a band 10 megacycles wide. Table A gives some typical values of the variations. Values are given for those degrees of mismatch which would produce standing wave ratios of 1, 1.4 and 2 db at the junctions. It will be seen that for lines between 100 and 200 feet in length the variations in some cases are greater than the limits achieved in the multistage amplifiers and other com-

TABLE A

$SWR_1 = SWR_2$ in *db* ($\alpha = 0$)	1	1.4	2
$r_1 = r_2$ ($\alpha = 0$)	0.058	0.082	0.115
Max. Amplitude Variation in *db*	0.058	0.116	0.233
Max. Delay Variation in 10^{-9} seconds			
$\ell = 100$ feet	1.33	2.67	5.33
$\ell = 200$ feet	2.67	5.33	10.66

ponents of the system. Furthermore, it is usually impractical to compensate or tune out these variations because they are functions of the electrical length of the transmission lines and subject to change with temperature, frequency, and other small mechanical and electrical changes. It would appear that with present techniques this may be one of the most serious sources of distortion in a long relay system.

VII. CONCLUSION

Various phases of our work on microwave repeater circuits have been discussed. Such a circuit, made up of several repeaters as described in the last section, may be looked upon as a four-terminal network having specified amplitude and delay characteristics and should be suitable for the transmission of any signals for which the characteristics are adequate whether they be television signals or those of one of the various forms of multiplex telephony. It is outside the scope of this paper to discuss the uses to which such a repeater circuit might be put or the terminal equipment that any particular service might require.

The New York-Boston repeater circuit has been built to provide the experimental field trials necessary to answer many remaining questions which deal with the performance of an actual circuit. The results of these trials will be reported in a separate paper.

Acknowledgements

The work which has been discussed in this report of necessity involved the combined efforts of many individuals. In addition to those already mentioned, practically every other member of the staffs of the Deal and Holmdel Radio Laboratories has made valuable contributions to the work, as have also many in other departments of the Bell Telephone Laboratories.

The development of the components of the New York-Boston circuit has been under the direction of Mr. Gordon N. Thayer.

In particular, we wish to acknowledge the support and stimulating advice of Dr. R. Bown under whose general direction the work progressed.