ENGINEERING CONSIDERATIONS for MICROWAVE COMMUNICATIONS SYSTEMS

GTE LENKURT

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PREFACE

This book is a lineal descendant of an earlier Lenkurt publication, "Microwave Path Engineering Considerations 6000-8000 MC", which was originally published in 1960 and re-issued in a slightly revised edition in 1961. The purpose of that publication was to assemble in one volume, in a readily usable and practical form, the basic information, principles, techniques and practices needed by an engineer engaged in the planning and engineering of line-of-sight paths for microwave communications systems.

The present volume retains essentially the same purpose, in an expanded, enlarged, and modernized version, reflecting the substantial changes which have taken place in the past decade. A much wider range of frequencies is covered, and new and extensive material is included on propagation, diversity, and reliability calculations. Also much expanded is the material on noise performance and noise calculation methods, and new material has been added on towers, transmission lines, and waveguides.

A preliminary edition of the present book was prepared in 1969 and a limited distribution made, with a view to obtaining comments and suggestions from the industry. Many valuable suggestions were received, most of which have been incorporated into the present edition to make it, we believe, a much improved work. The efforts of those who were kind enough to review the preliminary edition and send us their comments are gratefully acknowledged.

Although considerable effort has been made to eliminate errors, it is likely that some have survived, and we would appreciate having any of these called to our attention.

Robert F. White
Senior Staff Engineer
Systems Engineering Department

San Carlos, California
June, 1970

This second printing is unchanged from the 1970 edition except for correction of minor errors and the addition of a note at the bottom of Page 119.

R.F.W.

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I. INTRODUCTION

This book is intended to assemble in one volume, in a readily usable and practical form, a compendium of the best available information on the planning and engineering of line-of-sight microwave paths for communications systems. The emphasis is on techniques and practices, but a considerable amount of theoretical discussion is included as an aid in understanding the various phenomena which are important in microwave transmission.

Though the book will have its major value as a handbook for communications engineers engaged in microwave path planning, it is sufficiently general to be used by management people and by engineers in other fields, as an aid in understanding the characteristics of microwave communications systems.

Many formulas, charts, and figures are scattered throughout the work. Lists of the charts and the figures are included in the Table of Contents, and as an aid in day-to-day usage, selected formulas, charts, and figures have also been duplicated in appendices at the end of the book.

Throughout the text and in Appendices I and III, all material involving units of length is developed on the basis of feet and miles, the units in common use throughout North America. In order to make the book more useful to those who work in metric units, Appendix II provides metric versions of those formulas and equations which involve units of length. Also included in this Appendix are some suggested ways in which certain of the charts and figures can be used with meters and kilometers rather than feet and miles.

The book covers FM microwave systems which operate in one of a number of frequency bands between approximately 2 GHz and 16 GHz. The discussion is general so that it can be applied to any microwave equipment designed for operation in the above bands, provided due account is taken of differences in RF components, transmission lines, antennas, and propagation characteristics.

The use of frequency modulated microwave radio systems is widely recognized as a flexible, reliable and economical means of providing point-to-point communications facilities. These radio systems, when used with appropriate multiplex equipment, can carry from a few circuits up to a large number of voice, telegraph and data circuits. They can also be arranged to carry additional wide-band circuits for high-speed data, facsimile, or high-quality audio channels. Television is also carried by microwave radio, but, because of its wide baseband requirements, each video signal is usually carried on a separate radio channel.

Comparative cost studies usually prove a microwave system to be the most economical means for providing communication circuits where there are no existing cable or wire lines to be expanded. Where there are severe terrain or weather conditions to be overcome, the cost advantage becomes several-fold. For temporary facilities, and other applications where installation time is severely limited, the advantages of radio are obvious.

In applications where expandability is important, a microwave system can be installed initially with only a few carrier circuits. Then, as traffic increases, the capacity can be expanded by the addition of more channelizing equipment, or by paralleling radio equipment. Several radio channels can be arranged to use the originally installed antennas, waveguides, supporting structures, buildings and standby protection facilities.

II. MICROWAVE FREQUENCY BANDS

Most, if not all, microwave systems will be subject to regulation by the government of the country in which the system is to be located. In general, each country allocates specific bands of frequencies for specific services or for specific users. Within the United States the Federal Communications Commission (FCC) is the controlling authority for all systems except those operated by agencies of the Federal Government, the latter usually being placed in frequency bands separate from those controlled by FCC. In Canada, the licensing body is the Department of Communications. In many countries it is the Department of Posts and Telegraphs, or some similar entity. Most countries, other than the United States, follow the frequency allocations recommended by the International Radio Consultative Committee (CCIR).

Within the broad microwave portion of the radio frequency spectrum, the fixed allocations in effect at the time of preparation of this manual (1970) are given in the following tables:

Table A1. Microwave bands available to communications common carrier in the U.S.A. under Part 21 of the FCC Rules

BAND NAME	RANGE GHz	CENTER FREQ GHz	ATT'N IN dB AT 1.0 MILE	COMMENTS OR EMISSION LIMITATION
2 GHz	2.11 – 2.13 & 2.16 – 2.18	2.145	103.2	3500F9 †††
4 GHz	3.70 – 4.20	3.950	108.5	20,000F9 †
6 GHz	5.925 – 6.425	6.175	112.4	30,000F9 ††
11 GHz	10.7 – 11.7	11.20	117.6	50,000F9

† Shared with Satellite-to-Earth
†† Shared with Earth-to-Satellite
††† 2.11 to 2.12 shared with space telecommand transmit

Table A2. Microwave bands available for private and local government microwave systems within the U.S.A. under various Parts of FCC Rules. (Not all bands are available to all types of users.)

BAND NAME	RANGE GHz	CENTER FREQ GHz	ATT'N IN dB AT 1.0 MILE	COMMENTS OR EMISSION LIMITATION
2 GHz	1.85 – 1.99	1.920	102.3	8,000F9
	2.13 – 2.15 & 2.18 – 2.20	2.165	103.3	800F9
	2.45 – 2.50	2.475	104.5	Shared
6 GHz	6.575 – 6.875	6.725	113.1	10,000F9
12 GHz	12.2 – 12.7	12.450	118.5	20,000F9

Table A3. Microwave bands available for TV Auxiliary Services within U.S.A. under Part 74 of FCC Rules.

BAND NAME	RANGE GHz	CENTER FREQ GHz	ATT'N IN dB AT 1.0 MILE	COMMENTS OR EMISSION LIMITATION
2 GHz	1.99 – 2.11	2.050	102.8	STL, etc.
7 GHz	6.875 – 7.125	7.000	113.5	STL, etc.
12 GHz	12.7 – 13.25	12.975	118.9	STL, etc.
	12.7 – 12.95	12.825	118.8	CARS †

† Comm. Antenna Relay Service

Table A4. Microwave bands available for Federal Government Services within U.S.A.

BAND NAME	RANGE GHz	CENTER FREQ GHz	ATT'N IN dB AT 1.0 MILE
2 GHz	1.71 – 1.85	1.780	101.6
	2.20 – 2.29	2.245	103.6
4 GHz	4.40 – 5.00	4.700	110.0
7-8 GHz	7.125 – 8.40 †	7.750	114.4
14 GHz	14.40 – 15.25	14.825	120.0

† 7.25 – 7.30 GHz reserved for Satellite-to-Earth
7.975 – 8.025 GHz reserved for Earth-to-Satellite

Table A5. Microwave bands per CCIR Recommendations. (OSLO–1966)

REC. NO.	RANGE GHz	CENTER FREQ GHz	ATT'N IN dB AT 1.0 MILE	CAPACITY CHANNELS
283-1	1.7 – 1.9	1.808	101.7	60 & 120
	1.9 – 2.1	2.000	102.6	
	2.1 – 2.3	2.203	103.5	
382-1 or (279-1)	1.7 – 2.1	1.903	102.2	600 to 1800 or (300)
	1.9 – 2.3	2.101	103.0	
	3.8 – 4.2 †	4.0035	108.6	
	3.7 – 4.2 *†	3.950	108.5	
383-1	5.925 – 6.525 ††	6.175	112.4	1800 or equiv.
384-1	6.43 – 7.11	6.770	113.2	2700 or equiv.
385	7.125 – 7.425 **	7.275	113.8	60, 120, 300
	7.250 – 7.550 **	7.400	114.0	
	7.425 – 7.725	7.575	114.2	
	7.550 – 7.850	7.700	114.3	
386-1	8.200 – 8.500	8.350	115.0	960
	7.725 – 8.275 **	8.000	114.7	1800
387	10.7 – 11.7	11.200	117.6	960

† Shared with Satellite-to-Earth
†† Shared with Earth-to-Satellite
* Alternate for U.S.A.
** 7.250 – 7.300 GHz Reserved for Satellite-to-Earth
7.975 – 8.025 GHz Reserved for Earth-to-Satellite

Table A6. Microwave bands in Canada per Department of Communications Plan.

PLAN NO.	RANGE GHz	CAPACITY & USE
303	1.700 – 1.900	12–60 channels on interleaved plan 60–300 channels on main plan
304 *	1.900 – 2.300	600–1800 channels or TV
300	2.548 – 2.686	Instructional TV Service
302 *	3.550 – 4.200	600–2700 channels or TV
301	5.925 – 6.425 6.425 – 7.125	600–1800 channels or TV 300–960 channels or TV 600–2700 channels Studio TV Links, etc.
305	7.125 – 7.725	12–120 channels 300–960 channels 24 PCM channels One-way TV or Radar on certain RF channels Passive reflectors may be used Four-frequency plan
306 *	7.725 – 8.275	600–1800 channels or TV
– –	8.275 – 8.400 10.55 – 10.68	One-way TV or Radar Experimental PCM channels
387 CCIR Rec.	10.70 – 11.70	300–960 channels or TV Passive reflectors may be used Four-frequency plan

*System design objective should be 3.0 pW/km or better

NOTE: Assignments are given on the basis of use rather than type of user. Government policy is that no frequency bands will be designated common carrier, etc., although in actual practice the common carrier would normally use those bands marked with asterisk, to obtain performance desired.

III. ROUTE AND SITE SELECTION

A. Order of Procedure

As a starting point, it is assumed that preliminary facility planning (including operational requirements, traffic studies, expansion potential, reliability requirements and cost studies) has been completed to such a degree that the points to be served have been fixed, and the required system capacity has been determined.

Preliminary studies for site location can usually be made from maps and aerial photographs prepared by other agencies; however, the final site selections must be made from field surveys, and the profiles and notes thereby derived.

B. Sites

1. The Requirements

Terminal sites are more often than not locations of existing structures or facility terminals, but the intermediate sites are located with considerable emphasis on factors having to do with propagation over the intermediate paths, and possible interference from sources internal or external to the system. In some cases there may be problems in the use of an existing building for a terminal radio site, even though common maintenance, facility layout and economy tend to dictate such selection. If the building is of adequate height to mount antenna fixtures on the roof without fear of path blockage by other buildings, this will be an ideal terminal point. The possibility of future building construction along the path must, of course, be taken into consideration. The mounting of antenna fixtures on the roof, or constructing a tower on the roof, may require investigation into the architectural and structural plans of the building to determine whether the structure is adequate, and the cost of building modifications to accomplish the purpose must be considered. On rare occasions plans for future floor additions must be considered. It may be desirable to add the steel for one or more floors for the future addition, and then add the antenna structure on top. This plan must include maintenance access to the antennas and waveguide. Such a plan should be worked out having in mind the views of city officials who may require building facing for the vacant floors for appearance reasons.

Where additional height is required and building structure is such as to reject all of the above considerations, a separate tower on the building lot or adjacent to it may be the solution. In any event, a terminal must be the starting point on site selection.

Intermediate Repeater Sites

The choice of intermediate repeater sites is greatly influenced by the nature of the terrain between sites. In preliminary planning it may be assumed that, in relatively flat areas, the path lengths will average out between 25 and 35 miles for the frequency bands from 2 GHz through 8 GHz, with extremes in either direction depending upon terrain. In the 11 GHz and higher bands, the pattern of rainfall has a large bearing on path length. These factors will be covered in more detail under subsection C. Microwave Paths.

In all cases the sites should be as level as possible. The need for leveling should be considered as a part of the cost of preparing the site.

Since maintenance access is very important, an access road must be considered. In addition, the availability of AC power of suitable voltage, and the possibility that telephone facilities may be needed, enter into the site selection.

The possibility of interference, internal or external to the system, must be considered. Internal interference may take the form of overreach, junction or adjacent section interference. External interference may take the form of radar interference, interference from nearby radio systems of similar frequency, or interference induced from unfiltered lower frequency radio systems.

Site Considerations

There are a number of site considerations which must be investigated in the field survey, and the accumulated data should be recorded for subsequent use. Some of these are indicated in the following paragraphs.

1. A full description of each site by geographical coordinates, political subdivision, access roads and physical objects with which it can be identified. Geographical coordinates should be computed to the nearest second of latitude and longitude for the exact location recommended for the tower. Should this location be changed in later engineering work, the coordinates should be corrected accordingly.
   Note: This degree of accuracy is required primarily because it is specified in FCC Rules.

2. Any unusual weather conditions to be expected in the area, including amount of snow and ice accumulation, maximum expected wind velocity and range of temperatures.

3. A description of the physical characteristics of the site, indicating the amount of leveling required, removal of rocks, trees or other structures, etc.

4. The relationship of the site to any commercial, military or private airport within several miles. It is very important to determine the relationship of the site to the orientation of runways where planes may be taking off or landing. This information is needed to determine compliance with government regulations on potential obstructions to air traffic.

5. The mean sea level elevation of the site at the recommended tower location, and the effect on that elevation of any necessary leveling.

6. A full description or recommendation for an access road from the nearest improved road to the proposed building location.

7. There is a possibility that building code restrictions may be involved. Such sites should be avoided if practicable.

8. The nearest location where commercial electric power of suitable secondary or distribution voltage may be obtained, and the name and office location of the power company. In countries or locations where public power is furnished, similar information is required, but the procedures for contact may be somewhat different.

9. If telephone communication is desired, the nearest telephone facility should be indicated together with the name of the company and the type of service available.

10. Any other facts that can be determined at the time of the survey which might bear on the proposed construction.

C. Microwave Paths – General Appreciation of Path Influences

1. Description of Microwave Beam

For simplicity, the following discussion will treat the microwave beam in general as the line representing the longitudinal center of the beam or main lobe, particularly when discussing line-of-sight clearances. The microwave beam behaves much like a light beam insofar as atmospheric influences are concerned, and is subject to certain other external influences in a manner described in the following discussion.

Influence of Terrain and Obstructions

The microwave beam is influenced by the intermediate terrain between stations and by obstacles. It tends to follow a straight line in azimuth unless intercepted by structures in or near the path. In traveling through the atmosphere it usually follows a slightly curved path in the vertical plane, i.e. it is refracted vertically due to the variation with height in the dielectric constant of the atmosphere; generally slightly downward, so that the radio horizon is effectively extended. The amount of this refraction varies with time due to changes in temperature, pressure and relative humidity, which control the dielectric constant. This is illustrated in the lower portion of Figure 1.

At the point of grazing over an obstacle the beam is diffracted. There is a very small shadow area where some energy is redirected in a narrow and rapidly diminishing wedge toward total shadow. The angle enclosed by the wedge of diminishing energy behind the diffracting object, from full signal to full shadow, is extremely small. This phenomenon is illustrated in the upper portion of Figure 1. The principal point to be stressed here, is that when the centerline of the beam just grazes an obstacle, there is a loss of energy reaching the far antenna. The loss may be from six to twenty decibels, depending on the type of surface over which the diffraction occurs. It has been shown experimentally that a knife-edge diffraction obstacle will produce a loss of six decibels at grazing. A smooth surface, such as flat terrain or water, which actually follows the earths contour, will produce the maximum loss at grazing. Most obstacles normally found in the path will produce a loss somewhere between the above limits. Trees tend to produce a loss close to six decibels. In order to minimize diffraction losses, line-of-sight microwave paths are planned to have better than grazing clearance even under the most adverse atmospheric conditions.

Most physical objects in the line-of-sight will tend to block the beam, causing loss of signal at the receiver. Deciduous trees which may cause relatively less loss in winter, can totally block the path

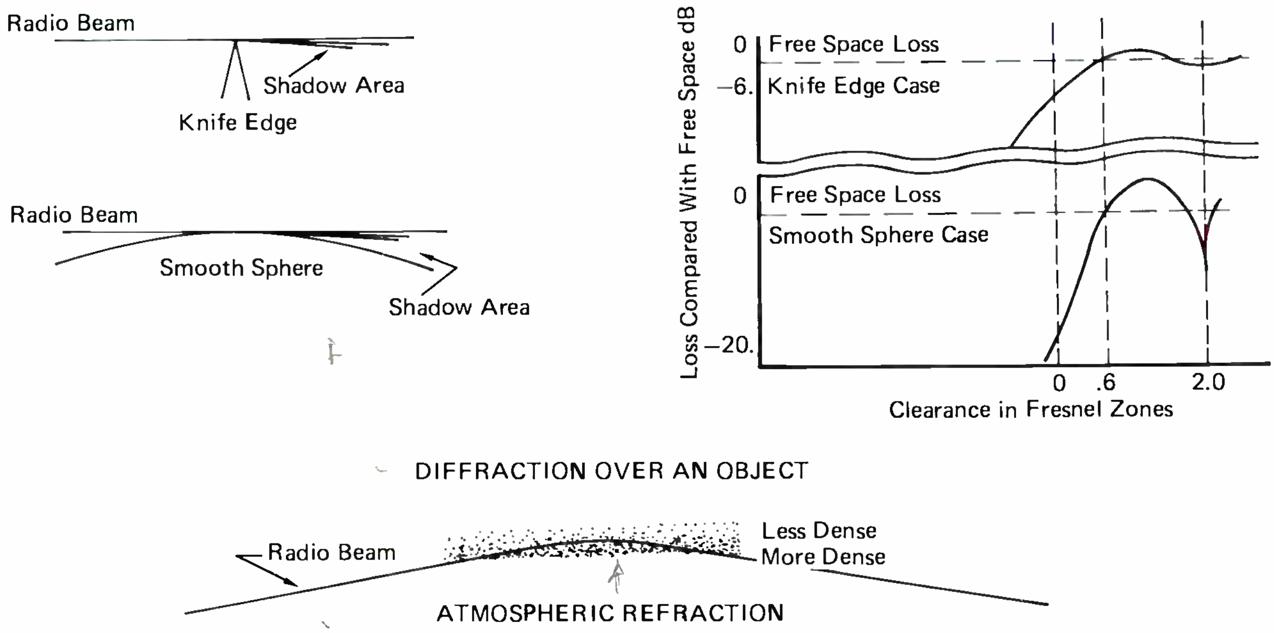


Figure 1. Diffraction vs Refraction

in summer when the leaves are out. In all cases, trees should be considered as blocking when in the path line, unless the beam has adequate clearance over the trees.

The beam can be reflected from relatively smooth terrain and water surfaces, just as a light beam is reflected by a mirror. Since the wavelength is very much longer than light waves, the criterion of smoothness is quite different. The criterion of smoothness is also quite different for very small angles of incidence than it is for large angles. This can be illustrated for visual light in the case of an asphalt highway. When viewed directly, the surface looks slightly rough and does not reflect light well; however, when viewed from a distance at a very small angle, it looks like a mirror or wet surface.

An important concept in analyzing microwave propagation effects, particularly those of diffraction, refraction, reflection, and the effects of terrain and obstructions, is that of the Fresnel zone. The first Fresnel zone radius is a kind of "rubber" unit, which is used to measure certain distances (path clearances in particular) in terms of their *effect* at the frequency in question, rather than in terms of feet. The 2nd and higher order Fresnel zones are also very important under certain conditions, such as highly reflective paths.

The Fresnel zones are a series of concentric ellipsoids surrounding the path. The first Fresnel zone is the surface containing every point for which the sum of the distances from that point to the two ends of the path is exactly one-half wavelength longer than the direct end-to-end path. The nth Fresnel zone is defined in the same manner, except that the difference is n half-wavelengths.

Since the cross-section of the Fresnel zones at any point along the path is a series of concentric circles completely surrounding the path, it is important to note that clearance requirements, expressed in Fresnel zones, apply to the sides and above as well as below the path. Formulas and graphs for calculating Fresnel zone radii are given in a later section.

Influence of Weather

Although a microwave beam is conventionally shown as a line, the actual method of propagation is as a wave front, and the important portion of the wavefront involves a sizeable transverse area.

In order to ensure free space propagation it is essential that all potential obstructions along a path are removed from the beam centerline by at least $0.6\ F_1$, where F_1 is the radius of the 1st Fresnel zone at the point of the obstructions.

For this reason, it is necessary to provide path clearance over intermediate objects which is somewhat greater than line-of-sight. Because refractive bending varies in cycles daily and changes erratically at times, the clearance over the intermediate terrain must be adjusted to minimize the losses at the extreme bending conditions. Normally, as mentioned previously, the beam is bent downward by atmospheric refraction so that the radio horizon is effectively extended. Nevertheless, at times the atmospheric conditions may be such that the beam is bent upward, effectively reducing the clearance over terrain in the path. These phenomena and Fresnel zones will be discussed more fully in a later section.

Influence of Rain and Fog at Higher Frequencies

At microwave frequencies up to the 6 and 8 GHz bands, rain attenuation as such is not considered sufficient to warrant special consideration in the design of the paths, except in very extreme situations. Under saturation rain conditions, a 30 mile path might suffer only a few dB attenuation at 6 GHz. Uniform fog conditions can be considered in much the same light. However fog conditions often result from atmospheric conditions such as temperature inversion, or very still air, accompanied by stratification; the former tends to negate clearances, and the latter causes severe refractive or reflective conditions, with unpredictable results. In areas where these conditions prevail, shorter paths and adequate clearances are recommended.

At microwave frequencies of 11 and 12 GHz or above, rain attenuation can be very serious. The amount of attenuation depends upon the rate of rainfall, the size of the drops and the length of exposure. Accordingly, in areas of heavy rainfall where extremely high reliability is required, short microwave paths are recommended. It should be noted that the rate of rainfall to be considered is not the annual total rainfall, but the instantaneous intensity at the time of occurrence. Thus, the west coast areas of Oregon and Washington in the United States, despite having frequent rain, are considered not likely to experience serious rain attenuation at 11 GHz for paths up to 30 miles, because the actual rate of rainfall at a given time is very low. For other coastal areas the reverse is often true.

Influence of Objects in Azimuth

The influence of objects in azimuth is not confined entirely to those which are directly in the path. While the microwave energy is concentrated in a fairly narrow beam, it tends to spread gradually as it is propagated through the atmosphere. There are also minor lobes of the antenna which, although having much less power than the main lobe, are transmitted in different directions. The corridor of free transmission required will run up to 230 feet at the center of a 40 miles path at 4 GHz for example. However, the influence of objects in azimuth can run well beyond this indication. Figure 2 illustrates the case of objects in azimuth. The potential problem with off-path objects is reflections, and these usually turn out to be from buildings. Energy traveling the longer reflection path lags behind the main beam. The most serious case is one of multiple reflections which might occur, for example, when a beam is transmitted down a street with tall buildings on both sides. In this case the delay is likely to be so great as to cause delay distortion in the baseband. A small round object is generally incapable of reflecting sufficient energy in one direction to cause trouble, as the reflection beam is diverse. However, a very large round object has been known to cause trouble. Such an object would be a large gas container or oil storage tank, running up to 150 feet and more in diameter.

Atmospheric Absorption

Atmospheric absorption due to oxygen and water vapor also exists. The magnitude of the effect is quite small at the lower frequencies (2-8 GHz), and is usually neglected. Even in the higher bands the effect is relatively small, but not entirely negligible. Since the amount of attenuation from this phenomenon is directly proportional to path length, it is usually significant only on longer paths. A table of absorption attenuation as a function of path length and frequency band is given in section IV.C.

D. Sources of Path Data

1. Maps

Maps are the principal sources of basic data, both for office study which usually precedes the field survey, and for the field survey itself. There are a number of types of maps which will be discussed herein. Experience has shown that maps covering a rather large area in the general territory to be surveyed, represent good work and record sheets which, when posted as the map survey progresses, illustrate the progress, general locations, angles and place names. A good map of this type is

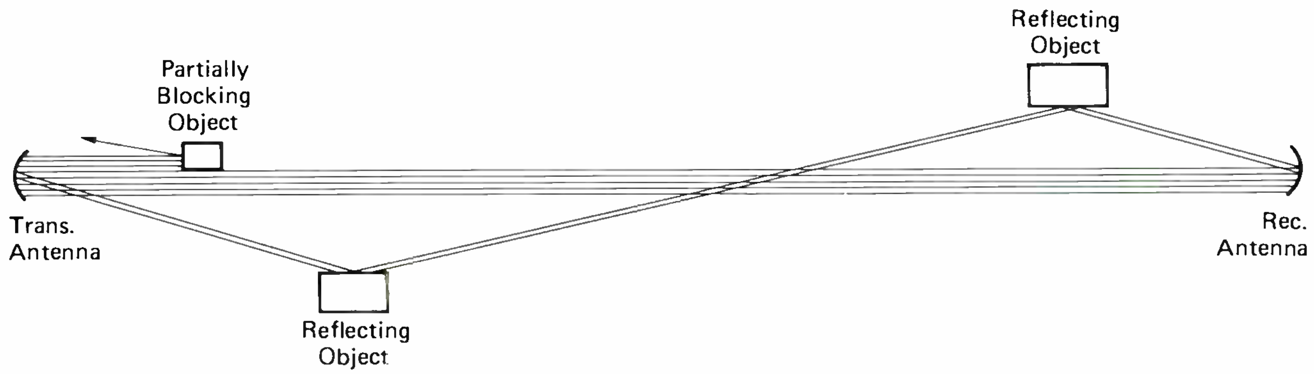


Figure 2. Blocking and Reflecting Effects in the Horizontal Plane

the Aeronautical Chart, which is published for most countries where commercial and private air navigation are the rule. In the United States they are published by the U.S. Coast and Geodetic Survey. In addition to showing large areas they show some topography in very large contours, and the air navigation routes. They may be ordered with the flight chart overlay which shows the established commercial airways. The Coast and Geodetic Survey publishes and distributes aeronautical charts of the United States, its Territories and Possessions. Charts of foreign areas are published by the USAF Aeronautical Chart and Information Center (ACIC), and are sold to civil users by the Coast and Geodetic Survey. A catalog of aeronautical charts is available from one of the following field offices, which will also supply the aeronautical charts desired on a specific order. The catalog will be supplied free of charge. The charts are sold at the prices listed in the catalog. Exact payment must accompany the order.

Chief, New York Field Office
Coast and Geodetic Survey
Room 1407 Federal Office Building
90 Church Street, New York, N.Y. 10007

Mid-Continent Field Director
Coast and Geodetic Survey
Room 1436 Federal Building
601 East 12th Street
Kansas City, Mo. 64106

West Coast Field Director
Coast and Geodetic Survey
Room 121 Courthouse
555 Battery Street
San Francisco, Ca. 94111

The Aeronautical Charts cover large areas of a state, and often several states on one chart. For a large part of the United States, they typically show elevations in contours of 500 or 1000 feet, and therefore are not useful in actual plotting of profiles. They show airports, airways, major aerial obstructions and large topographical features such as lakes and mountain ranges. If the terminal locations are plotted on the appropriate aeronautical chart(s), and the intermediate charts (if any) are spliced in, an overall view of the possible route is obtained. Generalizations can be made concerning features to be avoided and possible areas of search. As the map survey proceeds, the preferred and alternate sites arrived at by profiling from other maps (discussed below) can be plotted using the precise latitude and longitude of each. The composite chart then becomes a good overall worksheet and provides a quick check for the possibility of overreach or other interference (discussed in a later section).

The basic maps from which office profiles can be made are the topographic maps such as those published by the U.S. Geological Survey. These are to be found in quadrangles of different sizes depending upon the date of the survey and the area surveyed. They also show different elevation contour intervals in different areas, depending on the area, date of survey and size of quadrangle. These contour intervals generally range from 2.5 to 100 feet. Topographic maps in the United States based on surveys made prior to 1920 have been found to contain some errors both in elevations and locations of topographic details. In the absence of later surveys however, they can be used as a rough guide until specific field checks are made.

The Geological Survey has, for a number of years, been making surveys which will eventually cover all of the United States and Puerto Rico. The published maps covering the more recent surveys generally fall into one of the following sizes and scales:

7-1/2 Minutes of latitude and longitude; Scale 1:24,000 (1 inch = 2000 ft) or 1:31,680 (1 inch = 1/2 mile).

15 Minutes of latitude and longitude; Scale 1:62,500 (1 inch = approx. 1 mile).

30 Minutes of latitude and longitude; Scale 1:125,000 (1 inch = approx. 2 miles).

1 Degree of latitude and 2 degrees of longitude; Scale 1:250,000 (1 inch = approx. 4 miles).

There is for each state and for Puerto Rico an index circular showing all U.S.G.S. topographic maps distributed. They show the quadrangle location, name, survey date and publisher (if other than U.S.G.S.). There are also listed special maps and sheets with prices, map agents and Federal distribution centers, addresses of mapreference libraries, and detailed instructions for ordering topographic maps.

The index circulars are accompanied by a folder describing the topographic maps. They are furnished free on request and may be obtained from one of the following offices:

For maps East of the Mississippi and Hawaii

U.S. Geological Survey — Map Information
1200 South Eads Street
Arlington, Virginia 22202

For maps West of the Mississippi River, all of Louisiana and Alaska

U.S. Geological Survey — Map Information
Room 15426 Federal Building
1961 Stout Street
Denver, Colorado 80225

For Alaska they may also be obtained from

U.S. Geological Survey — Map Information
Room 108 Skyline Building
508 Second Avenue
Anchorage, Alaska 99501

The U.S.G.S. also maintains sales counters in Washington, D.C.; Denver, Colorado; Salt Lake City, Utah; Sacramento, San Francisco, Menlo Park and Los Angeles, California; and Anchorage, Juneau and Palmer, Alaska, where the maps can be purchased in person. The particular street addresses are subject to change from time to time, but can usually be found in the local telephone directory. There are also private agents who sell quadrangle maps at their own prices. Their names and addresses are listed in the state index circulars.

Accurate topographic maps are available for many areas of Canada. All of Canada is covered by maps published on the scales of 1:506,880 (1 inch = 8 miles) and 1:1,000,000 (1 inch = 15.783 miles). Coverage on other large scales is not complete. Many areas are covered by maps published on the scales of 1:50,000 (1 inch = 0.79 miles), 1:63,360 (1 inch = 1 mile), 1:126,720 (1 inch = 2 miles) and 1:253,440 (1 inch = 4 miles). The indices of these maps and the maps themselves may be purchased directly from:

Department of Energy, Mines and Resources
Surveys and Mapping Branch
615 Booth Street
Ottawa, Ontario, Canada

They may also be purchased at local stationers, but this is not a reliable source.

Additional maps may be obtained from the Department of Mines, Lands and Forests, or Department of Natural Resources of the Provincial Governments in the appropriate provincial capitals.

Aeronautical charts for Canada on scales of 8 miles to the inch and 16 miles to the inch, and air photos may also be obtained from the Department of Energy, Mines and Resources, surveys and mapping branch.

Additional maps which may be useful are U.S. Forestry Service maps, and strip maps of railroads, pipe lines, power companies and telephone companies.

County highway maps published by the state highway departments in the United States have been found to be useful in making the field surveys. They usually show man-made structures and are often more up-to-date on road information, than are many topographic maps. In addition, they frequently show some of the bench marks and occasionally secondary level points. These must be considered cautiously, as grades,culverts and poles on which they may be located are subject to change.

Comparable topographic mapping in other countries of the world is usually available.

2. Aerial Photography

Aerial photography is often useful in rough terrain because it can show more of the details of a prominent terrain feature than a topographic map, and also shows trees and other obstructions. An index map showing all Government and military aerial photography in a given area can be obtained by writing the Superintendent Of Documents, Washington D.C. 20402.

Aerial photography is also used in the process of preparing path profiles by the technique known as photogrammetry.

E. Path Profiles

After tentative antenna sites have been selected, and the relative elevation of the terrain (and obstacles) between the sites has been determined, a profile chart can be prepared. In some cases a complete profile will be necessary; in other cases only the end sites and certain hills or ridges need to be plotted.

1. Curvature

The relative curvature of the earth and the microwave beam is an important factor when plotting a profile chart. Although the surface of the earth is curved, a beam of microwave energy tends to travel in a straight line. However, the beam is normally bent downward a slight amount by atmospheric refraction. The amount of bending varies with atmospheric conditions. The degree and direction of bending can be conveniently defined by an equivalent earth radius factor, K. This factor, K, multiplied by the actual earth radius, R, is the radius of a fictitious earth curve. The curve is equivalent to the relative curvature of the microwave beam with respect to the curvature of the earth, that is, it is equal to the curvature of the actual earth minus the curvature of the actual beam of microwave energy. Any change in the amount of beam bending caused by atmospheric conditions can then be expressed as a change in K. This relative curvature can be shown graphically; either as a curved earth with radius KR and a straight line microwave beam, or as a flat earth with a microwave beam having a curvature of KR. The second method of plotting is preferred, because it (1) permits investigation (and illustration) of the conditions for several values of K to be made on one chart, (2) eliminates the need for special earth curvature graph paper, and (3) facilitates the task of plotting the profile. It is convenient to plot the profiles on regular 10 by 10 divisions to the inch, reproducible graph paper of the 11 by 17 inch or B size.

2. Scales

A horizontal scale of two miles to the inch has been found to be very convenient. It permits paths of up to 30 miles in length to be plotted on one sheet. For longer paths it is not difficult to trim and splice two sheets together with small pieces of transparent tape on the reverse side. (It is suggested that pieces of tape not over 1.5 inches long to be used so that tape shrinkage will not ruin the charts. This precaution should also be made when splicing maps together).

More than one vertical scale will be necessary to cover all types of terrain. A basic elevation scale of 100 feet to the inch has been found to be quite convenient for all cases where the changes in elevation along the path do not exceed 600 or 800 feet. For paths in hilly country, a more compressed scale of 200 feet to the inch is convenient, and for mountainous country, it may be necessary to use a vertical scale of 500, or 1000 feet to the inch. *It should be noted that if the distance scale is doubled, the height scale should be quadrupled to preserve the proper relationship.*

3. Equivalent Earth Profiles

A path profile plotted on rectangular graph paper with no earth curvature (as suggested above), and with the microwave beam drawn as a straight line between the antennas represents conditions when the beam has a curvature identical to that of the earth (i.e. there is no relative change in curvature between the beam and the earth) and the equivalent earth radius, K, is equal to infinity. This is one of the extreme conditions that must be investigated when making a study of the effect of abnormal atmospheric conditions on microwave propagation over a particular path. In order to complete a propagation study, it is necessary to show the path of the beam (relative to the earth) for other expected values of K. In all cases, it is of interest to study the path under normal atmospheric conditions when K is equal to 4/3.

The curvature for various values of K can be calculated from the following relationship:

$$h = \frac{d_1 \; d_2}{1.5\,K} \tag{1}$$

where h = the change in vertical distance from a horizontal reference line, in feet

d_1 = the distance from a point to one end of the path, in miles

d_2 = the distance from the same point to the other end of the path, in miles

K = the equivalent earth radius factor

For the K conditions of primary interest in path analysis, Equation (1) takes the following forms:

$$h_{(K = \infty)} = 0 \tag{1-A}$$

$$h_{(K = 4/3)} = \frac{d_1 \; d_2}{2} \tag{1-B}$$

$$h_{(K = 2/3)} = d_1 \; d_2 \tag{1-C}$$

$$h_{(K = 1)} = .67 \, d_1 \; d_2 \tag{1-D}$$

The forms 1-B and 1-C are particularly simple, useful, and easy to remember. Equation 1 and its derivative forms are basic to all microwave path engineering, and most transmission engineers immediately commit them to memory.

Figure 3 provides a graphical means of calculating and plotting a section of a parabolic arc which can be used as a base-line for a curved earth profile plot, or (inverted) as a curved ray template to be used with a flat earth profile plot. The generating equation is that of a parabola with its origin at the apex of the curve. The parameter d represents distance from center, along the x axis, in miles, while h gives distance along the y axis, in feet, as a function of the distance from the center (origin) and the K factor being used.

Equivalent earth profile curves have been calculated from the above formula for values of K normally investigated. These curves, plotted to a convenient scale, are shown in the full size version of Figure 4 inserted loose in the back of the book. This figure can be used to make templates of plastic or cardboard, or it can be used as an underlay to profiles that have been plotted on graph paper. When the final antenna heights have been selected, the path of the beam can be traced directly from the curves, as shown in Figure 5. Besides making sure that the correct scales are used, one other precaution is necessary when using the curves in this manner; it is necessary to keep the horizontal lines of the profile chart parallel with the horizontal lines on the curves. This will automatically insure that the correct portion of the curve is being used. (The reduced version of Figure 4 on page 14 is illustrative only).

Where extremely large differences in elevation exist along the path, it may be more convenient to plot the profile with respect to a curved earth with radius KR and to use a straight line to represent the microwave beam. The curves of Figure 4 can be used to establish a curved base-line, by inverting the curves and tracing the curved line for the desired value of K on 10 X 10 to the inch graph paper. The curved baseline is assigned an altitude which is the nearest hundred foot interval below the lowest altitude required for the profile. The path profile is then plotted above the baseline. A straight line drawn between the proposed antenna locations then indicates the path of the microwave beam for the chosen value of K.

Contrary to a widely-held belief, it is not necessary with either approach to locate the center point of the path at the apex of the curve to "balance" it about the center. If the curves are accurately constructed, the same results will be obtained whether the path is centered or offset; this characteristic of the parabolic arc can be shown mathematically as well as experimentally.

Another very useful method is to construct the curved baseline by calculation; this method allows the use of any convenient scales and any value of K. It also allows the use of millimeter-ruled paper if desired, giving somewhat greater resolution. In this method the "earth bulge" at a number of points along the path is calculated by equation (1) and is plotted above the bottom line of the profile paper. A smooth parabolic curve is then drawn through the points to provide the curved baseline for the profile. A value of K = 4/3 is usually chosen for such profiles, in which case the

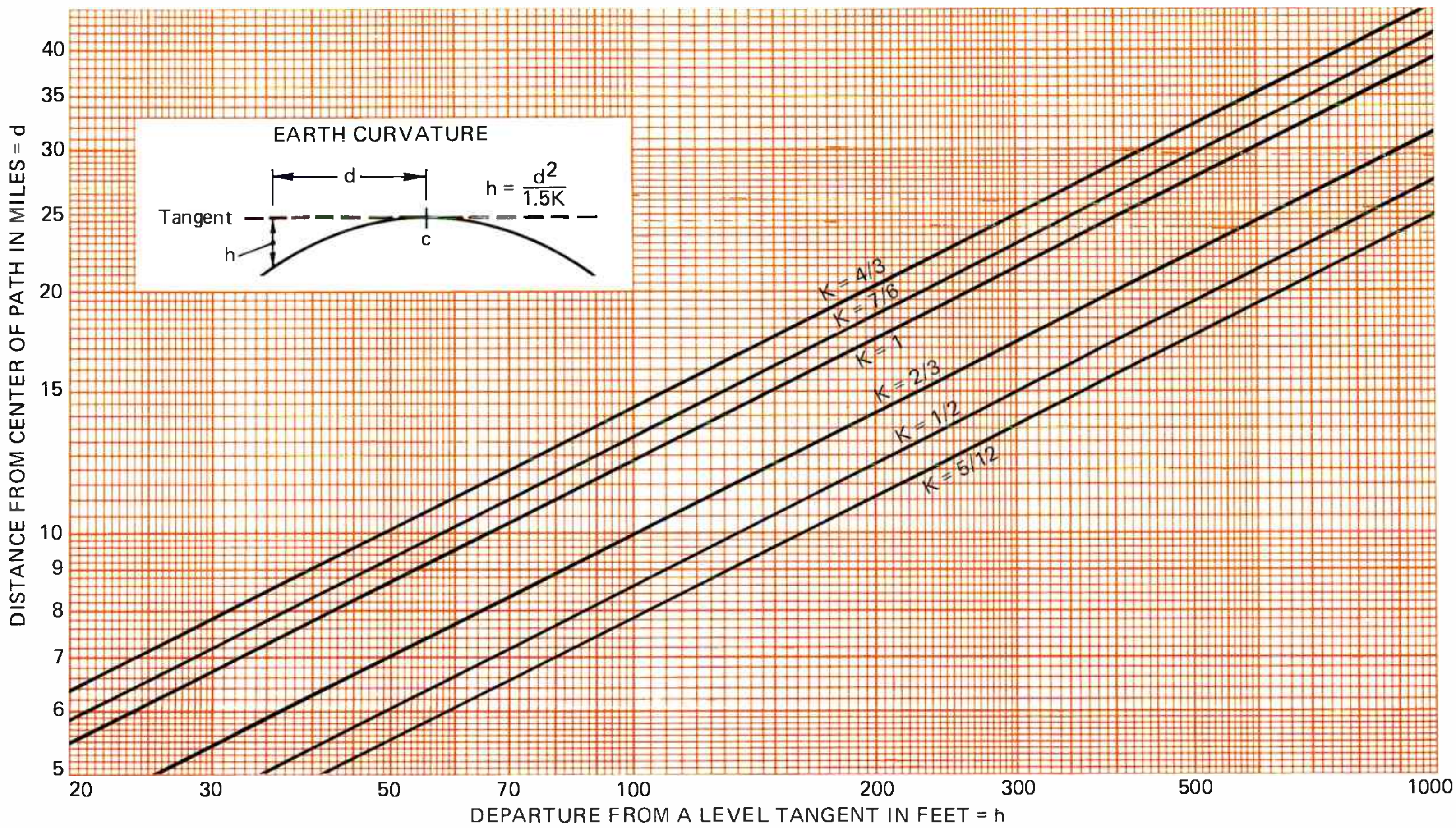


Figure 3. Earth Curvature for Various Values of K

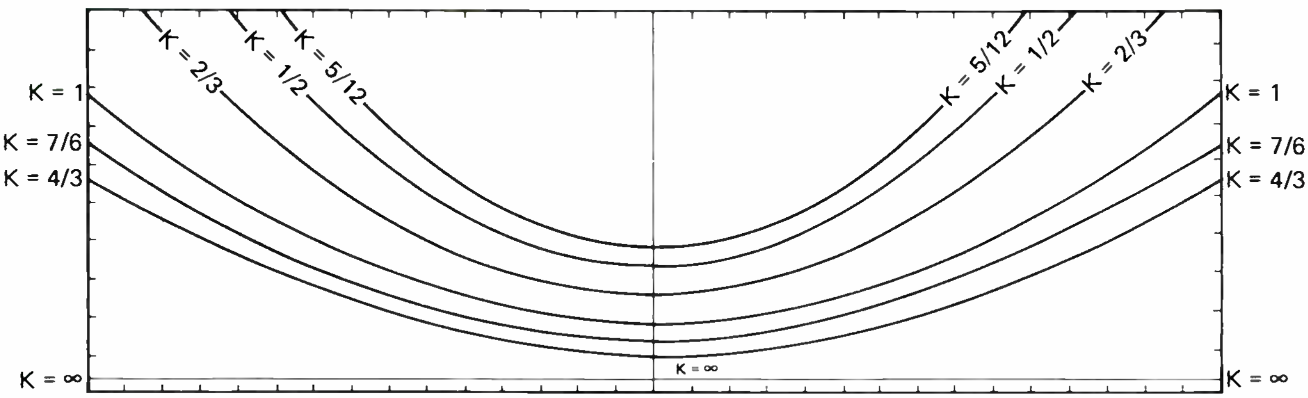


Figure 4. Equivalent Earth Profile Curves.
(Example only; see Appendix for full-scale working template of Fig. 4)

equation reduces to $h = \frac{d_1 d_2}{2}$. For K = 2/3 the equation is $h = d_1 d_2$, and the earth bulge at any point is just double that at K = 4/3, making it relatively easy to consider the effects of a change from K = 4/3 to K = 2/3.

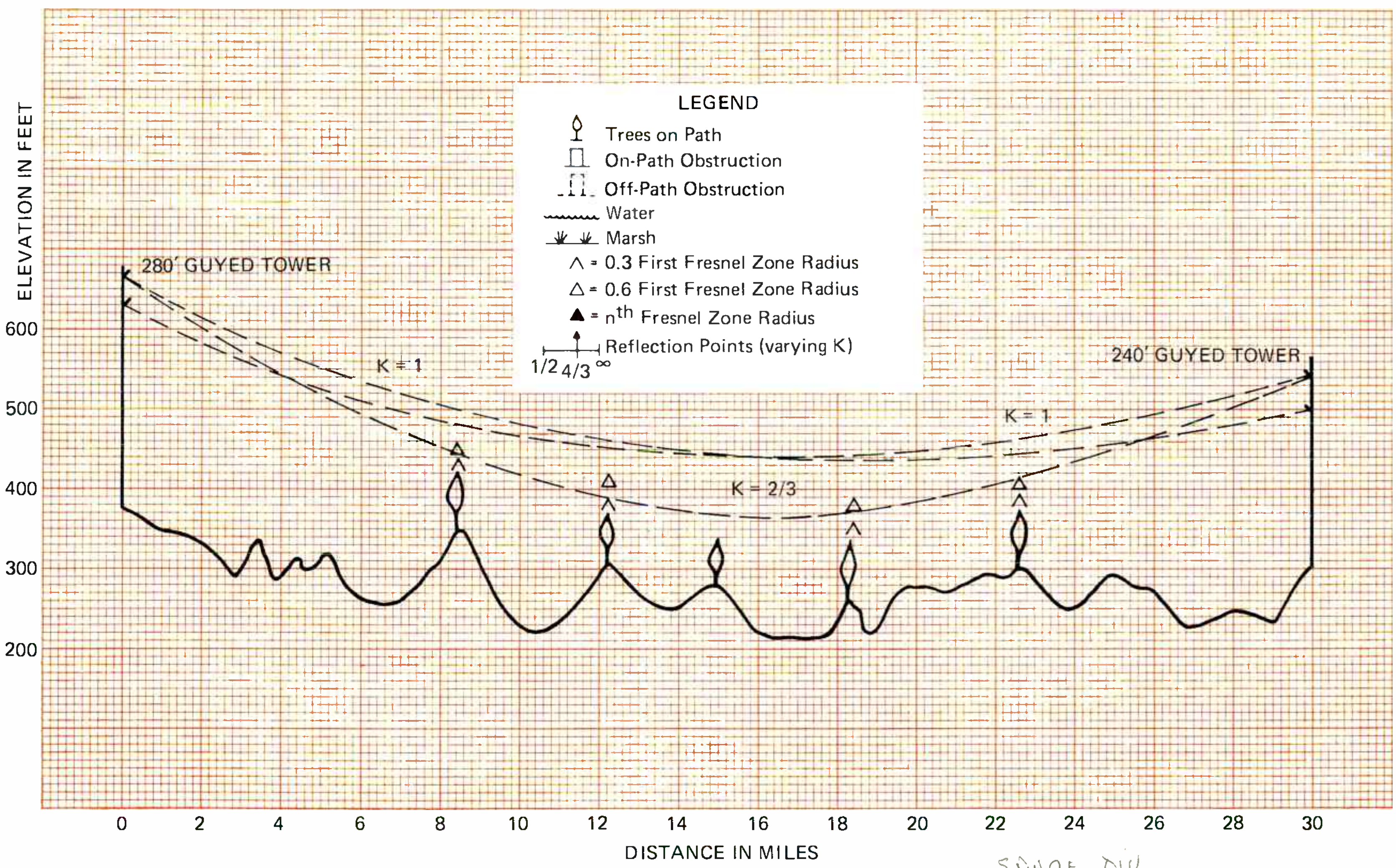
Figure 5 shows an example of a path profile on the flat earth basis. The curved beam paths are traced with the aid of the curves of Figure 4 (note that Figure 5 has been reduced in size; the original was on 10 X 10 to the inch graph paper). Figure 5 illustrates a space diversity arrangement with 40' vertical separation of the diversity antennas. Clearance criteria for this path were at least $0.3F_1$ clearance at K = 2/3 for top-to-top antennas and at least $0.6F_1$ clearance at K = 1 for top-to-bottom antennas.

Figure 6 shows an example of a profile plotted on a curved earth basis, using millimeter paper and a calculated curved baseline. This figure indicates a path with potential reflections from a water surface and also shows the analysis of the potential blocking of the reflected ray under conditions of K = 4/3, K = 2/3 and K = ∞. Figure 6 further illustrates the fact that when antennas are at different elevations with respect to the reflecting surface, the reflection point moves along the path as K varies. It is closest to the end with the lower antenna at K = ∞ (flat earth) and moves toward the center of the path as K decreases. Such paths are customarily analyzed by calculating the reflection point for K = ∞, K = 4/3 and K = 2/3, and investigating the blocking or screening of the reflected ray at each value. The ideal situation is to have adequate clearance at the lowest value of K, yet have the reflective ray blocked over the entire range of K. In some cases this situation can be achieved by appropriate choices of antenna heights so as to utilize terrain blocking, or to move the reflection point from water to a rough surface.

There is one other graphical analysis method, preferred by some engineers, which eliminates the need for either curved baselines or curved ray paths. In this method the path terrain profile is plotted, on any convenient scale, on a flat earth basis in a fashion similar to that shown in Figure 5.

The engineer then analyses the path to determine the potential obstructing points, and for each of these points makes a calculation of the earth curvature at the point, using Equation (1) and also of the desired Fresnel zone clearance, using Equation (4A) or Figure 15. The calculations must be made using the particular set of criteria which are to be applied. For example, the clearance criterion might be $0.6\ F_1$ at K = 1.0.

The sum of the calculated earth curvature and the calculated desired Fresnel zone clearance is added to the elevation of the top of the obstruction, and the point marked on the chart. The microwave beam, plotted as a straight line, must clear this point. Similar calculations are made and similar points marked above each of the potential obstructing points along the path. Tower heights must be determined so that a straight line between the antenna locations clears all of the marked points. Where more than one set of criteria are applied, as in the "heavy route" criteria described



SPACE DIV
CRITERIA 0.3F + K: 2/3 Top-Top
0.6F + K= 1 Top-BTM

Figure 5. Path Profile Example (Flat Earth)

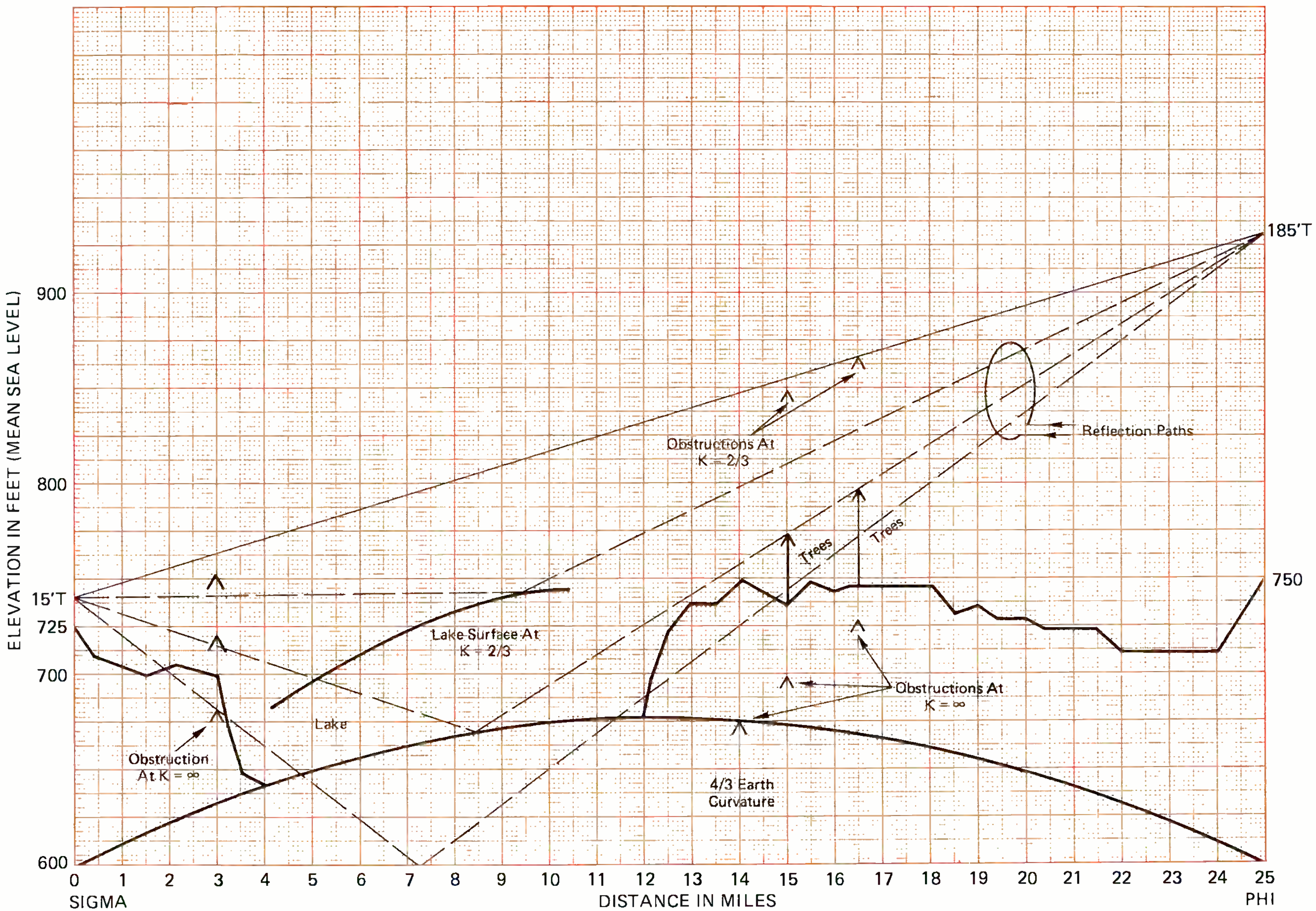


Figure 6. Path Profile Example (Curved Earth)

on Page 51, separate calculations and separate points must be marked for each set, over each obstruction, and again all points must be cleared by the line between antennas.

A skilled transmission engineer can usually reduce the number of points for which calculations need to be made to a relative few. For example, in the path of Figure 5 it is obvious that only five points are potential obstructions.

Though this method at first glance appears to be tedious, in practice it can be done very quickly and easily for most paths. It has the great advantage that any kind of rectilinear graph paper can be used, and that the most convenient scales can be chosen for each axis individually, since there is no interacting effect between the two scales. This method also provides a very useful means of double-checking, at critical points, the clearances as determined by any of the other methods.

4. Reflection Point Calculations

When $K = \infty$ (flat earth condition) there is a very simple relationship between the antenna heights and the distances from the respective ends to the reflection point (in miles). The relationship is:

$$\frac{h_1}{h_1 + h_2} = \frac{d_1}{d_1 + d_2} = \frac{d_1}{D}$$

where h_1 is the elevation of the lower antenna and h_2 the elevation of the higher antenna in feet above the reflecting surface, d_1 the distance in miles from the h_1 end to the reflecting point, and $d_1 + d_2 = D$ the path length in miles. This leads to the following expression:

For $K = \infty$

$$d_1 = nD, \text{ where } n = \frac{h_1}{h_1 + h_2} \qquad (2A)$$

For values of K other than infinity, and for unequal antenna elevations, the geometric relation involves cubic equations whose solution is somewhat cumbersome. However, graphical solutions have been worked out and are found in the literature in several different forms. A nomographic solution for the condition K = 2/3 is given in Figure 7A, and a similar solution for K = 4/3 in Figure 7B.

By the use of these charts, an approximate value of n for each condition can be determined and the corresponding values of d_1 calculated. These values, together with the calculated value for $K = \infty$ can then be used to plot the reflection point range on the path profile, as shown in the example of Figure 6.

Reflective path analysis can be carried out equally well and in somewhat simpler fashion using the flat earth, curved beam approach as depicted in Figure 5. With flat earth profiles, the reflection points for the three significant values of K are calculated and marked along the bottom line of the profile. The appropriate curved beam template for each of these values is then used to trace the beam path and determine whether the paths from antenna to reflective point are clear or obstructed.

The nomograph solution provides accuracy adequate for most work. Where greater accuracy is desired, the following relationships are useful:

For K = 2/3

$$\frac{h_1}{d_1} - d_1 = \frac{h_2}{d_2} - d_2 \qquad (2B)$$

After determining the approximate values of d_1 and d_2 by means of Figure 7A, substitute them in the above equation, together with h_1 and h_2. If the reflection point location is correct, the two sides will be equal. If they are not equal, increase d_1 by a small amount and decrease d_2 by the same amount. If this change causes the inequality to increase, reverse the procedure. Continue by iteration until a value is reached for which the two sides are equal or very closely so.

For K = 4/3

$$\frac{h_1}{d_1} - \frac{d_1}{2} = \frac{h_2}{d_2} - \frac{d_2}{2} \qquad (2C)$$

Use Figure 7B to determine the approximate values of d_1 and d_2 then proceed as described above to determine a more exact value.

The small shaded area in the lower left hand corner of each chart represents conditions for which the path would be below grazing and a true reflection point would not exist.

The parameters X and Y from Figure 7 can also be used to determine the value of K for which the path would be at grazing clearance. For this condition the relation is given by the expression:

$$K = \frac{1}{1.5\,(X + Y + 2\sqrt{XY})}$$

They can also be used to determine the location of the reflection point for the grazing condition from the expression:

$$n = \frac{1}{1 + \sqrt{Y/X}}$$

The range of 0 to 3.0 for the Y parameter and 0 to 2.0 for the X parameter on Figures 7A and 7B will cover most situations where location of the reflection point is needed. For example, on a path of 30 miles, a value of 3.0 for Y would correspond to a height of 2700 feet above the reflecting surface for h_2, and a value of 2.0 for X to a height of 1800 feet above the surface for h_1.

In some special situations it may be necessary to calculate reflection points for paths where one or both of these values are exceeded. Since the charts are rectilinear and all of the lines are straight, it is relatively simple to extrapolate beyond the chart to determine an appropriate value for n. The iterative process described above can be used to improve the accuracy, if desired.

5. Preliminary Map Survey

The preliminary map survey has for its objective the planning of one or more routes which might appear to be possible between the terminal points given, based on available data, and the plotting of profiles which are necessarily preliminary, for all of the indicated paths and alternates determined from the study.

Use of Maps

The prospects of arriving at the best possible route and sites are enhanced by the study of the greatest numbers of alternate possibilities. It is recommended that all available information bearing on the proposed route be first assembled. Large scale maps, such as the aeronautical charts, are suggested for the beginning of the study, and for keeping track of its progress and the site locations.

It is recommended that the terminal locations be first plotted on the aeronautical charts, if these are available, or on a large scale map. At this point it is advisable to make a determination of the maximum, minimum and possible average length of path to be considered. This will depend to a considerable extent on the frequency band to be used, type of service to be assigned to the system and the topography. Having made these determinations, it is suggested arcs be drawn representing these values from the first terminal, and the areas between the arcs be searched for possible first sites. The search at this point will include the topographic maps and any other maps or photography which can be searched for additional information. Having selected one or more possible sites for the first repeater point, preliminary profiles should be drawn. These will furnish a check of the practicability of using the sites selected.

Limitations

The ideal situation obtains when all of the topographic maps for a given path can be assembled on the same scale and spliced together accurately so that a straight line can be drawn between the plotted adjacent sites. An accurate scale in miles is then marked on the line and, following the line over the contours, it is then possible to tabulate the elevations with their appropriate mileages. It is essential that the tabulated elevations and mileages be sufficient to fully describe the profile when plotted. Where the line crosses a hilltop, it is reasonable to assume in the preliminary map study that the ultimate height, unless specifically marked, is half the contour interval higher than the nearest lower contour line.

Cases of Incomplete Mapping

Where some maps in a given path are to a different scale than others, a useful technique is to splice plain white paper to each map where the scale changes and project, by geographic coordinates, a known point or the next adjacent site so that the proper line can be drawn on the available map. The tabulated data should then be plotted on the type of profile desired. For the paths with incomplete topographic survey, it is desirable to plot the profile showing that portion of the elevations which is available. If the path appears to be a good one in the field survey, the profile may be completed from field data. The tentative regular and alternate sites should also be plotted by geographic coordinates on the large scale map, to show their relative position, and to assist in the interference and frequency coordination study.

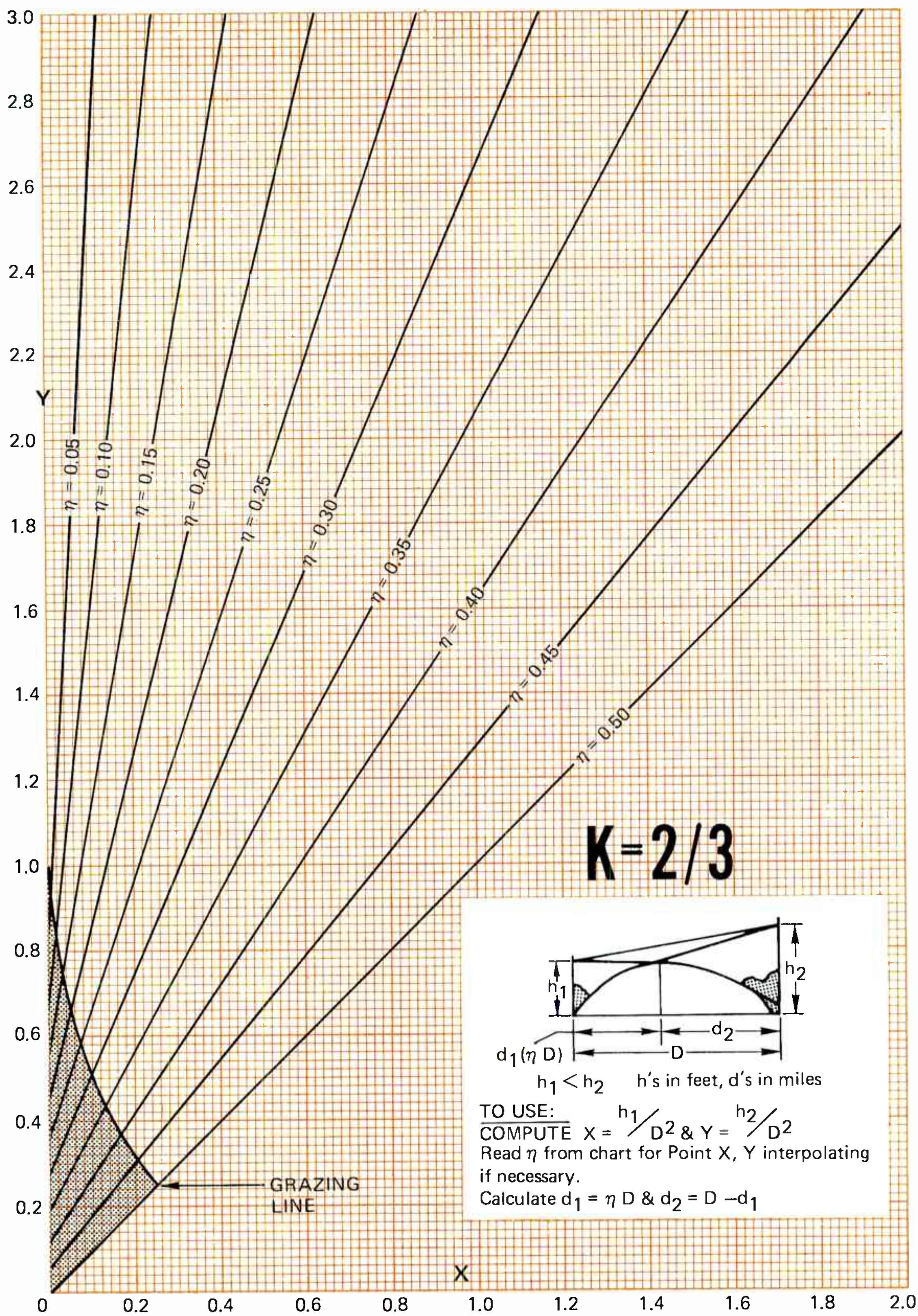


Figure 7A. Point of Reflection On Over-Water Microwave Path

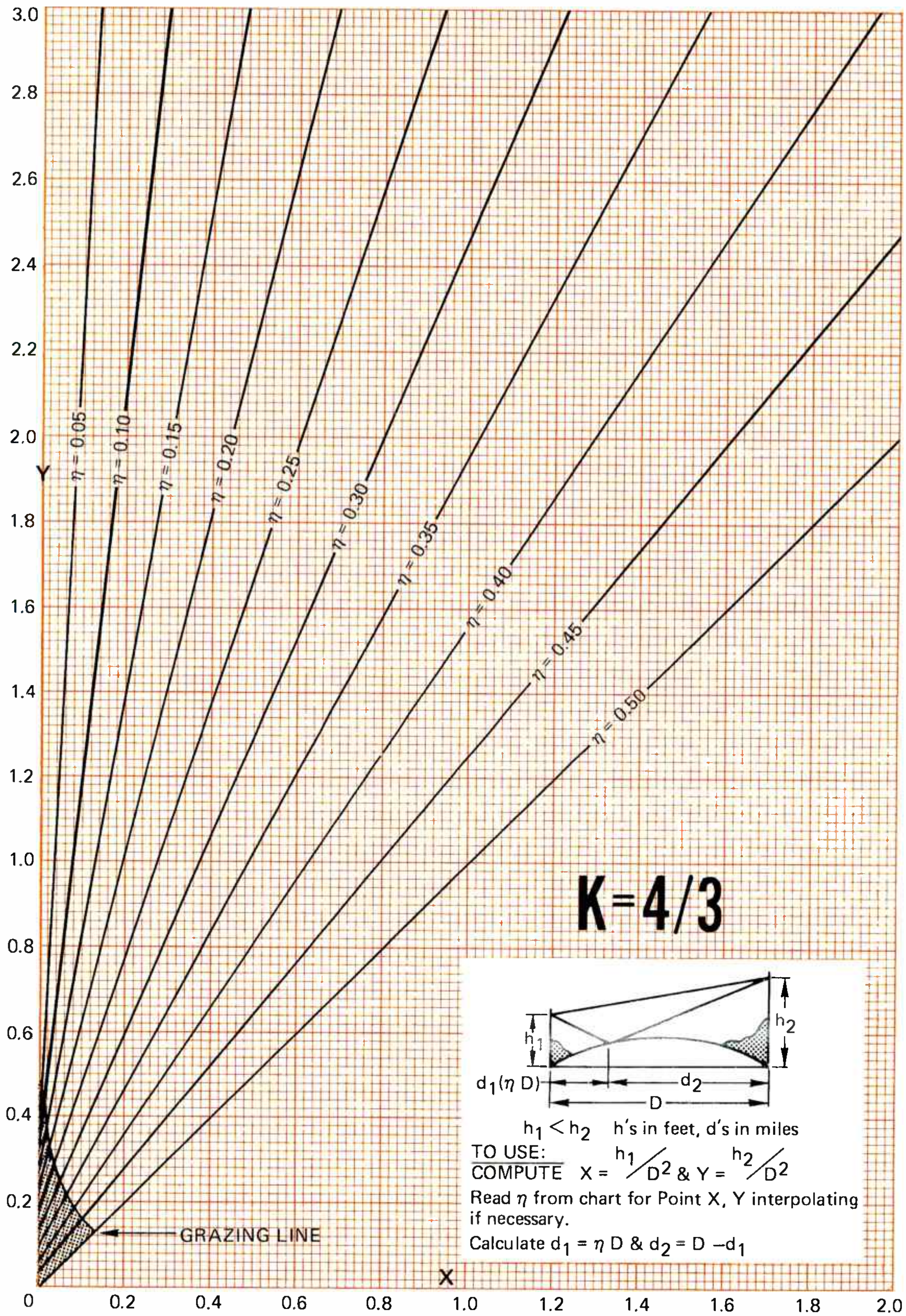


Figure 7B. Point of Reflection On Over-Water Microwave Path

Interference and Coordination
Preliminary Considerations

One very important consideration in engineering microwave systems is the avoidance of interference. Such interference may occur within the system itself, or from external sources. Both of these possibilities can be eliminated or minimized by good site selection when selecting the route, and by consideration of proper antennas and operating frequencies.

Frequency Band Division

Operation of a microwave communications system will generally be within one of the several frequency bands allocated for such services, as listed in Section II. For two-way operation, as is generally required in such systems, each band is divided in half, with the lower frequency half identified as "low band" and the other half as "high band". At any given station all transmitters are normally on frequencies in one band and all receivers in the other. Each path has a "Transmit High" station at one end and a "Transmit Low" station at the other end. This arrangement minimizes near-end crosstalk. In some bands a different arrangement is used.

It should be noted that radars can radiate substantial amounts of power in many of the commonly used frequency bands, even though the fundamental radar frequency is much lower. Suppression filters are available for the radars in most countries, but the burden of making certain they are installed may fall on the communications user.

Intra-System Interference

Interference within the system may be classified as overreach, adjacent section and spur interference. Overreach interference is illustrated in Figure 8, in which only the frequencies in one direction of transmission are represented. The situation in the opposite direction is identical. The problem is to avoid having frequency F_1, transmitted from A, received at D at a substantial level, when there is a fade condition for the F_1 signal received from C. The parameters which may be used to avoid this interference are; (a) a longer overreach path as compared to the direct C—D path, (b) antenna discrimination against the overreach path, and (c) earth blocking in the overreach path. If the latter must be relied on, the overreach path should preferably be blocked by at least 1000 feet when the overreach path is plotted on flat earth profile. Cases of less blocking should be analyzed individually.

Spur or junction interference, and adjacent section interference for systems using the so-called two-frequency plan, are illustrated in Figure 9. In all of these cases, far-end crosstalk is of concern because it is quantitatively dependent upon the discrimination of the antennas at the junction, both transmitting and receiving. Near-end crosstalk at the junction may become important when there are several channels on the main route, as frequency translations can result in interference to an adjacent receiver. The mechanisms of this type of interference are rather complex and need not be delineated at this point. The criterion in both near-end and far-end crosstalk is the signal-to-interference ratio, which is a function of antenna discrimination. For example, a periscope antenna arrangement would not be used at a junction or on main routes using the two-frequency plan because it typically has a relatively low front-to-back ratio, and may have some odd side lobes. On routes with only one radio channel, or two channels in frequency diversity arrangement with maximum inter-channel frequency spacing, periscopes may often be used with high towers for economic reasons.

External Interference

External interference cases were discussed superficially in (1) above, but require further clarification as to methods of approach and specific values. Radars typically radiate pulse energy, often in a 360 degree arc and at a very high energy level. Quite often the second or third harmonic, if unfiltered, can have an effective radiated pulse power in the order of +60 dBm (1 KW). As pointed out above, it is possible to determine whether harmonic filters are installed, and if not, such filters can usually be obtained. Experience in the United States indicates that the agency operating the radar will usually install the filter if the necessity is established. There is another precaution recommended in the case of radar, in that, regardless of frequency radiated, the input circuit of the microwave system should not see large amounts of peak pulse power. Input preselection filters for the microwave system vary in the amount of discrimination outside the pass band, but it is usually very high. Nevertheless it is possible to paralyze the input converter in extreme cases. Therefore it is recommended that the path from the radar to all

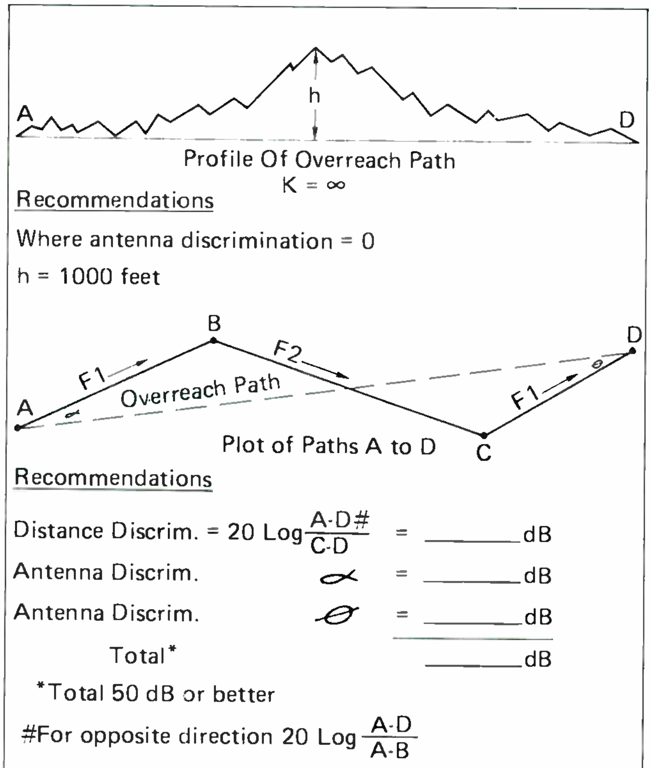


Figure 8. Overreach Interference Criteria

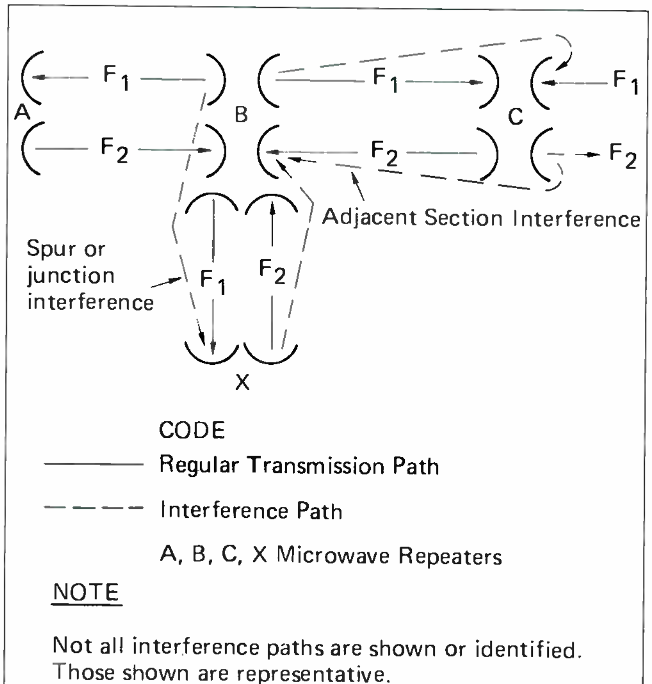


Figure 9. Adjacent Section and Junction or Spur Interference

microwave receivers within 50 miles be profiled at K = ∞ and, if possible, substantial earth blocking be arranged. In the unblocked case, it is recommended the path between radar and microwave receiver be in excess of ten miles, and the receiving antenna discrimination be at least 30 dB. Figure 10 illustrates the radar case.

In the case of paralleling or intersecting microwave systems the transmitter output power of the two systems is usually somewhat comparable, but where differences exist they must be taken into account. Additional criteria for interference coordination involve the parameters of distance, antenna discrimination, receiver sensitivity and selectivity. Interference considerations are two-way in this case, as the new system must not interfere with the existing system. The paralleling system problem is illustrated in Figure 11. Profiles made for possible earth blocking should be prepared for K = ∞.

6. The Field Survey

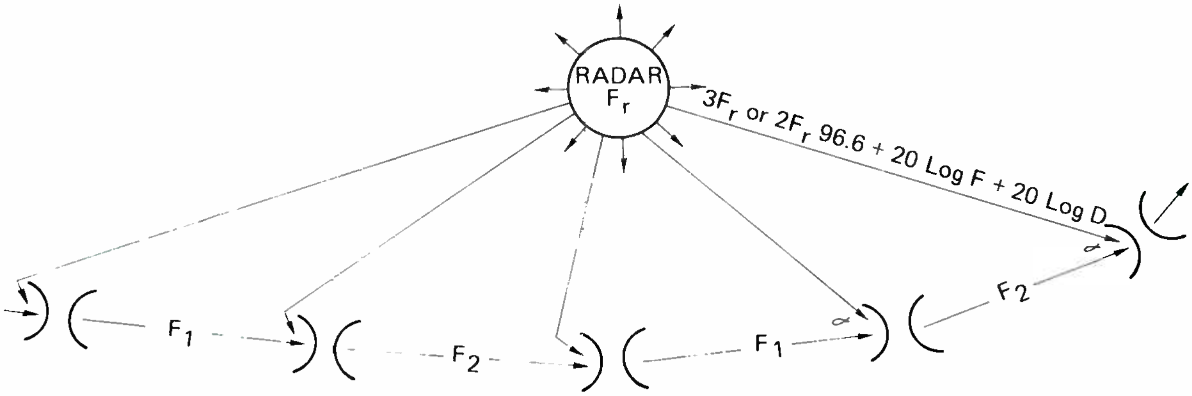
The field survey is much more than the phrase implies. Actual altimeter measurements, judgments of the actual terrain along each path and data concerning obstructions are recorded. The existence of paralleling or intersecting foreign systems and interference possibilities is indicated, and the various data concerning regular and alternate sites are obtained. The preliminary profiles made from the map study become a tool for the field investigation, and these are supplemented, corrected, or actually replaced as a consequence of the factual data obtained. In the absence of path tests, the information brought in from the field survey constitutes almost all of the factual data about the route, from which judgments can be made which will determine the service performance of the system when installed.

Instrumentation

The following instrumentation is recommended, although all items may not be needed in all cases

A good pair of binoculars, 7 x 35 or 8 x 50, with coated lens. (Larger diameter magnification does not ordinarily improve the results).

Two precision altimeters of the proper range for the terrain to be measured.



Computation for Unfiltered Radar

E.R.P.P. of 2nd or 3rd Harmonic	+60
Gain Of Receiving Antenna	G
	60+G
Path Loss Rad. 96.6 + 20 Log F +20 Log D	P
Antenna Discriminator at α Degrees	D
Filter & Waveguide Losses	L
Radar Harmonic Input (60+G-P-D-L)	R
Computed Microwave Signal Input	S
Signal-to-Interference Ratio (S—R)	S/I dB

Requirements

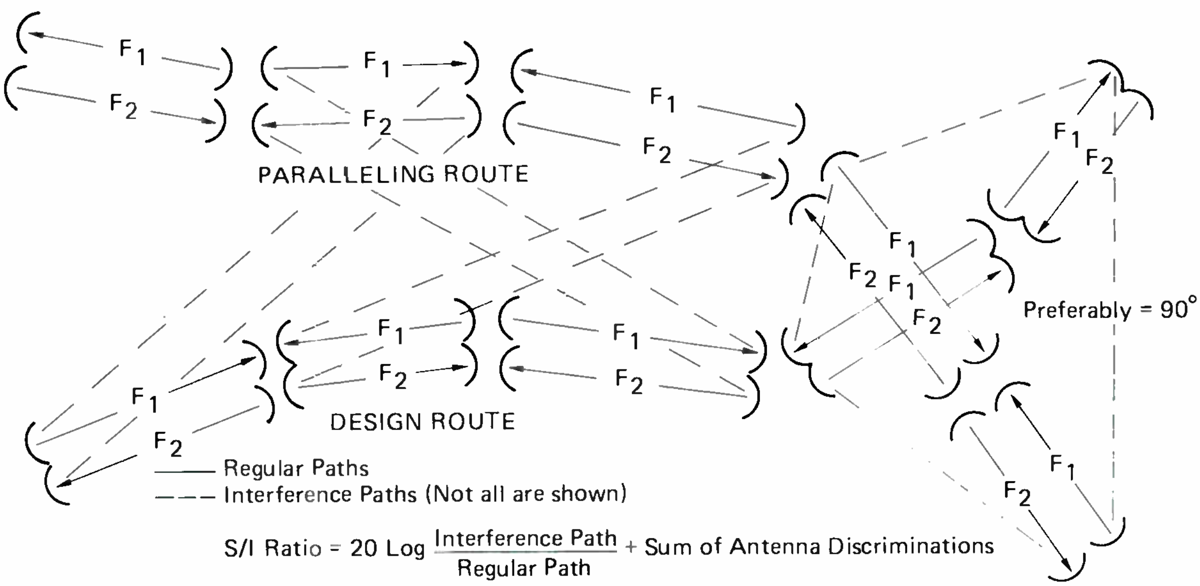
Harmonic Energy in Band — Message	30 dB
Television	45 dB

Fundamental-Ten Miles + 30 dB Antenna Discrimination

NOTE

For on-site microwave terminals at radar locations, or locations less than ten miles from a high powered azimuth operated radar, obstruction blocking is desirable.

Figure 10. The Radar Interference Case



——— Regular Paths
– – – Interference Paths (Not all are shown)

$$\text{S/I Ratio} = 20 \text{ Log } \frac{\text{Interference Path}}{\text{Regular Path}} + \text{Sum of Antenna Discriminations}$$

S/I Ratio Recommendation 66 dB for Co-channel

NOTE

Cross polarization advantage at angles close to 90° is zero and may be negative in some combinations.

Figure 11. Interference Coordination of Paralleling Systems

Both may be of the portable type, or one may be of the portable type and one the recording type. Care should be exercised in considering these instruments. Each precision altimeter should be accompanied by its own correction chart and an accurate thermometer. Suitable instruments may be obtained from one of the following:

American Paulin System
1524 Flower Street
Los Angeles, Calif. 90015

Wallace & Tiernan Inc.
25 Main Street
Belleville, N.J. 07109

A good hand level, and a pocket compass.

A good theodolite is useful in some instances to determine that an elevation measurement is on path. Visibility of both adjacent sites is necessary for success in such determination.

Two-way portable radio (VHF) capable of shadow area reception at 30 miles.

In rugged country, a precision mirror of 6 inch to one foot dimension is a good item for "flashing" paths. Special signaling mirrors as developed for military use are particularly good.

U.S.G.S. book or books showing established bench marks in area.

County highway maps (in U.S.A.) for the area. Scale 1/2 inch = 1 mile.

Pertinent topographic maps, marked for the preliminary profiles.

Things to be Avoided

There are several situations which should be avoided if practicable, particularly in the case of high density systems with multiple RF channels. These are as follows:

Over-water paths and paths over low, flat terrain. Where they cannot be avoided, the high-low technique to place the reflection point over rough terrain should be considered. A very low antenna in combination with some intermediate terrain blocking may be possible. Where the antenna is actually on the flat terrain, a very low antenna can be selective, particularly in the lower frequency bands. Refer to the discussion under PROPAGATION for further details.

Sites on high mountain ridges with low, flat terrain between. In high density systems, serious delay distortion may result. If it cannot be avoided by reselection or terrain blocking, it may be necessary to locate one site on the intermediate low flats.

Sites near high power radars.

Crossing of foreign system routes of similar frequencies at small angles or with near repeater stations.

Obstructions near the line-of-sight which may reflect energy from the transmitters into the receiving antennas. Paths along a city street between buildings are very bad.

Much can be done in the field search to eliminate propagation hazards which may affect service.

Methods of Operation

For running the profiles in the field (Altimeter surveys), it is desirable to have maps with accurate road information and having a scale of about 1/2 inch to the mile. In the U.S.A. the county highway maps serve this purpose very well. The maps for a particular radio path should be spliced together accurately, the radio sites specifically located and a straight line drawn between the sites. This line is then scaled in miles. Consider when doing so that the maps are sometime stretched in printing. This can be checked by using a scale across a number of road intersections to see if the mile points come up accurately. The elevations shown at road intersections should not be taken as accurate unless they agree with established U.S.G.S. bench marks which have not been moved. The maps should be posted with the location and elevation of established U.S.G.S. bench marks. They are then ready to be used as a guide in securing information for the profile.

The most accurate altimeter method of running levels for the profile is known as the two-altimeter method. The process involves placing both altimeters at the nearest bench mark and calibrating them exactly alike. The altimeters are based on the aneroid principle, so their readings will vary, even when on a fixed bench mark. The work should be done during stable weather conditions, and in the period from at least one hour after sunrise to one hour before sunset. One altimeter remains at the bench mark throughout the measuring period. If this is a manually read instrument, readings of temperature and the altimeter should be taken each five minutes until the roving altimeter returns. The readings of the two instruments should then be compared. If the readings differ by as much as five feet after temperature stabilization, the survey should be repeated after recalibration of the instruments. The difference may be due to the higher temperature in the car in which the roving altimeter is carried.

If one recording instrument is used, it can be placed at the bench mark and thereby save manpower. The principle of operation is the same as with two manually read instruments.

The roving altimeter is used to measure the elevations along the path. A record is made of the mileage at the measuring point, the temperature, the time, the trees and other obstructions, and any terrain features of interest in preparing the profile. The roving altimeter should be used on the portions of the path which are reasonably close to the bench mark. As many bench marks should be used as are necessary to accurately survey all of the path on this basis. The final measurement at the bench mark on each measuring trip provides a reasonable check on the data. The crew then moves to the next bench mark and proceeds in the same way. If there is only one bench mark near a path, it is desirable to create secondary bench marks, using the same method as for measuring on the path.

Corrections are made for temperature and approximate relative humidity, using the instructions in the handbooks furnished with the instruments. The computed elevations for the points measured are then determined. There should be enough points measured to fully describe the profile of the path, which is then prepared based on the method selected for the office profiles.

Another method which has been used to obtain information for the profile involves aerial photography over the path by stereopticon techniques. The relative elevations are then determined by stereopticon procedures. The accuracy is good, but the cost is relatively higher than the altimeter survey method. In a case involving a fairly large number of profiles, and where new forces might be required to handle the altimeter survey, this method might be investigated. It should be borne in mind however, that the field altimeter party brings in a great deal more information than elevation data, and such information is necessary in any event. A third method of profiling involves flying over the path with equipment which measures clearance (plane above terrain) using the radar principle. This is also relatively expensive and has certain limitations.

In mountainous country where all of the intermediate terrain in a path is rough and timbered, a profile is of little value if adequate clearance can be determined by other means. This can often be done by "flashing" the path. It involves two parties, preferably equipped with two-way VF radio for communication. Lacking the radio, it is often possible to coordinate by time. The party carrying the mirror should preferably be at the site opposite the direction from which the sun is shining, or two mirrors may be used; one at each site. The mirror is to reflect a beam of light toward the far site. The aim must be very good. The stability of the mirror when held on the beam by hand will be such that the flashing effect is automatic. To be effective, it is necessary for the flashing party to establish the exact direction to the opposite site, and to concentrate the flash along that line. One way of doing this is to drive stakes to mark the exact line and then, using the mirror, focus the sun's rays along the line of stakes, gradually raising the beam until it is level with the distant horizon. This process should be repeated at frequent intervals until the far party is known to have seen and identified the flashes. At the far site the flashes will appear very large. A transit set up at the far site can be very useful, both for observing the flashes and obtaining accurate path bearings. The terrain clearance can be roughly established visually, but not precisely. Any known ridges which appear close to the line-of-sight should be checked for elevation at the critical point and adequate clearance established by computation.

In checking path clearance by mirror flashing, or by any optical or visual methods, account should be taken of the fact that light rays have a slightly different curvature than do radio rays. A nominal value of K = 7/6 is usually taken as the "standard" for light. Figure 4 includes a curve for this value of K which can be used for evaluating clearances determined by optical methods.

In most rugged terrain, and particularly in high mountain areas, the almost continuous and shifting wind currents tend to produce a thoroughly mixed atmosphere which does not support atmospheric multipath fading of great depth. In some cases it is possible to achieve high reliability without diversity, or to use somewhat longer radio paths than in lower terrain, without endangering the fade margins for which the equipment was designed. However it should be noted that exceptions exist, and that in some cases and some mountain areas atmospheric effects can cause deep atmospheric fading.

Records and Reports

Information to be reported about the sites was covered in subsection "B. Sites" as data to be recorded. For the paths and alternate paths, complete profiles carefully plotted with obstructions and estimated antenna height requirements should be prepared, together with all facts or determinations with reference to possible interference. In addition, all pertinent supporting data, altimeter readings, maps and bench mark information should be furnished.

Final Profiles

The final profiles provide the basis for the microwave system engineering, including the final selection of paths, determination of final antenna elevations, selection of antenna sizes and configurations, and computation of the received signal strengths, fade margins and the system noise to be expected. The final profiles should reflect the final decisions on route, sites and antenna elevations.

Path Coordinates, Azimuths and Distances

After final selection of the precise locations for the tower or antennas, the latitude and longitude of each location should be very carefully determined. For systems under FCC jurisdiction, the rules require that the coordinates of the antenna or final radiating element be determined to the nearest second. This requires very careful scaling from the best available maps.

Path azimuths should be determined to the nearest tenth of a degree of arc and path distances to the nearest tenth of a kilometer. In order to achieve the required accuracies it is necessary to calculate the azimuths and the path distance from the coordinates of latitude and longitude at the two ends of the path. Because of the convergence of meridians, the azimuth at the two ends of the path will in general not be precisely 180° apart.

Two methods of calculation are in common use. The most common one is to use a computer to calculate azimuths and distances by solving the great circle spherical triangle comprised of the two end points and the north pole. The second, known as the "inverse position method", is adapted from a calculation method described in "Special Publication No. 8 — Formulas and Tables for the Computation of Geodetic Position", published by the Coast and Geodetic Survey. This method includes the use of tables which take into account the oblateness of the earth. Because of this, the inverse position method provides somewhat more accurate results than the great circle method, unless the latter includes a correction for non-sphericity.

Although either method provides results which are sufficiently accurate from a technical viewpoint, a recently inaugurated program by one FCC bureau (the Common Carrier bureau) uses the inverse position method, and it is therefore desirable that this method be used in calculations for applications filed with that bureau. Included in the Appendix to this publication is a memorandum giving a method of making azimuth and distance calculations by the inverse position method, suitable for hand calculations using a six-place logarithm table. The method detailed uses essentially the same format as that used in the FCC computer program and should give identical results. The memorandum includes sufficient extracted data from Special Publication No. 8 to be self-sufficient and not require use of the latter.

F. Path Tests

On line-of-sight microwave systems, path tests, when used, are made primarily for the analysis of problems related to the reflective characteristics of the path and, secondarily, to determine or provide a cross-check on the path clearance. They are most often made on routes with very heavy communications requirements or potential. Such routes typically use heavy horn reflector antennas which require changes in tower design when antenna elevations must be changed. To make such changes after the route is placed in service would be costly. Path testing prior to construction of the system helps to avoid such a possibility. A further consideration affecting the decision whether or not path tests are justified, is that large capacity microwave systems utilizing multiple RF channels in parallel, are considerably more susceptible to the effects of reflections than are systems using only a single working channel.

This type of test is a short term process, normally performed during daytime periods when atmospheric conditions are essentially normal and stable. It should not be confused with the type of path testing in which long term recordings are made of the signal strength along a path. Such tests are common in tropo-scatter or other beyond line-of-sight systems, but are rarely used in line-of-sight work. Long term fading characteristics, range of K variations to be encountered, and other things of that nature will not be normally determined in this line-of-sight type of path test.

Brief Description of the Tests

In the most effective form, the tests involve transmitting an unmodulated RF carrier over the path between adjacent sites and measuring the received power at various combinations of antenna elevations. The receive power measurements are then converted to path loss measurements, since the transmitted power and antenna gains are known. A test series made with both antennas moved in the same direction at the same rate, or with one antenna fixed and the other moving, is called a "height-loss run". The results are plotted with path loss as ordinate against antenna elevations as abscissa.

Information they Provide

By properly planning the height-loss runs, and by analysis of the results using a very accurate profile, it is possible to determine the following information about the path:

The location of reflection points in the path.

The value of K at the time of measurement.

The reflection coefficient.

The antenna elevations where free space loss is just obtained.

The best antenna elevations to minimize ground reflection fading.

Path tests of this type require the utmost precision in calibration of the equipment and in measuring the antenna heights.

Effects of K Variations

Accurate results can be obtained only when the value of K is stable. This can be expected over most terrain only during fair weather and in the daytime hours at least 1 to 2 hours after sunrise and before sunset. Changes in K during the height-loss runs would be reflected in incorrect values for all of the desired results listed above.

Cost Considerations

Path tests involve costly equipment and considerable manpower. For this reason, where single channel routes and light loading are involved, economics dictates reliance on the field accuracy of other methods. Accurate profiling and good judgment will go a long way toward establishing a good working system.

IV. OVERALL SYSTEM DESIGN

A. Purpose of Section

Given the fundamental plans and study results, and the profiles, data and other information obtained from the field investigation, the next order of procedure concerns the transmission engineering plans, calculations and design specifications. Some of this work, such as planning, is actually accomplished prior to the route selection work. After the field investigation is finished, the final calculations concerning antennas, interference, system and channel noise, distortion and propagation reliability are made. If the final sites have not been selected from the field data at this point, this must be done prior to much of the other work. Also at this point it may be important that information about the final choice of sites be made available to those responsible for obtaining the necessary property, as failure to obtain one site may affect some of the other selections.

1. Final Objective for a System

The final objective for any microwave system is that it provide the best distortion-free and interference-free service continuity for the type of service to be assigned, and within the framework of the available economics.

Overall reliability or service continuity involves not only equipment failure rates and power failures, but also the propagation performance of the individual paths. This involves antenna sizes and elevations, frequency or space separations in diversity systems, path lengths and frequency-attenuation relationships. It also includes fading margins which, in addition to path parameters, are affected by noise figure, transmitter power, and attenuation of waveguide and filter arrangements.

Distortion may occur in the radio path, but more often it occurs due to poor return loss of amplifier components, waveguide filters and antennas. Also the characteristics of switching devices and/or combiners are involved.

System noise is affected by the same things which, in addition to interference, can have an adverse affect on overall system performance.

2. Order of Presentation

The presentation within this section is designed to provide an orderly discussion which follows, insofar as practicable, the design-related priority for consideration of each topic. For example, the first item listed below should be considered, and any problems resolved before proceeding to the next item, as any changes resulting from it could otherwise cause considerable duplication of effort. The order of priority is as follows:

Interference and frequency coordination.

Propagation.

Noise and noise sources.

Distortion.

Equipment.

The above order of priority, and in fact most of the discussions in this section, are specifically applicable to high density systems involving multiple parallel RF channels over each path. Such systems often grow to the point where all, or nearly all, of the available channels in a full band are in use over every path of the system. In such situations the interference potential, both intra-system and with other systems occupying the same band, is very large.

Microwave systems involving only one working RF channel per path (or one working and one protection channel) have the same kinds of problems, but in most cases they have a much wider range of practical solutions. The system designer has a good deal more leeway in engineering such systems, even when they require the same degree of reliability and performance as the more complicated systems.

B. Interference and Frequency Coordination

1. How Interference Occurs

Radio system interference causing degraded transmission may be introduced through antennas, waveguides, cabling, radiation or by spurious products produced in the radio equipment itself. Interference introduced into the cabling or in the equipment can be prevented by good installation practices, including proper separation of high and low level cabling, proper grounding practices, shielding where necessary, and good equipment design and assembly. Interference introduced by coupling between waveguides in the same station is usually produced by radiation from waveguide and filter flanges which are not properly tightened, or which have been damaged and cannot be mated properly. There are cases on record where interferences of more than marginal character have existed for long periods, but were undiscovered until a particular critical service was assigned to the system.

Because a radio system depends upon the atmospheric medium for transmission, it is subject to interference from systems using the same medium. This includes not only other radio, radar and other devices, but also other parts of the same system. The interferences are broadly classified as external interference and self-interference. These were briefly discussed in Section III and illustrated in Figures 8 and 11, which provide some measurement criteria for those concerned with selecting the route. In this section it is the intent to discuss the subject more analytically, and thereby provide a somewhat more thorough understanding of the mechanisms involved. It is expected that this will be of assistance in solving or judging marginal cases.

Interference Mechanisms

The evaluation of interference can be complex and difficult because of the nature of the systems involved, and the complex nature of the signals. The mechanisms are varied. In the simpler cases they may be direct interference into the radio receiver. In other cases they may be spurious products or combinations of products which arrive at the receive input and produce a net resultant interference into the receiver intermediate frequency section. The latter may be frequency translations resulting from sum and difference products within the same system. In still other cases the receiver may see an identical signal to the regular signal, but which arrives later or earlier than the regular signal.

Effect on the System

Interference can produce beats or noise products in a radio receiver which have detrimental effects depending on the frequency, deviation, channel separation and linearity of the receiving medium, as well as the nature of the interfering signal. In some cases they combine with other frequencies in the system, including carrier sum and difference frequencies, to produce interference in a third radio channel. These products may hold up automatic gain control during critical fading periods, with serious effect on system noise. In almost all cases, noise in the baseband channels is an end product of the interference.

2. General Classifications

General classifications have been given broadly as self interference and external interference. Each of these broad categories can be broken down into a number of different types for discussion purposes.

Self Interference

Self interference has been classified in terms of overreach interference as illustrated in Figure 8, adjacent section interference and junction interference, with the latter two cases illustrated in Figure 9. Adjacent section interference also includes what is sometimes called same section interference. The only difference is the matter of whether the discrimination is principally the front-to-back ratio of the receiving antenna or the transmitting antenna. These cases were previously discussed because they involve the sites and radio paths.

There are also possible cases of self interference within the stations themselves. One of these is the result of carrier leak from a common carrier supply. In the C.C.I.R. plan for 6 GHz common carrier systems, eight channels of radio in each direction are recommended, each with a capacity of 1800 or more message channels. The successful operation of this system requires very tight limits on frequency tolerance. A common carrier supply is one way of assuring that all of the precise frequency relationships are maintained within the system. If, on the other hand, through radiating waveguide flanges and/or leaking filter connections, the high level carrier supply is introduced into the wrong input circuit, a translated signal as well as the regular signal can be introduced into the channel, with certain degradation resulting. Also, because of the close relationships between the channels, there may be a tone translation which results in interference to an adjacent channel and even to a third channel.

External Interference

The control of external interference depends in part on coordination, control, and sometimes compromise on radio channel assignments through, in most cases, direct negotiation. The frequency regulating body, such as the Federal Communications Commission in the United States, assigns the frequency spectrum by services and licenses specific channels within the designated bands. In most cases the public service communications bands are assigned in accordance with the recommendations of the Consultative Committee On International Radio (C.C.I.R.); however, certain priorities and other special requirements are recognized. Thus in the 4 GHz band, where transcontinental networks were established in the United States prior to C.C.I.R. recommendations, a different plan is used requiring a wider band. In the 2 GHz band and the bands between 6 and 11 GHz, the allocations used in the United States differ considerably from those used by other nations, and channel assignments are accordingly different. For all plans, adjacent channel discrimination within the immediate section is aided by cross-polarizing adjacent channels. For such channels transmitted and received by the same antennas, the discrimination is maximum; up to 20 dB per antenna. Cross-polarization discrimination varies widely at angles other than zero degrees between the interference signal and the desired signal. The pattern of variation is quite different for different types of antennas and in many cases the discrimination is actually negative when the angle is 90 degrees. In dealing with cross-polarized interfering signals, the calculations can be meaningless unless the cross-polarization discrimination characteristic of the receiving antenna is known. For example, there is no generalized discrimination curve for parabolic reflector antennas for any given frequency and, with a different feed, both the discrimination curves and the cross-polarization curves will differ. For a particular case it is recommended the manufacturer be requested to furnish this information about the antenna being used.

With antenna-reflector combinations (periscope antennas), angular discrimination up to about ±2 degrees will generally be about as good as the dish itself for a given polarization. For larger angles and cross-polarized interference signals, preconstruction predictions would probably not be valid, due to the odd side lobe patterns from the reflector. Dual-polarized periscope antennas have been used with success in specialized cases. Generally speaking, dual-polarized dishes are not as efficient as single polarized.

The criteria established for external interference depend upon many factors, and to some extent are arbitrary. They are typically expressed in one of two ways; (1) as an absolute value of the interfering signal power not to be exceeded, or not to be exceeded for more than some specified percentage of time, at the input to the interfered-with receiver or, (2) as a value of S/I ratio where S is the power of the desired signal at the receiver input and I is the power of the undesired signal. For co-channel operation (desired and undesired signals on the same frequency) typical objectives for the first type of criterion range from as low as −125 dBm to perhaps −100 dBm, and typical values for the second type of criterion from approximately 60 dB to 95 dB, depending upon a number of factors.

Figures 10 and 11 illustrate the method of calculating S/I ratios for external interferences. Typical criteria for co-channel objectives are shown on the figures and are applicable where desired and undesired signals are closely held to a small frequency difference. For slightly larger frequency differences, where the interfering signal falls within the first order sidebands of a heavily loaded system, the criterion for satisfactory operation may become more stringent by as much as 25 to 30 dB.

For off-channel interference (between adjacent channels for example) the criteria will be substantially affected by the selectivity characteristics of the particular receiver involved. Most manufacturers can provide curves or other data showing the allowable levels for interfering signals as a function of the frequency separation between desired and undesired signals. Tone translations can occur when a frequency received in one channel, well out from the carrier frequency, is modulated in a heterodyne repeater and radiated as a new tone in the next adjacent channel.

3. Effect of Interference on Different Types of Signals

The effect of interference varies, not only with the nature of the interfering signal, but also with the nature of the desired signal.

Voice, Data, Television

Voice channel interference to voice channel systems usually results in either intelligible cross-talk or burble, which is similar to what one might expect for voice frequency systems. All types of interference to data systems are generally capable of producing data errors. Interference to television systems is usually readily detected and analyzed by observing the video monitor and "A" oscilloscope.

Interference from voice systems or data systems produce typical overall patterns on the monitor which are easily recognized. Pulse interference, such as radar, produces scattered dashes on the monitor, which will be white on a black background and black on a white background. On IF heterodyne microwave systems the point of interference may be many miles away from the point where monitoring equipment is normally available. The known history of the system, or visual inspection of the route, may suggest where the interference is originating. In the final analysis, the system can be looped back at IF at different points, provided there is a return channel available, and monitored at the transmitting point. For one-way only channels, a portable FM transmitter and a video signal generator can be used at successive stations while the receiving terminal monitors. The standard window signal is best for interference analysis.

Tone Versus Pulse Interference

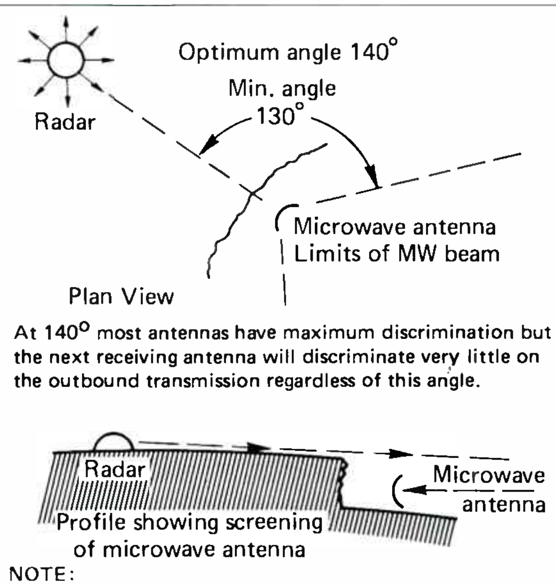
Tone or beat interference is detrimental in the frequency region in which it occurs, not only because of the direct interference with the desired signal, but also because of the background noise it produces, even at levels below the signal level. Pulse interference, such as radar signals, may not be noticed in individual message channels when the peak pulse power received is below the message channel level, because of the pulse spectrum distribution; however, when the interference peak pulse power is above or equal to the message level, the interference becomes intolerable. The breakover point for data is usually somewhat lower, depending upon the bandwidth and bit rate of the data signal. For television, the breakover point is approximately 15 dB below the equal level point. To provide fading margin, the approximate voice and television limits are −30dB and −45dB respectively and, for data, except for voice channel data, it is recommended the latter limit of −45dB be used.

4. Calculations for Interference Effect

Figures 10 and 11 illustrate the off-path radar and the parallel systems case respectively, and indicate the calculations that are appropriate to these exposures. There are, however, cases where it is necessary to terminate a microwave system on a radar site. In such cases, the microwave receiver is quite vulnerable to the radar transmitter signals unless special precautions are taken. The radar should be filtered for harmonics and spurious radiation that would otherwise interfere with the microwave receiver. Also, because of the peak pulse power of the radar, it is advisable to shield the microwave antennas from the radar beam if practicable. Figure 12 illustrates the case and provides some objectives. It will be noted that the receiving antenna at the far end of the microwave path will have practically zero discrimination against the radar signal. Obviously, this violates the rule stated in ROUTE AND SITE SELECTION in connection with off-path radars. At this point it is necessary to look at out-of-band losses of antennas, low frequency cut-off of waveguide, and preselection filter characteristics, for the fundamental. The harmonic filters can usually be counted on for about 60 dB rejection. If the harmonic is normally in the region of +60 dBm and the microwave transmitter has a one watt output, the in-band S/I ratio should be in the region of 73 dB for a 6 GHz system using a 10 foot transmitting antenna, the gain of which would be about 43 dB.

5. Satellite System Interference

Certain frequency bands are used on a shared basis both by terrestrial radio relay systems and



NOTE:
1. Alternately for filtered radar in L or S band use 6 or 11 GHz and long waveguide. Waveguide for these frequencies will cut off well above the radar fundamental. For X band or F, D radars research output frequency pattern.
2. Select receiving frequency not in harmonic spectrum of radar.

Figure 12. A Method of Screening Microwave Antennas When Required on a Radar Site

earth-satellite systems. Because of the widely differing parameters of the two systems, the potential for interference is high when earth stations and terrestrial radio relay systems in the same bands are located anywhere near to each other. The potential interference radius around an earth station may range as high as several hundred miles, depending on terrain factors and relative orientation.

The present international commercial satellite service uses the 5.925—6.425 GHz band for transmission from earth to satellite, and the 3.7—4.2 GHz band for transmission from satellite to earth. Because of the enormous (by line-of-sight standards) effective radiated power of the up-link transmission, the earth-station transmitters can interfere with terrestrial system receivers in the 6 GHz band to distances which can extend far beyond the horizon. The earth station receivers are much more sensitive than those of a terrestrial relay system and they can be interfered with by even the low-powered transmitters of terrestrial 4 GHz systems well beyond the horizon.

Although there is little need for sophistication in the calculations of microwave path losses within the horizon, the matter of path losses at ranges beyond the horizon is much more complicated. As a result, the problem of coordination between earth-satellite systems and terrestrial systems sharing the same bands is a rather complex consideration.

Within the United States, Part 25 of the FCC rules requires the earth station licensees or applicants to calculate coordination contours around each earth station, and also to calculate interference levels for all stations within these areas. After an earth station has been established, any user operating under Part 21 of the rules and desiring to establish a microwave station in the 6, 4 or 2 GHz bands must, if the proposed station falls within the coordination contours of an earth station, demonstrate to the Commission by suitable calculation, that the proposed station will neither interfere with (in the 4 GHz band), or be interfered with by (in the 6 GHz band), the earth station. The procedure for doing this may be quite complicated as it involves calculation of expected path losses on beyond-the-horizon paths.

The internationally established criteria, at the present time (1969), for allowable interferences from earth stations into terrestrial radio relay receivers are that the interference in any voice channel shall not exceed —63 dBm0p (500 pWp0) for more than 20% of any month, and —43 dBm0p (50,000 pWp0) for more than 0.005% of any month. It is assumed that there are no more than two such sources of interference in any system.

The corresponding criteria for allowable interferences from terrestrial radio relay transmitters into earth station receivers are that the interference in any voice channel shall not exceed —66 dBm0p (250 pWp0) for more than 20% of any month and —43 dBm0p (50,000 pWp0) for more than 0.01% of any month. It is assumed that there are no more than four sources contributing to the 20% value, and no more than three to the 0.01% value.

Beyond-the-horizon operation in the form of tropo scatter has been in use for many years, and much work has been done on methods of calculating path loss. This work is, however, primarily concerned with determining the average value and the highest value of transmission loss to be expected, with little or no concern about those periods when transmission loss drops to lower than normal values. But, in evaluating interference potentials, what is of concern is precisely the lowest value of transmission loss over the undesired path, since this results in the highest level of interference. The problem is to calculate how strong the interfering signal will be under supernormal propagation conditions, and for what percentage of the time such strong interference signal conditions can be expected to occur.

This is a much more difficult task indeed, and is complicated by the fact that operational data has not been oriented in the past toward shedding any light on this situation.

In recent years considerable work has been done on the problem, by a number of organizations. Much of the work is based on NBS Technical Note 101, Revised, January 1, 1967.

A COMSAT technical publication "Transmission Loss Calculations For Interference Evaluation in the 4 Ghz and 6 Ghz Shared Frequency Bands", dated September 15, 1967, provides a somewhat simplified method based on the NBS work (the latter is very lengthy and somewhat abstruse so that it is not easy to follow or apply).

A considerable number of CCIR documents deal with the coordination problem, for example:

CCIR Oslo, 1966 Vol IV Recommendations 355 through 359, Report 382 and 393.

CCIR Oslo, 1966 Vol II Reports, 243, 244 and 339.

Part 25 of the FCC rules also outlines a coordination calculation procedure.

One additional source of potential interference has begun to receive attention, but is so far not very amenable to calculations. That is the problem of scattered interference caused by rainstorms which happen to fall within common volumes illuminated by an earth station antenna and neighboring radio relay station antennas. Indications are that this can, in some cases, be quite serious. This problem is discussed in the above referenced CCIR Report 339, and will be treated in greater detail in a revised version of that report intended for inclusion in the next CCIR books. The subject has also been treated in several places in the technical literature.

Another coordination procedure which will probably become mandatory in the very near future is that of calculating or determining whether or not the extended beam of a terrestrial microwave antenna operating in any of the shared bands will intersect, within $\pm 2^{\circ}$, any portion of space which could be occupied by a satellite in a stationary equatorial orbit. This problem and the necessary calculation methods are touched on in the referenced CCIR Report 393, and more refined methods have been suggested in proposed draft revisions to that report. At any given terrestrial site there are only two azimuths at which such intersections can occur, and only in the near vicinity of these azimuths are detailed calculations needed, taking into account the effects of changes in the refractive index of the atmosphere and the angle of elevation of the terrestrial beam.

Earth stations are typically located in areas which are relatively isolated and, where possible, well shielded by surrounding terrain. As more and more earth stations come into the picture, the problems can be expected to increase.

C. Propagation

Under ROUTE AND SITE SELECTION there is contained a discussion intended to impart a general appreciation of path influences. The purpose of the present section is to cover the subject in more depth and to provide working criteria for establishing antenna elevations.

1. Variations in Signal Level due to Fading

Fading is a general term applied to loss of signal as seen by the radio receiver at its input. The term is intended to apply to propagation variables in the actual radio path. This section is concerned with actual fading, which is the change in path loss between the transmitter at one station and its normal receiver at the following station. These changes in path loss have to do with atmospheric conditions and the geometry of the path. This is true even though at times the complexity of conditions, and the impracticability of measuring precisely the parameters at the moment of difficulty, may result in erroneous judgments being made concerning cause and remedies.

Comparison with Carrier Systems

The effect of fading on radio paths is very much greater than the attenuation variables of open wire and cable carrier systems, which are primarily due to the effect of temperature variations in the metallic medium. Radio fading is caused by refractive, diffractive and reflective effects in connection with the atmosphere and fixed objects, which can result in defocusing, blocking and sometimes cancellation from multiple paths of varied lengths, due to the resultant variation in phase angles on arrival at the receiving antenna.

2. Ground, Sky and Space Waves

Considering the radio spectrum in its entirety, radio waves are propagated from one point to another in three principal forms; ground waves, sky waves and space waves. At a given frequency, these forms bear a specific relationship to each other, depending upon the characteristics of the media through which they travel. Ground waves, for example, depend on the reflection coefficient, induction field and secondary ground effects, as well as the ground conductivity.

General Effects of Frequency

Whereas the proximity of the ground has a profound effect on the propagation of radio waves, its effect varies substantially with frequency. If the space wave transmission (which itself varies with frequency) is taken as a standard, the surface wave diminishes much more rapidly with frequency, and is negligible in the microwave region. The induction field and secondary effects of the ground can usually be ignored, so that for the portion of the spectrum of interest here, there is only the direct wave and the reflected wave, reflections from fixed objects, or sky waves which are generally classified in this category although, theoretically at least, refraction may be the proper category.

3. General Nature of Microwave Propagation

Because the path of a radio beam is often referred to as line-of-sight, it is often thought of as a straight line in space from transmitting to receiving antenna. The fact that it is neither a line, nor is the path straight, leads to the rather involved explanations of its behavior, which attempt to give an understanding of the fundamentals of path propagation essential to the solution of the problems within each radio path.

Comparison with Light Waves

A microwave beam and a beam of light are similar in that both consist of electromagnetic energy, the difference in their behavior being principally due to the difference in frequency. Because a light beam is visible, it is easier to observe its behavior. So long as the similarities and differences can be classified, the comparison is useful. As a matter of fact, most of the characteristics of microwaves can be visually demonstrated with light waves, and in a very small space.

A basic characteristic of electromagnetic energy is that it travels in a direction perpendicular to the plane of constant phase; i.e. if the beam were instantaneously cut at right angle to the direction of travel, a plane of uniform phase would obtain. If, on the other hand, the beam entered a medium of non-uniform density and the lower portion of the beam traveled through the more dense portion of the medium, its velocity would be less than that of the upper portion of the beam. The plane of uniform phase would then change, and the beam would bend downward. This is refraction, just as a light beam is refracted when it moves through a prism.

The atmosphere surrounding the earth has the non-uniform characteristics of temperature, pressure and relative humidity, which are the parameters that determine the dielectric constant, and therefore the velocity of propagation. The earth atmosphere is therefore the refracting medium that tends to make the radio horizon appear closer or farther away. It also affects the path clearances in the manner discussed in the section on SITE AND ROUTE SELECTION. In the discussions following, it will be shown how it also affects other factors of judgment on the radio path.

4. Free Space Attenuation

Although the atmosphere and terrain over which a radio beam travels have a modifying effect on the loss in a radio path, there is, for a given frequency and distance, a characteristic loss. This loss increases with both distance and frequency. It is known as the free space loss.

Definition

Free space loss is defined as the loss that would obtain between two isotropic antennas in free space, where there are no ground influences or obstructions; in other words, where blocking, refraction, diffraction and absorption do not exist. An isotropic antenna is defined as one which radiates or receives energy uniformly in all directions. Although such an antenna is physically unrealizable, it provides a convenient reference point for calculations. Path loss charts for microwave tranmission are customarily prepared on the basis of free space loss between isotropic antennas, and antenna gains are specified with respect to the gain of an isotropic antenna. These gains may be easily applied to obtain the net loss from the waveguide out at the transmitter to the waveguide in at the receiver. This is often referred to as the net loss for the path.

Nature of Losses in Free Space

Radio energy is lost in space primarily because of the spreading of energy in the wavefront as it travels through space, in accordance with the inverse-square law. Only a small amount of the energy which is radiated from the transmitting antenna actually reaches the receiving antenna. The remainder is spread over areas of the wavefront outside the capture area of the receiving antenna.

Free Space Formula

The derivation of the formula for free space loss involves the isotropic radiator, from which the energy is transmitted equally in all directions. If one were to look instantaneously at the surface of constant phase at some point d distance from the source, it would appear as a sphere of radius d. If one were to intercept the energy impinging on a small portion of that surface with an area of A, the energy intercepted would bear a relationship to the total energy from the source as A bears to the total area of the sphere, which is $4\pi d^2$. This relationship represents the loss between a point source and an antenna whose "gain" in terms of A is equal to $\frac{4\pi A}{\lambda^2}$ where λ is wavelength. By appropriate substitutions and converting d to miles and frequency in GHz as an inverse function of wavelength, the loss between two isotropic antennas becomes:

$$A = 96.6 + 20 \log_{10} F + 20 \log_{10} D \qquad (3)$$

where A = free space attenuation between isotropics, in dB
F = frequency in GHz
D = path distance, in miles

For very short distances, such as between two points on the same tower, another formula is useful, but is usually stated verbally rather than mathematically. Simply stated, it is; for a distance equal to one wavelength the loss is 22 dB, and each time the distance is doubled, another 6 dB is added. For example; at 6 GHz, one wavelength is 0.05 meters and the loss for this distance is 22 dB. At 0.1 meter, the loss is 28 dB and at 0.2 meter, the loss is 34 dB. This progression builds up rapidly and can be used in connection with near-end crosstalk calculations where the antennas are separated on the tower. The two loss formulas can be shown to produce identical results at a given distance. Figure 13 is a set of curves showing free space losses at different frequencies and distances.

5. Terrain Effects

The effect of obstacles, both in and near the path, and the terrain, has a bearing on the propagation of radio energy from one point to another, even at microwave frequencies, where the ground wave does not enter into the calculations. The nature of these effects depends upon many things, including the position, shape and height of obstacles, nature of the terrain, and whether the effects of concern are primary or secondary effects.

Blocking, Cancellation of Out of Phase Signals

Where an obstacle is blocking, much depends upon whether it is totally or partially blocking, whether the blocking is in the vertical or the horizontal plane, and the shape and nature of the obstacle. The most serious of the secondary effects is reflection from surfaces in or near the path, including the ground. The problem in such cases, is that the reflected energy travels a separate and longer route than the main beam, and usually is not directly in phase with it. The result is often cancellation of the direct signal to an extent dependent on the relative powers of the direct and reflected signal, and the relative phase of the two signals arriving at the receiving antenna.

Nature of Obstruction Losses

It may be useful to discuss the effect of some commonly experienced obstacles to illustrate the nature of obstruction losses. Trees for example, cause dispersion of energy and affect the vertical clearance. At grazing they look similar to a knife edge in diffraction theory, and result in about a 6 dB loss. When they become obstructions they are normally considered to be totally blocking. At clearances above grazing there are usually no secondary effects of consequence above the point of free space transmission.

The effect of man-made obstacles depends entirely upon their shape and position, if microwave-transparent objects, which are few, are ignored. A large round container such as a gas storage reservoir, if partially in the path, causes both diffraction and dispersion as well as some blocking. It will also reflect a certain amount of energy off-path, where the wrong receivers may be affected to the point of serious interference. A more common object is a water tower, which may produce some of these effects at times when refractive conditions are such as to place the tank portion in the main beam.

Square or rectangular objects in the path or near it can be very destructive from the transmission standpoint, not only because of blocking effects, but also because diffraction occurs over and around them. The flat surfaces cause reflections. These are all, in effect, spurious signals and may cause both interference and signal cancellation. In

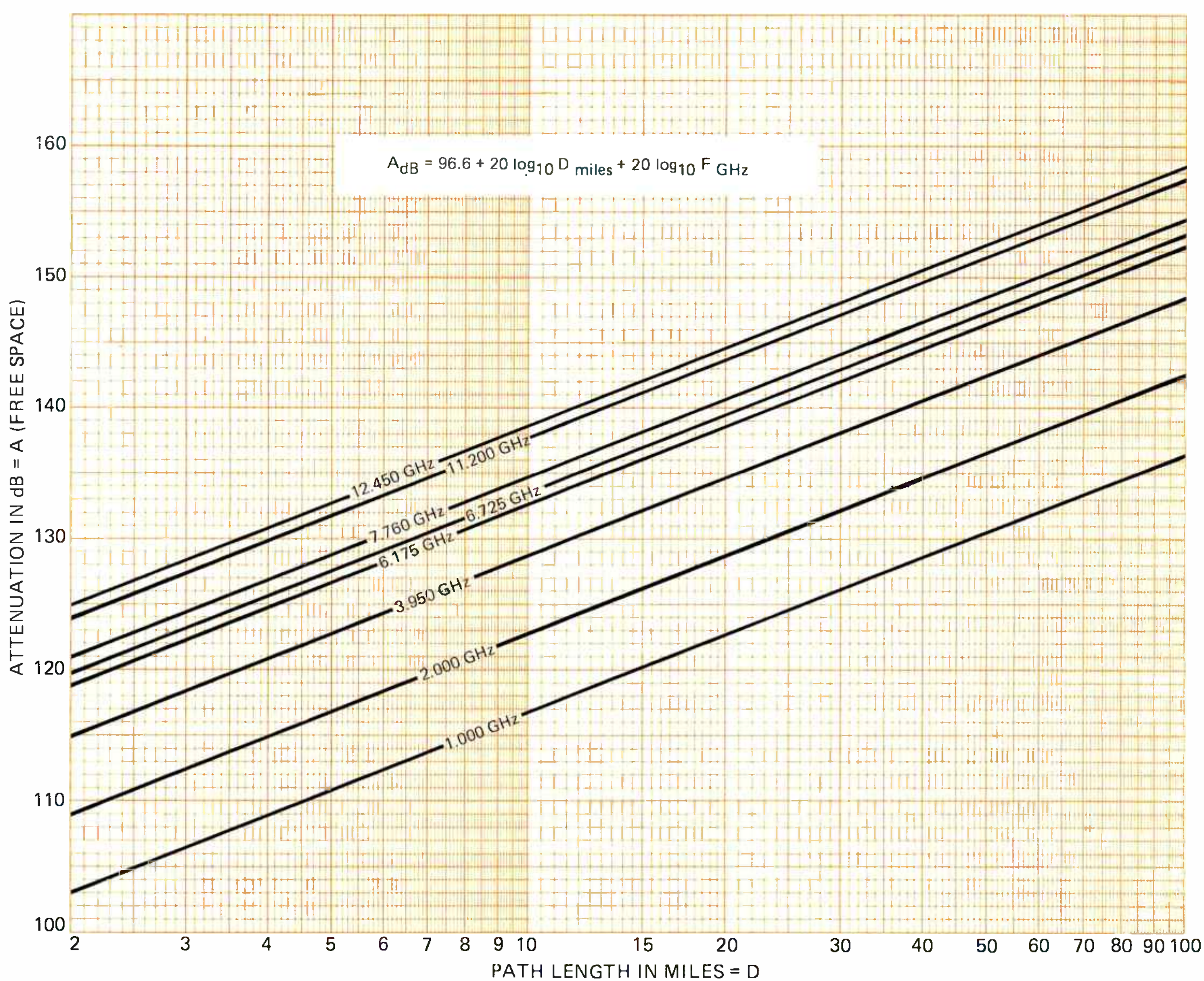


Figure 13. Free Space Attenuation Between Isotropic Antennas

cities, where this usually occurs, reflected energy has been known to interfere with a receiver from the opposite direction, or after several reflections, cause delay distortion. Billboards are also good reflectors. One particularly destructive type is the mechanically rotating sign. This results in energy being deflected in various directions as it rotates. In a working system it can usually be identified by comparing the rate of rotation with the rate of interference in the system.

Figure 14 is an interesting set of curves which illustrate the effect of the shape of an object over which a microwave beam passes. The curve marked R = 0 illustrates the knife edge diffraction case as shown in Figure 1. The curve marked R = −1.0 illustrates the smooth sphere diffraction case also shown in Figure 1. The significance of the negative reflection coefficient is the 180° phase shift which takes place for reflected energy at low angles. The curve marked R = −0.3 represents about the common experience on many paths. It should be pointed out that the criterion of smoothness is not always as obvious as it is, for instance, on water or a dry lake bed. The point is that within the "obstruction zone" the loss in signal depends substantially on the nature of the obstruction.

The curves of Figure 14 are illustrative only, and are not intended to be used for path loss calculation purposes, particularly in the obstruction zone. In situations where actual calculations of such obstructed zone losses are required for a particular path, it is recommended that the methods developed in NBS Technical Note 101, Revised, or other equivalent methods be applied.

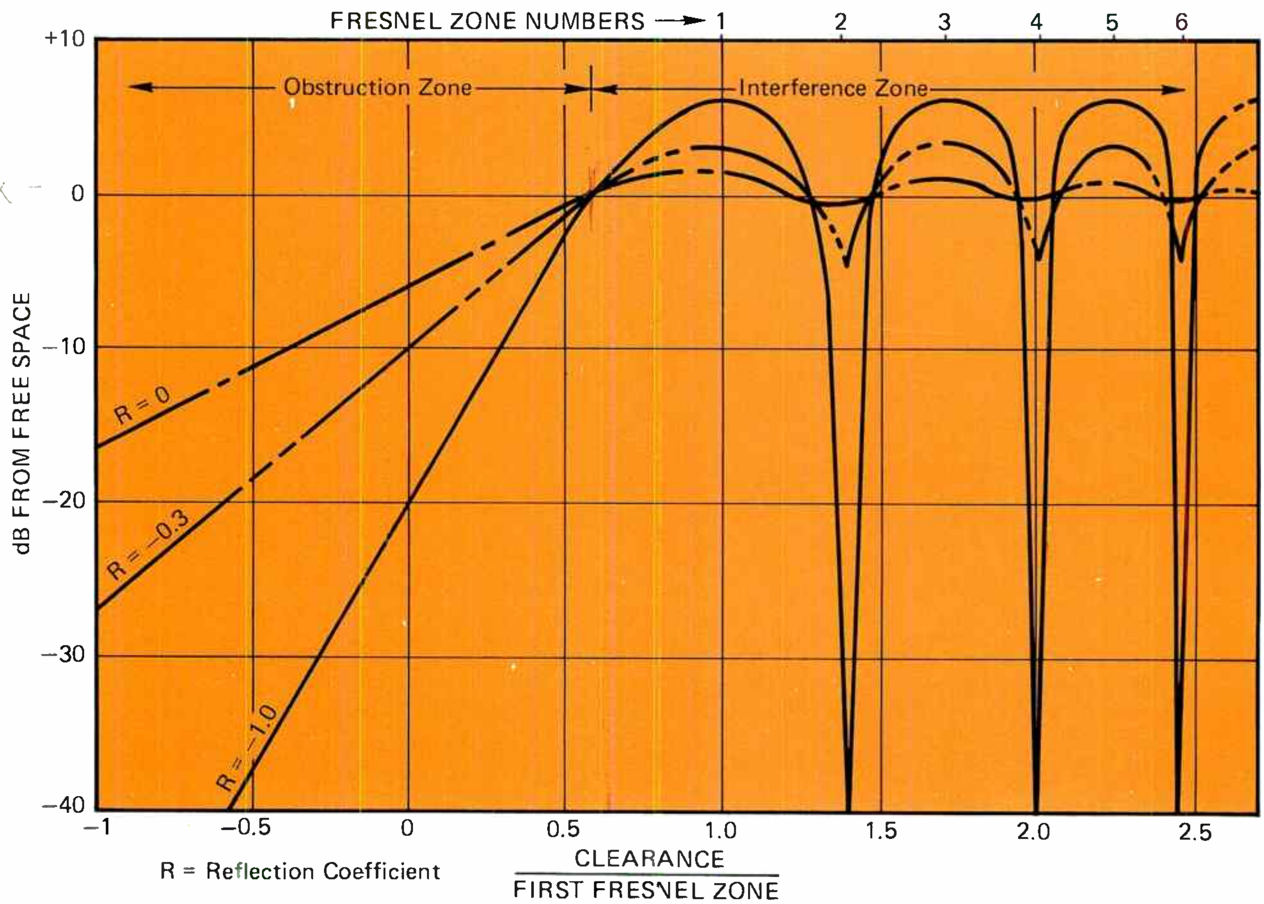


Figure 14. Attenuation vs Path Clearance

Development of Fresnel Zone Radii

Refer again to Figure 14. The solid curve, representing transmission losses when the reflection coefficient R is -1.0, is the type of curve to be expected when path height-vs-loss tests are made over flat, relatively smooth terrain such as dry lake bed, the salt flats of Utah or smooth water surface. The type of test represented is the typical path loss test, where transmitter and receiver are separated by a normal path length with smooth terrain between. The curve is the result of a height-loss run as described under ROUTE AND SITE SELECTION. As the transmitting and receivng antennas are raised to the point where free space loss is just obtained, and then further, the received signal reaches a peak value. This is because the direct signal from the transmitting to the receiving antenna, and the ground-reflected energy are in phase addition at the receiving antenna. Since it is known that, for the very low angles of incidence typical of microwave paths, there is a 180° phase delay $(\frac{\lambda}{2})$ at the reflection point, the total reflection path from transmitting antenna to reflection point to receiving antenna must be 1/2 wavelength $(\frac{\lambda}{2})$ longer than the direct signal path. If this is true, there must also be a point where the direct and reflected signals are in phase opposition. This first occurs when the reflection path is one wavelength longer than the direct signal path. With the 180° phase delay at the reflection point, the reflected signal is, in effect, one and one half wavelengths behind the direct signal at the receiving antenna, and in phase opposition to the direct signal. The first point of phase addition is the point where first Fresnel zone clearance over the reflection point obtains. If an imaginary line were drawn longitudinally on the path from the transmitting antenna to the receiving antenna, such that a reflection surface at any point on that line would produce first Fresnel zone addition, it would describe a very long, narrow ellipse. If this ellipse were rotated about the main beam as an axis (using the straight beam philosophy and the center of the beam) it would describe an ellipsoid of revolution, which is the locus of all possible surfaces which would produce first Fresnel zone addition. The condition therefore applies to objects at the side of the path as well as to ground reflections. The distance from the exact center of the main beam to the above line or surface is also known as the first Fresnel zone radius. It will be noted that the reflection point need not be in the middle of the path, but at any point which provides the necessary geometric relations.

The first Fresnel zone at any point in the path may be calculated from the following formula:

$$F_1 = 72.1\sqrt{\frac{d_1\ d_2}{fD}} \qquad (4A)$$

where F_1 = First Fresnel zone radius in feet

d_1 = Distance from one end of path to reflection point in miles

D = Total length of path in miles

d_2 = $D - d_1$

f = Frequency in GHz

Once again refer to Figure 14. The lowest antenna elevations which will produce cancellation of main and reflected signals are indicated by the null below Fresnel zone 2. The locus of points along the path where surfaces would produce the same condition describes a larger ellipse, and its ellipsoid describes the locus of all possible surfaces which would produce second Fresnel zone cancellation. As illustrated in Figure 14, the pattern of additive and canceling Fresnel intervals increases with increasing path clearance, but the differences keep getting smaller as the number increases. All of those which produce cancellation are even numbered zones. If the value for the first Fresnel zone is known and it is desired to calculate the n^{th} zone, where n is the Fresnel zone number, then:

$$F_n = F_1\sqrt{n} \qquad (4B)$$

Or to calculate any Fresnel zone directly:

$$F_n = 72.1\sqrt{\frac{nd_1\,d_2}{fD}} \qquad (4C)$$

Figure 15 is a curve which permits determination of the first Fresnel zone for any point on a path of given length for the center of the 5.925—6.425 GHz band. This is sufficiently accurate for the entire band.

Table B1 lists multiplying factors to allow Figure 15 to be used for each of the other commonly used bands. In each case the value of F read from the scale is to be multiplied by the factor listed for the band of interest. Also included is a table of square roots of integral numbers from 1

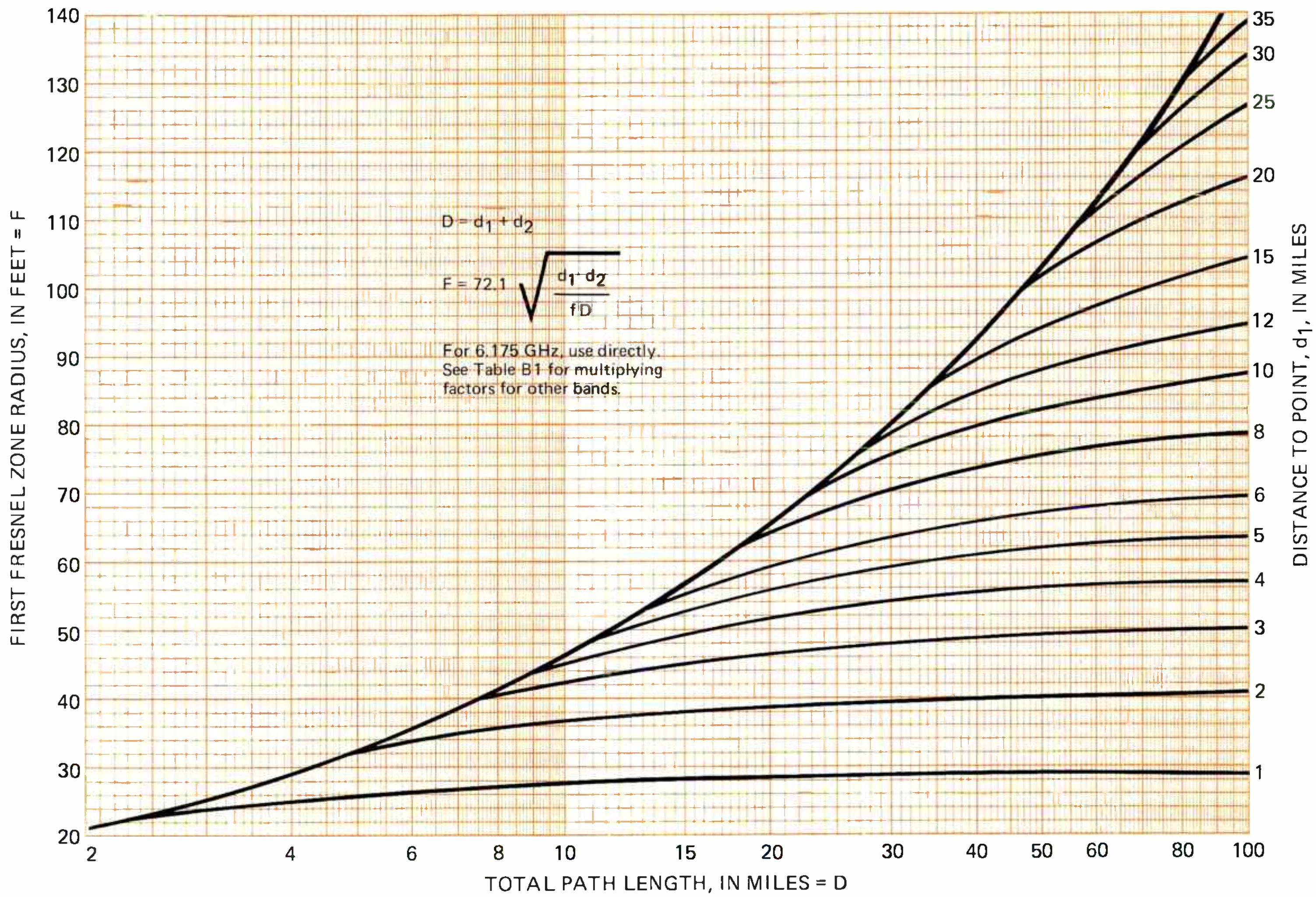


Figure 15. First Fresnel Zone Radius (6.175 GHz)

through 65 (Table B2), to facilitate calculation of the n^{th} Fresnel zone radius when the 1st Fresnel zone radius is known. To calculate the n^{th} zone radius, simply multiply F_1 by the square root of n.

At this point, the reader is reminded that the microwave beam travels as a wavefront of considerable tranverse cross-section. Specular reflection of such a wavefront requires a reflecting surface of a certain area, rather than a single point. The receiving antenna collects reflected energy from the reflection surface from ahead, beyond and to the sides of the theoretical "reflection point". When the terrain is perfectly flat and the reflection coefficient is −1.0, it is possible to receive exactly as much energy from the reflection path as from the direct signal path. Under these conditions, when they are in phase addition, the summation signal at the receiver will be 6 dB higher than would obtain with free space loss, or without the ground influence. When the two signals are in direct phase opposition, the loss can be very large, approaching cancellation. There are also some other effects which tend to distort the signals, and these are covered in a later section.

Table B1. Multiplying Factors Which Can Be Used To Convert Fresnel Zone Radii Calculated For 6.175 GHz To Other Bands.

BAND GHz	CENTER FREQUENCY	MULTIPLY BY
1.850 – 1.990	1.920	1.793
1.990 – 2.110	2.050	1.735
2.110 – 2.130 2.160 – 2.180	2.145	1.697
2.130 – 2.150 2.180 – 2.200	2.165	1.688
2.450 – 2.500	2.475	1.580
3.700 – 4.200	3.950	1.250
4.400 – 5.000	4.700	1.146
5.925 – 6.425	6.175	1.000
6.575 – 6.875	6.725	0.9582
6.875 – 7.125	7.000	0.9392
7.125 – 8.400	7.437	0.9112
	7.750	0.8926
	8.063	0.8751
10.700 – 11.700	11.200	0.7425
12:200 – 12.700	12.450	0.7043
12.700 – 12.950	12.825	0.6939
12.700 – 13.250	12.975	0.6899

Up to this point the discussion has been principally about a perfectly reflecting surface. This is not a common experience. Referring again to Figure 14, it will be noted that, as the reflection coefficient becomes less, the addition and cancellation effects diminish, so that for many paths clearance is the principal concern. One very interesting point about the curves however, is that regardless of the reflection coefficient, the elevations where free space loss first occurs are such that approximately $0.6F_1$ clearance for the center of the beam obtains over the reflection point. When path loss tests are made, the value of K at the time of the test can be determined with this knowledge and an accurate profile.

While, as mentioned above, many paths do not exhibit serious reflective characteristics, reflective phenomena are not confined to totally flat areas and water surfaces. There are many areas with rolling hills that are devoid of trees and brush, or undulating valleys between high hills, which will cause serious reflections. Our common experience with light waves tends to inhibit our judgment as to the criterion of smoothness, and also regarding the geometry of the path itself. The judgment of such areas by visual inspection alone is often inaccurate with regard to three important parameters. These are; the criterion of smoothness for the frequency to be used, the effective area of a possible reflecting surface, and the angles of incidence and reflection.

The reflection coefficient actually increases markedly as the angle of incidence becomes very small. This can be illustrated by viewing an asphalt highway surface at a considerable distance. The surface appears like glass and actually reflects the sun at this very small angle of reflection. Yet when viewed from directly overhead, the surface appears quite rough at light frequencies. Since microwaves have a very much longer wavelength than light waves, and the angles of incidence and reflection are very small, the surface can be very much rougher than the highway surface and still reflect very well. For instance, wheat stubble, a uniform growing crop, and flat prairies, can be very good reflectors at 2, 4 and 6 GHz. At 11 to 13 GHz the reflection coefficient appears to be somewhat reduced, probably because of the shorter wavelengths.

Table B2. Multiplying Factor For Determining F_n When F_1 Is Known. ($F_n = F_1\sqrt{n}$)

n	$\sqrt{n}$	n	$\sqrt{n}$	n	$\sqrt{n}$	n	$\sqrt{n}$	n	$\sqrt{n}$
1	1.000	16	4.000	31	5.568	46	6.782	61	7.810
2	1.414	17	4.123	32	5.657	47	6.856	62	7.874
3	1.732	18	4.243	33	5.745	48	6.928	63	7.937
4	2.000	19	4.359	34	5.831	49	7.000	64	8.000
5	2.236	20	4.472	35	5.916	50	7.071	65	8.062
6	2.449	21	4.583	36	6.000	51	7.141		
7	2.646	22	4.690	37	6.083	52	7.211		
8	2.828	23	4.796	38	6.164	53	7.280		
9	3.000	24	4.899	39	6.245	54	7.348		
10	3.162	25	5.000	40	6.325	55	7.416		
11	3.317	26	5.099	41	6.403	56	7.483		
12	3.464	27	5.196	42	6.481	57	7.550		
13	3.606	28	5.291	43	6.557	58	7.616		
14	3.742	29	5.385	44	6.633	59	7.681		
15	3.873	30	5.477	45	6.708	60	7.746		

There are at least two conditions which produce reflections quite similar to ground reflections, and are often classified with ground reflection phenomena. A growing crop such as alfalfa often, on level irrigated land, tends to collect a heavy dew in the early morning. The resulting "surface" between wet vegetation and dry air, as presented to microwaves, is highly reflective. Also in a fairly arid country where there is a small stream in a valley, there is often, at times of still air, a low ground fog. The upper surface of this fog represents an abrupt change in the dielectric constant and can be reflective.

6. Atmospheric Effects

The matter of establishing antenna elevations to provide minimum fading would be relatively simple were it not for atmospheric effects. The antennas could easily be placed at elevations to provide somewhere between free space loss and first Fresnel zone clearance over the predominant surface or obstruction, reflective or not, and the transmission would be expected to remain stable. Unfortunately, the effective terrain clearance changes, due to changes in the air dielectric, with consequent changes in refractive bending. The radio beam is almost never a precisely straight beam in fact, which is to say that using our straight beam concept, the condition for K = 1 is a rarity. Actually the most stable atmospheric condition exists, in most areas, during the daytime hours from 1 to 2 hours after sunrise to 1 to 2 hours before sunset, and during normal weather conditions. The variation in these limits morning and evening, are determined by local conditions. For instance, sunrise or sunset are quite different in mountainous terrain than they are at sea level, or in intermediate terrain. Between these daytime hours, for most areas, the refractive effect does not usually vary much from the K = 4/3 condition. During these daylight hours, rising convection currents and winds tend to produce a homogenenous atmosphere, in the sense that stratification in the air does not exist, and the refractive gradient resulting from the normal pattern of pressure, temperature and relative humidity is fairly uniform. Because this is the only reasonably stable refractive condition for any period, it is often referred to as standard. It is also the basis for the start of path analysis in many instances, because beam bending is considered in relation to this reference standard.

The Refractive Index

The radio refractive index of the atmosphere, n, is a number on the order of 1.0003, varying between 1.0 (free space, above atmospheric influence) and about 1.00045 at a maximum. For greater computational convenience, it is customary to utilize a term N, called "radio refractivity", which is defined as:

$$N = (n-1) \times 10^6 \qquad (5)$$

The "N" term would be zero in free space, and a number on the order of 300 at the earth surface.

The radio refractivity of air for frequencies up to 30 GHz is given as:

$$N = 77.6\frac{P}{T} + 3.73 \times 10^5 \frac{e}{T^2} \qquad (6)$$

where P is the total atmospheric pressure in millibars, T is the absolute temperature in degrees Kelvin, and e is the partial pressure of water vapor in millibars. The P/T term is frequently referred to as the "dry term" and the e/T^2 term as the "wet term".

By examination of Equation 6, it can readily be seen that, while pressure and relative humidity are direct factors in the refractive index, temperature as a function of N is the predominant factor. In the following discussion, the phenomenon of temperature inversion is covered. It is easily seen why this phenomenon is of concern in connection with radio propagation.

M Profiles

The discussion of atmospheric irregularities is often aided by the use of another term or symbol, M, which is called the "modified" index of refraction. It is defined in terms of the radio refractive index and the mean sea level elevation. The following formula is applicable:

$$M = (n-1) \times 10^6 + 4.8h \qquad (7)$$

where n = the radio refractive index
h = the height above sea level in hundreds of feet

In the normal atmosphere (where K = 4/3), M increases at a linear rate of about 3.6 units per hundred feet increase in altitude. When height is plotted as ordinate against M as abscissa, the plot is called an M profile. The slope of the M profile determines the degree of bending of the microwave beam in relation to the earth.

For discussion purposes it might be somewhat easier to follow a simplified equation for M, which results from combining equations 4 and 6. The term N, which is the variable fraction of the refractive index, then becomes the parameter against which the modified index M and its profile are compared. The new equation becomes:

$$M = N + 4.8h \qquad (8)$$

Figure 16 is a group of M profiles representing specific conditions which will be discussed in the following paragraphs.

Illustrations of Refraction as Related to the M Profile

Figure 16a illustrates the M profile representing the "standard" condition, where K = 4/3 and the slope of the profile is constant at 3.6 units per hundred feet.

If N decreases more rapidly than normal with increasing altitude, the slope of the M curve is steeper, the value of K is greater than 4/3, and the microwave beam follows the curvature of the earth more closely. If the value of M becomes constant with change in altitude, the microwave beam follows the curvature of the earth exactly, and K is equal to infinity. This condition is represented graphically by a vertical M profile, as shown for super-standard in Figure 16b. It would be represented on a path profile by a flat earth and a straight line microwave beam. If conditions become even more extreme, the M profile will have a negative slope, corresponding to a negative K, or a "concave earth" condition. An example of this condition is the rare instance when an overreach interference signal is recorded at a station, and yet the overreach path is obstructed by several hundred feet when the path profile is plotted on flat earth.

When the slope of the M profile is greater than normal, the refractive condition is referred to as super-standard or earth flattening, since the radio horizon distance is increased. When the slope of the M profile is less than normal, the refractive condition is known as sub-standard or earth bulging. This condition is reflected in Figure 16b.

In practice, linear profiles do not usually occur except near the standard profile, because weather factors usually change the shape of the M profile as well as its slope.

The effect of an abnormally high surface temperature, or increasing water vapor content with altitude, is shown in the M profile of Figure 16c. Such a sub-standard surface condition will result in curving the beam away from the earth, and this is called inverse beam bending or earth bulging. The effect is similar to that from a linear M profile with a slope less than normal (earth bulging), except that it is concentrated near the surface of the earth.

A rise in temperature with increasing height, or a decrease in water vapor content, or both, will produce the effect shown in Figure 16d. This is a slightly super-standard condition that will cause the beam to follow the curvature of the earth more closely (earth flattening).

When the changes in refractive index are most severe near the surface, the condition will be as shown in Figure 16e. This condition is known as a surface duct, because the beam will tend to stay within the surface and the elevation limit a, depending on the slope of the M profile near the surface. When the beam enters the duct at a small angle, it is bent until it is horizontal, and then turned downward by further bending.

Figure 16f is the M profile of an elevated duct, the upper limit of which is formed by the upper limit of the super-standard or inversion layer from a to b, and the lower limit by the sub-standard layer from b to c. Under these conditions, the beam will tend to remain within the duct limits from a to c, due to the bending toward the center of the duct. Concentration of radio energy within a duct will cause an increase in received signal when both the transmitting and receiving antennas are within the duct. Obviously this effect cannot be relied on for satisfactory propagation, because the conditions producing it are subject to change. The terms trapping, super-refraction and guided propagation are also employed to describe duct phenomena.

M profiles which are essentially linear are significant primarily because of their effect on path clearance. The non-linear profiles, in addition to affecting path clearances, also give rise to conditions leading to atmospheric multipath effects.

K Factors

The discussion of refractive effects as given here is principally for use as background information to aid in understanding the mechanisms and principles involved in the complex phenomenon of propagation through the atmosphere.

Despite the great amount of work which has been done in collecting and analyzing data on the variations with time of the refractive index, and its gradient at many locations throughout the world, the available data based on meteorological measurements is still of relatively limited value in determining the values of K to be used in engineering line-of-sight microwave paths.

Three K values are of particular interest in this connection

(1) Minimum value to be expected over the path. This determines the degree of "earth bulging" and directly affects the requirements for antenna height. It also establishes the lower end of the clearance range over which reflective path analysis must be made, in the case of paths where reflections are expected.

(2) Maximum value to be expected over the path. This leads to greater than normal clearance and is of significance primarily on reflective paths, where it establishes the upper end of the clearance range over which reflective analysis must be made.

(3) Median or "normal" value to be expected over the path. Clearance under this condition should be at least sufficient to give free space propagation on non-reflective paths. Additionally, on paths with significant reflections, the clearance under normal conditions should not fall at or near an even fresnel zone.

Of these three values of K, only the median value can be predicted with any degree of confidence from available meteorological data.

The minimum value chosen for K must, for a highly reliable path, be an extreme value which will be passed for only exceedingly small percentages of the time. Experience has indicated that, for actual microwave paths, the effective K over the entire path reaches a very high or very low value for a much smaller percentage of time than would be indicated by the distribution of K values as found by meteorological measurements at single points. The most probable explanation is that the unusual conditions causing these extreme values are unlikely to occur over more than a small part of the path at any given instant. In any event, the correlation between limiting K values as found in practice with those based on meteorological data has been found to be very low, and the best guide to choice of K values to be used in path engineering is past history and experience in the field.

The K factor corresponding to an atmosphere with a linear gradient of refractive index dn/dh can be calculated by the equation:

$$K = \frac{1}{1 + \frac{a}{n} \cdot \frac{dn}{dh}} \qquad (9A)$$

where a is the true radius of the earth, n the radio refractive index, and dn/dh is the gradient of n with respect to height in the portion of the atmosphere affecting the path.

In this equation, the variations of n itself are too small to have significant effect, and n can be taken as 1.0003 for all practical purposes. This leaves dn/dh as the significant variable affecting the value of K.

It is more convenient to consider the gradient of N units instead of the gradient of n units, and, when dN/dh is substituted for dn/dh, with the appropriate correction factor of 10^{-6}, and the value of 6370 kilometers for a and 1.0003 for n are substituted in (9A), the following equation is obtained:

$$K = \frac{157}{157 + \frac{dN}{dh}} \qquad (9B)$$

where dN/dh is the N gradient per kilometer.

From (9B) it can be calculated that an N gradient of —40 units per kilometer would give a K of 4/3 (the so-called standard atmosphere), an N gradient of —157 units per kilometer would give a K of infinity (super-normal refraction), and an N gradient of about +79 units per kilometer would give a K of 2/3 (sub-normal refraction).

In addition to the linear variations in the gradient, there are periods of considerable non-linearity such as just after sundown, particularly in coastal areas. At that time the lower strata of air is cooling; the relative humidity is increasing, the n gradient is increasing, and the value of K increases. This pattern often holds until around midnight, after which the relatively warmer earth surface, and the cooler air above it, appear to produce an irregular pattern in which values of K smaller than the daytime values are not unusual.

There is also a seasonal pattern. The maximum values of K are usually greater in summer than in winter, with the minimum values for summer and winter running about the same. There is, therefore, a greater excursion of K in summer than in winter. This, together with its effect on the radio path, has given rise to the generally prevalent feeling that fading is a summer phenomenon. This is not exclusively true, as there are radio paths which fade in winter as well as summer, but in most areas the summer fading is likely to be much more frequent and sometimes more severe.

Weather Fronts

Weather fronts moving through a particular area during the hot summer months are usually accompanied by sudden cooling in the lower atmosphere. The result is a rather abrupt change in the air dielectric to a lower value, with a reduction in the n gradient and an increase in the value of K. This is a very temporary change. The value of K changes according to the storm condition while it is in progress, which may vary widely depending upon whether saturation rain, high winds, hail, etc., obtain. After the storm, the value of K usually returns to something in the order of its normal value for the time of day and the area. The drop in barometric pressure preceding a storm front by some hours, is in itself not a large variant in the value of K, although it does affect the refractive index as such. Referring to Equation 6, it is estimated by the National Bureau Of Standards that the "dry term" $\frac{77.6P}{T}$ accounts for at least 60% of the value of N.

Rain Attenuation

Attenuation of a microwave signal due to rainfall or snow along the path, is present to some degree at all microwave frequencies, but the effect is so small as to be insignificant, at least in comparison to the other types of fading, for the bands of 8 GHz and lower. But at higher frequencies, the excess attenuation due to rain increases rather rapidly and, in the bands above about 10 GHz, is great enough to significantly affect path length criteria, except in areas of very light precipitation.

The degree of attenuation is a function of a number of variables including the frequency band, size and shape of the drops, and the distribution of

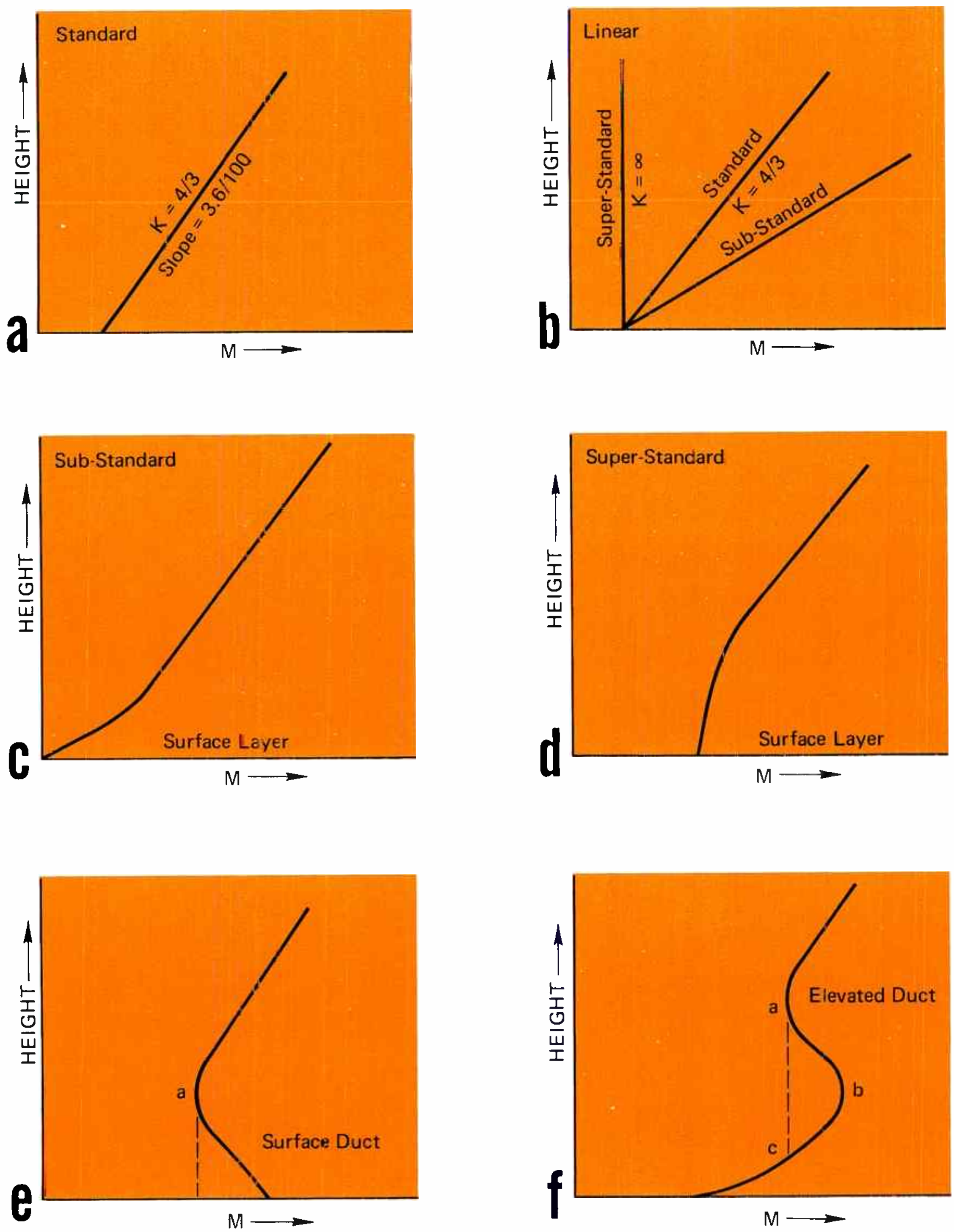


Figure 16. Typical M Profiles

rain (in terms of its instantaneous intensity) along the path. What is important is not the total amount of rain which falls over an extended period, but rather the maximum instantaneous intensity of fall which is reached at any given moment, and the size of the area over which the high intensity cell extends at that moment. Most available rainfall statistics cast little light on these matters and, consequently, are of only limited value in estimating the magnitude of rain attenuation effects. Figure 17 gives excess path loss in decibels per mile versus rainfall rate in inches per hour, for several frequency bands. It is based on the theoretical work of Ryde and Ryde. Figure 18, taken from a CCIR document, provides somewhat similar information in the form of excess path loss in decibels per kilometer versus frequency in GHz, for a number of rainfall rates. This figure also includes curves for attenuation in fog or cloud, which, though substantially lower than that of heavy rainfall, can nevertheless reach measureable values at the higher frequencies. Hathaway and Evans, in an article published in Communications and Electronics, January, 1959, discussed both the theoretical and practical aspects of the rain attenuation problem and provided application data for the 11 GHz band, applicable to the continental United States, which is still one of the best available sources. Figure 19, adapted from that article, divides the U.S. into eight geographical areas in ascending order of effect of rain attenuation, and Figure 20 gives the estimated outage time in hours per year versus path length in miles for each area. The latter figure is based on 11 GHz paths with a 40 dB fade margin. Figure 20 can be used for 13 GHz by reducing the mileage figures on the path length scale by approximately 30%. Dropping the fade margin from 40 dB to 35 dB would increase the expected outage time by approximately 25%, while increasing it from 40 dB to 45 dB would decrease the expected outage time by approximately 15%.

The degree of severity of the rain attenuation problem, depends very critically on the degree of reliability which is established as an objective. Present day reliability objectives for highest reliability systems are such as to require per path reliabilities on the order of 99.99% to 99.9999%, depending on the number of hops involved. This means that total outage objectives for a path may range from 0.01% to as little as 0.0001%. On an annual basis, 0.01% amounts to approximately 53 minutes per year, while 0.0001% would amount to only about 30 seconds per year. Comparing these values to the yearly outage times shown on Figure 20, one can see that in the heavier rain areas, the second objective could not be met at all, and the first could be met only with very short paths. In the light rain areas such objectives might be met with even relatively long paths.

On the other hand, there are certain types of service (for example, CATV relaying) in which somewhat lower reliabilities, 99.9% for example, are considered quite acceptable, and in such cases relatively long paths may be practical, even in areas with high rainfall.

Two things to bear in mind in connection with rain attenuation are that (1) multipath fading does not occur during periods of heavy rainfall, so the entire path fade margin is available to combat the rain attenuation, and (2) neither space diversity nor in-band frequency diversity provide any improvement against rain attenuation.

One thing has been well established; that cross-band frequency diversity, with one channel in a 6 GHz band and the other in an 11 or 12 GHz band, is entirely practical, even in the heaviest rain areas and for highest reliability requirements. The reason is that, since multipath fading is unlikely during heavy rainstorms, the 6 GHz path can carry the service during such periods without needing any diversity protection. The degree of equipment protection is reduced slightly, but not by a significant amount.

Apart from the rain attenuation problem, the 11 to 14 GHz bands have excellent characteristics and, despite the rain limitation, they have proven very useful in practice. They are lifesavers in areas where congestion has used up all available frequencies in the lower bands. In addition to their usefulness in cross-band applications, they can be used for spur routes or very short paths in the heavy rain areas, and almost unrestrictedly in very light rain areas. They are extremely valuable for hops into or in the vicinity of earth stations, since they avoid the coordination problems associated with the shared 4 and 6 GHz bands. In any given situation, the effects of rain attenuation can be reduced by raising the fade margin, shortening the paths, or both.

Fog

When fog forms, either by nocturnal cooling of the ground or by the flow of warm air over cool

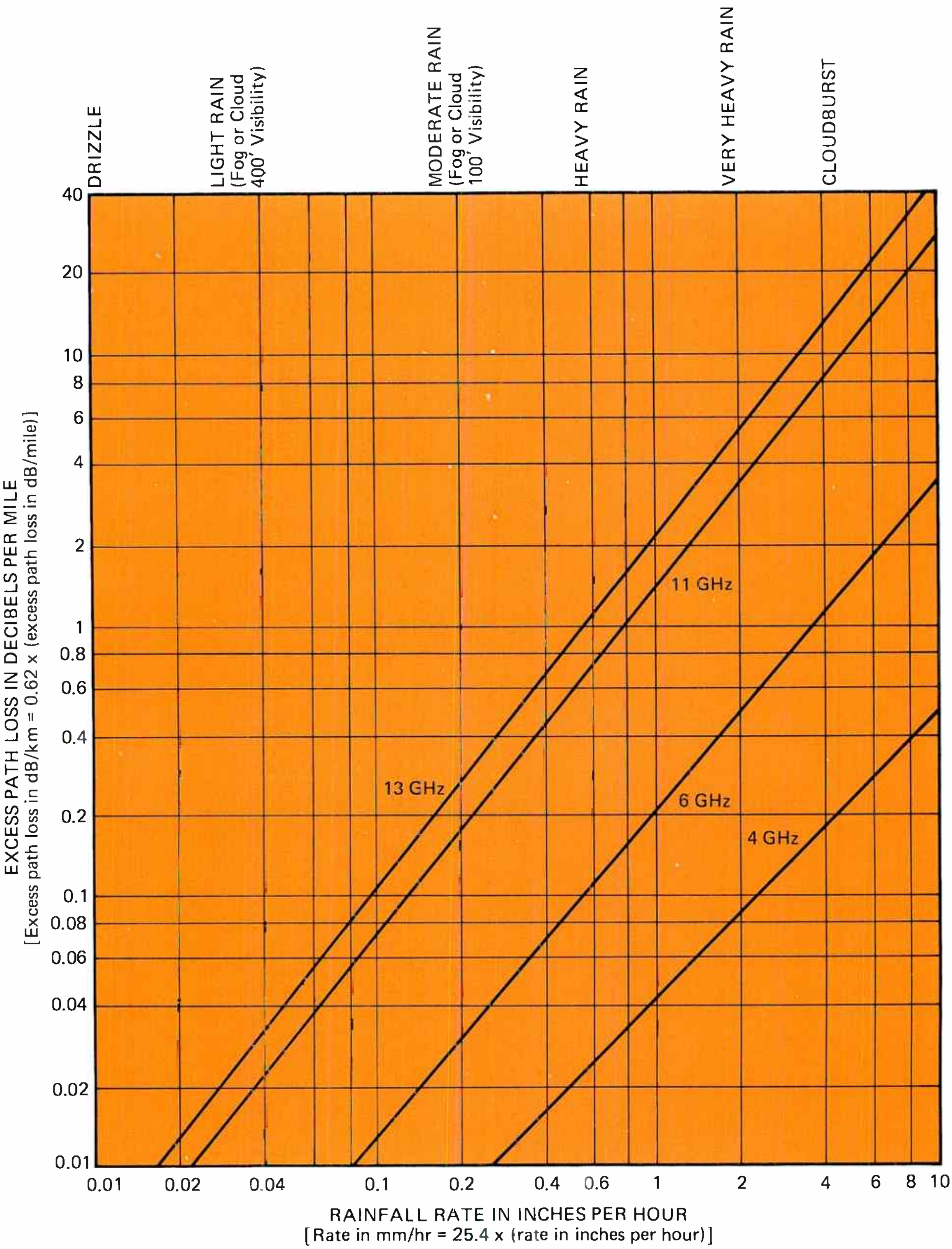
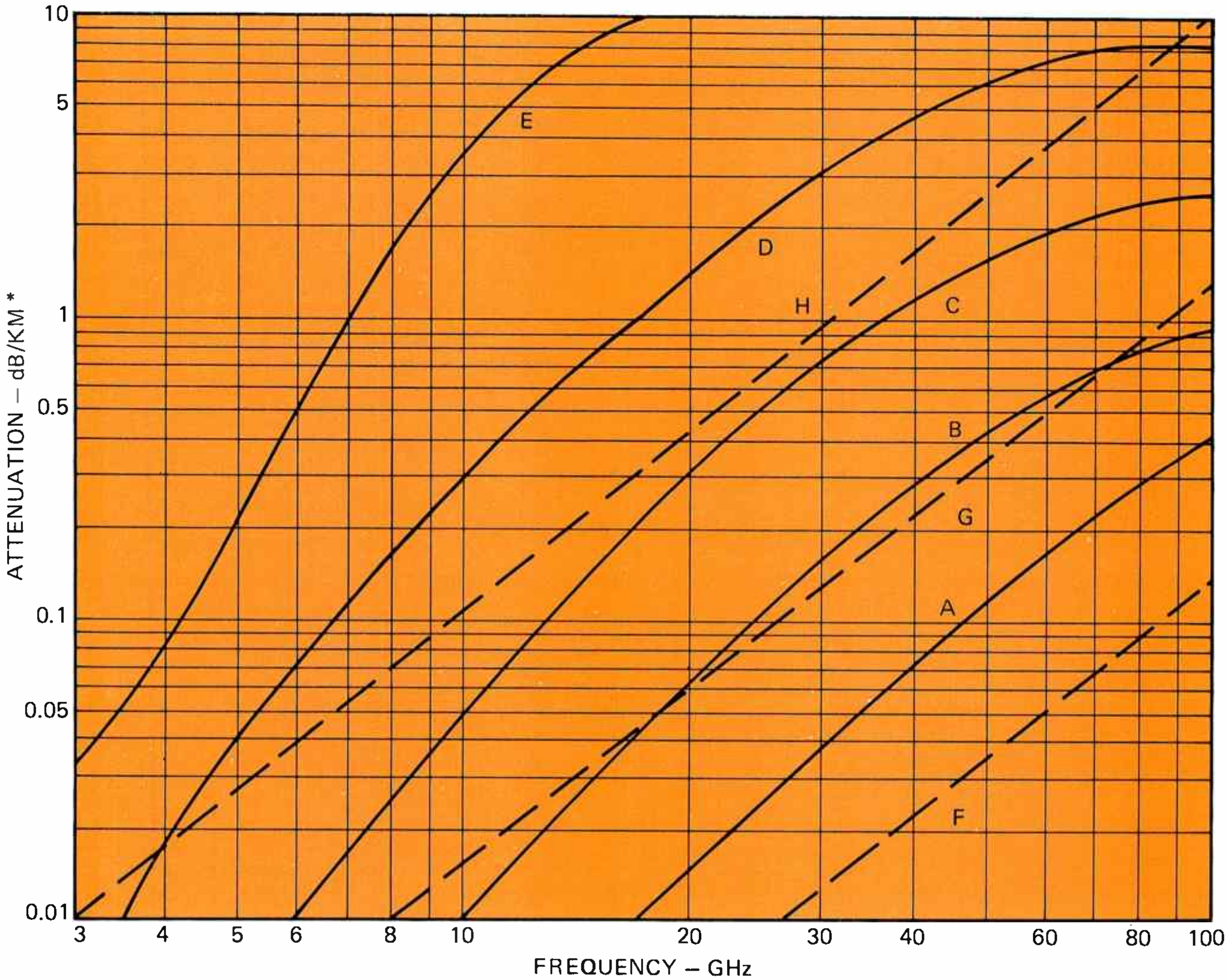


Figure 17. Rain Attenuation vs. Rainfall Rate (Theoretical, after Ryde and Ryde)



Attenuation in rainfall intensity of:
A, 0.25 mm/hr (drizzle) – .01 in/hr
B, 1.0 mm/hr (light rain) – .04 in/hr
C, 4.0 mm/hr (moderate rain) – .16 in/hr
D, 16 mm/hr (heavy rain) – .64 in/hr
E, 100 mm/hr (very heavy rain) – 4.0 in/hr

Attenuation in fog or cloud:
F, 0.032 gm/m^3 (visibility greater than 600 meters)
G, 0.32 gm/m^3 (visibility about 120 meters)
H, 2.3 gm/m^3 (visibility about 30 meters)

*Attn – dB/mile = 1.61 x (Attn in dB/km)

Figure 18. Attenuation Due To Precipitation (after CCIR)

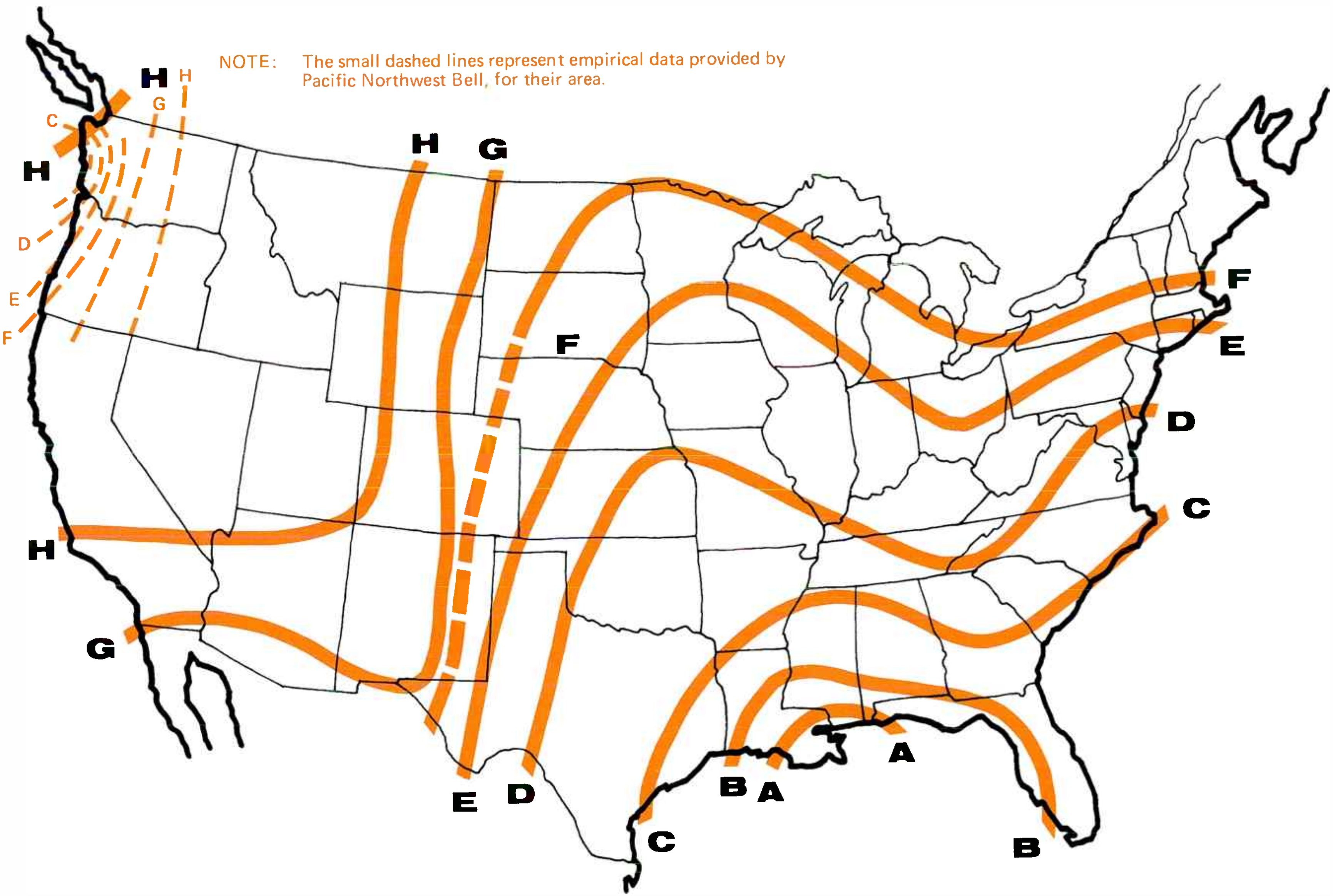


Figure 19. Contours Of Constant Path Length For Fixed Outage Time.

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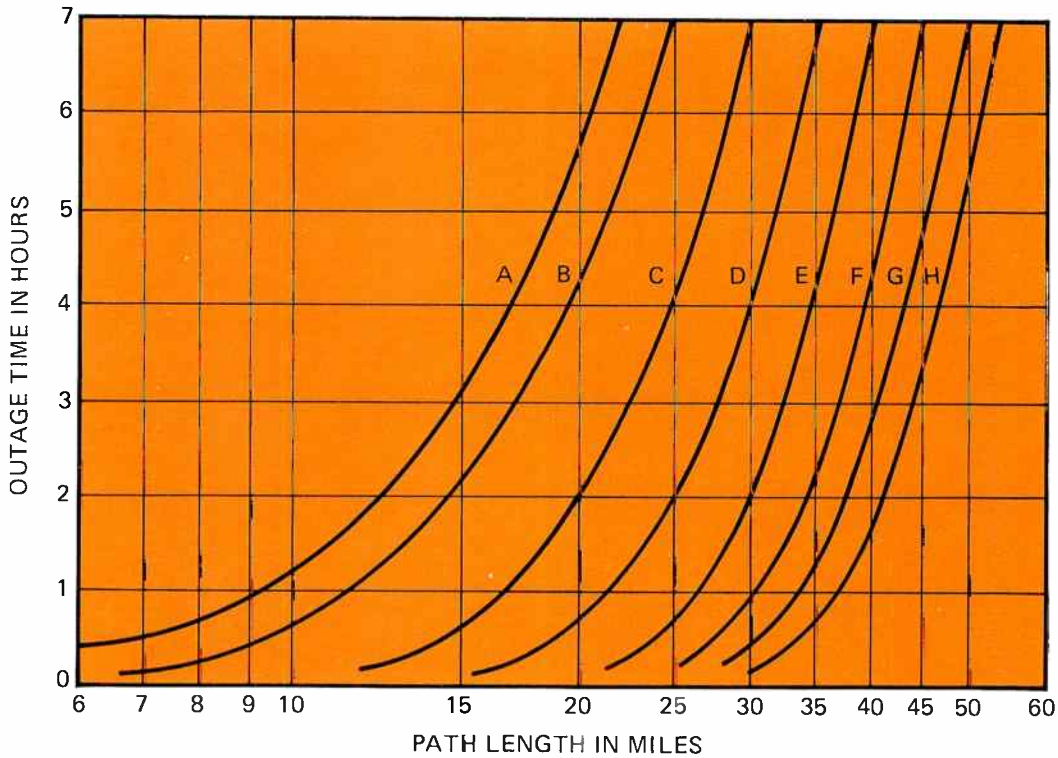


Figure 20. **Expected Outage Time In Hours Per Year vs. Path Length In Miles For Various Areas of the United States. (Based on 11 GHz paths with 40 dB fade margin; for 13 GHz paths, reduce path lengths by 30%. For 45 dB fade margin, decrease outage time by 15%; for 35 dB fade margin, increase it by 25%.)**

ground, the total amount of water in the air remains substantially the same, but part of the water condenses into minute droplets. Its contribution to the refractive index is then substantially less than when it is in the form of vapor. The weather effect normally accompanying a fog is very still air, and some temperature inversion. These are the conditions for a sub-standard surface layer with an M profile as in Figure 16c. The result is to reduce the effective beam clearance at all frequencies and, in extreme cases, will put the system out of service. The term "earth bulging" is often applied to the refractive effect.

Fog which occurs very close to the ground in the early morning, usually in a valley immediately over a small stream, has quite another effect. In this case, the normal beam is in the clear, but the surface at the fog layer is a smooth strata which forms a good reflector of microwave energy.

Attenuation by Atmospheric Gases

The principal gaseous absorption is by oxygen and water vapor. The attenuation due to oxygen is relatively constant in the 2 GHz to 14 GHz frequency range, and is slightly under 0.01 dB/mile at 2 GHz and slightly above at the 14 GHz end. Water vapor absorption, on the other hand, is highly dependent on the frequency (as well as the density of water vapor). It is extremely low at 2 GHz, on the order of 0.0002 dB/mile, and still negligible, on the order of 0.002 dB/mile, in the 8 GHz range. But at about 14 GHz it is approximately equal to the oxygen absorption, at about 0.01 dB/mile, and at frequencies in the vicinity of 20 GHz it is up to nearly 0.2 dB/mile. These data are taken from NBS Monograph No. 92, "Radio Meteorology", and Table C is also based on that source.

Table C. Excess Attenuation Due To Atmospheric Absorption

PATH LENGTH MILES	ATTENUATION – dB				
	2-4-6 GHz	8 GHz	10 GHz	12 GHz	14 GHz
20	0.20	0.26	0.32	0.38	0.48
40	0.40	0.52	0.64	0.76	0.96
60	0.60	0.78	0.96	1.14	1.44
80	0.80	1.04	1.28	1.52	1.92
100	1.00	1.30	1.60	1.90	2.40

It is seen that the assumption of "free space" propagation through the atmosphere is reasonably well justified for paths up to about 50 miles from 2 through 8 GHz, or for paths up to about 20 miles for 10 through 14 GHz, but for longer paths, the gaseous absorption loss should be taken into consideration. Unlike free-space loss, this loss is directly proportional to the length of the path. The water vapor absorption will change with the density of water vapor in the atmosphere, consequently the bands of 10 GHz and up, which have significant amounts of attenuation from this source, will vary with the vapor density.

7. Clearance Criteria

For practical calculation purposes, based on the experience of many users, K is usually considered to fall within a range from about infinity in the supernormal direction (flat earth condition) to about 2/3 in the subnormal direction, with normal or "standard atmosphere" being taken as K = 4/3.

Excursions beyond these upper and lower limits do occur in some areas, but on rare occasions and usually for quite small time intervals. Published data based on point meteorological soundings show considerably higher percentages of time at the extreme ends, particularly in some difficult areas (hot, humid) such as the Gulf Coast, but practical experience indicates that the extreme values shown by such point measurements apply only to a small area and do not accurately represent what is likely to happen along an entire path at a given instant.

The choice of clearance criteria for a microwave route or path is an important one, since it can profoundly affect both the cost and the quality of performance. It is desirable on the one hand to get the antennas high enough so that even under subnormal refractive conditions there will still be adequate signal strength. Yet if the antenna heights are greater than actually needed, there can be an unwarranted increase in system cost and — for paths with significant ground reflections — an increase in multipath and ground reflective fading.

Normal Non-Reflective Paths

Although there are some variations, there are two basic sets of clearance criteria which are in common use in microwave communications systems. One is a "heavy route" set used for those systems with the most stringent reliability requirements, the other a "light route" set used for systems where some slight relaxation of the requirements can be made. The following are typical clearance criteria:

For "heavy route", highest-reliability systems
At least $0.3F_1$ at K = 2/3 and $1.0F_1$ at K = 4/3, whichever is greater. In areas of very difficult propagation, it may be necessary also to ensure a clearance of at least grazing at K = 1/2. (For 2 GHz paths above 36 miles, substitute $0.6F_1$ at K = 1.0).

Note that the evaluation should be carried out along the entire path and not just at the center. Earth bulge and Fresnel zone radii vary in a different way along the path, and it often happens that one criterion is controlling for obstacles near the center of the path and the other is controlling for obstacles near one end of the path.

For "light-route" systems with slightly less stringent reliability requirements
At least $0.6F_1$ + 10 feet at K = 1.0.

At points quite near the ends of the paths, the Fresnel zones and earth bulge become vanishingly small, but it is still necessary to maintain some minimum of perhaps 15 to 20 feet above all obstacles.

The "heavy route" criteria are on the conservative side except in the more difficult propagation areas, and undoubtedly in some cases will result in greater heights than actually needed. But it is difficult to predict accurately just what can happen on a given path under all conditions, and these criteria are well backed by experience. It should be noted, however, that even the heavy route criteria will not guarantee complete protection against the rare and unpredictable "blackout" fading which occurs in some areas.

Note: *A theory has been proposed recently at Lenkurt which, if experience backs it up, might provide protection against some forms of "blackout" fading. The theory is based on an assumption that even in difficult propagation areas the extreme gradients necessary to produce such blackout situations are likely to occur only very near the earth's surface. Some evidence seems to indicate that a layer of perhaps 100 to 150 feet above the earth's surface might encompass most of this difficult area. The theory then suggests that in areas where this kind of propagation anomaly is known to exist or might be expected, an additional criterion of at least 150' clearance above the earth's surface, all along the path, for a K value of 1.0 might be applied. Obviously on relatively short paths with no other clearance problems, this would require a considerable increase in tower heights, over those needed to meet the other criteria.*

Reflective Terrain

For non-reflective paths it is necessary only to provide sufficient clearance to ensure free-space propagation under normal conditions, and to meet minimum acceptable clearance under subnormal conditions. The same considerations apply to reflective paths, but with the added requirement to study the path for supernormal conditions, at least up to $K = \infty$ in most cases.

As discussed in earlier sections, it is possible in some situations to choose sites and antenna heights so as to provide screening or blocking of all potential reflective paths. In other situations this may not be possible. Where local conditions permit, the so-called high-low technique can be used to advantage to minimize the effect of reflections. There are some variations in this plan, and some pitfalls. Basically, the plan involves one adjacent repeater on a mountain ridge or very high point, and the other on a very low point, even on the flats if such is the nature of the surface. The point is, that provided the difference in elevation is sufficient considering the length of each path, the reflection "point" (center of locus of reflection) is moved in close to the low stations. The elevation of the antennas on the flats is quite low, and is computed to receive direct and not more than first Fresnel zone reflected energy most of the time. The transmit side works the same way, since the two transmission directions are symmetrical geometrically. As the value of K changes, the reflection point moves. As K becomes larger it moves toward the low site, and when it becomes smaller it moves toward the high site. A principal objective is to establish the low antenna elevation so that second and higher order Fresnel zone energy, that might otherwise reach the low antenna (receiving side), is blocked by the curvature of the earth at any expected values of K except near infinity.

Figure 7 can be used to locate the reflection point for a given path under various values of K. However, for this special case, because of the small value of x (or h_1), it is advisable to derive approximations from the curve, and then use the iterative procedure described in the text to find the exact values of n for various K. The critical parameters are the relative elevations h_1 and h_2, path length, frequency and the size of the antenna aperture. By experimenting with Figure 7, the limiting values for path geometry can be easily established in a given case. The high-low technique is a good solution where applicable, if the operating frequencies are in the 2 or 4 GHz range. At 6 GHz and above, the smaller Fresnel intervals in relation to the sizes of antenna aperture required for good transmission, would render the scheme questionable at 6 and 7 GHz without other external structures, and impracticable at 11 to 13 GHz.

The high-low technique can be used at all microwave frequencies if the reflection point, under all values of K, can be placed in rough non-reflective terrain. It should be noted that under this arrangement, the reflection point moves over a considerable distance with different values of K. For instance; a 30 mile path, with antennas at the sites at elevations of 10 feet and 1000 feet (relative to the reflective surface), will have the reflection point for K = 2/3 at 1.35 miles from the low site, but for $K = \infty$ the point will be approximately 0.3 miles from the low site. In many cases, advantage may be taken of the existence of low ridges or other irregular obstacles. Figure 6 illustrates such a case.

Open, Rolling Terrain

In open terrain where a path may have trees and other obstacles at some points, and barren, rolling hills at others, the obstacles and high terrain should be considered for clearance based on the criteria indicated above. Additionally, the barren hills should be considered as possible reflectors, and determinations made within the assumed limits of K as to whether serious even zone reflections could be experienced. Frequently adjustments in antenna elevations can be made to eliminate such possibility without seriously affecting terrain clearance at the assumed lower limit of K. There is an economy in considering the reflection possibility when original antenna elevations are established, even though only a small percentage of the cases considered may turn out to be seriously reflective.

8. Delay Distortion

Delay distortion may be caused by the radio path, waveguide system or the radio equipment. The end product is noise distortion in the message, data or television channel assigned to the baseband. It can be particularly destructive of data service assigned to part of the baseband, when message or data service is assigned to other parts of the baseband, because of the cross-modulation noise peaks that cause data errors.

In the propagation path, delay distortion is caused by reflected energy which reaches the receiving antenna, but is delayed by a number of wavelengths as compared to the direct signal. In this case it is not the instantaneous phase, but the actual time delay, that causes the delay distortion. It can be detected by delay sweep instrumentation, and is measured in nanoseconds. The critical amount of delay that can be tolerated without serious distortion depends upon the top frequency of the baseband. As discussed under ROUTE AND SITE SELECTION, the typical situations to avoid in order to minimize delay distortion in the propagation path, are paths which are mountain top to mountain top with low, flat terrain between, and paths that go through areas of tall buildings; particularly where the path aligns with the street. All paths which may result in terrain clearance of F_{60} (60th Fresnel zone) or more over flat terrain, should be examined for possible delay distortion. Building reflections are difficult to compute in advance, but situations which might indicate this possibility should be avoided because of reflection fading and interference, as well as delay distortion.

Waveguide echoes are another source of delay distortion. They result from impedance mismatches or irregularities. From the systems engineering viewpoint, the principal precautions are avoiding excessive waveguide lengths, and minimizing the amount and number of flexible waveguide sections. The latter have a tendency under the stresses of installation or maintenance to produce mismatches which, combined with slight mismatches at the antenna, will produce round trip echoes, causing delay distortion.

9. Fading

The beam of microwave energy is not a single line, but a wavefront extending for a considerable distance about the center line. Since the index of refraction under normal atmospheric conditions is lower at the top of the wave front and higher at the bottom, and since velocity is inversely proportional to the index of refraction, the upper portion of the wavefront under such conditions will travel slightly faster, with the result that the wavefront as it moves along the path will tend to have the top tilted more and more forward. Since the direction of beam travel is always perpendicular to the wave-front, the beam itself will be bent downward, thus increasing the apparent clearance. The amount of bending is actually very slight on a percentage basis, but is sufficient to cause significant variations.

Under certain atmospheric situations there can be even greater than normal negative N gradients (supernormal, or "earth flattening" type), or others in which the N gradients become less negative, or even positive. In the latter situation the lower part of the wavefront will travel faster, and the beam will be bent upward, reducing the apparent clearance. This is the subnormal, or "earth-bulging" type.

Most of the time these gradients in the lower atmosphere are essentially linear. These linear variations affect clearance, and are also important when the path is reflective, but they do not produce atmospheric multipath situations.

However, when non-linear gradients such as shown on c and d, and particularly e and f of Figure 16 occur, it is possible for multiple paths, in addition to the direct path, to exist within the atmosphere itself, independent of any reflecting surface.

These "kinks" in the atmosphere can occur when conditions are such that stratified layers with different gradients may lie on top of one another, much like a layer cake. The familiar smog-producing temperature inversion is an example.

These conditions typically are most likely to occur on hot, still, humid wind-free nights when temperature inversions are in existence. Under normal daytime conditions, temperature is greatest near the ground and decreases with altitude, a condition which leads to convection with the rising air keeping the atmosphere well mixed. But at night, radiation can cool the ground more rapidly than the air, and the temperature may then increase with increasing altitude. This is a stable condition and allows the stratifications to occur. These conditions can also happen in daytime, but are much less likely.

When the appropriate conditions exist on a path, it is possible to get the so-called atmospheric multipath, in which 2, 3, 4 or even many more distinct signal paths may exist between transmit antenna and receive antenna. Under these conditions the received signal is the vector sum of the various components, all of which are varying in phase in a random manner, and usually in amplitude as well. In such a situation there will be short intervals in which the various vectors will effectively cancel each other to produce a null. It is this phenomenon, often accentuated by some ground reflection complications, which causes most of the fast, very deep fading experienced on many microwave links.

Fading due to ground reflection phenomena is not confined to water and perfectly smooth, flat surfaces such as dry lake beds or salt flats, although such surfaces approach the classical reflection coefficient of -1, indicating a perfect reflector. Our common experience will show that two different surfaces will reflect visible light even though, when examined under a microscope, one is much rougher than the other. Such is also the case with microwave energy, which will be reflected by different surfaces with somewhat different reflection coefficients. Moreover, the effective reflection coefficient is affected by angle of incidence, and by the wavelength of the microwave energy. Additionally, from a quantitative standpoint, the effective area of the reflecting surface with just the right angle of incidence, is a measure of the reflected energy reaching the receiving antenna. Experience has shown that rolling prairie, such as that existing in some of the midwestern United States, can have a fairly high reflection coefficient, and may produce fades from one reflection surface, as shown by path tests, of 20 to 25 dB when 4 and 6 GHz CW test signals are used. Two or more reflection surfaces in a path, which, under some combinations of antenna elevations and specific values of K, produce coincident even zone reflections, can cause even deeper fades.

Growing crops such as alfalfa, can produce serious fading with early morning dew, especially if on low, flat irrigated land. Wheat stubble (depending partly on the planting orientation) and new wheat in the early growing period, provide insufficient roughness to break up a reflection pattern. All low, flat areas, and rolling country without trees, heavy brush or obstacles, should be considered for ground reflection possibilities when analyzing a path for the determination of antenna elevation, and for space diversity intervals. The combination of ground reflection fading and atmospheric multipath fading can be particularly severe.

This brief and very sketchy discussion of a very complicated process is intended only to give a general familiarity with the phenomena causing the fading effects. Ordinarily in line-of-sight microwave work it is not necessary to calculate or measure the actual variations in the atmosphere or the index of refraction.

Insofar as the clearance portion is concerned, the accepted practice is to assume a range of variations based on documented experience, and to select a set of clearance criteria appropriate to the type of service and the area.

The treatment of multipath fading is also based largely on experience. This type of fading has been found to follow distributions which are generally related to the well-known Rayleigh distribution. The latter or some modification is widely used in estimating propagation reliability, and in estimating the improvements which are attainable by the use of suitable diversity methods.

It is generally agreed that multipath fading tends to be greater on long paths than on short ones, and also to be somewhat greater at the higher frequencies. The Rayleigh distribution is often taken as the limiting value for multipath fading on line-of-sight paths with adequate clearance. One way of estimating reliability is to make a "worst case" assumption that a path will have continuous Rayleigh-distributed fading. This distribution has a slope of 10 dB per decade of percentage of time. A path with this fading distribution would have 20 dB fades for 1.0% of the time, 30 dB fades for 0.1% of the time, and 40 dB fades for 0.01% of the time. Continuous Rayleigh-distributed fading is unlikely to occur on most paths, and the assumption is therefore very much on the conservative side. It does, however, allow some leeway for the effects of combinations of attenuation fading and multipath fading, which can be much worse than either one alone.

The incidence of multipath fading varies not only as a function of path length and frequency, but also as a function of climate and terrain conditions. In the most favorable areas, for example paths in dry windy mountainous areas, it may be essentially non-existent. Hot, humid coastal areas typically have a high incidence of multipath fading, and inland temperate areas are somewhere in between. Flat terrain along a path tends to increase the probability of fading, while irregular or hilly terrain tends to reduce it.

Over the course of the years a number of approaches have been developed for calculating or estimating the distribution of deep fading for a microwave path.

In a following section we develop methods of calculating the annual outage probability for a line-of-sight microwave path, as a function of the pertinent parameters and conditions of the path.

These methods are based on relatively new experimental and theoretical results reported by W.T. Barnett, of Bell Telephone Laboratories, at an URSI meeting in Washington, D.C., in April, 1969, and by Arvids Vigants, also of Bell Laboratories, in several published papers. (See Page 119).

Barnett's work was in two parts. One described ways of calculating the outage due to fading on a non-diversity path as a function of terrain, climate, path length, and fade margin. The other gave formulas for calculating the effective improvement achievable by frequency diversity, as a function of the spacing interval and the frequency band. Vigants' work gave formulas for calculating the effective improvement achievable by vertical space diversity, as a function of the spacing in feet, the path length, and the frequency.

Since these studies were based on a relatively limited number of paths, the generalization to other paths, other frequencies, and other areas involves some degree of risk. It may be noted, however, that the Barnett and Vigants formulas give results which are considerably more pessimistic (conservative) than similar formulas developed by Japanese investigators and reported in the literature.

10. Propagation Reliability and Diversity Considerations

Table D provides a simple means of translating a given system reliability percentage into terms which are more easily related to experience. For example, the 99.99% value would correspond to about 53 minutes of outage time per year, while the 99.9999% value would amount to only about 32 seconds per year. The latter value is typical of per path objectives for the highest reliability systems.

It was indicated in a previous section that diversity techniques, when properly applied, can reduce the effects of multipath fading on line-of-sight systems to insignificance. Whether or not the considerable expense of providing diversity is justified will depend very critically on the nature of the communications and the degree of outage which is acceptable.

By providing adequate path clearance to essentially eliminate outages due to earth blocking (which diversity does not help in any event), and by providing fade margins of 40 dB or more, it is possible to achieve per path propagation reliabilities, with respect to Rayleigh-distributed fading, on the order of 99.99% or better without diversity. For many types of service this may be adequate. But for long systems, and particularly for systems carrying data, it is almost mandatory to employ diversity if very high reliability is needed.

A point of considerable significance in connection with multipath fading, is that its potential for causing data errors is a function of the *number* of interruptions as well as total interruption time. A great many very short interruptions because of deep multipath fades would be far worse than the same total time if it were only a single interruption. As a result of this phenomenon, and the greatly increased use of systems for data as well as voice, diversity protection may be desirable even if the total time reliability objectives did not require it. Diversity reduces the number as well as the total time of the multipath propagation outages.

Table D. Relationship Between System Reliability And Outage Time

RELIABILITY %	OUTAGE TIME %	OUTAGE TIME PER		
		YEAR	MONTH (Avg.)	DAY (Avg.)
0	100	8760 hours	720 hours	24 hours
50	50	4380 hours	360 hours	12 hours
80	20	1752 hours	144 hours	4.8 hours
90	10	876 hours	72 hours	2.4 hours
95	5	438 hours	36 hours	1.2 hours
98	2	175 hours	14 hours	29 minutes
99	1	88 hours	7 hours	14.4 minutes
99.9	0.1	8.8 hours	43 minutes	1.44 minutes
99.99	0.01	53 minutes	4.3 minutes	8.6 seconds
99.999	0.001	5.3 minutes	26 seconds	0.86 seconds
99.9999	0.0001	32 seconds	2.6 seconds	0.086 seconds

Diversity Arrangements

For routes which have two or more parallel working channels (for example, the TD2 or TH mainline routes), the protection arrangements are of the so-called 1-for-N or 2-for-N type, where one protection channel protects on a sectional basis against either equipment failures or selective fading in any one of several working channels. In effect, this is a form of frequency diversity, though considerably more efficient in usage of spectrum than the straight frequency diversity arrangement, which requires two RF channels for each working path. There are at present no practical ways of using space diversity alone to provide both equipment and propagation protection on a multiple channel switching section, consequently the 1-for-N or 2-for-N systems are the only available method of handling this situation.

For systems requiring only one working RF channel, as typical of most industrial systems and a great many of the military systems, the most commonly used basic protection methods are:

Frequency diversity, either in-band or cross-band, applied on a per hop basis, with post-detection combining or selection.

Vertical space diversity, applied on a per hop basis, with post-detection combining or selection.

Hybrid diversity, a special combination of frequency and space diversity.

The frequency diversity arrangement provides full and simple equipment redundancy, and has the great operational advantage of two complete end to end electrical paths, so that full testing can be done without interrupting service. Its disadvantage is that it doubles the amount of spectrum required. Also it is sometimes prohibited by the licensing authority; it is not, for example, available to industrial users in the U.S.A.

The space diversity arrangement can also provide full equipment redundancy (when automatically-switched hot standby transmitters are used), but does not provide a separate end to end operational path. Because of the requirement for additional antennas and waveguide, it is more expensive than the frequency diversity arrangement. However, it provides efficient spectrum usage, and extremely good diversity protection, in many cases substantially greater than obtainable with frequency diversity, particularly when the latter is limited to small frequency spacing intervals.

The somewhat specialized form of diversity, which we have called "hybrid diversity", consists of an otherwise standard frequency diversity path, in which the two T—R pairs at one end of the path are separated from each other, and connected to separate antennas which are vertically spaced as in space diversity. This arrangement provides a space diversity effect in both directions; in one direction because the receivers are vertically spaced, and in the other direction because the transmitters are vertically spaced. This arrangement combines the operational advantages of frequency diversity with the improved diversity protection (particularly on reflective or difficult paths) of space diversity. It has, of course, the same disadvantage as ordinary frequency diversity, in that it requires two RF frequencies to obtain one working channel.

Because of growing congestion in the microwave bands, the use of vertical space diversity has increased tremendously in the past few years. This change has come about largely as a result of its demonstrated effectiveness in long haul industrial systems in the United States. Space diversity is particularly effective against ground or water-reflective fading, and can even be arranged in particular instances to provide "anti-correlated" fading, in which a fade on one diversity half is accompanied by an actual signal rise on the other diversity half. Space diversity is also quite effective against atmospheric multipath fading.

The amount of improvement against interference fading (an all-inclusive name for multipath or selective fading) which will be provided by diversity, depends on the distribution of the fading on each diversity half, and on the degree of correlation between the two distributions. If each diversity half has Rayleigh-distributed fading, the correlation coefficient between the two halves can be any value from 0 to 1. A correlation coefficient of 0 would mean completely independent fading on the two halves, and a correlation coefficient of 1.0 would mean that the two halves faded identically. A coefficient of 0.0 corresponds to uncorrelated or independent fading. It is often referred to as a "Rayleigh-squared" distribution, since, in this case, the probability that both halves will simultaneously fade below a given level, is equal to the square of the probability that either half alone will fade below that level.

The diversity improvement at the higher fade margins is startlingly large for uncorrelated fading, and remains surprisingly good even when the degree of correlation is high. For example; at a fade margin of 40 dB, diversity with 0.99 correlation would improve the path reliability by a factor of 100, from 99.99% to 99.9999%. At this same fade margin, diversity with 0.0 correlation would show a calculated improvement by a factor of 10,000!!

In order to estimate or calculate the reliability of a diversity system it is necessary to know or make some assumption about the value of the correlation coefficient, for the particular path conditions.

Until fairly recently it has been a common practice to assume essentially zero correlation for frequency diversity with spacings of 5% or more, and a correlation of around 0.8 for the more common spacing interval of 2%. Barnett's data, supported by that reported by others, indicates that the correlation is far higher and the diversity improvement far less than would correspond to these values.

On the other hand, Vigants' data on space diversity improvement indicates that the commonly assumed 100 to 1 improvement, with a 40 dB fade margin and spacings of about 30 to 40 feet, is a valid and somewhat conservative assumption.

With 40 dB fade margins, the 100 to 1 improvement indicated by this assumption is sufficient to provide estimated path reliabilities of 99.9999%, even if there were continuous Rayleigh fading on every path.

It should be noted that considerably lower space diversity correlation coefficients, with consequently much larger diversity improvements, have been reported in the literature. For example, a Japanese report to CCIR gave a semi-empirical equation for space diversity correlation coefficient which produces coefficients on the order of 0.6 to 0.7 for the spacings suggested in this document, and even lower coefficients for greater spacings. At the 40 dB fade margin level, such correlations would indicate diversity improvements in excess of 1000 to 1; an order of magnitude greater than our conservatively assumed value of 100 to 1.

In frequency diversity operation, band allocations and frequency patterns are relatively rigid and there is little freedom to choose or vary the amount of spacings. Typical diversity spacings in most microwave bands are on the order of 2%, and even less in some cases.

Space diversity applications allow essentially full freedom of choice as to spacing interval, subject only to economic or physical limitations. The method of choosing a spacing interval depends primarily on a judgment as to whether the significant multipath fading on the particular path is likely to be of the atmospheric type or the ground reflected type. For the atmospheric type of multipath, the diversity improvement tends to increase as the vertical separation increases, very rapidly at first and then more slowly. With this situation it is not necessary to use discrete, calculated spacings, but simply to ensure that the spacing is at least equal to some chosen minimum value. Experience has indicated that, on most overland paths, excellent diversity will be obtained with minimum vertical spacing intervals of about 60' at 2 GHz, 45' at 4 GHz, 30' at 6 GHz, and 15' to 20' at 12 GHz. Larger intervals can be expected to provide even better diversity action, but may impose undesirable problems in tower heights, clearances, etc. A spacing of 40' in the 6 GHz bands, chosen from physical as well as propagational considerations, has been widely used by Lenkurt with extremely good results.

Reflective Path Diversity Spacings

For overwater paths with unblocked reflective paths, or for overland paths with known surfaces of high reflectivity, it is possible to calculate discrete vertical spacings which will provide improved diversity action against ground-reflected multipath fading. In fact, if this is the only kind of fading on a path, spacings can often be chosen so that over a considerable range of variation of K, one or the other of the two diversity signals will always be close to the free space value. The basic principle is to choose an interval such that when the signal on one diversity half is at or near a null point, the signal on the other diversity half will be at or near one of the adjacent maximum points. For example; referring to Figure 14, we might choose diversity antenna heights on a particular path in such a way that, under normal atmosphere ($K = 4/3$), the upper antenna has 5th Fresnel zone clearance over the reflection point, and the lower antenna has 4th Fresnel zone clearance. This would mean that under normal conditions the upper antenna would have a strong signal and the lower one would have a faded signal, the depth depending on the amplitude of the reflection. But if K moved in either direction, the signal on the upper antenna would go down and that on the lower antenna would rise. When the upper antenna hit a null at either the 6th or 4th zone, the lower would be near a maximum at the 5th or the 3rd zone. In practice, it is common to use spacings which, at $K = 4/3$, are somewhat less than that between an adjacent even and odd zone. The purpose is to avoid a spacing which might, as K increases toward infinity and the separation between Fresnel zones gets smaller and smaller, reach a point where it corresponds to the separation between two even zones, and thus allow both antennas to be in a deep simultaneous fade.

The following procedure can be used to calculate the desired vertical spacing interval to protect against reflected path fading up to $K = \infty$. The procedure provides calculations for a spacing which, under normal atmospheric conditions ($K = 4/3$), will have the upper and lower path clearances differ by one Fresnel zone or less, and for the super-refraction condition of $K = \infty$, will have the two clearances differ by no more than 85% of *two* Fresnel zones. The latter condition prevents both paths from simultaneously reaching an even Fresnel zone null, up to $K = \infty$. Both situations are calculated, and the smaller of the two spacings is chosen. A difference of one Fresnel zone at $K = 4/3$ provides "anti-correlation", that is, one signal goes up as the other goes down.

Step A: Establish the antenna heights needed to meet the required path clearance criteria.

Step B: Let h_1 = the difference in elevation between the antenna at one end of the path and the reflecting surface, h_2 = the difference in elevation between the antenna at the other end and the reflecting surface, and h_t = the height of the transmitting antenna above a plane tangent to the earth at the point of reflection.

Step C: For $K = \infty$, first consider the h_1 end as the transmitting site and calculate a tentative value for the receiver spacing at the *other end* of the path, using the following formula: (This formula makes the reflected path 1.7 half-wavelengths — slightly less than two half-wavelengths — longer than the direct path at $K = \infty$.)

$$\Delta h_2 = \frac{2.2 \times 10^3 \ D}{F_{GHz} \times h_t} \qquad (10A)$$

where Δh_2 = diversity spacing at the h_2 end, in feet.

h_t = height of the *transmitting* antenna, in feet, at the h_1 end, above a plane tangent to the earth at the point of reflection. (For "flat earth", $K = \infty$, this is simply the elevation above the reflection point and $h_t = h_1$.)

F_{GHz} = frequency in GHz

D = path length in miles

Step D: In most cases, the final spacing will be that calculated in Step C, but to ensure that it does not result in spacings which would give more than a one-half wavelength path difference at K = 4/3, the following calculation is also made using the same symbols as in (10A)

$$\Delta h_2 = \frac{1.3 \times 10^3 \ D}{F_{GHz} \times h_t} \qquad (10B)$$

Note, however, that h_t in this formula will not be equal to h_1, because of the earth curvature.

In order to determine this h_t, it will be necessary to locate the reflection point for K = 4/3, by the methods described in an earlier section. h_t will then be equal to h_1 minus the earth curvature corresponding to the distance between the reflection point and the h_1 end.

That is, $h_{t1} = h_1 - \frac{d_1^2}{2}$.

where d_1 is the distance, in miles, from the h_1 end.

Step E: The desired vertical spacing at the h_2 end will be the smaller of the two calculated values of Δh_2.

Step F: The above steps cover one direction of transmission and give a calculated spacing between receive antennas at one end of the path. Except in the case of symmetrical antenna heights, the spacings will be different at the two ends of the path. Step F, therefore, consists of repeating Steps A through D for the other direction of transmission.

Cautionary Note:

Depending on the path configuration, the above calculations can, at times, produce spacings which are either very large or very small. In such cases the method may become impractical, and one would simply revert to choosing spacings adequate to provide protection against atmospheric multipath. Such arbitrary spacings will almost always also provide a large measure of protection against ground reflected multipath, or a combination of the two types.

11. Methods of Calculating the Probability of Outages due to Propagation (See Page 119)

These methods are based on the previously mentioned work of Barnett and Vigants. In this section we will, for mathematical convenience, use fractional probability (per unit) rather than percentage probability, and will deal with the "unavailability" or outage parameter, designated by the symbol U. The "availability" parameter, for which we use the symbol A, is given by (1—U). "Reliability", in percent, as commonly used in microwave circles, is given by 100A, or 100 (1—U).

Non-Diversity Annual Outages

Let U_{ndp} be the non-diversity annual outage probability for a given path.

We start with a term r, defined by Barnett as follows:

$$r = \frac{\text{actual fade probability}}{\text{Rayleigh fade probability } (= 10^{-F/10})} \qquad (11A)$$

For the worst month

$$r_m = a \times 10^{-5} \times (f/4)^{1.5} \times D^3 \qquad (11B)$$

D = path length in miles
f = frequency in GHz
a = 4: for very smooth terrain, including overwater,
1: for average terrain, with some roughness,
¼: for mountainous, very rough, or very dry.
F = fade margin, to the "minimum acceptable" point, in dB.

Over a year

$$r_{yr} = b \times r_m \qquad (11C)$$

b = 1/2: Gulf coast or similar hot, humid areas,
1/4: normal interior temperate or northern,
1/8: mountainous or very dry

By combining the three equations and noting that U_{ndp} is equal to the "actual fade probability" for a given fade margin F, we can write

$$U_{ndp} = r_{yr} \times 10^{-F/10} = b \times r_m \times 10^{-F/10},$$

or (12)

$$U_{ndp} = a \times b \times 1.25 \times 10^{-6} \times f^{1.5} \times D^3 \times 10^{-F/10}$$

The product of the terrain and climate factors, a x b, in this equation, ranges from a maximum of 4 x 1/2 = 2, for very smooth paths and hot, humid climate, to a minimum of 1/4 x 1/8 = .031, for mountainous or very rough, dry paths. This is a range of 64 to 1 in outages, between the worst and the best situations. "Normal" or average paths, with some roughness would have a x b = .25, halfway between the two extremes.

The method we will use here for calculating the outages for a system with diversity will be to calculate separately the non-diversity outage for the path, and a "diversity improvement factor", for which we will use the symbol I. The diversity outage or fade probability will be given by:

$$U_{div} = \frac{U_{ndp}}{I} \qquad (13)$$

Frequency Diversity Improvement Factor

Barnett has defined two "frequency diversity improvement factors", one for the 4 GHz common carrier band and another for the 6 GHz common carrier band. Those are experimental formulas derived by curve fitting of actual measured data in the deep fade regions.

The formulas are:

$$I_{fd(4)} = 1/2 \times \left[\frac{\Delta f}{f}\right] \times 10^{F/10} \qquad (14A)$$

$$I_{fd(6)} = 1/4 \times \left[\frac{\Delta f}{f}\right] \times 10^{F/10} \qquad (14B)$$

where f is the frequency and Δf the diversity spacing, and
F is the fade margin in dB.

Unfortunately Barnett's data, though extensive, covers only these two bands, and there is nothing to indicate how to extend it to other bands. Furthermore, his data at present covers only paths of average length (25 to 30 miles or so) and do not indicate whether there is any distance dependency.

The two formulas do indicate that a given frequency spacing in percent gives twice as much improvement at 4 GHz as at 6 GHz. This is a startling and unexpected phenomenon, and if the same dependence continues on to the higher bands would indicate progressively poorer diversity performance with increasing frequency.

As an "educated guess", the following formulas (with no experimental data) are suggested:

$$I_{fd(7\text{-}8)} = 1/8 \times \left[\frac{\Delta f}{f}\right] \times 10^{F/10} \qquad (14C)$$

$$I_{fd(11\text{-}12)} = 1/12 \times \left[\frac{\Delta f}{f}\right] \times 10^{F/10} \qquad (14D)$$

Cross Band Diversity Improvement Factor

Since Barnett's work doesn't cover this situation, judgment must again be used. Experience indicates that for 6/11 or 6.7/12.4 GHz cross-band situations, the diversity improvement is at least as good as that achieved by 4% in-band spacing at 6 GHz. This corresponds to an improvement of 100 to 1, assuming fade margins of 40 dB. Hence, with 40 dB fade margins we can simply assume:

$$I_{cbd} = 100 \tag{15}$$

Space Diversity Improvement Factor

Vigants has defined a "space diversity improvement factor" which is a function of the path length, the frequency, the vertical spacing, and the fade margin. (He called it a 'fade reduction factor', but we will use the term 'improvement factor' to be consistent with the similar term in Barnett's work on frequency diversity).

In a modified form, Vigants' improvement factor can be written as:

$$I_{sd} = \frac{7.0 \times 10^{-5} \times f \times s^2 \times 10^{\overline{F}/10}}{D} \tag{16}$$

where
- f = frequency in GHz
- s = vertical antenna spacing, in feet, between centers
- D = path length in miles
- $\overline{F}$ = fade margin associated with the second antenna. The barred F is introduced to cover the situation where the fade margins are different on the upper and lower paths. In such a case F will be taken as the larger of the two fade margins and will be used in calculating U_{nd} for the path. $\overline{F}$ will be taken as the smaller and will be used in the calculation of I_{sd}.

Hybrid Diversity Improvement Factor

Experience indicates that the improvement in hybrid diversity systems is mainly due to the space diversity effect. Consequently, we assume that:

$$I_{hybrid} = I_{sd} \tag{17}$$

and the improvement factor is calculated as if the path were straight space diversity.

Correlation Coefficients

Although our calculations do not require the correlation coefficients, they are of some theoretical interest. The correlation coefficient is related to the improvement factors as follows:

$$k^2 = 1 - \frac{I}{10^{F/10}} \tag{18}$$

where
- k^2 is the correlation coefficient
- I the diversity improvement factor, and
- F the fade margin.

Note: All of the above formulas are valid only in the situations where the calculated value of I is at least 10. If the calculations indicate that I is less than 10, the diversity improvement will be somewhat better than shown by the calculations. A further point: the formulas apply only to the deep fading regions, that is, fades of 20 dB or more. They cannot be used to calculate the incidence of low level fading.

Example: To illustrate the method, consider a 30 mile path with average terrain, with some roughness, in an inland temperate climate, operating at a frequency of 6.7 GHz with a fade margin of 40 dB. we will make calculations for this path without diversity, with 2% frequency diversity, and with 40 foot vertical space diversity.

Non-Diversity Case

Using (12), with the proper values for the various factors, we have

$$U_{ndp} = 1 \times 1/4 \times 1.25 \times 10^{-6} \times (6.7)^{1.5} \times 30^3 \times 10^{-4} = .0000148$$

This would correspond to an A of .9999852, or a "reliability" of 99.99852%.

Frequency Diversity Case

Using (14C), we have

$$I_{fd(7\text{-}8)} = 1/8 \times (.02) \times 10^4 = 25$$

so that, substituting in (13),

$$U_{fdp} = \frac{.0000148}{25} = .00000059$$

Thus, the calculated "frequency diversity improvement" is 25 to 1, and the calculated reliability for frequency diversity is 99.999941%. (If we had used (14B) instead of (14C), the calculated improvement factor would have been 50, twice as great as shown).

Space Diversity Case

Using (16), we have

$$I_{sd} = 7.0 \times 10^{-5} \times 6.7 \times 1600 \times 10^4 \times \frac{1}{30}$$
$$= 250$$

and substituting in (13), we have

$$U_{sdp} = \frac{.0000148}{250} = .000000059$$

Thus, the calculated "space diversity improvement" is 250 to 1, and the calculated reliability for space diversity is 99.9999941%.

The superiority of space diversity for this situation is clearly evident, but it is important to note that despite the 10 to 1 difference, the reliability associated with the frequency diversity is still extremely high, and would be quite adequate for very high reliability systems.

On the other hand, if the path were located in the "worst case" propagation area, all of the calculated outages would be increased by a factor of 8, which would put the frequency diversity reliability at 99.99953%, perhaps slightly marginal for ultra-high reliability systems.

Equations 11 through 18 provide a complete analytical fading model which can easily be set up for routine machine computation. It is also relatively easy to do the computations by hand. Hand computation can be facilitated by preparing tables of several of the factors. For example, the factor $(f)^{1.5}$ is 2.82 for 2 GHz, 8 for 4 GHz, 11.2 for 5 GHz, 15.4 for 6.2 GHz, 17.4 for 6.7 GHz, and 37.5 for 11.2 GHz, and so on. The factor $10^{F/10}$ is equal to approximately $.63 \times 10^4$ for a 38 dB fade margin, $.8 \times 10^4$ for a 39 dB fade margin, 1×10^4 for a 40 dB fade margin, and so on. The factor $\frac{\Delta f}{f}$ is simply the frequency diversity spacing, in per cent, divided by 100. By making up a few tables covering the pertinent range of parameters, the calculations can be done easily and quickly, using slide rule or logarithms. Extreme precision in the calculations is not warranted. Path lengths can be rounded to the nearest mile, and fade margins to nearest dB.

As an aid in visualizing the relationships between the various parameters and illustrating some of the characteristics of the fading probability distributions predicted by these theories and equations, two charts are presented. In both cases a frequency of 6.7 GHz is assumed.

Figure 21 shows the probability of outage vs. fade margin for non-diversity paths of various lengths, and for the same paths with 40' vertical space diversity. The Rayleigh and the Rayleigh-square distributions are also given, for information. Also, Figure 21 is for an "average-average" path. That is, a x b = 1 x 1/4 = 0.25.

A point of considerable interest, and one which differs significantly from some previous theories, is that the slope of the fading probability distribution for non-diversity paths in the deep fading region is the same as the Rayleigh slope, that is, 10 dB per decade of probability, and that the slope of the simultaneous fading probability distribution for diversity paths in that region is the same as the Rayleigh-squared slope, that is, 5 dB per decade of probability.

Figure 22 shows, for the same path and frequency, the probability of outage vs. path length for non-diversity with a 40 dB fade margin, and for the same paths with a 40 dB fade margin and 40' vertical space diversity.

Multiline Systems

The previous analysis applies only to diversity systems of the 1 for 1 type, with the diversity applied on a hop by hop basis. A more complex situation exists with respect to propagation outages in multiline systems. These systems have two or more working RF channels and one (sometimes

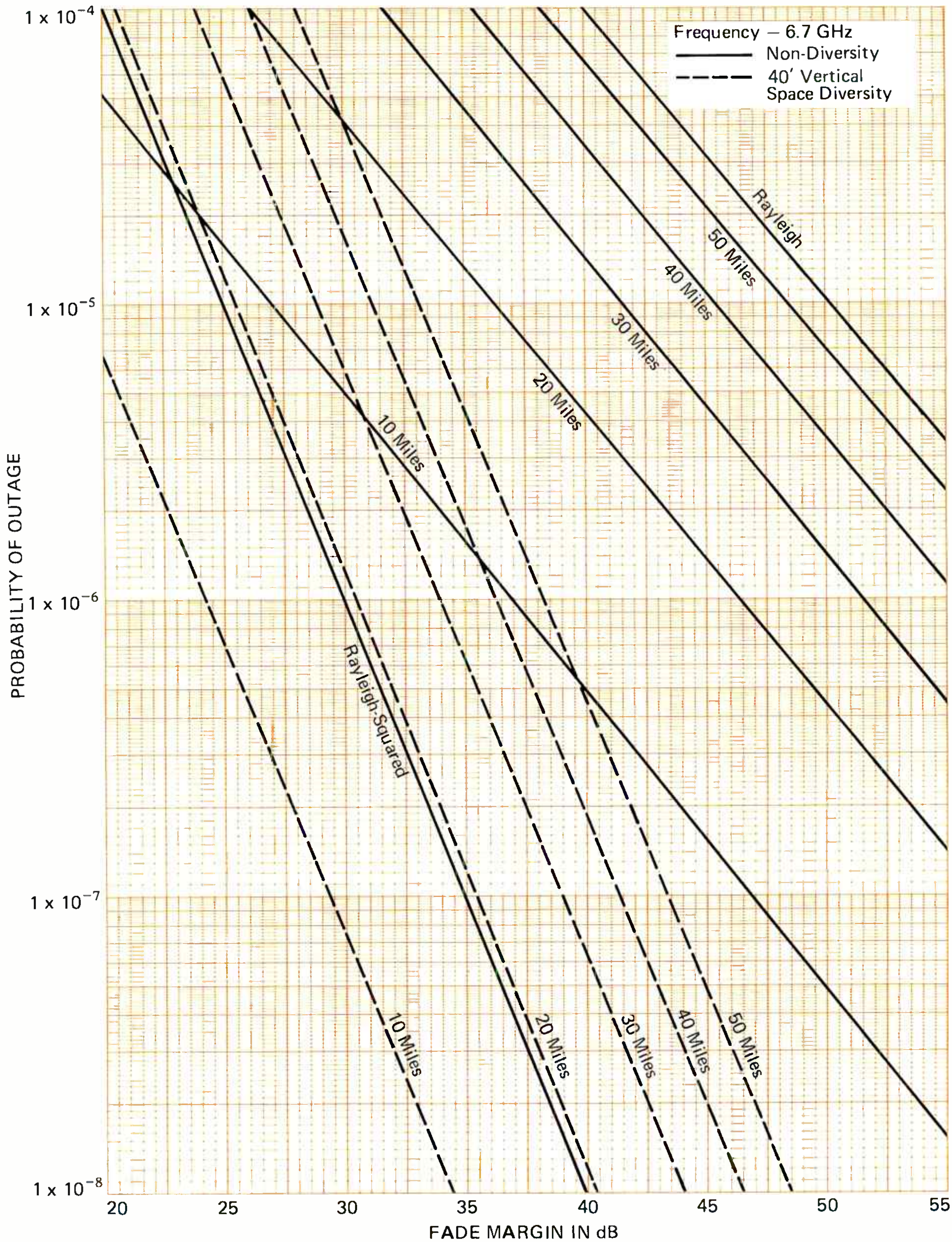


Figure 21. Outage Probability vs. Fade Margin for 6.7 GHz paths of Various Length, Average Terrain and Climate (after Vigants & Barnett)

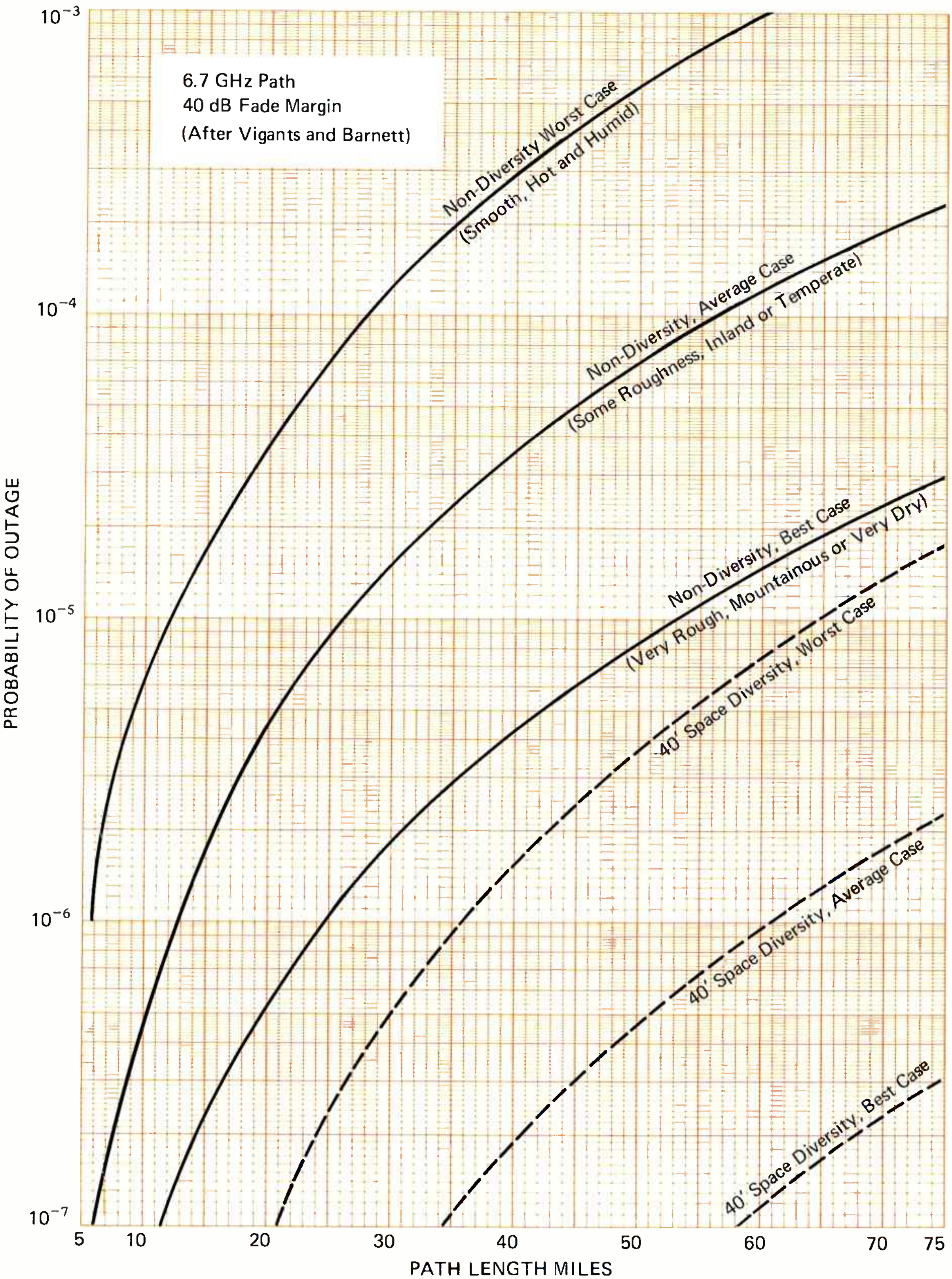


Figure 22. Outage Probability vs Path Length for a 6.7 GHz Path with 40 dB Fade Margin

two) protection RF channel operating in parallel over the path, each on its own frequency. The protection channels, besides giving protection against equipment failure, provide a form of frequency diversity protection against propagation failure.

The amount of diversity protection depends among other things on the frequency spacing between the closest spaced channels, and in the case of fully equipped systems in the 4 or 6 GHz common carrier bands, adjacent channels may be separated by as little as 0.5% in frequency. Considering two such closely spaced channels as a "frequency diversity pair", it is clear that the probability of simultaneous fading to threshold could be relatively high, that is, the fading highly correlated between the two channels.

The analysis of the complete propagation effects in multiline systems, particularly those utilizing 2 protection channels and arranged with switching sections comprising several hops in tandem, is too complex to be treated in this book. Comprehensive treatments can be found in the literature.

There is one aspect of the multiline situation which we can usefully treat. That aspect is the effect on the diversity action which results from switching or combining only at the end of a number of hops in tandem, rather than on every hop. The difference between the two lies in the fact that in the tandem situation, outages can be caused by simultaneous inter-hop as well as simultaneous intra-hop fades to thresholds. For example, a simultaneous fade to threshold on Channel 1 of Hop A and Channel 2 of Hop B could cause an outage in this situation, but not if the diversity were applied on a per hop basis. If the same degree of correlation existed for the deep fading on inter-hop combinations as on the intra-hop combinations, the outage with diversity applied end to end on an N-hop system would be N times that of the same system with diversity applied over each hop. But in actuality the fading distributions on inter-hop combinations, even for adjacent hops, must be essentially independent (zero correlation) because each path at a given instant has a unique set of multipath conditions which cannot possibly be the same as those of the other path. We thus have the situation that the fading for intra-hop combinations is highly correlated (k^2 in the range of .99 and higher) while the inter-hop combinations have zero or at least very low correlations. Stated another way, the "diversity improvement factors" for inter-hop combinations at a 40 dB fade margin would be in the range of 10,000 to 1, so that outages attributable to an inter-hop combination would be considerably less than 1% of the outages attributable to an intra-hop combination.

Thus, though there is a slight increase in outage probability for the tandem arrangement, it is far too small to be of any significance and from the propagation point of view either system can be treated, analytically, as if the diversity were on a per hop basis. In either case the total calculated outages for an N-hop system will simply be N times the calculated single hop outages. (In the case of very high reliability systems, the possibility of diversity outages occurring at the same instant on two different hops is extremely small and so can be neglected.)

Non-Selective Fading

The diversity improvement factors apply only to fading which is caused by some sort of multiple path conditions, since this sort of fading exhibits frequency and space selectivity. The improvement does not apply to those types of fading which are non-selective. On well-engineered paths (adequate clearance and fade margins) outages due to such non-selective fading are very unlikely to occur, but when and if they do occur they can have a significant effect on the overall availability because they tend to last much longer than multipath deep fades and because they receive no diversity improvement. No attempt has been made to include them in the analytical models, since no adequate prediction methods are available for them and since they do not apply to the great majority of paths.

Other Treatments

Several pertinent articles on reliability and diversity have been published in recent years in the IEEE Transactions on Communications Technology. Among them: February, 1966 — Barnett on reliability; December, 1966 — Abraham on reliability; August, 1967 — Makino and Morita on space diversity; February, 1968 — White on space diversity; December, 1968 — Vigants on space diversity.

D. Noise Performance

The noise performance of a communications system is one of the most significant parameters, with strong effects on many phases of system engineering.

1. Total Noise

The total noise in any derived channel is composed of noise contributions of several types including; thermal, intermodulation, echo path distortion, interference, and noise from the multiplex equipment.

Thermal Noise

Thermal noise is caused by random current variations in every portion of the electronic equipment and is present whether or not a modulating signal is being applied. One portion of the thermal noise, often called intrinsic or idle noise, is that generated in the transmitter and in the late stages of the receiver. It is independent of receiver input level, and is the limiting noise performance level which could be measured between terminals under conditions of no modulating signal and a very strong RF input signal.

The more important portion of the thermal noise includes noise generated by the antenna resistance, plus the noise generated in the front end circuits of the receiver. This noise undergoes amplification within the receiver, along with the RF carrier, and as a result of the FM process the noise at the output of the receiver will vary inversely with the RF carrier input level. For RF signals above the FM improvement threshold (to be discussed later), the thermal noise at the output of the receiver will decrease 1 dB for each 1 dB increase in RF signal input, up to the point at which the intrinsic noise, unrelated to the RF carrier, becomes controlling.

Intermodulation Noise

Intermodulation noise is created whenever the complex modulating signal passes through any kind of non-linearity of phase or amplitude in the transmission facility. It is present only when the system is being modulated, and increases with the level of the modulating signal. Such factors as the total number of active message channels, the level of signaling and data tones, and the individual speech levels, determine the level of the baseband signal, or the baseband load on the system.

Echo Distortion Noise

Echo distortion noise is a form of intermodulation noise which is created when delayed echo signals are present in the FM portion of the system. Its magnitude is a complex function of the relative magnitude of the delayed signal, its absolute delay with respect to the main signal, the amount of loading present, the width of the baseband, and the relative position of the channel in the baseband. It is more significant the higher the channel frequency, the greater the echo amplitude and the greater the delay. In high density systems with long waveguide runs (or long IF cable runs), it is necessary to maintain close control of impedance matching throughout (low VSWR of antennas, equipment, waveguide runs and all junctions) in order to keep distortion to acceptably low levels.

Path reflections can, in some unusual cases, also produce echo signals with sufficiently long delay as to cause significant path distortion. Such a condition could occur on a path with a relatively strong reflection and an abnormally large amount of clearance, or as a result of a double bounce off one building to another building and thence to the distant antenna. With good equipment design and proper system layout, path distortion noise can be kept to the very low levels needed to meet overall noise objectives.

Atmospheric and Man-Made Noise

The contribution to system noise from atmospheric and most forms of man-made noise is very small at microwave frequencies and can be neglected. However, interfering signals from other microwave systems, or from the spurious radiations of high powered radars, can produce noise in a microwave system. This form of noise must be kept to insignificant levels by proper equipment, systems design and by appropriate coordination of frequency usage in any geographical area, as discussed in more detail elsewhere.

Multiplex System Noise

The multiplex system is also a source of noise, but its noise contribution is not affected by fading, so for any given configuration it is relatively fixed. The amount of noise contributed by the multiplex system under loaded conditions, is a characteristic of the equipment, and can be determined either from the manufacturer's specifications, or from actual measurements on terminals connected back-to-back.

2. Noise Units

The most commonly specified noise parameter is that of noise power in a voice channel. It is defined and specified in a number of ways. Noise power units in common use include:

dBrnc (dB above reference noise, C-message weighting. Reference noise is equivalent to a 1,000 hertz tone at −90 dBm.)

dBa (dB above reference noise-adjusted, F1A weighting. Reference noise adjusted is equivalent to a 1,000 hertz tone at −85 dBm.)

pWp (picowatts of noise power, psophometrically weighted. 1.0 pWp is equivalent to a 800 hertz tone at −90 dBm.)

dBmp (psophometrically weighted noise power in dB, with respect to a power level equivalent to a 800 hertz tone at 0 dBm.)

S/N (signal-to-noise ratio in dB, either unweighted or with a specified weighting.)

The first four units as defined above represent absolute values of noise. In order to make them meaningful with respect to an actual circuit, it is necessary to take into account the relative level at the point of measurement. For measurements made at a point of 0 relative level, the absolute and relative values will be the same. Consequently, it is customary to express objectives and measurements in equivalent noise power at a point of zero relative level (0 dBm0 point), and to identify such objectives or measurements as dBrnc0, dBa0, pWp0 or dBm0p, respectively.

The dBrnc is the present noise unit used in the telephone industry in the United States. The dBa is no longer in general use in the telephone industry, but is still common among industrial users. The pWp and dBmp are international units based on CCIR recommendations. Table E shows the approximate relationships between these various noise units. Because of differences in weighting curves, the correlations are in some cases valid only for "white noise". The correlations shown, which are rounded off to integral values, are the ones in widest use, though slightly differing correlations are also found in the literature.

In most practical cases the variations due to the round-offs are unimportant, but where fractional dB's are significant (in meeting guaranteed performance requirements, for example), the following correlations are more precise:

$$\text{dBrnc0} = 10 \log_{10} \text{pWp0} + 0.8 = \text{dBa0} + 6.8 \tag{19}$$

$$= \text{dBm0p} + 90.8 = 88.3 - \text{S/N}$$

S/N

The conditions for precision are a frequency band of 300-3400 Hz, a square channel response, an ideal noise meter, and that the noise be essentially "white noise".

Note: The parameter "noise power ratio" (NPR) has had wide usage as another way of describing noise performance. But it cannot be considered as a "noise unit" in the same sense as the other parameters listed here, because its relationship to noise in the derived channel is not constant but depends on the relationship between the noise loading ratio and the bandwidth ratio being used. For this reason the use of NPR as a means of specifying noise performance is not recommended since it often leads to confusion. S/N − BLR − NLR

3. Determination of System Noise

The receiver "front end" noise is of particular significance in microwave system engineering because of its effect on thresholds and fading margin, in addition to its effect on overall noise. Fortunately, this type of noise and its effects in a derived channel are readily calculable from a knowledge of certain system parameters, including receiver noise figure (F) in dB, receiver IF bandwidth in megahertz at the 3 dB points (B_{MHz}), the per-channel deviation (Δf), the center frequency of the derived channel at the top of the baseband (f_{ch}), the effect of emphasis (if used), and the desired voice channel weighting characteristic.

Note: The receiver noise figure F and the various equations for noise developed in this and later sections are all based on the assumption that the noise temperature of the area in the field of view of the antenna is approximately 290° Kelvin. This is a sufficiently accurate approximation for any microwave path between two terrestrial points. But a different approach (using noise temperatures instead of noise figure) is needed for situations where the antenna "sees" a region of space with a much lower noise *temperature, such as an earth station antenna looking up at a high angle toward a satellite. For such situations the formulas in this book will give correct results if the noise figure F as used here is replaced by an "operational noise figure" defined as* $F_{op} = 10 \log_{10} \left(\frac{T_a + T_e}{290}\right)$, *where* T_a *is the noise temperature, in degrees Kelvin, associated with the antenna and* T_e *the noise temperature of the equipment. For the case where* $T_a = 290°$ *Kelvin,* F_{op} *is identical to F.*

Receiver Thermal Noise

The starting point for receiver thermal noise calculations is the thermal noise generated in the antenna resistance. For terrestrial microwave systems with an assumed effective antenna noise temperature of 290° Kelvin, the antenna noise transferred to the receiver has been calculated to be —174 dBm per cycle of bandwidth, or —114 dBm per megacycle of bandwidth. In a perfect receiver this would be the only source of "front end" noise, but any actual receiver will itself contribute additional noise, which will raise the equivalent noise input by the dB value of the receiver noise figure. Total equivalent noise input (N) in dBm can then be calculated as:

$$N = -114 + 10 \log_{10} B_{MHz} + F \qquad (20)$$

This is one kind of threshold, and is often called "detection threshold", "absolute noise threshold" and similar expressions. It should be clearly understood that, in an FM microwave system, this threshold does not represent a usable signal level. The true working threshold, often called the FM improvement threshold or the FM breaking point, occurs when the power of the signal is approximately 10 dB higher than the power of the noise. At this point the peaks of the signal begin to exceed the peaks of the noise and FM quieting begins. For input signals higher than this level, the thermal noise in a derived channel will decrease 1 dB for each 1 dB increase in RF input level.

If the input signal drops below the FM threshold, the noise in the derived channel rises quickly to an intolerable level. Consequently most microwave receivers are arranged to mute when the input level drops below this point. The maximum available fade margin in such a receiver is, therefore, the difference in dB between the normal unfaded signal and the FM improvement threshold. The FM improvement threshold (T_{FM}) can be calculated as:

$T_{FM} = N + 10db$

$$\therefore \quad T_{FM} = -104 + 10 \log_{10} B_{MHz} + F \qquad (21)$$

This is the point at which the RF carrier-to-noise ratio (C/N) is equal to 10 dB. It is notable that the FM threshold is independent of baseband frequency, deviation, emphasis, etc., but the noise at the FM threshold in a derived channel, which *is* a function of these parameters, is indeterminate until these parameters are known or specified.

The choices for these other parameters, as now reasonably well standardized by international usage, are such as to make the noise in a derived channel, at the FM threshold, fall approximately at, or slightly higher than, the level considered to be the maximum tolerable noise for a telephone channel in the public network. By present standards, this maximum is considered to be 55 dBrnc0 (49 dBa0). In industrial systems, a value of 58 dBrnc0 (52 dBa0) is commonly used as the maximum acceptable noise level. per equation (19) dBrnc0 = 88.3 − S/N ∴ S/N = 88.3 − 58 ≈ 30 dB

The noise in a derived voice channel resulting from the receiver equivalent input noise can be calculated, for RF inputs above the FM threshold, as:

$$\text{dBrnc0} = -C - 48.1 + F - 20 \log_{10} \Delta f/f_{ch} \qquad (22A)$$

$$\text{dBa0} = -C - 54.1 + F - 20 \log_{10} \Delta f/f_{ch} \qquad (22B)$$

$$S/N_{dB,\ flat} = C + 136.1 - F + 20 \log_{10} \Delta f/f_{ch} \qquad (22C)$$

$$\text{pWp0} = \log_{10}^{-1} \left[\frac{-C - 48.6 + F - 20 \log \frac{\Delta f}{f_{ch}}}{10} \right] \qquad (22D)$$

where
- C = RF input power in dBm.
- F = receiver noise figure in dB, referred to the point at which input power is established.
- Δf = peak deviation of the channel for a signal of test tone level.
- f_{ch} = center frequency occupied by the channel in the baseband.

∴ f_{ch} ↑, noise ↑

Table E. Noise Unit Comparison Chart.

dBrnc0	dBa0	pWp0	dBm0p	S/N_{dB}	dBrnc0	dBa0	pWp0	dBm0p	S/N_{dB}
0	–6	1.0	–90	88	34	28	2520	–56	54
1	–5	1.3	–89	87	35	29	3162	–55	53
2	–4	1.6	–88	86	36	30	3981	–54	52
3	–3	2.0	–87	85	37	31	5012	–53	51
4	–2	2.5	–86	84	38	32	6310	–52	50
5	–1	3.2	–85	83	39	33	7943	–51	49
6	0	4.0	–84	82	40	34	10,000	–50	48
7	1	5.0	–83	81	41	35	12,500	–49	47
8	2	6.3	–82	80	42	36	15,850	–48	46
9	3	7.9	–81	79	43	37	19,950	–47	45
10	4	10.0	–80	78	44	38	25,200	–46	44
11	5	12.6	–79	77	45	39	31,620	–45	43
12	6	15.8	–78	76	46	40	39,810	–44	42
13	7	20.0	–77	75	47	41	50,120	–43	41
14	8	25.2	–76	74	48	42	63,100	–42	40
15	9	31.6	–75	73	49	43	79,430	–41	39
16	10	39.8	–74	72	50	44	100,000	–40	38
17	11	50.1	–73	71	51	45	125,900	–39	37
18	12	63.1	–72	70	52	46	158,500	–38	36
19	13	79.4	–71	69	53	47	199,500	–37	35
20	14	100	–70	68	54	48	252,000	–36	34
21	15	126	–69	67	55	49	316,200	–35	33
22	16	158	–68	66	56	50	398,100	–34	32
23	17	200	–67	65	57	51	501,200	–33	31
24	18	252	–66	64	58	52	631,000	–32	30
25	19	316	–65	63	59	53	794,300	–31	29
26	20	398	–64	62	60	54	1,000,000	–30	28
27	21	501	–63	61	61	55	1,259,000	–29	27
28	22	631	–62	60	62	56	1,585,000	–28	26
29	23	794	–61	59	63	57	1,995,000	–27	25
30	24	1000	–60	58	64	58	2,520,000	–26	24
31	25	1259	–59	57	65	59	3,162,000	–25	23
32	26	1585	–58	56	66	60	3,981,000	–24	22
33	27	1995	–57	55					

Table E shows the relationship between five commonly used units for expressing noise in a voice band channel. In the first four columns, the units represent weighted noise at a point of zero relative level. In the fifth column the "S" represents a tone at zero relative level, and the "N" represents unweighted noise in a 3 kHz voice channel, therefore, S/N is the dB ratio of test tone to noise.

The table is based on the following commonly used correlation formulas, which include some slight round-offs for convenience. Correlations for Columns 2, 3 and 4 are valid for all types of noise. All other correlations are valid for white noise, but not necessarily for other types.

$$\text{dBrnc0} = 10 \log_{10} \text{pWp0} = \text{dBa0} + 6 = \text{dBm0p} + 90 = 88 - S/N$$

In FM systems without emphasis, Δf has a constant value, regardless of the baseband frequency occupied by the channel. In this case, the equations show that the noise will be higher in the higher frequency channels, the increase being at 6 dB per octave. Thus the thermal noise is worst in the top channel, and that channel is typically used for system noise calculations.

In order to provide a more even distribution of noise across the baseband, emphasis networks are often used to increase the deviation at the higher frequencies and decrease it at the lower frequencies.

Most present day microwave systems are designed around the parameters recommended by CCIR for certain standard configurations. For example, per channel deviations of 200 kHz rms (282.8 kHz peak) are used for systems of 300, 600 or 960 voice channels, and per channel deviations of 140 kHz rms (200 kHz peak) for systems of 1200 and 1800 voice channel capacities. When CCIR emphasis is used, this deviation applies only to the channel at the crossover point of the emphasis curve; at 0.608 f_{max}, where f_{max} is the top baseband frequency. Channels near the bottom of the baseband will have deviations approximately 4 dB lower, and the channel at the very top of the baseband, a deviation 4 dB higher than the reference deviation.

Even with emphasis, the thermal noise is greatest in the top channel, consequently, it is customary to make calculations for that channel only. Table F lists values for the $20 \log_{10} \Delta f/f_{ch}$ factors for the top measuring slot (also standardized by CCIR) for the various channel configurations, both with and without emphasis.

The thermal noise parameter used in video applications is a broadband "signal-to-noise ratio" (S/N), which is defined as the ratio of the peak-to-peak signal to the rms thermal noise in the video baseband. The S/N ratio is dependent on the receiver input level, the noise figure, the video bandwidth, the peak deviation, the de-emphasis characteristics if used, and the weighting function.

The following formulas can be used to calculate the video S/N ratio in dB. They assume a peak deviation of 4 MHz (this is generally standard everywhere) and a video bandwidth of 4.3 MHz. They would not be correct for other bandwidths and deviations.

$$S_{ptp}/N_{rms} = C - F + 118 \quad \text{(22E)}$$
(unweighted, unemphasized)

$$S_{ptp}/N_{rms} = C - F + 126.5 \quad \text{(22F)}$$
(EIA emphasis,
EIA color weighting)

The latter formula is applicable to most television transmission in North America.

CCIR at this time (1970) has not yet standardized either emphasis or weighting for color TV. For monochrome television, CCIR Recommendation No. 421-1 describes several different systems. Monochrome weighting network characteristics are included, and monochrome emphasis network characteristics are given in Recommendation No. 405. The following equation can be used to calculate the video S/N ratio for the various CCIR monochrome systems. The first constant term in each equation represents the unemphasized, unweighted S/N value, and the second constant term is the combined effect of weighting and emphasis. The equations also take account of the fact that CCIR defines the "signal" to exclude the synchronizing pulses, unlike North American practices.

$$S_{p\text{-}p}/N_{rms} = C - F + A \quad \text{(22G)}$$

(CCIR monochrome emphasis and weighting)

$$A = \begin{cases} 119.5 + 13.7 \text{ (405 lines, 3 MHz)} \\ 115.7 + 17.3 \text{ (525 lines, 4 MHz — Japan)} \\ 112.8 + 16.2 \text{ (625 lines, 5 MHz)} \\ 110.5 + 18.1 \text{ (625 lines, 6 MHz)} \\ 112.8 + 13.5 \text{ (819 lines, 5 MHz)} \\ 103.8 + 16.1 \text{ (819 lines, 10 MHz)} \end{cases}$$

Practical Threshold

The "practical threshold", or minimum acceptable RF input level point, cannot be lower than the FM improvement threshold, but may be higher if it is established as an arbitrary value of noise in the top channel. As an example, consider a 960 channel system with CCIR deviation and emphasis, with a receiver noise figure of 10 dB and an IF bandwidth of 32 MHz. We can calculate the FM threshold as $T_{FM} = -104 + 10 \log 32 + 10 = -78.9$ dBm. We can then use this level as the value of C in (22A) to calculate the derived channel noise level at the FM threshold, for example:

Table F. Standard CCIR 20 $\log_{10} \Delta f/f_{ch}$ Factors For Top Slot

SYSTEM CHANNELS	TOP SLOT	WITHOUT EMPHASIS	WITH EMPHASIS
120	534 kHz	–5.52 dB	–1.82 dB (120 channel emphasis)
300	1248 kHz	–12.9 dB	–9.2 dB (300 channel emphasis)
420*	1722 kHz	–15.7 dB	–12.0 dB (420 channel emphasis)
600	2438 kHz	–18.7 dB	–15.0 dB (600 channel emphasis)
960	3886 kHz	–22.8 dB	–19.1 dB (960 channel emphasis)
1200	5340 kHz	–28.5 dB	–24.8 dB (1200 channel emphasis)

NOTE: 200 kHz rms per channel deviation for all except 1200 channel system, which is 140 kHz rms per channel deviation.

*Not a CCIR Standard, but widely used in U.S.A. industrial systems.

$$\text{dBrnc0} = -(-78.9) - 48.1 + 10 - (-19.1)$$
$$= 59.9$$ = the channel noise when C/N = 10 dB

If we are using a value of 55 dBrnc0 as the maximum allowable limit, it is evident that the "practical threshold" will be at an RF input which is 4.9 dB above the FM threshold, or -74 dBm. This illustrates the point that the threshold may be controlled either by the FM threshold or by the channel noise, whichever requires the higher signal.

Fade Margin

Fade margin is the dB difference between the "practical threshold" level and the normal signal level. Most line-of-sight microwave systems are engineered with fade margins in the range of 35 to 40 dB or more. The high normal signal levels are only partly to provide protection against fading. Even if there were no fading, they would still be needed, in most cases, to meet basic system noise objectives.

Figure 23A shows characteristic curves of per channel thermal noise for typical receiver configurations, as a function of RF input level. The curves are calculated for a receiver with a noise figure of 10 dB. They can be used for other values by shifting the input scale 1 dB upward for each dB increase in noise figure above 10 dB, or 1 dB downward for each dB of decrease in noise figure.

Lines corresponding to the limiting values established by the two types of threshold are shown on the figure. The FM threshold lines correspond to four commonly used IF bandwidths. The 12 MHz IF is typical for systems limited to 300 channels, the 22 MHz IF is typical for systems with 300 to 600 channels and a 32 MHz IF is typical for systems with 600 channels or color TV. For systems of 960 channels through 1800 channels, 40 MHz IF bandwidths are commonly used.

In order to meet overall noise objectives, most microwave paths are engineered with normal signal levels chosen to limit receiver thermal noise to about 14 to 16 dBrnc0. Figure 23A shows that, with a 10 dB noise figure, this will dictate receiver input levels of about —40 dBm for the 300 channel system, about —37 dBm for the 600 channel system, about —33 dBm for the 960 channel system, and about —28 dBm for the 1200 channel system. When very stringent noise objectives are required, receiver inputs up to 5 dB higher than the above may be essential.

System Loading and its Effect on Noise

At the low end of the RF input range, the thermal noise is the only significant noise source, but at the high end of the range, intermodulation noise and intrinsic noise are significant contributors, and provide a lower limit to the total noise curve.

For 600 ...

dBrnc0 = $-C - 48.1 + F - 20 \log_{10} \Delta f/f_{ch} = -C - 48.1 + 10 - 20 \log$ 200 kHz / 2438 kHz
$= -C - 48.1 + 10 - (-15.0)$
$= -C - 23.1$
$= -C - 23.1 \Rightarrow C = -38.1$ dBm

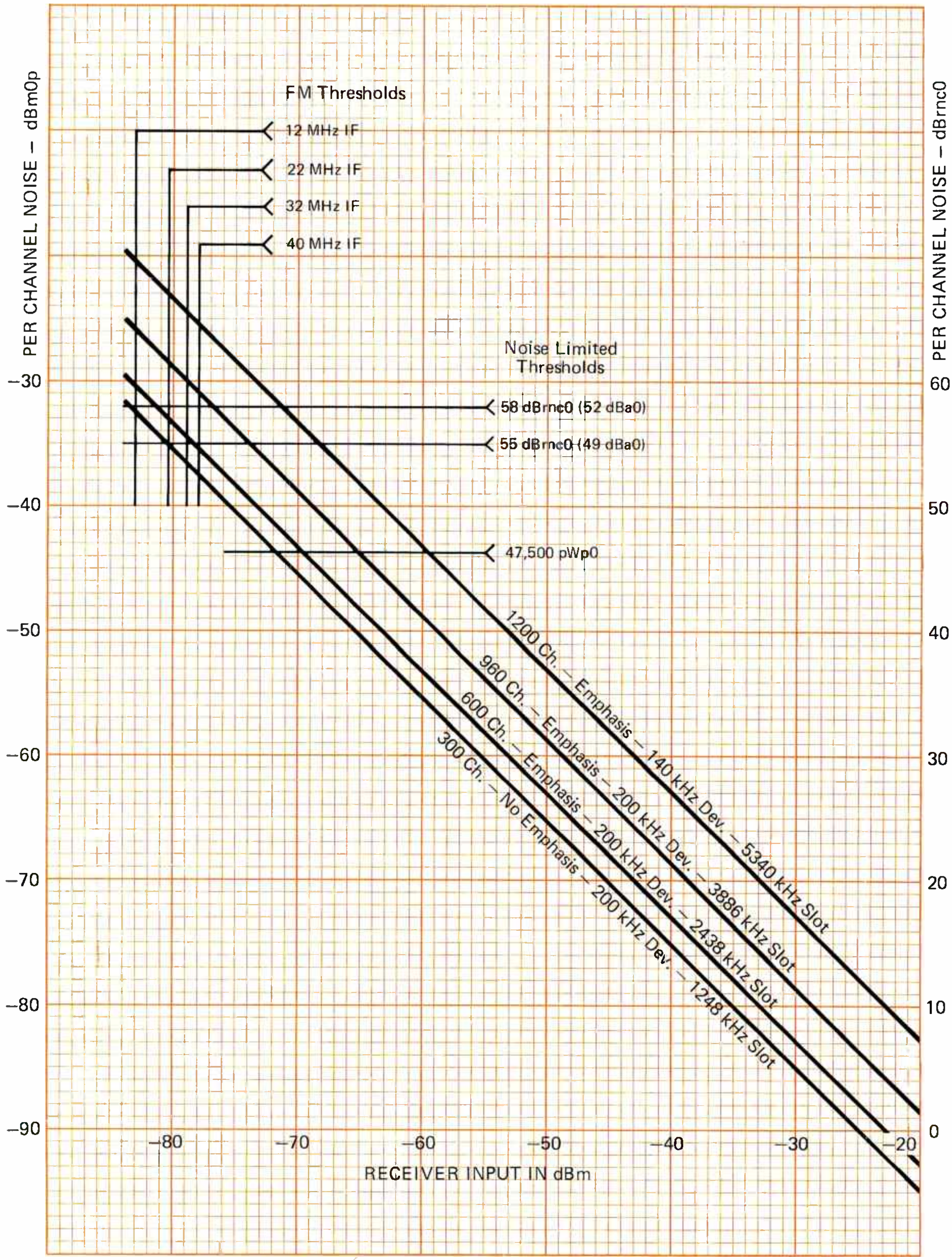


Figure 23A. Receiver Thermal Noise. (10 dB Noise Figure Assumed)

The amount of intermodulation and intrinsic noise is a characteristic of the equipment and can be determined either from the manufacturer specifications, or from actual measurements on systems; unlike receiver thermal noise, it is not readily calculable from known system parameters. The intermodulation noise is, however, a function of the system loading as well as equipment parameters.

It is now standard practice to specify the loading capacity of a microwave system, and its performance, in terms of "white noise" loading applied to the used portion of the baseband at a level chosen to make its peak values equivalent to those of a multichannel telephone load.

Voice Loading

Microwave systems are commonly designed and rated for the equivalent loads as established by CCIR and given by the following equations:

$$P = (-15 + 10 \log_{10} N)\ \text{dBm0} \quad \text{(when } N = 240 \text{ or more)} \tag{23A}$$

$$P = (-1 + 4 \log_{10} N)\ \text{dBm0} \quad \text{(when } N = 12 \text{ to } 240\text{)} \tag{23B}$$

where P = equivalent noise loading power
N = number of SSBSC channels

The expressions within the parentheses give the ratio, in dB, of the equivalent noise load power to test tone power, and are often called "Noise Loading Ratio" or NLR.

The second equation reflects the fact that the peak-to-rms factor for the smaller number of channels is higher than that of white noise, so that a higher value of rms white noise power must be provided, in this case, to obtain the same peak value as that of the voice channels. Beyond the 240 channel point, the white noise and voice are considered to have the same peak factor, so that the rms values of the voice load and the equivalent white noise load are equal.

The choices of per channel deviation, IF bandwidth, etc., are made to provide reasonably good balance between thermal and intermodulation noise at normal signal levels. Thermal noise performance can be improved by increasing the per channel deviation, but at the expense of worsened intermodulation noise because increasing the deviation increases the loading. The CCIR recommended parameters are well suited to systems carrying principally public telephone traffic, but may not be optimum for systems with high percentages of the channels devoted to data.

A slightly different loading formula is in widespread use in the U.S.A. telephone industry. This loading, which was established by the Bell System as a result of the latest available measurement data, is given — for multi-channel systems of relatively high capacity — by:

$$P = (-16 + 10 \log_{10} N)\ \text{dBm0} \tag{23C}$$

It represents 1 dB lighter loading than the corresponding CCIR formula, and reflects a reduction in talker volumes from previous values.

"Military" Loading

Standards currently in use or proposed by the Defense Communications Agency (DCA) for U.S. military systems specify an equivalent loading 5 dB higher than CCIR, in order to allow essentially unrestricted use of data at relatively high levels. The corresponding formula is given by:

$$P = (-10 + 10 \log_{10} N)\ \text{dBm0} \tag{23D}$$

This means an equivalent loading power approximately 3.2 times as great as that for CCIR loading. Thus, a 300 channel system with this type of loading must be engineered essentially as if it were a 960 channel system with CCIR loading.

Composite Loading

Equations (23A) and (23B) given above, are based on systems which are primarily used for voice transmission, though they include an allowance for signaling tones and for a small percentage of the SSBSC channels to be used for telegraph or data multiplex.

When relatively large numbers of the channels are used for such data services, the calculation of equivalent loading becomes more complex. Usual practice is to calculate separately the equivalent white noise loading for the number of channels used for voice, and the equivalent rms power in dBm0 of all the tones used for transmitting data over the system, then to sum the two rms powers on a power additive basis to obtain an equivalent total white noise loading power and noise loading ratio. This is justifiable because the peak to rms factor for a composite signal consisting of a number of tones, is closely equivalent to that of white noise when the number exceeds about 16 tones.

As an example, consider a 600 channel system with 500 channels devoted to voice, 40 channels carrying data at a level of —12 dBm0 per channel, and 60 channels, each carrying 20 submultiplexed telegraph carriers at —21 dBm0 per carrier. The equivalent load of the voice is, from equation (23A), —15 + 10 log 500 = +12.0 dBm0; the equivalent load of the data is —12 + 10 log 40 = + 4.0 dBm0, and the equivalent load of the telegraph carriers is —21 + 10 log 20 + 10 log 60 = +9.8 dBm0.

Summing dB Powers

Table G provides a simple means of summing (or subtracting) non-coherent powers expressed in dB form. Using this table, we can calculate that the summation of the three loads of +12.0, +4.0 and +9.8 dBm0, determined in the preceding paragraph, is approximately +14.5 dBm0.

Since the equivalent noise loading for a 600 channel system is only +12.8 dBm0, it is evident that the +14.5 dBm0 load would be some 1.7 dB higher than that of an equivalent 600 channel voice system. It would then be necessary to determine whether the system could actually carry the extra load with an acceptable level of performance, and without exceeding bandwidth restrictions. If not, the load could be reduced by any of several expedients. One would be to lower the levels of the data and the telegraph channels until the total rms power of data tones occupying a SSBSC channel does not exceed -15 dBm0. Another would be to eliminate some of the voice channels. A third would be to reduce the per-channel deviation, seeking a more optimum balance between thermal and intermodulation noise for the new conditions. It is beyond the scope of this work to discuss all the ramifications associated with this problem.

Intermodulation Noise

Equipment intermodulation noise is not a directly calculable quantity. Its value is usually specified by the manufacturer, at the normal design loading of the system and for specified conditions. On a measurement basis, it can be determined by noise loading measurements on terminals connected back-to-back, with the RF received signal set sufficiently high so as to reduce front-end thermal noise to insignificance. The resulting reading will then give the equipment intermodulation plus intrinsic noise. If desired, the latter can be determined by repeating the measurement with loading removed.

ECHO NOISE

Echo distortion noise, as described earlier, is a form of intermodulation noise which is produced by delayed echo signals, created usually by a combination or combinations of impedance mismatches in waveguides, antennas, and the equipment itself. It can also occur at IF, from cable and equipment mismatches and, more rarely, as a result of long-delayed echoes in the path itself.

The effects of echo distortion are a very complicated function of a number of things such as the number of mismatches, the magnitudes of the mismatches, the distance between them, the velocity of propagation in the guide, the number of channels, the per channel deviation, the total loading, the presence or absence of emphasis, etc. Because of the complexity, exact calculations are very difficult.

The curves of Figure 24 can be used to calculate an approximate value of echo distortion noise, for certain standard channel arrangements. To use the curves, it is necessary to know the equipment return loss, and to know (or to assume) a composite value for the combined return losses of antenna-plus-waveguide at the frequencies of interest. The latter are assumed to be lumped at the antenna location. As an example of the use of the chart in Figure 24, make the following assumptions:

A 960 channel system, in the 6 GHz band, with 100' of waveguide. Equipment return loss of 28 dB. Antenna-plus-waveguide return loss of 26 dB (equivalent to a VSWR of 1.1).

From the chart, the dBrnc0 constant for 960 channels at 100' is 70.5, and the loss of 100' of waveguide at 6 GHz is approximately 2 dB. Hence, the noise in the top channel in dBrnc0 is equal to;

$$70.5 - 28 - 26 - 4 = 12.5 \text{ dBrnc0}$$

Table G. Summation Or Subtraction Of Non-Coherent Powers.

This table can be used for summing the powers of two non-coherent signals expressed in dB form. It can also be used for power subtraction.

P_a and P_b represent two powers whose summation is P_s: in all cases P_a is taken as the larger of the two powers.

To sum two powers, calculate $P_a - P_b$, locate the resulting value in Column 1, then add the corresponding value in Column 2 to P_a to obtain P_s, the desired sum.

To subtract one power from another, treat the larger one as P_s and the smaller one as P_a (if it is within 3 dB of P_s) or P_b (if it is more than 3 dB below P_s).

In the first case, calculate $P_s - P_a$, locate the resulting value in Column 2, then subtract the corresponding value in Column 3 from P_s to obtain P_b, the desired remainder.

In the second case, calculate $P_s - P_b$, locate the resulting value in Column 3, then subtract the corresponding value in Column 2 from P_s to obtain P_a, the desired remainder.

When more than two powers are to be summed or subtracted, iteration can be used.

Example: Summation

To add + 10.0 dBm0 to +8.7 dBm0
$10.0 - 8.7 = 1.3$
1.3 in Column 1 falls between 1.2 and 1.4 so the value from Column 2 is
$(2.45 + 2.37)/2 = 2.41$.
So $P_s = + 10.0 + 2.41 = + 12.41$ dBm0

Example: Subtraction

To subtract −15.0 dBm from −10.0 dBm0
$-10.0 - (-15.0) = 5$, treating −10.0 as P_s and −15.0 as P_b we locate 5 in Column 3 as very near to 5.035, so we subtract the corresponding value in Column 2, 1.635, from −10.0 to obtain P_a.
So $P_a = -10.0 - 1.635 = -11.6$ (rounded).

Col. 1	Col. 2	Col. 3
$P_a - P_b$	$P_s - P_a$	$P_s - P_b$
0.0	3.010	3.010
0.2	2.911	3.111
0.4	2.815	3.215
0.6	2.721	3.321
0.8	2.629	3.429
1.0	2.539	3.539
1.2	2.451	3.651
1.4	2.366	3.766
1.6	2.284	3.884
1.8	2.203	4.003
2.0	2.124	4.124
2.2	2.048	4.248
2.4	1.974	4.374
2.6	1.902	4.502
2.8	1.832	4.632
3.0	1.764	4.764
3.2	1.698	4.898
3.4	1.635	5.035
3.6	1.573	5.173
3.8	1.513	5.313
4.0	1.455	5.455
4.2	1.399	5.599
4.4	1.345	5.745
4.6	1.293	5.893
4.8	1.242	6.042
5.0	1.193	6.193
5.2	1.146	6.346
5.4	1.100	6.500
5.6	1.056	6.656
5.8	1.014	6.814
6.0	0.973	6.973
6.5	0.877	7.377
7.0	0.790	7.790
7.5	0.710	8.210
8.0	0.639	8.639
8.5	0.574	9.074
9.0	0.515	9.515
9.5	0.461	9.961
10.0	0.414	10.414
11.0	0.331	11.331
12.0	0.266	12.266
13.0	0.216	13.216
14.0	0.170	14.170
15.0	0.135	15.135
16.0	0.108	16.108
17.0	0.086	17.086
18.0	0.068	18.068
19.0	0.054	19.054
20.0	0.043	20.043
25.0	0.016	25.016
30.0	0.004	30.004
∞	0.000	∞

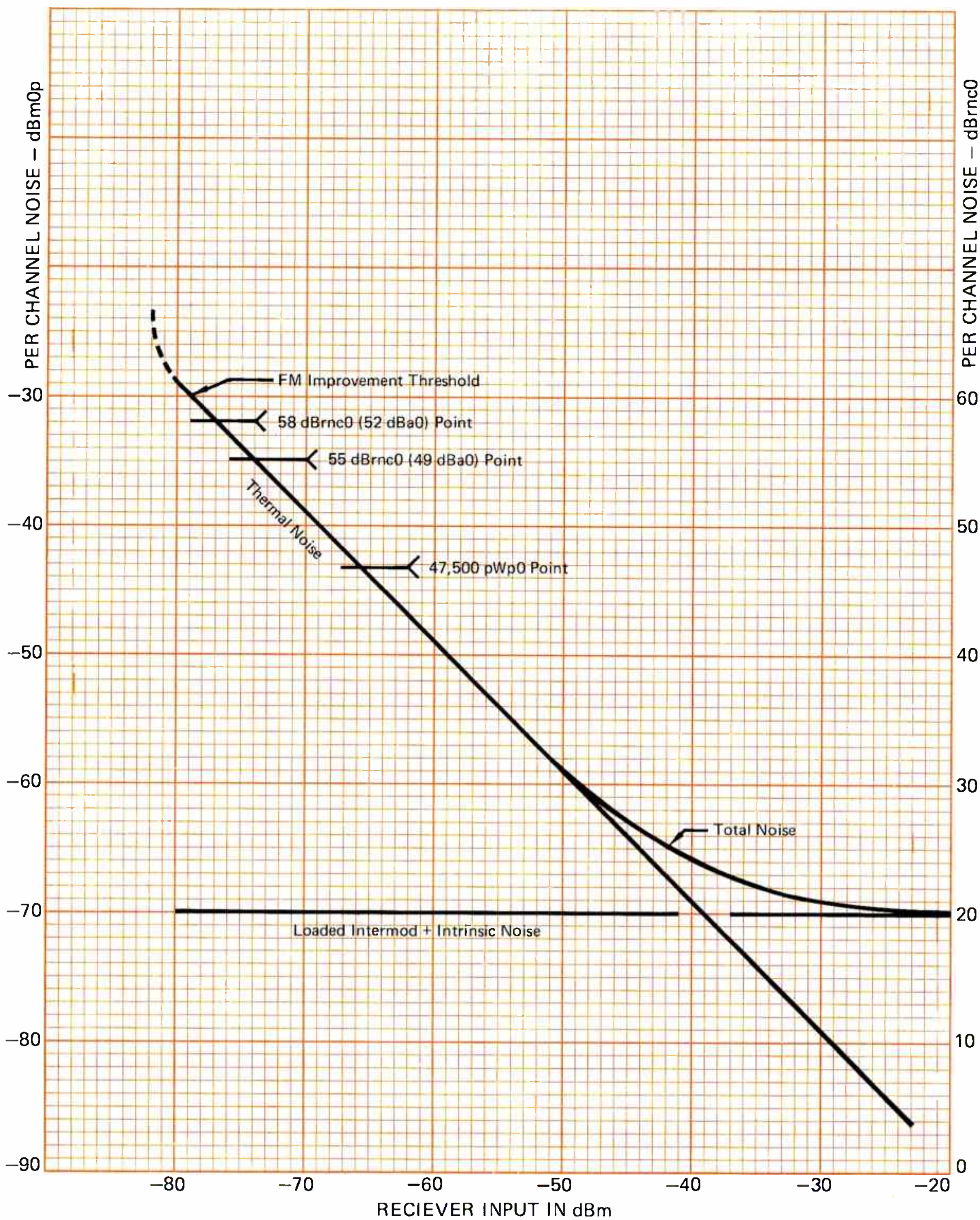


Figure 23B. Typical Receiver Noise Curve. Top Slot Of 960 Channel System; 10 dB Noise Figure; 200 kHz Deviation; CCIR Emphasis, +14.8 dBm0 Load.

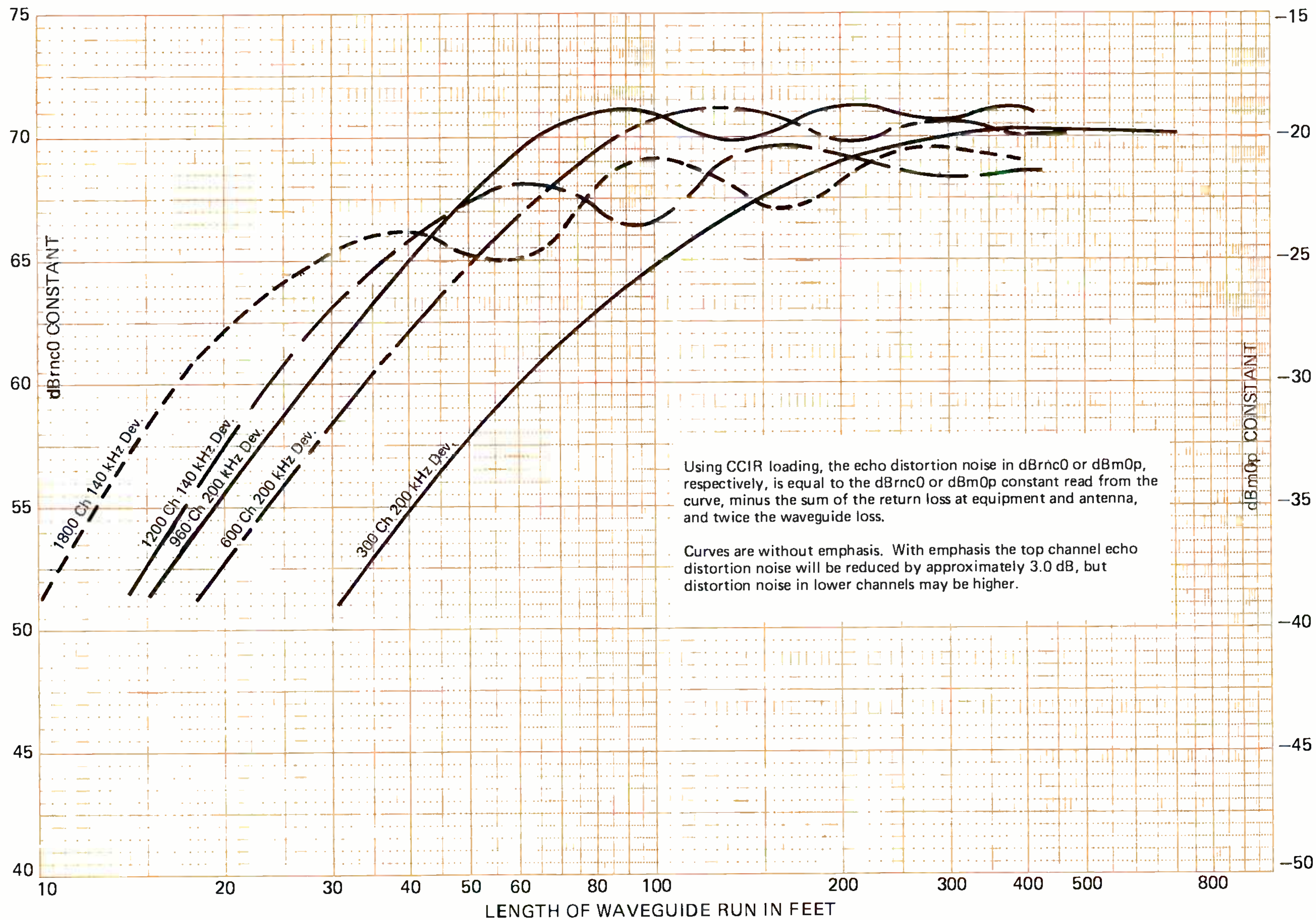


Figure 24. Echo Distortion Noise.

From Table E, in a previous section, this can be found to be equivalent to 18 picowatts, psophometrically weighted, if that noise unit is to be used. If emphasis is used, the noise will be 3 dB lower, or 9.0 pWp0.

It will be seen from the chart, that the echo distortion noise constants increase rapidly with guide length until a plateau is reached, beyond which they change very little as the guide gets longer. Because of the effect of waveguide attenuation in reducing the noise, for any given configuration there will be a peak point in the noise curve, at about the knee of the curve, for which the noise will be a maximum. For longer or shorter lengths it will be less.

Return loss in dB can be calculated as

$$20 \log_{10} \frac{\text{VSWR} + 1}{\text{VSWR} - 1}$$

or taken from the following:

VSWR	R.L.	VSWR	R.L.	VSWR	R.L.
1.02	40.1	1.07	29.4	1.15	23.0
1.03	36.6	1.08	28.3	1.20	20.8
1.04	34.1	1.09	27.3	1.25	19.0
1.05	32.2	1.10	26.4	1.30	17.8
1.06	30.7	1.12	24.9	1.40	15.4

Note: The significant parameter in microwave systems work is return loss rather than VSWR. The continued use of VSWR in describing the impedance characteristics of microwave antennas, waveguides and components is a holdover from the days when the only available measurement technique, the slotted line, gave results directly in VSWR. With modern sweep generators and reflectometer techniques the measurement gives return loss directly, and it is somewhat pointless to first convert it to VSWR and then reconvert it to return loss. It would be simpler if ratings were simply stated in terms of return loss rather than VSWR, by the manufacturers of microwave equipment and components. Some companies have already taken steps in this direction.

Per Hop Total Noise

If the level of intermodulation noise plus intrinsic noise for the top channel of a given system at the desired loading is known (it usually must be obtained from the manufacturer), this noise and the receiver thermal noise can be plotted on a chart and added on a power basis, to construct an overall curve showing loaded noise as a function of RF signal level for the particular equipment and conditions. Figure 23B is an example of such a curve. It should be borne in mind that such a curve will be valid only for the particular parameters and conditions on which it is based. Note also that Figure 23B does *not* include echo distortion noise, which must be treated separately for each case. See later examples.

The curve of Figure 23B is for a "baseband" or remodulating type of equipment, and includes the noise contributions of the FM modulator, demodulator and associated baseband equipment. In heterodyne systems these elements are not included with the radio equipment, so their loaded noise contributions must be determined separately, and added in as another contributor to the system noise. The radio portion of a heterodyne equipment will have the same thermal noise as an equivalent remodulating equipment, but its intermodulation and intrinsic noise will be lower because of the absence of remodulating and baseband equipment.

4. System Noise Objectives

Because of the wide variety of communications systems, and the many ways in which a channel can be used, there is no way to establish a simple and universal noise objective which would be optimum for all systems. A very strong trade-off relationship exists among system capacity, system costs, and system noise performance. Each communications user must, in some manner, evaluate and establish his own requirements in the light of these relationships. Choosing noise objectives higher than really needed can often cause an inordinately large increase in system cost, while setting them too low can seriously limit the usefulness and expandability of the system.

Since the noise power in a multi-hop system is approximately equal to the power summation of the noise powers of the individual hops, the noise objectives must, of course, include a distance factor. Consequently, most noise objectives are established as proportional to the length of the system. This is a reasonable assumption for long systems, since hop lengths tend to average out, but is difficult to apply to short systems, and particularly to those with short average hop lengths.

The following paragraphs list some noise objectives in current use at the time of preparation of this manual. It should be borne in mind that these values are design objectives and not "standards" or specification requirements, though they are often interpreted in that way.

CCIR/CCITT International Circuits

CCIR and CCITT establish hypothetical reference circuits 2500 kilometers long, and relate their noise objectives to these reference circuits, or to systems closely similar to them.

The basic design objective for a voice channel of the hypothetical reference circuit is 10,000 pWp0 mean noise power in any hour, of which 2,500 pWp0 is allocated to the multiplex equipment and 7,500 pWp0 to the transmission line (microwave system in this case). For systems between 280 and 2500 kilometers in length, the noise contributed by the line is considered proportional to length, i.e. 3 pWp0 per kilometer.

For real systems, the following somewhat more complicated formulas are established:

Systems of 50 to 840 kilometers; 3 pWp0 per kilometer plus 200 pWp0.

Systems of 840 to 1670 kilometers; 3 pWp0 per kilometer plus 400 pWp0.

Systems of 1670 to 2500 kilometers; 3 pWp0 per kilometer plus 600 pWp0.

CCIR does not attempt to define noise performance for systems shorter than 50 kilometers, since such systems are not amenable to any pWp0 per kilometer formula and must be treated as special cases. The above formulas are for the noise contributed by the microwave or transmission system only, and do not include the multiplex noise. The latter must be added in to obtain total noise on the system.

The CCIR/CCITT objectives were originally developed for circuits over coaxial cables. It is now recognized that they are not well suited to microwave systems because of the "in any hour" requirement which, if strictly interpreted, could mean the "worst hour" of the "worst month" of the "worst year". This has two significant disadvantages. One is that there is no practical way to determine in advance (and hardly any practical way after the fact) just what the relationship is between mean noise in the worst hour and mean noise under normal conditions for any microwave system. The other is that, even if there were such a way, it would in many cases be economically grotesque to perhaps double the cost of the system to avoid allowing a relatively modest increase in noise during a very few hours of the year.

The subject has been under discussion for some time, but as yet no changes have been agreed on. Most CCIR-conforming countries have adopted an informal and pragmatic approach, which is to assume that the mean noise in the worst hour will be that corresponding to the noise with all hops in the system faded by approximately 5 dB below their normal free space level. This, of course, requires that in some cases, the hops be engineered for roughly 5 dB stronger signals than would otherwise be needed, but at least it establishes a definable and measurable condition.

The CCIR noise objectives assume that the system is loaded, either with traffic or with white noise, at the levels indicated by the appropriate CCIR loading formulas. In addition to the "mean noise in any hour" criterion, CCIR requires that the same numerical value of noise, measured on a "one-minute-mean" basis, not be exceeded for more than 20% of any month. This, too, is conventionally handled by a simulated fade of 5 dB below the median signal. This value is justified by the fact that there is a difference of 5 dB between the median (50%) level of a Rayleigh-distributed signal and the 80% level, as well as by experimental data.

U.S.A. Public Telephone Networks

Noise objectives in the U.S.A. are not as rigidly or officially formulated as in the CCIR, and are subject to fairly frequent changes. Table H gives what we believe to be objectives commonly applied

Table H. Typical U.S.A. Noise Objectives.

TRUNK LENGTH IN MILES	TYPE OF TRUNK INTERTOLL* dBrnc0	TOLL CONNECTING* OR TANDEM dBrnc0	DIRECT* dBrnc0
0–50	30	32	35
51–100	31	33	36
101–200	33	35	38
201–400	35	37	40
401–1000	37	39	42
1001–1500	38	40	43
1501–2500	41	43	46
2501–4000	43	45	47

*SEE NOTE ON PAGE 80.

at the present time. They are for complete trunks, from the outgoing switch in one office to the incoming switch in the next office, and therefore, include multiplex noise and any other noise sources, as well as that of the microwave. They are given in terms of dBrnc0, measured during the busy hour (but with the microwave system in a normal or unfaded condition).

Two points are worth mentioning. One is that, unlike CCIR, these objectives assume that any microwave systems are in the unfaded condition, which prevails for the overwhelming percentage of the time. Another is that the most stringent objectives are applied to intertoll circuits, and other types of circuits have considerably relaxed objectives. Intertoll circuits in the public networks may be tandemed with a considerable number of similar ones in making up very long circuits, and it is for this reason that their performance must be so good. Private microwave systems do not generally have this problem to contend with (at least not to that degree), consequently a private user looking to telephone company objectives as a guide to his desired microwave performance could, in many cases, use the more relaxed objectives associated with toll connecting trunks or even in some cases, those of direct trunks.

Note: The objectives as tabulated in Table H represented the practices at the time this book was first prepared. It is our understanding that present practices (1970) remove the distinction between intertoll, toll connecting, tandem, and direct trunks, and apply the objectives of the "intertoll" column to all trunks. The reason is that with modern telephone industry switching and automatic routing practices the clearcut distinctions between the various types of trunks tends to disappear. We have left the table intact, however, because of its usefulness for other applications.

Another slight difference between the CCIR and U.S.A. approach, is that the latter generally assumes an equivalent white noise busy hour loading of −16 dBm0 per channel (Bell System Standard) instead of the −15 dBm0 per channel CCIR value.

Industrial Systems

A commonly used objective for industrial systems, is 32 dBa0 (equivalent to 38 dBrnc0) busy hour loaded noise, for a system 1,000 miles in length (34 hops). This objective includes the multiplex noise contribution.

"Military" Systems

Current or proposed DCA standards for the long term median noise on real systems are as follows (where L is the length of the hop or system in *nautical* miles)

Section Length (L)	Allowable Noise
$L > 151$ nmi	3.33 L pWp0
$27 < L < 151$ nmi	(2.76 L + 85.5) pWp0
$L < 27$ nmi	160 pWp0

As presently stated (1970) these are specifically identified as "worst hour" objectives. An EIA ad hoc committee has recommended to DCA that the "worst hour" statement be deleted and the objectives specified and measured with all hops faded 3 dB below their normal calculated signal level. The reasons are the same ones discussed in the section on CCIR objectives.

Note that the military objectives apply with the system loaded per Equation (23D), that is, −10 dBm0 equivalent noise loading per channel.

Television Transmission Systems

The random noise parameter used to specify noise for television transmission, is the ratio of the peak-to-peak video signal to the weighted rms noise in the band occupied by the television signal. Various objectives are in existence.

EIA Standard RS-250A recommends a weighted S/N ratio minimum of 59 dB for a single hop system, and a minimum of 56 dB for a multihop system. This is applicable for either color or monochrome transmission.

CCIR Recommendation 421 lists, for the 2500 kilometer hypothetical reference circuit, various values ranging from 50 dB to 57 dB, depending upon the system. This recommendation is for monochrome only. In general, systems engineered to meet voice channel noise performance requirements for 600 channels or more, will provide adequate noise performance for television transmission.

Note: The EIA and the CCIR definitions of S/N differ by a factor of 3 dB, because EIA defines the "signal" as the total video signal, whereas CCIR defines it as the picture signal, exclusive of synchronizing pulses. Other differences arise as a result of widely differing emphasis and weighting networks, so the EIA and CCIR S/N ratios cannot be directly compared.

5. Noise Allowable for Small Percentages of Time

A second type of noise objective establishes a maximum allowable value of noise, beyond which the circuit is considered unusable. In a microwave system, this value of noise defines the "practical threshold" of the receivers, and those periods of time during which the receiver is below the practical threshold (and usually muted) are by definition "propagation outages".

This parameter thus establishes the bottom limit of the fade margin range, and is closely tied in with system reliability.

Line-of-sight microwave systems are invariably engineered with very large fade margins, and often with diversity protection, so that the percentages of time during which any one hop will be at, or below, threshold are very small. Consequently, it is unlikely that two hops of even a long system will be simultaneously below threshold. This means that the expected propagation outage time of a multihop system will be equal to the sum of the expected propagation outage times of all the individual hops. Consequently, only a single value of maximum noise is used, regardless of the number of hops, but the allowable time percentage during which it can be experienced, *does* increase with the number of hops, and is approximately proportional to system length.

CCIR/CCITT

The CCIR objective for the 2500 kilometer hypothetical circuit, is that the one-minute mean noise power should not exceed 47,500 pWp0 for more than 0.1% of any month, with proportionately smaller allowable percentages down to 280 kilometers.

The same objective is applied to real circuits from 280 to 2500 kilometers in length, but for real circuits of any length up to 280 kilometers, a single percentage value of (280/2500) x 0.1%, or 0.0112%, is specified.

Again, the CCIR objective is adapted from older objectives developed for cable systems, and is not well suited for microwave systems. One reason is that it is in terms of "one-minute mean noise", that is, noise integrated over a period of one minute. This is reasonable for systems in which noise changes are slow with respect to a minute, such as cable systems, but it is inappropriate for the rapidly changing noise associated with deep fading in a microwave system. Consequently it is a difficult parameter to predict and measure. A second objection is that this value of noise is not really high enough to justify considering it a "mute" or "propagation outage" point, since the circuits would be quite usable with 8 or 9 dB greater noise.

No solution satisfactory to CCIR has yet been found to this problem, which is still under study by CCIR/CCITT. For further details of CCIR noise objectives, reference should be made to CCIR Recommendation No. 395-1.

> *Note: CCIR Recommendation No. 393-1, which applies only to the hypothetical reference circuit, has an additional short term noise objective which is omitted completely from the objectives in 395-1 for real systems.*
>
> *The additional objective is that the noise power on the 2500 kilometer reference circuit should not exceed 1,000,000 pW0* unweighted *(with an integrating time of 5 ms) for more than 0.01% of any month.*
>
> *This value would correspond to 562,000 pWp0 of weighted noise power, or about 57.5 dBrnc0.*
>
> *It is a relatively easy parameter to predict and measure, and is a satisfactory value to be used as a muting or "propagation outage" point.*
>
> *It thus appears that it would be a much more suitable CCIR noise parameter to use in real systems than the controversial 47,500 pWp0 parameter discussed above.*
>
> *It is very similar in all respects, to the short term allowable noise objectives as used in the USA and described in the following two paragraphs.*

U.S.A. Public Telephone Networks

The maximum allowable noise level is presently established at 55 dBrnc0. It is to be measured with a short time constant meter (on the order of milliseconds), and does represent the muting or outage level, consequently it is a much more realistic parameter than the comparable CCIR objective.

The allowable time percentage objectives depend on the type of system. For systems of the long-haul type (specifically intertoll) an objective of 0.02% for a 4,000 miles system is applied, with proportionally shorter times for shorter systems. For systems of the short-haul type, an objective of 0.02% for systems of any length up to about 200 miles is used. *Both types of objective are such as to require, in general, some form of diversity to protect against multipath fading.*

Industrial Systems

The most commonly used value of maximum allowable noise in industrial systems is 52 dBa0. This corresponds to 58 dBrnc0, and thus allows 3 dB higher noise before muting, than the current telephone objective described above. Until fairly recently, 52 dBa0 was the objective of the telephone industry as well.

Allowable time percentage objectives vary rather widely, depending on the type of system and the usage. Many industrial users have reliability objectives no less stringent than those applied to long haul telephone systems. Meeting such objectives will, in general, also require diversity to protect against multipath fading.

"Military" Systems

Current or proposed DCA standards state that the short term mean noise power, with an integrating time of 5 ms, on any 4 kHz (nominal) channel shall not exceed 316,000 pWp0 for more than an accumulated 2 minutes in any month or more than 1 minute in any hour over any hop.

316,000 pWp0 is equivalent to 55 dBrnc0 or 49 dBa0.

Television Transmission

EIA RS-250A suggests that a weighted S/N ratio of 33 dB be considered as the "outage threshold" beyond which the noise will be unacceptable. No time percentage is suggested. A color-weighted S/N ratio of 37 dB is also widely used as a practical threshold for video.

V. EQUIPMENT

It is obvious from the preceding discussions that there is a very close inter-relationship between the characteristics of the various items of the equipment to be used, and the engineering choices and performance parameters of the paths themselves.

Thus it is desirable, in fact almost essential, that the path survey engineers have enough advance knowledge of the frequency bands to be considered (often only one, but in some cases more than one), the kind of service, the number of channels (both present and future) to be accommodated by the system, the kind of performance and reliability criteria desired, and the pertinent parameters of the microwave equipment to be used (for example, transmitter output power, receiver noise figure and bandwidth, per channel deviation etc.), to allow an intelligent approach to the problem of path engineering. Many choices are involved in path selection, and choices made without a thorough knowledge of all the pertinent circumstances may not be the best ones.

A. Radio Equipment

Microwave systems can range from as little as 5 or 10 miles to distances as long as 4,000 miles. Facility requirements can be relatively small, requiring structures and equipment for only a light route, or they may be very heavy, requiring multi-channel, heavy route layout with sophisticated switching. They can be constructed for nominally good service during certain limited hours of the day with considerable economy, or they can be built for a very high quality of service on a 24 hour a day, year-in and year-out basis.

Some systems are of a "through" type, with all or almost all of the channels going end-to-end, while others require multiple access, with dropping and inserting of channels at most, if not all, repeater points. The latter is very typical in industrial systems, in which long haul and short haul are almost always combined in a single radio channel.

The two types of FM microwave equipment in common use are the IF heterodyne type and the baseband, or remodulating, type. The IF heterodyne type, by eliminating demodulation and remodulation steps at repeaters, contributes the least amount of distortion, and is the preferred choice for systems handling exclusively, or almost exclusively, long-haul traffic, with little or no requirement for drop and insert along the route. The heterodyne type is also preferable for systems carrying color TV, if more than a few hops are involved.

Equipment of the baseband or remodulating type is widely used for short haul or for distributive systems in the telephone industry, and for either short or long haul industrial systems. The great flexibility for drop and insert, plus maintenance advantages, are the determining factors. Heterodyne systems are inherently at a considerable disadvantage in such applications.

Apart from the choice between heterodyne or remodulating equipments, some other primary considerations in the selection of the best radio equipment for a particular system include: (a) characteristic of the end-to-end baseband facility, including bandwidth, frequency response, loading capability, noise figure and noise performance; (b) the amount of radio gain available, as determined by transmitter power output and receiver noise characteristic; (c) operating frequency band, and required frequency spacing between radio channels, as determined by transmitter deviation, receiver selectivity and frequency stability; (d) primary power requirements and options available; (h) supervisory functions available, including order wire, alarms and controls; (f) equipment reliability, including availability of redundant versions such as frequency diversity, 1-for-N or 2-for-N multiline switching, hot standby, or hot standby at transmitters and space diversity at receivers; and, (g) provisions for testing and maintenance.

With the rapidly changing nature of the state of the art, and the continuing development of new equipments and upgrading of old ones, specific data on microwave equipment characteristics can become outdated in very short order. Consequently we have not included such data in this manual. Rather, the user or engineer should rely on up-to-date data obtained from the manufacturer.

B. RF Combiners

A variety of methods are used to combine a transmitter and receiver, or several transmitters and receivers, for operation over a single antenna system. These methods make use of waveguide switches, hybrids, filters, phasors, isolators, circulators, and other devices that have a certain amount of inherent attenuation. This attenuation must be included when measuring transmitter power and receiver sensitivity, or it must be taken into account when calculating net path loss. Because of the wide variety of devices and possible combinations, it is necessary to rely on the manufacturer specifications for attenuation data. These data should clearly state the point at which transmitter power and receiver sensitivity are specified and measured, and the loss of any device that is to be located between this point and the antenna.

C. Towers

Towers and tower problems have a significant effect on many microwave path engineering choices. The microwave path engineer needs a considerable prior knowledge of the limitations which the characteristics of towers (and of antennas and waveguides as well) impose, and the resulting restrictions in his freedom of choice, in order to be equipped to do his job properly.

For example, the engineer must have a fairly clear idea about how high he can economically go with towers, before he can tell how long the paths can be, or conversely, whether a given path can be achieved in one hop. From the material in previous sections, it is easy to calculate that a 30 mile path on relatively flat terrain could call for towers on the order of 250 feet at each end; if there were hills or trees in the middle, this could easily go up to 300', 350' or even more, in order to achieve the desired clearances. These are not unreasonable heights for stations located in the country, or in areas where plots sufficiently large to allow guyed towers exist. But, if a station is in a built-up downtown area, as end terminal stations often are, guyed towers are usually out of the question. Self-supported towers of such heights are extremely expensive, also require quite a bit of real estate for the foundation, and often may be prohibited by local codes or other legal reasons. Another important consideration is the proximity to airports or air lanes, which brings in the possibility of government restrictions on permissible tower height. Some foreknowledge of the potential tower limitations is mandatory before the problem can be attacked knowledgeably.

The two generic types of tower are guyed, and self-supporting. For very short towers there is not much cost difference, but as heights go up, the cost of the self-supported types increase more or less exponentially, while that of the guyed towers, which have a constant cross-section, increases more or less linearly. So where high towers are required, there are very strong incentives toward the use of guyed towers, provided there is sufficient space to allow them.

Figures 25A, 25B and 25C show the required areas for various types of tower. Though these drawings are representative, other types of tower, particularly the large, heavy structures used with high-density systems, will have different base arrangements and area requirements. An examination of the figures will show the large difference in ground requirements. Where self-supported towers cannot be avoided, considerable study is warranted in layout of a microwave system, to try to keep the paths such that excessive heights are not required.

There are several things which are frequently queried in connection with towers, and with the quoting and furnishing of towers by a manufacturer. Some of these are:

Soil conditions.
Wind loading.
Local building codes and restrictions.

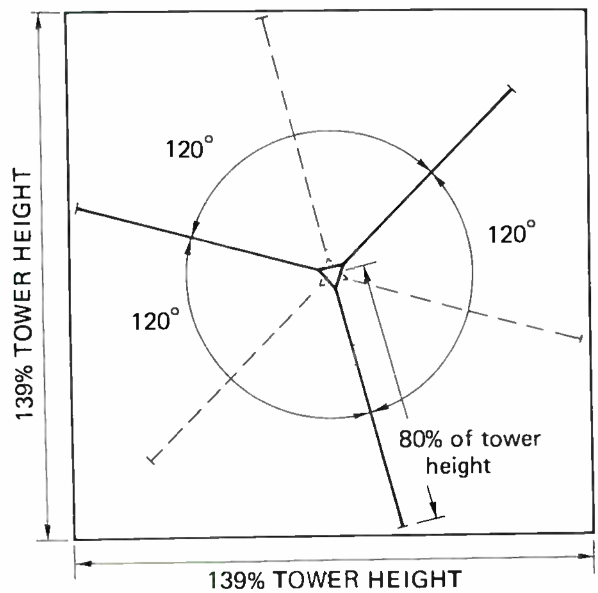
Unless accurate and specific information about soil characteristics (on the exact site of a proposed tower) are available, the quoted costs are almost invariably based on "standard soil" as defined in EIA Standard RS-222A.

If it turns out that soil conditions are non-standard, for example very rocky and requiring difficult excavation, or with low load bearing capability requiring extra large bases, the additional cost incurred because of such unusual conditions is usually stipulated to be billed to the customer. In some unusual situations, this may amount to a sizeable increase in tower cost.

Wind loading creates some misunderstandings, mainly in definitions. EIA Standard RS-222A on steel towers, and its companion RS-195A on antennas, which are normally used in specifying and determining wind loadings, contain recommendations and tables for different areas of the country. Figure 26 is a map of the U.S.A. divided into three loading zones. Zone A has a recommended minimum of 30 lbs./sqft, Zone B 40 lbs./sqft, and Zone C 50 lbs./sqft. The latter is typical of southeast coastal areas in the hurricane belt.

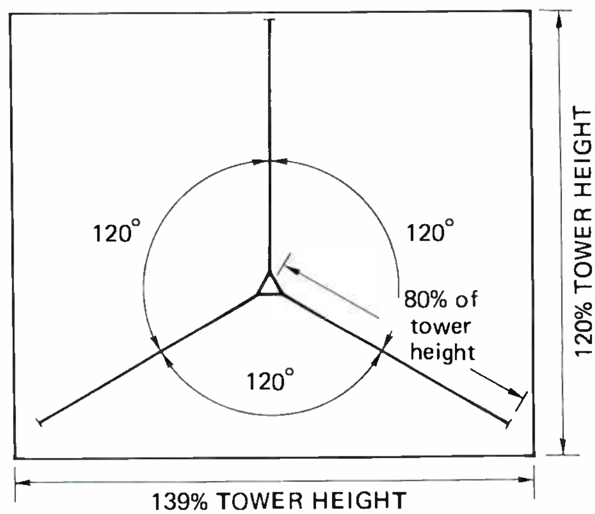
This wind loading is the "design" or "elastic limit" wind loading, though neither of these terms is very satisfactory. The point is that this design

PREFERRED AREA



Height (feet)	Area (feet)	Approx. Acreage
400	565 x 565	7.35
380	537 x 537	6.62
360	510 x 510	5.97
340	482 x 482	5.33
320	455 x 455	4.75
300	428 x 428	4.20
280	400 x 400	3.67
260	372 x 372	3.17
240	344 x 344	2.71
220	316 x 316	2.29
200	288 x 288	1.90
180	260 x 260	1.55
160	232 x 232	1.25
140	205 x 205	.97
120	177 x 177	.72
100	149 x 149	.51
80	121 x 121	.34
60	94 x 94	.20

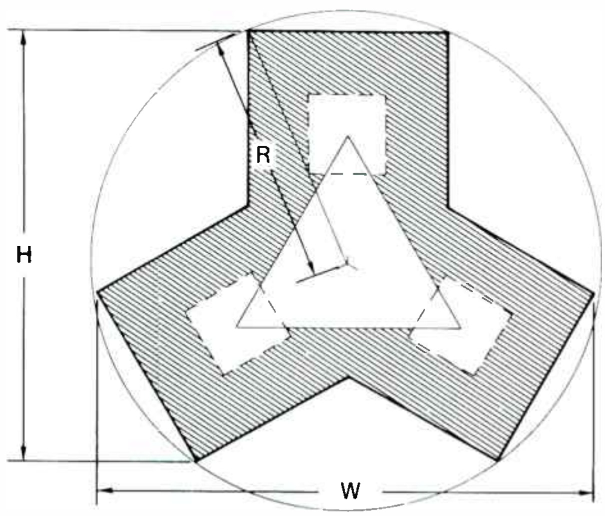
MINIMUM AREA



Height (feet)	Area (feet)	Approx. Acreage
400	489 x 565	6.34
380	465 x 537	5.73
360	441 x 510	5.17
340	417 x 482	4.62
320	393 x 455	4.10
300	369 x 428	3.64
280	345 x 400	3.18
260	321 x 372	2.74
240	297 x 344	2.35
220	273 x 316	1.98
200	249 x 288	1.64
180	225 x 260	1.34
160	201 x 232	1.08
140	177 x 205	.84
120	153 x 177	.62
100	129 x 149	.44
80	105 x 121	.29
60	81 x 94	.17

NOTE: Areas shown in tables include 80% guying plus additional allowance for guy anchors. Where space is restricted, special tighter guying can sometimes be used.

Figure 25A. Approximate Area Required For Guyed Tower.

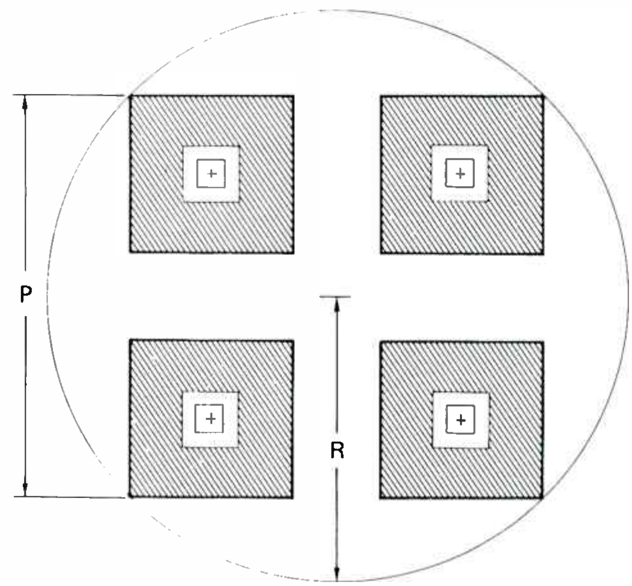


APPROXIMATE DIMENSIONS (Feet)

Tower Height	R	W	H
25	9.5	17.3	16.1
50	13.8	26.6	23.4
75	15.7	30.2	26.6
100	17.3	33.5	29.3
125	19.4	37.2	32.8
150	21.9	42.0	37.1
175	24.4	46.8	41.3
200	26.4	50.8	44.6
225	28.4	54.5	48.0
250	30.5	58.7	51.2
275	32.4	62.1	54.8
300	34.8	66.8	59.0
325	36.4	70.0	61.6
350	38.0	73.1	64.4

(Shaded portion represents rupture area)

Figure 25B. Approximate Area Required for 3-Leg Self Supported Tower.



APPROXIMATE DIMENSIONS (Feet)

Tower Height	R	P
50	15.4	21.8
75	17.8	25.3
100	20.5	28.9
125	22.8	32.2
150	25.4	35.9
175	27.9	39.5
200	30.6	43.3
225	33.2	46.9
250	35.8	50.6
275	38.1	53.8
300	40.8	57.6
325	42.6	60.0
350	44.5	64.0

(Shaded portion represents rupture area)

Figure 25C. Approximate Area Required for 4-Leg Self Supported Tower

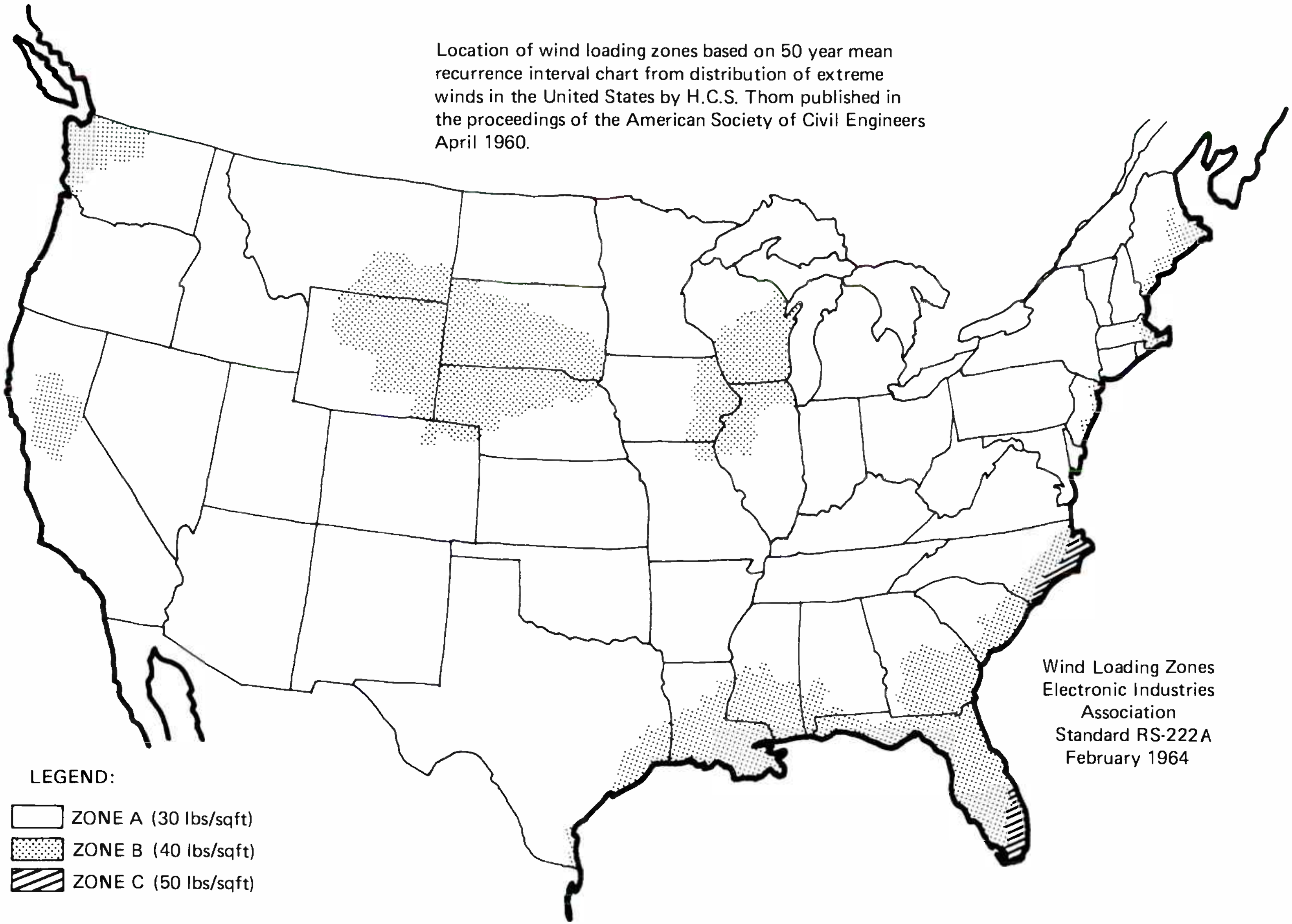


Figure 26. EIA Wind Loading Zones In The U.S.A.

loading is *not* associated with operational requirements (tower twist and sway), so if a wind velocity of this magnitude exists, the tower might twist or sway enough to substantially degrade the signal, but it should return to normal, or very nearly so, when the wind stops.

In addition to the "design" loading, which is the name-loading by which a tower is described, EIA specifies an *operational* loading for which the twist and sway limits of the tower should not degrade the signal by more than 10 dB. A table is included in RS-222A and RS-195A giving representative values for allowable twist and sway for various beamwidths.

> *Please note carefully:* The EIA operational minimum loading is based on 20 lbs./sqft wind load, regardless of what the design loading may be. If operational minimums higher than 20 lbs./sqft are desired, it is necessary for the user to call them out specifically so that the tower manufacturer can be asked to quote on a stronger tower, usually at a higher cost.

Wind loading increases as the square of the actual wind velocity. Expressed as a formula, $P = KV^2$, where P is the pressure in pounds per square foot, K is the wind conversion factor, and V is the actual wind velocity in miles per hour. EIA recommends the use of 0.004 as a nominal value of K for pressures on the projected areas of flat surfaces. Using this value for K, the approximate wind velocities corresponding to several common tower loadings are:

20 lbs/sqft = 71.0 mph
30 lbs/sqft = 86.0 mph
40 lbs/sqft = 100.0 mph
50 lbs/sqft = 112.5 mph

The loading on a tower depends very critically on the sizes, shapes, locations and relative positions of all the antennas, reflectors, waveguides and other paraphernalia which are mounted on it. A very important point is to review the possibility that additional paths or antennas may have to be added in the future, since such considerations could significantly affect the design of the initial tower.

A very important consideration in some areas is that of ice forming on towers and antennas. It is of particular significance in the case of mountaintop sites, especially those where moisture-laden winds may sweep up one side of the mountain and deposit a large portion of their moisture content near the top. In such cases almost unbelievable quantities of ice can build up on antenna and tower structures. In areas subject to icing, tower and antenna structures must be designed to carry the additional loading caused by the weight of the ice, plus added wind loading resulting from the increased surface areas.

Local building codes within cities may impose more severe restrictions on loading than those of EIA.

Roof mounted towers are sometimes required. These, if of any great size, will require careful evaluation of the structural adequacy of the building to support the proposed tower.

Within the United States, the FAA and FCC both have complex regulations covering tower heights, depending on proximity to airports or airlanes. They also have requirements on painting, lighting, and in some cases, obstruction marking of towers. Similar government entities in other countries regulate tower heights and markings within their jurisdictions, usually in accordance with recommendations of the ICAO (International Civil Aviation Organization).

D. Waveguide and Transmission Lines

Waveguide and transmission line is important not only for its loss characteristics, which enter into the path loss calculation, but also for the degree of impedance matching attainable, because of the effect on echo distortion noise. The latter becomes extremely important with high-density systems having long waveguide runs.

In the 2-GHz bands coaxial cable is usually used, and, except for very short runs, it is usually of the air dielectric type. Typical sizes are 7/8", with an attenuation of about 2 dB/100', and 1-5/8", at about 1.1 dB/100'. It is normally ordered in the exact lengths required, with factory installed and sealed terminal connectors. When the larger sized cable is used, it is desirable to reduce to 7/8", with a suitable transition, for flexibility in connecting to the radio equipment. In some cases similar treatment may be needed at the antenna end, though generally the use of a rigid right-angle connector will allow sufficient flexibility for antenna orientation.

The other bands use waveguide almost exclusively, one of three basic types; rigid rectangular, rigid circular, and semi-flexible elliptical. The elliptical type is of continuous construction, while the other types come in sections with flanges. Short sections of flexible waveguide are also used for the connections to the antennas and to the equipment. In all cases it is desirable to keep the number and length of flexible sections as small as possible, since they tend to have higher losses and poorer VSWR than the main waveguide types.

1. Rectangular Guide

Rigid rectangular waveguide is the most commonly used, with oxygen-free, high-conductivity copper (OFHC) the recommended material. The types and approximate characteristics are as follows:

4 GHz band	WR 229 is standard for most installations. It has a loss of approximately 0.85 dB/100'.
6 GHz bands	WR137 is normally used. It has a loss of approximately 2.0 dB/100'. In cases where, due to high towers, a reduced transmission loss is required, transitions can be supplied for use with WR159, which has a loss of about 1.4 dB/100'.
7-8 GHz	WR112 is normally used. Attenuation is approximately 2.7 dB/100'.
11 GHz	WR90 is normally used. Attenuation is approximately 3.5 dB/100'.
12-13 GHz	WR75 is normally used. Attenuation is approximately 4.5 dB/100'.

For the most critical applications, where extremely low VSWR is required to meet stringent noise performance specifications, special precision waveguide, manufactured to very tight tolerance, is recommended.

2. Circular Guide

Circular waveguide has the lowest loss of all, and in addition, it can support two orthogonal polarizations within the single guide. It is also capable of carrying more than one frequency band in the same guide. For example, WC281 circular guide is normally used with horn reflector antennas to provide two polarizations at 4 GHz and two polarizations at 6 GHz. But circular guide has certain disadvantages. It is practical only for straight runs, requires rather complicated and extremely critical networks to make the transition from rectangular to circular, and can have significant moding problems, when the guide is large enough to support more than one mode for the frequency range in use. Consequently, though circular waveguide is available in several different sizes, and its low losses make it attractive, it is recommended that it be used with considerable caution.

3. Elliptical Guide

Semi-flexible elliptical waveguide is available in sizes comparable to most of the standard rectangular guides, with attenuations differing very little from the rectangular equivalents. The distinctive feature of elliptical guide is that it can be provided and installed as a single continuous run, with no intermediate flanges. When very carefully transported and installed it can provide good VSWR performance, but relatively small deformations can introduce enough impedance mismatch to produce severe echo distortion noise.

The most commonly used types and their approximate characteristics are as follows:

4 GHz band:	EW-37,	approximately .85 dB/100'.
6 GHz bands:	EW-~~59~~, 64	approximately 1.75 dB/100'.
7-8 GHz:	EW-71,	approximately 2.5 dB/100'.
11 GHz band:	EW-107,	approximately 3.7 dB/100'.
12-13 GHz bands:	EW-122,	approximately 4.5 dB/100'.

In all types of waveguide systems it is desirable to keep the number of bends, twists, and flexible sections to a minimum. It is also vitally important to use great care in installation, since even very slight misalignments, dents, or introduction of foreign material into the guides can create severe discontinuities.

E. Antenna Systems

Highly directional antennas are used with point-to-point microwave systems. By focusing the radio energy into a narrow beam that can be directed toward the receiving antenna, the transmitting antenna can increase the effective radiated power by several orders of magnitude over that of a non-directional antenna. The receiving antenna also, in a manner analogous to that of a telescope, can increase the effective received power by a similar amount.

Although gain is the primary characteristic, there are other antenna characteristics which are of importance in communications systems. Antenna beam-width, side-lobe magnitudes, off-axis radiation and sensitivity patterns, and polarization discrimination are of great significance for frequency coordination purposes. Impedance match (usually expressed as VSWR, though return loss is a much more useful parameter) across the band to be used is of great importance in situations where echo distortion is significant. Consequently, it is no longer sufficient merely to select an antenna system for optimum gain efficiency.

Nevertheless, it is basic that the antenna system must have enough gain so that the desired net path loss between transmitter output and receiver input is attained. The required antenna gains are determined by a calculation which involves a knowledge of the transmitter output power, fixed losses of waveguides, circulators, hybrids, radomes, and any other items between the transmitter and its antenna, and between the receiver and its antenna, the unfaded path attenuation, and the receiver strength needed to give the required noise performance and fade margin. The calculations are usually formalized and recorded in a "Path Data Sheet". An example is given in Section VI.

The gain of an antenna is expressed in dB relative to the gain of an isotropic antenna, which is a theoretical omnidirectional antenna, with a gain which by definition would be 1, or 0 dB. At a given operating frequency, the gain of an antenna (either transmitting or receiving) is a function of the effective area and is given by:

$$G = 10 \log_{10} (4\pi A e/\lambda^2) \qquad (24)$$

where G = gain over isotropic, in dB
A = area of antenna aperture
e = antenna efficiency
λ = wavelength at operating frequency, in same units as A

The following paragraphs list some of the more commonly used antennas or antenna systems, with some descriptive comments.

1. Direct Radiating Antennas

Parabolic Antennas

This type of antenna consists of a parabolic dish, illuminated by a feed horn at its focus. Available in a wide variety of sizes, with diameters of 2', 4', 6', 8', 10' and sometimes 12' and 15' in most frequency bands.

The simplest form is with single plane polarized feed, which can be either vertical (V) or horizontal (H). Others have dual polarized feeds (DP), with separate V and H connections. DP's usually have a bit less gain than single polarized, because of the more complex feedhorns.

Off-beam discrimination is reasonably good, but front-to-back ratios on the order of 45 to 50 dB maximum are generally not adequate for back-to-back transmission (or reception) of the same frequency in both directions. Often available with special low VSWR feeds (on the order of 1.05 to 1).

The gain efficiencies of most commercially available parabolic antennas are in the order of 55 to 65%. With 55% efficiency, the gain of a parabolic antenna is given by:

$$G = 20 \log_{10} B + 20 \log_{10} F + 7.5 \qquad (25)$$

where G = gain over isotropic, in dB
F = frequency in GHz
B = parabola diameter in feet

Although this formula can be used for estimating purposes, actual gain should be determined from the manufacturer published specifications. In the higher bands (11 to 13 GHz particularly), and with cross-band and other complex feed horn systems, the efficiencies can often be considerably lower than given above, and actual gains may easily be 1 to 2 dB lower than that given by equation (25).

The half-power beamwidth of a parabolic antenna is given approximately by:

$$\phi = 70/FB \tag{26}$$

where ϕ = half-power beamwidth in degrees
F = frequency in GHz
B = parabola diameter in feet

Side-lobes and front-to-back ratio are caused by imperfect illumination of the parabola, phase errors introduced by the feed, and irregularities in the reflecting surface. Antenna patterns, usually available from the manufacturer, give the radiation in all directions on both principal planes relative to the main beam. Such patterns are necessary for interference studies.

In using such antenna patterns for interference studies, it should be kept in mind that it is not just the response of the receiving antenna to the opposite polarization that is important. A transmitting antenna on one polarization will radiate power on the other polarization as well, in accordance with *its* cross-polarization pattern. If the transmitting antenna has poor cross polarization discrimination at the angle looking toward a receiving antenna which is oppositely polarized, no amount of discrimination at the latter will reduce the component it receives from the cross radiation of the transmit antenna. Thus, in an interference analysis, one must take as the polarization discrimination, the worst combination of transmitter direct to receiver opposite, or transmitter opposite to receiver direct.

High-Performance or Shrouded Antennas

These are similar to the common parabolic types, except that they include a cylindrical built-out shield which helps to improve the front-to-back ratio, and the wide-angle radiation discrimination.

Shrouded antennas are usually available as either single polarized or double polarized. Gain efficiency is usually slightly poorer than that of the simple parabolas.

They are substantially bulkier, heavier, and more expensive than the ordinary parabolas. However, they can provide front-to-back ratios on the order of 65 dB, sufficient in many cases, to allow back-to-back transmission of the same frequency in both directions. This is the primary reason for their use. Special feeds are usually required for very low VSWR (1.05 to 1 or so) applications.

Cross-Band Parabolic Antennas

These are parabolic antennas with feeds designed to permit operation in two widely separated bands (for example, 6 and 11 GHz). Because of the very complex and critical feed assemblies, these antennas typically have somewhat reduced gain (a dB or so) and poorer VSWR than single-band antennas. Problems have also been experienced with interaction between the two feed systems, since the 6 GHz waveguide can propagate spurious 11 GHz modes. A commonly used method of solving the interaction problem is to insert in the 6 GHz feed-line, at a point as close as possible to the antenna, a filter which attenuates the 11 GHz energy while allowing the 6 GHz energy to pass through essentially without loss.

Horn Reflector Antennas

The horn reflector (cornucopia) antenna has a section of a very large parabola, mounted at such an angle that the energy from the feed horn is simultaneously focused and reflected at right angles. The standard Bell System horn antenna is about the equivalent of a 10' parabola insofar as gain is concerned. But it has much higher front-to-back ratios (on the order of 70 dB or more); sufficient to allow operation in two directions (or more) from a station on the same frequencies.

It has good VSWR characteristics, and, with suitable coupling networks (which are quite complex), can be used for multiband operation on both polarizations. However, there are some moding problems, particularly at the higher frequencies, which, if uncorrected, can cause severe distortion.

Disadvantages are that this antenna is very big, heavy,and complex as to mounting, and quite expensive not only for the antenna itself, but for its effect on mounting and tower costs. Almost no flexibility in choice of sizes, though at least one smaller size of horn reflector is currently available.

The following table can be used for preliminary estimates of antenna gains in the various bands. Final gains should be those guaranteed by the manufacturers of the particular antenna to be used.

Table I. Antenna Gains For Estimating Purposes.

Plane Polarized Parabolic Antennas. (DP's, HP's and Cross-band somewhat lower)							
Diameter in feet	Gain Relative To Isotropic – dB						
	2 GHz	4 GHz	6 GHz	7 GHz	8 GHz	11 GHz	13 GHz
4	25.5	– –	35.2	35.9	37.0	40.3	41.3
6	29.0	35.0	38.7	39.4	40.6	43.8	44.8
8	31.5	37.3	41.1	41.9	43.1	46.0	47.3
10	33.5	39.3	43.0	43.9	45.2	47.7	48.5
12	– –	40.8	44.6	45.5	46.7	– –	– –
15	– –	42.6	46.0	46.9	48.7	– –	– –
Horn Reflector Antennas							
8X8 (Std)	– –	39.4	43.0	– –	– –	47.4	– –
6 (Circ)	– –	35.7	39.4	– –	– –	43.8	– –

2. Periscope Antenna Systems

In many cases, usually where considerable height is required or waveguide runs are difficult to make, a periscope antenna system is used. It consists of a parabolic radiator, generally at or near ground or building level, illuminating a reflector at the top of the tower. Gain is a complex function of the antenna and reflector sizes and separation, the frequency, and the geometric relationships.

A periscope system allows waveguide runs to be kept short, thus reducing losses and improving the situation with respect to echo distortion, since the latter is a function of waveguide length as well as impedance match. With suitable choices of combinations, and over certain ranges of separations, antenna-reflector combinations can give net gains equal to, or even greater than, the gain of the parabola alone, thus in effect eliminating most of the losses due to waveguide.

For this reason the periscope system is often, from the point of view of overall net gain, the most efficient antenna system, particularly where high towers are needed. The periscope system has had very wide usage in the bands from 6 GHz up, but it is relatively inefficient at 2 GHz, and has had very little usage at 4 GHz.

It is generally accepted, though with some dissenting opinions, that periscope antenna systems do not have particularly good off-beam discrimination characteristics, and for this reason their use in high density systems and in very congested areas leaves something to be desired. Consequently, there has been something of a trend toward increased usage of direct radiating antennas, despite the added loss and echo distortion problems associated with the long waveguide runs, which they typically require.

In their most common configuration, periscope antenna systems have the illuminating dish at the bottom of the tower, directly under the reflector. Assuming a horizontal path, the latter will then be mounted so that its face makes an angle of 45° with the horizontal. In some cases, for example space diversity, physical relationships may dictate a "skewed" periscope arrangement, with the dish moved off to one side, and sometimes forward or backward along the path, with respect to the distant end.

In such cases, the reflector will typically have to be rotated around its vertical axis so that its normal no longer points directly along the path; the tilt angle in general will no longer be 45°, but will depend on the actual configuration.

Calculations of the tilt, amount of rotation away from the path, and the effect of the "skew" on the gain are somewhat complicated. Graphical solutions can be made using Lenkurt drawing EEH-20020, a copy of which is included in the Appendix section.

In current practice, the reflectors used in periscope systems may be rectangular, rectangular with some or all of the corners truncated, or elliptical. The surface of the reflector may be flat, or it may be constructed so that it can be adjusted to have a curvature approximating that of a section of a large paraboloid, with the focus at the radiator.

A number of theoretical studies have been published on periscope antenna system gains, and these, together with experimental data and results from systems in practical use, provide the basis for gain curves commonly employed by microwave manufacturers and users. It should be borne in mind that in all the studies a number of simplifying assumptions had to be made to render the mathematical processes tractable. The results therefore cannot be regarded as exact.

Curved reflectors provide somewhat greater gain than flat reflectors over most separation ranges, and are believed to have somewhat better discrimination characteristics. There are differences of opinion about the merits of the curved reflectors, but Lenkurt's experience has been that when they are properly installed and aligned, and the curving very carefully done, they perform as predicted. Best results have been found to be obtained by carefully adjusting the curvature, prior to installation, to calculated values which will cause the reflector surface to closely approximate a paraboloidal section with its focus at the illuminating dish.

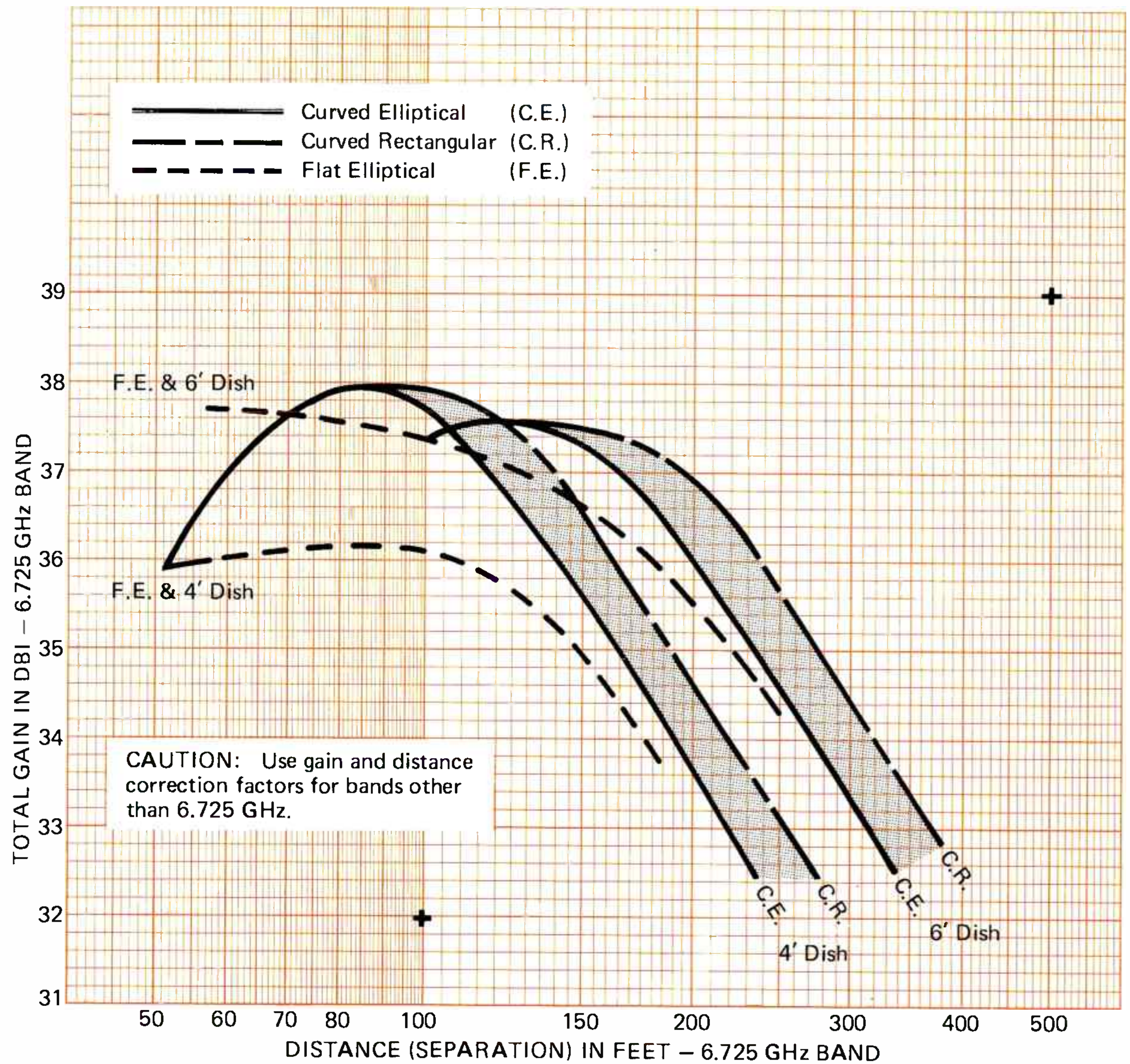
Figures 27A through 27D are charts of gain-vs-separation distance for various combinations of antennas and reflectors. Each chart covers one of the standard reflector sizes, together with several antenna sizes. Each chart covers three types of reflectors. The heavy solid lines are for curved elliptical reflectors (C.E.) and the broken solid lines are for curved rectangular reflectors (C.R.), while the dashed lines are for flat elliptical reflectors (F.E.). In the case of the curved reflectors, the surfaces are assumed to be adjusted to an accurate approximation of a paraboloid with the focus at the illuminating antenna. For small separation distances the gain of the flat reflectors will exceed that of the curved reflectors, and it will be noted that the lines for the curved reflectors are truncated at that point in each case.

For the case of the curved reflectors, it will be noted that for each antenna-reflector combination there is — to the right of the maximum point — a bifurcated and shaded section bounded on the left by the (C.E.) line and on the right by the (C.R.) line. In this area to the right of the maximum point, the greater size of the rectangular reflector comes into play and results in a somewhat greater gain. In the case of small separations, the "corner" areas of the rectangular reflector are no longer as effective, and may even become slightly detrimental. Consequently we have, for conservatism, ignored the extra size of the rectangular reflector in the area to the left of the maximum point, treating them the same as the elliptical.

When using the curves, the heavy solid lines should be used to the left of the maximum, whether the reflector is elliptical or rectangular. To the right of that point, the heavy solid line should be used for elliptical reflectors, and the broken solid line for fully rectangular reflectors. Reflectors with two corners clipped can be estimated between the two lines, depending on the degree of clipping. Those with all four corners clipped can be treated the same as elliptical. At separations where a flat reflector exhibits greater gain it should, of course, be preferred to a curved reflector.

The gains as given by the charts are for the standard periscope arrangement, with the reflector at a 45° angle and located directly (or nearly so) above the illuminating dish. In this case the projected area of the reflector is equal to .707 x (true area) and is approximately square for the rectangular type and circular for the elliptical type. For skewed arrangements the projected shape will change, and in some cases the projected area also. Moving the dish straight out to one side at a right angle to the path leaves the projected area unchanged. A dish location forward from the reflector increases the projected area and a location backward from it decreases the projected area.

Skewed shots also make it more unlikely that the correct parabolic curvature can be achieved. For all these reasons it is necessary to use considerable caution when applying the curves of Figure 27 to skewed arrangements, particularly those for curved reflectors.



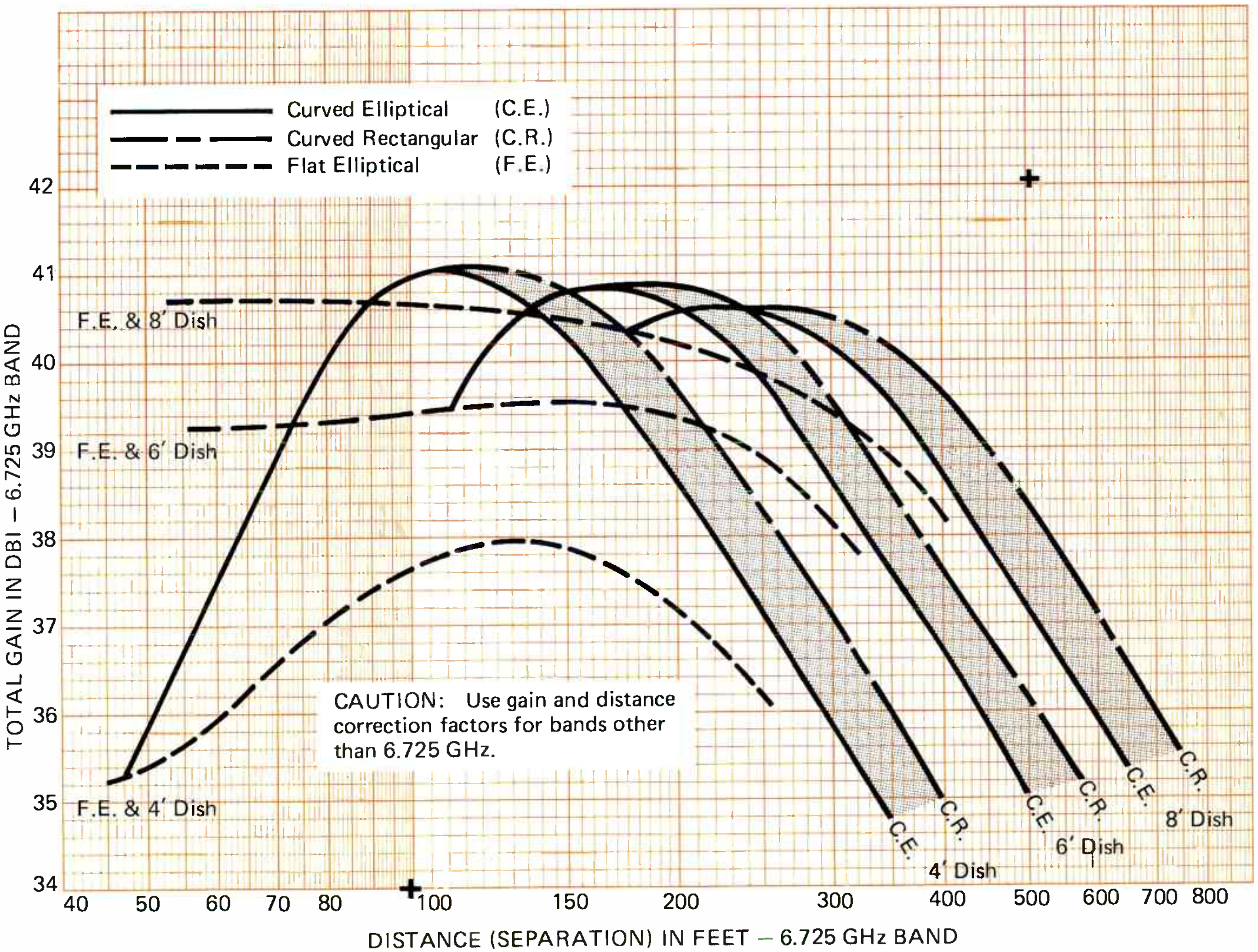
Band GHz	Gain Factor	Distance Factor
1.92	−10.8	3.50
2.15	− 9.9	3.13
3.95	− 4.6	1.70
4.7	− 3.1	1.43
6.175	− 0.7	1.09
6.725	—	—
7.0	+ 0.4	.96
7.437	+ 0.9	.91
8.0	+ 1.5	.84
11.2	+ 4.5	.60
12.45	+ 5.4	.54
14.8	+ 6.9	.46

True Gain = Chart Gain + G Factor

$$\text{True Dist.} = \frac{\text{Chart Dist.}}{\text{Dist. Fact.}}$$

Chart Dist. = True Dist. x Dist. Factor

Figure 27A. Periscope Gain Curves for 6'x8' Reflectors



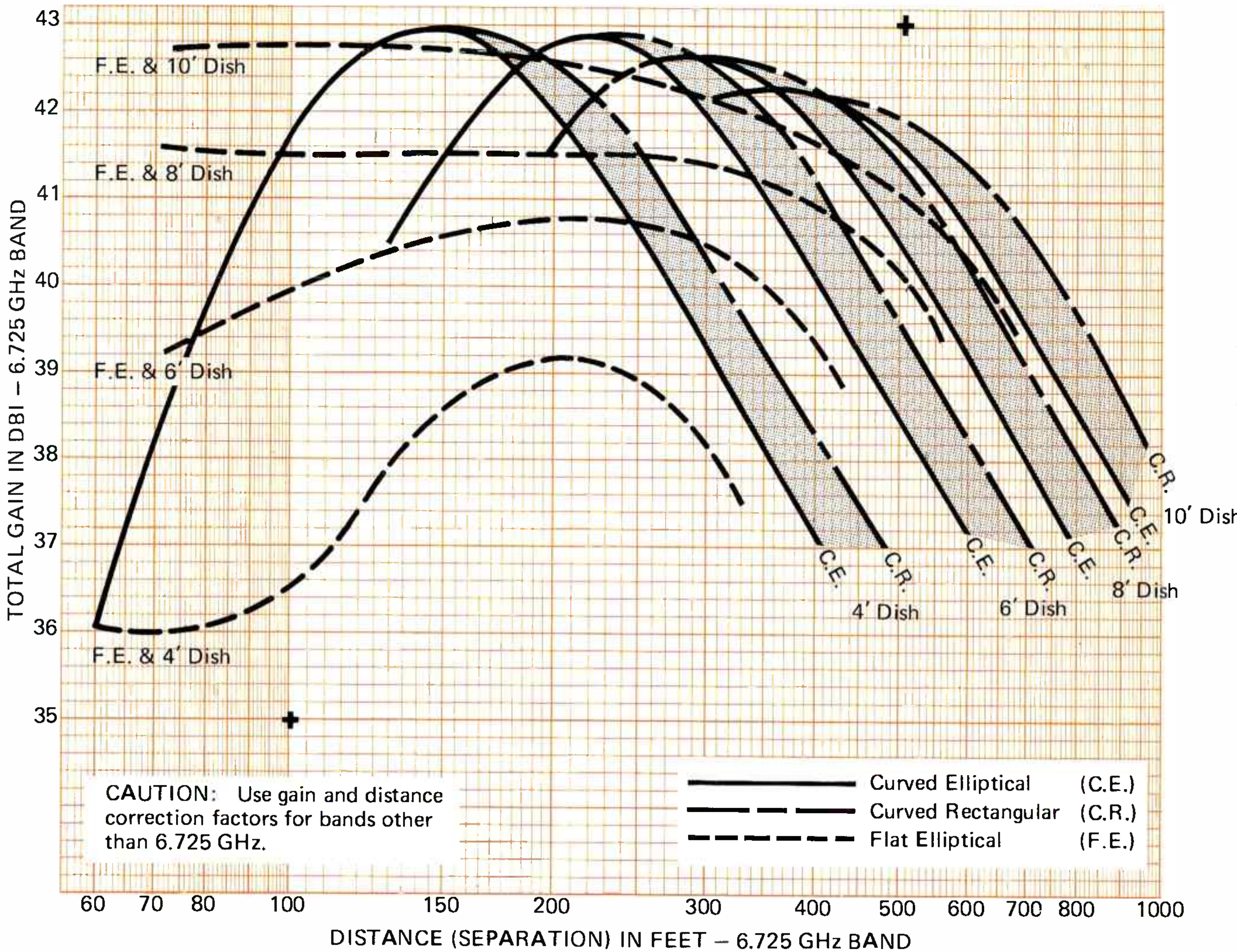
Band GHz	Gain Factor	Distance Factor
1.92	−10.8	3.50
2.15	− 9.9	3.13
3.95	− 4.6	1.70
4.7	− 3.1	1.43
6.175	− 0.7	1.09
6.725	–	–
7.0	+ 0.4	.96
7.437	+ 0.9	.91
8.0	+ 1.5	.84
11.2	+ 4.5	.60
12.45	+ 5.4	.54
14.8	+ 6.9	.46

True Gain = Chart Gain + G Factor

True Dist. = $\dfrac{\text{Chart Dist.}}{\text{Dist. Fact.}}$

Chart Dist. = True Dist. x Dist. Factor

Figure 27B. Periscope Gain Curves for 8’x12’ Reflectors



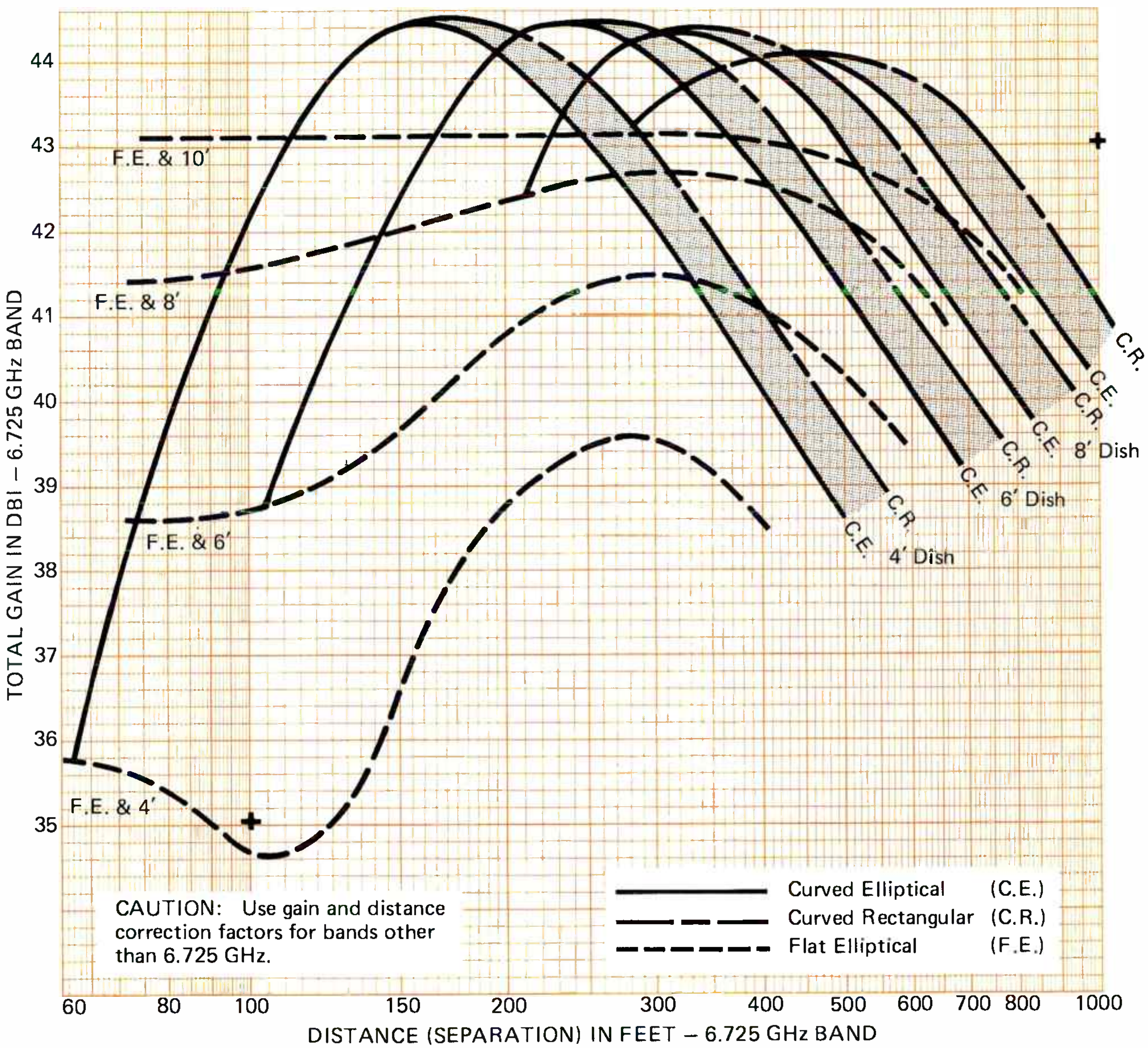
Band GHz	Gain Factor	Distance Factor
1.92	–10.8	3.50
2.15	– 9.9	3.13
3.95	– 4.6	1.70
4.7	– 3.1	1.43
6.175	– 0.7	1.09
6.725	–	–
7.0	+ 0.4	.96
7.437	+ 0.9	.91
8.0	+ 1.5	.84
11.2	+ 4.5	.60
12.45	+ 5.4	.54
14.8	+ 6.9	.46

True Gain = Chart Gain + G Factor

$$\text{True Dist.} = \frac{\text{Chart Dist.}}{\text{Dist. Fact.}}$$

Chart Dist. = True Dist. × Dist. Factor

Figure 27C. Periscope Gain Curves for 10’x15’ Reflectors



Band GHz	Gain Factor	Distance Factor
1.92	–10.8	3.50
2.15	– 9.9	3.13
3.95	– 4.6	1.70
4.7	– 3.1	1.43
6.175	– 0.7	1.09
6.725	–	–
7.0	+ 0.4	.96
7.437	+ 0.9	.91
8.0	+ 1.5	.84
11.2	+ 4.5	.60
12.45	+ 5.4	.54
14.8	+ 6.9	.46

True Gain = Chart Gain + G Factor

True Dist. = $\frac{\text{Chart Dist.}}{\text{Dist. Fact.}}$

Chart Dist. = True Dist. x Dist. Factor

Figure 27D. Periscope Gain Curves for 12'x17' Reflectors

The charts are intended to represent the total gain of the periscope system when the efficiency of the illuminating dish is 55%, but have been derated by about 0.5 dB to allow for some of the variables. If antennas with gain efficiencies lower than 55% are used, a further derating should be made. *Also when radomes are used, the radome loss must be deducted from the overall gain, if it is not accounted for in the fixed losses.*

The charts give directly the gain of the periscope system in dBi (dB with respect to an isotropic antenna) at 6.725 GHz, the center frequency of the 6.575 to 6.875 GHz industrial band, as a function of the separation distance between the dish and the reflector. Thus for this band the charts can be used directly.

In order to allow the charts to be used for other frequency bands, a table of correction factors is provided covering the center frequencies of most of the bands of interest. For each frequency, a gain factor and a distance factor are given.

The use of the gain factors is straightforward and self-evident. They are simply added algebraically to the gains shown by the chart. For example, a chart gain reading of 38 dB would, at 1.92 GHz, give 38 — 10.8 = 27.2 dB. Conversely, a true gain of 27.2 dB at 1.92 GHz would correspond to an apparent (chart) gain of 27.2 — (-10.8) = 38.0 dB.

The use of the distance factor, however, is not quite so self-evident, and some care is needed to make sure that it is properly used. The factors as given in the table are such that they are used as multipliers when converting from the true distance (at some frequency other than 6.725) to the chart distance, and as divisors when converting from the chart distance to the true distance. The reason for setting it up this way is that the transmission engineer normally will know the true distance and will want to calculate the chart distance, and most people find multiplication somewhat easier than division.

It is very important to recognize that the true separation distance at frequencies *lower* than 6.725 GHz will be *less* than those shown on the charts, and the true separation distances at frequencies *higher* than 6.725 GHz will be *greater* than those shown. The point is emphasized because there is a natural tendency to assume the opposite, because of the fact that the lower frequencies have longer wavelengths. A good mnemonic device is: "A higher frequency means greater gain and greater separation, and a lower frequency means lower gain and lower separation". To illustrate the point, consider a 10' antenna with a 12' x 17' reflector. Figure 27D indicates that such a combination will give a gain of about 44.1 dB at a separation distance of 450 to 500 feet, at 6.725 GHz. But at 1.92 GHz the gain would be reduced to 44.1 — 10.8 = 33.3 dB, and the separation distance would by reduced to about 130 to 140 feet. Furthermore, at this relatively short separation distance, a rather large curvature would be required on the reflector face, a difficult mechanical problem. These facts indicate rather clearly why periscope systems are seldom used in the lower frequency bands.

In the practical case the transmission engineer usually knows the frequency band, the true separation distance and the required gain for the antenna system. The following example illustrates the procedure:

Assume; Band, 7.437 GHz. Separation distance 280'. Required gain 41 dB. From the tables, the corresponding chart distance will be 280 x .91 = 255' and the corresponding chart gain will be 41 — 0.9 = 40.1 dB.

One possible combination, from Figure 27B, would be an 8' dish with an 8' x 12' reflector which could be either C.E. or C.R. Other combinations could also be used, for example, a 6' dish with a C.R. reflector.

Though the chart presentation method used here is somewhat less simple to use than separate sets of charts for each band, it makes it possible to cover 12 frequency bands with only four charts instead of nearly fifty. Also the tables can be easily extended to frequencies other than those shown. For a frequency of X GHz, the required distance factor is simply $\frac{6.725}{X}$, and the gain factor is $20 \log_{10} \frac{X}{6.725}$.

It is also a simple matter, using 2-decade semi-logarithmic paper of the same type used in these charts, to prepare a separate set of charts covering any desired frequency band, by an overlay and tracing process. To facilitate this process, two small crosses are placed at appropriate spots on each of the graphs. By calculation using the above factors, one can determine where the points corresponding to those crosses should fall on the new sheet and place a pair of dots at those points. The two points are then overlaid exactly on the two crosses and the various curves traced on the new sheet.

The gain-separation charts have been derived from various published theoretical studies, and slightly modified toward the conservative side. They have been used successfully to calculate gains for systems in several — but not all — of the bands covered in the tables. The charts are published here for information only, and Lenkurt makes no guarantee of their accuracy.

It should be remembered that the requirements on trueness of reflector face, and on allowable deflections, become much more severe as the frequency is increased, and are quite severe in the bands of 11 GHz and higher. Twist and sway requirements at these higher frequencies also impose strong restraints on tower design.

Sway requirements require special consideration in periscope systems, because the effects of tower sway (but not twist) are doubled as a result of the reflection. Thus the sway tolerances are twice as stringent as those of an equivalent direct radiator. RS-195A provides a table with representative values of twist and sway as a function of antenna beamwidth, as well as a graph of antenna and reflector beamwidth as a function of antenna and reflector size and the frequency band. The half-power beamwidth of an elliptical reflector is approximately 60/FW degrees, and that of a rectangular reflector approximately 52/FW degrees, where F is the frequency in GHz and W is the width in feet of the narrow dimension of the reflector.

One potential problem with periscope systems and, to a lesser extent, with passive repeaters, is that of a "sneak" echo path existing between the illuminating dish and the distant end. Unless this direct path is thoroughly blocked by intervening terrain, some signal may get through directly, particularly under super-refractive conditions, and, if its magnitude is sufficient, can cause serious intermodulation problems. The direct path will be shorter (usually by about the distance of the antenna-reflector separation) and it will be a "leading" instead of a "trailing" echo, but the distortion effect is the same. Since echo signals with long delays can cause significant intermodulation, even if they are 50 to 60 dB below the main signal, it is apparent that a good deal of blocking is required. In some cases it has been found expedient to place metal shields on the path side of the illuminating dish to cut down the direct signal.

F. Radomes

Horn reflector and shrouded types of antennas usually include integral radomes, whose losses are taken into account in the manufacturer's published gain figures. Parabolic dish gains, however, usually do *not* include radome losses, and, if radomes are to be used, their losses must be ascertained (from the manufacturer catalog) and added in with the other fixed losses. The amount of loss may vary from less than 0.5 dB for a typical unheated radome at 6 GHz, to almost 2.0 dB for a typical heated radome in the high frequency bands.

Radomes also can be expected to degrade the VSWR as compared to that of the antenna without a radome. This becomes very significant in situations where very low VSWR's are needed to control echo distortion. In some cases radomes have been found to create highly reflective "spikes" at particular frequencies, and if these coincide with a used RF channel frequency, the results can be a high degree of distortion in that channel.

G. Passive Repeaters

Where a direct microwave path cannot be established between two points because of some geographical or man-made obstacle, it is sometimes possible to establish a path by way of a passive repeater. The function of such a passive is as a "beam redirector" to pass the microwave beam around or over something which would obstruct its direct path. A firm requirement is that there be radio "line-of-sight", with adequate clearances, between the passive and each of the end points.

There are two general types of passive repeaters in common use. One consists of two parabolic antennas connected back-to-back through a short piece of waveguide. Because it is relatively inefficient, this type of passive repeater is seldom used except on extremely short paths. The other, and more common type, is the flat "billboard" type metal reflector, which acts as a microwave mirror. With surfaces of adequate flatness it is close to 100% efficient, as compared to about 55% efficiency for antennas. Furthermore, the passive reflector acts both as a receiving antenna and a retransmitting antenna, and its "gain" is therefore applied twice.

Billboard passives fall into two basic configurations, depending on the geometric relationships. If the site of the passive repeater is off to one side, or behind one terminal, so that the included angle between the two paths at the reflector is less than about 130° (the smaller the angle the better), a single billboard can be used. This is the most common application.

However, if the only available location happens to be more or less in line with the path, a double billboard may be needed, consisting of two reflectors usually fairly close together and geometrically arranged to reflect the beam at the proper angles. Double billboards are applicable in situations where the effective change in beam direction at the passive repeater is to be less than about 50°.

Billboard reflectors are available in a variety of sizes, up to as large as 40' x 60'. They have typically been used mainly at frequencies of 6 GHz and higher. They can be used at lower frequencies, but because of the vastly increased gain at the higher frequencies, and the fact that the added gain factor appears twice for the billboards while the added path loss factor appears only once, the billboards are far more efficient at the higher frequencies.

Passive repeater gain calculations are somewhat complicated. For single billboards, things are simple if the billboard is in the far field of both antennas. In such a case the antenna gains and the billboard gains are independent and do not interact with each other. Then we simply calculate a total path loss which is the sum of the two separate path losses, and from it subtract the sum of the two antenna gains and the two-way gain of the reflector, in order to arrive at the net end-to-end path loss through the reflector.

The following formula can be used to calculate the free-space, two-way gain of a single passive billboard:

$$G = 22.2 + 40 \log_{10} F + 20 \log_{10} A + 20 \log_{10} \cos \alpha \qquad (27)$$

where
- G = two-way gain in dB
- F = frequency in GHz
- A = actual area of the passive in square feet
- α = one-half of the included angle between the two paths at the passive. (This last term in effect converts A into its projected or effective area).

A useful formula for "rule of thumb" determination of the approximate boundary of the near zone for an antenna-reflector combination is:

$$d = \frac{2B^2}{\lambda} = \frac{2FB^2}{0.984} \approx 2FB^2 \qquad (28)$$

where
- d = distance between antenna and reflector in feet
- B = diameter of dish, or widest projected dimension of the reflector, whichever is larger, in feet
- λ = wavelength in feet
- F = frequency in GHz

Figure 28A gives the two-way free-space gains of various sizes of passive as a function of frequency and the included horizontal angle.

When the passive is so close to one end that it is in the near-field of that antenna, as indicated by equation (28), the antenna and reflector gains are no longer independent, but react with each other in such a way that the net gain would be reduced. In this case the above methods cannot be used, since they would give overly optimistic results.

One way of attacking the near-field situation, is to treat the antenna and the nearby passive in the same fashion as a "periscope" antenna system. In this case, a "correction factor" is calculated and applied to the gain of the antenna, to obtain the net gain of the periscope combination. Since this gain is referred to the location of the *reflector*, the "path" in this method is simply that from the reflector to the more distant end. The shorter path simply disappears from the calculation. Figure 28B provides curves for deriving a "periscope" correction factor.

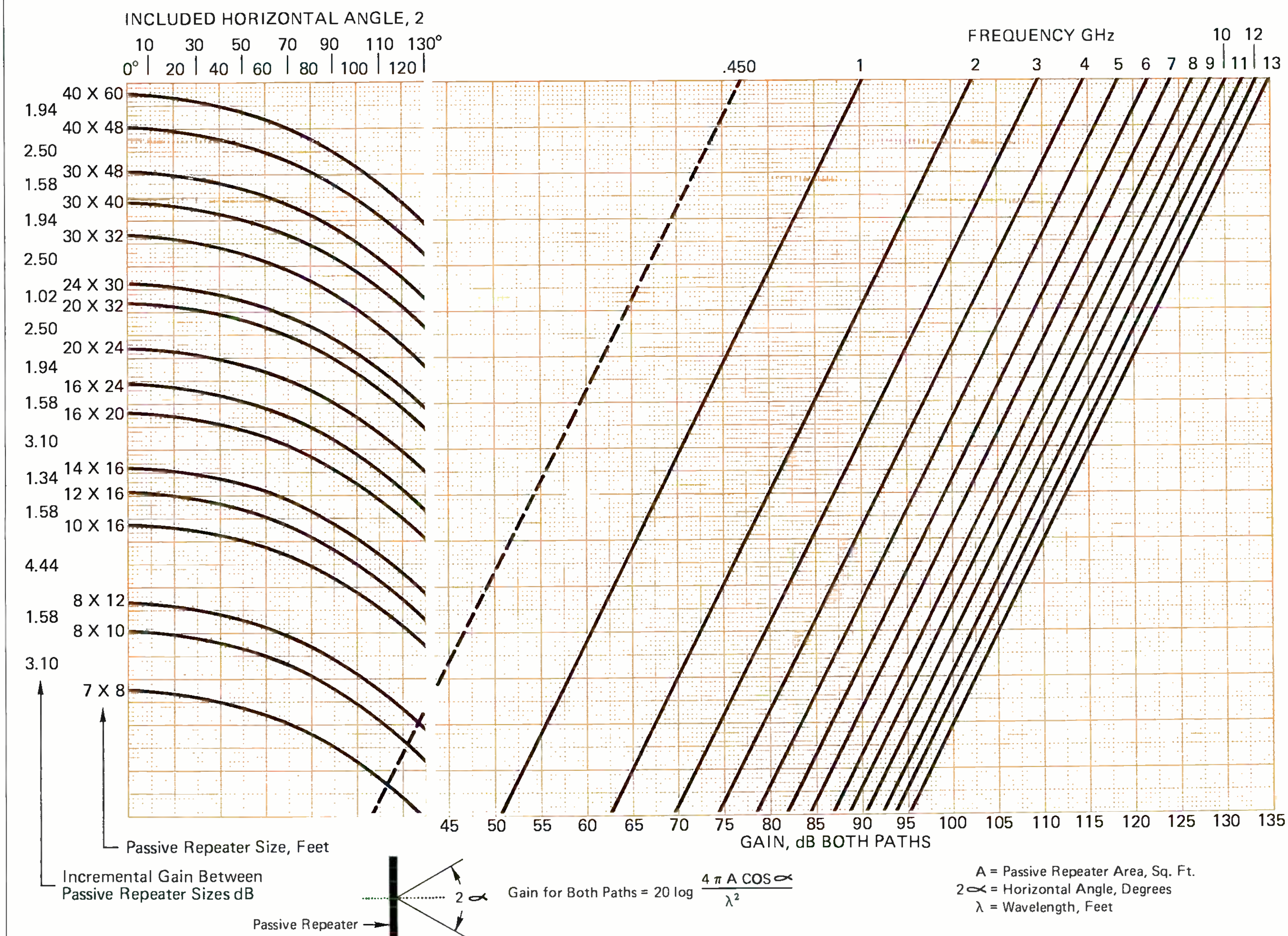


Figure 28A. Passive Repeater Gain Chart

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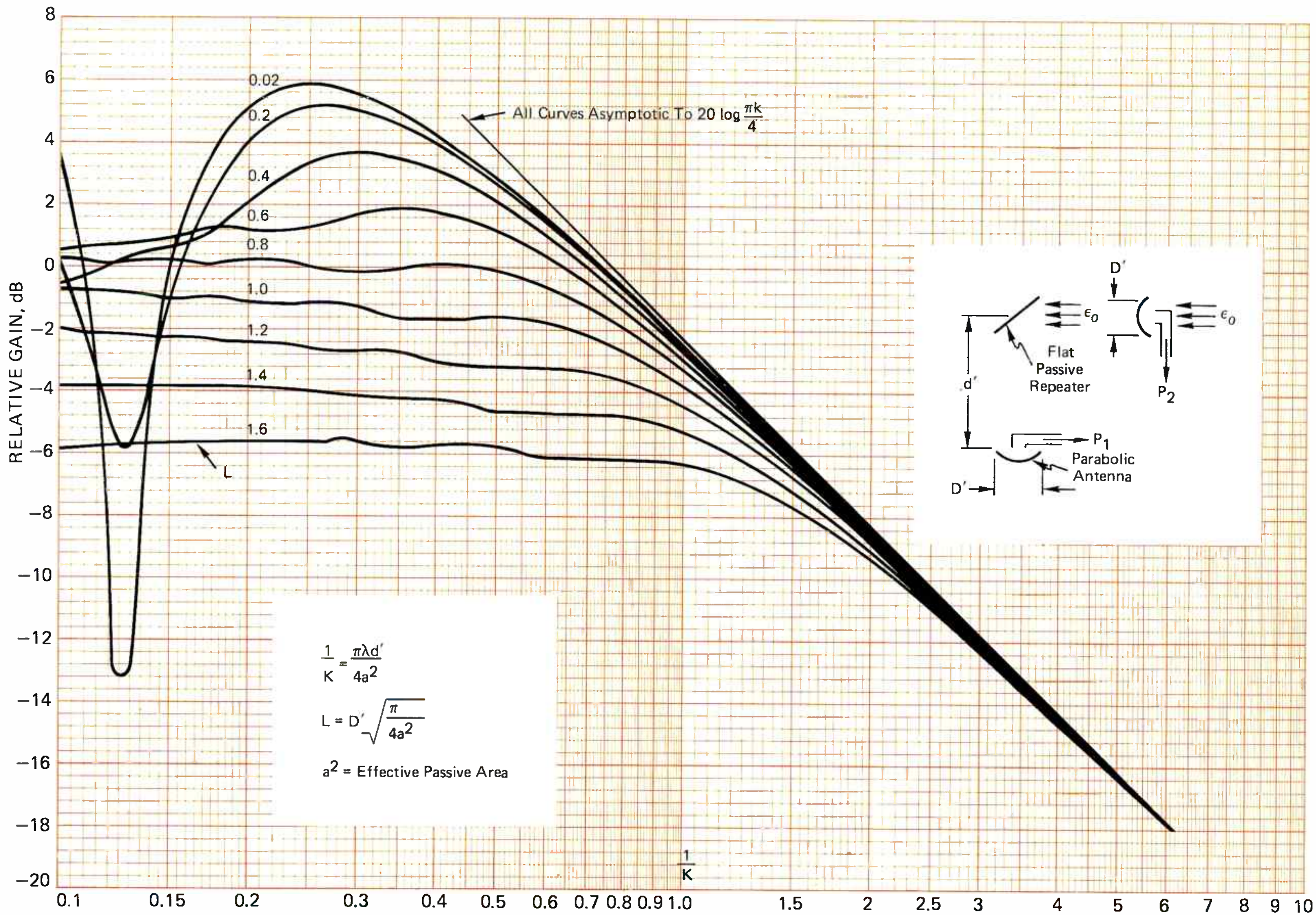

Figure 28B. Antenna-Reflector Efficiency Curves

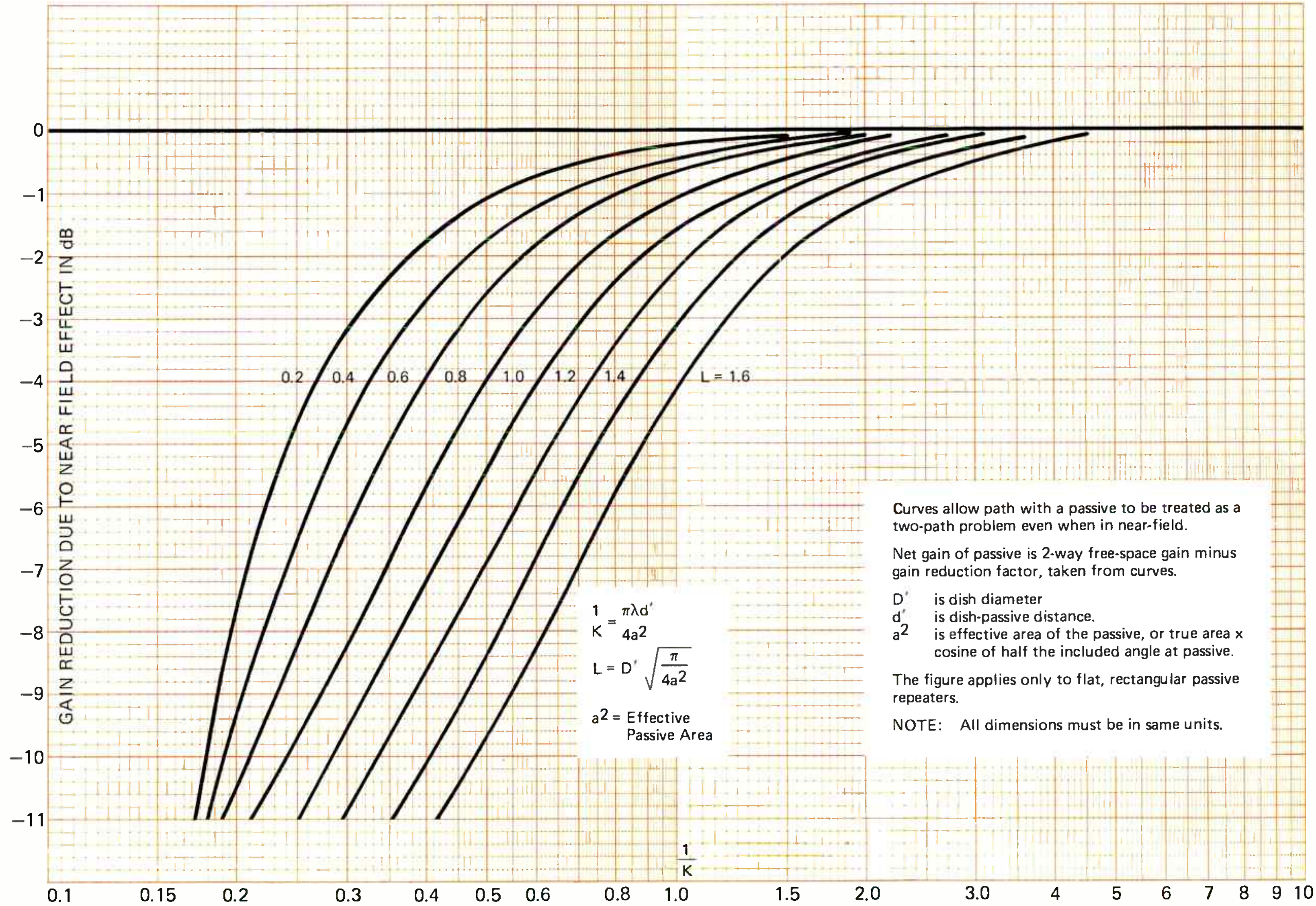

Figure 28C. Passive Repeater Gain Correction; When Passive In Near Field

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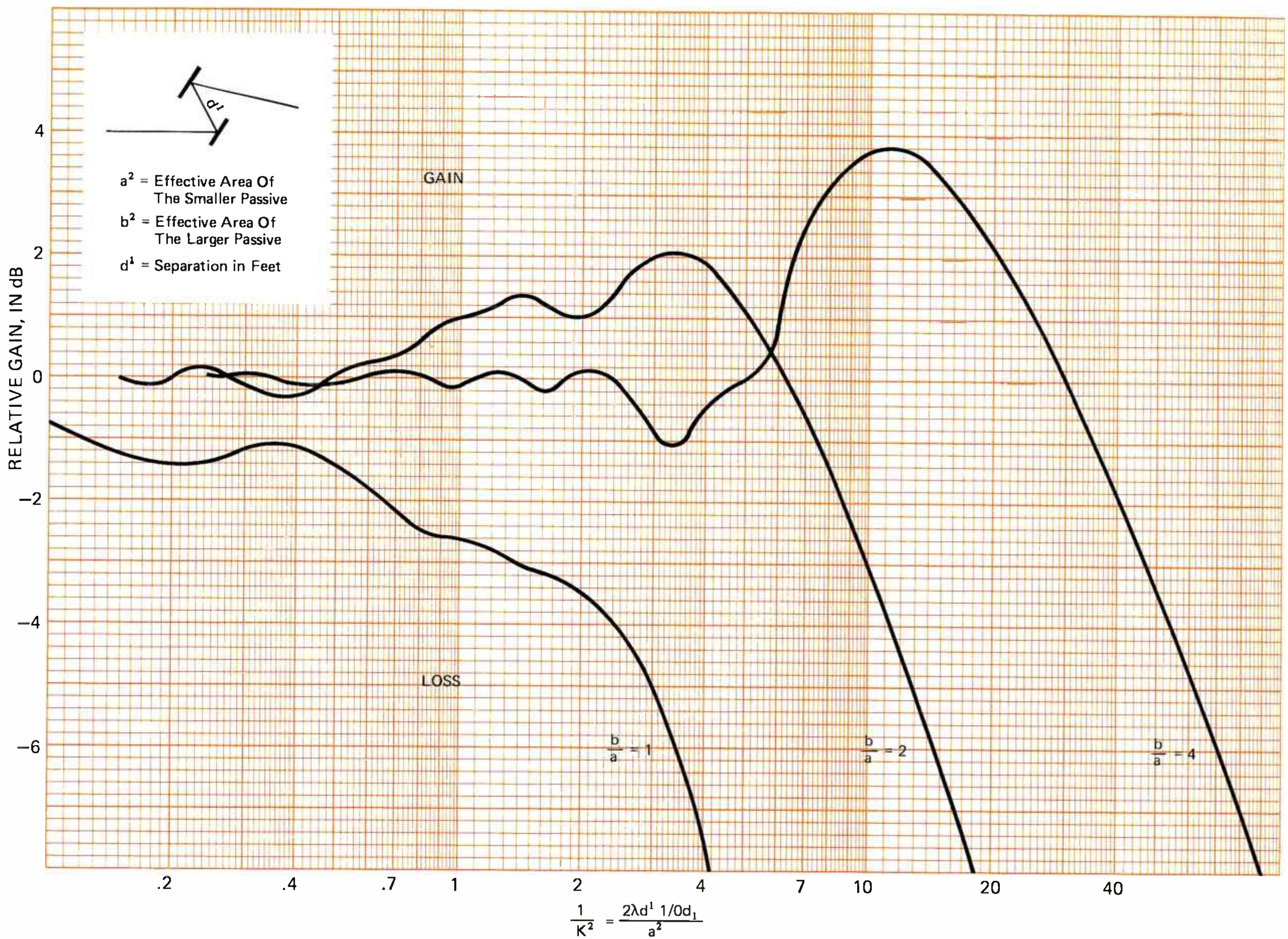


Figure 28D. Double Passive Repeater Efficiency Curves.

An alternative method of handling the near-field situation is provided by the curves of Figure 28C, which are an adaptation from the curves of Figure 28B. Figure 28C derives a gain reduction factor which, when applied to the 2-way free-space passive gain, gives the equivalent net gain when the problem is treated on a two-path basis. The big advantage of this approach is that it allows the passive repeater to be shown in the same way (as a separate location) and treated in the same way on the path calculation sheets, whether it is in the far field or the near field.

Closely spaced double passive repeaters are often used in situations where the transfer angle is less than about 50 degrees. In such cases, the free-space gain of the combination of the two passives is equal to the 2-way free-space gain for the effective area of the *smaller* of the two billboards (in most, but not all cases, they are made the same size), minus a reduction in gain which can be calculated from the "double passive repeater efficiency curves" of Figure 28D. If the double passive is close to one end of the path, near-field correction must also be applied.

Note that the parameters 1/K and L in Figures 28B and 28C are dimensionless. This means that if the antenna and reflector dimensions are in feet, the separation distance must also be in feet. Any consistent set of units can be used.

The following example is included to illustrate the calculation procedures:

Assume: 6 GHz band. 20' x 30' passive, with an included angle of 102°. 10' dishes at each end of path. Short leg 0.5 miles. Long leg 25 miles.

$$1/K = \frac{\pi \lambda d'}{4\,a^2} = \frac{3.14 \times 0.984 \times 0.5 \times 5280}{4 \times 6 \times 20 \times 30 \times \cos 102^\circ/2} \approx 0.9$$

$$L = D'\sqrt{\frac{\pi}{4\,a^2}} = 10\sqrt{\frac{3.14}{4 \times 20 \times 30 \times \cos 102^\circ/2}} \approx 0.48$$

From equation (3), short leg attenuation is 106.1 dB;
long leg attenuation is 140.1 dB.
From equation (27), free-space gain of passive is 105.0 dB.
Assume free-space gain of 10' dishes is 43.1 dB each.
Waveguide and fixed losses omitted for simplicity.
From equation (28), or by inspection of 1/K and Figure 28C it is apparent that the passive is in the near field of the close-in antenna.

Calculation using Figure 28B

We find for the calculated values of 1/K and L, a correction factor of -1.8 dB. Using this in a "periscope" type formula, we find

Net Path Loss = 140.1 − 43.1 − (43.1 − 1.8) = 55.7 dB

The quantity in brackets is the net gain of the periscope combination.

Calculation using Figure 28C

We find for the calculated values of 1/K and L a correction factor of −0.7 dB. Using this in a "two-path" formula, we find

Net Path Loss = 140.1 + 106.1 − 43.1 − 43.1 − (105.0 − 0.7) = 55.7 dB

As seen, either method should give the same result, within the accuracy with which the charts can be read.

Although passives are extremely useful in many cases, it should be remembered that they have their limitations. They are seldom practical in flat country, and even in hilly country, fairly favorable conditions are usually needed.

The effectiveness of a passive repeater is an inverse function of the *product* of the lengths of the two paths, rather than the sum of their lengths as one might suppose. Thus it is highly desirable to keep one of the paths very short. When high density systems with stringent noise performance objectives are involved, it is something of a rule of thumb that the product of the two path lengths in miles should not exceed about 30 (in the bands of

6-8 GHz) or about 50 (in the bands of 11-13 GHz) unless extremely large passives can be used. This is by no means an absolute restriction, and each case should be considered on its own merits, taking into account the system parameters and the required performance. With low density systems, or systems where somewhat greater noise can be tolerated, products considerably in excess of these figures can be acceptable.

When passives are used, it is often necessary (though not always) to settle for somewhat lower signal strengths than would be achieved if the same total path length could be spanned directly. In high density systems, where very strong signals may be required, this may limit or prevent the use of passive repeaters in some applications.

With very large passives, and particularly at 11 GHz, the beamwidths may become so narrow as to raise problems in maintaining sufficient rigidity under all conditions, and there may even be a possibility that unusual atmospheric conditions may bend the reflected beam far enough to cause a significant signal loss.

Nevertheless, the passive repeater, where the topography permits, and where it is suitable in other respects for the particular application, can be a very useful and efficient tool. Its big advantage is that it requires no maintenance, and thus can be used in spots which are too inaccessible or too expensive for an active repeater.

There appears to be no evidence of any tendency toward increased multipath fading because of a passive in a path. In fact, a path broken into two segments by a passive repeater might experience less multipath fading than a direct path of the same total length.

Although passive repeaters in themselves do not appear to contribute significant intermodulation noise, there are situations where sufficient signal may be reflected from hillsides, terrain, or even trees in the vicinity of the passive to create a delayed "echo" path with respect to the main signal. In some cases the result may be an intolerably high level of intermodulation noise. Thus, in any passive application the possibility of such reflections should be evaluated. Small passives with small included angles are particularly vulnerable to this problem.

A slightly different but related problem can occur with passive repeaters which are essentially "in line" with the path (either double billboards or back-to-back parabola types). The potential echo path here is by diffraction of the direct signal, over or around the obstruction.

H. Wire Line Entrance Links

In some circumstances it is possible to achieve substantial economies or to solve otherwise difficult problems by an arrangement in which the multiplex or terminal location and the actual radio site are physically separated by some distance and are interconnected by wire line entrance links of some kind.

Such links are operated on a four-wire basis, with separate pairs or separate cables for the two directions of transmission.

Entrance links operating at baseband usually utilize twin-conductor cables of the type used for video transmission. The range of distances may be from a few hundred feet up to a mile or more, depending on a number of factors. Except for very short links, or small channel densities, equalization and amplification are required for level and slope coordination.

With heterodyne systems it is also possible to separate the modems and the RF equipments, and operate wire line entrance links at IF frequencies, over coaxial cables. Here, too, amplification and equalization may be required, and the distance which can be spanned will be relatively limited unless intermediate repeaters are used.

Because of the many considerations involved and the rather special nature of these applications, a full treatment of wire line entrance links is beyond the scope of this book. However the microwave engineer doing path transmission work needs to be aware of these possibilities since they may affect his choice of sites or methods.

VI. CALCULATIONS FOR A MICROWAVE SYSTEM

A. Path Data Sheets

The path data sheet provides a formalized way of determining and recording all the parameters affecting the overall transmission loss equation. It is a useful tool for preliminary work, as well as for recording this and other pertinent data for future reference.

A separate data sheet can be completed for each path, or the data for a number of paths can be combined into a single running sheet. However, in the latter case, the data and calculations are for the individual paths and *not* for the overall system.

Figure 29 is an example of a completed path data sheet. The following discussion of various items on the sheet is intended to illustrate some of the details of the system planning, as well as the calculation methods.

The heading indicates that this is a one hop system, operating in the 5925-6425 MHz frequency band, with a design capacity of 960 channels using Lenkurt 78A microwave and 46A multiplex.

The data in Items 2, 3 and 4 we assume were determined during the path survey, which also produced a path profile and other information, allowing the engineer to determine the tower heights shown in Item 5, based on the desired clearance criteria, which we can assume to have been the "heavy route" criteria described in Section IV.C, Part 7.

From the disparity in tower heights at the two ends, it appears either the path was non-symmetrical, or that the situation in Alpha was such that a high tower at that location was impractical.

Items 7 and 8 were determined by calculation from Items 2 and 3, and the path attenuation then calculated by Equation (3), or read from Figure 13, and entered as Item 9.

Items 10 through 15 record, separately for each end of the path, the collective dB losses in all items of fixed loss appearing between the equipment connection flange and the antenna connection flange; plus the fixed loss of the radome if one is used, and its loss is not already included as part of the antenna gain figure.

In the system design stage, the exact waveguide layout is usually not known, so it is necessary to make reasonable estimates as to the amount and types of guide to be used; to be corrected later to reflect the "as built" condition. The fixed losses, though they appear small in comparison to the path attenuation, are vitally important to the system loss and gain equation, and must receive very careful consideration. Their importance can be understood by recalling that an increase of 3 dB in the fixed losses is equivalent to cutting the transmitter power in half, or an increase of 6 dB in fixed losses is the equivalent of doubling the path length.

It is obvious that, in the system design state, the items from 10 on are not developed in the order in which they appear in the data sheet. In fact, there are strong interactions between many of the items, and the actual selection usually involves evaluating several different combinations to find the one most suitable for the particular circumstances.

Rather than discuss the remaining items in order, we will discuss them in a sequence in which the transmission engineer might have developed them, in a series of steps:

1. He ascertains that the fade margin he is required to provide, or desires to provide, is 40 dB to the 55 dBrnc0 point. He enters the value of 55 dBrnc0 in the parentheses of Item 30, to establish the practical threshold point. He also tentatively enters 40 dB in Item 31. This will be changed later to 39.9 dB as a result of the final choices and calculations.

2. He ascertains, from the manufacturer specifications, or from a curve such as Figure 23A or 23B, that the RF input required to give 55 dBrnc0 in the top (worst) channel is −74 dBm. He enters this value in Item 30.

3. By algebraically adding Item 31 to Item 30, he determines a tentative value of −34 dBm as the received signal needed to give a 40 dB fade margin. However, he also ascertains from the manufacturer specifications that the recommended median receive signal level for 960 channel operation is −33 dBm. Since this latter value is higher than the −34 dBm he calculated, he tentatively enters −33 dBm in Item 27. It will be changed later to −34.1 dBm as a result of the final choices.

MICROWAVE PATH DATA CALCULATION SHEET

CUSTOMER ALPHA TELEPHONE COMPANY

PROJECT NO. XXX FREQUENCY 6175 MHz

SYSTEM ALPHA TO BETA EQPT 78A

LOADING +14.8 dBm0 (960 CHANNELS OF 46A)

			ALPHA	BETA
1	SITE		ALPHA	BETA
2	LATITUDE		34° 19' 01"	33° 57' 01"
3	LONGITUDE		84° 53' 52"	84° 39' 57"
4	SITE ELEVATION	Ft.	125	240
5	TOWER HEIGHT	Ft.	80	250
6	TOWER TYPE		SS	GUYED
7	AZIMUTH FROM TRUE NORTH		152°11	332°14
8	PATH LENGTH	Miles	28.55	
		KMS	45.94	
9	PATH ATTENUATION		141.5	
10	RIGID WAVEGUIDE WR137	Ft.	100	25
11	FLEXIBLE WAVEGUIDE	Ft.	5	5
12	WAVEGUIDE LOSS	dB	2.5	1.0
13	CONNECTOR LOSS	dB	0.5	0.5
14	CIRCULATOR OR HYBRID LOSS	dB	–	–
15	RADOME LOSS, TYPE*	dB	0.5u	0.5u
16	TOTAL FIXED LOSSES	dB	3.5	2.0
17	TOTAL LOSSES	dB	147.0	
18	PARABOLA HEIGHT	Ft.	75	15
19	PARABOLA DIAMETER	Ft.	10	6
20	REFLECTOR HEIGHT	Ft.	–	245
21	REFLECTOR SIZE, TYPE**	Ft.	–	10x15C
22	PARABOLA-REFL. SEP.	Ft.	–	230
23	ANTENNA SYSTEM GAIN	dB	43.0	41.9
24	TOTAL GAINS	dB	84.9	
25	NET PATH LOSS	dB	62.1	
26	TRANSMITTER POWER	dBm	+28.0	
27	MED. RECEIVED POWER (±2 dB)	dBm	–34.1	
28	RECEIVER NOISE THRESHOLD	dBm		
29	THEORETICAL RF C/N RATIO	dB		
30	PRACTICAL THRESHOLD (55dBrnc0)	dBm	–74.0	
31	FADE MARGIN (TO PRAC. THRES.)	dB	39.9	
32	RELIABILITY, SPACING***	%		
33	PROFILE NUMBER		1	

*U – Unheated, H – Heated, F – Feed Heater
**F – Flat, C – Curved
***F – Frequency Diversity, S – Space Diversity, Q – Space & Freq. Diversity, N – Non-Diversity,
(Reliability figures are for Rayleigh Distributed Fading Only)

NOTES:

Engineer: R.F.W. Date: 6/1/69

Figure 29. Microwave Path Data Calculation Sheet

4. From the manufacturer specifications, he determines that the transmitter has a minimum output power of +28 dBm, and this is entered in Item 26.

5. By subtracting Item 27 from Item 26 (algebraically) he determines a maximum allowable value of 61 dB for the net path loss, and tentatively enters 61 dB in Item 25.

6. By subtracting the tentative 61 dB in Item 25 from the 141.5 dB in Item 9, he determines that the total antenna gains, minus the fixed losses, must be at least 80.5 dB, in order to produce the desired value of net path loss.

7. At this point, if he has not already done so, he will make a tentative selection of the type of antenna systems to be used. If other considerations are not controlling, the choice will probably be based on the best combination, considering gain-efficiency and economics. However, frequency congestion or other considerations might have precluded the use of a periscope system, or dictated the choice of some specific antenna arrangements.

   We assume in this case, that the choice is a direct radiating parabola at Alpha, mounted at the top of the tower, and a periscope antenna system at Beta.

8. Having chosen the antenna system, he must then make a reasonably close estimate of the amount of waveguide required, and all other applicable fixed loss items. In this case he chooses WR137 rigid waveguide, and enters the estimated lengths in Item 10. He also enters estimated lengths of flexible guide, for connecting to equipment and antenna, in Item 11. He calculates waveguide losses, using 2.0 dB per 100 feet for the rigid and 0.1 dB per foot (a typical value) for the flex, and enters the total waveguide losses in Item 12.

   "Connector loss", Item 13, is a catchall item for small losses associated with pressure windows, bends and flanges. The value of 0.5 dB per end shown here, is a safe, probably conservative, estimate for most waveguide runs.

   In this case there are no circulators or hybrids external to the equipment, so no entry is made in Item 14.

   Radome loss, Item 15, will of course depend on the type, as discussed elsewhere. The value of 0.5 dB for an unheated radome is typical in this band.

9. He adds up the fixed losses and enters them in Item 16; adds the 5.5 dB total fixed losses to the 141.5 dB path attenuation of Item 9, and then enters the result, 147.0 dB, in Item 17.

10. He now subtracts the tentative value of 61 dB of Item 25 from the 147.0 dB total losses of Item 17, and obtains a value of 86.0 dB as a tentative value of the required total gains; entering this value temporarily in Item 24.

11. He divides 86.0 by two to obtain a preliminary value of 43.0 dB as the required antenna gain at each end of the path. (It is usually, though not always, most economical to have the antenna gains divided about equally.)

12. He determines that a gain of 43.0 dB at 6175 MHz will require at least a 10' parabolic antenna. In this case he enters the gain figure as taken from Table I, 43.0 dB, as the gain of the Alpha antenna in Item 23, and enters 10' in the Alpha column for Item 19.

13. By subtracting this 43.0 dB from the tentative 86.0 dB of Item 24, he finds that 43.0 dB gain will be needed from the antenna system at the other end. From Figure 27, using the −0.7 dB gain factor and the 1.09 dB distance factor for the 6.175 band, he determines that a 12' x 17' reflector would be needed to meet the 43.0 dB true gain requirement.

    At an apparent (chart) distance of 230 x 1.09 = 251', a 6' dish and either a C.R. or a C.E. reflector would give an apparent (chart) gain of about 44.5 dB, or a true gain at 6.175 GHz of 44.5 − 0.7 = 43.8, somewhat better than the objective. If the requirements were deemed absolute, this would be his probable choice. However, in this case we assume that the engineer for some reason does not want to use the very large and heavy 12' x 17' reflector. Instead he examines the next lower size, a 10' x 15', and determines that at the apparent distance of 251' a 6' dish and a 10' x 15' C.E. reflector will give an apparent gain of about 42.6 dB, or a true gain of 41.9 dB at 6.175 GHz. After entering this value in Item 23 and carrying out the necessary calculations, he finds that the median received signal will be —34.1 dBm, some 1.1 dB lower than the recommended level of —33.0 dBm, and the fade margin will be 39.9 dB instead of the desired 40 dB. The 0.1 dB difference in fade margin is quite insignificant (unless he is faced with a fixed and immutable requirement) and, since noise performance is also found to be satisfactory, the choice of a 6' dish and a 10' x 15' reflector seems to be a good one.

14. Having run through his preliminary calculations and made his final choices, the engineer now enters a 6' parabola under Beta in Item 19 and a 10' x 15' curved reflector under Item 21. (Item 18, the parabola height, Item 20, the reflector height and Item 22, the parabola-reflector separation, will have been determined and entered sometime prior to Step 13.) He enters 41.9 dB under Beta in Item 23, changes the tentative 86.0 dB in Item 24 to the final value of 84.9, subtracts this from Item 17 to obtain the final value of 62.1 dB for Item 25; subtracts 62.1 dB from the +28.0 dBm of Item 26 to obtain the final median received signal level of —34.1 dBm, then subtracts from this the —74.0 dBm0 of Item 30, to obtain the final fade margin of 39.9 dB for Item 31.

The process has been described in considerable detail to illustrate the methods used, and some of the choices.

Item 28, the "receiver noise threshold", and Item 29, the "theoretical RF C/N ratio", have deliberately been left blank in this example, since they play no part in the choices or calculations. They are on the sheet mainly for historical reasons, and because user specifications occasionally call for them. Item 28 is the absolute noise threshold as given by Equation (20). It is 10 dB lower than the "FM Improvement Threshold" given by Equation (21). Note that, in this example, the FM threshold is of no importance in the calculations, since the practical threshold determined from noise considerations is at a considerably higher level. This can be seen from Figure 23, which shows that the FM threshold would fall at about —79 dBm, assuming a 32 MHz IF bandwidth, or at —78 dBm with a 40 MHz bandwidth.

Item 32, "reliability", has also been left blank in the example for a different reason. This item will serve as a basis for discussion in the next section.

A microwave hop which includes a passive repeater, requires a somewhat more complicated approach in the path data sheet. For Items 1 through 9, and the pertinent items from 17 through 23, it is treated as a two-path system, but for Items 24 and after it is treated the same as a one-path system.

Space diversity hops also are more complicated than the example, because they have two separate antenna and waveguide systems at each end of the path, with different characteristics and, in some cases, different gains. Details of both antennas and guides are shown, and the subsequent calculations are made using the one with the lower gain for conservatism.

B. System Reliability Estimates

In estimating the overall reliability of a microwave system, one must first estimate the reliability of each hop. Consequently, we can discuss the reliability of our one-hop example system of Figure 29, and then show how the results could be extended to multihop systems.

1. Reliability with Respect to Multipath Fading

From Figure 29 we know that this is a path 28.55 miles long, with a fade margin of 39.9 dB, and that it operates in the 6 GHz common carrier band. To calculate the propagation reliability on a non-diversity basis we need to assign values to Barnett's a and b factors. The path and the elevations are fictitious, but the coordinates which were used would put it in northwest Georgia. This fits the "inland, temperate" category, and if we also assume that the path is of average terrain, with some roughness, we can use the "average-average" value of 1 for a and 1/4 for b.

Using (12), we have:

$$U_{ndp} = 1 \times 1/4 \times 1.25 \times 10^{-6} \times 15.4 \times 29^{3}$$

$$\times 10^{-4} = .0000118, \text{ by slide rule calculation.}$$

(Note that to simplify the calculations we have rounded off the path length to 29 miles and the fade margin to 40 dB. In view of the approximations and uncertainties in the theory and the fading models themselves, no greater precision is warranted. We should also round off the results to 2 significant figures, to be consistent).

Thus, the calculated propagation outage, on a non diversity basis is approximately 99.9988%, for this single hop system. If the system were never expected to grow beyond one hop, this figure might be quite acceptable. However, in view of the high channel density and the strong likelihood that many of the channels will be used for data, diversity would almost certainly be considered mandatory. Furthermore, future growth would most likely be such as to indicate a need for more than one working channel, so the initial system would probably be engineered as a 1-for-N system, with only one working channel and the protection channel equipped initially.

With a 1 for 1 system, the path would operate as a straight frequency diversity system insofar as propagation is concerned. Consequently we can use (14B) to calculate the "frequency diversity improvement factor". Assume that the diversity spacing is 2%. We then have:

$$I_{fd(6)} = 1/4 \times (.02) \times 10^4 = 50$$

and, using (13)

$$U_{fdp} = \frac{.000012}{50} = .00000024$$

so that the reliability is 99.999976%, an acceptable value for even the highest reliability systems. It would amount to about 8 seconds of outage per year.

If this one hop were part of a multihop system, we would make similar estimates of the per hop reliability percentages for each hop in the system. In order to obtain the system figure, we would first find the outage or "unreliability" percentage for each hop by subtracting its reliability percentage from 100. Since the individual hop outages are each extremely small, it is unlikely that such outages will occur in any two hops simultaneously, so the equivalent outage or "unreliability" of the entire system can be determined by adding up the outages of the individual hops. The system outage estimate is then subtracted from 100 to obtain the equivalent system propagation reliability.

As an example, consider a 10 hop system with each hop having an estimated propagation reliability of 99.9999% as above. The outage percentage for 10 hops would be 10 x (0.0001), or 0.001%, and the system estimated reliability is 99.999%. In short, the outage time would be directly proportional to the number of hops.

The above analysis applies only to systems in which diversity is applied to each path individually. In 1-for-N systems or 2-for-N systems, where 1 protection channel provides both propagation and equipment protection for several working channels, usually over a multihop switching section, reliability calculations become much more complex, and a much more critical evaluation of the correlation of fading among the various paths would be needed. A paper by Abraham, in the IEEE Transactions on Communication Technology, December, 1966, provides a thorough analysis of reliability in high-density microwave systems. By contrast, in the simple per-path diversity systems, the diversity advantage is so great that there is little need for precise calculations.

2. Reliability with Respect to Non-Selective Fading

There is no realistic way of calculating the magnitude of this kind of fading. It is therefore assumed that, when large fade margins and conservative clearance criteria are used as they were in this example, non-selective fading to threshold will either be non-existent, or that such effects as it has will be lumped in with the estimated multipath fading. There is one notable exception, a rare type of non-selective fading known as "blackout" fading, which occasionally occurs in some areas. It can wipe out all microwave frequencies in the area for a period ranging from a few minutes to an hour or more. It is not amenable to frequency diversity, nor to ordinary space diversity, though space diversity with extremely large spacing (hundreds of feet) might counteract it. There is no known practical or economical way to combat it, so in areas where it does occur, it is normally simply treated as a "catastrophic outage" rather than a propagation outage.

We call attention again to previous cautionary statements about any over-reliance on the absolute accuracy of estimates of propagation reliability. Such estimates are, however, extremely useful if their limitations are recognized.

3. Equipment Reliability Considerations

Microwave equipment of good modern design is inherently capable of providing quite high reliability, even on a non-redundant basis. But in order to meet the extraordinarily stringent objectives imposed by the need for extremely high-reliability long haul systems, some form of redundant or standby equipment is required.

Frequency diversity systems or space diversity systems with hot-standby transmitters each provide essentially 100% redundancy for the microwave equipment, and 1-for-N or 2-for-N multiline systems provide essentially 100% redundancy which is shared among the working channels.

Systems of this kind, with standby as well as working channels operating continuously and equipped with monitor, alarm, and automatic transfer facilities, provide the basic elements for ultra-high reliability. Equally important in achieving such reliability are the kind and quality of the maintenance given to the system.

These techniques for engineering and operating microwave systems to achieve extremely high reliabilities are well-known and well-developed, and they can be applied with confidence based on the demonstrated effectiveness of such techniques on existing systems.

But the techniques of making a priori predictions of the reliability performance of proposed new systems have been less well-developed and have played little part in the developments leading to highly reliable systems. Most of these developed methods have been empirical, as a result of accumulated experience.

Section IV-C-(11) provided means for calculating the probability of a propagation outage for a microwave path, as a function of a number of pertinent parameters. The calculations were developed in terms of a parameter "U", called "unavailability", also referred to as the probability of outage.

This unavailability parameter is simply the ratio, defined over any given period of time, of the outage, or "downtime" to the total, or "downtime" + "uptime". Thus;

$$U = \frac{\text{Down-time}}{\text{Total-Time}} = \frac{\text{Down-time}}{\text{Down-time + Uptime}}$$

The most natural period of interest in connection with microwave systems is the year, which is 8760 hours long, and the most widely used way of expressing propagation reliability is in terms of the unavailability or outage ratio, considered over a period of a year. Since fading phenomena in microwave systems generally follow a yearly cycle, the statistics for one year can be expected to be generally representative of those for any other year, and this parameter is thus a good measure of long term reliability. (In some cases it is also useful or desirable to look at U on a short term basis, for example, the worst month or the worst hour of the year). Another easy-to-understand way of specifying reliability is by the number of hours of downtime or outage per year. Thus we have:

$$U = \frac{\text{Downtime in hours}}{8760}$$

For example, an outage or downtime of .876 hours (about 53 minutes) per year would give

$$U = \frac{.876}{8760} = 10^{-4} = .0001$$

which corresponds to an annual "availability" A of .9999, or 99.99%; the term availability in reliability engineering circles has the same meaning as the term 'reliability' as commonly used by microwave engineers.

As discussed in an earlier section, per hop reliability objectives of 99.9999% are not uncommon for modern long-haul systems. This would require a value of U of .000001, corresponding to .00876 hours or about 32 seconds per year. Where these objectives include all sources of outage, it means that equipment outages also must be held to comparable levels.

(Note: It is worth mentioning that although it has only recently been recognized, CCIR objectives for allowable short term noise not to be exceeded for specified very small percentages of time, do not *include the effects of outages due to equipment failure, but only the performance during those periods of time when the equipment is in operating condition. The problem of total reliability or availability has only recently come under study by CCIR and is in a preliminary stage.)*

Reliability engineering techniques have been developed for calculating, a priori, the probability of successful operation of a component, unit, or system over any given time period, as a function of the number of components involved and their failure rates, the latter being assumed to be constant with time ("Early failures" and wear-out failures are excluded). Source data for failure rates is usually some document such as MIL HDBK 217-A, or some comparable industry data.

By starting with some basic source data, and making a number of assumptions, it is possible to calculate for each unit or item of equipment comprising a microwave system a parameter "mean time between failures", abbreviated as MTBF, usually expressed in hours. To the degree that the calculated MTBF is valid, it represents the average period of time the unit will operate without failure, considered over an indefinitely long period of time. (Actual field data on measured MTBF's are of course much preferable, but are difficult to obtain with high reliability equipments because failure rates are so low).

If the MTBF of the total equipment comprising a microwave hop is known, and if another parameter called "mean time to repair", abbreviated MTTR, is known or can be assumed, the equipment unavailability or outage ratio for the hop is given by:

$$U = \frac{\text{MTTR}}{\text{MTTR} + \text{MTBF}} \tag{29}$$

In high reliability systems MTBF >> MTTR, and the following approximation, which greatly simplifies the mathematics, is quite accurate:

$$U = \frac{\text{MTTR}}{\text{MTBF}} \tag{30}$$

To illustrate, suppose that the MTBF of the hop equipment is 10,000 hours, that is, on the average a little less than one failure per year, and that the MTTR, that is, the average period of outage after such a failure, is 1 hour. We then have:

$$U = \frac{1}{10{,}000} = .0001$$

corresponding to an "availability" of 99.99%. Mathematically we could bring this up to the 99.9999% level by reducing the MTTR to .01 hours, that is, about 36 seconds, but in practice this is an impossibility. In fact, when one considers that microwave systems consist in general of a number of unattended stations spaced a considerable distance apart, and that failures can occur at any hour of day or night, week-ends and holidays included, it seems highly optimistic even to assume a value as low as one hour for the "mean time to repair", including travel time, time to diagnose the trouble, make repairs or change out units, and get the system back in service. Even two hours will be an overly optimistic assumption in areas where repeater sites are remote or difficult of access.

If one must assume a relatively long average repair time, such as an hour or more, the only way to achieve an extremely low value of U is to greatly increase the denominator of the equations, the MTBF.

The relationship is very simple. In order to get an availability of 99.9999% we must have a U of .000001, or 10^{-6} and the MTBF must be 1,000,000 times the MTTR. If the repair time is 1 hour, the MTBF must be 1,000,000 hours, and if the repair time is 2 hours, the MTBF must be 2,000,000 hours.

These are truly staggering figures. Converted to years, the MTBF of 2,000,000 hours means that *on the average* there will be one failure on the hop every 228 *years*. Surprisingly enough, it is possible, if certain requisites are met, to show calculated per hop MTBF's of this order of magnitude, by the use of redundancy in the form of 100% operational standby. The requisites are: high (but not astronomical) basic reliability in the equipment comprising each side of the redundant pair; monitor, alarm and switching facilities which detect and report any failures and automatically switch to the good side if the other side has failed; a sparing and maintenance program which results in relatively quick repairs of all "minor" or one-side failures, in order to keep the time during which the path is operating non-redundantly as low as possible. (The "redundancy improvement factor", somewhat similar to the diversity improvement factors, is essentially the ratio of total time over a given period to the non-redundant time during that period.)

The following equation gives the calculated MTBF of a fully redundant block (for example, a one-way frequency diversity hop) in terms of the MTBF of the total equipment comprising one complete side of the redundant block, and the mean time to repair and restore a one-side failure:

$$m = \frac{M^2}{2\,T_1} \qquad (31)$$

where
- m is the MTBF of the redundant block
- M is the MTBF of one complete side
- T_1 is the mean time to repair and restore a one-side failure.

As an example, consider a hop with redundant equipment having a one-side MTBF of M = 10,000 hours, or 10^4 hours, and assume a value of 10 hours for T_1. We then have:

$$m = \frac{10^8}{20} = 5{,}000{,}000 \text{ hours}$$

as the MTBF of the redundant block.

The "redundancy improvement" is given by:

$$I_{red} = \frac{M}{2\,T_1} \qquad (32)$$

which in this case would be 10,000/20 or 500 to 1.

(Caution: This analysis excludes any equipment outside the fully redundant portion, and also assumes that switching is perfect and that there is no degree of commonality in the joining and splitting devices. In practical situations these other effects usually come into play and may reduce the true overall MTBF by an order of magnitude or more, below that calculated by the above equations. The analysis also is based on an assumption that failures on the two sides are random and are completely independent and uncorrelated, an assumption which may not always be correct. For these and other similar reasons reliability calculations of this type should be viewed with considerable caution).

Equipment Availability Calculations

The calculations can be formally extended to provide a means of calculating the availability for a redundant block (in terms of its time complement, the unavailability or outage ratio U), as follows:

$$U_r = \frac{T_2}{m} = \frac{2\,T_1 T_2}{M^2} \qquad (33)$$

where U_r is the outage ratio for the redundant block
m, M and T_1 are as in (31) and
T_2 is the mean time to repair and restore the *system* after a both-side failure (an actual system outage).

T_2 is simply the mean time to get one or the other side repaired and back in service, and thus is a somewhat complicated parameter.

A calculation approach which is simple and convenient mathematically is to assume each of the two one-side failures which cause the system outage is repaired independently, as if the other did not exist, and that the repair time for each is T_1. Under this assumption, it can be shown that:

$$T_2 = \frac{T_1}{2} \qquad (34)$$

giving:

$$U_r = \frac{2T_1}{M^2} \times \frac{T_1}{2} = \left(\frac{T_1}{M}\right)^2 \qquad (33A)$$

Although the above assumption is a convenient one, it is unrealistic when applied to real systems, where a true system outage generates much more pressure to make a quick repair and get the system back in operation.

If, instead of a "robot repairman" assumption as above, we consider a real life situation where the maintenance man will immediately be aware when a "both side" failure has occurred and will then assess the situation and act to repair the side which can be restored in the shortest time, it can be shown that T_2 can be derived from the following:

$$T_2 = T_3 - \frac{T_3^{\,2}}{2\,T_1} \qquad (35)$$

where T_2 and T_1 are as defined in (33) and T_3 is the "mean time to repair and restore one of the two one-side failures causing an outage", using all available resources and maximum effort to reduce travel time, but also assuming that the two failures occurred simultaneously or that at the time of the second failure no start had been made on repairing the first failure.

giving:

$$U_r = \frac{2\,T_1 \left(T_3 - \frac{T_3^{\,2}}{2\,T_1}\right)}{M^2} = \frac{2\,(T_1 T_3 - \frac{T_3^{\,2}}{2})}{M^2} \qquad (33B)$$

(Note that if the "robot repairman" assumption is made, T_3 becomes equal to T_1 and 33B reduces to 33A, as it should).

Carrying on with the previous example, assume M = 10,000 hours, T_1 = 10 hours, and further assume that T_3 = 3 hours.

The simple assumption of (33A) would give:

$$U_r = \left(\frac{10}{10^4}\right)^2 = 10^{-6} = .000001$$

or an "availability" of 99.9999%.

The "real life" assumption of (33B) would give:

$$U_r = \frac{2\,(10 \times 3 - \frac{3^2}{2})}{10^8} = \frac{51}{10^8} = .00000051$$

or an availability of 99.999949%.

(Note: in a one-way frequency diversity hop there is only one redundant block. But in a hot standby system, or a hot standby transmitter-space diversity receiver system, there are two separate and independent redundant blocks, one for transmitters and one for receivers. In this case the calculations would be made separately for each of the redundant blocks, and the resulting $U_{r(transmit)}$ and $U_{r(receive)}$ added together to get $U_{r(total)}$.

Availability as a Parameter

Although the derivations of "availability" calculation methods (through the complementary parameter unavailability or U) are formally and mathematically correct, there are some very serious drawbacks when one attempts to use this parameter as a measure of the reliability performance, with respect to equipment, of a highly reliable redundant system.

One basic requirement which must be met if the U parameter is to be useful is that the period of interest must be sufficiently long as to provide a reasonably good statistical sample of events (each individual outage constituting an event).

In considering propagation outages, over a period of a year there might be perhaps two-thousand individual one-side events each lasting perhaps 3 or 4 seconds on the average (one side fade below threshold), and with appropriate diversity perhaps 10 individual both-side events each lasting perhaps 2 seconds (both sides simultaneously faded below threshold), giving an "annual outage" of 20 seconds. Thus the period of a year is long enough to get a reasonable statistical sample, and calculated annual outages in this range could exist on a real system so that the calculated values could be tested and measured in the field. So the calculated U or the annual outage is a good parameter in this case.

But in the case of the equipment example discussed above, an entirely different situation exists. The *number* of events is very small, and the *length* of each event is very large. If the MTBF, as calculated in the example, is 5,000,000 hours, there will be *on the average*, over a sufficiently long time, about 1 failure every 570 years. Even if the hop were to be operated for a thousand years, we would have only a modest statistical sample of about 2 "events". For periods shorter than a millenium, U would have little meaning, and for periods of a year or even of the entire useful life of a microwave hop (15 to 20 years) it would have essentially no meaning at all.

Looking at it another way, if the mean time between failures is 570 years, we will have — on the average — one 570th of a failure every year. Mathematically if it takes 5 hours to repair one failure, it will take $\frac{5}{570}$ hours to repair one 570th of a failure. Five 570ths of an hour amount to about 32 seconds, confirming the previously calculated value of annual outage corresponding to this example.

Though mathematically acceptable, fractional failures cannot exist in real life, and as a result neither availability, unavailability, or annual outage is a suitable measure for equipment reliabilities in this range. These quantities are useful and meaningful with respect to propagation reliability, but — as shown above — have little or no value with respect to equipment reliability. If one observed such an equipment over the course of a year, there would be a very high probability of no outages at all. But in any year in which an outage *did* occur, the unavailability would be extremely high and in fact would use up all the outages allowable for a 570 year period. There is thus no way at all of checking such calculations in the field.

What Alternatives?

The value of m as calculated from (31), preferably divided by 8760 to give the "mean time between failures, in years" is a good measure of reliability for a redundant block.

Another possibility is to use the following equation to calculate the probability that the redundant block will operate without failure over a period of time t.

$$R(t) = e^{-t/m} \qquad (36)$$

For a period of a year, this is:

$$R(8760) = e^{-8760/m} \qquad (37)$$

When m is very much greater than t, (36) is accurately given by:

$$R(t) = 1 - t/m \qquad (36A)$$

and (37) by

$$R(8760) = 1 - 8760/m \qquad (37A)$$

For our previous example, with an m of 5,000,000 hours, (37A) gives a probability of .99825, or 99.825%, that the hop will operate without failure for a period of a year, under the assumed conditions.

This is perhaps the best way of expressing equipment reliability in such circumstances. Although neither the R parameter nor the m parameter from which it is derived can be tested against field experience (because even the full expected life of the equipment is too short a time frame for it), the two quantities M and T_1 in the expression (31) which gives the value of m *are* testable. Data taken over one hop for several years or over a number of hops for one year on the number of one-side failures and the average time taken to repair them can be directly checked against the calculated estimates, and can be used in (31) and (37) instead of the a priori calculated values. In this way an indirect check on the m and R parameters can be made.

It is clear also that propagation reliability and equipment reliability are completely different in character and should be treated separately rather than lumped together. Propagation reliability is well and usefully described either by an availability parameter or by an annual outage parameter, but equipment reliability is not.

The complete reliability parameter describing such a hop would be in two parts, a "U" or availability parameter for propagation and an "R" or probability of no outage for a year as the equipment parameter.

4. Power Reliability Considerations

The reliability of the power source supplying the equipment, is an extremely important factor in overall reliability. Systems operating directly from a primary AC power source will be subject to an outage whenever the power fails, even for a few seconds. Even in areas with highly reliable primary power, this is ordinarily not adequate to meet normal system reliability standards. Consequently, almost all microwave systems requiring high reliability are arranged with some provision for locally generated standby power, on a "no break" basis. The most satisfactory, and the most commonly used method, is to operate the microwave equipment from a storage battery plant, which is continuously charged or "floated" to maintain it at full capacity. The battery itself will take care of all short outages, and a standby generating plant is provided to switch on in case of longer outages.

C. Noise Performance Calculations

1. Microwave Noise

Just as in the case of reliability, the starting point for system noise performance calculations is to calculate the noise for each individual hop, then use these results to calculate the system noise.

We will first calculate the noise performance for the hop of Figure 29 by the North American method, then by the CCIR method as used internationally.

North American Method

In the North American method, the noise is calculated for the condition of an unfaded signal and busy hour loading. From a curve such as Figure 23B (which is appropriate for the particular equipment used in this example), the total noise can be read directly as a function of the received signal strength. In this case it is seen to be at approximately 21.5 dBrnc0 for a received signal of —34.1 dBm. Or, one could read separately the intrinsic plus intermodulation noise as 20 dBrnc0 and the terminal noise at —34.1 dBm input as 15 dBrnc0, and combine the two noises on a power basis to obtain 21.2 dBrnc0, approximately the same result, using Table G.

CCIR Method

In the CCIR method, it is common practice to; (1) show the intermodulation and thermal noises separately, (2) make the thermal noise calculation with the signal faded by approximately 5 dB, to simulate the "any hour" requirement, and (3) show the noise values in pWp0, which can then be added directly to obtain the total noise. Again, we use the curves of Figure 23B. The intrinsic + intermod noise of 20 dBrnc0 or —70 dBm0p can be converted to pWp0 using Table E, and is found to be 100 pWp0.

The thermal noise is read for a receive level of —39.1 dBm instead of —34.1 dBm, and is also found to be approximately 20 dBrnc0 or 100 pWp0.

The total noise for the hop is then 200.0 pWp0.

We can again refer to Table E to find that this is 23.0 dBrnc0; about 1.5 dB higher noise than calculated for the same system using the North American method.

This serves to illustrate why systems engineered using the CCIR approach often must be provided with several dB higher median signal levels, in order to be able to meet noise requirements under an assumed condition of a 5 dB fade. The extra dB are often very expensive to implement.

Although curves such as Figure 23B are useful and simple to use, they are not essential to noise calculations. An equally accurate calculation can be made by calculating the thermal noise from the known system parameters, using Equation (22), and adding in the intrinsic + intermodulation noise, which must be obtained from the manufacturer's specifications.

In either the North American approach or the CCIR approach, the estimated microwave-contributed noise for a multihop system is obtained by adding up the individual hop noises on a power basis. This is particularly easy to do when noises are given in pWp0, since they can be added directly. When the individual noises are expressed in dBrnc0 or dBa0, it is necessary to combine them two by two, using Table G, and continue the process until all the powers have been included. If all hops have the same noise power, the noise power of an N-hop system will be 10 $\log_{10}N$ dB higher than the per hop noise power, or N times the per hop power in picowatts.

Power addition of noise powers on a multihop system is based on two conditions; (1) that the equipment design is such that the odd-order intermodulation products, some of which tend to add on a voltage rather than a power basis, are very much lower than the even-order intermodulation products, and (2), that in very long systems, provisions are made to break up the baseband pattern at intervals of about 10 hops (most commonly by random supergroup and group interconnections at the sectionalizing point in high-density telephone systems, or by similar interconnections or by filter sectionalizing in industrial systems). If these conditions are met, it is considered that the system noise will be essentially power additive. Note that if either of these two conditions is not met, the assumption of power addition of multi-hop noise is not necessarily correct. Depending on the relative conditions, the power in such cases can be expected to combine on a basis somewhere between power and voltage addition.

2. Echo Distortion Noise

For illustrative purposes, assume a value of 28 dB equipment return loss, and a value of 26 dB for the lumped waveguide-antenna return loss. From this data, and a knowledge that the waveguide at Alpha is 100' long (approximately 2 dB loss), and that at Beta is 25' long (approximately 0.5 dB loss), we can use Figure 24 to obtain an estimated value of echo distortion noise.

Alpha:	Noise = 70.5 — 28 — 26 —4
	= 12.5 dBrnc0 = 18.0 pWp0
Beta:	Noise = 58.5 — 28 — 26 — 1
	= 3.5 dBrnc0 = 2.2 pWp0
Total:	Noise
	= 13.0 dBrnc0 = 20.2 pWp0

The 13.0 dBrnc0 is the estimated echo noise in the top channel of a system without emphasis. If the emphasis is used, as it normally would be, the echo noise would be reduced by some 3 dB, to about 10.0 dBnrc0.

Adding the echo noise to the 21 dBrnc0 radio equipment noise contribution, would increase the latter by slightly less than 0.5 dB. Though such an increase would be of little significance in a one-hop system, the accumulation of a number of such contributions in a long system *would* be significant.

Echo noise can be reduced, dB for dB, by improving either the equipment or the antenna system return losses, and of course, the noise will increase if the return losses decrease.

Echo distortion noise is calculated for each end of each path separately, and in multihop systems, all of these noises are added on a power additive basis, to the other system noises.

3. Multiplex Noise

This is normally simply taken from the manufacturer specifications, and added in with the other system noise contributors. For illustrative purposes, consider a loaded multiplex noise contribution of 23 dBrnc0, or 200 pWp0 per channel.

4. Total Noise Estimates

North American Method

Radio equipment noise, busy hour loading	= 21.0 dBrnc0
Echo distortion noise, busy hour loading	= 10.0 dBnrc0
Multiplex noise, busy hour loading	= 23.0 dBrnc0
Total noise for complete trunk	= 25.3 dBrnc0

This value is some 4.7 dB better than the objective given in Table H for 0 — 50 mile intertoll trunks.

CCIR Method

Equipment noise, busy hour loading (5 dB faded)	= 200.0 pWp0
Echo distortion noise	= 10.0 pWp0
Total microwave noise	= 210.0 pWp0

CCIR allowance for the microwave noise (does not include multiplex) would be approximately 3 x 50 + 200 = 350 pWp0. (For paths less than 50 kilometers, as this one just is, the noise allowance is treated as if the path were 50 kilometers.)

The calculated noise is approximately 2 dB better than the objective.

In specifying, listing and calculating noises from a great many sources, which are assumed to be additive on a power basis, the picowatt is a very convenient noise unit, since the total noise is simply the arithmetic sum of the individual noises.

On the other hand, the logarithmic noise units are much more meaningful as to the actual effects of noise changes, and a logarithmic form is the only convenient form to use in system measurements and maintenance. The North American approach is to use units in dB form for both purposes, simply tolerating the slight inconvenience in adding powers expressed that way. The CCIR method generally uses picowatts for specifications and calculations, but reverts to some logarithmic form, such as the dBm0p, for measurements and maintenance usage. There are advantages and disadvantages to both approaches.

AN IMPORTANT POST SCRIPT ABOUT THE PROPAGATION CALCULATIONS

As stated on Page 55, the methods of calculating propagation outages used in this book are based on previously reported work by W. T. Barnett and Arvids Vigants of Bell Telephone Laboratories.

At the June, 1970 International Communications Conference in San Francisco Mr. Barnett gave an updating report on his work, based on the continuing study and analysis which has gone on in the past year. Based on this further study, Barnett now proposes a modification in the frequency term in the equation describing non-diversity fading. The modification changes the frequency term in Equation (11B) from $(f/4)^{1.5}$ to $(f/4)$.

The effect of this modification is to make the calculated outage a linearly increasing rather than an exponentially increasing function of frequency. Outages calculated according to the original equations, as used in this book, can be multiplied by $(4/f)^{0.5}$ to convert them to outages in accordance with the modified formula. The value of this modifying factor is: 2 GHz 1.4; 4.0 GHz 1.0; 5.0 GHz 0.9; 6.2 GHz 0.8; 6.7 GHz 0.77; 7.8 GHz 0.72; 8.7 GHz 0.68; 9.5 GHz 0.65; 11.2 GHz 0.60; 12.4 GHz 0.57.

As can be seen, the maximum change resulting from the modification is limited to about 40% difference in the calculated outage, and for the bands of most interest –those from 4 GHz up– the outages calculated by the original methods are more conservative (greater indicated outage) than the modified version. Consequently we have not considered it necessary to modify the methods proposed in the book.

Those who desire to use the modified form can do so by applying the modifying factor discussed above, or can make the following changes in the book itself:

In Equation (11B) which appears on Page 59, change the frequency term from $(f/4)^{1.5}$ to $(f/4)$.

In Equation (12) which appears on Page 60 and again in Appendix I on Page A2, change the frequency term from $f^{1.5}$ to f, and change the constant factor from 1.25 to 2.5.

In the metric version of Equation (12), which appears in Appendix II on Page B2, change the frequency term from $f^{1.5}$ to f, and change the constant factor from 3.0 to 6.0.

It is emphasized that the experimental and theoretical studies are continuing, and other modifications may very well be reported at some future date.

NOTE:
It is important to note that graphs, formulas and methods for calculating outages throughout the book are all for one way outage. To calculate two way outages it is necessary to double the calculated multipath and equipment outages. Outages due to rain or to non-selective fading do not have to be doubled since they occur simultaneously in both directions of transmission.

APPENDIX I
USEFUL FORMULAS AND EQUATIONS

		Text Reference Page Number
Earth Curvature		
$h = \dfrac{d_1 \times d_2}{1.5K}$	(1)	12
$h\ (K = \infty) = 0$	(1A)	12
$h\ (K = 4/3) = \dfrac{d_1 d_2}{2}$	(1B)	12
$h\ (K = 2/3) = d_1 d_2$	(1C)	12
$h\ (K = 1) = .67\, d_1 d_2$	(1D)	12
h in feet, d's in miles		
Reflection Point Relations		
For $K = 4/3$; $\dfrac{h_1}{d_1} - \dfrac{d_1}{2} = \dfrac{h_2}{d_2} - \dfrac{d_2}{2}$	(2C)	17
For $K = 2/3$; $\dfrac{h_1}{d_1} - d_1 = \dfrac{h_2}{d_2} - d_2$	(2B)	17
For $K = \infty$; $d_1 = nD$ where $n = \dfrac{h_1}{h_1 + h_2}$	(2A)	17
h in feet, d and D in miles		
Path Attenuation		
$A_{dB} = 96.6 + 20 \log_{10} F_{GHz} + 20 \log_{10} D_{miles}$	(3)	35
Fresnel Zones		
1st zone; $F_1 = 72.1 \sqrt{\dfrac{d_1 d_2}{F_{GHz} D}}$	(4A)	38
nth zone; $F_n = F_1 \sqrt{n} = 72.1 \sqrt{\dfrac{n\, d_1\, d_2}{F_{GHz} D}}$	(4C)	38
F_1 in feet; d and D in miles		

		Text Reference Page Number
Fictitious Earth Radius		
$$K = \frac{157}{157 + \frac{dN}{dh}}$$	(9B)	44
$\frac{dN}{dh}$ is gradient in N units per kilometer		
Recommended Clearance Criteria		
Heavy-route, or highest reliability systems:		51
At least $0.3F_1$ at $K = 2/3$ and at least $1.0F_1$ at $K = 4/3$, whichever requires the greater heights. (In areas of very difficult propagation, it may be necessary also to ensure a clearance of at least grazing at $K = 1/2$).		
All criteria should be evaluated along entire path.		
Light-route, or medium reliability systems:		51
At least $0.6F_1 + 10$ feet at $K = 1.0$		
Vertical Diversity Spacing For Reflective Paths		
$$\Delta h_2 = \frac{2.2 \times 10^3\ D}{F_{GHz} \times h_t}$$ or (see Text)	(10A)	59
$$\Delta h_2 = \frac{1.3 \times 10^3\ D}{F_{GHz} \times h_t}$$	(10B)	59
h's in feet, D in miles		
Fading Outages and Diversity Improvement Factors		
$$U_{ndp} = a \times b \times 1.25 \times 10^{-6} \times f^{1.5} \times D^3 \times 10^{-F/10}$$	(12)	60
$$U_{div} = \frac{U_{ndp}}{I}$$	(13)	60

		Text Reference Page Number

Fading Outages and Diversity Improvement Factors (Continued)

$I_{fd(4)} = 1/2 \times \left[\frac{\Delta f}{f} \right] \times 10^{F/10}$	(14A)	60
$I_{fd(6)} = 1/4 \times \left[\frac{\Delta f}{f} \right] \times 10^{F/10}$	(14B)	60
$I_{fd(7\text{-}8)} = 1/8 \times \left[\frac{\Delta f}{f} \right] \times 10^{F/10}$	(14C)	60
$I_{fd(11\text{-}12)} = 1/12 \times \left[\frac{\Delta f}{f} \right] \times 10^{F/10}$	(14D)	60
$I_{CBD} = 100$	(15)	61
$I_{SD} = \frac{7.0 \times 10^{-5} \times f \times s^2 \times 10^{\overline{F}/10}}{D}$	(16)	61
$I_{hyb} = I_{SD}$	(17)	61
$k^2 = \frac{I}{10^{F/10}}$	(18)	61

Noise Threshold

$N = -114 + 10 \log_{10} B\ (IF)\ MHz + F_{dB}$	(20)	68
FM Improvement Threshold		
$T_{FM} = -104 + 10 \log_{10} B\ (IF)\ MHz + F_{dB}$	(21)	68

Noise Units Correlation

$$dBrnc0 = 10 \log_{10} pWp0 = dBa0 + 6 = dBm0p + 90 = 88 - \frac{S}{N_{flat}}$$ 69

(Precise: $dBrnc0 = 10 \log_{10} pWp0 + 0.8 = dBa0 + 6.8 = dBm0p + 90.8 = 88.3 - \frac{S}{N_{flat}}$) 67

		Text Reference Page Number
Thermal Noise In Derived Channel		
$\text{dBrnc0} = -C - 48.1 + F_{dB} - 20\log_{10}\frac{\Delta f}{f_{ch}}$	(22A)	68
$\text{dBa0} = -C - 54.1 + F_{dB} - 20\log_{10}\frac{\Delta f}{f_{ch}}$	(22B)	68
$\frac{S}{N}\,\text{dB, flat} = C + 136.1 - F_{dB} + 20\log_{10}\frac{\Delta f}{f_{ch}}$	(22C)	68
$\text{pWp0} = \log_{10}{}^{-1}\left[\frac{-C - 48.6 + F - 20\log\frac{\Delta f}{f_{ch}}}{10}\right]$	(22D)	68
Video Noise		
$\frac{S_{ptp}}{N_{rms}}$ (unweighted, unemph) $= C - F_{dB} + 118$	(22E)	70
$\frac{S_{ptp}}{N_{rms}}$ (EIA, emph, EIA Color wtg) $= C - F_{dB} + 126.5$	(22F)	70
Noise Loading		
CCIR $P = (-15 + 10\log_{10}N)$ dBm0 when, N = 240 or more	(23A)	73
CCIR $P = (-1 + 4\log_{10}N)$ dBm0 when, N = 12 to 240	(23B)	73
Bell $P = (-16 + 10\log_{10}N)$ dBm0	(23C)	73
Military $P = (-10 + 10\log_{10}N)$ dBm0	(23D)	73
Parabola Gain		
$G_{dB} = 10\log_{10}\left(4\pi A\frac{E}{\lambda^2}\right)$	(24)	90
$G_{dB} = 20\log_{10}B_{ft} + 20\log_{10}F_{GHz} + 7.5$	(25)	90

		Text Reference Page Number
3-dB Beamwidth		
$\phi \approx \dfrac{70}{F_{GHz}\ B_{ft}}$ for parabola	(26)	91
$\phi \approx \dfrac{60}{F_{GHz}\ B_{ft}}$ for elliptical reflector		99
$\phi \approx \dfrac{52}{F_{GHz}\ W_{ft}}$ for rectangular reflector		
2-Way Free-Space Billboard Gain		
$G_{dB} = 22.2 + 40 \log_{10} F_{GHz} + 20 \log_{10} A_{sqft} + 20 \log_{10} \cos \alpha$	(27)	100
Approximate Near Zone Boundary		
$d_{ft} = 2\ F_{GHz}\ B^2{}_{ft}$	(28)	100

APPENDIX II
USEFUL FORMULAS AND EQUATIONS IN METRIC FORM

In this section we provide metric equivalents of those formulas and equations of Appendix I which involve units of length expressed in feet or miles. The symbols used are identical to those used in the English-unit versions except that those units which were in feet are now in meters, and those which were in miles are now in kilometers. In short, all h's are in meters and all d's in kilometers.

We also provide in this section suggested methods of adapting some of the English-unit charts for use with the equivalent metric units.

We have omitted from this section those equations which do not involve units of length, or which are identical in the two systems.

	Text Reference Page Number
Earth Curvature (Metric)	
$h = \frac{d_1 \, d_2}{12.75K}$	12
$h(K = \infty) = 0$	12
$h(K = 4/3) = \frac{d_1 \, d_2}{17}$	12
$h(K = 2/3) = \frac{d_1 \, d_2}{8.5}$	12
h in meters, d's in kilometers	
Reflection Point Relations (Metric)	
For K = 4/3: $\frac{h_1}{d_1} - \frac{d_1}{17} = \frac{h_2}{d_2} - \frac{d_2}{17}$	17
For K = 2/3: $\frac{h_1}{d_1} - \frac{d_1}{8.5} = \frac{h_2}{d_2} - \frac{d_2}{8.5}$	17
h's in meters, d's in kilometers	
Path Attenuation (Metric)	
$A_{dB} = 92.4 + 20 \log_{10} F_{GHz} + 20 \log_{10} D_{KMS}$	35

	Text Reference Page Number

Fresnel Zones (Metric)

1st zone:	$F_1 = 17.3\sqrt{\dfrac{d_1\ d_2}{F_{GHz}\ D}}$	38
nth zone:	$F_n = F_1\sqrt{n}$	38

F_1 in meters, d's and D in kilometers

Vertical Diversity Spacing for Reflective Path (Metric)

$\Delta h_2 = \dfrac{127 \times D}{F_{GHz} \times h_t}$, or	59
$\Delta h_2 = \dfrac{75 \times D}{F_{GHz} \times h_{t_1}}$	59
$h_{t_1} = h_1 - \dfrac{{d_1}^2}{17}$ (See text)	59

h's in meters, d and D in kilometers

Fading Outages and Diversity (Metric)

$U_{ndp} = a \times b \times 3.0 \times 10^{-7} \times f^{1.5} \times D^3 \times 10^{-F/10}$	60
$I_{sd} = \dfrac{1.2 \times 10^{-3} \times f \times s^2 \times 10^{\overline{F}/10}}{D}$	61

f in GHz, s in meters

D in KMS, $\overline{F}$ in dB

The other equations in this section are unchanged.

Parabola Gain (Metric)

$G_{dB} = 20 \log_{10} B + 20 \log_{10} F_{GHz} + 17.8$	90

B is in meters

	Text Reference Page Number

3-dB Beamwidth (Metric)

$\phi \approx \frac{22}{F_{GHz}\,B}$	for parabola	91
$\phi \approx \frac{19}{F_{GHz}\,B}$	for elliptical reflector	99
$\phi \approx \frac{16}{F_{GHz}\,W}$	for rectangular reflector	99

ϕ in degrees, B and W in meters

2-Way Free-Space Billboard Gain (Metric)

$$G_{dB} = 42.9 + 40 \log_{10} F_{GHz} + 20 \log_{10} A + 20 \log_{10} \cos \alpha$$ 100

A is in square meters

Approximate Near Zone Boundary (Metric)

$$d_{mtrs} = 6.6 F_{GHz}\, B^2$$ 100

B is in meters

Suggestions for use of Figures

Figure 3:
Earth Curvature for Various Values of K.

The generating equation is

$$h = \frac{d^2}{12.75\ K}$$

where h is in meters and d in kilometers.

If the distances on the d scale are considered to represent kilometers rather than miles, the value read on the "departure" scale must be multiplied by 0.118 to obtain h in meters.

Figure 4:
Though the curves of Figure 4 were designed for use with miles, feet, and 10-division-per-inch rectangular graph paper, with a basic scale of 1" horizontal = 2 miles, 1" vertical = 100', it turns out that by good fortune they can also be used with kilometers, meters, and 10-division-per-centimeter paper, with a basic scale of 1 cm horizontal = 1 kilometer, 1 cm vertical = 7.5 meters, and also with a scale of 1 cm horizontal = 2 kilometers, 1 cm vertical = 30 meters.

A very slight error is incurred, amounting to about one part in 400 in the corresponding elevation readings, but it is considerably lower than the accuracy with which the charts themselves can be read so is inconsequential.

Alternatives are use of the direct calculation methods described in the text, using the metric versions of (1B), (1C) or (1D), as required, or to use (1) or the metric generating equation for Figure 3 to construct metric templates with other scale values.

Figure 7:
The reflection point charts can be used directly using

$$X = \frac{8.5\ h_1}{D^2} \text{ and } Y = \frac{8.5\ h_2}{D^2}$$

where the h's are in meters and the D's in kilometers.

Figure 13:
The free-space attenuation chart can be used directly by considering D to be the path length in kilometers, and subtracting 4.2 dB from the attenuation reading.

Figure 15:
The 1st Fresnel Zone radius chart can be used by considering the D and the d's to be in kilometers, and dividing the value read on the F scale by 4.17 to obtain F in meters.

The multiplying factors of Tables B1 and B2 can then be applied to convert the readings for frequency bands other than 6.7 GHz, or for higher zones.

Figures 21 and 22:
Convert the distances shown in miles to kilometers by multiplying by 1.609. The charts can then be read directly.

Figure 24:
The best way to use this chart is to convert waveguide length in meters to feet by multiplying by 3.28, then enter the chart with the reading in feet.

Figure 25:
The dimensions shown in feet can be converted to meters by multiplying by 0.3048. Areas shown in acres can be converted to square meters by multiplying by 4047.

Figure 27:
Since these charts are for specific sizes of standard reflectors and antennas dimensioned in feet, they could not be directly used for metric-dimensioned reflectors and antennas.

Figure 28A:
This chart is also for specific sizes of reflector dimensioned in feet, and could not be used directly for metric-dimensioned reflectors.

Figures 28B, C, D:
These charts are dimensionless, and require only that all dimensions be in the same units.

Inverse Position Azimuth and Path Distance Calculations. The method produces distances in both kilometers and feet, so can be used in either set of units.

APPENDIX III
USEFUL TABLES AND FIGURES

Table C. Excess Attenuation Due To Atmospheric Absorption. P-51

PATH LENGTH MILES	ATTENUATION – dB				
	2-4-6 GHz	8 GHz	10 GHz	12 GHz	14 GHz
20	0.20	0.26	0.32	0.38	0.48
40	0.40	0.52	0.64	0.76	0.96
60	0.60	0.78	0.96	1.14	1.44
80	0.80	1.04	1.28	1.52	1.92
100	1.00	1.30	1.60	1.90	2.40

Table D. Relationship Between System Reliability And Outage Time P-56

RELIABILITY %	OUTAGE TIME %	OUTAGE TIME PER		
		YEAR	MONTH (Avg.)	DAY (Avg.)
0	100	8760 hours	720 hours	24 hours
50	50	4380 hours	360 hours	12 hours
80	20	1752 hours	144 hours	4.8 hours
90	10	876 hours	72 hours	2.4 hours
95	5	438 hours	36 hours	1.2 hours
98	2	175 hours	14 hours	29 minutes
99	1	88 hours	7 hours	14.4 minutes
99.9	0.1	8.8 hours	43 minutes	1.44 minutes
99.99	0.01	53 minutes	4.3 minutes	8.6 seconds
99.999	0.001	5.3 minutes	26 seconds	0.86 seconds
99.9999	0.0001	32 seconds	2.6 seconds	0.086 seconds

Table F. Standard CCIR 20 $\log_{10} \Delta f / f_{ch}$ Factors For Top Slot

P-71

SYSTEM CHANNELS	TOP SLOT	WITHOUT EMPHASIS	WITH EMPHASIS
120	534 kHz	–5.52 dB	–1.82 dB (120 channel emphasis)
300	1248 kHz	–12.9 dB	–9.2 dB (300 channel emphasis)
420*	1722 kHz	–15.7 dB	–12.0 dB (420 channel emphasis)
600	2438 kHz	–18.7 dB	–15.0 dB (600 channel emphasis)
960	3886 kHz	–22.8 dB	–19.1 dB (960 channel emphasis)
1200	5340 kHz	–28.5 dB	–24.8 dB (1200 channel emphasis)

NOTE: 200 kHz rms per channel deviation for all except 1200 channel system, which is 140 kHz rms per channel deviation.

*Not a CCIR Standard, but widely used in U.S.A. industrial systems.

Table H. Typical U.S.A. Noise Objectives

P-80

TRUNK LENGTH IN MILES	TYPE OF TRUNK		
	INTERTOLL* dBrnc0	TOLL CONNECTING* OR TANDEM dBrnc0	DIRECT* dBrnc0
0–50	30	32	35
51–100	31	33	36
101–200	33	35	38
201–400	35	37	40
401–1000	37	39	42
1001–1500	38	40	43
1501–2500	41	43	46
2501–4000	43	45	47

*SEE NOTE ON PAGE 80

Table I. Antenna Gains for Estimating Purposes

P-92

Plane Polarized Parabolic Antennas. (DP's, HP's and Cross-band somewhat lower)

Diameter in feet	Gain Relative To Isotropic – dB						
	2 GHz	4 GHz	6 GHz	7 GHz	8 GHz	11 GHz	13 GHz
4	25.5	– –	35.2	35.9	37.0	40.3	41.3
6	29.0	35.0	38.7	39.4	40.6	43.8	44.8
8	31.5	37.3	41.1	41.9	43.1	46.0	47.3
10	33.5	39.3	43.0	43.9	45.2	47.7	48.5
12	– –	40.8	44.6	45.5	46.7	– –	– –
15	– –	42.6	46.0	46.9	48.7	– –	– –
Horn Reflector Antennas							
8X8 (Std)	– –	39.4	43.0	– –	– –	47.4	– –
6 (Circ)	– –	35.7	39.4	– –	– –	43.8	– –

Table E. Noise Unit Comparison Chart.

dBrnc0	dBa0	pWp0	dBm0p	S/N_{dB}	dBrnc0	dBa0	pWp0	dBm0p	S/N_{dB}
0	–6	1.0	–90	88	34	28	2520	–56	54
1	–5	1.3	–89	87	35	29	3162	–55	53
2	–4	1.6	–88	86	36	30	3981	–54	52
3	–3	2.0	–87	85	37	31	5012	–53	51
4	–2	2.5	–86	84	38	32	6310	–52	50
5	–1	3.2	–85	83	39	33	7943	–51	49
6	0	4.0	–84	82	40	34	10,000	–50	48
7	1	5.0	–83	81	41	35	12,500	–49	47
8	2	6.3	–82	80	42	36	15,850	–48	46
9	3	7.9	–81	79	43	37	19,950	–47	45
10	4	10.0	–80	78	44	38	25,200	–46	44
11	5	12.6	–79	77	45	39	31,620	–45	43
12	6	15.8	–78	76	46	40	39,810	–44	42
13	7	20.0	–77	75	47	41	50,120	–43	41
14	8	25.2	–76	74	48	42	63,100	–42	40
15	9	31.6	–75	73	49	43	79,430	–41	39
16	10	39.8	–74	72	50	44	100,000	–40	38
17	11	50.1	–73	71	51	45	125,900	–39	37
18	12	63.1	–72	70	52	46	158,500	–38	36
19	13	79.4	–71	69	53	47	199,500	–37	35
20	14	100	–70	68	54	48	252,000	–36	34
21	15	126	–69	67	55	49	316,200	–35	33
22	16	158	–68	66	56	50	398,100	–34	32
23	17	200	–67	65	57	51	501,200	–33	31
24	18	252	–66	64	58	52	631,000	–32	30
25	19	316	–65	63	59	53	794,300	–31	29
26	20	398	–64	62	60	54	1,000,000	–30	28
27	21	501	–63	61	61	55	1,259,000	–29	27
28	22	631	–62	60	62	56	1,585,000	–28	26
29	23	794	–61	59	63	57	1,995,000	–27	25
30	24	1000	–60	58	64	58	2,520,000	–26	24
31	25	1259	–59	57	65	59	3,162,000	–25	23
32	26	1585	–58	56	66	60	3,981,000	–24	22
33	27	1995	–57	55					

Table E shows the relationship between five commonly used units for expressing noise in a voice band channel. In the first four columns, the units represent weighted noise at a point of zero relative level. In the fifth column the "S" represents a tone at zero relative level, and the "N" represents unweighted noise in a 3 kHz voice channel, therefore, S/N is the dB ratio of test tone to noise.

The table is based on the following commonly used correlation formulas, which include some slight round-offs for convenience. Correlations for Columns 2, 3 and 4 are valid for all types of noise. All other correlations are valid for white noise, but not necessarily for other types.

$$\text{dBrnc0} = 10 \log_{10} \text{pWp0} = \text{dBa0} + 6 = \text{dBm0p} + 90 = 88 - S/N$$

Table G. Summation Or Subtraction Of Non-Coherent Powers.

This table can be used for summing the powers of two non-coherent signals expressed in dB form. It can also be used for power subtraction.

P_a and P_b represent two powers whose summation is P_s: in all cases P_a is taken as the larger of the two powers.

To sum two powers, calculate $P_a - P_b$, locate the resulting value in Column 1, then add the corresponding value in Column 2 to P_a to obtain P_s, the desired sum.

To subtract one power from another, treat the larger one as P_s and the smaller one as P_a (if it is within 3 dB of P_s) or P_b (if it is more than 3 dB below P_s).

In the first case, calculate $P_s - P_a$, locate the resulting value in Column 2, then subtract the corresponding value in Column 3 from P_s to obtain P_b, the desired remainder.

In the second case, calculate $P_s - P_b$, locate the resulting value in Column 3, then subtract the corresponding value in Column 2 from P_s to obtain P_a, the desired remainder.

When more than two powers are to be summed or subtracted, iteration can be used.

Example: Summation

To add + 10.0 dBm0 to +8.7 dBm0
10.0 – 8.7 = 1.3
1.3 in Column 1 falls between 1.2 and 1.4 so the value from Column 2 is (2.45 + 2.37)/2 = 2.41.
So P_s = + 10.0 + 2.41 = + 12.41 dBm0

Example: Subtraction

To subtract –15.0 dBm from –10.0 dBm0
–10.0 – (–15.0) = 5, treating –10.0 as P_s and –15.0 as P_b we locate 5 in Column 3 as very near to 5.035, so we subtract the corresponding value in Column 2, 1.635, from –10.0 to obtain P_a.
So P_a = –10.0 – 1.635 = –11.6 (rounded).

Col. 1	Col. 2	Col. 3
$P_a - P_b$	$P_s - P_a$	$P_s - P_b$
0.0	3.010	3.010
0.2	2.911	3.111
0.4	2.815	3.215
0.6	2.721	3.321
0.8	2.629	3.429
1.0	2.539	3.539
1.2	2.451	3.651
1.4	2.366	3.766
1.6	2.284	3.884
1.8	2.203	4.003
2.0	2.124	4.124
2.2	2.048	4.248
2.4	1.974	4.374
2.6	1.902	4.502
2.8	1.832	4.632
3.0	1.764	4.764
3.2	1.698	4.898
3.4	1.635	5.035
3.6	1.573	5.173
3.8	1.513	5.313
4.0	1.455	5.455
4.2	1.399	5.599
4.4	1.345	5.745
4.6	1.293	5.893
4.8	1.242	6.042
5.0	1.193	6.193
5.2	1.146	6.346
5.4	1.100	6.500
5.6	1.056	6.656
5.8	1.014	6.814
6.0	0.973	6.973
6.5	0.877	7.377
7.0	0.790	7.790
7.5	0.710	8.210
8.0	0.639	8.639
8.5	0.574	9.074
9.0	0.515	9.515
9.5	0.461	9.961
10.0	0.414	10.414
11.0	0.331	11.331
12.0	0.266	12.266
13.0	0.216	13.216
14.0	0.170	14.170
15.0	0.135	15.135
16.0	0.108	16.108
17.0	0.086	17.086
18.0	0.068	18.068
19.0	0.054	19.054
20.0	0.043	20.043
25.0	0.016	25.016
30.0	0.004	30.004
∞	0.000	∞

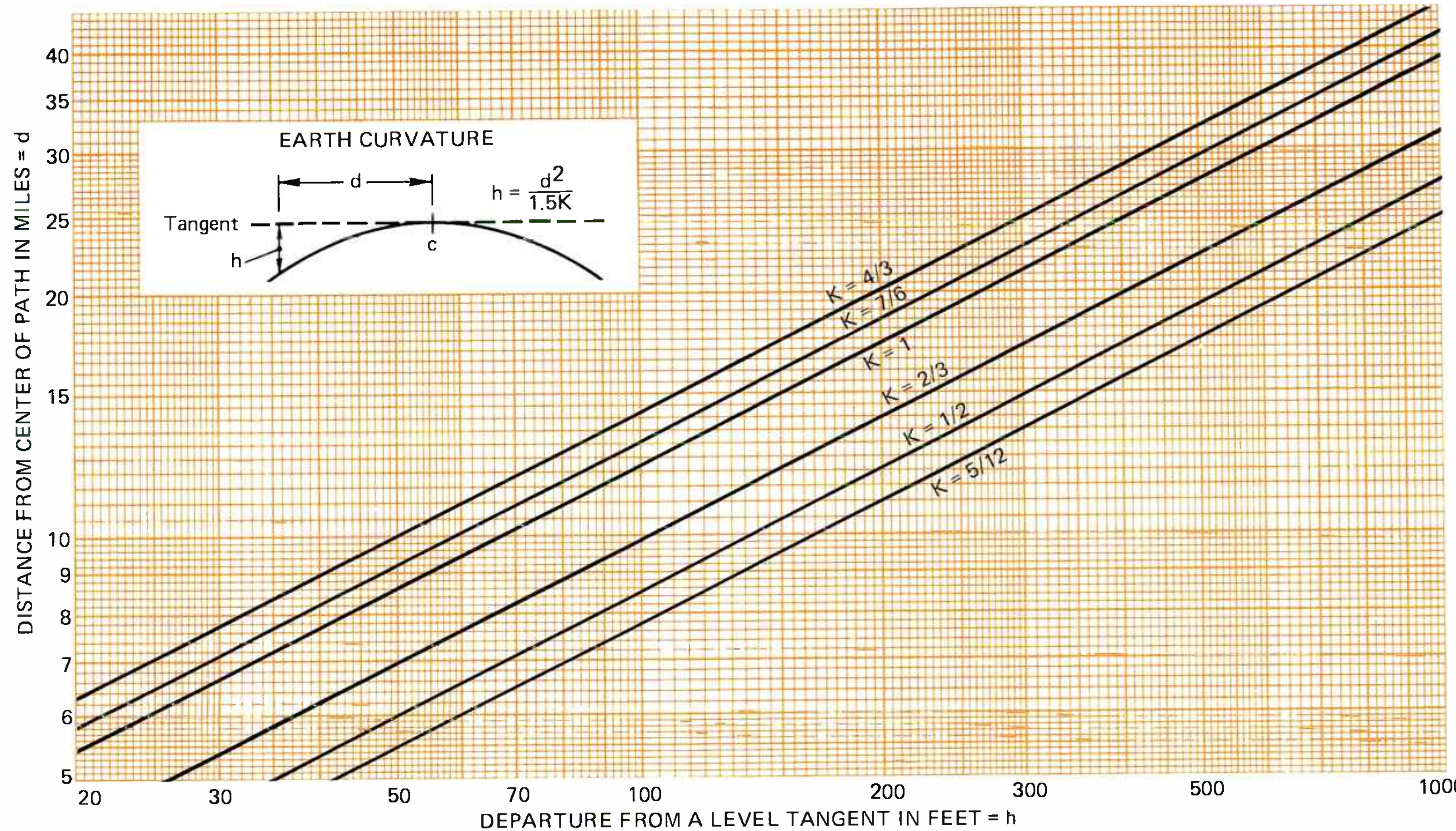


Figure 3. Earth Curvature for Various Values of K

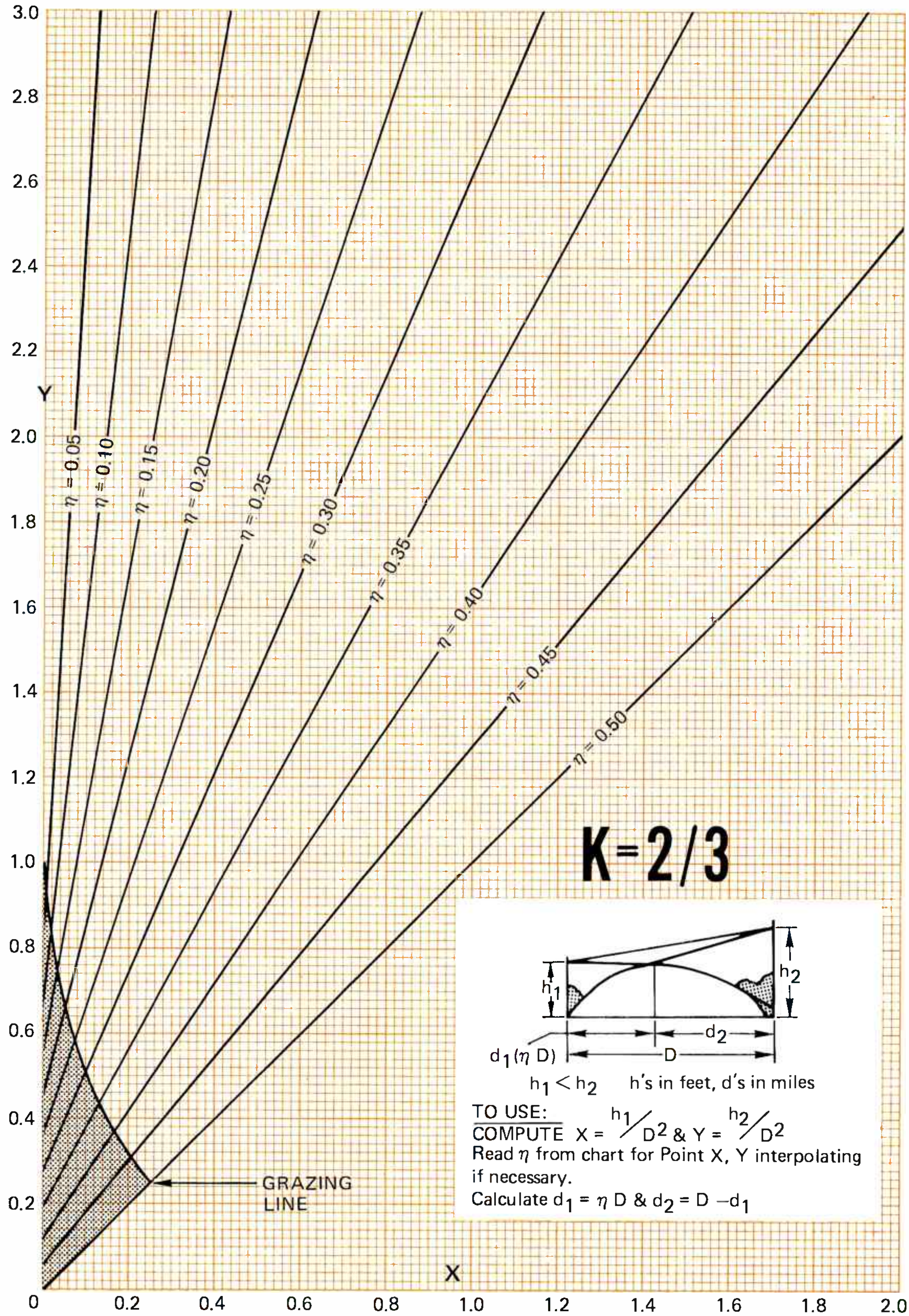


Figure 7A. Point of Reflection On Over-Water Microwave Path

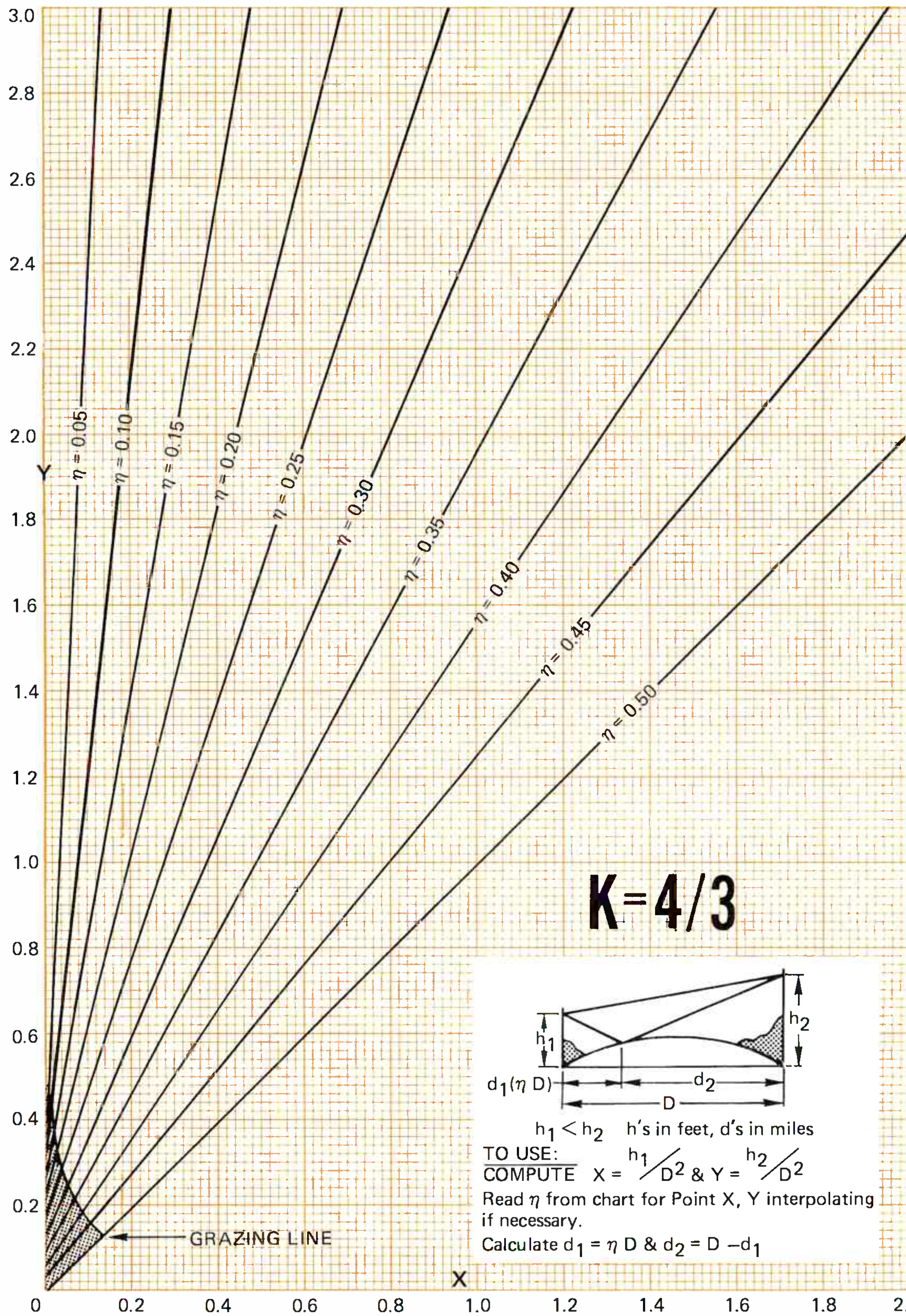


Figure 7B. Point of Reflection On Over-Water Microwave Path

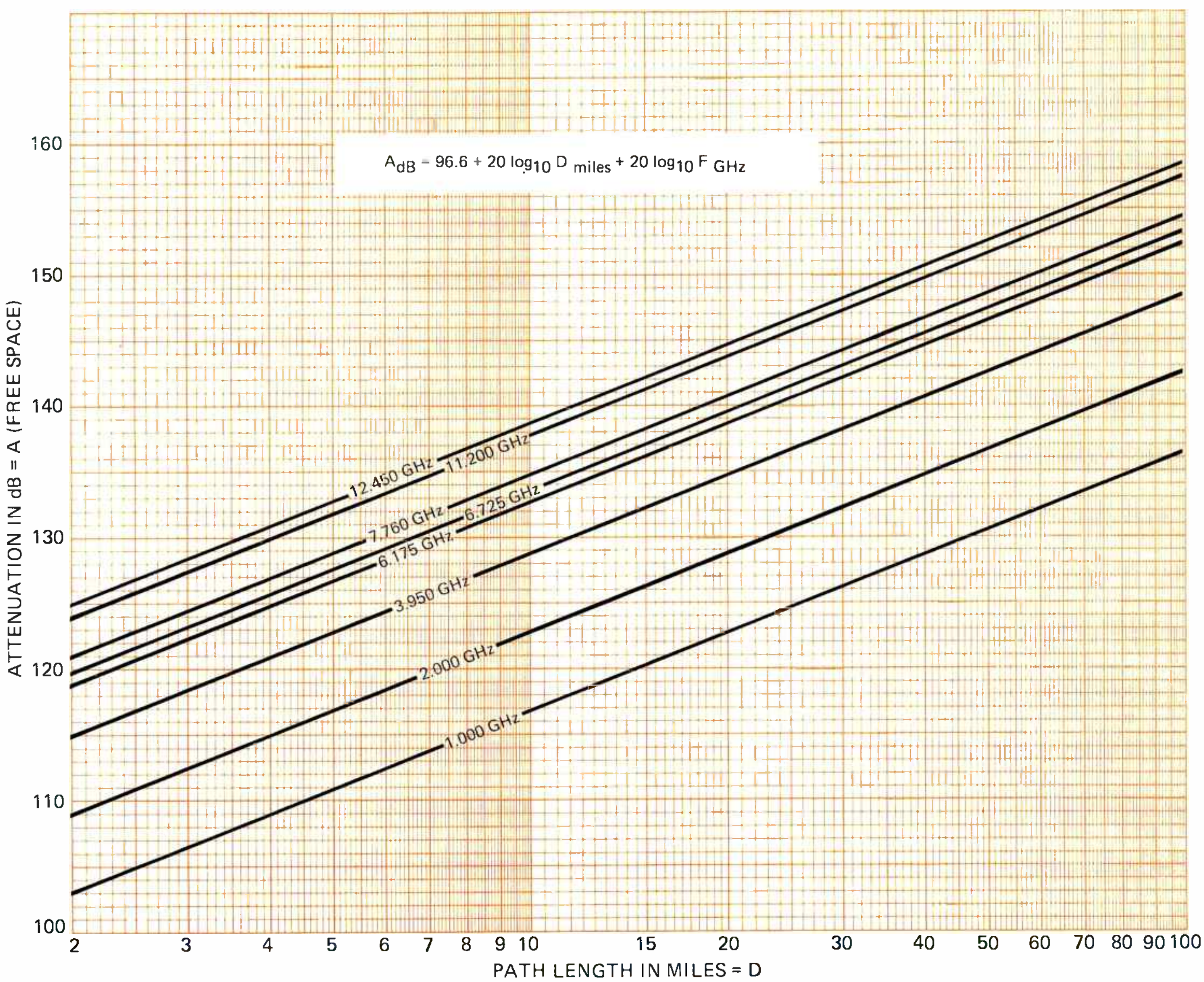


Figure 13. Free Space Attenuation Between Isotropic Antennas

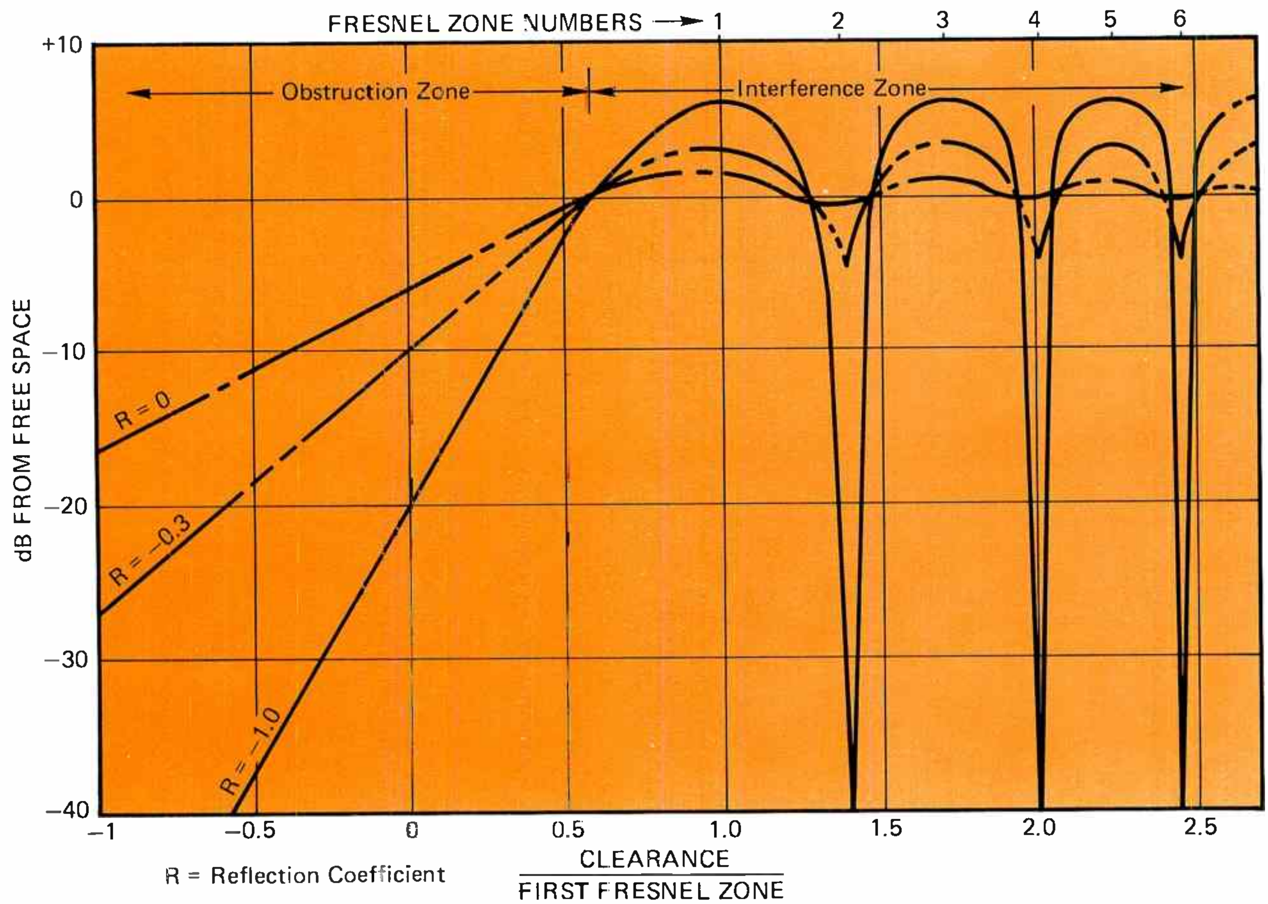


Figure 14. Behavior of Attenuation vs. Path Clearance for Various Types of Obstruction

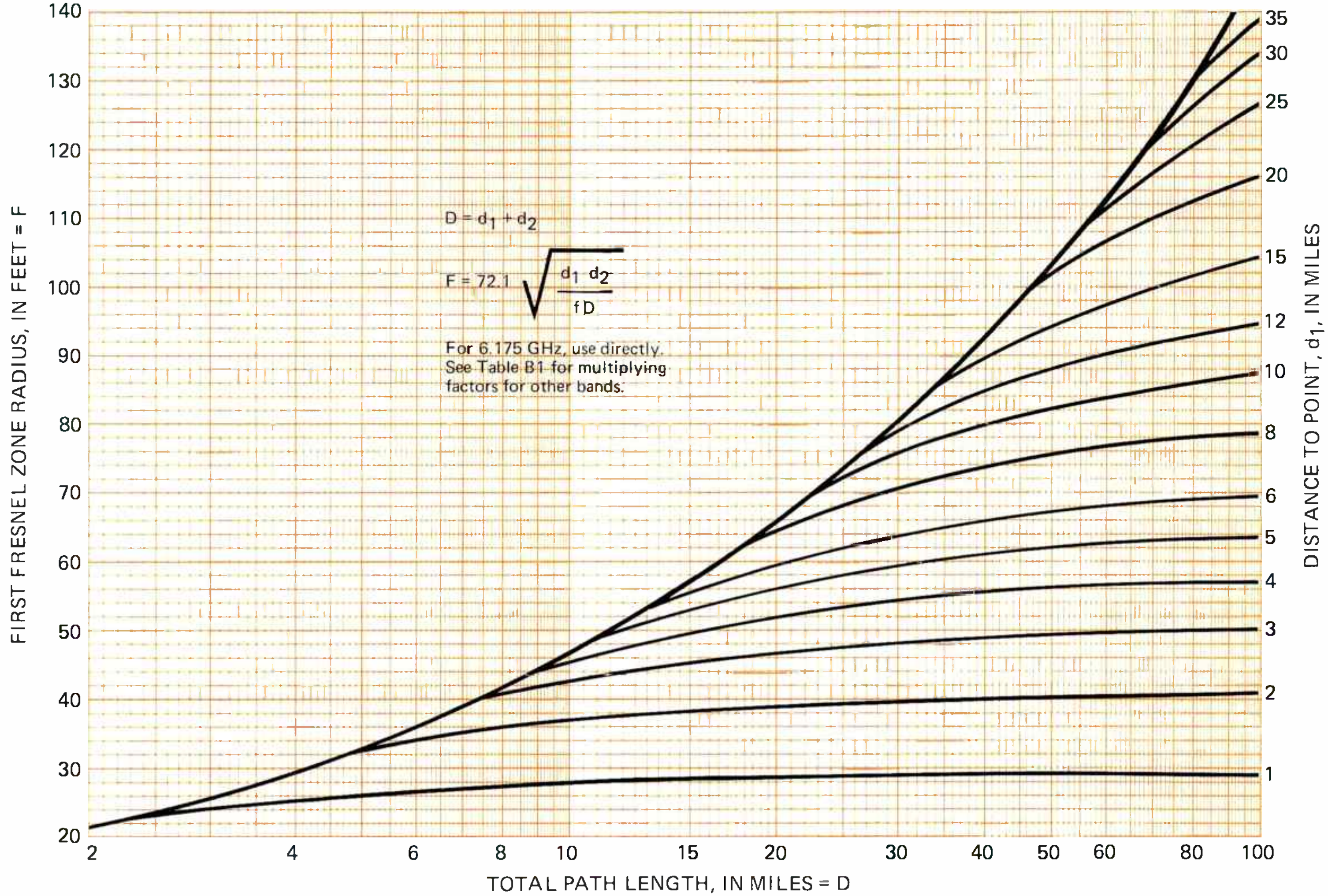


Figure 15. First Fresnel Zone Radius (6.175 GHz)

Table B2. Multiplying Factor For Determining F_n when F_1 is Known. ($F_n = F_1 \sqrt{n}$) P-40

n	$\sqrt{n}$	n	$\sqrt{n}$	n	$\sqrt{n}$	n	$\sqrt{n}$	n	$\sqrt{n}$
1	1.000	16	4.000	31	5.568	46	6.782	61	7.810
2	1.414	17	4.123	32	5.657	47	6.856	62	7.874
3	1.732	18	4.243	33	5.745	48	6.928	63	7.937
4	2.000	19	4.359	34	5.831	49	7.000	64	8.000
5	2.236	20	4.472	35	5.916	50	7.071	65	8.062
6	2.449	21	4.583	36	6.000	51	7.141		
7	2.646	22	4.690	37	6.083	52	7.211		
8	2.828	23	4.796	38	6.164	53	7.280		
9	3.000	24	4.899	39	6.245	54	7.348		
10	3.162	25	5.000	40	6.325	55	7.416		
11	3.317	26	5.099	41	6.403	56	7.483		
12	3.464	27	5.196	42	6.481	57	7.550		
13	3.606	28	5.291	43	6.557	58	7.616		
14	3.742	29	5.385	44	6.633	59	7.681		
15	3.873	30	5.477	45	6.708	60	7.746		

Table B1. Multiplying Factors Which Can Be Used To Convert Fresnel Zone Radii Calculated For 6.175 GHz To Other Bands. P-41

BAND GHz	CENTER FREQUENCY	MULTIPLY BY
1.850 – 1.990	1.920	1.793
1.990 – 2.110	2.050	1.735
2.110 – 2.130 2.160 – 2.180	2.145	1.697
2.130 – 2.150 2.180 – 2.200	2.165	1.688
2.450 – 2.500	2.475	1.580
3.700 – 4.200	3.950	1.250
4.400 – 5.000	4.700	1.146
5.925 – 6.425	6.175	1.000
6.575 – 6.875	6.725	0.9582
6.875 – 7.125	7.000	0.9392
7.125 – 8.400	7.437	0.9112
	7.750	0.8926
	8.063	0.8751
10.700 – 11.700	11.200	0.7425
12.200 – 12.700	12.450	0.7043
12.700 – 12.950	12.825	0.6939
12.700 – 13.250	12.975	0.6899

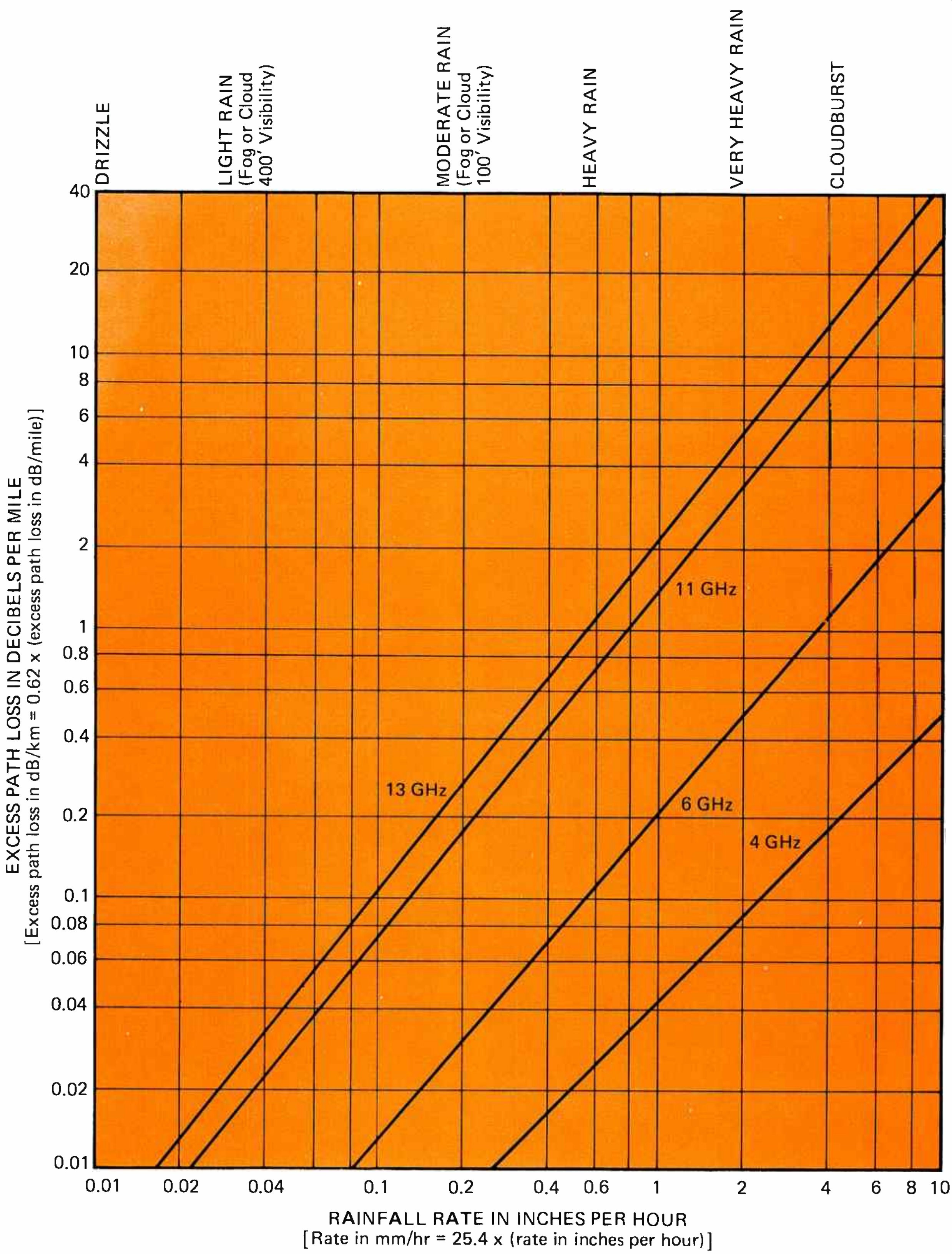
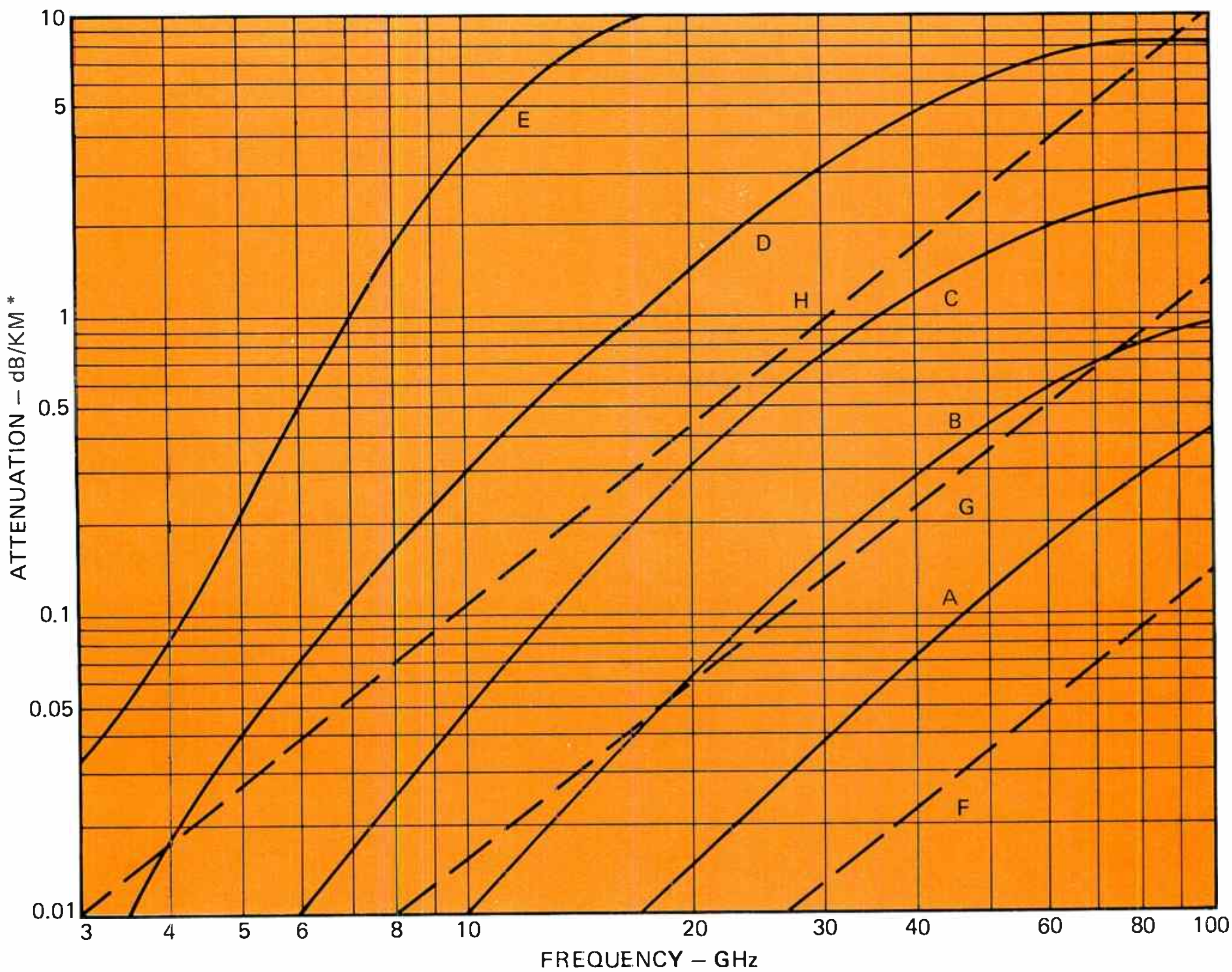


Figure 17. Rain Attenuation vs Rainfall Rate (Theoretical, after Ryde and Ryde)



——— Attenuation in rainfall intensity of:

A, 0.25 mm/hr (drizzle) — .01 in/hr
B, 1.0 mm/hr (light rain) — .04 in/hr
C, 4.0 mm/hr (moderate rain) — .16 in/hr
D, 16 mm/hr (heavy rain) — .64 in/hr
E, 100 mm/hr (very heavy rain) — 4.0 in/hr

— — — Attenuation in fog or cloud:

F, 0.032 gm/m^3 (visibility greater than 600 meters)
G, 0.32 gm/m^3 (visibility about 120 meters)
H, 2.3 gm/m^3 (visibility about 30 meters)

*Attn – dB/mile = 1.61 x (Attn in dB/km)

Figure 18. Attenuation Due To Precipitation (after CCIR)

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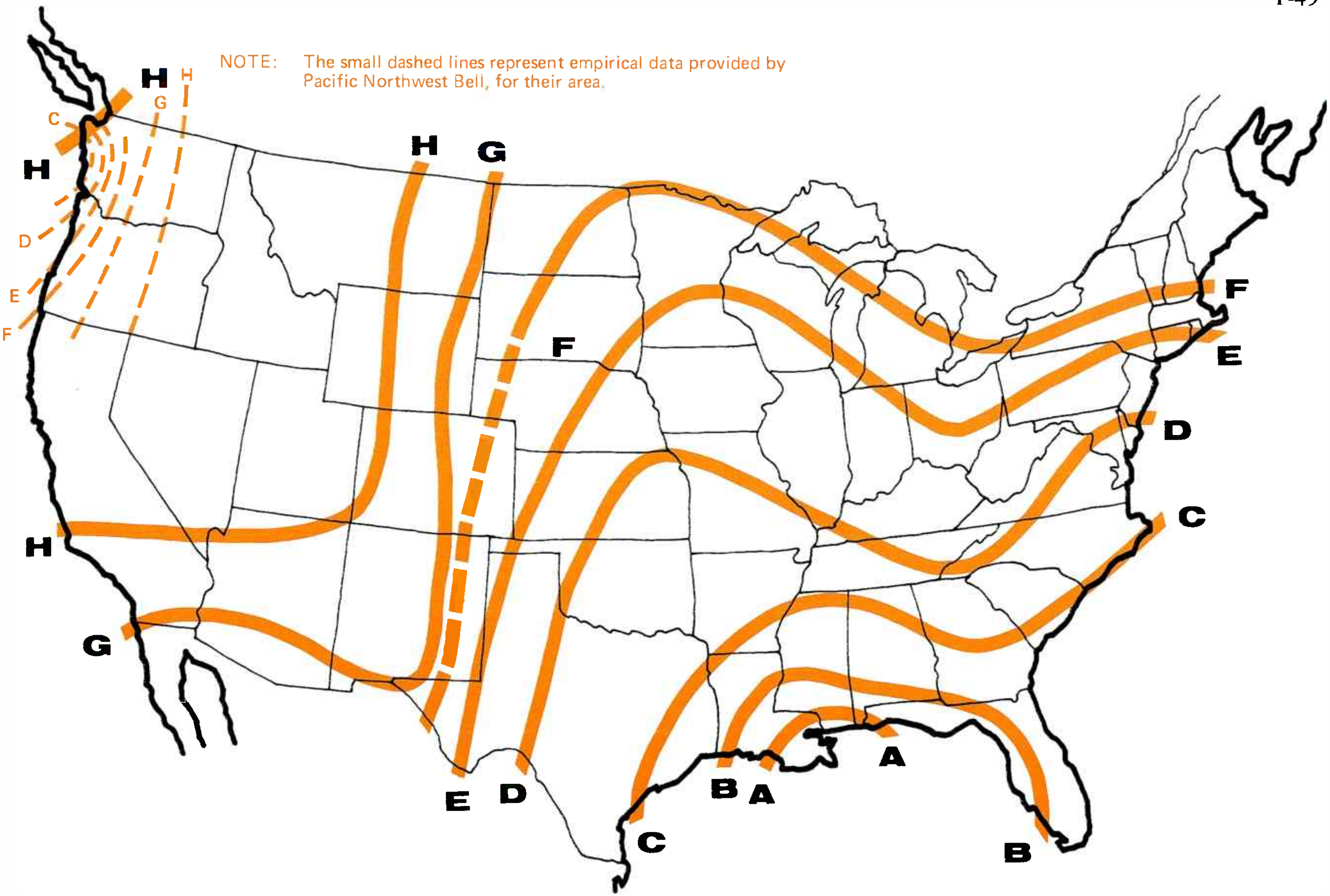


Figure 19. Contours Of Constant Path Length For Fixed Outage Time.

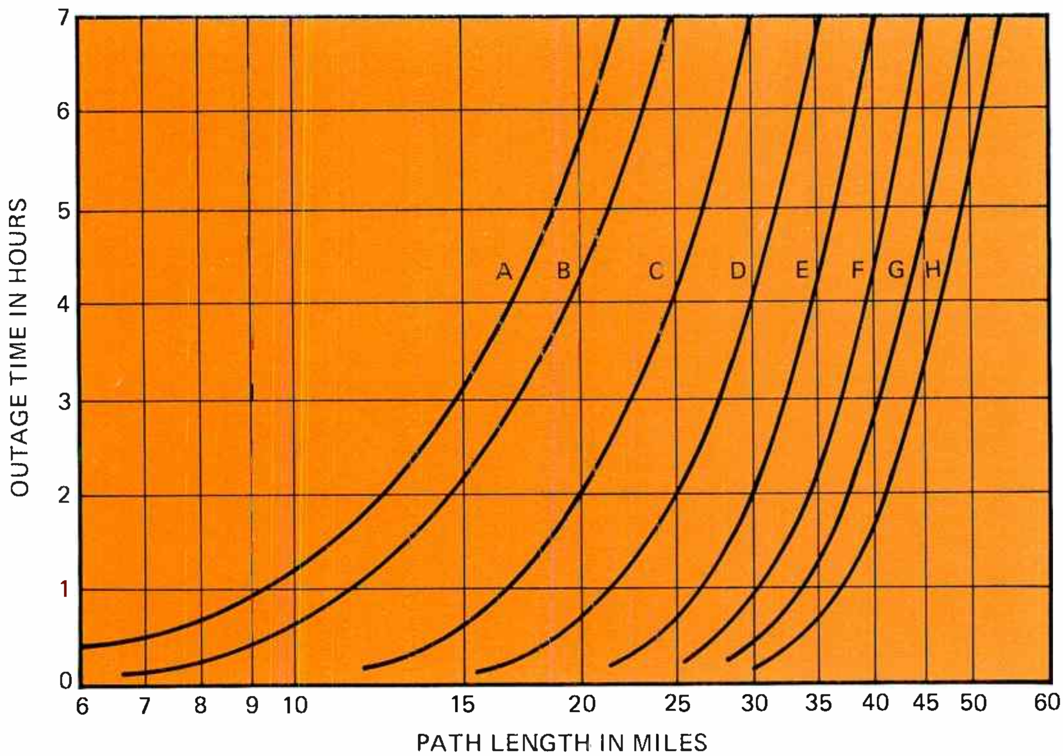


Figure 20. Expected Outage Time In Hours Per Year vs. Path Length in Miles For Various Areas of the United States. (Based on 11 GHz paths with 40 dB fade margin; for 13 GHz paths, reduce path lengths by 30%. For 45 dB fade margin, decrease outage time by 15%; for 35 dB fade margin, increase it by 25%).

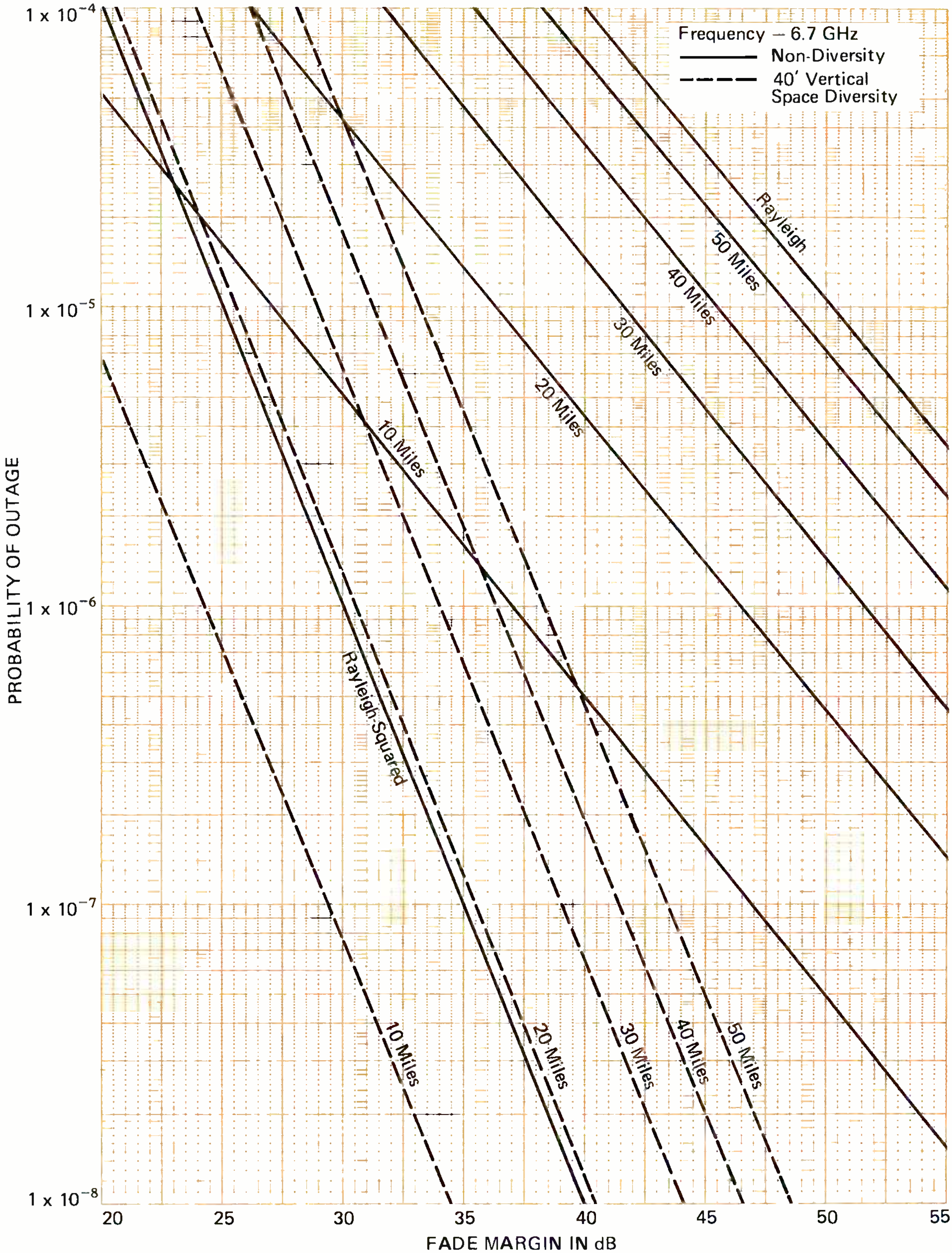


Figure 21. Outage Probability vs. Fade Margin for 6.7 GHz paths of Various Length, Average Terrain and Climate (after Vigants & Barnett)

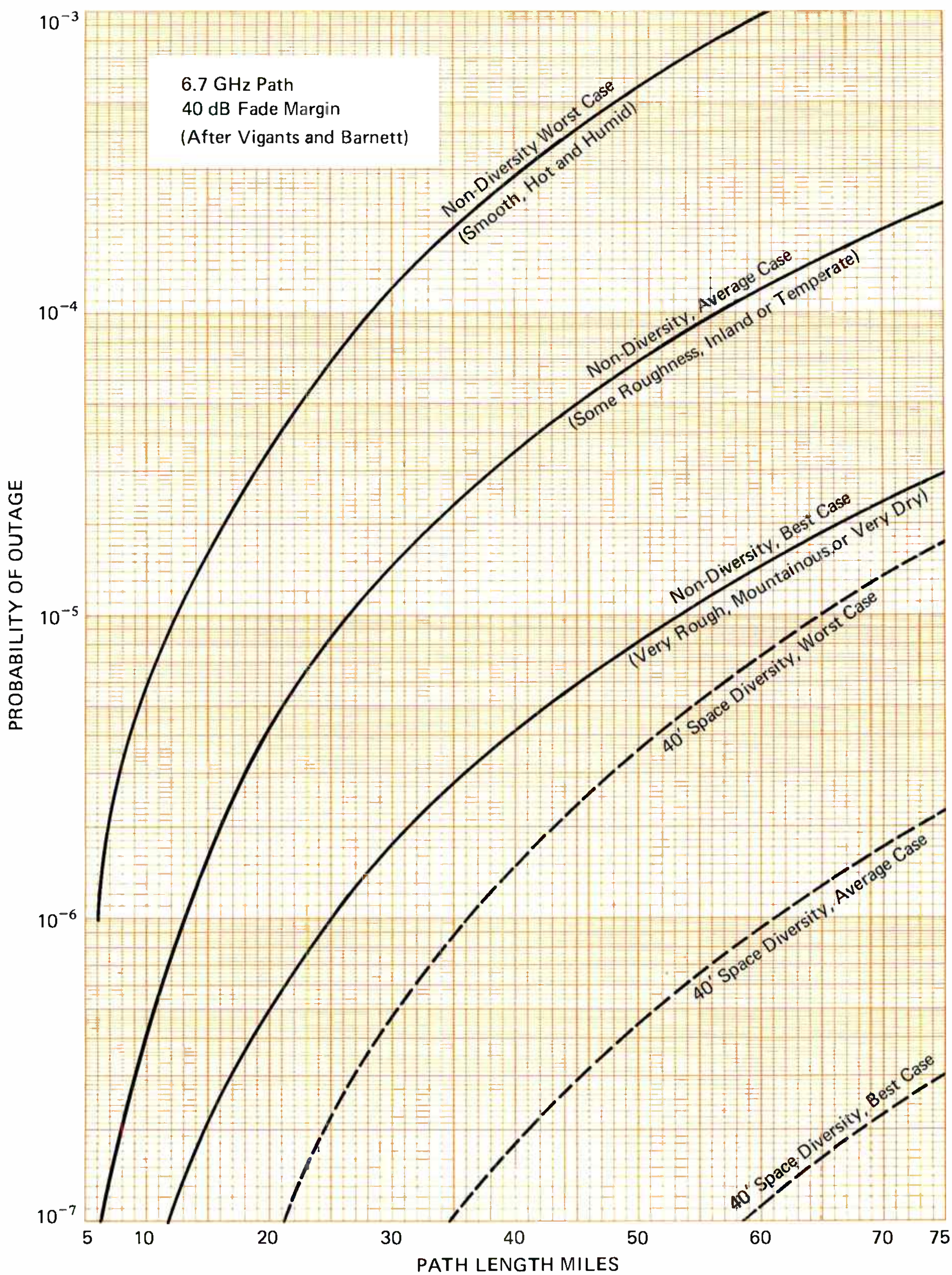


Figure 22. Outage Probability vs Path Length for a 6.7 GHz Path with 40 dB Fade Margin

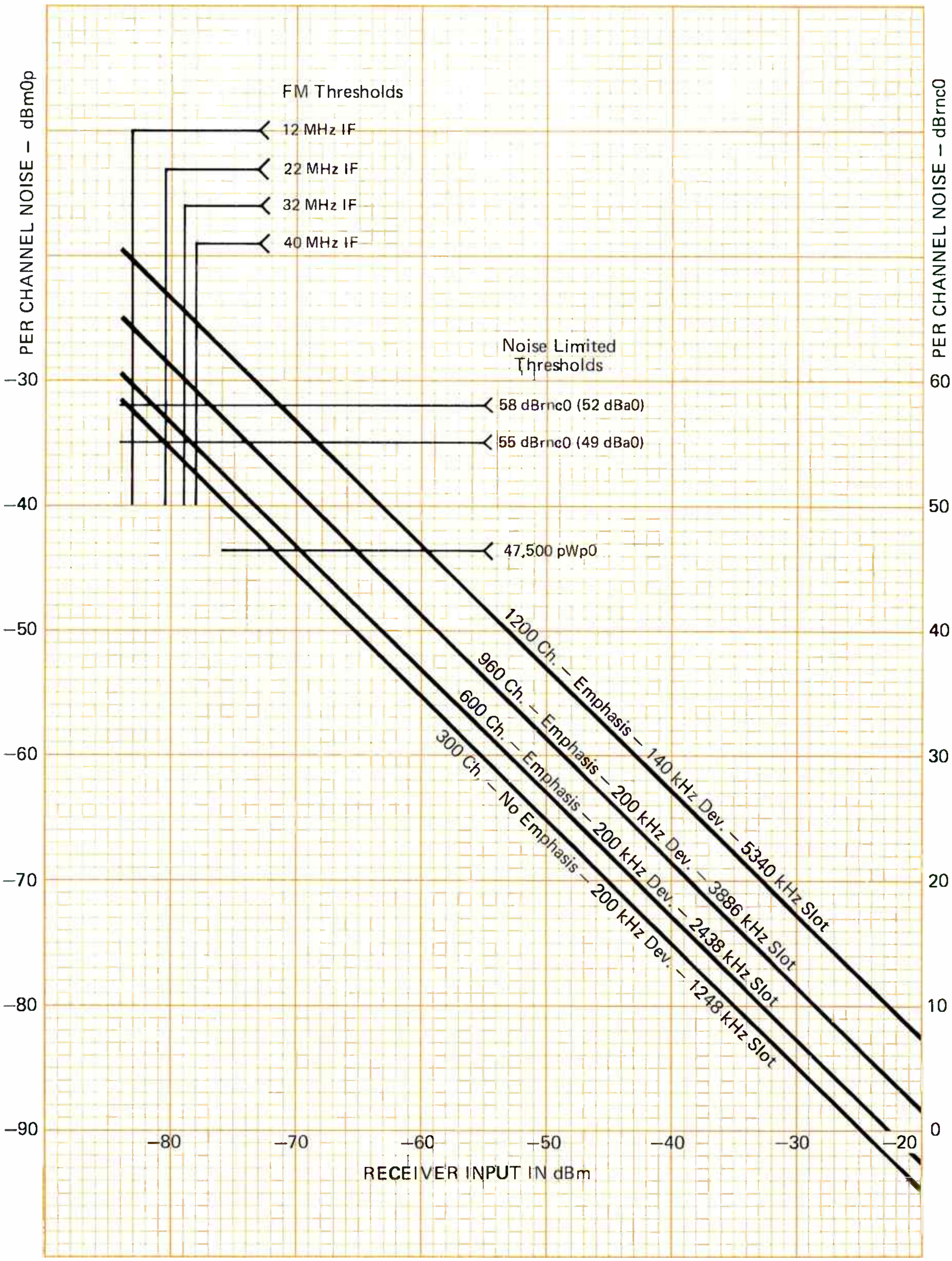


Figure 23A. Receiver Thermal Noise. (10 dB Noise Figure Assumed)

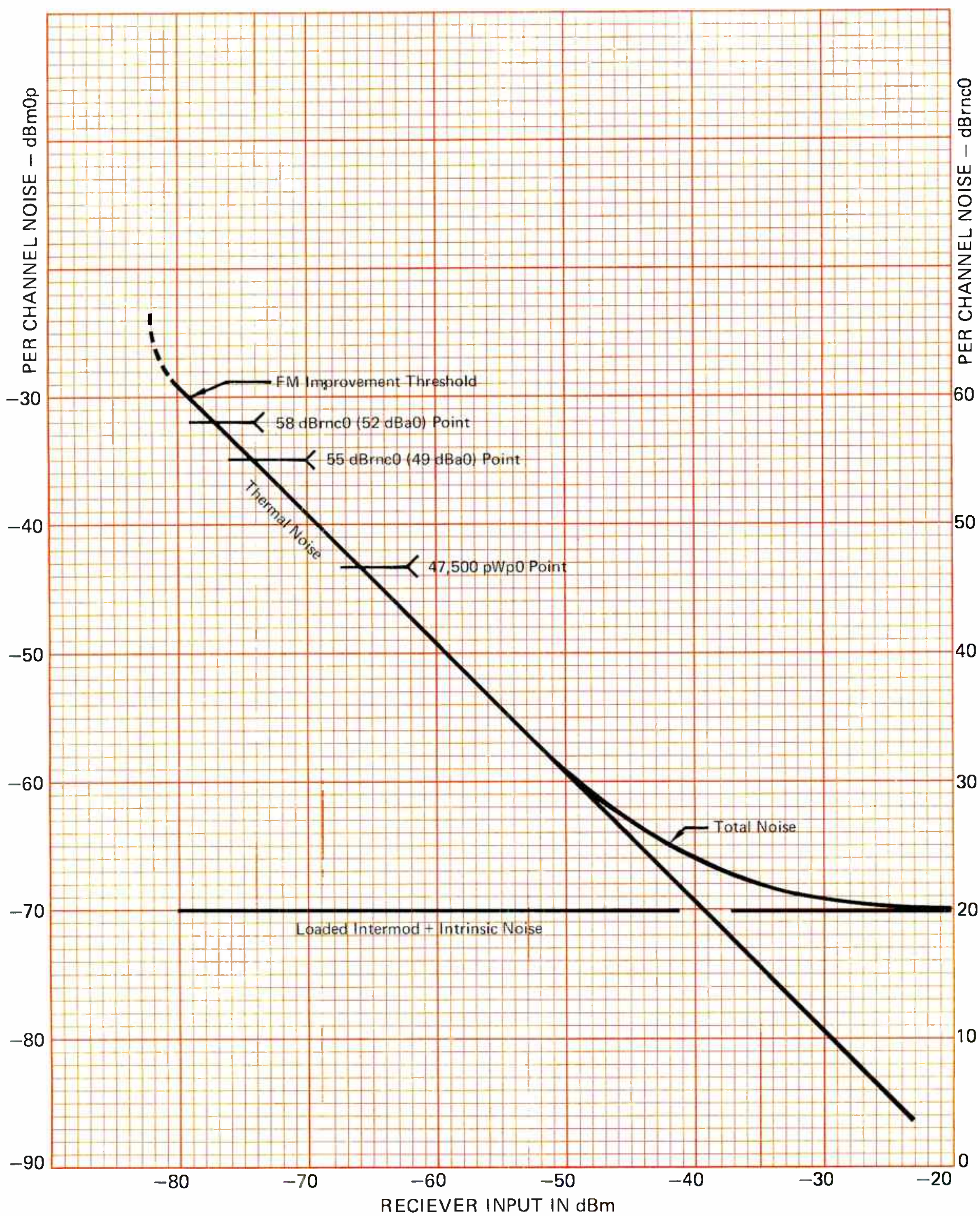


Figure 23B. Typical Receiver Noise Curve. Top Slot of 960 Channel System; 10 dB Noise Figure; 200 kHz Deviation; CCIR Emphasis, +14.8 dBm0 Load.

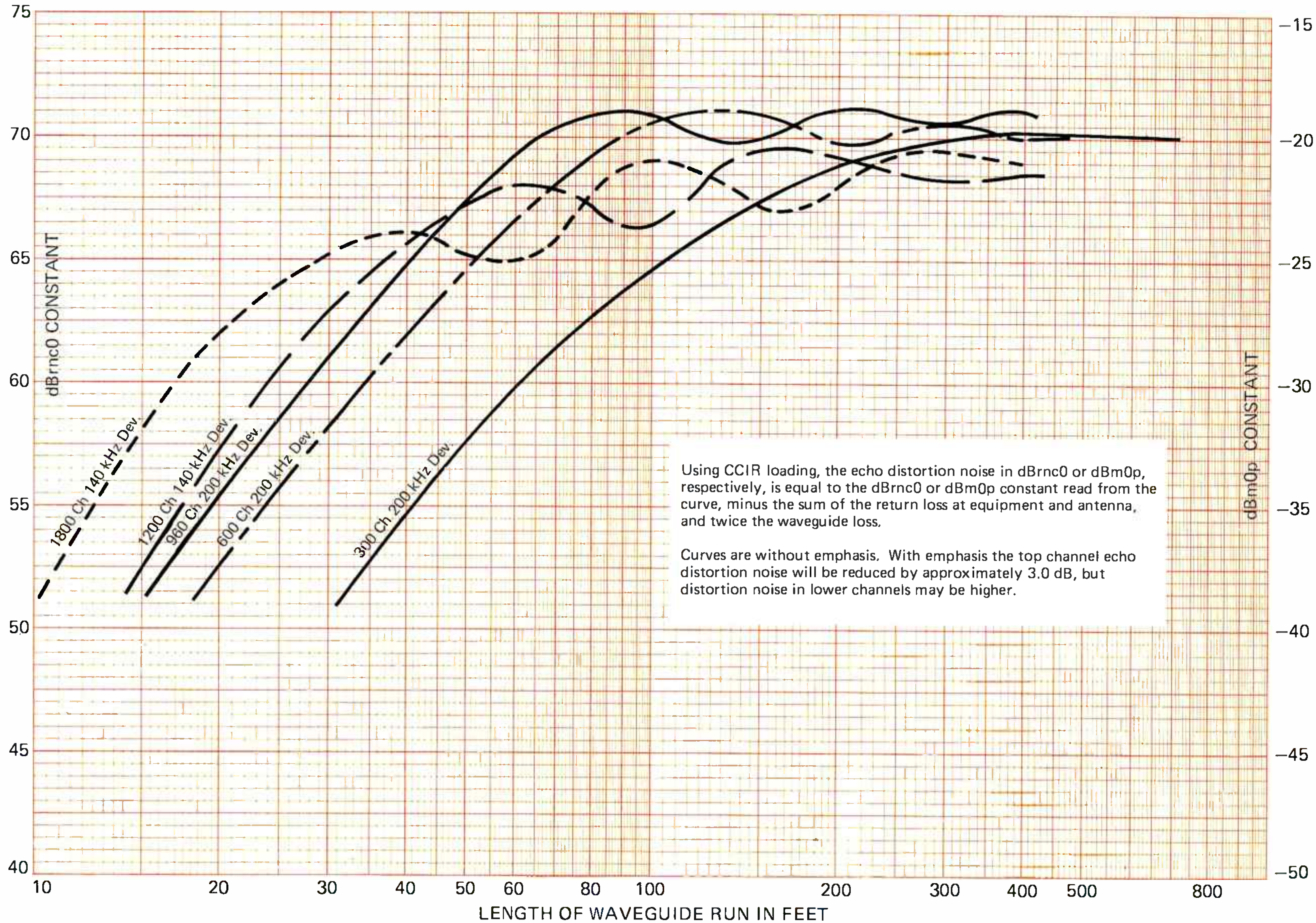
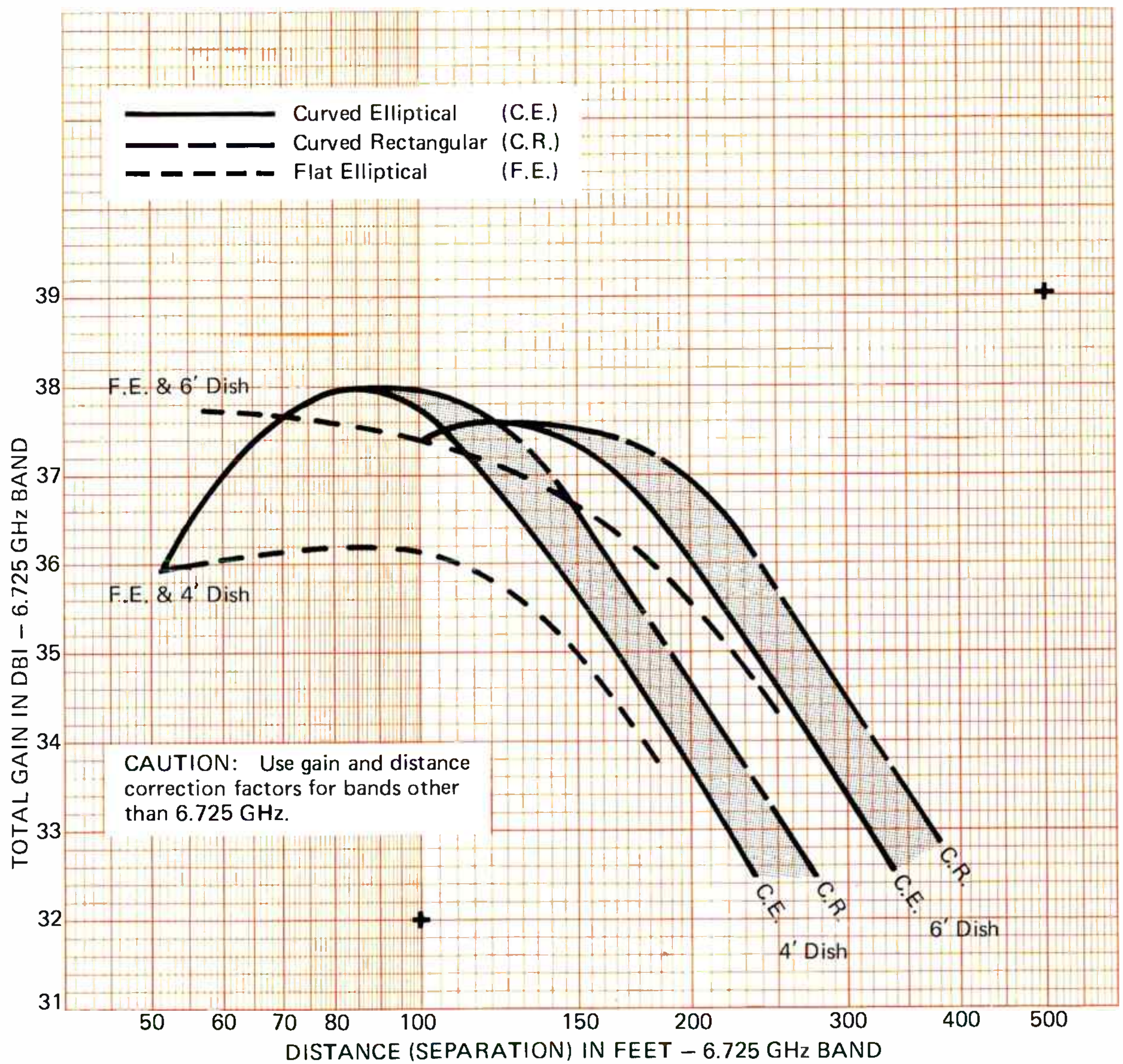


Figure 24. Echo Distortion Noise



Band GHz	Gain Factor	Distance Factor
1.92	–10.8	3.50
2.15	– 9.9	3.13
3.95	– 4.6	1.70
4.7	– 3.1	1.43
6.175	– 0.7	1.09
6.725	–	–
7.0	+ 0.4	.96
7.437	+ 0.9	.91
8.0	+ 1.5	.84
11.2	+ 4.5	.60
12.45	+ 5.4	.54
14.8	+ 6.9	.46

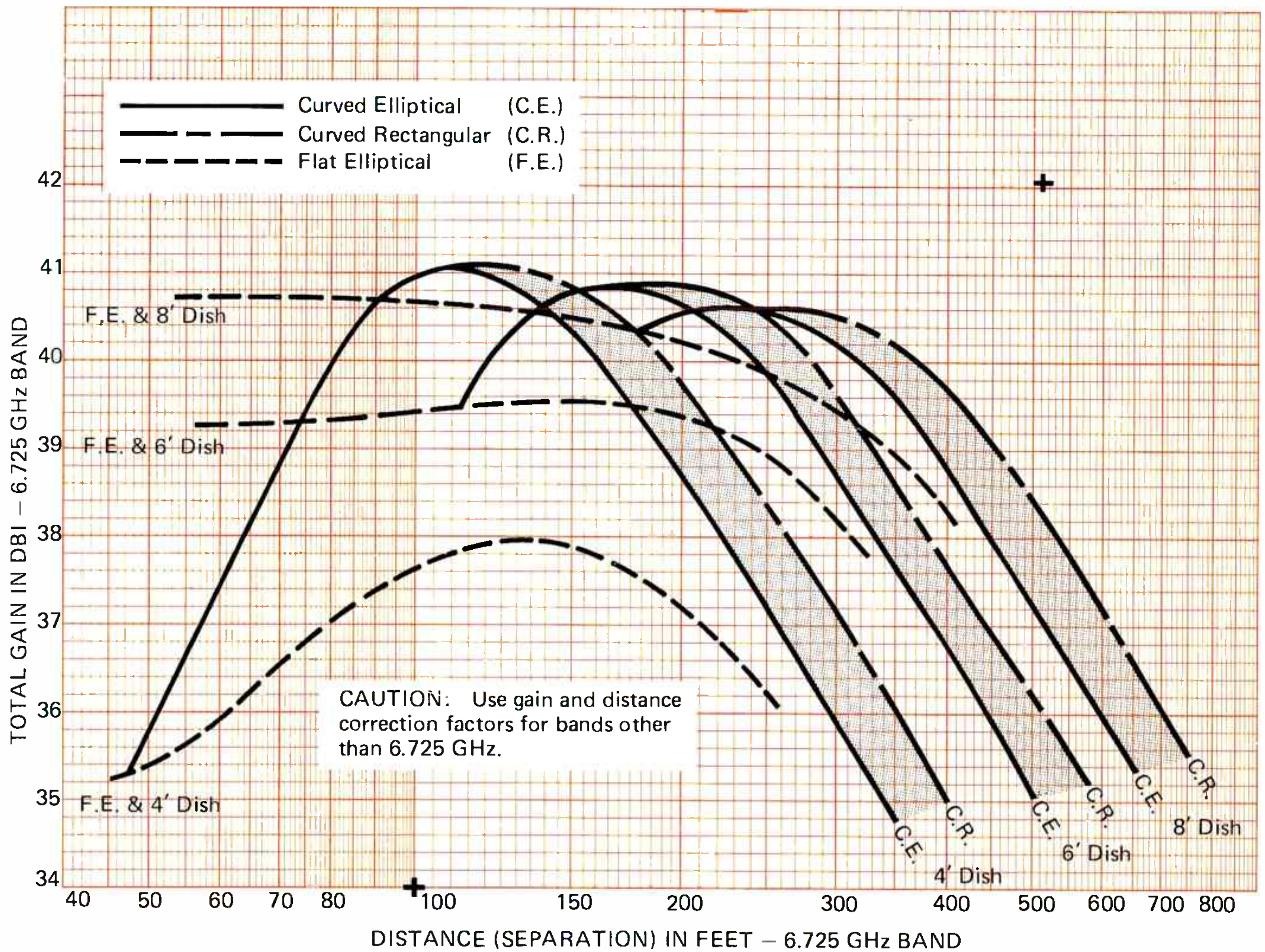
True Gain = Chart Gain + G Factor

$$\text{True Dist.} = \frac{\text{Chart Dist.}}{\text{Dist. Fact.}}$$

Chart Dist. = True Dist. × Dist. Factor

Figure 27A. Periscope Gain Curves for 6'x8' Reflectors

C21



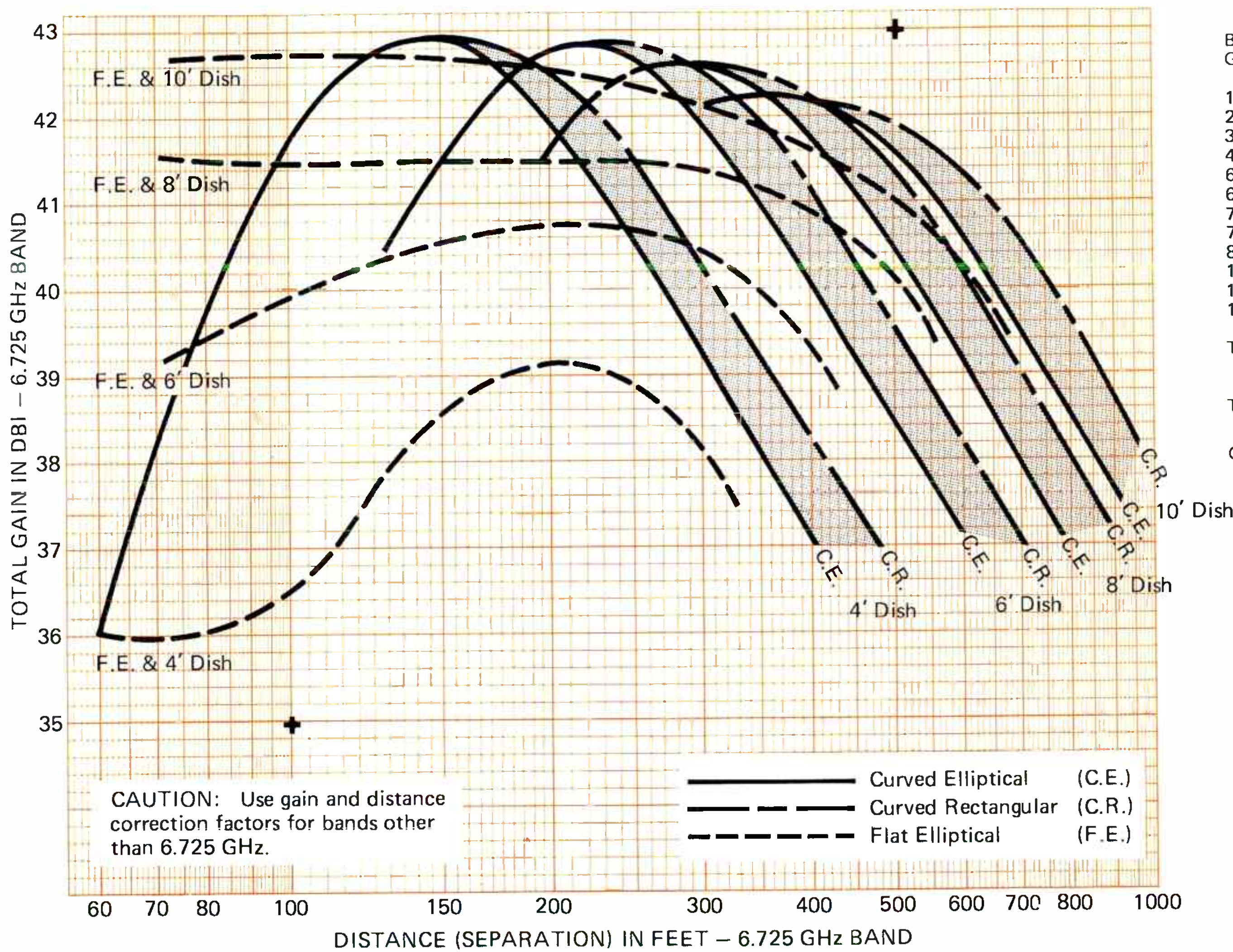
Band GHz	Gain Factor	Distance Factor
1.92	−10.8	3.50
2.15	− 9.9	3.13
3.95	− 4.6	1.70
4.7	− 3.1	1.43
6.175	− 0.7	1.09
6.725	–	–
7.0	+ 0.4	.96
7.437	+ 0.9	.91
8.0	+ 1.5	.84
11.2	+ 4.5	.60
12.45	+ 5.4	.54
14.8	+ 6.9	.46

True Gain = Chart Gain + G Factor

True Dist. = $\frac{\text{Chart Dist.}}{\text{Dist. Fact.}}$

Chart Dist. = True Dist. x Dist. Factor

Figure 27B. Periscope Gain Curves for 8'x12' Reflectors



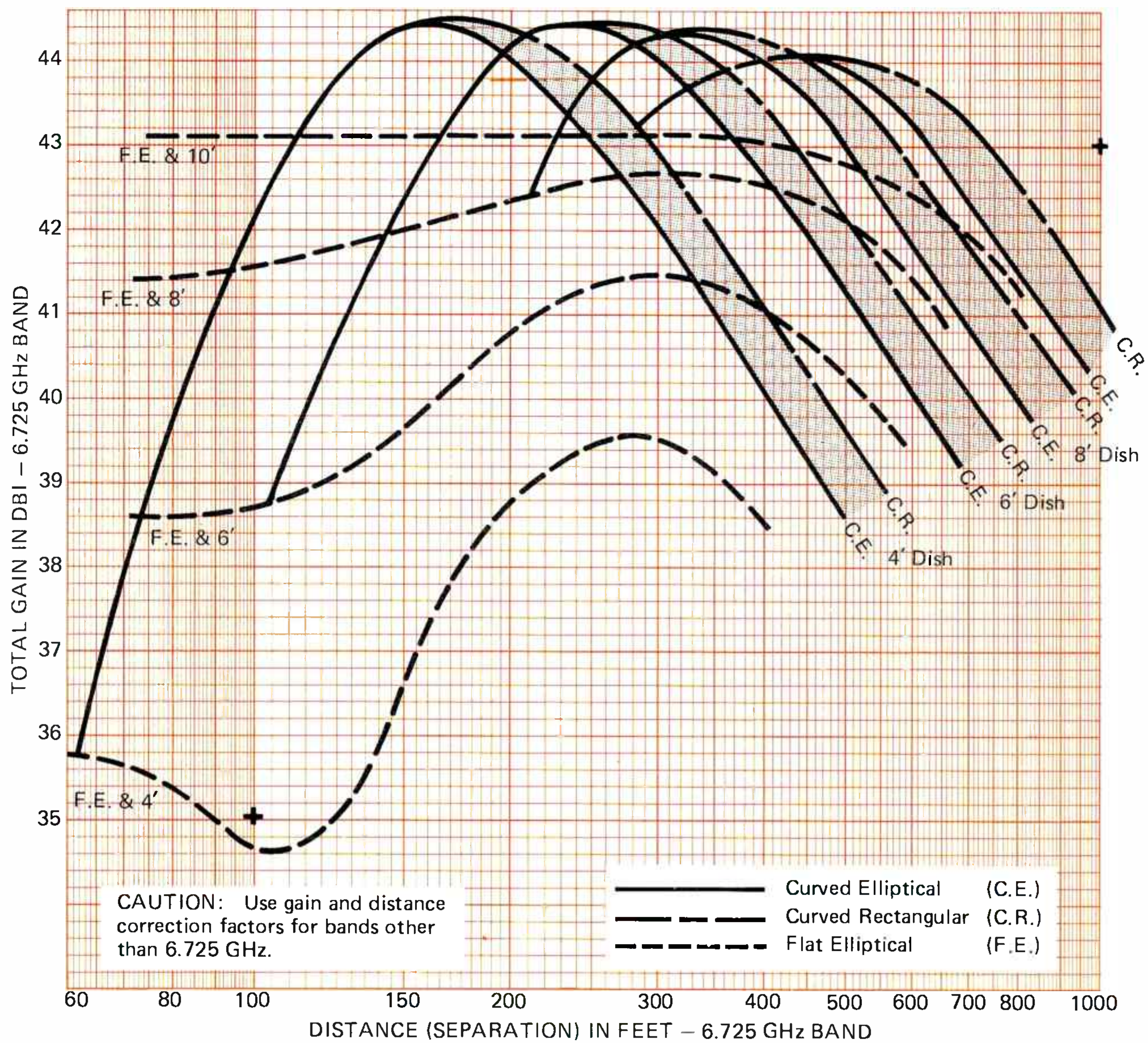
Band GHz	Gain Factor	Distance Factor
1.92	−10.8	3.50
2.15	− 9.9	3.13
3.95	− 4.6	1.70
4.7	− 3.1	1.43
6.175	− 0.7	1.09
6.725	–	–
7.0	+ 0.4	.96
7.437	+ 0.9	.91
8.0	+ 1.5	.84
11.2	+ 4.5	.60
12.45	+ 5.4	.54
14.8	+ 6.9	.46

True Gain = Chart Gain + G Factor

$$\text{True Dist.} = \frac{\text{Chart Dist.}}{\text{Dist. Fact.}}$$

Chart Dist. = True Dist. x Dist. Factor

Figure 27C. Periscope Gain Curves for 10'x15' Reflectors



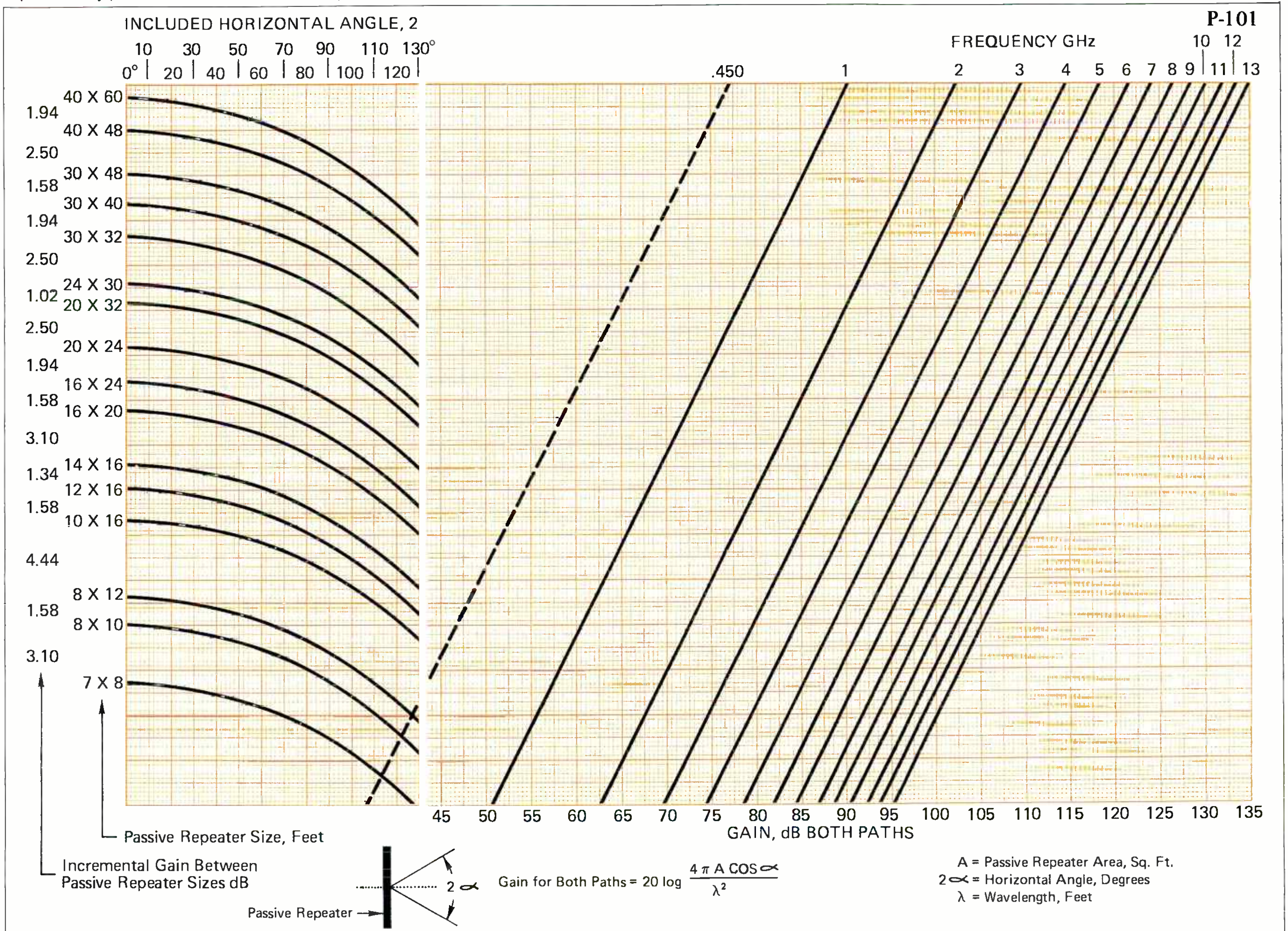
Band GHz	Gain Factor	Distance Factor
1.92	−10.8	3.50
2.15	− 9.9	3.13
3.95	− 4.6	1.70
4.7	− 3.1	1.43
6.175	− 0.7	1.09
6.725	—	—
7.0	+ 0.4	.96
7.437	+ 0.9	.91
8.0	+ 1.5	.84
11.2	+ 4.5	.60
12.45	+ 5.4	.54
14.8	+ 6.9	.46

True Gain = Chart Gain + G Factor

True Dist. $= \dfrac{\text{Chart Dist.}}{\text{Dist. Fact.}}$

Chart Dist. = True Dist. x Dist. Factor

Figure 27D. Periscope Gain Curves for 12'x17' Reflectors



Gain for Both Paths = 20 log $\frac{4\pi A \cos\alpha}{\lambda^2}$

A = Passive Repeater Area, Sq. Ft.
2α = Horizontal Angle, Degrees
λ = Wavelength, Feet

Figure 28A. Passive Repeater Gain Chart

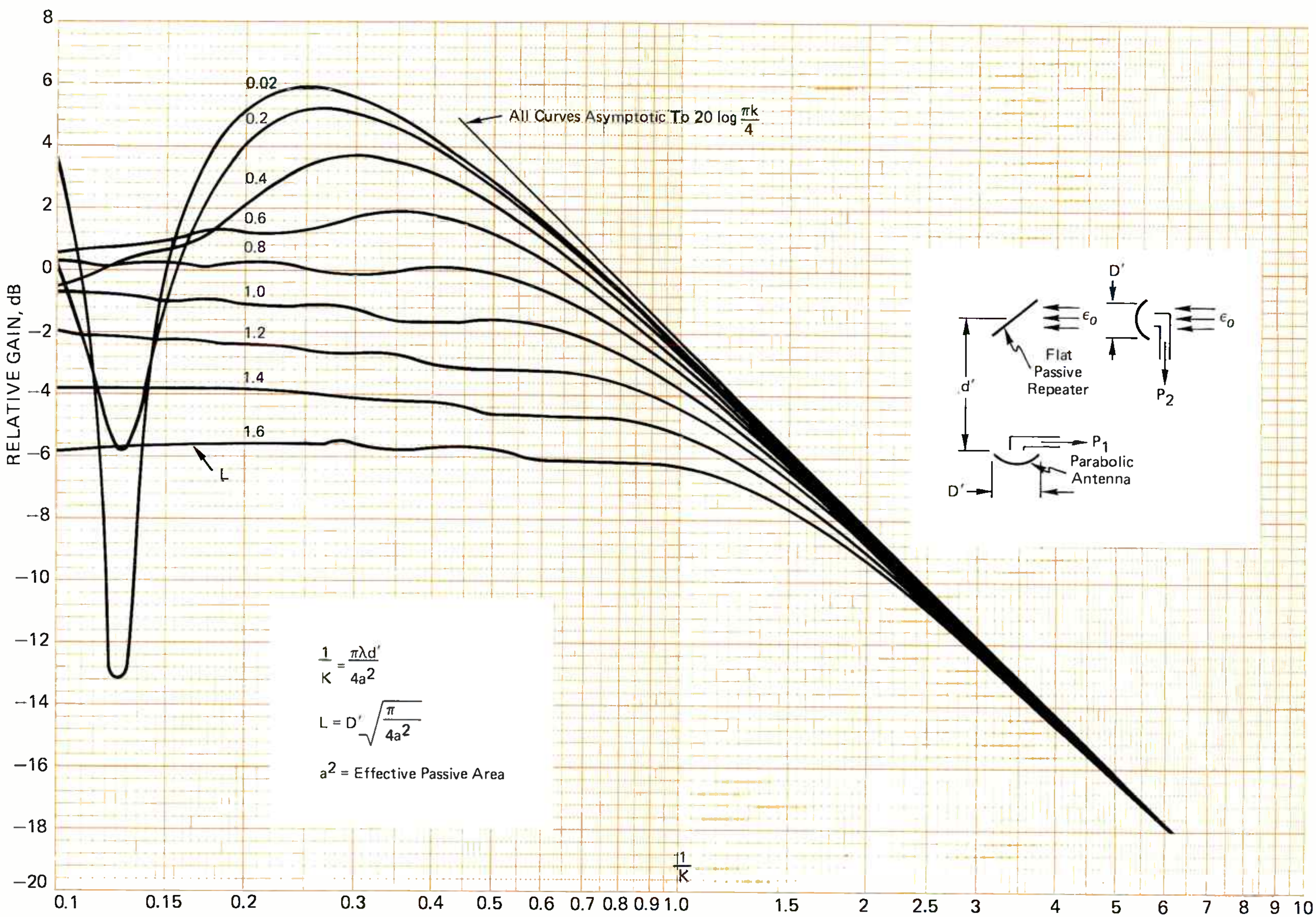


Figure 28B. Antenna-Reflector Efficiency Curves

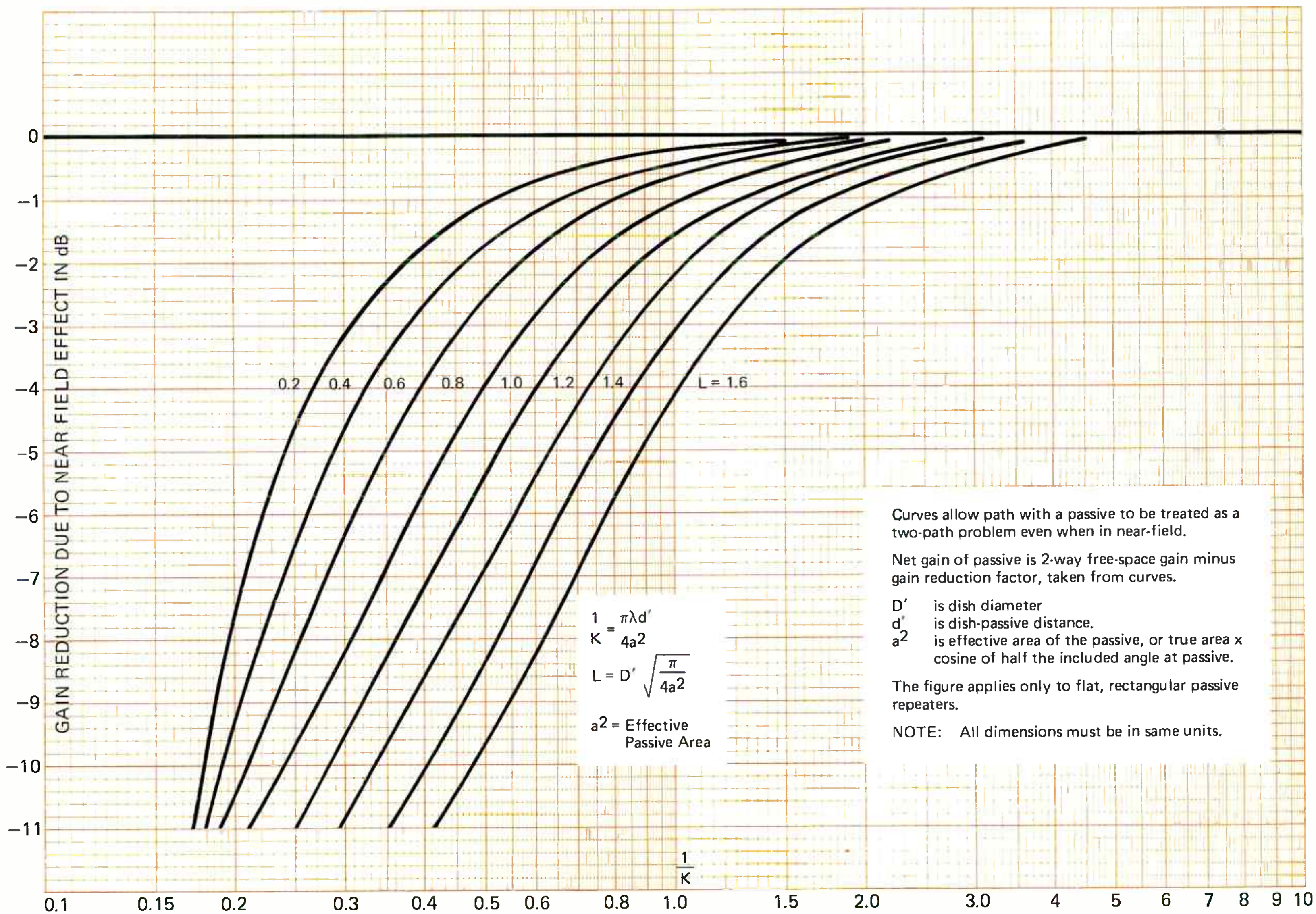


Figure 28C. Passive Repeater Gain Correction; When Passive In Near Field

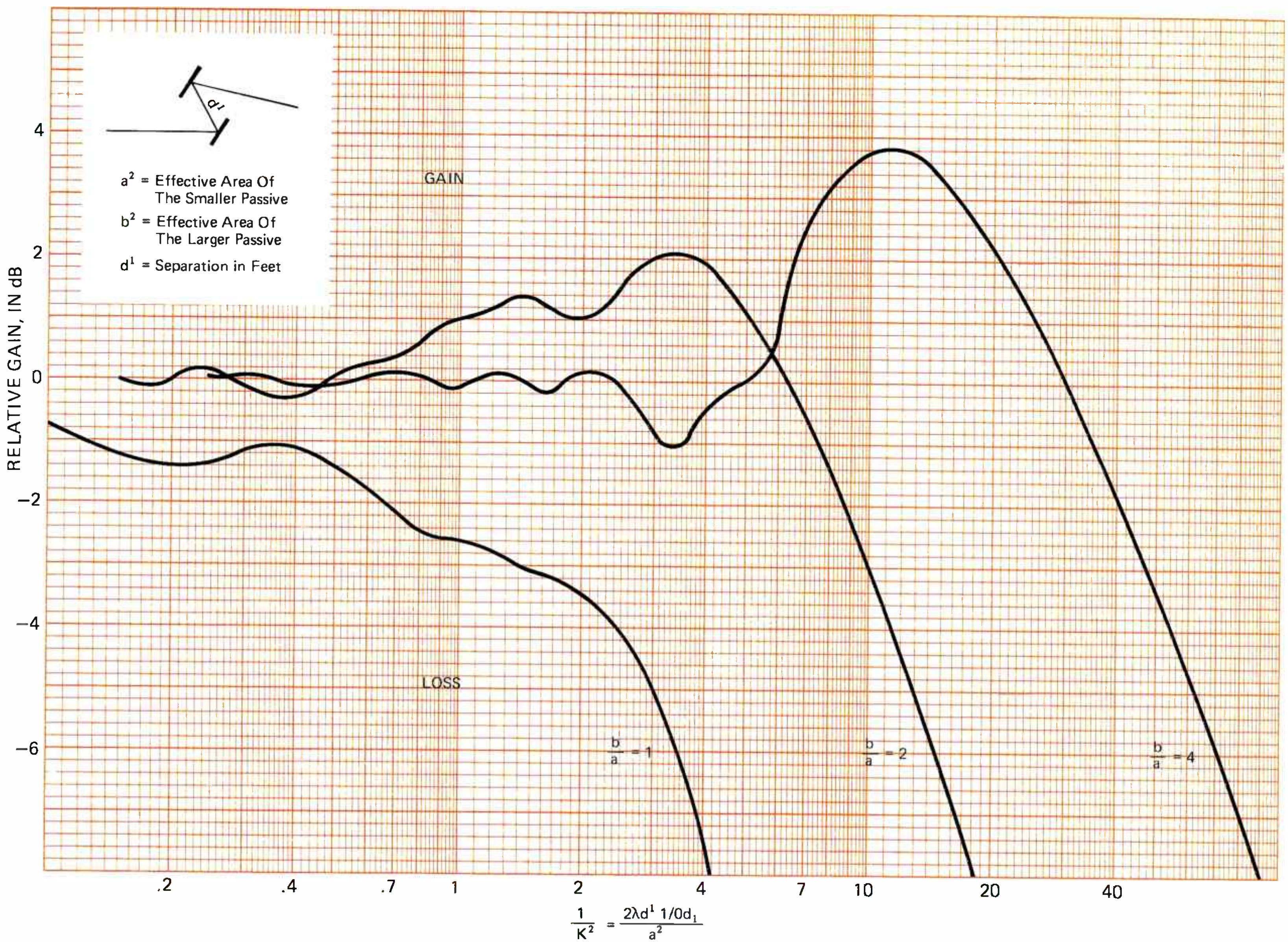


Figure 28D. Double Passive Repeater Efficiency Curves.

This memorandum describes a method for making accurate calculations of the true azimuths, at each end of a path, and the path distance, for microwave paths whose end coordinates (latitudes and longitudes) are accurately known.

The method is an adaptation from the "Inverse Position Computation" on page 14 of "Special Publication No. 8", Coast and Geodetic Survey; Formulas and Tables for the Computation of Geodetic Positions. Factors (log B and log A) from certain tables in that publication are used in the computation. The tables list these factors for every minute of latitude from 0° to 72°, but this degree of accuracy is not needed for these calculations. Following the FCC approach, we have extracted from the table only those values of log B and log A corresponding to integral degrees of latitude from 0° to 72°, and the necessary values are given in the attached Table "Extracts From Special Publication No. 8," so the publication itself is not required.

The listed values are log A_m and log B_m/A_m. Log B_m alone is not used in our computation, so it is not listed in the table. The subscript 'm' indicates that the values are taken for the average latitude of the path, the nearest value listed in the table being the one selected.

The errors introduced by using these factors to the nearest degree, rather than to the nearest minute of latitude, will not exceed a few seconds in azimuth and substantially less than a tenth of a mile in distance, a degree of accuracy more than adequate for path calculation purposes. Also, since the FCC's program takes the same approach, it should give results identical to that program, and thus avoid having them change the applied for azimuths and distances.

Note that the log A_m factor is always negative, the characteristic in all cases being $\bar{8}$.

Calculations are best made using six-place logarithm tables, though five-place tables will give reasonably accurate results.

Since the method takes into account the oblateness of the earth, it gives more precise values than an uncorrected great-circle calculation method. (For paths longer than about 75 miles, the great-circle method should be used.)

Two calculation work sheets are included, one a blank which can be used to make reproductions, and the other with a sample calculation to illustrate the method.

NOTE: The method will fail when the two stations have *exactly* the same longitude. In this case, one station will have an azimuth of 0° and the other an azimuth of 180°. For the distance calculation, log S_{meters} will be equal to log $\Delta\,\phi$sec minus log A_m minus log B_m/A_m. Also, when the two stations have *exactly* the same latitude, angle w is equal to 0° and need not be calculated.

Inverse Position Azimuth and Path Distance Calculation Sheet

		Latitude	Longitude
Station West	ADAIRSVILLE	34° 19' 01"	84° 53' 52"
Station East	MARIETTE	33° 57' 01"	84° 39' 57"
Difference,		$\Delta\phi$ 22' 00"	$\Delta\lambda$ 13' 55"
Convert to seconds		$\Delta\phi_{sec}$ 1320"	$\Delta\lambda_{sec}$ 835"

$\phi_m = \phi_{smaller} + \frac{\Delta\phi_{sec}}{2}$ = 33° 57' 01" + 11' 00" = 34° 08' 01"

Azimuth Calculations

log B_m/A_m(1)	.002029	
log cos ϕ_m	$\overline{9}$.917890	add
log $\Delta\lambda_{sec}$	2.921686	
	2.841605	
log $\Delta\phi_{sec}$	3.120574	subt.
log cot w =	$\overline{9}$.721031	

Then w = 62° 15' 11"

Calculate (slide rule adequate)

$C = \frac{\Delta\lambda}{2}(\sin\phi_m)$

$= (\frac{13'55"}{2}) \times .559$

$= 3' 54"$

Use w and C to calculate azimuths from following table:

Case 1 Northern Hemisphere
Sta E north of Sta W

Az at W is 90° − w − C
Az at E is 270° − w + C

Case 2 Northern Hemisphere
Sta E south of Sta W X

Az at W is 90° + w − C
Az at E is 270° + w + C

Case 3 Southern Hemisphere
Sta E north of Sta W

Az at W is 90° − w + C
Az at E is 270° − w − C

Case 4 Southern Hemisphere
Sta E south of Sta W

Az at W is 90° + w + C
Az at E is 270° + w − C

Distance Calculations

log cos ϕ_m	$\overline{9}$.917890	add	
log $\Delta\lambda_{sec}$	2.921686		
	2.839576		
log A_m (1)	$\overline{8}$.509267	subt.	
	4.330309		
log cos w	9.667983		
log S_{mtrs}	4.662326	subt.	
S_{mtrs}			45,954
	$\overline{6}$.793350		
log. .000621	1.455676	add	
log S_{miles}			
S_{miles}	28.555		

		90° 00' 00"
	± w	+ 62 15 11
	± C	− 3 54
Az at W =		152° 11' 17"
		270° 00' 00"
	± w	+ 62 15 11
	± C	+ 3 54
Az at E =		332° 19' 05"

Path length 28.56 miles 45.95 kms

(1) Log B_m/A_m & log A_m from attached table, for tabulated latitude nearest to ϕ_m.

Inverse Position Azimuth and Path Distance Calculation Sheet

		Latitude		Longitude
Station West	______	° ' "		° ' "
Station East	______	° ' "		° ' "
		° ' "		° ' "
Difference,	$\Delta\phi$		$\Delta\lambda$	
Convert to seconds	$\Delta\phi_{sec}$	______	$\Delta\lambda_{sec}$	______

$\phi_m = \phi_{smaller} + \frac{\Delta\phi_{sec}}{2}$ = ___° ___' ___" + ___' ___" = ___° ___' ___"

Azimuth Calculations

log B_m/A_m(1) | add
log cos ϕ_m | add
log $\Delta\lambda_{sec}$ ______
log $\Delta\phi_{sec}$ ______ | subt.
log cot w =

Then w = ___° ___' ___"

Calculate (slide rule adequate)

$C = \frac{\Delta\lambda}{2}$ (sin ϕ_m)

$= (\frac{\quad}{2})$ x ______

= ___' ___"

Use w and C to calculate azimuths from following table:

Case 1 Northern Hemisphere
Sta E north of Sta W

Az at W is 90° – w – C
Az at E is 270° – w + C

Case 2 Northern Hemisphere
Sta E south of Sta W

Az at W is 90° + w – C
Az at E is 270° + w + C

Case 3 Southern Hemisphere
Sta E north of Sta W

Az at W is 90° – w + C
Az at E is 270° w – C

Case 4 Southern Hemisphere
Sta E south of Sta W

Az at W is 90° + w + C
Az at E is 270° + w – C

Distance Calculations

log cos ϕ_m | add
log $\Delta\lambda_{sec}$ ______
log A_m (1) ______ | subt.
log cos w ______
log S_{mtrs} | subt.
S_{mtrs} ______
log. .000621 $\bar{6}.793350$ | add
log S_{miles}
S_{miles} ______

90° 00' 00"
± w
± C
______
Az at W = ___° ___' ___"

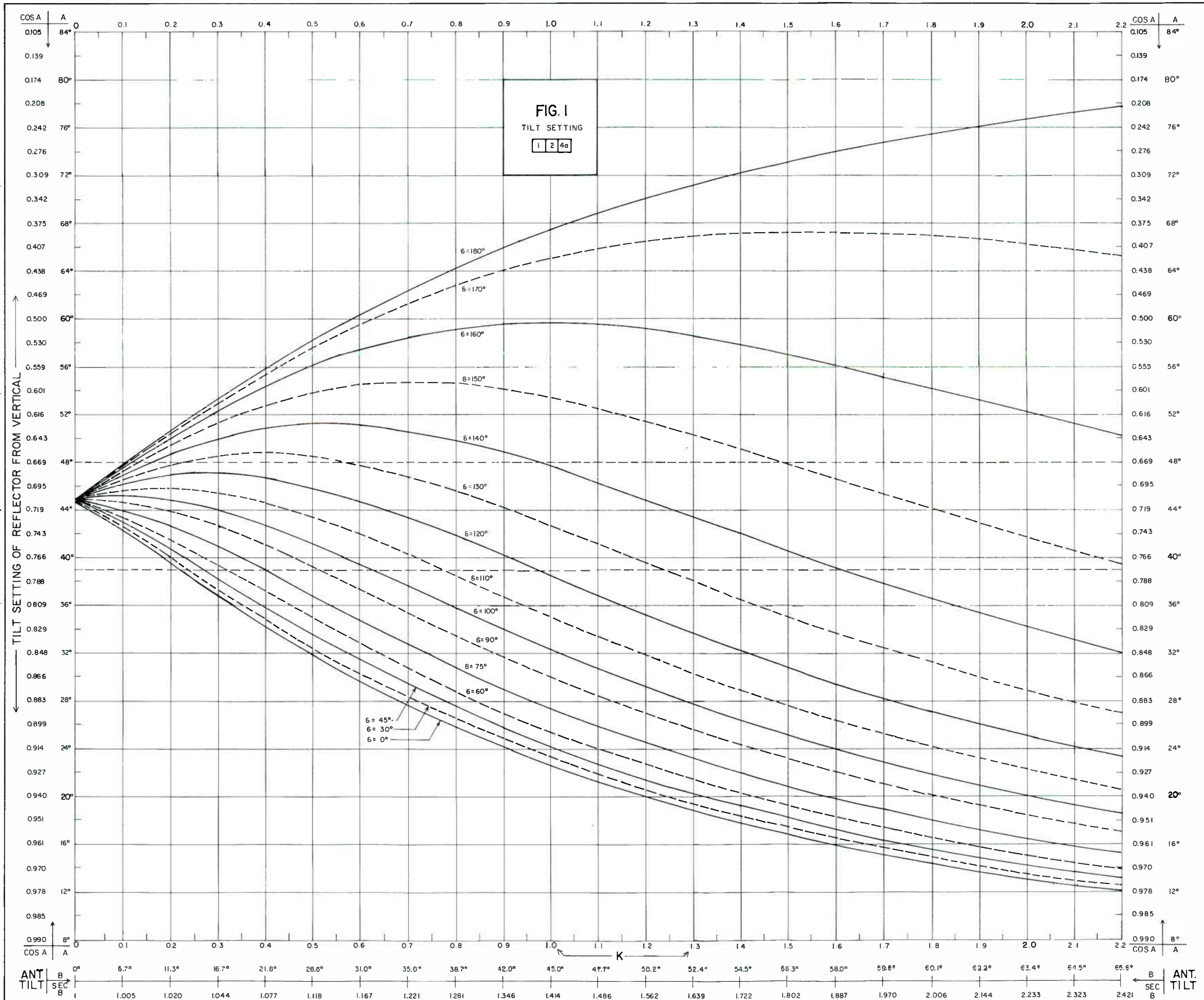
270° 00' 00"
± w
± C
______
Az at E = ___° ___' ___"

Path length ______ miles ______ kms

(1) Log B_m/A_m & log A_m from attached table, for tabulated latitude nearest to ϕ_m.

Latitude (degrees)	Log A_m	Log B_m/A_m	Latitude (degrees)	Log A_m	Log B_m/A_m
00	$\bar{8}$.509727	.002949	37	$\bar{8}$.509194	.001884
01	726	.002949	38	169	.001834
02	725	.002946	39	144	.001784
03	723	.002941	40	118	.001733
04	719	.002935	41	093	.001683
05	715	.002927	42	066	.001631
06	711	.002917	43	042	.001580
07	705	.002906	44	$\bar{8}$.509016	.001529
08	698	.002893	45	$\bar{8}$.508990	.001477
09	691	.002878	46	965	.001426
10	682	.002861	47	939	.001374
11	673	.002843	48	913	.001323
12	663	.002823	49	888	.001272
13	652	.002801	50	862	.001221
14	641	.002778	51	837	.001170
15	628	.002753	52	812	.001120
16	615	.002726	53	787	.001071
17	601	.002698	54	762	.001021
18	586	.002669	55	738	.000973
19	571	.002638	56	714	.000925
20	555	.002606	57	690	.000877
21	538	.002572	58	667	.000830
22	520	.002537	59	644	.000784
23	502	.002501	60	621	.000739
24	483	.002463	61	599	.000695
25	464	.002424	62	578	.000652
26	444	.002384	63	557	.000610
27	423	.002343	64	536	.000568
28	402	.002301	65	516	.000528
29	381	.002258	66	496	.000489
30	359	.002214	67	478	.000452
31	336	.002169	68	459	.000415
32	313	.002123	69	442	.000380
33	290	.002077	70	425	.000346
34	267	.002029	71	409	.000313
35	243	.001981	72	393	.000282
36	218	.001933			

Computation Factors Extracted From Special Publication No. 8



DISTR
L-M23

EED-20020

REVISIONS

SYM	ZONE	DESCRIPTION	DATE	APPROVED

NOTES:

1. THE DEFINITION OF SYMBOLS USED ON THIS DRAWING ARE AS FOLLOWS, REFER TO DETAILS A AND B:

   A = ANGLE BETWEEN THE REFLECTOR FACE AND THE VERTICAL SUPPORTING STRUCTURE (USUALLY A TOWER).
   B = ANGLE BETWEEN A HORIZONTAL LINE PERPENDICULAR TO THE REFLECTOR'S VERTICAL SUPPORTING STRUCTURE AND THE LINE PERPENDICULAR TO THE ANTENNA FEEDHORN.
   D = DISTANCE BETWEEN ANTENNA AND REFLECTOR.
   δ = DELTA. ANGLE BETWEEN R AND P'.
   ΔØ = DELTA PHI. THE CHANGE IN THE POLARIZATION ANGLE CAUSED BY REFLECTING THE RADIO BEAM. ΔØ IS A CORRECTION FACTOR.
   EA = EFFECTIVE AREA OF A REFLECTOR.
   H = VERTICAL SEPARATION BETWEEN THE CENTER OF THE REFLECTOR AND THE CENTER OF THE ANTENNA.
   K = RATIO OF R TO H. NB: R AND H MUST BE EXPRESSED IN THE SAME UNITS.
   N = THE NORMAL TO THE REFLECTOR'S FACE.
   N' = THE PROJECTION OF THE REFLECTOR NORMAL ONTO THE GROUND.
   P' = THE PROJECTION OF THE RADIO PATH AZIMUTH ONTO THE GROUND.
   R = HORIZONTAL SEPARATION BETWEEN THE CENTER OF THE REFLECTOR AND THE CENTER OF THE ANTENNA.
   θ = ANGLE BETWEEN THE NORMAL OF THE REFLECTOR AND THE SITE'S RADIO PATH AZIMUTH, I.E., N' AND P'. θ IS A CORRECTION FACTOR.

2. TO USE THE THREE FIGURES ON THIS DRAWING, FIRST OBTAIN THE VALUE OF δ, H AND R FROM THE SITE PLOT PLAN, NEXT DETERMINE K (K = R/H). INTERPOLATE AS REQUIRED FOR VALUES OF δ AND K NOT PLOTTED ON THE DRAWING. THEN,

   a. FROM THE DETERMINED VALUE OF K, MOVE VERTICALLY UPWARD TO THE APPROPRIATE VALUE OF δ.
   b. FROM THE INTERSECTION OF K AND δ, MOVE HORIZONTALLY TO READ THE REQUIRED VALUE OF A, θ OR ΔØ, FIGURES 1, 2 OR 3 RESPECTIVELY.
   c. FROM THE DETERMINED VALUE OF K, MOVE VERTICALLY DOWNWARD TO READ THE APPROPRIATE VALUE OF B.

3. ON FIGURE 3, SCALE X IS ASSOCIATED WITH THE SOLID δ CURVES; SCALE Y IS ASSOCIATED WITH THE DASHED δ CURVES.

4. a. THE VALUES OF A AND B, REFLECTOR AND ANTENNA TILT ANGLES, SHALL BE SHOWN ON THE SITE PLOT PLAN BY THE ASSOCIATED ITEM.
   b. FROM THE VALUE OF θ, DETERMINE THE AZIMUTH OF N AS FOLLOWS:
      SUBTRACT θ FROM THE AZIMUTH OF P' IF THE ANTENNA IS TO THE RIGHT OF P' AS SHOWN IN DETAIL B.
      ADD θ TO THE AZIMUTH OF P' IF THE ANTENNA IS TO THE LEFT OF P'.
   c. AFTER ESTABLISHING THE PROPER POLARIZATION SETTING FOR THE ANTENNA FEEDHORN PER TEM 13 (PAGE 5 AND FIGURE 6), ROTATE THE ANTENNA FEEDHORN, AS VIEWED FROM THE REAR OF THE ANTENNA, BY ΔØ AS FOLLOWS:
      CCW IF THE ANTENNA IS TO THE RIGHT OF P' AS SHOWN IN DETAIL B.
      CW IF THE ANTENNA IS TO THE LEFT OF P'.

5. TO DETERMINE THE GAIN OF A PARTICULAR ANTENNA-REFLECTOR COMBINATION, THE EFFECTIVE AREA (EA) OF THE REFLECTOR AND THE SEPARATION (D) BETWEEN THE ANTENNA AND THE REFLECTOR MUST BE KNOWN.
   a. EA = ACTUAL AREA OF REFLECTOR X COS θ X COS A
   b. D = H X SECANT B

REF: EEM-40, ISSUE 3 OR LATER
TEM-13

INITIAL USE:		CKD	DFTG: AYN	DATE 2-3-70
			APPL: RAC	
ENGINEER R.A. COMAS	DATE 2-3-70	APPD	STDS:	
DRAFTSMAN R.R. LUTAP			TRANS:	
			APPL: RAC	2-3-70
		DRAWING RELEASE		

MICROWAVE
ANTENNA/REFLECTOR
TILT, AZIMUTH & CHANGE IN POLARIZATION

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DWG ISSUE: 1 | SHEET 1 OF 3

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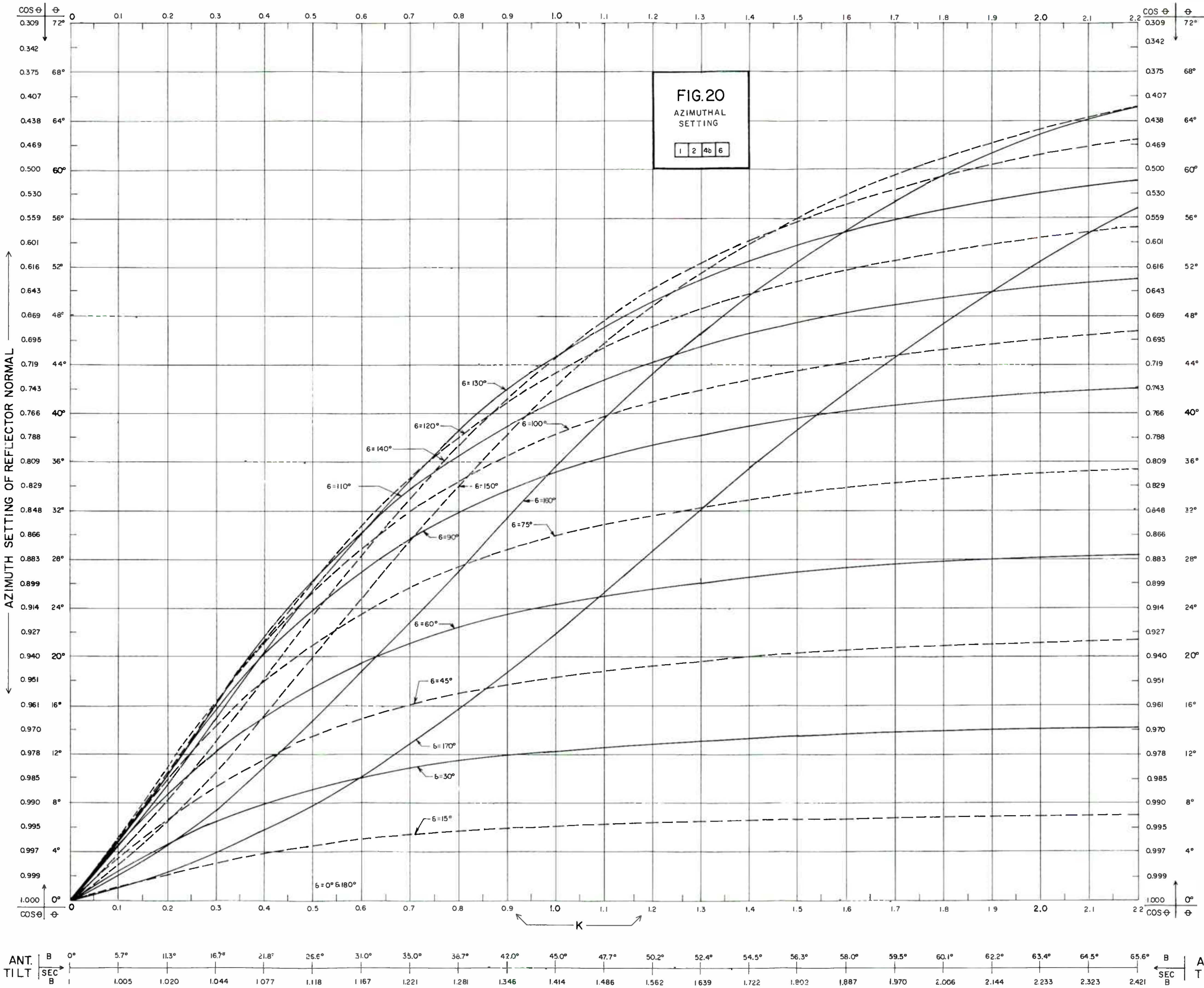

INITIAL USE:		CKD	DFTG: AYN	DATE
			APPL: RAC	2-3-70
ENGINEER R. A. COMAS	DATE 2-3-70	APPD	STDS:	
			TRANS:	
DRAFTSMAN R. R. LUTAP			APPL: RAC	2-3-70
		DRAWING RELEASE		

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ANTENNA/REFLECTOR
TILT, AZIMUTH & CHANGE IN POLARIZATION

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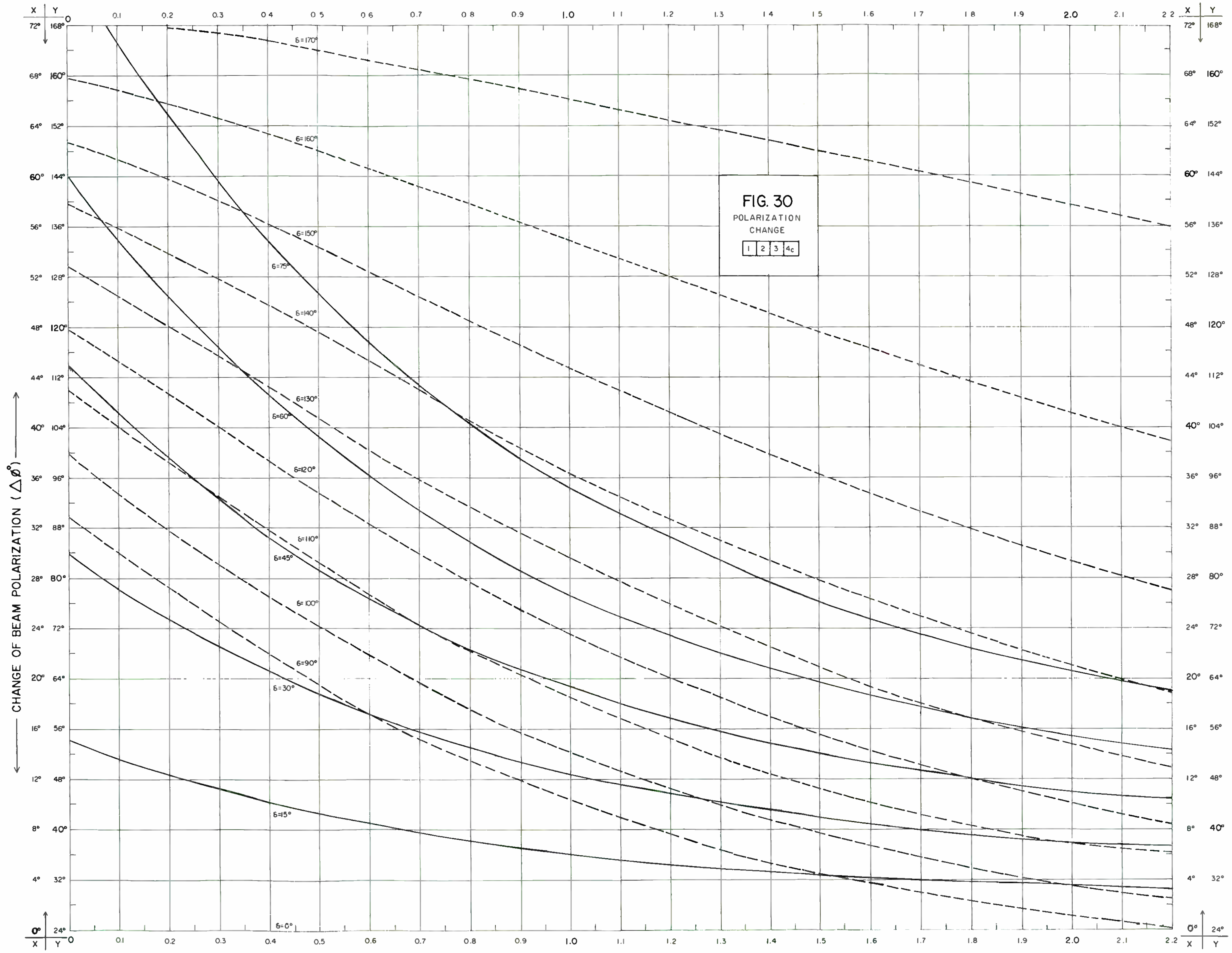
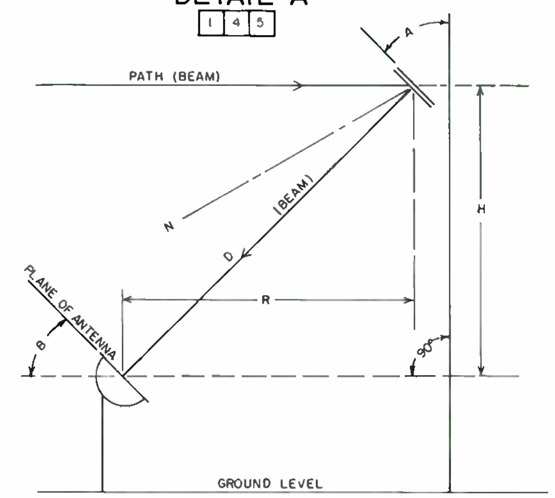
SYM	ZONE	DESCRIPTION	DATE	APPROVED



ANT. TILT																							
B	0°	5.7°	11.3°	16.7°	21.8°	26.6°	31.0°	35.0°	38.7°	42.0°	45.0°	47.7°	50.2°	52.4°	54.5°	56.3°	58.0°	59.5°	60.1°	62.2°	63.4°	64.5°	65.6°
SEC B	1	1.005	1.020	1.044	1.077	1.118	1.167	1.221	1.281	1.346	1.414	1.486	1.562	1.639	1.722	1.802	1.887	1.970	2.006	2.144	2.233	2.323	2.421

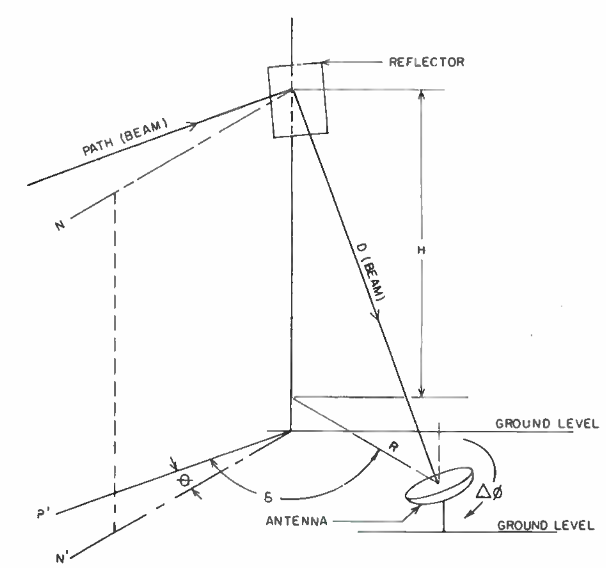
DETAIL A

1 4 5



DETAIL B

1 4 5



INITIAL USE:		CKD	DFTG: AYN	DATE
			APPL: RAC	2-3-70
ENGINEER R. A. COMAS	DATE 2-3-70	APPD	STDS:	
DRAFTSMAN R. R. LUTAP			TRANS:	
			APPL: RAC	2-3-70
			DRAWING RELEASE	

MICROWAVE
ANTENNA/REFLECTOR
TILT, AZIMUTH & CHANGE IN POLARIZATION

CODE IDENT NO.	SIZE	
83744	D	EED-20020

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CUT ALONG THIS LINE TO REMOVE.

Figure 4. Equivalent Earth Profile Curves

0—1069—3