NASr-21(02)

MEMORANDUM
RM-5785-NASA
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THE TECHNOLOGY POTENTIALS FOR SATELLITE SPACING AND FREQUENCY SHARING

J. L. Hult, S. J. Dudzinsky, N. E. Feldman,
J. D. Mallett, N. C. Ostrander and E. E. Reinhart

PREPARED FOR:

NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

The RAND Corporation
SANTA MONICA • CALIFORNIA

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This research is sponsored by the National Aeronautics and Space Administration under Contract No. NASr-21. This report does not necessarily represent the views of the National Aeronautics and Space Administration.

The RAND Corporation
1700 MAIN ST. • SANTA MONICA • CALIFORNIA • 90406

Published by The RAND Corporation

PREFACE

This Memorandum reports the results of one phase of a continuing study of the technological potential of communication satellites. It is hoped that documentation at this time will make a timely contribution to important pending decisions about the use of orbital locations and frequencies for communication relay.

This work represents the efforts of a number of contributors. The material in Section II and a companion publication, RM-5786-NASA, Radio Relay System Performance in an Interference Environment, were authored by E. E. Reinhart, with contributions and assistance from C. R. Lindholm and E. Bedrosian. Appendix A and the material on antennas are due to J. D. Mallett, with contributions from C. V. Baker on conventional antennas and W. Doyle on adaptive arrays. S. J. Dudzinsky provided the material on polarization discrimination and Appendix C on propagation factors and precipitation scatter interference. N. C. Ostrander provided the material in Appendix B on earth-to-satellite geometry factors, and N. E. Feldman, the material in Appendix D on frequency dependence of low-noise and output devices. J. L. Hult was responsible for the remaining material and the coordination of the study.

SUMMARY

SCOPE

This Memorandum describes in a fresh and reasonably comprehensive way the important factors influencing satellite spacing and frequency sharing. In particular, it derives the system parameter constraints that could improve the compatibility of and benefits for satellite and microwave relay systems sharing a common spectrum. Three types of stations must be considered in analyzing such spectrum sharing: terrestrial microwave repeaters and terminals, satellite repeaters, and earth-station terminals. The receivers at each of these three types of stations can be interfered with directly by the transmitters of each type. Thus there is a total of nine possible direct interference modes for this sharing situation. In addition, there are two potentially important scatter interference modes that involve the intersection of the main beam of an earth station with the main beam of a microwave relay station. If rain or other scatterers are contained within the common volume of intersection, scatter interference may be experienced either in the mode involving scatter of earth-station transmissions to a microwave relay receiver or in the reverse mode.

Some of the "possible" interference modes would not be experienced in current operations because of the separation between the frequency bands used to transmit up to satellites and those used to transmit from satellites down to earth. However, the possibility of doubling the equivalent spectrum available makes attractive the use of some portions of the spectrum for both up and down links to satellites. For this reason the potential for this type of usage is included in this Memorandum.

The first part of this Memorandum outlines the technical and technological background needed to derive suitable specifications for high-performance system design, which includes satisfactory system compatibility for spectrum sharing. If the reader is not interested in the details of the technical background, he can turn directly to the part of the report that treats satellite spacing and frequency sharing.

SUMMARY OF TECHNICAL BACKGROUND

In order to establish a quantitative measure of the compatibility of systems sharing the use of a common spectrum, it is necessary to:

- establish a quantitative measure of output message quality
- relate this to the receiver input signal parameters
- quantify the effects of antenna, propagation, and geometric factors

OUTPUT MESSAGE QUALITY

Communication messages may be categorized into two general classes: analog, e.g., voice; and digital, e.g., teletype. This should not be confused with the fact that either type of message may be communicated using either analog or digital types of modulation. For analog messages the receiver output message quality objectives may be established by subjective determinations and specified in terms of output signal-to-noise ratios; for digital messages the output quality may be measured by error probabilities. Usually the output message quality specifications will be expressed in terms of limits to be maintained for selected percentages of time. The most common unwanted signal is white thermal noise, which is used in this report as a reference unwanted signal with which the effects of other types of unwanted signal may be compared.

RECEIVER TRANSFER IMPROVEMENT

In transferring the wanted and unwanted signals from the input of a receiver to its output, a number of operations may be performed on them, including demodulation, demultiplexing, and a variety of signal-processing operations. The improvement of the wanted relative to the unwanted signals in transferring through the receiver can be described by the ratio of an appropriate measure of output message quality to the input wanted-to-unwanted-signal ratio C/X. For analog modulation systems this transfer improvement is referred to as the receiver transfer characteristic (RTC). It should be distinguished from the "improvement factor" commonly used to compare the performance of frequency relative to amplitude modulation.

BANDWIDTH EXPANSION RATIO (W/B)

An important parameter influencing the transfer improvement and the spectrum cost of transmitting information is the bandwidth expansion ratio W/B, which is the ratio of the transmitting spectrum bandwidth W to the bandwidth B of the information that is transmitted. Figure i illustrates how the receiver transfer improvement varies with W/B in the presence of the reference white-noise type of unwanted signal for some representative types of messages and modulations. Single-sideband suppressed carrier amplitude modulation (SSB) is the base case with W/B = 1. The curves for frequency modulation (FM) provide a continuous range of RTC values as a function of design choice for W/B. At the cost of using more transmitting bandwidth the total transmitting power and the effects of interference or noise can be significantly reduced by increasing W/B as long as this does not result in reducing C/X below the FM threshold of about 10 dB. For both SSB and FM, there is a linear relationship between the output quality and C/X. An increase by some factor in the input ratio will appear as an increase in the output quality by the same factor. This is not the case for digital modulation methods. The points for noncoherent phase-shift keying (NCPSK) illustrate the performance of a typical high-quality pulse code modulation (PCM) system. (They represent four-level NCPSK with 7, 8, and 9 bits per PCM sample.) The receiver transfer improvement indicated is that obtained with the input at threshold. Any increase in C/X above threshold will cause only a slight increase in output quality, which (if C/X is above threshold) is almost completely determined by quantization noise.

UNWANTED SIGNALS DIFFERENT FROM THERMAL NOISE

When the unwanted signal has a different spectral distribution from thermal noise, the transfer improvement may differ from that illustrated in Fig. i. This is especially pertinent for interference from systems with significant bandwidth expansion that share a common band. Such systems typically have greater power concentration in the center of their channels so they may prove more effective as interference

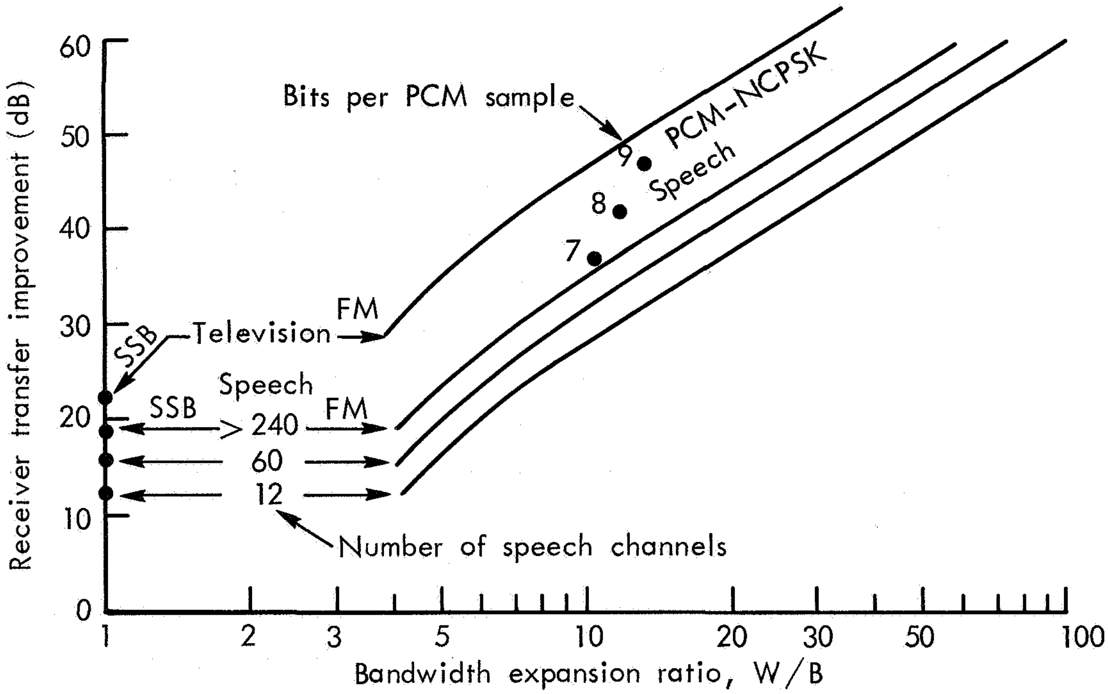

Fig. i — Receiver transfer improvement with white-noise interference

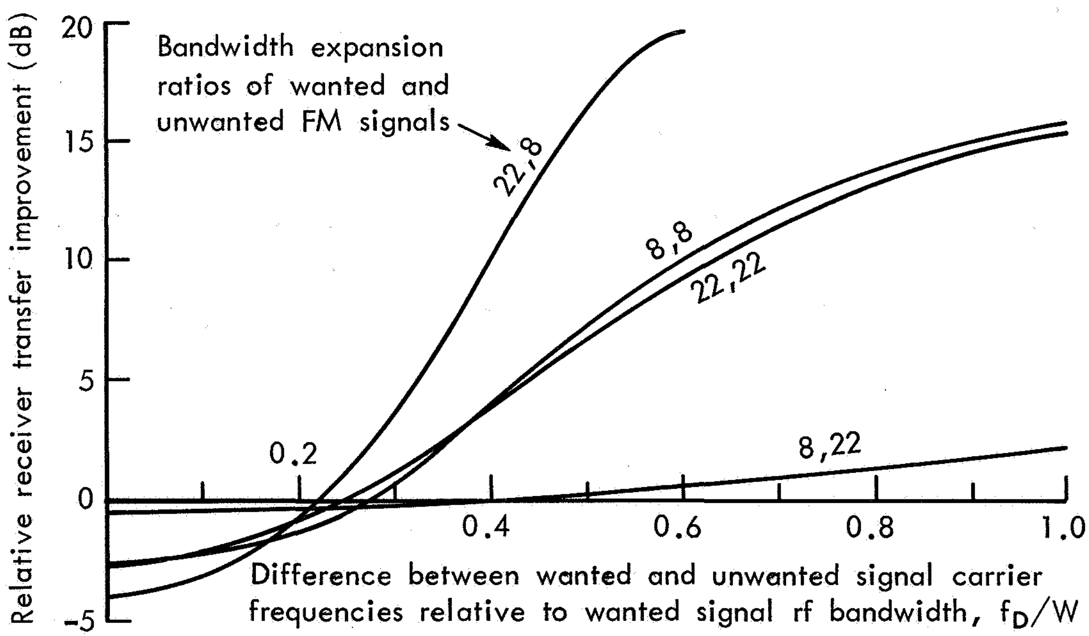

Fig. ii — Receiver transfer improvement for interference between wideband FM signals relative to that with white-noise interference than an equal thermal-noise power for co-channel operation, but less effective if the band centers are displaced from each other. This is illustrated for wideband FM signals in Fig. ii. If interleaving greater than 0.25 of the wanted transmitting bandwidth is used, transfer improvement greater than for the reference thermal-noise case is obtained. Thus if interleaving of bandwidths is employed, the reference transfer improvement is a conservative assumption; it will be used in all further considerations in this Memorandum.

MULTIPLE LINKS AND NOISE ENTRIES

Communication relay systems may consist of many links in tandem (a transcontinental microwave chain may have more than 100 repeaters in series). Unwanted signals can enter anywhere in such a system, and each link will usually include a variety of different unwanted signals. The wanted and unwanted signals in any receiver are transmitted to the next receiver and combined with other unwanted signal entries, etc., until the terminal receiver is reached. The way in which the unwanted signals accumulate depends on the type of repeaters and the details of the environment for each link. However, if a given type of interference can make multiple entries into a relay system, the tolerance per entry is obviously reduced. In considering interference to terrestrial microwave relay systems in this report, the conservative assumption is made that there are 140 entries for each type of interference. This might be approached under special circumstances for extreme communication distances; however, for many situations it provides as much as an additional 20 dB of margin for operation against the interference under consideration.

MARGINS AND PROTECTION RATIOS

If a relay system is designed to deliver the desired message quality under ideal conditions, an ideal required C/X is obtained. When appropriate system margins have been added to protect the output message quality against fading or attenuation of the wanted signal and any other degradation of C/X that the system is designed to accommodate, the required C/X including margins is called a protection ratio. The total output noise due to all types of unwanted signals is usually partitioned and allocated to the various types of unwanted signals and their numbers of entries. A typical terrestrial microwave relay system design might allocate ten percent of the total unwanted signal to interference from earth stations and satellites. With the conservative assumption that there may be as many as 140 entries in the transmission of an extremely long-distance message, typical protection ratios for this type of interference range from 70 dB (for the TH system) through 58 dB (for the TD-2 system) to 50 dB (for a hypothetical system with greater W/B). For satellite relays and earth stations where larger W/B is used, typical protection ratios against one mode of interference might range from 20 dB to 40 dB. The value assumed in this Memorandum for determining parameter constraints for compatible spectrum sharing is 30 dB, which is about 10 dB greater than is needed for large W/B and conservative margin allowances.

ANTENNA PATTERNS

The orbital spacings that can be used between satellites when they share a common spectrum will be strongly influenced by the earth-station antenna patterns as determined by their size (diameter to wavelength ratio D/λ) and illumination. In order to illustrate the influence of typical antennas on satellite spacing and system design, some reference antenna patterns were selected, two of which are illustrated in Fig. iii. The parabolic antenna represents what might be expected for the average value of the sidelobes from typical high-quality production design of this type of antenna. Similarly the "uniform" illumination circular array represents typical performance for this antenna type. In order to specify the effects of both D/λ and the illumination with a single measure, an equivalent uniform size $(D/\lambda)_u$ is introduced. This is the equivalent uniform illumination circular aperture that will permit the same satellite spacing performance as the antenna in question.

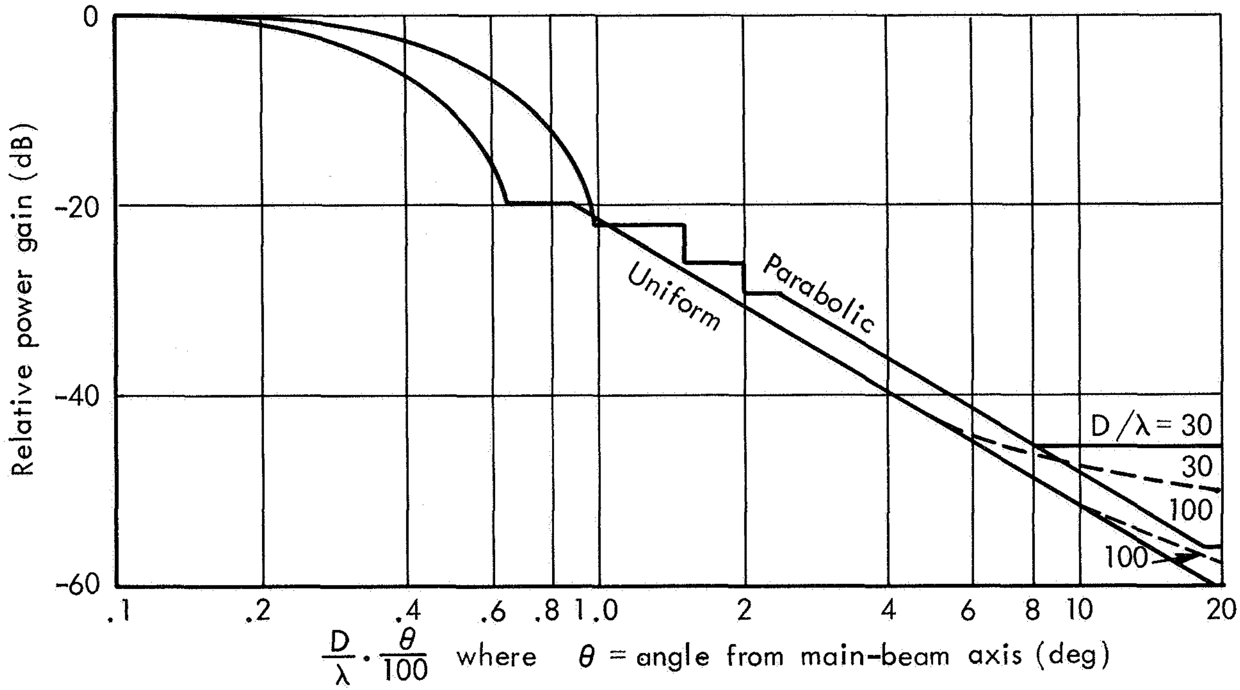

Fig. iii—Antenna model patterns

ADAPTIVE ARRAYS

In addition to the fixed patterns illustrated in Fig. iii, an attractive future possibility is the adaptive or adapted array for which the phasing and illumination functions are adjusted to minimize the total interference received from unwanted directions. The outstanding discrimination performance that appears possible with relatively small adaptive apertures compared with conventional antennas might eventually completely change system design for satellite spacing and frequency sharing. Adaptive arrays offer particular promise: for earth terminals requiring links to many different satellites simultaneously, for small antennas where size or cost limits the number of elements that can be used and where strong rejection of a small number of interfering signals is needed, and for microwave relays to adaptively couple two antennas in height diversity to limit the depth of multipath fading.

However, until other antennas have been replaced, system design must be based on the weakest antenna that has to be protected in service. The specifications that will be derived in this Memorandum for compatible sharing will assume the use of conventional antennas.

POLARIZATION DISCRIMINATION

It appears feasible in practice for satellite relay systems to provide polarization discrimination of 20 to 30 dB (or more with adaption) in the antenna main beam. Thus if the systems are designed to operate with smaller C/X than the polarization discrimination, it should be possible to double the effective spectrum available by using it independently on two orthogonal polarizations.

EARTH-TO-SATELLITE GEOMETRY

A reference earth-to-satellite geometry model is used for interference calculations. It consists of an earth station at 45° latitude linked with a wanted equatorial synchronous satellite on the same longitude and a full orbit of unwanted synchronous satellites at equal longitudinal spacings on both sides of the wanted satellite. If the earth station is at a different latitude, or if the wanted satellite is at a different longitude, there will be a change in the relative ranges and apparent angular separations between the satellites, and the influence of the horizon cutoffs will differ. However, these variations can be accommodated with a 1-dB margin in the protection ratios for interference between earth stations and satellites.

REFERENCE GROUND-ENVIRONMENT MODELS

In order to test how the various system parameters influence interference, it is necessary to use a suitable reference ground-environment model. The model chosen was that of maximum theoretical density of ground stations for a given minimum separation between stations uniformly distributed on the earth's surface. This model is a conservative limiting case for a given minimum station separation, and the interference that might be experienced in practice would only approach that predicted if the density of sharing approached that of the model.

SUMMARY OF COMPATIBLE SATELLITE SPACING AND FREQUENCY SHARING

Using the technical background that has been outlined, it is possible to derive the joint constraints imposed on the various system parameters for tolerable interference and to define the technological potentials for satellite spacing and frequency sharing.

INTERFERENCE FROM SATELLITES TO EARTH STATIONS, MODE 1

This is the mode that is examined in greatest detail. The other modes are then more easily treated.

Effects of Earth-station Antenna Patterns

If all the satellites are equally spaced and have equal equivalent isotropically radiated power (EIRP) in any bandwidth in the direction of an earth station, the required satellite spacing to isolate them for independent use of a common spectrum by the earth-station receiving antennas is primarily determined by these antenna sidelobe patterns. Using the previously described reference earth-satellite geometry model and antenna patterns, the curves of X/C versus required satellite spacing σ of Fig. iv are obtained. When the antenna patterns themselves fall off rapidly with angle, the closest interfering satellites tend to dominate the total interference, and X/C falls off approximately as the antenna pattern. In Fig. iv, X/C is approximately proportional to $1/\sigma^3$ and the interference contribution of all the satellites beyond the nearest neighbors is less than 0.2 of the total for uniform spacing and EIRP. The importance of the antenna size is obvious from Fig. iv, since $\sigma \propto \lambda/D$.

Tradeoffs Involving W/B

With the help of Figs. i and iv it can be shown to be advantageous for both total power and information capacity to increase W/B (and the number of satellites) until threshold is reached or until the satellite spacing crowds the mainlobe of the antenna. Smaller required C/X values would also permit greater multiplication of the independent beams possible with directive antennas on board the satellites. However,

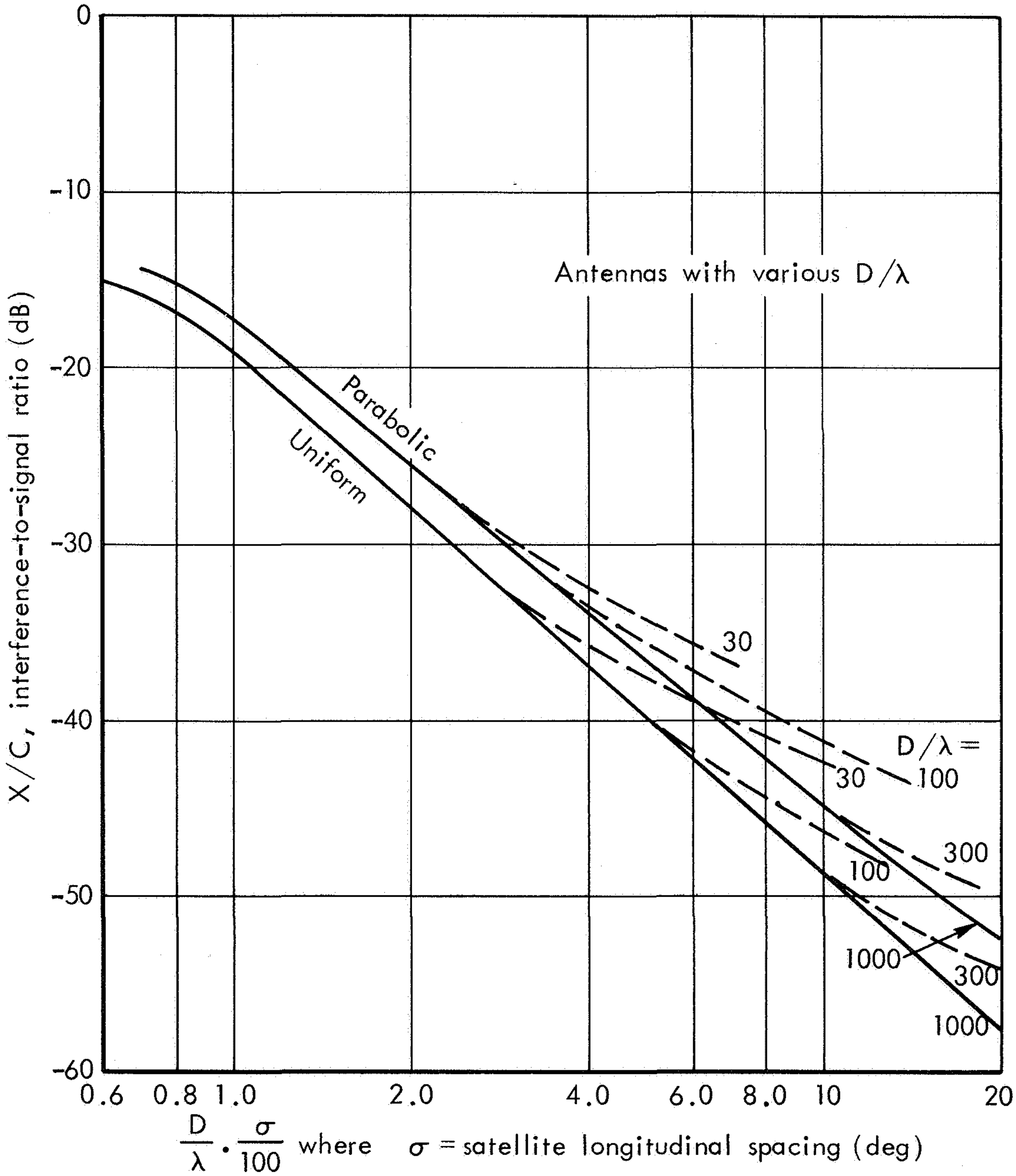

Fig. iv—X/C versus satellite spacing

some of these possible increases in total system capacity can be achieved only at greater cost per circuit. It is important, therefore, in establishing satellite systems to use large enough values for parameters such as W/B and D/λ that will permit the orbital capacity growth that may be needed during the projected life of the system.

Differences in Satellite EIRPs

For this mode of interference, only the relative values of EIRPs are involved in determining X/C, i.e., X/C and σ are independent of the absolute values of the satellite EIRPs. If the EIRPs of the satellites differ, the exact solution for the minimum orbital spacings becomes very complicated. However, as a rough rule the spacing $\sigma_{a,b}$ required between any two adjacent satellites "a" and "b" will increase over σ_o, the spacing required when these parameters are the same, with any differences in EIRP or earth-station antenna size. An important factor in determining the spacing $\sigma_{a,b}$ is the spacing quotient $q_{a/b}$, i.e., $\sigma_{a,b} \propto 1/q_{a/b}$ where $q_{a/b}$ is proportional to the cube root of the ratio of the EIRPs that is less than unity, i.e., that is protecting the weaker system. Thus if a number of satellites with different spacing quotients are to be mixed, the most efficient use of orbit space would order them to minimize the number and the differences of adjacent satellites with different spacing quotients. As a rough guide, a doubling of the spacing between adjacent satellites is needed for each 9 dB of difference in their EIRPs.

Effects of Satellite Antenna Patterns

Previous discussion indicated how earth-station antennas in conjunction with satellite spacing could separate or isolate links so that a given spectrum could be used independently many times (each time with a different satellite) to the same or different earth locations. A similar type of link isolation can be provided with larger and more directive antennas on board the satellites. This may be used to augment the isolation provided by the earth-station antennas, or to provide enough isolation to earth stations at different locations so that the same spectrum can be used independently many times, in this case from the same or different satellites. Since the isolations by the earth-station antennas and by the satellite antennas can be made independently, the possible multiplication of the spectrum use is a product of the two. In addition, the larger satellite apertures enable a given EIRP from the satellite with less transmitted power, and allow the up link to operate with less earth-station EIRP and interference potential to terrestrial systems.

Margins Required

It is necessary to add a variety of margin factors to the ideal C/X required at the receiver input in order to obtain a satisfactory protection ratio against this interference mode. A 1-dB margin would accommodate independent longitudinal variations about assigned positions of 8 percent of the nominal spacing. This should be adequate for this factor including the effects of any uncertainties in antenna pointing. A 3-dB margin is probably needed for variations in the satellite antenna gain over the designed area of coverage. A 2-dB margin is probably needed for differences in satellite power output because of aging or component differences. At least 2 dB of margin is required to take care of the variations in earth-satellite geometry factors previously discussed, the probability of interference from sidelobe peaks contributing more than the average, and the effects of differences in the attenuation and fading of the wanted compared with the unwanted signals. Thus a total margin of at least 8 dB seems appropriate to include in the protection ratio for this mode of interference.

Coordination Benefits

Coordination of earth-station receiving antenna size is especially important. Establishing some minimum $(D/\lambda)_u$ and making it as large as possible is important in determining the minimum satellite spacing achievable and consequently the total information capacity of a given spectrum and orbit space for satellite relays. The modulation properties of the signals should be coordinated for greatest benefit. This would involve using comparable bandwidth expansions and interleaving the frequencies between adjacent satellites. Also, increasing W/B to enable the use of the smallest practical protection ratios will permit closer satellite spacings and more independent beams from the satellites, thus increasing the total information capacity of satellite relays for a given spectrum and orbit space. In addition, the opportunity exists for two independent systems and allocations on two independent polarizations if the systems are designed for required C/X smaller than 20 to 30 dB.

The relative EIRPs of adjacent satellites is an important factor in the spacing required between them. If their locations and EIRPs can be coordinated to minimize the number and the differences of adjacent satellites with different EIRPs, the most efficient use of orbit space will be obtained.

INTERFERENCE FROM EARTH STATIONS TO SATELLITES, MODE 2

Most of the characteristics of Mode 1 also have application to this mode. Only the differences will be summarized here. Antenna patterns again play a prominent role; however Mode 2 involves the ratio of the gains of separate earth-station transmitting antennas, whereas Mode 1 involved the ratios of the gains of the same earth-station receiving antenna in the various directions of the satellites. This distinction will probably be of most concern in establishing margins for propagation attenuation. Since the propagation paths in the earth's atmosphere may be quite different for the illuminations of adjacent satellites in this case, the full magnitude of the possible variations in propagation attenuation must be added as a margin in the protection ratio for Mode 2. This may mean that a significantly larger $(D/\lambda)_u$ will be required in the up link than in the down link if the satellite spacing is not to be limited by the up link at frequencies above 10 GHz.

It will be beneficial to coordinate earth transmitting stations to establish some minimum $(D/\lambda)_u$ and to make this minimum large enough so that the up link does not limit the satellite spacing achievable. The EIRP per unit of bandwidth for all earth-station transmitters using the same satellite transponder should be equalized. For frequencies above 10 GHz it may not be feasible to use two polarizations independently on the same frequency for the up link unless they emanate from the same earth station. The coordination that will be beneficial for other system parameters are similar to those for Mode 1 and the two modes should be made compatible.

OTHER DIRECT INTERFERENCE MODES

The seven other direct interference modes are less complicated and can be summarized very briefly.

Satellite-to-satellite Interference, Mode 3

This mode can be significant only if the same frequency bands are used for both up links and down links. Such operations may be practical for some future applications and could double the bandwidth available in prime portions of the spectrum. It would be relatively straightforward to design the satellite antennas so that this mode of interference need not limit operations of this type.

Interference Between Earth Stations, Mode 4

Again, this mode is of no consequence unless the same frequency bands are used for both up links and down links. It can be shown that this mode of interference could be tolerated within the constraints that other interference modes will impose anyway for reducing earth-station EIRP and increasing satellite aperture and EIRP. However, as is the case for some of the other interference modes, there is a restriction to some minimum earth-station spacing for compatible applications because of this type of interference. For mobile and other applications where it may not be possible to limit or restrict the minimum transmitter and receiver spacings, it may not prove feasible to operate with up links and down links in the same frequency bands.

Interference from Terrestrial to Earth Stations, Mode 5

This interference mode is very similar to Mode 4. Even if the earth station is imbedded in a maximum density matrix of microwave relay stations with a nominal minimum spacing of 40 km, no new constraints are imposed on the systems over those of Mode 4.

Interference Between Terrestrial Microwave Relays, Mode 6

This mode involves only the microwave relays, and most common carrier practice employs adequate antenna directivity to cope conservatively with this interference mode for the maximum theoretical station density distribution for conventional minimum spacings between stations.

Interference from Earth to Terrestrial Stations, Mode 7

This mode of interference in a maximum theoretical station density environment would primarily constrain the earth-station transmitting power to within about 20 dB of the transmitting power of the microwave relays in the same frequency band.

Interference from Satellites to Terrestrial Stations, Mode 8

This mode is most easily controlled by constraining the directions of pointing microwave relay antennas. If these antennas are sited so that the sum of their gains in the directions of all the satellites radiating in their frequency bands is less than 10, they could tolerate significantly greater satellite EIRP than is currently recommended by the CCIR. For these conditions the TD-2 system could tolerate satellite EIRPs of 75 dBW in 500 MHz, and for the TH system 80 dBW in 500 MHz. Current CCIR recommendations provide a limit equivalent to 62 to 68 dBW for the EIRP in 500 MHz.

Interference from Terrestrial Stations to Satellites, Mode 9

This mode also is most easily controlled by constraining the directions of pointing microwave relay antennas. If these antennas are sited so that the average value of all their transmitting antenna gains in the direction of a satellite is less than isotropic, up to 10^5 microwave relays could be within view of the satellite without requiring the earth-station transmitting power to exceed by more than 20 dB that of a microwave relay in the same band. With larger satellite receiving apertures limiting the field of view of the earth, the interference from the maximum theoretical density of microwave relays with conventional minimum spacings could be tolerated with earth-station transmitting powers comparable with those of the microwave relays.

SCATTER MODES OF INTERFERENCE

The precipitation scatter modes of interference are the most difficult to evaluate as to constraints on system design to meet performance objectives for satellite and microwave relay systems sharing the same frequency bands. The principal reason for this is that the expected spatial distribution of precipitation and its statistics for the localities of interest are essentially unknown, particularly at the higher weather altitudes.

The potentially most important geometry for scatter interference between microwave relay and earth stations is for coupling through scatterers in the common volume of main-beam intersection of the two systems. If a maximum theoretical density of microwave relays is assumed with a spacing of 40 km between stations, a limiting potential for this type of interference is obtained. In this environment it would probably be feasible to coordinate site locations and beam orientations to avoid main-beam intersections within two station spacings separation and below 1-km altitude. Beyond these ranges and altitudes the coordination to avoid intersections soon becomes impractical. Even so, there appears to be enough leeway in feasible satellite aperture and allowable EIRP to design around the scatter interference for the densest sharing environment with 40-km spacing between microwave relays. However, experimental data are needed to better establish the design specifications that will ensure the compatible sharing of the spectrum as constrained by scatter interference in a particular location.

COMBINED CONSTRAINTS FROM ALL MODES

Ignoring the economic factors and considering only technological constraints from all modes indicate that most of the burden of accommodation to an intense sharing environment (between microwave relays and satellite relays with both up links and down links in the same frequency bands) rests on the design of the satellite relay system. The indicated directions of change to relieve the interference problems (increased satellite aperture and EIRP, increased bandwidth expansion W/B, and earth-station antennas as large as is practical with decreased transmitting power) also tend to increase the total system capacity and should therefore be logical long-term goals anyway.

The single important added constraint imposed on the microwave relays is that of pointing coordination to avoid the directions of synchronous orbits. The coordination indicated seems about the minimum acceptable and could be evolved gradually with the buildup of all the systems. It does not seem to be an excessive requirement for the great benefits that could be derived with the coordinated sharing of the spectrum. This pointing coordination should not be hard to accommodate at low through mid-latitudes. However, at latitudes above 60° the excluded directions become so large that it may not be feasible to implement a specific requirement for microwave relays with the coordination restrictions. In these regions it may be necessary to choose either microwave <u>or</u> satellite relays for a system operating at the shared frequencies.

FREQUENCY DEPENDENCE OF LOW-NOISE AND OUTPUT DEVICES

The current hardware for low-noise and power output applications rapidly degrades with increasing frequency above about 10 GHz with respect to performance and availability, and the costs increase sharply. However, there does not appear to be any fundamental reason why development effort and production demand could not provide hardware with only a modest dependence on frequency up through 100 GHz. Thus eventually neither availability, nor performance, nor cost of the hardware is apt to be decisive in the choice of frequencies in the range up through 100 GHz for advanced communication satellite systems.

REQUIREMENTS FOR A VARIETY OF TYPES OF SERVICE

The most demanding applications with respect to outage seem to be for trunk line carrier services. The requirements for these services can most easily be satisfied (especially from a propagation standpoint) with spectrum below about 15 GHz for satellite links. It should be possible to use common shared frequency bands for both up and down satellite links, even though different antennas or satellites would be required for the up and the down links on the same frequency. However, this would be feasible only for applications where some minimum separation of terminals could be specified and controlled. It would not be practical for many mobile uses or for the density of usage that might be involved in receiving broadcast. These latter applications could best be satisfied with one-way usage, and such usage could be shared by all satellite relay services--mobile, fixed, and broadcast; however, purely terrestrial services involving transmitters and receivers on the same frequencies without minimum separation restrictions should be excluded for efficient use of these bands.

The greatest need for services with unrestricted terminal separations also involves the smallest antenna sizes. Mobile systems have mechanical aperture size limitations for operation, and broadcast receivers for homes would require small apertures in order to be economically feasible for so many receivers. Since the total global system capacity, as limited by satellite spacing, depends on the diameter in wavelengths of the earth terminal antennas, the greatest potential capacity for a given bandwidth as well as larger bandwidths can be obtained by choosing as short a wavelength as is practical. Also most of these applications are less demanding on service continuity than trunk line carrier applications, and can better tolerate the propagation degradations that may be experienced at these higher frequencies.

There is a further requirement for most mobile applications, which is to be able to point the antenna to accommodate rapid vehicle attitude changes. At the higher frequencies a given adequate aperture size (e.g., 1-m diameter) becomes fairly directive, necessitating a rapid pointing or tracking capability, whereas at very low frequencies (e.g., 500 to 1000 MHz) the same aperture size requires very little pointing. This difference between using the two extremes in frequency is further accentuated if simultaneous linking with a number of satellites in different directions is required. The technical problem of achieving this economically in the near term is much more formidable at the higher than at the lower frequencies. If it proves technically feasible to superimpose down links on current allocations in the range 470 to 960 MHz without restricting or interfering with the applications of existing allocations, this portion of the spectrum will seem extremely attractive for many mobile, broadcast, and austere terminal applications.

ILLUSTRATIVE SPECTRUM USE

There are many ways in which the spectrum might be partitioned and shared to provide enormous communication capacities. This illustration is one way in which it might be done with minimal jeopardy to vested practice, providing many times the capacity currently used or contemplated.

In portions of the spectrum between 3.4 and 15.35 GHz, there is 7.6 GHz of spectrum currently allocated to fixed and mobile or to communication satellite services. This is the prime portion of the spectrum most suitable for trunk line carrier and other very demanding services. It could be exploited most advantageously by sharing it between fixed and both up and down link communication satellite services. It should be possible to use this spectrum for the communication satellite services on two independent polarizations, and with high-gain satellite antennas, the spectrum capacity could be multiplied many times by independent beams from each satellite. If large earth-station antennas are used, many satellites at close spacings could be used independently. It is estimated that a margin of about 10 to 15 dB for propagation attenuation might be required at the 15 GHz end of this spectrum to ensure against system performance degradation 0.01 percent of the time in areas of moderately high precipitation.

The remaining portions of the spectrum that might be shared by satellite relays in this illustration involve one-way usage for all types of satellite relay service--mobile, fixed, and broadcast. The down links for one band might be from 470 to 960 MHz with the corresponding up links in the frequency region between 1430 and 2300 MHz. This usage might be designed to be superimposed on current allocations without mutual interference, and to serve air mobile, land mobile, direct broadcast, and other austere earth-station antenna applications. This spectrum could be used independently on two polarizations, and with large satellite antennas the spectrum could be reused with independent beams to different earth terminal areas. The number of satellites used would depend on the total power and coverage needed, and their orbital locations and spacings would not be critical.

In the spectrum between 17.7 and 50 GHz there is a total of more than 25 GHz that is currently allocated primarily to fixed and mobile services, or unallocated, that might be partitioned into exclusive satellite up and down bands for all kinds of satellite service, including fixed, air and land mobile, and direct broadcast. Since this part of the spectrum is subject to much larger atmospheric attenuations, more expensive redundant and diversity techniques are needed to achieve modest performance and continuity objectives. Space diversity might be used with fixed systems to meet most performance objectives with margins of 10 to 30 dB for propagation attenuation in this part of the spectrum. And, versatile direction diversity with redundant satellites might be used with mobile and direct broadcast applications with comparable margin allowances for propagation attenuation.

And finally, above 80 GHz (80 to 100 GHz, 130 to 160 GHz, and 200 to 300 GHz), there is 150 GHz of spectrum that is currently unallocated and that might be partitioned into exclusive synchronous satellite up and down bands for all kinds of satellite service. It would seem most appropriate, however, to emphasize the use of these latter bands with aircraft or satellites above much of the atmosphere.

All of the one-way bands above 17 GHz could be used independently on two polarizations, could provide adequate isolation to permit close spacing of satellites with earth terminal antennas of very modest physical size, and would permit further spectrum multiplication with high-gain (though physically small) antennas on the satellites. Thus, there is an enormous potential capacity available at the higher frequencies for applications that can accept or avoid the effects of large atmospheric attenuations that may occur.

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I. INTRODUCTION AND PROBLEM DESCRIPTION

Synchronous satellites offer great promise for communication relays and other information links to earth. However, there is growing concern that the uncoordinated proliferation of such satellites could appropriate or preempt desirable orbital locations and frequency bands in a grossly inefficient manner. It is the purpose of this Memorandum to explore the technological limits to the orbital and spectral capacity that might be achieved under various coordination conditions and system configurations.

Much of the spectrum that has been allocated (or that may be allocated in the future) for use with satellite relays is (or could be) shared with microwave relay or other systems. The intensity with which the spectrum and orbital space may be exploited and shared without intolerable interference depends on the values of the many parameters defining the systems involved. In the important case of spectrum sharing between communication satellite and microwave relay systems there are nine possible direct interference modes and two potentially important scatter interference modes as illustrated in Fig. 1-1.

The nine direct interference modes are derived from considering three types of receivers (terrestrial microwave, satellite, and earth station), each of which can experience an interference mode from the transmitters at each of the three types of stations. The two potentially important scatter interference modes, as illustrated in Fig. 1-1, involve the intersection of the main beam of an earth station with the main beam of a microwave relay station. If rain or other scatterers are contained within the common volume of main-beam intersection, scatter interference may be experienced either in the mode involving the earth-terminal receiver or in the mode involving the terrestrial receiving terminal, both of which may be viewing the illuminated common volume.

Some of the "possible" interference modes would not be experienced in current operations because of the separation between the frequency bands used to transmit up to satellites and those used to transmit from

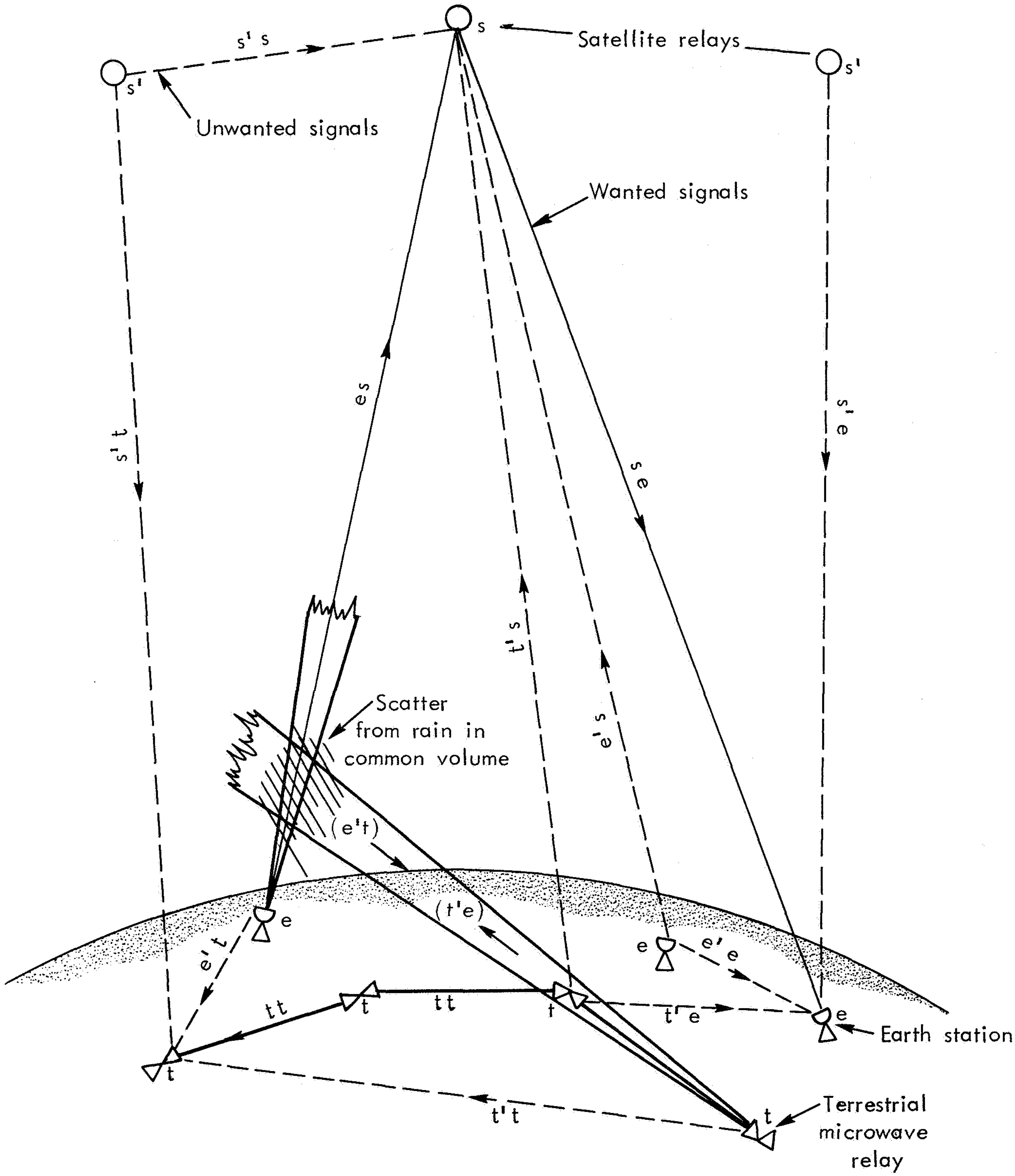


Fig. 1-1—Interference modes with shared frequency operation of satellite and terrestrial relay systems

satellites down to earth. However, the possibility of doubling the equivalent spectrum available makes attractive the use of some portions of the spectrum for both up and down links to satellites. For this reason the potential for this type of usage is included in this Memorandum.

The wanted carrier power C_{se} transmitted from a satellite to the input of an earth-station receiver for free-space propagation conditions can be indicated with the symbols used in this Memorandum as follows:

$$C_{se} = \frac{P_s \, G_{s\rightarrow e,e}}{4\pi d_{se}^2} \cdot \frac{\lambda^2 \, G_{e\leftarrow s,s}}{4\pi}$$

where subscripts s and e refer to satellite and earth stations respectively; P_s is the transmitted power at s; d_{se} is the distance from s to e; λ is the wavelength; $G_{x\rightarrow y,z}$ is the antenna gain in the direction of z when it is located at x and pointed toward y. The subscript arrow away from x indicates use in a transmitting mode and toward x in a receiving mode. Any subscript not needed for clarity may be omitted.

The unwanted signal power $X_{s'e}$, transmitted from another satellite s' to the input of the earth-station receiver can be expressed in a similar manner, and the ratio $X_{s'e}/C_{se}$ can then be obtained. However, it will be convenient to use a factor $M_{s'e}$ to sum up the interference from all the unwanted satellites, multiplying that received from the nearest or strongest unwanted satellite signal for this mode of interference. Also if the fractions of the unwanted and wanted signal powers that are received (compared to what would have been the case with free-space propagation) are represented by F_x and F_c respectively, we obtain the following expressions:

$$\frac{\sum_{s'} X_{s'e}}{C_{se}} = \frac{M_{s'e} X_{s'e}}{C_{se}} = \frac{P_{s'} \, G_{s'\rightarrow e'e}}{P_s \, G_{s\rightarrow e,e}} \cdot \frac{M_{s'e} \, d_{se}^2 \, F_x}{d_{s'e}^2 \, F_c} \cdot \frac{G_{e\leftarrow s,s'}}{G_{e\leftarrow s,s}} \simeq \frac{M_{s'e} \, G_{e\leftarrow s,s'}}{G_{e\leftarrow s,s}}$$

This is typical of the quantitative relationships between the various system parameters for each mode of interference. For the mode illustrated previously, the relatively simple final approximation results if all the satellites are equal in terms of effective radiated power in the direction of the earth station.

In Section II of this Memorandum, the relations between the ratio of unwanted to wanted signal power and output signal quality are developed and applied to determine the protection ratios C/X needed in theory and in practice for each link of a communication system. These relationships involve multiple links and noise entries and the effects of signal fading; modulation, bandwidth, and power tradeoffs; and multiplexing and a variety of signal-processing factors. Since the protection ratios depend on the detailed characteristics of both the wanted and unwanted signals, representative reference situations are developed that illustrate typical system and performance objectives to be expected in current and projected environments.

In Section III, the other factors quantitatively related to X/C are characterized. Reference antenna patterns and geometric models are presented, with which limiting intense-sharing situations and others can be tested. The possibilities of exploiting polarization discrimination are presented, and pertinent propagation factors are introduced.

Section IV develops the constraints on the various system parameters that will permit compatible satellite spacing and frequency sharing for each interference mode and for the combined modes. These constraints include the coordination that will be beneficial with respect to the relative and absolute power outputs of the various types of stations, the modulations and bandwidths used in the various systems, the antenna specifications and employments, and the locations of the various stations.

Finally, Section V presents the potential for shared exploitation of orbits and frequencies, including the dependence of hardware factors on frequency, and the potential demands for various types of service. A way of partitioning and using portions of the spectrum to obtain enormous communication capacity is presented to illustrate the potential of this limited resource.

II. MESSAGE OBJECTIVES, PROTECTION RATIOS, AND THEIR INTERRELATIONSHIPS

All of the communication systems of interest in this Memorandum may be viewed as special cases of an n-link radio-relay system. When unwanted rf signals (thermal noise, interference, distortion, etc.) enter the repeaters and the terminal receiver of such a system, corresponding types of noise appear at the single-channel outputs of the system and impair the quality of the messages they carry.

This section will first outline the way in which message quality in any selected channel of a radio-relay system depends on the type and magnitude of the unwanted rf signals entering the system. Then the message quality objectives to be met by current and future systems will be introduced and applied to determine the constraints they impose on the permissible levels of unwanted rf signals. Detailed background material will be found in a separate Memorandum.*

OUTPUT MESSAGE QUALITY VERSUS WANTED AND UNWANTED SIGNAL LEVELS

Output message quality is measured by the single-channel signal-to-noise ratio S_{ch}/N_{ch} in the case of analog messages, and by the error probability p_{ch} in the case of digital messages. For both types of message, output quality is a function of the ratios C_i/X_i, $i = 1, \ldots, n$, where C_i is the average power of the wanted signal which carries the selected channel, and X_i is the average power of the total unwanted rf signal in the wanted signal passband. Both C_i and X_i are measured at or referred to the input of the i^{th} repeater.

In general X_i will be the sum of contributions from a number of sources including thermal noise, nonlinear distortion, and radio frequency interference (RFI) from transmitters other than the $(i-1)^{st}$ repeater. The functional form of the relationship between output message quality, as described by S_{ch}/N_{ch} or p_{ch}, and the input wanted-to-unwanted-signal ratios C_i/X_i depends primarily on the nature, analog or digital,

*E. E. Reinhart, Radio Relay System Performance in an Interference Environment, The RAND Corporation, RM-5786-NASA, October 1968.

of the message and of the modulation method used for transmission.

Analog Messages and Analog Modulated Systems

When analog messages such as frequency division multiplexed (FDM) telephone and television signals are transmitted using analog modulation methods such as amplitude modulation (AM) and frequency modulation (FM), the relation between output message quality and the wanted-to-unwanted-signal ratios is particularly simple, providing that these ratios are much larger than unity. This simplicity arises from the fact that the output signal-to-noise ratio is proportional to the input signal-to-noise ratio not only for the analog demodulator which recovers the FDM baseband from the carrier, but also for the equipment at the terminal receiver which recovers the single-channel messages from the baseband.

Two types of radio-relay system should be distinguished according to whether the carrier is demodulated to baseband at each repeater (baseband repeaters) or merely translated in frequency without demodulation (rf or i-f repeaters). Thus, in a system of baseband repeaters, the baseband signal-to-noise ratio S_i/N_i at the i^{th} repeater exceeds the wanted-to-unwanted-signal ratio C_i/X_i by a factor of proportionality R_i called the demodulator transfer characteristic (DTC). Similarly, at the terminal receiver, the single-channel output signal-to-noise ratio S_{ch}/N_{ch} exceeds the effective baseband signal-to-noise ratio S_n'/N_n' by another proportionality factor R_{ch} called the channelizing transfer characteristic (CTC). Since noise due to unwanted signals accumulates at baseband, N_n'/S_n' is given by the sum of the N_i/S_i, and the desired relation between output message quality and the wanted-to-unwanted-signal ratios is

$$\frac{S_{ch}}{N_{ch}} = R_{ch}\,\frac{S_n'}{N_n'} = \frac{R_{ch}}{\sum\limits_{i=1}^{n} N_i/S_i} = \frac{R_{ch}}{\sum\limits_{i=1}^{n} \frac{1}{R_i}\,\frac{X_i}{C_i}} \qquad \text{Baseband repeaters} \qquad (2\text{-}1)$$

In a system employing rf or i-f repeaters, the only demodulation occurs at the terminal receiver. Hence the baseband signal-to-noise ratio S_n/N_n exceeds the effective wanted-to-unwanted-signal ratio C_n'/X_n' at the terminal receiver input by R_n, the DTC for the terminal demodulator. Since unwanted signals accumulate at rf rather than at baseband, X_n'/C_n' is given by the sum of the X_i/C_i, and the desired relation between output message quality and the wanted-to-unwanted-signal ratios is

$$\frac{S_{ch}}{N_{ch}} = R_{ch}\frac{S_n}{N_n} = R_{ch}R_n\frac{C_n'}{X_n'} = \frac{R_{ch}R_n}{\sum_{i=1}^{n} X_i/C_i} \qquad \text{rf or i-f repeaters} \qquad (2\text{-}2)$$

It is important to note that due to the linearity of Eqs. (2-1) and (2-2) exactly similar relations may be written for each of the components of the total unwanted signals and the corresponding component of output channel noise.

The products $R_{ch}R_i$, $i = 1, \ldots, n$, which appear in Eqs. (2-1) and (2-2) are called the receiver transfer characteristics (RTC), or interference reduction factors. In the case of the terminal receiver ($i = n$), for example, the RTC is the factor by which the output signal-to-noise ratio in the selected channel exceeds the effective wanted-to-unwanted-signal ratio at the terminal receiver input.

The numerical values of the DTC, CTC, and RTC depend on a variety of factors including the type of message, its position in the baseband, the number of channels in the baseband, the type of signal processing to which the message or the baseband is subjected, the particular modulation method used, the modulation index, and finally on the statistical nature of the unwanted signals. Since it is not practical to provide values appropriate for all interesting combinations of these factors, values of the DTC and CTC will be given for a reference case of wide applicability together with an indication of the departures from this case characteristic of other types of unwanted signal.

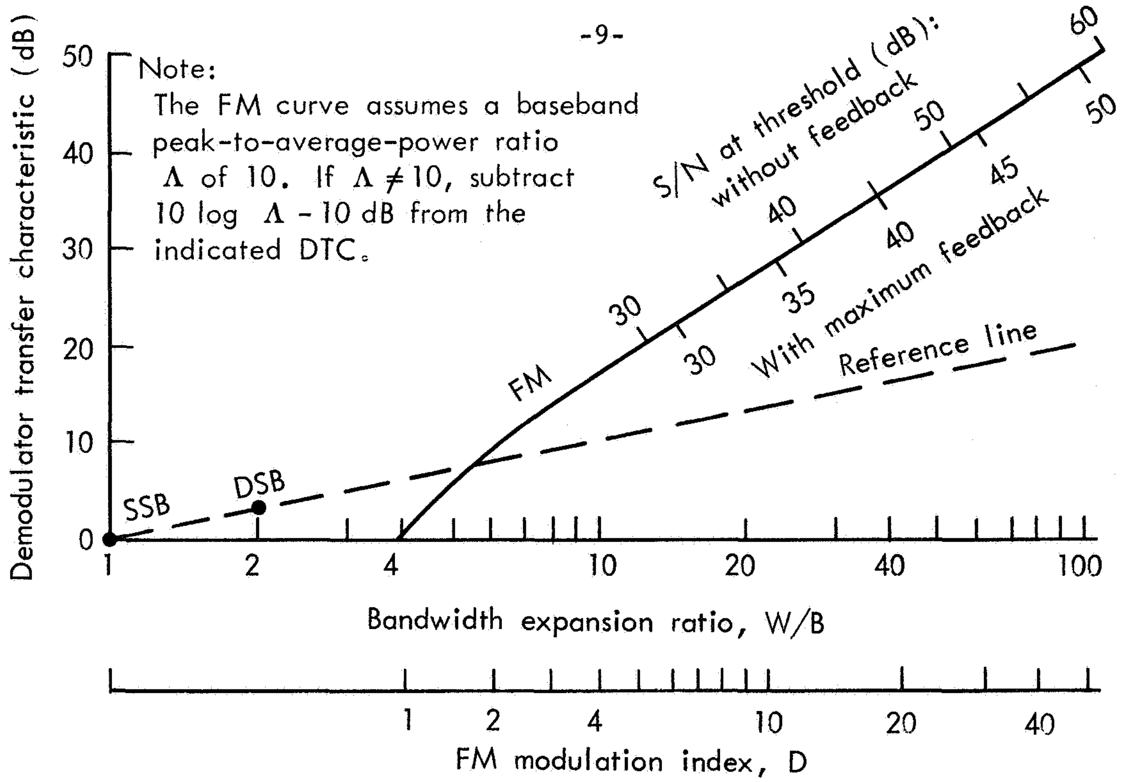
Reference Unwanted Signal. In the reference case, the unwanted signal is white noise. The values of DTC and CTC for this case are thus exact when the unwanted signal is thermal noise and are good approximations when the unwanted signal is rf interference with a more or less flat power spectrum across the passband of the wanted signal receiver.

Figure 2-1a shows the DTC in dB for the double- and single-sideband suppressed carrier forms of amplitude modulation (DSB and SSB), and for FM plotted against the ratio W/B of rf to baseband bandwidth. For FM, this ratio may be expressed in terms of the FM modulation index D using Carson's rule

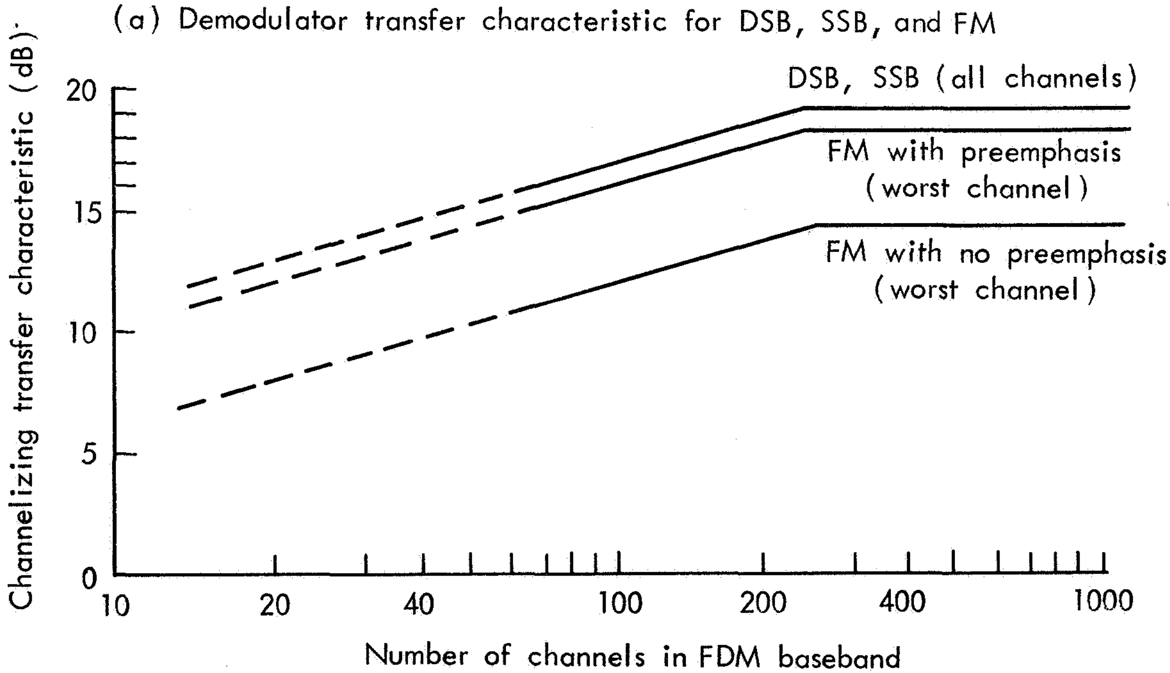
$$W/B = 2(D + 1) \tag{2-3}$$

It is apparent that with FM, increasing the modulation index will increase the DTC and hence reduce the amount of carrier power C needed to yield a specified baseband signal-to-noise ratio S/N in the face of a given unwanted signal power X. However, this process can only be continued until C/X is reduced to the FM threshold which, for a conventional FM receiver, is about 10 dB. With a feedback FM receiver, the threshold can be reduced, but again there is a limit. The scales on the FM curve in Fig. 2-1a relate the S/N at threshold to the corresponding maximum values of D or W/B, both with and without feedback.

Also shown in Fig. 2-1a is a reference line indicating the number of dB which must be subtracted from the DTC to obtain the amount by which the baseband signal-to-noise ratio exceeds the ratio of the wanted signal power to the white-noise power contained in the baseband, rather than the rf, bandwidth. Measured relative to the reference line, the DTC is thus simply the "improvement factor" relative to an SSB system--i.e., for systems of the same carrier power facing white noise of the same density, it gives the amount by which the baseband signal-to-noise ratio of the system in question will exceed that of an SSB system. Alternatively, for a given baseband signal-to-noise ratio, the improvement factor gives the amount by which the carrier power can be reduced below that of an SSB system. Note that DSB offers no improvement over SSB, and FM is actually less efficient than SSB and DSB for bandwidth expansion ratios less than 5.6 ($D < 1.8$).



(a) Demodulator transfer characteristic for DSB, SSB, and FM



(b) Channelizing transfer characteristic for FDM telephony

(c) Channelizing transfer characteristic for TV: $R_{ch} = \begin{cases} 22 \text{ dB; DSB, SSB} \\ 29 \text{ dB; FM with preemphasis} \end{cases}$

Fig.2-1—Transfer characteristics for white-noise unwanted signal

Figure 2-1b shows the CTC in dB for telephone messages as a function of the total number of channels in the FDM baseband. When added to the baseband signal-to-noise ratio, these values of the CTC directly yield the test-tone-to-psophometrically-weighted noise in a 3.1-kHz band. Note that the CTC depends on which modulation method is used and with FM, on whether or not the baseband is preemphasized. It also depends on the position of the channel within the baseband; the values shown for FM are for the worst or highest frequency channel.

Finally, Fig. 2-1c gives the CTC for a television channel assuming a gaussian amplitude distribution for the video baseband. When added to the baseband signal-to-noise ratio, the values shown yield the ratio of peak-to-peak video signal power (excluding synch pulses) to weighted noise in the 4.2-MHz video baseband.

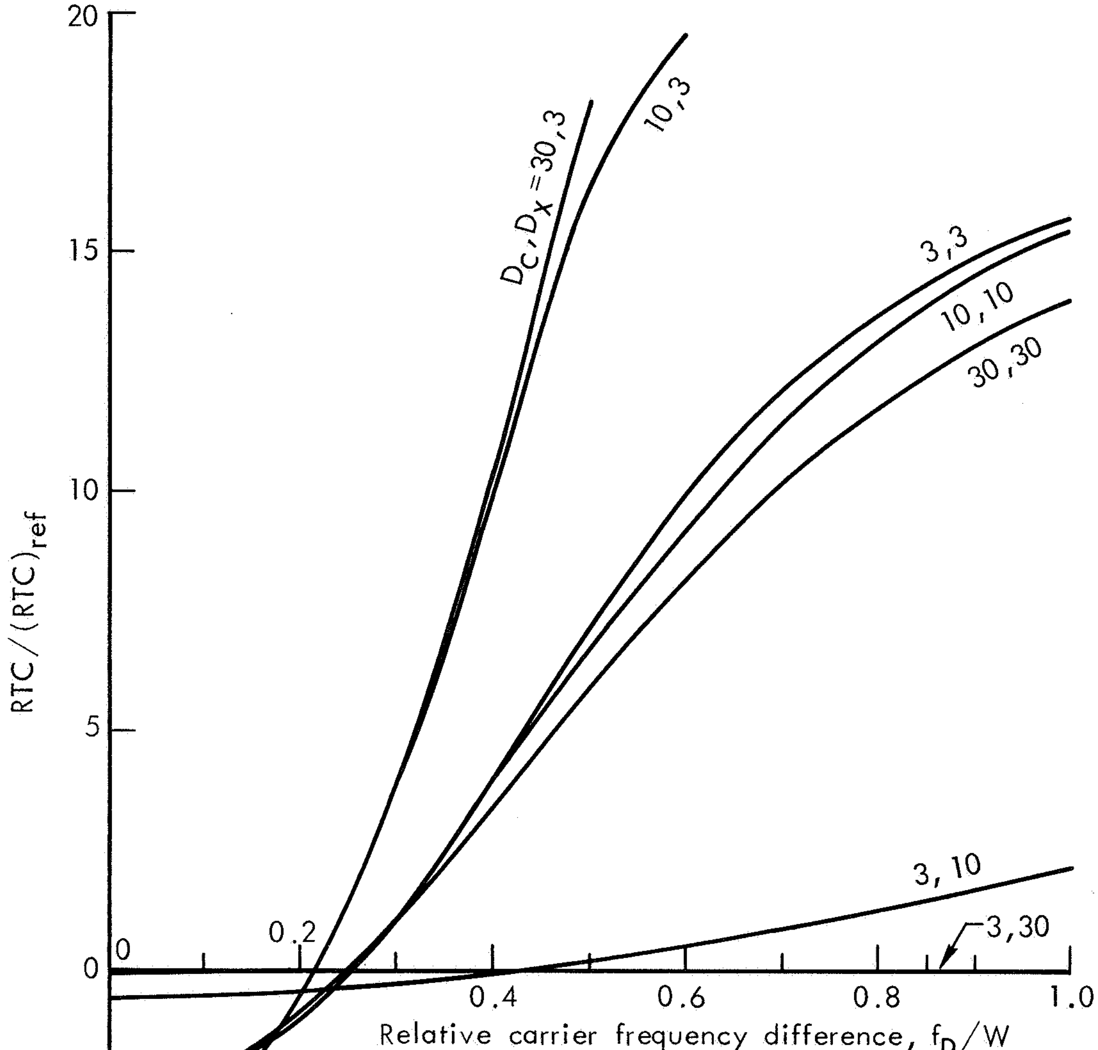
The RTC in dB for any baseband and for any of the illustrated modulation methods may now be found for the reference white-noise case by adding the appropriate values of DTC and CTC read from Fig. 2-1. To illustrate, consider a 60-channel preemphasized telephone baseband transmitted using FM with a bandwidth expansion ratio of 10. The DTC is seen to be 17 dB, and the CTC for the worst channel is 14.5 dB, leading to an RTC of 31.5 dB. This means, for example, that if a test-tone-to-noise ratio of 52.5 dB is required at the channel output, an effective rf wanted-to-unwanted signal ratio of at least 21 dB is needed at the terminal receiver input. As another example, if a preemphasized TV channel is transmitted using FM with a bandwidth expansion ratio of 10, the RTC is 17 + 29 = 46 dB. Thus the effective rf wanted-to-unwanted-signal ratio required for a 56-dB peak-to-peak output signal-to-noise ratio is only 10 dB.

Arbitrary Unwanted Signal. When the unwanted rf signal is not thermal noise, or does not have a uniform power spectrum across the bandpass of the wanted signal receiver, the values of the DTC and CTC will depart from those given in Fig. 2-1 for the white-noise reference case. The power spectrum of the noise at the demodulator output will no longer have the simple frequency dependence that is characteristic of a white-noise input--i.e., independent of frequency for the amplitude modulation methods and for phase modulation (PM), and parabolic for FM.

Instead, the output noise power spectrum will depend on the power spectra of both the wanted and the unwanted signals and on the difference f_D between their carrier frequencies. As a result, certain channels in a multichannel baseband will contain considerably more noise than they do in the case of a white-noise unwanted signal.

The output channel signal-to-noise ratio with an arbitrary unwanted signal relative to that for a white-noise unwanted signal of the same power is given by the ratio between the corresponding RTCs. Figure 2-2 displays this ratio as a function of carrier frequency difference for interference between wideband ($D \geq 3$) FM signals. Note that when the wanted and unwanted signals are strictly co-channel ($f_D = 0$), the RTC with an unwanted FM signal is always smaller (worse) than it is for the reference case. The RTC then increases as the carrier frequency difference f_D increases. Indeed, for interference between similar wideband FM signals, interleaving wanted and unwanted carrier frequencies with a frequency difference equal to half the rf bandwidth ($f_D = W/2$) is characterized by a worst channel RTC from 6 to 7 dB larger (better) than the reference case. Interleaving with $f_D = W/4$ produces an RTC equal to that of the reference case. It follows that with carrier frequencies interleaved at a spacing no smaller than one-quarter of the rf bandwidth, the output channel noise will be no greater than that due to white-noise interference having the same total power as that produced by the actual interfering carriers in the wanted signal passband.

As previously noted, output channel interference is particularly bad (RTC especially low) in certain channels when the unwanted signal includes a large spectral component (as it will with low index PM and FM), or when the modulating signal is small (as in periods of low traffic in the case of a telephone system). Conversely, for a given wanted-to-unwanted-signal ratio, the RTC generally improves (becomes larger) as the modulation index of the unwanted signal is increased. Indeed, it approaches the white-noise result as the unwanted signal becomes more uniform within the wanted signal passband. This is illustrated by the case $D_c = 3$, $D_x = 30$ in Fig. 2-2, where the subscripts c and x refer to the wanted and unwanted signals respectively.



D_C = modulation index of wanted signal

D_X = modulation index of unwanted signal

f_D = difference between carrier frequencies of wanted and unwanted signals

W_C = rf bandwidth of wanted signal

Fig. 2-2 — Receiver transfer characteristic for interference between wideband FM signals relative to that for white-noise interference

Digital Messages and Analog Modulated Systems

Digital messages such as computer data, teletype, and certain forms of facsimile can be, and commonly are, transmitted over analog modulated radio-relay systems. For this purpose, the digital message may be filtered or encoded to modify its spectral characteristics and then multiplexed with other digital signals using either digital or analog modulation methods to obtain a baseband which, for transmission purposes, is treated as an analog signal. For example, from 12 to 16 teletype signals may be combined into a 4-kHz FDM baseband by frequency-shift keying suitably spaced audio-frequency subcarriers; the result is then treated as an ordinary telephone message. Similarly, binary data up to a few thousand bits per second may also be treated as a 4-kHz voice signal after 4-level or 8-level phase-shift keying (PSK) of an audio-frequency subcarrier.

One of the problems in transmitting digital data over analog facilities is the danger of overloading the linear circuits which they constitute; another is that digital messages are much more vulnerable than analog messages to phase distortion. In the case of high-speed binary data, these problems are overcome by first subjecting the digital baseband to various types of coding and filtering and then by using a modulation method such as vestigial-sideband amplitude modulation (VSB) which transmits the carrier for phase reference but at a greatly reduced level.

The quality of the digital message at the output of the system is of course a function of the signal-to-noise ratio S_{ch}/N_{ch} in the analog channel which carried it. This ratio may be found with the aid of an appropriate RTC, but the conversion from S_{ch}/N_{ch} to digital error probability is governed by a relation that normally is highly nonlinear. Qualitatively, this relation will have the same form as the digital demodulator functions to be discussed in a moment in connection with digital modulated systems. But quantitatively, the relation will depend on the nature of the digital messages and on the details of the signal-processing, modulation, and multiplexing techniques used with them.

Analog Messages and Digital Modulated Systems

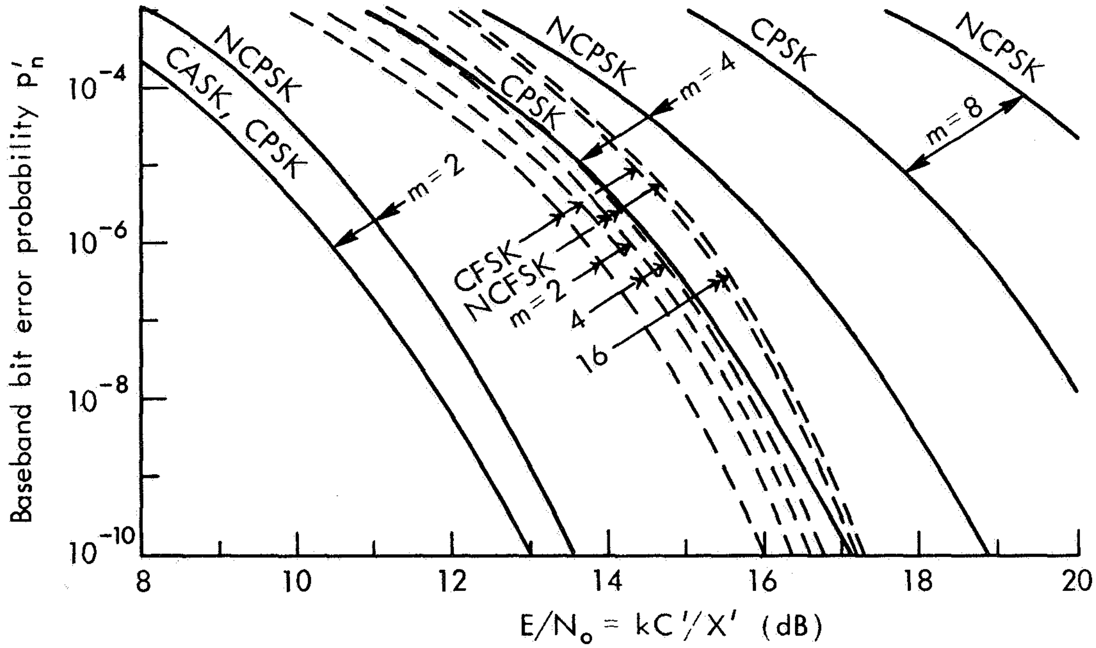
In order to transmit an analog message over a radio-relay system which employs a digital modulation technique, the message waveform is first converted to digital form, i.e., the message is processed or encoded into a sequence of discrete levels. Thus encoded, the message may be combined with others using time division multiplexing (TDM) to form a digital baseband. Alternatively, the analog messages may be combined into a baseband using FDM and then converted to digital form. In either case, the baseband signal is then used to modulate or "shift" the amplitude, frequency, or phase of a sinusoidal carrier among a set of m discrete values. These modulation methods are the m-ary digital equivalents of AM, FM, and PM and are abbreviated ASK, FSK, and PSK, where the "SK" stands for "shift keying."

As in an analog modulated radio-relay system, the baseband may be recovered at each repeater (baseband repeaters) or the modulated carrier may simply be amplified and shifted to a new carrier frequency (rf or i-f repeaters). After recovery at the terminal receiver, the baseband is demultiplexed into individual channels and the single-channel messages are converted back from digital to analog form.

The relation between output message quality and the wanted-to-unwanted-signal ratios at the repeater inputs may be found for a digital system in the same manner as for an analog system. The result is considerably more complicated, however, because the quality of a digital baseband, as measured by the error probability at the demodulator output, is a highly nonlinear function of the input wanted-to-unwanted-signal ratio.

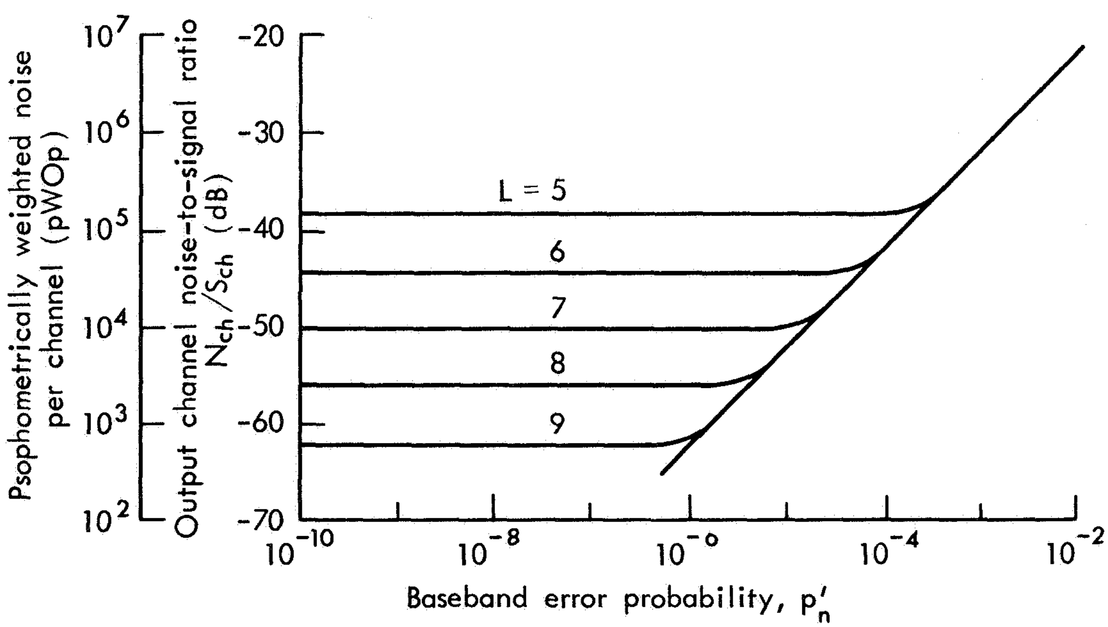
The procedure will be illustrated and the necessary functions presented graphically for the same white-noise unwanted signal used as a reference case for analog modulated systems.

Reference Unwanted Signal. In a digital system employing baseband repeaters, the probability p_i of baseband errors due to a white-noise unwanted signal X_i at the i^{th} repeater input depends on C_i/X_i in the manner shown for several digital modulator-demodulator combinations in Fig. 2-3a. The total baseband error probability p_n' at the output of



E = wanted signal energy per baud
N_o = one-sided noise spectral density
m = number of modulation levels
k = pulse shape factor
C'/X' = effective wanted-to-unwanted-signal ratio

(a) Demodulator functions for m-ary digital modulation methods



(b) Channelizing functions for PCM-encoded telephone channels

Fig.2-3—Demodulator and channelizing functions for white-noise unwanted signal the demodulator in the terminal receiver is then the sum of the p_i. If the baseband comprises pulse-code modulated (PCM) TDM telephone channels, the output signal-to-noise ratio S_{ch}/N_{ch} will depend on p'_n as shown in Fig. 2-3b.

If the digital system employs rf or i-f repeaters, the baseband is not recovered until the carrier reaches the terminal demodulator, and Fig. 2-3a now gives the total baseband error rate p'_n in terms of the effective wanted-to-unwanted-signal ratio C'_n/X'_n at the terminal receiver input. Here, as with an analog system, C'_n/X'_n may be obtained as the reciprocal of the sum of the reciprocals of the C_i/X_i. Finally, Fig. 2-3b again shows how S_{ch}/N_{ch} depends on p'_n.

Since they characterize the operation of digital demodulators, the functions shown in Fig. 2-3a will be called demodulator functions. Note that in general, the demodulator function depends on which parameter is modulated, on the number m of discrete values the parameter may take, and on whether the demodulator is coherent or not, as indicated respectively by the letters C or NC preceding the abbreviation for the modulation method.

The functions displayed in Fig. 2-3b are called channelizing functions because they describe the operation of the terminal equipment that recovers the single-channel messages from the demodulated baseband. The values of the PCM channelizing functions shown in the figure have been adjusted so that the output channel signal-to-noise ratios they predict may be compared directly with the message quality objectives for FDM telephony. The defining equation is

$$10 \log \frac{N_{ch}}{S_{ch}} = 10 \log \left(4p'_n + 4^{-L}\right) - 8 \text{ dB} \tag{2-4}$$

Where p'_n is the total baseband error rate and L is the number of bits per PCM level.

In this expression, the term $4p'_n$ represents the contribution of noise due to digital transmission errors, whereas the term 4^{-L} is the noise contribution due to the quantizing process. It is apparent from

Fig. 2-3a that the error noise contribution can be reduced to any desired level by increasing the wanted-to-unwanted-signal ratio. However, there is little point to increasing C/X much beyond the threshold value defined by the condition that the error noise is equal to the quantizing noise. For higher values of C/X, the quantizing noise contribution is dominant, and the output signal quality remains constant at a value determined by L.

In comparing digital modulation methods with each other and with analog methods, the rf bandwidth occupied by the modulated carrier is equally as important as the required wanted-to-unwanted-signal ratio. The ratio W/B of rf bandwidth to the total information bandwidth of the analog signals depends on the analog-to-digital-conversion method as well as the modulation method, and on the pulse shape which describes the transitions from one digital level to another. For L-bit PCM and m-ary digital modulation, the bandwidth expansion ratio is given by

$$\frac{W}{B} = \begin{cases} 2kL/\log_2 m & \text{ASK, PSK} \\ 2(m-1+k)\ L/\log_2 m & \text{NCFSK} \\ (m-1+2k)\ L/\log_2 m & \text{CFSK} \end{cases} \qquad (2\text{-}5)$$

where k measures the effect on bandwidth of pulse shape. The theoretical minimum bandwidth corresponds to $k = 1$, whereas $1.5 < k < 2$ represents good engineering practice.

Just as it is possible for analog modulated systems to combine the DTC and CTC to obtain an overall receiver transfer characteristic, so for digital modulated systems the demodulator and channelizing functions may be combined to describe the wanted-to-unwanted-signal ratio required to yield a prescribed output message quality. Figure 2-4 shows the result of effecting this combination for PCM-encoded telephone basebands transmitted by binary and by 4-level CPSK with $k = 1.5$. To facilitate comparison with FM, the curves are drawn to show the amount by which the output channel signal-to-noise ratio exceeds the threshold input wanted-to-unwanted-signal ratio as a function of the corresponding rf-to-baseband-bandwidth ratio. The corresponding quantity for an FM system is the RTC which depends on the number of channels in the baseband as well as on the bandwidth expansion ratio as shown in Fig. 2-4. It will be noted that for PCM with 7 or more bits per sample, CPSK offers both power and rf bandwidth savings relative to FM.

Arbitrary Unwanted Signal. When the unwanted signal does not have the characteristics of white noise, error probabilities can be expected to be somewhat lower than for a white-noise signal of the same average power. This will certainly be true when the unwanted signal comprises modulated carriers having sinusoidally distributed amplitudes. The rate of increase of error probability with unwanted signal power is sufficiently fast, however, that the difference in error probability does not permit a significant reduction in wanted signal power. For system analyses, therefore, the demodulator functions for white noise will be used for arbitrary unwanted signals as well.

Digital Messages and Digital Modulated Systems

When digital messages are transmitted by systems employing digital modulation methods, the output message quality is given by the channel error probability. This is normally equal to the baseband error probability so that the channelizing function is equal to unity. Hence, the relation between output message quality and the input wanted-to-unwanted-signal ratios is described entirely by the demodulator functions as given in Fig. 2-3a for several digital modulation methods. The appropriate measure for the rf bandwidth in this case is given by its ratio to the bit rate f_{bit} of the digital baseband

$$\frac{W}{f_{bit}} = \begin{cases} k/\log_2 m & \text{ASK, PSK} \\ (m-1+k)/\log_2 m & \text{NCFSK} \\ \frac{1}{2}\,(m-1+2k)/\log_2 m & \text{CFSK} \end{cases} \qquad (2\text{-}6)$$

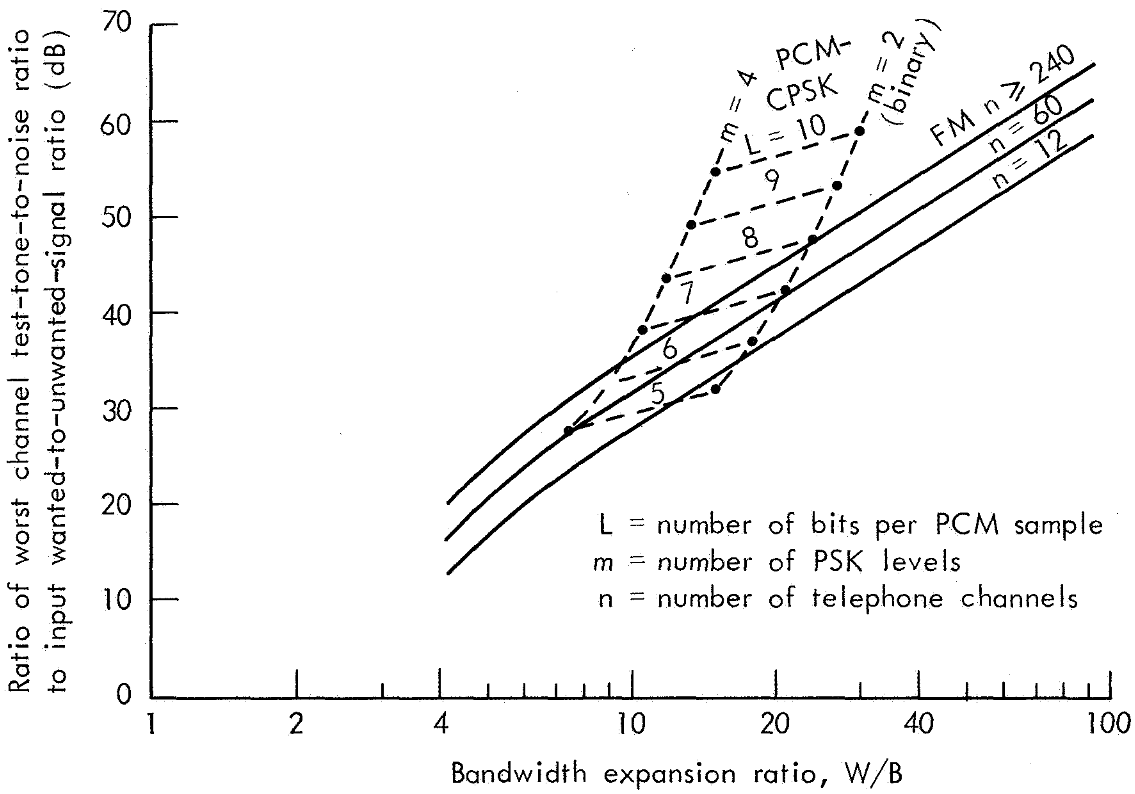


Fig.2-4—Comparison of interference reduction capability of FM with preemphasis to PCM-CPSK at threshold when used for multichannel telephony

MESSAGE OBJECTIVES AND PROTECTION RATIOS

Having established the relation between output message quality and the input wanted-to-unwanted-signal ratios C_i/X_i for the worst channel of a radio-relay system, it is only necessary to substitute a specific value for the minimum allowable message quality in order to obtain a constraint on the allowable values of the C_i/X_i. The output quality specification will be called the message objective, and the corresponding minimum values of the C_i/X_i will be called protection ratios.

In determining appropriate protection ratios, however, it is important to note that the message objective constrains the wanted-to-unwanted-signal ratios collectively rather than individually. For example, in an analog modulated system of i-f repeaters, it is apparent from Eq. (2-2) that the individual ratios C_i/X_i may have any values as long as the sum of their reciprocals does not exceed the product of the applicable RTC and the maximum permissible output noise-to-signal ratio.

In order for a message objective to determine individual protection ratios, it is necessary to divide it into contributions assignable to the individual links of the system, and, within a link, to the individual interfering sources. In practice, message objectives are not ordinarily specified in such detail. Typically, the objective will indicate only the total output noise and the component due to a specified type of rf interference. It is then up to the system designer to further subdivide the objective between repeaters and individual sources.

The problem of determining protection ratios is further complicated by the fact that, due to path loss variations and other causes, the wanted-to-unwanted-signal ratios, and hence the output message quality, will vary in time--i.e., fade. In order to allow for propagation-induced fading, message quality objectives are normally specified in the form of probability distributions which show the fraction of time that specified quality levels should be maintained. The objectives do not make allowance for unplanned differences in equipment parameters and for long-term changes in signal levels due, for example, to equipment aging. These factors may be taken into account by adding appropriate margins to the calculated protection ratios.

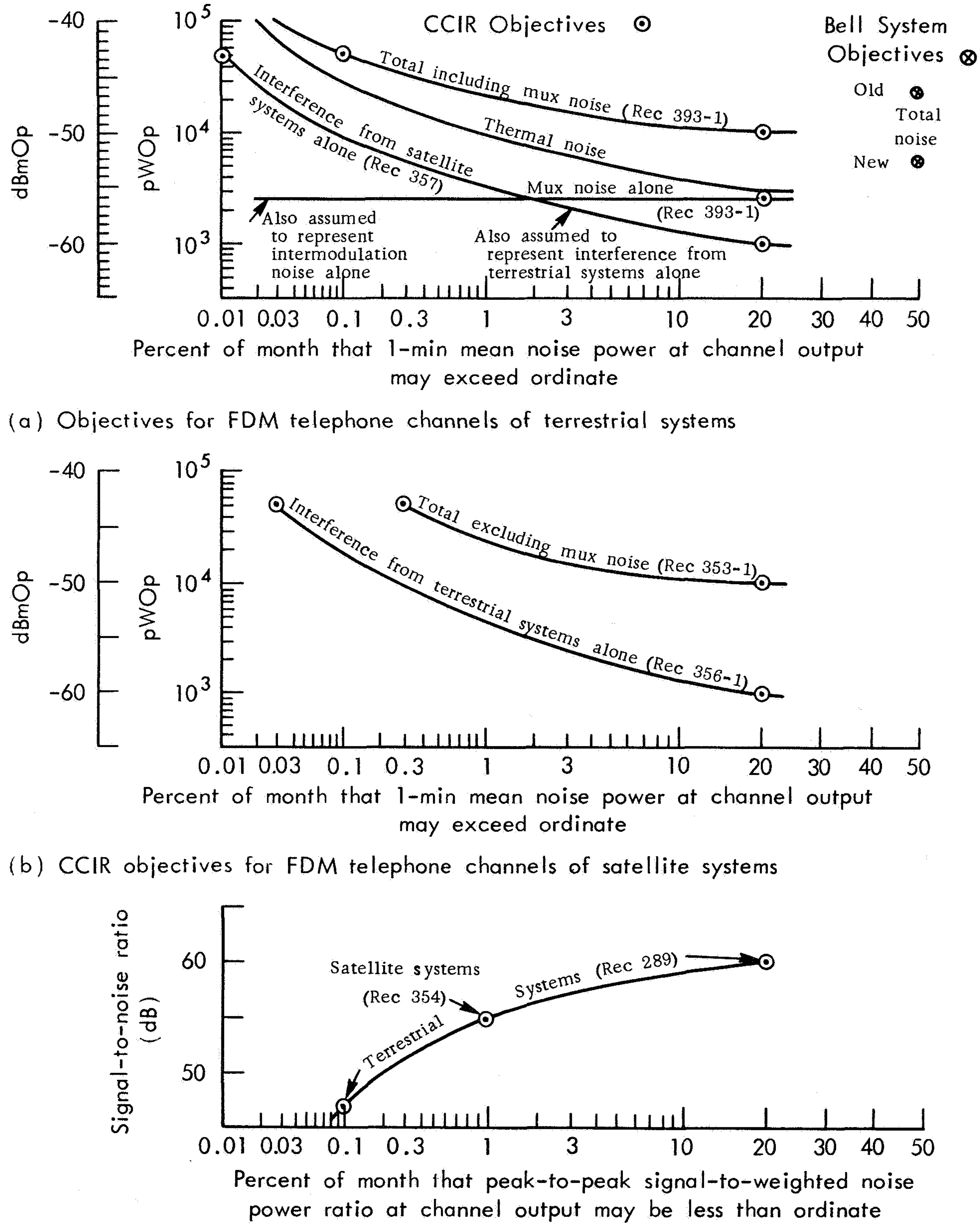
Message Objectives

Many of the foregoing remarks are illustrated by the objectives for telephone and TV messages shown in Fig. 2-5. These are based primarily on recommendations of the International Radio Consultative Committee (CCIR) for so-called hypothetical reference circuits. Also shown for comparison at 50-percent probability are the old and new Bell System telephone objectives for total noise in the absence of fading.

Note that in the case of FDM telephone channels (Figs. 2-5a and 2-5b) the CCIR has specified only the total noise and the components of noise originating from multiplexing equipment and from interference due to sources outside the system, as indicated by the circles at 0.01 percent, 0.1 percent, and 20 percent for terrestrial-relay systems and at 0.03 percent, 0.3 percent, and 20 percent for satellite relay systems. All of these objectives represent one-minute averages. Thus, the percentages on the abscissa scale may be thought of in terms of the number of minutes per month to which they correspond. For example, 0.01 percent is 4.3 min, 0.1 percent is 43 min, etc.

The objectives shown in Figs. 2-5a and 2-5b for other components and for other percentages of time represent reasonable subdivisions and extrapolations of the recommended values. For this purpose it has been assumed that in the absence of fading, the thermal noise and full-load intermodulation noise are equal to each other, and also that the interference from terrestrial systems is the same as the interference from satellite systems. Finally, the distribution of fading for both components of interference has been assumed to be the same as that for thermal noise.

In specifying telephone objectives, it is customary to indicate the desired signal-to-noise ratio S_{ch}/N_{ch} by giving the allowable value of N_{ch} as a so-called point of 0 relative level where, by definition, S_{ch} = 1 mW. The CCIR normally gives N_{ch} in units of pW0p (picowatts at a point of 0 relative level, psophometrically weighted), whereas the Bell System uses dBrnC (dB above reference noise of 1 picowatt, C message weighted at a point of 0 relative level), or dBa0 (dB above reference noise adjusted at a point of 0 relative level). In Fig. 2-5a and 2-5b all of these units have been converted to dBm0p (dB above a milliwatt at a point of 0 relative level, psophometrically weighted) using the



(a) Objectives for FDM telephone channels of terrestrial systems

(b) CCIR objectives for FDM telephone channels of satellite systems

(c) CCIR objectives for 525-line TV channels

Fig.2-5—Noise and interference objectives identities

$$N_{dBm0p} = 10 \log N_{pW0p} - 90 = N_{dBa0} - 84.5 = N_{dBrnC} - 90.5$$

It is also useful to note that N_{dBm0p} is numerically equal to the weighted-noise-to-test-tone ratio $(N_{ch}/S_{ch})_{dB}$ since $S_{ch} = 0$ dBm0.

The CCIR objectives for television channels shown by the circles in Fig. 2-5c apply to the ratio of peak-to-peak video signal (excluding synchronizing pulses) to the 1-sec average power of continuous random noise, using the CCIR-recommended weighting network. If a fading distribution similar to that assumed for telephone systems is adopted, the television objectives for satellite and terrestrial systems are seen to lie on the same curve.

The principles involved in using the message objectives of Fig. 2-5 to derive protection ratios are the same for satellite systems as for terrestrial radio-relay systems, but basic differences in the two types of systems affect the relative importance of the factors which must be taken into account. Thus, a typical long-haul microwave relay system has a large number of links (e.g., n = 140 for a transcontinental system), each having quite similar equipment and path parameters, and each subject to severe signal fading during certain times of the day. A typical satellite link, however, has only two links (n = 2), which are likely to have very different equipment parameters but exhibit practically no diurnal fading.

Protection Ratios for Terrestrial FM Radio-relay Systems

Consider an n-link microwave relay system carrying at least 240 FDM telephone channels using i-f type repeaters and low index FM without feedback. Such a system is typical of those employed by the Bell System for transcontinental telephone and TV transmission. Protection ratios at the repeater inputs are needed for both the thermal noise and the RFI components of the unwanted signal. For this purpose, it will be assumed that in the absence of fading, all repeaters have the same apparent carrier power input and that the fading on individual links, while uncorrelated between links, is governed by the same distribution relative to the free-space or unfaded carrier level.

Thermal Noise. The protection ratio for thermal noise may be found by using the experimentally determined single-link fading distribution in conjunction with Eq. (2-2) and the transfer characteristics for the reference case to predict the fading distribution of 1-min mean noise power at the output of the worst channel. The protection ratio is then that unfaded value of the single-link carrier-to-noise ratio which makes the predicted output distribution lie as close as possible to the objective thermal noise distribution without exceeding it at any point.

If the predicted distribution matches the objective at all points, the thermal noise protection ratio may be determined entirely from the message objectives that apply in the absence of a fade. Thus, using the subscript o to indicate unfaded values, and setting $C_i/X_i = (C/X)_o$, Eq. (2-2) gives for the thermal noise components of X_i and N_{ch}

$$\left(\frac{C}{X}\right)_o = n\left(\frac{C_n'}{X_n'}\right)_o = \frac{n}{R}\left(\frac{S_{ch}}{N_{ch}}\right)_o \qquad (2\text{-}7)$$

where $(C_n'/X_n')_o$ is the effective carrier-to-noise ratio at the terminal receiver, and R is the RTC for the reference unwanted signal, as obtained by adding the DTC and CTC in dB for FM as given in Fig. 2-1.

For example, with a modulation index D = 1.5, Fig. 2-1 gives R = 23.5 dB. Hence, setting n = 140 and using the 20-percent thermal-noise objective $10 \log(S_{ch}/N_{ch})_o = 55$ dB from Fig. 2-5a, Eq. (2-7) gives, for the thermal-noise protection ratio,

$$10 \log(C/X)_o = 21.5 - 23.5 + 55 = 53 \text{ dB}$$

This is to be compared with the unfaded single-link carrier-to-noise ratio of 50 to 53 dB obtained in practice with the TD-2 system whose modulation index with a 600-channel baseband is about 1.5.

The effect of changing the FM modulation index or bandwidth expansion ratio from that assumed in the example may be obtained with the aid of Fig. 2-1. Thus, if the index is reduced from 1.5 to 0.5, the RTC will decrease from 23.5 dB to 11.5 dB, and the protection ratio must be increased from 53 dB to 65 dB. This is close to the 63-dB carrier-to-noise ratios typical of the TH system whose modulation index with 1860 channels is about 0.5. Conversely, doubling the modulation index will increase the RTC by about 8 dB and permit the protection ratio to be relaxed to 45 dB.

In this connection, however, it is important to note from Eq. (2-2) that when the carrier-to-noise ratio on a single link fades by a factor ρ, the effective carrier-to-noise ratio at the terminal receiver input is reduced to

$$\left(\frac{C_n'}{X_n'}\right)_F = \frac{1}{\rho + n - 1}\left(\frac{C}{X}\right)_o \tag{2-8}$$

If the fade is deep, ρ will be much larger than n - 1 and it follows that $(C_n'/X_n')_F$ will be approximately equal to the apparent carrier-to-noise ratio $(C/X)_o/\rho$ on the faded link.* Therefore, in choosing the modulation index, the protection ratio should exceed the sum of the FM threshold and the maximum allowable single-link fade measured in dB; otherwise the terminal input carrier-to-noise ratio would drop below threshold during the fade. For example, if the maximum fade is $F = 10 \log \rho = 35$ dB, the modulation index should not exceed the value for which the protection ratio is equal to F + 10 = 45 dB.

The preceding calculations of thermal-noise protection ratios using the unfaded noise objective assumed that the predicted distribution of 1-min output noise power matched the thermal-noise objectives at all points. This method of calculation is still valid when the predicted fading is less severe than that implied by the objective, but when it is more severe, the noise objective corresponding to small percentages of the month is the dominating consideration and leads to somewhat higher

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*Comparing this with Eq. (2-7), it is seen that $(C_n'/X_n')_F = (n/\rho)(C_n'/X_n')_o$; i.e., during a single-link fade, the terminal input carrier-to-noise ratio fades by the factor ρ/n, or $F - 10 \log n$ dB where $F = 10 \log \rho$.

values of protection ratio. For example, if actual fading is comparable to that observed in certain parts of the TD-2 system, it can be shown that the protection ratios must be increased by about 8 dB over those calculated from the unfaded objectives.

Interference. In computing protection ratios for RFI in terrestrial systems, it should be recognized that the wanted signal at the repeater inputs will usually be the principal parameter to exhibit fading. Thus a distribution quite similar to that for the carrier-to-thermal-noise ratio should be obtained for the carrier-to-interference ratio and the calculation procedure will be the same except that (1) the RTCs will differ from reference case values to the extent that the power spectrum of the interfering signals departs from that of white noise, and (2) the unfaded carrier-to-interference ratios may not be the same at all repeaters.

To take these differences into account, Eq. (2-2) is written in component form

$$\frac{S_{ch}}{N_{ik}} = R_{ik} \frac{C_i}{X_{ik}} \tag{2-9}$$

where N_{ik} is the component of the total output channel noise N_{ch} caused by an unwanted signal of type k and power X_{ik} entering the system at repeater i, and R_{ik} is the RTC for unwanted signals of this type.

In order to use this equation to calculate interference protection ratios, it is of course necessary to subdivide the total interference objectives among the repeaters and the different interference sources. Such a subdivision, consistent both with the 20-percent CCIR objectives shown in Fig. 2-5a and with commercial practice for intra-system interference, is shown in Table 2-1A. The corresponding interference protection ratios, calculated from these objectives on the assumption that fading is no worse than that implied by the objectives for the smaller percentages of the month, are also presented in Table 2-1A for the same FM modulation indices used to illustrate the thermal-noise protection ratios--i.e., D = 0.5, 1.5, 3. It will be noted that the TV objectives lead to protection ratios that are less stringent than those imposed by the objectives for telephone messages.

Table 2-1

EXAMPLES OF MESSAGE OBJECTIVES AND PROTECTION RATIOS
FOR TERRESTRIAL FM RADIO-RELAY SYSTEMS

A. FDM Telephone

Type of Unwanted Signal	Assumed Number of Entries	20-percent Message Objective (pW0p)		Protection Ratio (dB) FM Modulation Index D		
		Per Entry[a]	Total	0.5 (TH)	1.5 (TD-2)	3
Mux equipment	...	...	2500[b]	..	..	..
Intermodulation	...	...	2500	..	..	..
Thermal noise	140	21.4	3000	65	53	45
RFI from terrestrial systems	140	7.1	1000	70	58	50
RFI from satellite systems	140	7.1	1000[b]	70	58	50
Total	1	10,000	10,000[b]	38.5[c]	26.5[c]	18.5[c]

B. Television

Type of Unwanted Signal	Assumed Number of Entries	20-percent Message Objective (dB)	Protection Ratio (dB) FM Modulation Index D		
			0.5 (TH)	1.5 (TD-2)	3
Thermal noise	140	60[b]	58.5	46.5	38.5
RFI from satellite systems	140	65	63.5	51.5	43.5
Total	1	55	32[c]	20[c]	12[c]

[a] An entry in this Memorandum is considered as the sum of all the unwanted signals of a given type at one repeater.

[b] CCIR recommended value.

[c] Minimum effective wanted-to-unwanted-signal ratio at terminal receiver input.

For simplicity, white-noise reference case values of the RTCs were used in computing all of the protection ratios. This is a good approximation for interference from terrestrial stations, and it also holds quite well for interference from satellite systems if carrier frequencies are spaced no closer than one-quarter the rf bandwidth. The protection ratios for the latter interference are of course quite dependent on how the total 1000-pW0p objective for this kind of interference is partitioned--i.e., on how the 1000 pW0p is divided between earth stations and satellite repeater sources and how many entries of each kind are assumed. The result in Table 2-1 illustrates the simple case where this objective is divided equally among all 140 repeaters.

Effect of Changing the Message Objectives. It is interesting to speculate on some of the design changes that would be possible if the message objectives were modified. For example, if in Table 2-1, the total objectives for thermal noise and for interference from satellite systems were to be interchanged, the protection ratios for thermal noise would all be increased by 5 dB while those for interference could be reduced by the same amount. The new protection ratio for thermal noise could then be achieved, for example, by increasing the transmitter power at all of the repeaters by 5 dB. This change in repeater power would not affect the observed carrier-to-interference ratios for intrasystem interference, and so the protection ratio for this type of interference would continue to be met. But the carrier-to-interference ratios observed for interference from satellites would be enhanced by 5 dB. Since the protection ratio for this type of interference is now 5 dB lower than before, satellite EIRP could be raised by nearly 10 dB.

Satellite Systems

As previously noted, a satellite relay system has only two links to consider, and fading on these links is normally less severe than that indicated in the CCIR noise objectives. On the other hand, the output noise objective may not be divided equally between the up link and the down link. This is typically the case for the thermal and intermodulation components, but for the cases of the RFI components that will be illustrated, they are divided equally between up link and down link.

The C/X ratios required for any specified partition of the output noise objectives for a satellite system may be calculated from Eq. (2-9). Thus, designating the up link by the subscript i = 1, and the down link by i = 2, the C/X ratios for the thermal noise component of the unwanted signals are

$$\left(\frac{C_1}{X_{iN}}\right)_o = \frac{1}{R_{iN}}\left(\frac{S_{ch}}{N_{iN}}\right)_o \qquad i = 1,2$$

where R_{iN} and $\left(S_{ch}/N_{iN}\right)_o$ are respectively the RTC and the unfaded message objective for thermal noise entering the system at the receiver for link i. Similarly the C/X ratios for the other components of the unwanted signal may be obtained.

The calculated C/X ratios are not designated protection ratios in this case since they do not include a variety of margin factors that must be added in a practical system. The appropriate margins and protection ratios will be discussed and illustrated in Section IV.

The C/X ratios for RFI for telephone and television channels using the FM parameters proposed for the INTELSAT III system and for the domestic pilot program of the Communication Satellite Corporation are indicated in Table 2-2.

The unfaded message objectives assumed for both systems are displayed in Table 2-2A for telephone channels and in Table 2-2B for television. The former are consistent with a published breakdown of the total CCIR objective of 10,000 pW0p used in planning the INTELSAT III system.

As with the examples used to illustrate the calculation of protection ratios for terrestrial systems, reference noise values of RTC are used in all cases. This is believed to be a conservative assumption for a model environment in which carrier frequency interleaving makes the total interference spectra uniform across the passband of any given receiver. In INTELSAT III, three capacities of rf channels are provided for telephone messages. These employ wideband FM to carry FDM basebands of 24, 60, and 132 channels in nominal rf bandwidths of 5, 10, and 20 MHz respectively. The proposed FM modulation indices are such that the

Table 2-2

EXAMPLES OF MESSAGE OBJECTIVES AND PROTECTION RATIOS

FOR SATELLITE FM SYSTEMS

A. FDM Telephone

Unwanted Signal	20-percent Message Objective pW0p		C/X (dB)		20-percent Message Objective RFI Emphasis (pW0p)		C/X (dB)
	Total	Per Entry[a]	INTELSAT III	Pilot Program Trunk (D = 1.25)	Total	Per Entry	Minimum
IM, thermal, and other noise	9000	...	....	..	7000	....	..
RFI	1000[b]	500	18.5	42	3000	1500	12
Total	10,000[b]	10,000	5.5[c]	29[c]	10,000[b]	10,000	6[c]

B. Television

Unwanted Signal	20-percent Message Objective (dB)		C/X (dB)		20-percent Message Objective RFI Emphasis (dB)		C/X (dB)
	Total	Per Entry	INTELSAT III D = 3.75	Pilot Program D = 3	Total	Per Entry	Minimum
Thermal noise	60[a]	..	..	..	65	..	..
RFI	65	68	23	25	60	63	12
Total	55	55	10[c]	12[c]	55	55	6[c]

[a] An entry in this Memorandum is considered as the sum of all the unwanted signals of a given type at one repeater.

[b] CCIR recommended value.

[c] Minimum effective wanted-to-unwanted-signal ratio at terminal receiver input.

reference RTC is about 44.5 dB for all three capacities. If the 1000-pW0p objective for RFI is divided equally between up link and down link, a C/X of 18.5 dB is obtained as indicated in Table 2-2A.

The trunk message channels proposed for the domestic satellite system would carry 1800-channel telephone basebands using narrowband FM with a modulation index of about 1.25. The RTC for such a circuit and white-noise interference is about 21 dB leading to the C/X for RFI of 42 dB indicated in Table 2-2A.

For TV transmission with INTELSAT III, three quality levels are available. The highest quality level uses an FM modulation index of 3.75 and requires a nominal 40-MHz rf channel. The reference value of the RTC for such a channel is 45 dB, which, when combined with the message objectives used for terrestrial TV transmission, yields the C/X for RFI of 23 dB shown in Table 2-2B. The C/X for RFI of 25 dB is obtained for TV transmission over the pilot system which proposed an FM modulation index of 3.

Also shown in Table 2-2 is a partition of the total message objectives for telephone and for TV which would emphasize the RFI components, i.e., allocate about one-third of the total objective to RFI and divide it equally between the up and the down links.

The final minimum total C/X of 6 dB shown in Table 2-2 could be obtained with a system that uses an appropriate modulation index and corresponding RTC. Although this apparent value is about 3 dB below threshold, it is assumed that either only half of the total nominal objective will be experienced, or that future standards may be changed to half of the nominal total objectives indicated. This type of consideration appears to have been used in the planning for INTELSAT III which yields a total C/X of 5.5 dB for FDM telephone channels.

III. ANTENNA, PROPAGATION, AND GEOMETRIC FACTORS

Antenna, propagation, and geometric factors play a prominent role in the separation or isolation of different communication links sharing the same spectrum. The satellite spacings and the number of satellites within view of an earth station that can use the same spectrum independently without producing excessive interference will depend critically on the directivity patterns of the earth-station antennas as determined by their size (diameter/wavelength or D/λ) and illumination. Significant isolation may also be achieved by appropriate use of polarization discrimination. Finally, propagation factors such as multipath fading and precipitation attenuation can become important in most systems, and geometric factors such as separation distances and orientations of directive antennas can become critical in the compatibility of systems that share frequencies.

REFERENCE ANTENNA PATTERNS

In order to illustrate the effects of typical antennas on satellite spacing and system design, some reference antenna models were selected and their patterns for a $D/\lambda = 100$ are illustrated* in Fig. 3-1. The CCIR standard is a conservative peak envelope of current low-noise antenna practice recommended by CCIR for interference calculations. The "parabolic" antenna represents what might be expected from typical production design with an illumination function falling to 0.25 power at the perimeter and a 10-percent feed-blocking factor. The pattern represents the average value to be expected for the sidelobe region and is asymptotic to 8 dB below isotropic at large angles off the mainlobe on the front side of the antenna. The "uniform" illumination circular array represents what might be expected if the performance is limited by a random 20-percent rms excitation error for each element in the array. The pattern again represents the average value of the sidelobes. A similar random error and average value of sidelobes is assumed for the Taylor 40 dB-5 array and the

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*See Appendix A for further illustrations and details.

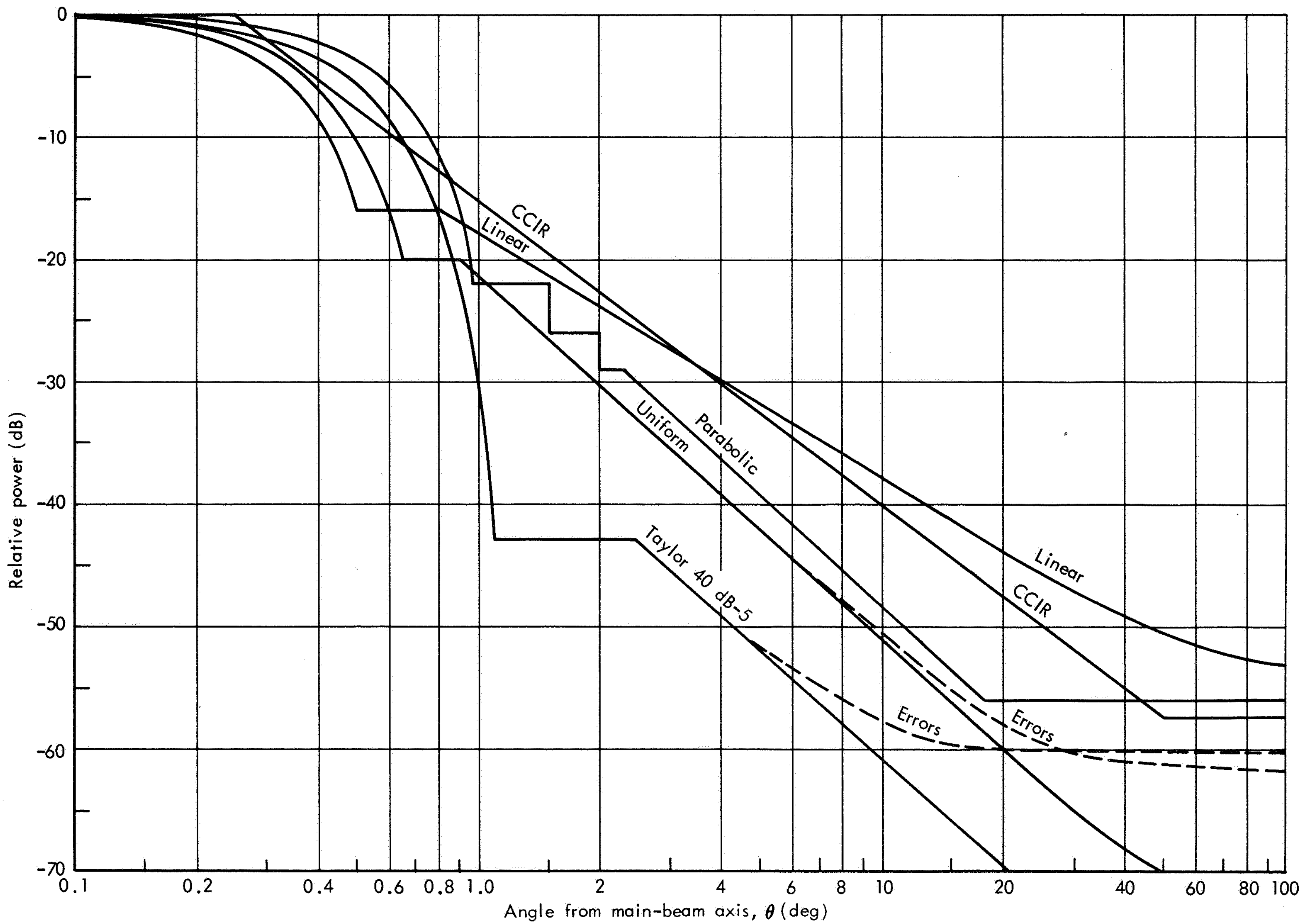

Fig. 3-1 — Antenna model patterns for D/λ = 100 uniformly illuminated "linear" array. All patterns are felt to be conservative performance estimates of future production possibilities.

In applying the reference patterns the question will arise as to how much margin should be allowed in the protection ratios to account for the variations in the antenna patterns above the value for the average sidelobe. If a situation is considered involving only one interfering source in the near sidelobes, the theoretical peak interference power would be twice the average sidelobe level, so a 3-dB margin addition would seem justified. However, if there are many interfering sources, particularly if they are spaced a number of beamwidths apart, and if bandwidths exceeding a few percent are used, no additional margin may be necessary. For the many closely spaced satellites illustrated in this Memorandum, a 2-dB margin addition is considered a conservative allowance.

In addition to the fixed patterns already discussed, an interesting future possibility is the adaptive or adapted array for which the phasing and illumination functions are adjusted to minimize the total interference received from unwanted directions. This technique can establish nulls in discrete interfering directions and can also minimize interference diffused over wide sidelobe angles. The rejection is limited by the bandwidth, the number of interfering sources, and the number and spacing of antenna elements. Also, it may be more effective if the wanted signal is coded in some manner to facilitate the adapting process.*

All the individual antenna patterns have two distinct, though interacting, regions of interest: the mainlobe and sidelobe regions. The mainlobe is characterized by very steep edges with the overall width for a given D/λ varying by about a factor of two over the range of illumination functions of interest for the applications considered. The uniform illuminations have the narrower mainlobes, while illuminations tapering away from the center may be used to lower the sidelobe levels at the expense of broadening the mainlobe by nearly a factor of two in the extreme cases considered here.

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*See Appendix A for further details of adaptive arrays.

The antenna patterns that have been illustrated are "free-space" patterns unmodified by the earth or other environment. The effects of reflections or scattering from the environment can significantly modify the free-space patterns--particularly at the lower backlobe levels. This may be most relevant for microwave relays that have been designed to have very low free-space backlobes. If these stations illuminate the earth or scattering objects with the antenna mainlobe or principal sidelobes, and if the illuminated earth can be seen from satellite locations, the effective coupling between the satellites and the microwave relays may be equivalent to significantly larger backlobes than free-space patterns would indicate.

POLARIZATION DISCRIMINATION AND PROPAGATION FACTORS

The use of polarization discrimination could be exploited to double the effective spectrum available if appropriate system design choices are made. Circular polarization for satellite communications has some obvious advantages over the method where one or both of the antennas are linearly polarized. However, it is not generally possible to obtain perfect circular polarization because of the depolarizing nature of the reflector in dish antennas and other hardware and propagation factors. Thus elliptical polarization with unequal axes must be considered.

Figure 3-2 illustrates the limits on polarization discrimination as a function of the difference in axial ratios of the polarization of the incident signal and the receiving antenna. The maximum discrimination is obtained when the major axes of the unwanted incident signal and the receiving antenna are orthogonal, and the minimum occurs when the major axes are aligned. The orthogonal case applies for all axial ratios, and the aligned case applies where the larger axial ratio of either the signal or receiving antenna is that marked on the curve. If the axial ratios of the polarization of the incident unwanted signals from satellites and the axial ratio of polarization for the earth receiving antenna can be maintained orthogonal and within 2 dB of each other, or if they can both be maintained within 1 dB regardless of their orientation, a discrimination of better than about 19 dB would be assured. On the other hand, if these

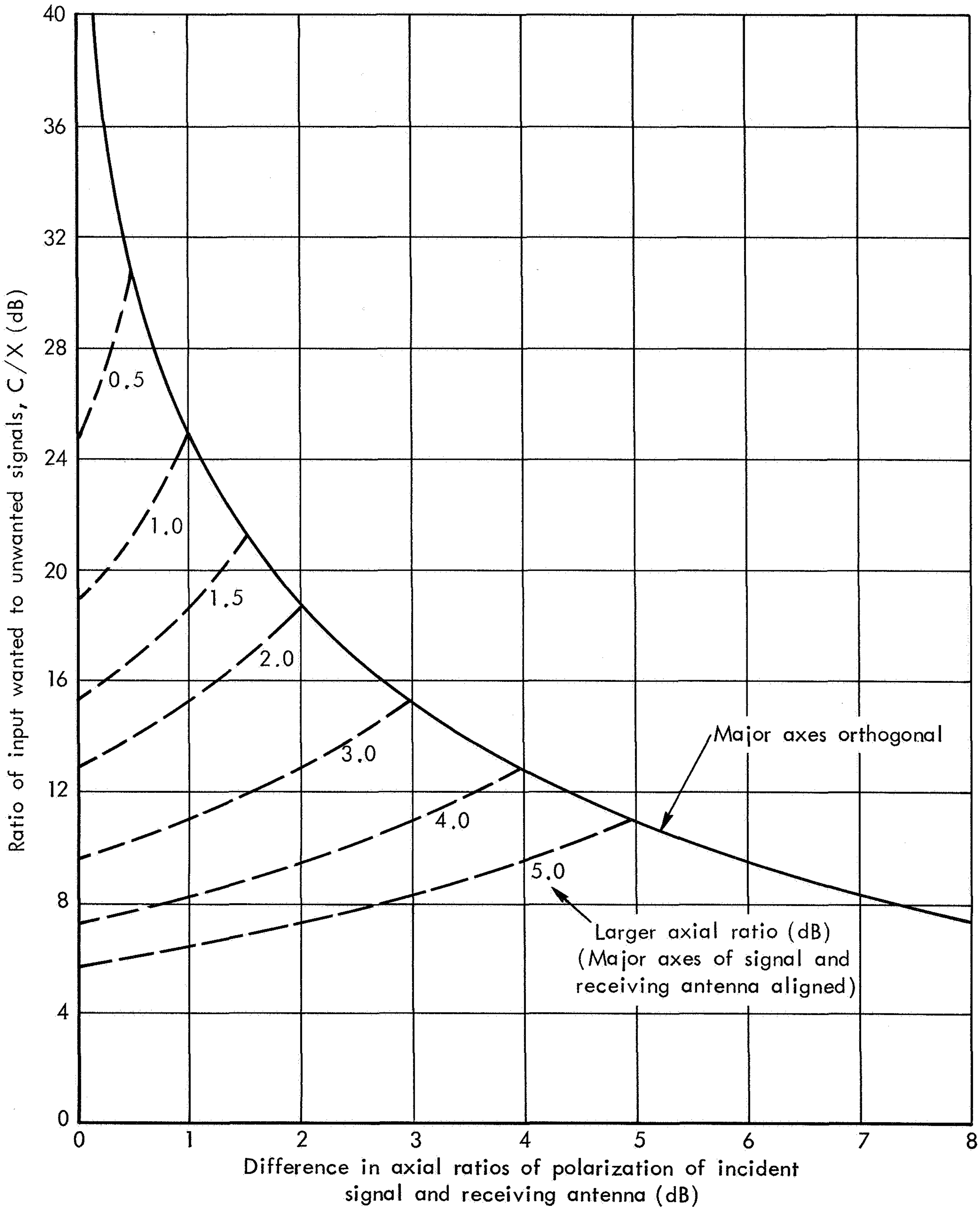

Fig.3-2—Polarization discrimination

axial ratios could be maintained orthogonal and within 0.5 dB of each other, more than 30 dB of discrimination could be assured. In practice it should be feasible to maintain operating specifications somewhere between these two so as to provide between 19 and 30 dB of polarization discrimination <u>over the main beam</u> (or within 1/3 radian of the axis for wider beams) without excessive cost in equipment design and maintenance. The sidelobes are assumed to be adequately rejected by the angular antenna response characteristics so as not to require significant additional polarization discrimination in sidelobe directions.*

As another alternative, an adaptive system might be used to maintain the axes of the polarization of the unwanted incident signal and the receiving antenna orthogonal while adjusting the magnitude of the axial ratio of the receiving antenna to minimize the unwanted signal. This technique should make feasible very high polarization discrimination.

Thus it seems feasible to provide polarization discrimination of 20 to 30 dB (or more with adaption), so that if modulation and other system characteristics are chosen to provide the output signal quality desired with input wanted-to-unwanted signal ratios in this range, it should be possible to double the effective spectrum available by using it independently on two orthogonal polarizations.

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*The curves of Fig. 3-2, which were determined under the assumption that the axial ratios of the wanted and unwanted incident signals are equal, also account for the fact that there is some discrimination against the wanted signal, unless the axial ratio of the wanted signal is equal to that of the receiving antenna and the major axes are aligned. However, since for axial ratios less than 2 dB the discrimination against the wanted signal is only a few tenths of a dB regardless of the orientation of the polarization axes, the curves would be altered only negligibly if the axial ratios of the wanted and unwanted signals were unequal. For detailed theoretical background on polarization discrimination refer to H. G. Booker (ed.), et al., "Techniques for Handling Elliptically Polarized Waves with Special Reference to Antennas," <u>Proc. IRE</u>, Vol. 39, No. 5, May 1951, pp. 533-552; G. Sinclair, "The Transmission and Reception of Elliptically Polarized Waves," <u>Proc. IRE</u>, Vol. 38, No. 2, February 1950, pp. 148-151; and M. G. Morgan and W. R. Evans, Jr., "Synthesis and Analysis of Elliptical Polarization Loci in Terms of Space-Quadrature Sinusoidal Components," <u>Proc. IRE</u>, Vol. 39, No. 5, May 1951, pp. 552-556.

The polarization discrimination discussed above will not be affected by propagation factors in most satellite applications of interest, because the propagation paths of the wanted and unwanted (to be discriminated against) signals are practically identical and are affected in the same way. Thus no additional propagation margin needs to be added to the C/X that must be obtained in order to take advantage of polarization discrimination. However, in order to accommodate propagation vagaries such as multipath fading or precipitation attenuation, appropriate margins must be included in the protection ratios to ensure other systems performance objectives. This tends to become more and more important with increasing frequency, and techniques such as space diversity to alleviate weather-induced attenuation* and adaptive antennas to reduce multipath fading** become attractive.

EARTH-TO-SATELLITE GEOMETRY FACTORS***

For any given distribution of synchronous satellites, the differences in C/X at various earth-station locations are due primarily to the variations in the angular separations between a wanted satellite signal and its nearest unwanted neighbors. The angular separations are determined largely by the distance from the earth station to the desired satellite, a distance fully prescribed by specifying the satellite elevation above the horizon. Variations of C/X are thus determined primarily by apparent elevation of the satellite above the horizon. For the extreme situations of vertical and horizontal viewing from the earth with the angularly most sensitive antennas, the net excursion of C/X is about 2 dB. The reference earth-to-satellite geometry model used for interference calculations illustrated in this Memorandum consists of an earth station at 45° latitude looking at a wanted equatorial synchronous satellite on the same longitude and a full orbit of unwanted synchronous satellites at

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*See Appendix C for further details.

**See Appendix A for further details.

***See Appendix B for details.

equal longitudinal spacings on both sides of the wanted satellite. This reference case is near the midpoint of the variation of the C/X with earth-station location,* and an additional margin allowance of 1 dB is adequate to handle other geometries from this latitude.

Another earth-to-satellite geometry factor that may become important for system design to reduce interference is concerned with the relative alignment of terrestrial microwave relay antennas and the satellite relay antennas. If the microwave relay sites are chosen so that none of their antennas point toward synchronous orbits, the interference potential can be greatly reduced. If, in addition, the satellite antennas are sufficiently directive so as to be able to significantly reduce their gain toward the earth's horizon, the interference potential can be further reduced. Such techniques are useful for low to mid-latitudes where relatively small bands of east-west microwave relay azimuths need to be excluded. However, above latitudes of about 60° it may be necessary to exclude so much north-south azimuth that sharing the same frequencies between microwave relays and satellite relays at these latitudes may not be feasible. In fact the technological possibilities of designing the satellite antenna patterns to provide uniform (to within 3 dB) coverage below 60° latitude, while adequately rejecting latitudes above 60°, seem rather remote. Thus most of the burden of antenna alignment for reducing interference in shared frequency bands must be placed on the terrestrial microwave relays, and at the higher latitudes these microwave relays may be limited to usage only with links within narrow east-west azimuth bands.

REFERENCE GROUND-ENVIRONMENT MODELS

In order to test how the various system parameters influence interference and the possibilities for sharing frequencies between terrestrial microwave relay stations and communication satellite earth terminals, it will be necessary to use suitable reference ground-environment models. One such reference model provides the maximum ideal density of ground stations for a given minimum separation d between stations. This is a

*See Appendix B for details.

uniform distribution arranged in a triangular grid of hexagonal patterns so that each station has 6 nearest neighbors at the minimum separation distance, 12 more stations within twice the minimum separation distance, 18 additional stations in the interval between two and three minimum separation distances, etc. It will be assumed that these locations represent the ideal maximum combined distribution density of terrestrial and earth stations using a common frequency band.

Another ideal distribution model that may be of interest for some applications would be one based on a pattern of squares with the minimum separation distance d forming the side of a square. The density of the ideal square distribution is $\frac{1}{2}\sqrt{3}$ times that from the triangular grid of hexagonal patterns. Figure 3-3 illustrates the manner in which interference would accumulate for the two ideal distributions discussed, assuming isotropic radiators on a flat plane with radiated power attenuating as $1/(\text{distance})^2$ from the sources. The accumulated interference from the square distribution approaches approximately $\frac{1}{2}\sqrt{3}$ times that from the triangular distribution, and as a first approximation the accumulation out to a distance of n minimum separation distances can be represented by $I_n \simeq C_1 + C_2 \log n$, where C_1 and C_2 are constants for any particular distribution. Thus, the interference does not accumulate very rapidly, and any additional attenuation beyond free-space propagation, e.g., atmospheric absorption or beyond-the-horizon diffraction, will normally quickly dampen to insignificance the contribution of the more distant sources. This may be illustrated by assuming a minimum spacing of 40 km (typical of microwave relay spacing in the 4 and 6 GHz bands) with isotropic sources on line-of-sight towers. Then with either the triangular or square distribution in representative terrain, the total power density at any one site from all directions would conservatively be less than six times that from the nearest station. Even with exceptional terrain and anomalous propagation it is very unlikely that the above factor would be exceeded for a minimum separation of 40 km.

These reference ground-environment models should be treated as limiting cases of high-density frequency sharing, and the interference that might be experienced in practice would only approach that predicted if the intensity of sharing approaches that of the models.

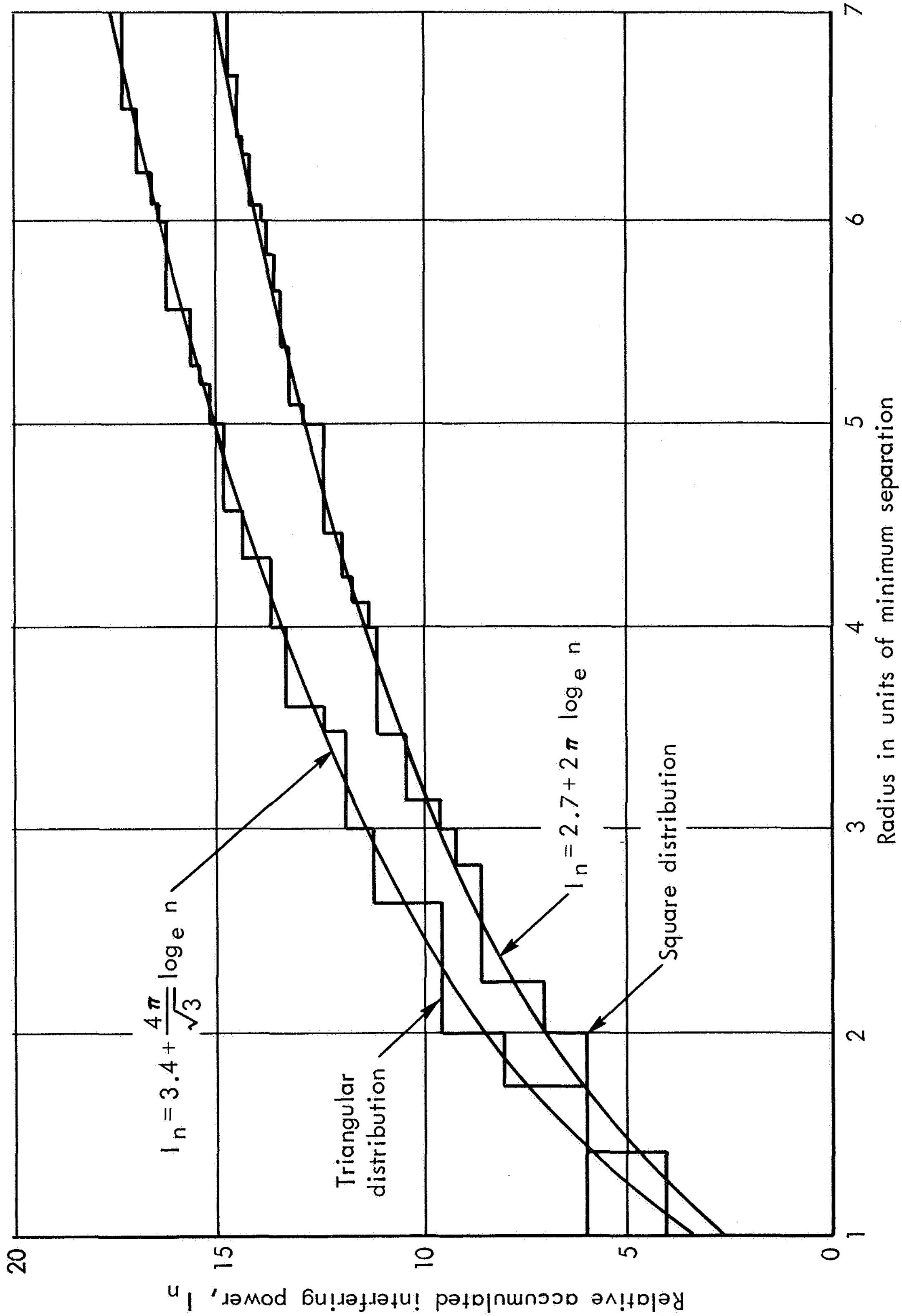

Fig.3-3—Ground-enviroment models

IV. PARAMETER CONSTRAINTS FOR TOLERABLE INTERFERENCE

The type of relationship indicated in Section I between X/C and the other parameters of the systems will be used in this section to determine the constraints on the various parameters for tolerable interference. With the performance objectives, the protection ratios and their interrelationships developed in Section II, and the other factors presented in Section III, the parameter constraints for tolerable interference in each interference mode are determined. For convenience in referencing, the modes will be numbered arbitrarily in the order in which they are discussed.

INTERFERENCE FROM SATELLITES TO EARTH STATIONS, MODE 1

The interference mode that will be considered in greatest detail is the single-link interference from unwanted satellite signals to the earth station receiving the wanted satellite signal. As the definition of symbols in Section I indicated,

$$\frac{\sum_{s'} X_{s'e}}{C_{se}} = \frac{P_{s'} G_{s'\to e',e}}{P_s G_{s\to e,e}} \cdot \frac{M_{s'e} d^2_{se} F_x}{d^2_{s'e} F_c} \cdot \frac{G_{e\leftarrow s,s'}}{G_{e\leftarrow s,s}} \simeq \frac{M_{s'e} G_{e\leftarrow s,s'}}{G_{e\leftarrow s,s}} \qquad (4\text{-}1)$$

The simple approximation holds very well for most situations in which the equivalent isotropically radiated power (EIRP) in any given bandwidth in the direction of the earth station is the same from every satellite. This situation will be elaborated in more detail than the others, although the effects of differing EIRP (whether inadvertent or by design) will also be treated.

Effects of Earth-station Antenna Patterns

The approximation of Eq. (4-1) principally involves the earth-station receiving antenna patterns. The satellite spacings that are derived from these patterns are obtained as follows: The reference earth-satellite geometry model is used, i.e., the earth station is at 45°

latitude with its antenna pointed at the wanted satellite on the same meridian, and there is a full ring of synchronous equatorial satellites at uniform longitudinal spacing providing interfering signals. All satellites are assumed to have the same EIRP in any given bandwidth in the direction of the earth station, but only those satellites above the local horizon are assumed to contribute interference. The ratios of the sum of unwanted signals to wanted signal (interference-to-signal ratios) for the reference antenna patterns of Section III are then determined as a function of satellite longitudinal spacing. These are illustrated in Fig. 4-1 for $D/\lambda = 300$.*

Figure 4-2 is a similar illustration for a linear adaptive array with $D/\lambda = 10$. Each point on the curve for an adaptive array represents a different illumination function. Thus the adaptive curves represent families of antennas of the same size, each of which is adapted to give the best signal-to-interference ratio for the satellite spacing involved. The Taylor 40 dB-5 array curves for $D/\lambda = 10$ are included for comparison with the adaptive arrays.

Several potential applications for which adaptive arrays appear especially attractive are outlined in Appendix A. The outstanding discrimination performance that appears possible with relatively small apertures compared with conventional antennas might completely change system design for satellite spacing and frequency sharing. However, until their performance has been proven in practice and the conventional antennas have been replaced, system design specifications must be based on the weakest conventional antenna whose performance must be protected in service. Thus the remainder of the discussions in this section will be concerned with parameter constraints for tolerable interference using conventional antennas.

An important limiting factor to consider for any of the curves of Figs. 4-1 or 4-2 is that any source of noise or interference which becomes competitive with the satellite sources under consideration but which is independent of satellite spacing will provide an asymptotic level for which increased spacing will no longer yield decreasing X/C.

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*Figures for other D/λ values and details for their use are given in Appendix A.

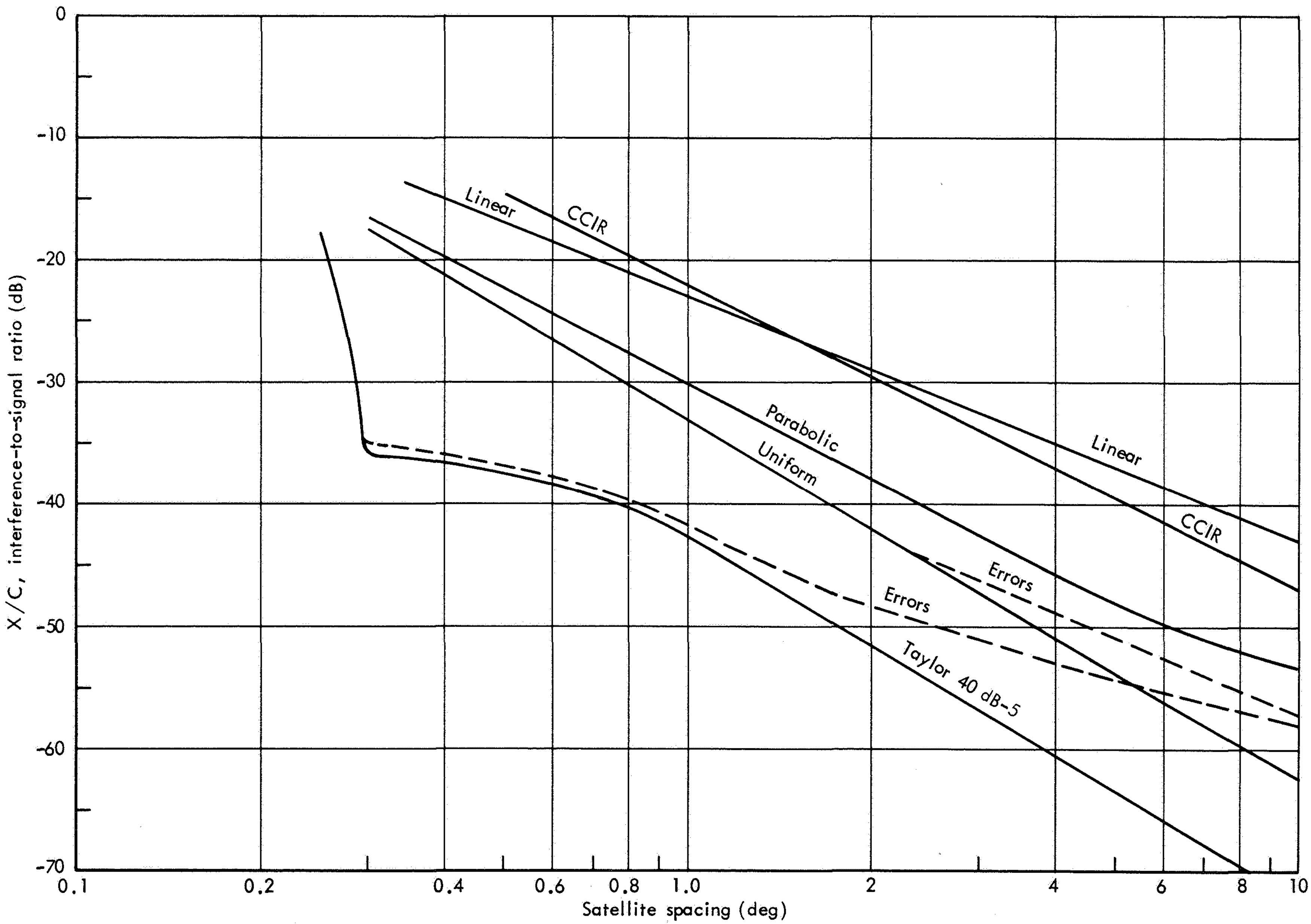

Fig.4-1—X/C versus satellite spacing for antennas with $D/\lambda = 300$

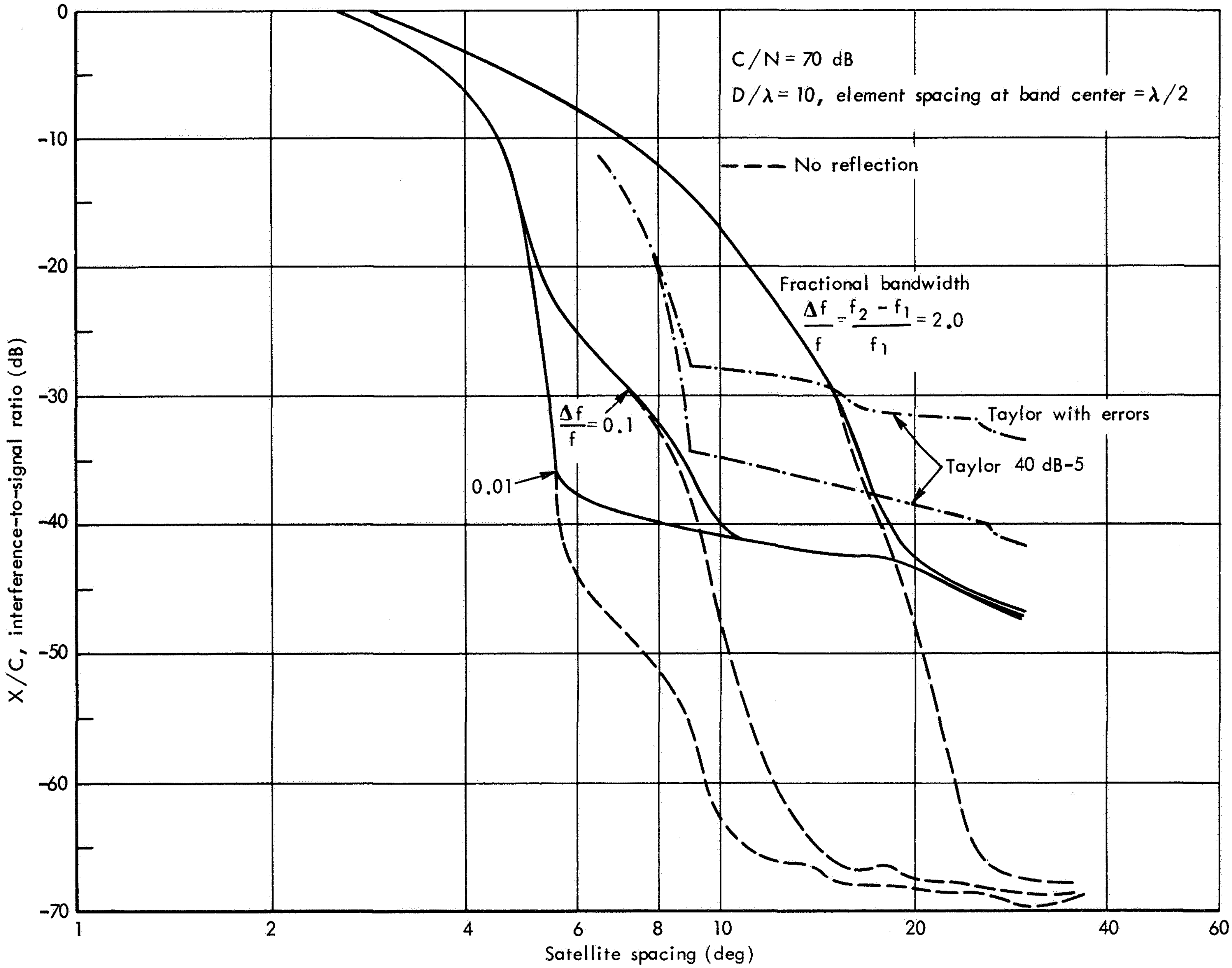

Fig. 4-2 — X/C versus satellite spacing for adaptive arrays

This behavior is illustrated for the adaptive arrays of Fig. 4-2, where receiver noise is assumed to be 70 dB below the signal.

Some general features of the curves for conventional antennas should be pointed out. If the interfering satellites are equally spaced and the closest ones are in directions corresponding to a flat portion of the antenna pattern (see figures in Appendix A), then X/C will fall off with satellite spacing σ approximately as $1/\sigma$. This is due to the change in the number of satellites contributing from that portion of the pattern and is illustrated in the region of the flat shelf for the Taylor 40 dB-5 antennas. On the other hand, if the antenna patterns themselves fall off rapidly with angle in the region of the nearest interfering sources, then these closest sources will dominate the total interference and X/C will fall off approximately as the antenna pattern. This is the case for the "linear" ($\sim 1/\sigma^2$), the CCIR ($\sim 1/\sigma^{2.5}$), and the "uniform" ($\sim 1/\sigma^3$) antennas as is illustrated in Fig. 4-1.

The $1/\sigma^3$ behavior is typical of many situations of interest and will be used to illustrate a variety of effects of antenna patterns on satellite spacing. For example, when $X/C \propto 1/\sigma^3$, the interference contribution of all the satellites beyond the nearest neighbors on each side of the wanted satellite is less than 0.2 of that from the nearest neighbors (i.e., $M_{s/e} < 2.4$) for uniform spacing and EIRP.

It is clear from Fig. 4-1 (and similar illustrations for other D/λ in Appendix A) that both D/λ and the type of illumination are important in determining the satellite spacing performance of the system. In order to specify this performance with a single measure, an equivalent uniform size $(D/\lambda)_u$ will be introduced. This is the equivalent uniform illumination circular aperture that will permit the same satellite spacing performance. It can be obtained by multiplying the actual D/λ by the spacing factor required to shift the actual X/C curve to the spacing indicated for the uniform antenna for the same desired value of X/C. For example, in Fig. 4-1 for X/C = -30 dB, the spacing value of 1.0° for the parabolic antenna must be multiplied by 0.8 to obtain the spacing for the uniform antenna for X/C = -30 dB. Thus the equivalent uniform size of the parabola would be $(D/\lambda)_u = 0.8(300) = 240$ at X/C = -30 dB.

Tradeoffs Involving Bandwidth Expansion

It may be desirable to use modulation methods that increase bandwidth expansion W/B in order to reduce the total power requirements or to increase the total information capacity of the system. For example, greater bandwidth expansion would permit lower C/X values but would reduce information transfer with a given spectrum per satellite. However, the lower permissible C/X values would make possible closer satellite spacings, and hence a larger number of satellites. Whether or not a greater total information capacity becomes possible depends on whether the X/C changes more for a given percentage change in satellite spacing or in W/B.*

Referring to Fig. 2-1 shows that there is no advantage in going from SSB or DSB to low index FM ($D < 1.8$) for more total information capacity or less total power (no portion of any curve of X/C versus σ in Fig. 4-1 has a negative slope of less than unity, so there will be no information capacity advantage in going to a greater W/B point in Fig. 2-1 unless it is above a slope of unity, i.e., above the reference line when referred to SSB or DSB). However, there can be an increase in total information capacity by greater bandwidth expansion in either FM or PCM if the fractional decrease in satellite spacing exceeds the fractional increase in W/B. If it is desirable for other reasons to use FM or PCM, then it will be advantageous for both total power and information capacity to increase W/B until threshold is reached for the output message quality desired or until the satellite spacing crowds the nearest unwanted satellites up to the mainlobe of the antenna. This advantage occurs because the slope of the FM transfer characteristic of Fig. 2-1 is greater than 3 and no portion of any curve outside of the mainlobe in Fig. 4-1 is steeper than -3.

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*Lower X/C values would also permit greater multiplication of the independent beams possible with directive antennas on board the satellites. If this technique is used it will nearly always be advantageous to reduce C/X through bandwidth expansion in order to increase the total information capacity of all the satellites. However, the justification developed at this point for greater W/B is based entirely on an increase in the number of satellites.

The above advantages for greater W/B are heightened by three other factors. First, if large enough W/B is used to reduce the required C/X to less than 20 to 30 dB, it should be possible to achieve enough polarization discrimination to use two polarizations independently and thus double the effective spectrum available. Second, if the frequencies of alternate satellites are interleaved by one-fourth to one-half of W, improvement factors even greater than for the reference white-noise case could be achieved, making increased W/B seem even more attractive. Third, the advantages of increasing the number of independent beams possible from each satellite could dominate all other factors.*

If the net increase in total system capacity achieved by increasing W/B is small, its cost may seem high, since the increase in the number of satellites must be greater than the increase in W/B, and the cost of the satellite component of the system will be approximately proportional to the number of satellites. Some increases in total system capacity can be achieved only at greater cost per circuit. Thus if there is no crowding of orbital space, or if only a very few satellites are used with any earth station, it would probably be less costly to minimize the number of satellites and to not increase W/B for greater total combined orbital capacity. This may also be the case for the earth-station receiving antenna size. When many earth receiving stations are involved in a system, the total system cost can usually be reduced by decreasing the size and cost of the earth-station receiving antennas. However, when the earth antennas are used to isolate the signals from different satellites, reducing antenna size will be in conflict with achieving closer satellite spacing and greater orbital capacity. It is important, therefore, in establishing satellite systems to use values for parameters such as W/B and earth-station antenna size that will permit the orbital capacity growth needed for all the systems that will share the orbit and spectrum during the projected life of the system.

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*See footnote on previous page.

Differences in Satellite EIRPs

For this mode of interference it is clear from Eq. (4-1) that only the relative values of EIRPs are involved in determining X/C, i.e., it is independent of the absolute values of the satellite EIRPs. If all EIRPs are multiplied by the same factor, both X/C and the required satellite spacings remain unchanged. If the EIRPs differ for satellite systems sharing a given orbit and spectrum, the exact solution for the minimum orbital spacings becomes very complicated. However, when the nearest neighbors dominate the interference (as was previously indicated to be the case in most situations of interest, and particularly if $X/C \propto 1/\sigma^3$), a simple approximate expression is obtained. In its simplest form it applies when the same protection ratios, W/B, and interleaving, are used with the adjacent satellites so that the X/C required is that obtained with white noise. The spacing required between any two satellites depends primarily on the spacing quotient which involves $(D/\lambda)_e$ and the ratio of the EIRPs of the two satellites. The larger the spacing quotient the smaller the spacing that is possible.

$$\left.\begin{aligned} \sigma_{a,b} &= \sigma_o (D/\lambda)_{uo}/q_{a/b} \\ q_{a/b} &\equiv (D/\lambda)_{ua}\rho_{a/b}^{1/3} \leq (D/\lambda)_{ub}\rho_{b/a}^{1/3} \equiv q_{b/a} \end{aligned}\right\} \qquad (4\text{-}2)$$

where $\sigma_{a,b}$ is the spacing needed between adjacent satellites s_a and s_b; σ_o is the spacing needed if all satellites were equally spaced with equal EIRP and all the earth stations had equal equivalent uniform antenna size $(D/\lambda)_{uo}$; $\rho_{a/b}$ is the ratio of the EIRP of s_a to that of s_b in the passband of s_a; and $(D/\lambda)_{ua}$ is the smallest equivalent uniform size antenna using s_a. The satellite with the smaller spacing quotient $q_{a/b}$ is the weaker system to be protected. It is designated s_a and the other s_b. Equation (4-2) applies only in the direction giving the larger spacing, i.e., protecting the weaker system s_a.

Equation (4-2) is most applicable to situations where the interference contribution from the two sides of s_a are comparable, i.e., when the spacing quotients in the two directions are comparable. If there is a sequence of satellites of ascending or descending strength so that a satellite within the sequence experiences most of its interference from only one side, then the spacing toward the interference side might be reduced to as little as 0.8 of that indicated by Eq. (4-2) when essentially all the interference is from the nearest neighbor on that side. However, for a long series of ascending satellite strengths, satellites beyond the nearest neighbor may also contribute significant interference which should be taken into account in determining the spacing.*

Thus, as a good approximation, the spacing required between any two adjacent satellites will differ from the reference spacing σ_o only if there is a difference in the spacing quotient of the adjacent satellites or their other nearest neighbors. In most cases there will be a greater spacing than σ_o required between satellites of different spacing quotients in order to protect the performance of the system using the satellite with the smaller spacing quotient. Thus, if a group of satellites each with the same spacing quotient is to be mixed with another group of satellites with a different common spacing quotient, the most efficient use of orbit space would minimize the number of adjacent satellites with different spacing quotients.

For example, if there are two groups each of n satellites with all parameters being identical except for a 9-dB difference in the EIRPs of the satellites in the two groups, and if the two groups are placed adjacent to each other so that there is only one pair of adjacent satellites

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*A simple procedure for determining the approximate minimum spacings for any specified ordering of satellites is to begin at a point of maximum satellite strength, i.e., equal or smaller spacing quotients from both sides. Then taking either nearest neighbor, the spacing to it can be determined by using Eq. (4-2) with s_a as the weaker satellite and s_b as the stronger one. If the next satellite s_c on the other side of s_a is significantly weaker than s_a, then σ_{ab} could be reduced by a factor as small as 0.8 if s_c makes no significant interference contribution to the use of s_a. In the above situation if σ_{ca} is now determined by Eq. (4-2) it may be necessary to increase this spacing somewhat because of the interference contribution of s_b. It may also be possible to decrease it slightly if there is no significant interference contribution from the other side of s_c. This procedure can be continued until a satellite with minimum spacing quotient has been reached after which the procedure should be initiated again at a point of maximum satellite strength. In this way all the spacings can be determined to give the approximate minimum that could be achieved with the specified ordering of the satellites.

with different EIRPs, the minimum orbital space occupancy would be approximately $2n\sigma_o$ between the first and last of the 2n satellites. On the other hand, if the satellites of the two groups were interleaved so that there was a 9-dB difference in EIRP between every pair of adjacent satellites, all the spacings would need to be $2\sigma_o$, and the total orbital occupancy would be $4n\sigma_o - 2\sigma_o$, or nearly twice that of the previous case for $n >> 1$.

As a rough general rule, if there are uncertainties about the relative EIRPs or spacing quotients of adjacent satellites, and assuming $X/C \propto 1/\sigma^3$, the spacing allowance that should be made between the two satellites is a doubling for each 9-dB equivalent of EIRP difference possible. For example, if an 18-dB difference is possible, the spacing should be approximately $4\sigma_o$.

Effects of Satellite Antenna Patterns

So far satellite spacing has been considered in conjunction with directive earth-station receiving antennas in order to separate or isolate links, so that any given spectrum can be used independently many times (each time with a different satellite) for the same or different earth locations. A similar type of link isolation can be provided with directive antennas on board the satellites. These may be used to augment the isolation provided by the earth-station antennas (by reducing the EIRP of unwanted satellites in the direction of the particular earth station). Or it may be used to provide enough isolation to earth stations at different locations so that the same spectrum can be used independently many times, in this case from the same or different satellites to different earth locations. Since the isolations by the earth-station antennas and by the satellite antennas can be made independently, the multiplication of spectrum use can be made a product of the two.

The total angle over which the satellites can be spaced from a single earth location is less than π radians in <u>one</u> dimension. The total angle over which the beams from one satellite can be spaced over the earth's surface is less than 1/3 radian in <u>two</u> dimensions. Thus for a given size antenna, a greater spectrum multiplication factor is potentially available from a synchronous satellite than from an earth station for D/λ greater than about 100. This is one reason why large antennas on board the satellites will become attractive for multiplying the effective spectrum available for communication relay. In addition, the large satellite apertures enable a given EIRP from the satellite with less transmitted power, and allow the up link to operate with less earth station EIRP and interference potential to terrestrial systems.

Margins and Protection Ratios Required

It is necessary to add a variety of margin factors to the ideal C/X required at the input to the receiver in order to arrive at a satisfactory practical protection ratio for operation against this interference mode:

- Satellite position uncertainties can very easily be converted into a power margin if X/C is proportional to some exponent of σ. For example, if $X/C \propto 1/\sigma^3$, and if every satellite can independently vary about its assigned longitudinal position by only a small percentage of the nominal spacing, X/C can have as much as three times this percentage change. In other terms, independent longitudinal variations about assigned longitudinal positions of 8 percent of the nominal spacing can be accommodated by adding a 1-dB additional margin to the protection ratio. Future control capability should provide such accuracy without excessive cost. The 1-dB margin should in fact be adequate for this factor and should include the effects of any uncertainties in antenna pointing.
- Satellite antenna gain may vary by as much as 3 dB over the designed coverage, especially if narrow beams are employed and many beams are used to cover an area. Thus at least a 3-dB margin is needed for this factor.
- The satellite relay power output may "age" or differ in performance over its design life to provide an uncompensated variation in output power between satellites which could probably

be limited to 2 dB without excessive design cost. Hence a 2-dB margin could be allowed for this factor.

- The probability of the sidelobe peaks rather than average value effectively determining the interference may justify adding a margin (see Section III) to the ideal X/C when using average value curves to determine satellite spacing. This margin should vary from about 2 dB for closest spacing near the mainlobe to 0 dB at greater than 10 times this spacing.
- Earth-to-satellite geometry factors may require an additional 1-dB margin (see Section III).
- Propagation attenuation and fading may require a margin factor for this mode of interference. However, it needs only accommodate the difference in attenuation and fading between the wanted satellite and the sum of the two nearest unwanted satellites. Below 10 GHz this difference is negligible and even at higher frequencies where large attenuations may be experienced, the difference is likely to be only a very small fraction of the total attenuation if the satellites are closely spaced.
- The protection ratios required for this interference mode can now be determined by adding the appropriate margin (e.g., 8 dB) to the input C/X values for any specified system. Using the values indicated in Table 2-2 for INTELSAT III and COMSAT's Pilot Domestic Satellite proposal and the X/C versus satellite spacing curves of Appendix A, the spacing values indicated in Table 4-1 are obtained. The "parabolic" antenna curves were used, and the $(D/\lambda)_e$ values indicated in Table 4-1 were assumed for the nominal antenna sizes of the earth receiving stations. The values for satellite spacing with minimum C/X RFI emphasis were obtained by adding an 8-dB margin to the C/X minimum for a message objective partitioning emphasizing RFI from Table 2-2. With interleaving of frequencies the W/B required to achieve the C/X minimum with white noise would be a conservative value. Table 4-1 illustrates that different services with the same satellite may result in different allowable spacings. For the pilot domestic program, the trunk message service to the larger

antennas requires a larger spacing than TV to the smaller antennas because of the smaller bandwidth expansion used for the trunk message service.

Table 4-1

ILLUSTRATIVE SATELLITE SPACINGS

System	$(D/\lambda)_e$	RFI X/C (dB)	RFI Protection Ratio (dB)	Satellite Longitudinal Spacing (deg)	Satellite Spacing with Minimum C/X RFI Emphasis (deg)
			INTELSAT III		
Message	320	-18.5	26.5	0.7	0.39
TV #3 Quality	320	-23	31	1.15	0.39
			Pilot Domestic Proposal		
Trunk Message	320	-42	50	5.7	0.39
TV	100	-25	33	4.1	1.25

Summary for this Mode of Interference

The characteristics of this mode of interference and the constraints it imposes for greatest capacity may be summarized as follows:

The Earth Stations. They should be coordinated to minimize differences in spacing quotients. This will establish some minimum $(D/\lambda)_u$ for a given usage. Making the minimum $(D/\lambda)_u$ as large as possible is especially important in determining the minimum satellite spacing achievable and consequently the total information capacity of a given spectrum for synchronous satellite relaying. This mode of interference is independent of the number or relative location of the earth receiving stations or coordination of their use; i.e., each receiving earth station needs to coordinate only with the transmitting earth station as to which satellite it is to use and when it is to use it.

The Communication Satellite System. The modulation properties of the signals should be coordinated for greatest benefit, which would involve using comparable bandwidth expansions and interleaving the frequencies between adjacent satellites. Also, increasing the bandwidth expansion ratio to enable the use of the smallest practical protection ratios will usually permit closer satellite spacings and more independent beams from the satellite and thus will make important increases possible in the total information capacity of a given spectrum for synchronous satellite relaying. In addition, the opportunity exists for two independent systems and allocations on two independent polarizations if the systems are designed to use ideal C/X ratios smaller than 20 to 30 dB.

The Satellites. This interference mode is independent of the absolute value of the satellite EIRP as long as it dominates the unwanted signals. The satellite spacing needed is determined by the spacing quotient, which depends on the relative values of the EIRPs of adjacent satellites. It is therefore most efficient in orbital space consumption to order the satellites so as to minimize the number of satellite adjacencies with different spacing quotients--to avoid, as much as possible, interleaving of disparate satellite systems if the spacing quotients for the satellite systems sharing a given spectrum cannot be standardized. The use of high-gain antennas on board the satellites would allow the isolation of many links to earth; this would permit a further important multiplication of the information capacity of a given spectrum from each satellite. It should require only 1 dB of margin to accommodate up to 8 percent of the nominal spacing of independent variation about assigned longitudinal positions.

INTERFERENCE FROM EARTH STATIONS TO SATELLITES, MODE 2

This interference mode involves interference from unwanted earth stations to the satellite receiving the wanted earth-station signal. Again, following the notations and procedures indicated in Section I,

$$\frac{\sum_{e'} X_{e's}}{C_{es}} = \frac{P_{e'}\, G_{e'\to s',s}}{P_e\, G_{e\to s,s}} \cdot \frac{M_{e's}\, d^2_{es}\, F_x}{d^2_{e's}\, F_c} \cdot \frac{G_{s\leftarrow e,e'}}{G_{s\leftarrow e,e}} \simeq \frac{M_{e's}\, G_{e'\to s',s}}{G_{e\to s,s}} \qquad (4\text{-}3)$$

The simple approximation is very nearly the same as Eq. (4-1) and it holds under a similar set of conditions. However, there is an important distinction in that this relation involves the ratio of the gains of separate earth-station transmitting antennas, whereas Eq. (4-1) involves the ratios of the gains of the same earth-station receiving antenna in the various directions of the satellites. This distinction will probably be of most concern in establishing margins for propagation attenuation. Since the propagation paths in the earth's atmosphere will be quite different for the illuminations of adjacent satellites in this case, the full magnitude of the possible variations in propagation attenuation must be added as a margin in the protection ratio for this mode of interference. This may mean that a significantly larger spacing quotient,* or $(D/\lambda)_u$, will be required in the up link than in the down link if the satellite spacing is not to be limited by the up link at frequencies above 10 GHz. Most of the other characteristics discussed for the reciprocal interference mode have application to this mode also and they will only be summarized here.

Earth Stations. This mode depends only on the relative and not the absolute values of the earth-station EIRPs as long as it dominates the unwanted signals. The earth stations should be coordinated to minimize differences in spacing quotients. This will establish some minimum $(D/\lambda)_u$ for a given usage. Making the minimum $(D/\lambda)_u$ large is especially important in determining the minimum satellite spacing achievable. It should be designed so as to be compatible with the spacing determined for the reciprocal link. In the 4- and 6-GHz bands the frequency of the up link being higher than the down link provides enough larger D/λ, using the same size dish to compensate for the margin allowance difference for precipitation attenuation in the two cases. Thus compatible satellite spacing can be achieved for both modes with the same earth-station antenna. This may not be possible for other situations involving higher frequencies or smaller ratios of frequencies, up link to down link. The EIRP per unit of bandwidth for all earth-station transmitters using the same satellite transponder should be equal. The transmissions should be coordinated between earth stations

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*The spacing quotient is defined in the same way as for the reciprocal case of Eq. (4-2), except that the EIRPs involved are those illuminating the respective satellites rather than emanating from them.

so that each satellite is illuminated only once with the full EIRP in each frequency band. The total wanted information bandwidth transmitted to each satellite cannot exceed the wanted information bandwidth relayed back toward earth.

The Communication Satellite System. The system characteristics indicated in the summary of the reciprocal mode also apply to this mode, including a requirement that the characteristics for the two modes must be mutually compatible. However, there is a further qualification on one feature, i.e., the opportunity exists for two independent systems and allocations on two independent polarizations if the systems are designed for protection ratios smaller than 20 to 30 dB. In this case it is the protection ratio, including all the margins (particularly for propagation attenuation), that is involved rather than the ideal X/C ratios, unless the two polarizations over any frequency band emanate from the same earth station. For the latter situation the specifications revert to those of the reciprocal mode, i.e., involving ideal X/C ratios. For frequencies above 10 GHz it would probably not be feasible in most cases to use two polarizations independently on the same frequency for the up link unless they emanate from the same earth station. Thus this condition might be an appropriate specification on the use of two independent polarizations.

The Satellites. The same satellite characteristics as those previously indicated in the summary of the reciprocal mode also apply to this mode, except that the wanted EIRP illuminating a satellite should now be substituted for the EIRP emanating from that satellite in the reciprocal mode.

OTHER DIRECT INTERFERENCE MODES

The seven other direct interference modes are less complicated and need less elaboration than the two already treated, which are highly involved with satellite spacing. These remaining modes will be treated very briefly, only by highlighting those features which might significantly constrain one or more of the parameters or limit frequency-sharing operations.

Satellite-to-satellite Interference, Mode 3

The equation for interference mode 3 is

$$\frac{\Sigma_{s'} X_{s's}}{C_{es}} = \frac{P_{s'}\, G_{s'\to e,s}}{P_e\, G_{e\to s,s}} \cdot \frac{M_{s's}\, d^2_{es}\, F_x}{d^2_{s's}\, F_c} \cdot \frac{G_{s\leftarrow e,s'}}{G_{s\leftarrow e,e}} \tag{4-4}$$

This mode can be significant only if the same frequency bands are used for both up links and down links. However, as will be indicated later, such operations may be practical for some future applications and would double the bandwidth available in prime portions of the spectrum. Although they may not be currently designed to operate in this manner, it would be relatively straightforward to design the satellite antennas so that this mode of interference need not limit operations of this type. For example, $G_{s'\to e,s} \simeq G_{s\leftarrow e,s'} \lesssim 10^{-1}$; $G_{s\leftarrow e,e} \gtrsim 10^{2}$; $G_{e\to s,s} \gtrsim 10^{5}$; $P_{s'} \simeq P_e$; $M_{s's}\, F_x/F_c \simeq 2$; and $\Sigma_{s'} X_{s's}/C_{es} \simeq 10^{-3}$. Then $d_{s's} \gtrsim d_{es}\, \sqrt{2}/10^{3}$, or the minimum satellite spacing could be less than 0.1° longitude, which is considerably less than the spacing limitations imposed by other interference modes for the foreseeable future. Analysis of the other interference modes will demonstrate that this mode may be dismissed without imposing conflicting constraints on any of the system parameters.

Interference Mode 4 Between Earth Stations

The equation for interference mode 4 is

$$\frac{\Sigma_{e'} X_{e'e}}{C_{se}} = \frac{F_{e'}\, G_{e'\to s,e}}{P_s\, G_{s\to e,e}} \cdot \frac{M_{e'e}\, d^2_{se}\, F_x}{d^2_{e'e}\, F_c} \cdot \frac{G_{e\leftarrow s,e'}}{G_{e\leftarrow s,s}} \tag{4-5}$$

Again, this mode is of no consequence unless the same frequency bands are used for both up links and down links. However, it will be shown that this mode of interference could be tolerated within the constraints that other interference modes will impose anyway for reducing earth station EIRP and increasing satellite EIRP. Thus this mode of interference would not prevent the shared use of a given frequency band for both up links and down links. In order to illustrate this, first consider only one other interfering earth station so that $M_{e'e}\, F_x/F_c \simeq 1$. If ${}_{e'\to s,e}G \simeq {}_{e\leftarrow s,e'}G \lesssim 10^{-1}$; ${}_{s\to e,e}G \gtrsim 10^2$; ${}_{e\leftarrow s,s}G \gtrsim 10^5$; $P_{e'} \simeq P_s$; and $X_{e'e}/C_{se} \simeq 10^{-3}$ then $d_{e'e} \gtrsim d_{se} \cdot 10^{-3}$ so the minimum earth station spacing could be less than 40 km. If the product of the satellite transmitting antenna gain and the earth-station receiving antenna gain is increased by a factor of ten, this mode of interference could be tolerated for $M_{e'e}\, F_x/F_c \simeq 10$ with a minimum earth-station spacing of 40 km and a maximum ideal density of these earth stations in the reference ground-environment model described in Section III. The constraints from this mode of interference are subordinate to those which will be indicated for modes yet to be discussed if some minimum earth-station spacing such as 40 km is acceptable for the applications under consideration. For mobile and other applications where it may not be possible to limit the minimum transmitter and receiver separations in this way, it may not prove feasible to operate with up links and down links in the same frequency band--i.e., the down link may need to exclude possibilities of close approaches between transmitters and receivers in the same frequency band.

Interference from Terrestrial to Earth Stations, Mode 5

This interference mode, involving interference from microwave relays to earth stations, is very similar to the previous one between earth stations:

$$\frac{\sum_{t'} X_{t'e}}{C_{se}} = \frac{P_{t'}\ {}_{t'\to t,e}G}{P_s\ {}_{s\to e,e}G} \cdot \frac{M_{t'e}\, d_{se}^2\, F_x}{d_{te}^2\, F_c} \cdot \frac{{}_{e\leftarrow s,t'}G}{{}_{e\leftarrow s,s}G} \qquad (4\text{-}6)$$

If $P_{t'}\ {}_{t'\to t,e}G \lesssim P_{e'}\ {}_{e'\to s,e}G$ the same minimum spacing could be achieved with d_{te} as with $d_{e'e}$ for the previous mode. Thus the same constraints on the product of the satellite transmitting antenna gain and the earth-station receiving antenna gain apply for this case, as do the limitations for possible applications.

Interference Mode 6 Between Terrestrial Microwave Relay Stations

In order to illustrate the extreme limiting case for interference potential, the ground geometry model described in Section III for maximum packing density of microwave relays will be assumed for this interference mode. For convenience in illustration for some of the other modes, the minimum separation distance d_{tt} between stations will be assumed to be 40 km or 0.001 of the distance to synchronous orbit. For the maximum theoretical density each station has 6 nearest neighbors at spacings of 40 km. The wanted signals involve one or two of these nearest neighbors while all the remaining stations can contribute unwanted signals with an upper bound to the rate of accumulation of these signals with distance behaving as indicated in Fig. 3-3. The equation describing this interference mode is

$$\frac{\sum_{t'} X_{t't}}{C_{tt}} = \frac{P_{t'}\; G_{t'\to t',t}}{P_t\; G_{t\to t,t}} \cdot \frac{M_{t't}\; d_{tt}^2\; F_x}{d_{t't}^2\; F_c} \cdot \frac{G_{t\leftarrow t,t'}}{G_{t\leftarrow t,t}} \qquad (4\text{-}7)$$

For the ideal geometry and equal transmitting powers this X/C is independent of the power outputs. Also, it is relatively independent of separation distance d_{tt} as long as the horizon limitations maintain nearly constant $M_{t't}$. For most microwave systems such as the TD-2 and TH,* a generous allowance would give

$$\frac{\sum_{t'} X_{t't}}{C_{tt}} \cdot \frac{d_{t't}^2\; F_c}{M_{t't}\; d_{tt}^2\; F_x} \approx 10^{-8} \approx \left(\frac{{}_tG_{t,t'}}{{}_tG_{t,t}}\right)^2 \qquad (4\text{-}8)$$

---

*See Section II for protection ratios and Section III for geometric factors.

This means that the ratio of main-beam-to-far-sidelobe antenna gains should be greater than 40 dB. Such a condition can be satisfied for antennas with $G \gtrsim 30$ dB and is better assured the larger the gain. Most current microwave systems conservatively satisfy this condition.

Interference Mode 7 from Earth to Terrestrial Stations

The extreme illustration of this mode of interference would be to replace all the unwanted microwave relay stations of the model for the previous mode with earth stations. Then Eq. (4-7) would be transformed to

$$\frac{\sum_e X_{et}}{C_{tt}} = \frac{P_e \; G_{e\rightarrow s,t}}{P_t \; G_{t\rightarrow t,t}} \cdot \frac{M_{et} \; d_{tt}^2 \; F_x}{d_{et}^2 \; F_c} \cdot \frac{G_{t\leftarrow t,e}}{G_{t\leftarrow t,t}} \tag{4-9}$$

and, following Eq. (4-8),

$$\frac{P_e}{P_t} \; \frac{{}_eG_{s,t} \; {}_tG_{t,e}}{{}_tG_{t,t}^2} \simeq 10^{-8}$$

Since ${}_eG_{s,t} \simeq {}_tG_{t,e} \simeq 10^{-1}$ and ${}_tG_{t,t}^2 \simeq 10^8$ for most current systems in the bands of interest,

$$P_e \lesssim 10^2 \; P_t$$

This condition can be satisfied for future systems, and probably will need to be satisfied anyway to avoid excessive precipitation scatter interference in shared bands at higher frequencies.

Interference Mode 8 from Satellites to Terrestrial Stations

Using the same ground geometry as in modes 6 and 7 for maximum ideal density of microwave relays, the equation for mode 8 becomes

$$\frac{\sum_s X_{st}}{C_{tt}} = \frac{P_s \, G_{s\to e,t}}{P_t \, G_{t\to t,t}} \cdot \frac{M_{st} \, d_{tt}^2 \, F_x}{d_{st}^2 \, F_c} \cdot \frac{G_{t\leftarrow t,s}}{G_{t\leftarrow t,t}} \tag{4-10}$$

This mode is most easily illustrated with a specific microwave relay system. The TD-2 system will be used for this purpose, and it will be assumed that first,

$$\frac{d_{tt}^2 \, F_x}{d_{st}^2 \, F_c} \lesssim 10^{-6}$$

since the protection ratio has already included a margin for F_c; second, that the TD-2 antennas avoid pointing at synchronous orbits so that at any single TD-2 receiving antenna, $G_{t\leftarrow t,s}$ will never be greater than 10 toward any satellite, will be greater than 1 for less than 10 satellites, and will be greater than 10^{-1} for less than 100 satellites, so that $M_{st} \, G_{t\leftarrow t,s} \lesssim 10$; third, that $P_t \, G_{t,t}^2 \approx 3 \times 10^8$ watts over a 500 MHz band; and fourth, that (from Section II) the protection ratio for this mode is less than 60 dB. Then <u>each</u> satellite EIRP in the 500-MHz band (with interleaving of frequencies so as to be the equivalent of white noise) is limited to

$$P_s \, G_{s\to e,t} \lesssim 3 \times 10^7 \text{ W} \quad \text{or} \quad 75 \text{ dBW in 500 MHz}$$

A similar exercise for the TH system yields

$$P_s \, G_{s\to e,t} \lesssim 10^8 \text{ W} \quad \text{or} \quad 80 \text{ dBW in 500 MHz}$$

Current CCIR recommendations for maximum power flux density produced at the surface of the earth from a space station provide a limit equivalent to 62 to 68 dBW for the EIRP for the same conditions.

Interference Mode 9 from Terrestrial Stations to Satellites

The equation for this interference mode is

$$\frac{\sum_t X_{ts}}{C_{es}} = \frac{P_t \, G_{t\to t,s}}{P_e \, G_{e\to s,s}} \cdot \frac{M_{ts} \, d^2_{es} \, F_x}{d^2_{ts} \, F_c} \cdot \frac{G_{s\leftarrow e,t}}{G_{s\leftarrow e,e}} \tag{4-11}$$

In order to illustrate the requirements on earth station EIRP, it will be assumed that $(d^2_{es} \, F_x)/(d^2_{ts} \, F_c) \simeq 1$, and $(\sum_t X_{ts})/(C_{es}) = 10^{-3}$. Then

$$P_e \, G_{e\to s,s} \gtrsim P_t \, G_{t\to t,s} \cdot \frac{M_{ts} \, G_{s\leftarrow e,t}}{G_{s\leftarrow e,e}} \cdot 10^3$$

If it is specified that the average value of $G_{t\to t,s}$ will be less than unity for the microwave relays within the area on the earth's surface covered by the satellite receiving beamwidth, and that less than 10 percent will have values as high as 10, the following constraints are obtained:

$$\frac{P_e \, G_{e\to s,s}}{P_t} \gtrsim 10^3 \cdot \frac{M_{ts} \, G_{s\leftarrow e,t}}{G_{s\leftarrow e,e}} \tag{4-12}$$

where M_{ts} is the number of microwave relays within the area covered by $G_{s\leftarrow e}$.

Thus with full earth coverage for $G_{s\leftarrow e}$, and with $G_{e\to s,s} \simeq 10^6$, $P_e/P_t \gtrsim 10^{-3} \, M_{ts}$. Then if $M_{ts} > 10^3$, P_e must be greater than P_t in their shared frequency band. On the other hand, to illustrate the effect of large $G_{s\leftarrow e}$, with 50-dB antenna gain (30 dB greater than for full earth coverage) only 200 microwave relays would be within the beamwidth for maximum theoretical density with a microwave relay spacing of 40 km. Then $(P_e \, G_{e\to s,s})/(P_t) \gtrsim 2 \cdot 10^5$, so that P_e could be made equal to P_t with a more modest value of $G_{e\to s,s}$.

SCATTER MODES OF INTERFERENCE

The precipitation scatter modes of interference are the most difficult to evaluate as to constraints on system design to meet performance objectives for satellite and microwave relay systems sharing the same frequency bands. The principal reason for this is that the expected spatial distribution of precipitation and its statistics for the localities of interest are essentially unknown, particularly at the higher altitudes. It is not too difficult to parameterize the scatter interference in ways that bound the effects so as to considerably reduce uncertainties, and this has been done in Appendix C. However, there still may be order-of-magnitude uncertainties in estimates that might be made about the influence of precipitation scatter interference in any locality.

The potentially most important geometry for significant scatter interference between satellite and microwave relay systems is for coupling through a common volume of main-beam intersection as illustrated in Figs. 1-1 and C-4. In this geometry the microwave relay beams are close to the horizon, whereas satellite relay beams are likely to be a good fraction of 90 minus latitude in degrees above the horizon, and any common volume of intersection is apt to be considerably closer to the earth station than to the microwave relay.

As a limiting model for maximum potential opportunity for intersections, consider the maximum theoretical density of microwave relays previously discussed with a spacing of 40 km and with an earth station substituted for one of the microwave relays. If the microwave relays each have two back-to-back beams with beamwidths of 1 deg which are randomly oriented in azimuth, their main beams would occupy less than 1/3 of 1 percent of the atmospheric volume at 0.1-km altitude but would occupy almost all of it above 13-km altitude. The probability of main-beam common volume for a vertical earth-station beam would approach unity by an altitude of about 7 km (350 km to microwave relay station), and for a beam 30° above the horizon by an altitude of about 3.5 km (240 km to microwave relay station). Thus it would probably be feasible to coordinate site locations and beam orientations to avoid main-beam intersections within two station spacings separation and below 1-km altitude. However, beyond these ranges and altitudes the coordination to avoid intersections is likely to be more difficult than designing the systems to be compatible with main-beam intersections. It is with this background that the scatter interference through main-beam intersection is investigated in Appendix C.

The scatter interference effects are calculated as a function of rain rate, separation between stations, and frequency. Two extreme spatial distributions are considered: one, that would maximize the scatter interference effects, assumes that the rain is contained in the common volume of main-beam intersection only; and the other, that would dampen any scattering effects by attenuation, assumes that the rain rate exists throughout the entire volume. A simple procedure is described in Appendix C which permits interpolating the effects of any mixture between these two extremes. The important questions are: What is the worst mixture that occurs the maximum acceptable fraction of the time for any locality? How do the effects vary with frequency? With station separation? With rain rate?

If an arbitrary, though plausible assumption is made that for no consequential period of time will there exist significant precipitation in a common volume of main-beam intersection without the presence of the equivalent of an average of at least 1 to 10 percent of the common-volume rain rate throughout the volumes between the intersection and the stations, a fairly simple general conclusion is obtained: Attenuation dominates the scatter. For increasing rain rate, station separation, and frequency above small values of these parameters, scatter interference has decreasing influence.

Even if the assumption about rain distribution does not hold in practice, there appears to be enough leeway in feasible satellite aperture and allowable EIRP to design around the scatter interference problem for the densest sharing environment with 40-km spacing between microwave relays. However, experimental data are needed to better establish the design specifications that will ensure the compatible sharing of the spectrum as constrained by scatter interference in a particular location.

There is some risk of the momentary effects of scatter interference from foreign objects (birds, aircraft, etc.) in the common volume of main-beam intersection. By coordinating siting to avoid low-altitude ($\lesssim$ 300 m) common volumes, the risk from birds and most frequent small object interceptions can be controlled. To minimize the brief high-lighting effects of large aircraft which could be troublesome out to ranges as great as 40 km, the earth stations should be located to avoid intercepting the busier air lanes.

COMBINED EFFECTS OF INTERFERENCE FROM ALL MODES

The constraints on the system parameters for tolerable interference in each interference mode have been determined, allowing the conservative protection ratios for terrestrial microwave relays that are used in current practice. For the satellite relays and earth stations, the protection ratio assumed for determining parameter constraints was 30 dB, which is about 10 dB greater than is needed for large bandwidth expansion ratios and conservative margin allowances. With these assumptions the combined effects of interference from all modes may be interpreted as follows. Modes 1 and 2 involve only the satellite and earth-station system. Increasing capacity benefits can be derived for increased $(D/\lambda)_e$, for increased RTC and bandwidth expansion, and for increased $(D/\lambda)_s$. It will also be advantageous for achieving greater capacity to use each frequency band independently on two polarizations and for both up links and down links. The coordination and standardization of the orbital and spectrum usage will also greatly enhance the benefits that can be derived. Mode 3 primarily involves the satellites which can easily be designed to make this interference mode of no consequence. Modes 4, 5, and 7 are best handled by reducing P_e to be comparable with P_t in its bandwidth and consequently increasing the satellite apertures. Mode 6 involves only the microwave relays, and conventional design practice uses large enough ${}_tG_t$ to conservatively take care of this problem. Modes 8 and 9 require coordination of the pointing of microwave relay antennas to avoid directions of synchronous satellites. Mode 9 may also require a conflictingly large value of P_e

unless larger apertures than for earth coverage are used on the satellites. With the pointing coordination of microwave relay antennas assumed for modes 8 and 9, considerably larger EIRPs could be allowed per satellite than current CCIR recommendations without changing the noise objectives or system design of microwave relay practice. The interference from scatter modes can be most easily alleviated by increasing satellite EIRP and decreasing earth-station transmitting power. However, more experimental data are needed to specify system design that will ensure adequate performance in the presence of scatter interference.

Thus, most of the burden of accommodation to an intense sharing between microwave relays and satellite relays with both up links and down links in the same frequency bands rests on the design of the satellite relay system. The indicated directions of change to relieve the interference problems (increased satellite aperture and EIRP, increased RTC and bandwidth expansion, and decreased earth-station transmitting power) also tend to increase the total system capacity and should therefore be logical long-term goals anyway.

The single important constraint imposed on the microwave relays is that of pointing coordination to avoid the directions of synchronous orbits. The coordination indicated seems about the minimum acceptable and could be evolved gradually with the buildup of all the systems. It does not seem to be an excessive requirement for the great benefits that could be derived with the coordinated sharing of the spectrum. This pointing coordination should not be hard to accommodate at low through mid-latitudes. However, as indicated in Appendix B, at latitudes above 60°, the excluded directions become so large that it may not be feasible to implement a specific requirement for microwave relays with the coordination restrictions. In these regions it may be necessary to choose between microwave <u>or</u> satellite relays for a system operating at the shared frequencies.

V. FREQUENCY-SHARING POTENTIAL OF SYNCHRONOUS ORBITS

The most beneficial utilization of every portion of the available spectrum will ultimately be needed to satisfy the demands for this limited (though nondepleting) resource. Each portion of the spectrum has different characteristics which must be exploited with hardware whose performance and cost is frequency dependent. This section will examine the dependence of low-noise and output devices on frequency and explore the potential spectrum needs for various types of service. Then illustrative sharing possibilities will be presented that match the long-term potential needs to the beneficial use of the characteristics of the various portions of the spectrum.

DEPENDENCE OF LOW-NOISE AND OUTPUT DEVICES ON FREQUENCY*

The equivalent noise temperature of low-noise devices and the efficiency of power output devices degrade gradually with increasing frequency over the 0.1 to 10 GHz range and only somewhat more rapidly in the 10 to 100 GHz range. Power output above 10 GHz tends to fall approximately as $1/f^2$ for most solid-state devices. The important exception, of course, is the transistor, which has a transition region at about 2 to 4 GHz and has little merit above 10 GHz as either a low-noise amplifier or as an output amplifier due to the limitations of current technology.

Current low-noise devices such as microwave transistor, tunnel diode, and uncooled parametric amplifiers are sophisticated and costly, but they can become commonplace and inexpensive via current laboratory technology for fabricating microwave hybrid integrated circuits if an application justifying high-volume production were to develop. Thus, although a single uncooled parametric amplifier at 4 to 10 GHz currently sells for about $4500 to $13,000 including the pump source, integrated units selling for $500 to $700 now appear feasible at the 100 to 500 quantity level, and units selling for $10 to $50 may become feasible at the 10 million unit level.

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*See Appendix D for details.

Despite the currently low cw power output of 0.1 to 10 W for IMPATT, Gunn, and LSA devices around 10 GHz, they already challenge varactor-multiplier chains and have eclipsed the tunnel diode oscillator. Injection locking provides gains of 10 to 30 dB as the locking range or frequency deviation decreases from 100 to 5 MHz. Techniques for stacking diodes, for forming series-parallel arrays on a chip, and for hybrid combining offer increased outputs of from 2 to 64 times that of the single device with little loss in efficiency. Thus solid-state devices may displace many traveling wave tubes (TWTs) within a few years for space applications. However, TWTs can make available kilowatts of microwave power from 2 to 100 GHz by hybrid or space combining. The overall efficiency of TWT amplifiers declines from a typical value of about 25 to 35 percent at 2 GHz to about 12 to 17 percent at 100 GHz. Such a slow decrease is not opt to dominate the choice of frequency.

If there is only a small frequency dependence of the cost of low-noise preamplifiers (as postulated in Appendix D), then eventually neither availability, nor performance, nor cost is apt to determine the choice of frequencies in the range 0.1 to 100 GHz for advanced communication satellite systems.

REQUIREMENTS FOR A VARIETY OF TYPES OF SERVICE

The most demanding applications with respect to outage seem to be for trunk line carrier services. The requirements for these services could most easily be satisfied (from a propagation standpoint) with spectrum below about 15 GHz for satellite links. It should be possible to use common shared frequency bands for both up and down satellite links, even though different antennas or satellites would be required for the up and the down links on the same frequency. However, this would be feasible only for applications where some minimum separation of terminals could be specified and controlled. It would not be practical for many mobile uses or for the density of usage that might be involved in receiving broadcast. These latter applications could best be satisfied with one-way usage, and such usage could be shared by all satellite relay services--mobile, fixed, and broadcast; however, purely terrestrial services involving transmitters <u>and</u> receivers on the same frequencies without minimum separation restrictions should be excluded for efficient use of these bands.

The greatest need for services with unrestricted terminal separations also involves the smallest antenna sizes. Mobile systems have mechanical aperture size limitations for operation, and broadcast receivers for homes would require small apertures in order to be economically feasible for so many receivers. Since the total global system capacity, as limited by satellite spacing, depends on the diameter in wavelengths of the earth terminal antennas, the greatest potential capacity for a given bandwidth, as well as for larger bandwidths, can be obtained by choosing as short a wavelength as is practical. Also most of these applications are less demanding on service continuity than trunk line carrier applications, and can better tolerate the propagation degradations that may be experienced at these higher frequencies.

There is a further requirement for most mobile applications, which is to be able to point the antenna to accommodate rapid vehicle attitude changes. At the higher frequencies a given adequate aperture size (e.g., 1-m diameter) becomes fairly directive, necessitating a rapid pointing or tracking capability, whereas at very low frequencies (e.g., 500 to 1000 MHz) the same aperture size requires very little pointing. This difference between using the two extremes in frequency is further accentuated if a navigation system is included which requires simultaneous linking with a number of satellites in different directions. The technical problem of doing this economically in the near term is much more formidable at the higher than at the lower frequencies. If it proves technically feasible to superimpose down links on current allocations in the range 470 to 960 MHz without restricting or interfering with the applications of existing allocations, this portion of the spectrum will seem extremely attractive for many mobile, broadcast, and austere terminal applications.

By using large bandwidth expansions it should be possible to provide adequate signal strength for the desired message quality without an observable effect on conventional TV reception. For the opposite problem of rejecting strong local TV signals, adaptive arrays with small numbers of elements appear to offer a very promising approach.

ILLUSTRATIVE SHARING POSSIBILITIES

There are many ways in which the spectrum might be partitioned and shared to provide enormous communication capacities. The purpose of this illustration is to show one way in which it might be done with minimal jeopardy to vested practice, providing many times the capacity currently used or projected for the foreseeable needs.

In three portions of the spectrum below 15 GHz (3400 to 4990 MHz, 5725 to 8500 MHz, and 10.7 to 15.35 GHz), there is 7.6 GHz of spectrum currently allocated to fixed and mobile or to communication satellite services. This is the prime portion of the spectrum most suitable for trunk line carrier services. It could be exploited most advantageously by sharing it between fixed and both up and down link communication satellite services. If system designs allowed for tolerable interference as previously indicated, it should be possible to use this spectrum on two independent polarizations, and if high-gain satellite antennas are used, the spectrum capacity could be multiplied again many times by independent beam positions for each satellite. In addition, the number of satellites that could be used independently would be proportional to the equivalent size of the smallest earth antenna using those satellites. If applications involving quite different earth antenna sizes are contemplated, it would be more efficient to separate them by satellite, frequency, and/or polarization so as not to limit the spacings of the satellites used by the larger earth antennas. For example, if the up and the down frequencies were reversed between the applications with earth antennas of different size, the satellites for the two applications could be spaced independently. It is estimated that a margin for propagation attenuation of 10 to 15 dB might be required at 15 GHz to ensure against system performance degradation 0.01 percent of the time in areas of moderately high precipitation.

The remaining portions of the spectrum that might be shared by satellite relays in this illustration would involve one-way usage for all types of satellite relay service--mobile, fixed, and broadcast. The down link for one band might be from 470 to 960 MHz with the corresponding up links in the frequency region between 1430 and 2300 MHz. This arrangement could accommodate, for example, 120 MHz for air mobile, 120 MHz for land mobile, and 250 MHz for direct broadcast or other austere earth antenna applications. Each of these portions could be used independently on two polarizations. The principal reuse of the spectrum in this case might be accomplished by using large antenna gains on the satellites to isolate communications to different areas at the earth terminals. There would be very little dependence on isolation by earth antennas for satellite spacing, i.e., the number of satellites would be determined by the total power and coverage needed and their orbital locations would not be critical.

In addition, between 17 and 50 GHz (17.7 to 23 GHz, 25.25 to 31.3 GHz, and 36 to 50 GHz) there is a total of more than 25 GHz of spectrum currently allocated, primarily to fixed and mobile services, or unallocated that might be partitioned into exclusive satellite up and down bands for all kinds of satellite service, including fixed, air and land mobile, and direct broadcast. Space diversity for fixed services may provide adequate system performance to meet trunk line carrier objectives for satellite relay in bands between 17 and 50 GHz with margins of 10 to 30 dB for propagation attenuation. For air and land mobile and direct broadcast, versatile direction diversity with satellite redundancy may also provide adequate system performance for these systems in the same bands with margins of 10 to 30 dB for propagation attenuation.

And finally, above 80 GHz (80 to 100 GHz, 130 to 160 GHz, and 200 to 300 GHz), there is 150 GHz of spectrum that is currently unallocated and that might be partitioned into exclusive synchronous satellite up and down bands for all kinds of satellite service. It would seem most appropriate, however, to emphasize the use of these latter bands with aircraft or satellites above much of the atmosphere.

All of the one-way bands above 17 GHz could be used independently on two polarizations, could provide adequate isolation to permit close spacing of satellites with very modest physical size earth terminal antennas, and would permit further spectrum multiplication with high-gain (though physically small) antennas on the satellites. Thus, there is an enormous potential capacity available at the higher frequencies for applications that can accept or avoid the effects of large atmospheric attenuations that may occur.

Appendix A

REFERENCE ANTENNAS AND THE EFFECTS OF SATELLITE SPACING

The earth-station antenna is particularly important in determining the amount of interference that is received from unwanted satellites, and hence in determining the satellite spacing needed to limit the interference to any specified value. There are many possible types of antennas and many illuminations for most types. In order to illustrate the range of performances potentially available, typical patterns for a variety of conventional antennas are defined as reference antenna patterns. Then the performances of these antennas are illustrated in terms of interference-to-signal ratios as a function of satellite spacing. Finally, the potential performance of adaptive arrays is illustrated under various conditions to demonstrate some of the attractive features of this unconventional type of antenna.

CONVENTIONAL ANTENNA PATTERNS

The reference antenna patterns chosen are illustrated in Figs. A-1 through A-4, where the power relative to the maximum on the main-beam axis is plotted as a function of the angle away from the axis. The mathematical form for the power pattern is usually a function of $u = (\pi D/\lambda) \sin \theta$, where D is the antenna diameter, λ is the wavelength, and θ is the angle from the axis. When it is of this form, a universal power pattern can be constructed as a function of u for any D/λ. However, the CCIR reference is not of this form; nor are the errors or boundary corrections for any of the patterns, so they have been illustrated for a range of values of D/λ. In order to use the illustrated patterns for arbitrary desired D/λ, the pattern with the nearest D/λ may be selected and the θ values multiplied by the ratio of the illustrated to the desired D/λ. This will give adequate accuracy for most applications except where "errors" or limitations below isotropic are involved. The CCIR patterns will of course always be subject to D/λ interpolation errors, which increase with the difference between the desired and illustrated D/λ.

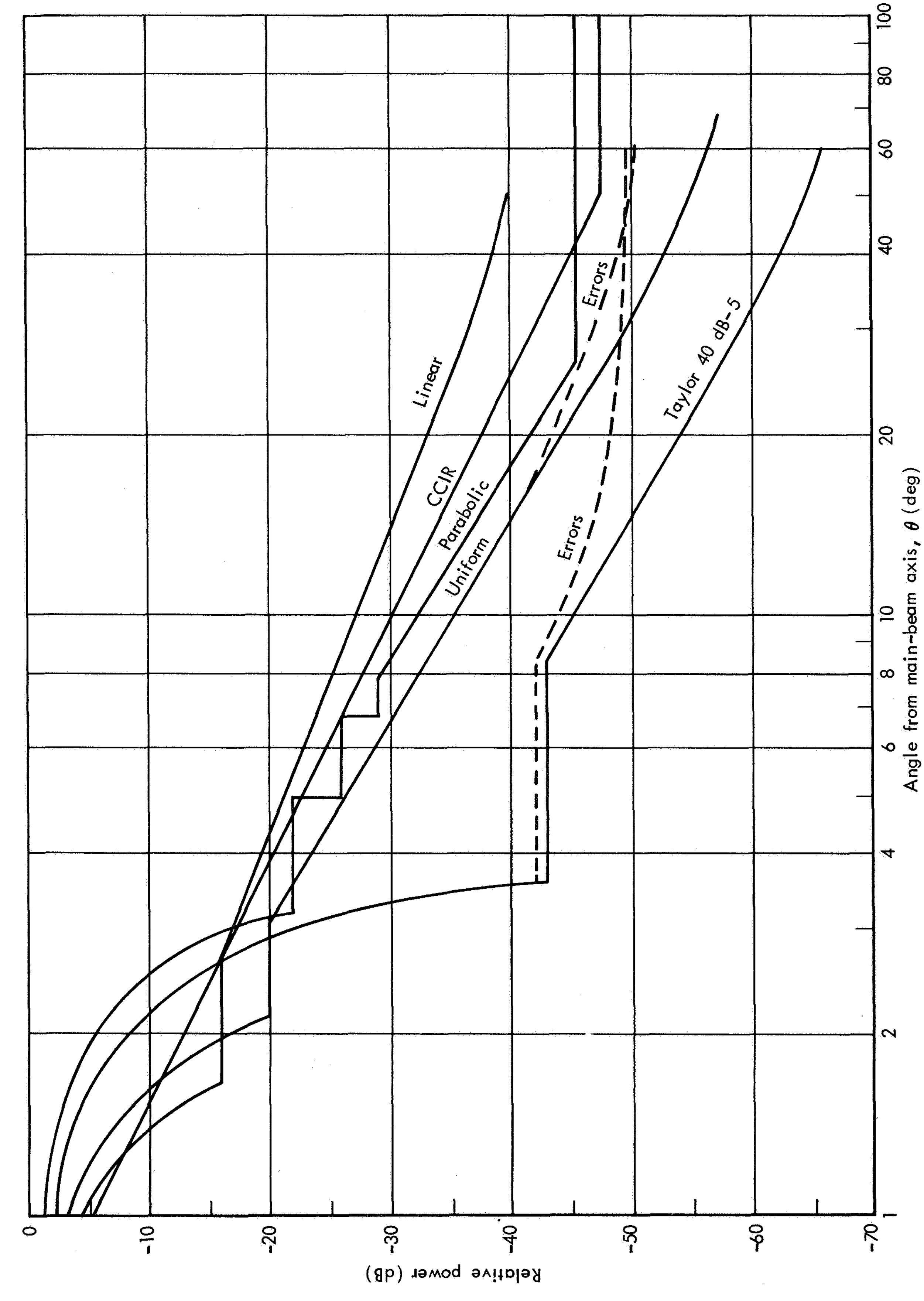


Fig. A-1—Antenna model patterns for $D/\lambda = 30$

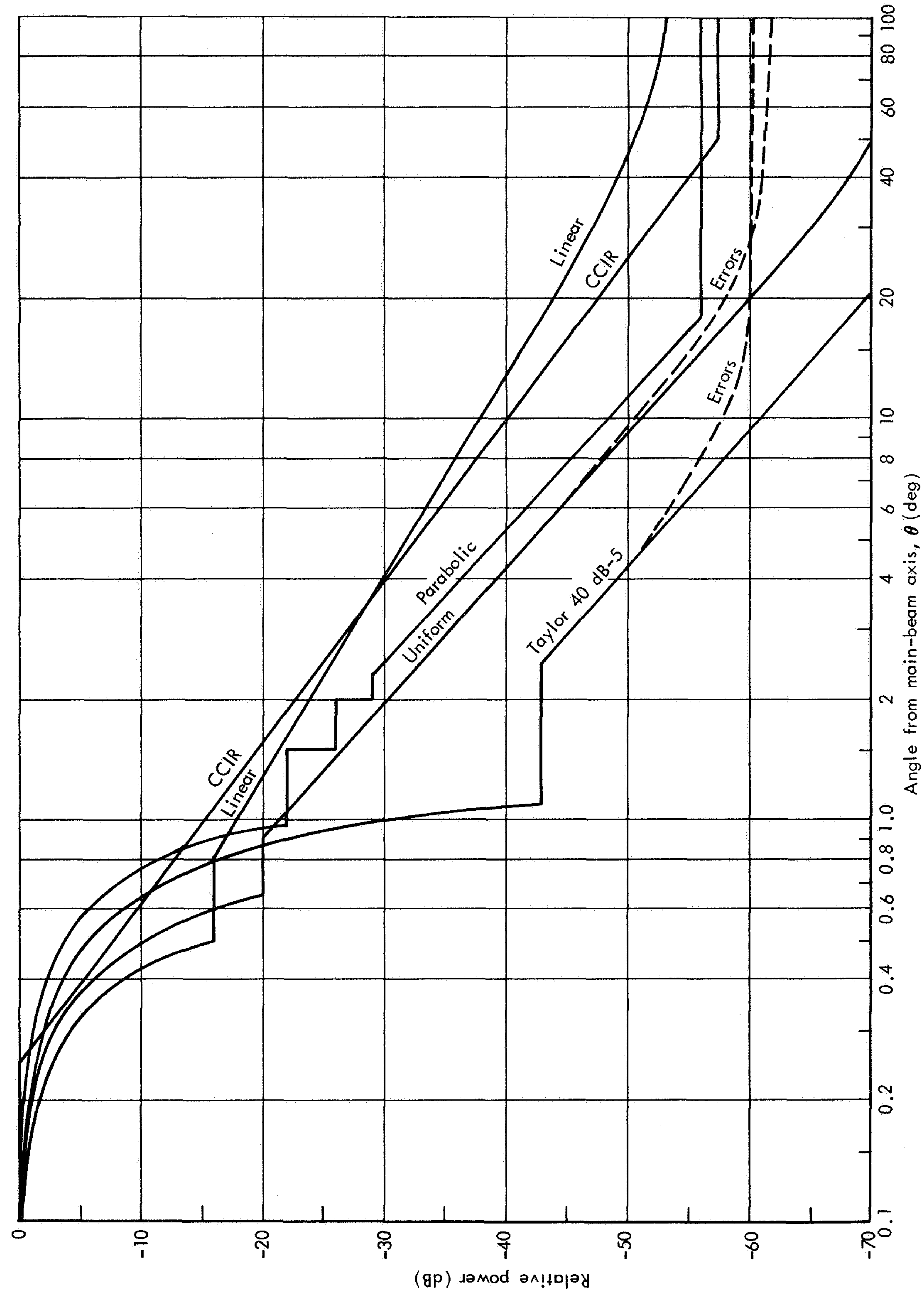


Fig. A-2—Antenna model patterns for $D/\lambda = 100$

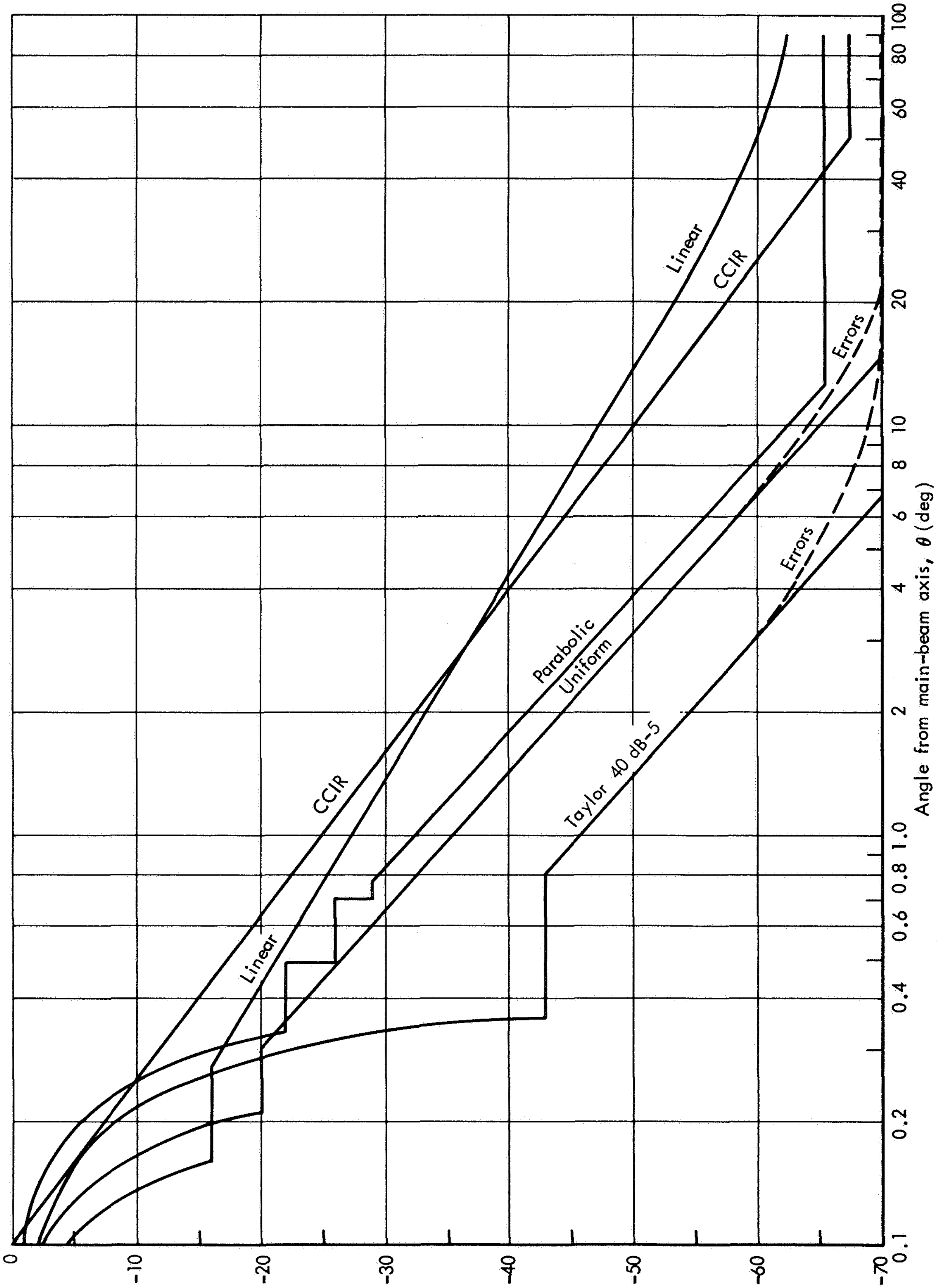


Fig. A-3—Antenna model patterns for $D/\lambda = 300$

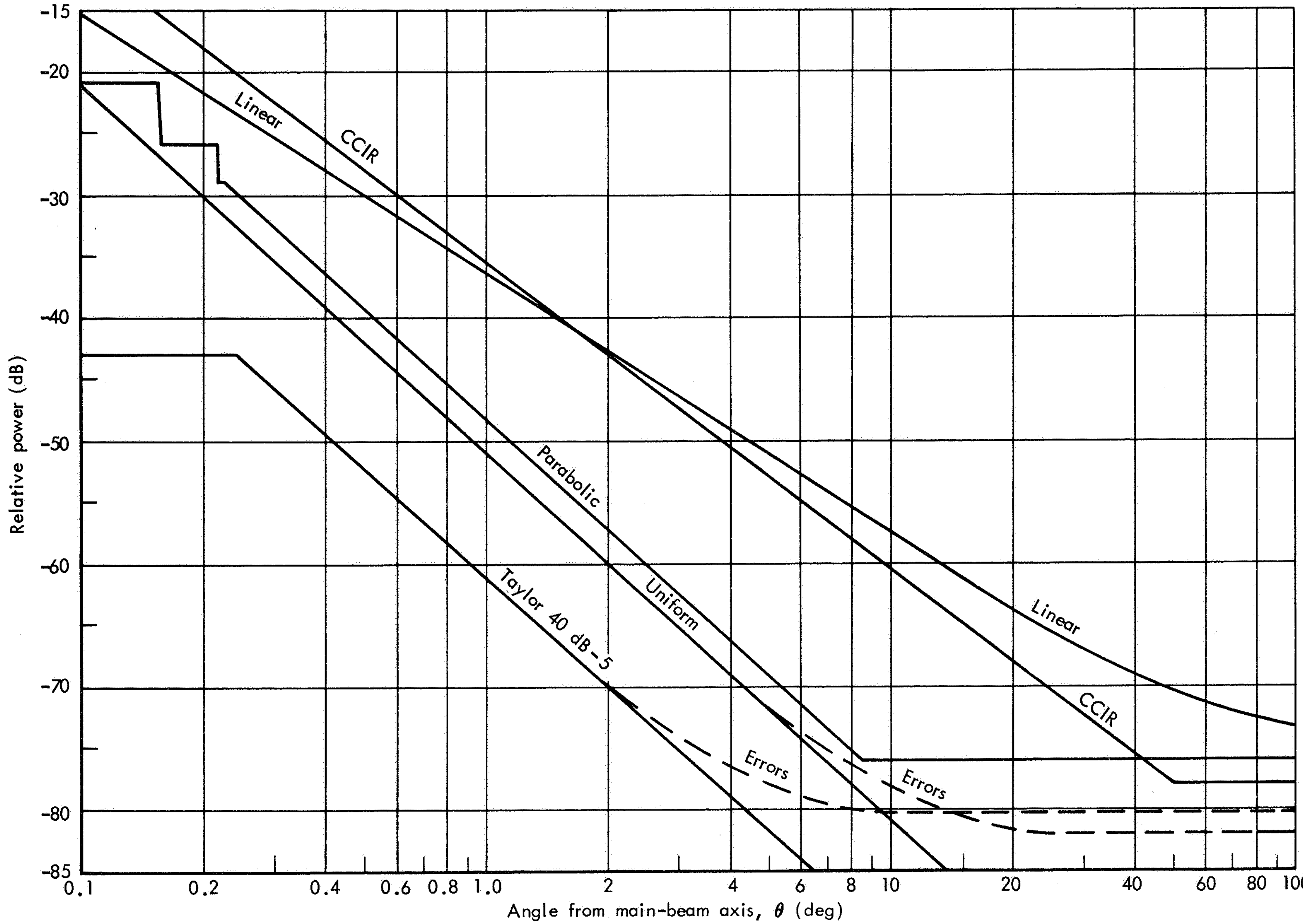


Fig.A-4—Antenna model patterns for $D/\lambda = 1000$

The "parabolic" reference pattern illustrates what could be considered as typical of a circular parabolic reflector antenna, with an illumination falling to 0.25 power at the perimeter and a 10-percent blockage factor. The curve represents the average value of the sidelobes by dropping in steps through the first three sidelobes and then as $1/u^3$ through the remainder of the sidelobes through points 3 dB below the peaks of all the sidelobes. This reference antenna, with an efficiency of 60 to 70 percent, has a pattern limited at wide angles to about 8 dB below isotropic.

The "uniform" reference pattern is that of a uniformly illuminated circular aperture. It steps through the first sidelobe and then falls as $1/u^3$ through the remainder of the sidelobes through points 3 dB below the peaks of all the sidelobes; it thus represents the average value of the sidelobes. The "error" curve shows the effects of an average 20 percent rms excitation error on each element of a large array with uniform illumination.

The "Taylor 40 dB-5" reference pattern[1,2] represents the practical limit of current antenna technology for low sidelobes. This model is designed to limit the first five sidelobe peaks to 40 dB below the main beam. The reference pattern represents the average value of the sidelobes. It is flat at 43 dB below the main beam through the first five sidelobes and then falls as $1/u^3$ through the average value of the remaining sidelobes. The "error" curve again represents the effects of an average 20-percent rms excitation error on each element of a large array designed for this pattern.

The "linear" reference model represents a uniformly illuminated line source. It steps through the first sidelobe and then falls as $1/u^2$ through the average value of all the sidelobes.

The final reference model is the CCIR standard for interference calculations. This standard is given as $G = 32 - 25 \log \theta$ (dB), with respect to isotropic, with a lower limit of 10 dB below isotropic. Since all our reference patterns are given relative to the maximum along the axis of the main beam, the CCIR antenna pattern is somewhat inconsistent and does not scale very well with the other patterns. It is probably very conservative over most of its range and represents an empirical envelope of the sidelobes for some low-noise antennas.

SATELLITE INTERFERENCE USING CONVENTIONAL EARTH-STATION ANTENNAS

The interference-to-signal ratios X/C with the reference antenna patterns were computed using the reference earth-satellite geometry model. This assumes an earth station at 45^o latitude with its antenna axis pointed at the wanted satellite on the same meridian. The interference is from a full ring of synchronous equatorial satellites at uniform longitudinal spacing. All satellites are assumed to have the same EIRP in any given bandwidth in the direction of the earth station, but only those satellites above the local horizon are assumed to contribute interference. The results of these computations for the reference antenna patterns of Figs. A-1 through A-4 are given in Figs. A-5 through A-8. Further interpretation of the interference computations for conventional antennas is made in Section IV.

INTERFERENCE DISCRIMINATION WITH ADAPTIVE ARRAYS

It has been shown that feedback circuitry like that in Fig. A-9 can adjust the complex weights of the elements of an array to give very nearly the best possible signal-to-interference ratio for any configuration of elements and interfering sources.[3,4] Because the adjusting signal is derived from the output and fed back to the input, an adaptive array not only seeks the optimum illumination function but also corrects what otherwise might be its own element-spacing errors, making near optimum performance possible without precise antenna design or construction. One may wish to know then:

- o How well can an adaptive array perform?
- o What are its limitations?
- o How practical is it to implement?

To answer the first question, a computer program was written to find the optimum performance for various arrays with interfering satellites arranged as indicated in the reference geometry. The results are illustrated in Fig. A-10. Internal receiver noise 70 dB below the wanted signal was assumed to limit the attainable C/X at about the limit of computer accuracy. Solid lines show the effect of a diffusely reflecting earth on a square array adaptive in the plane of the satellites but with a fixed

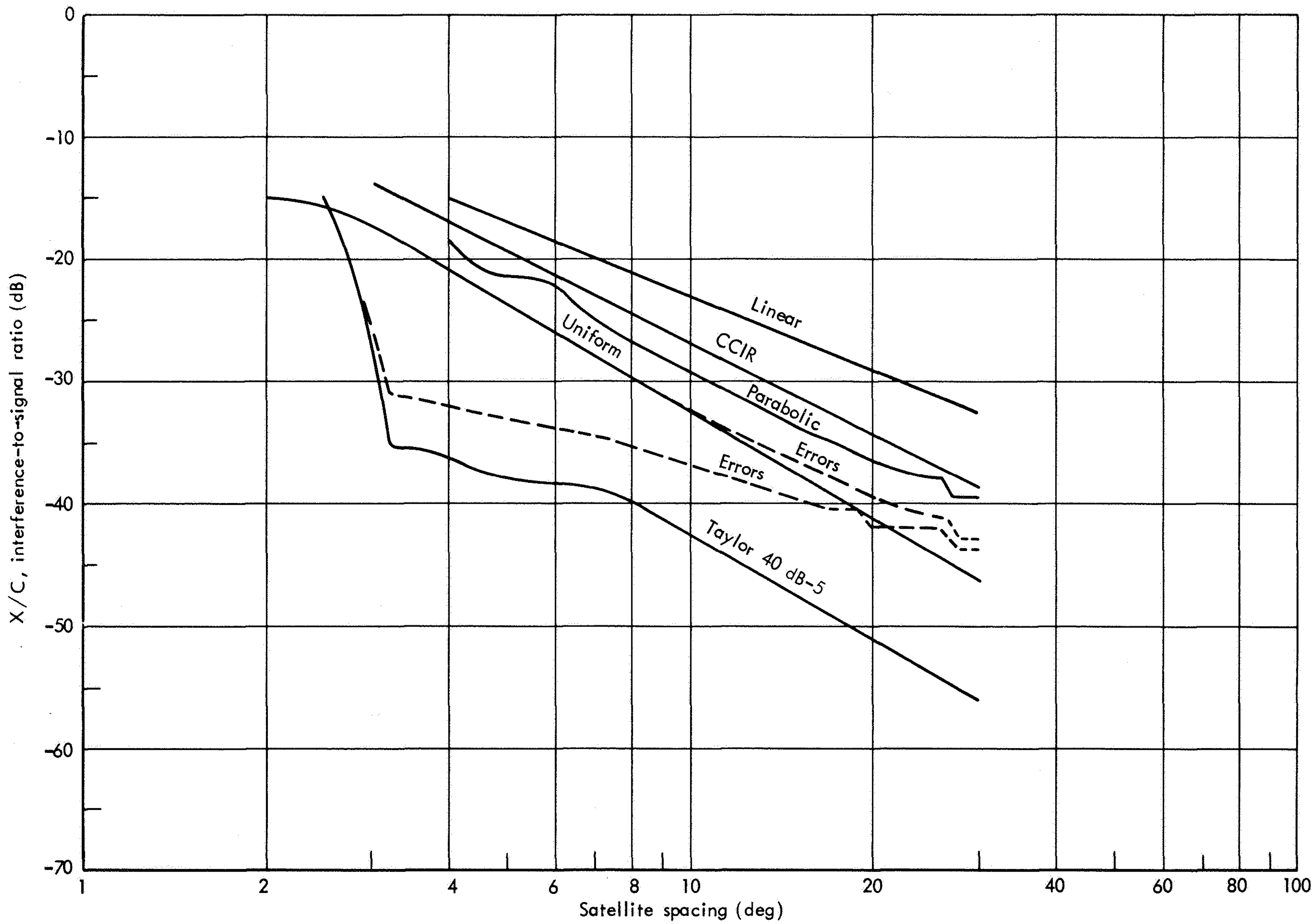


Fig.A-5—X/C versus satellite spacing for antennas with $D/\lambda = 30$

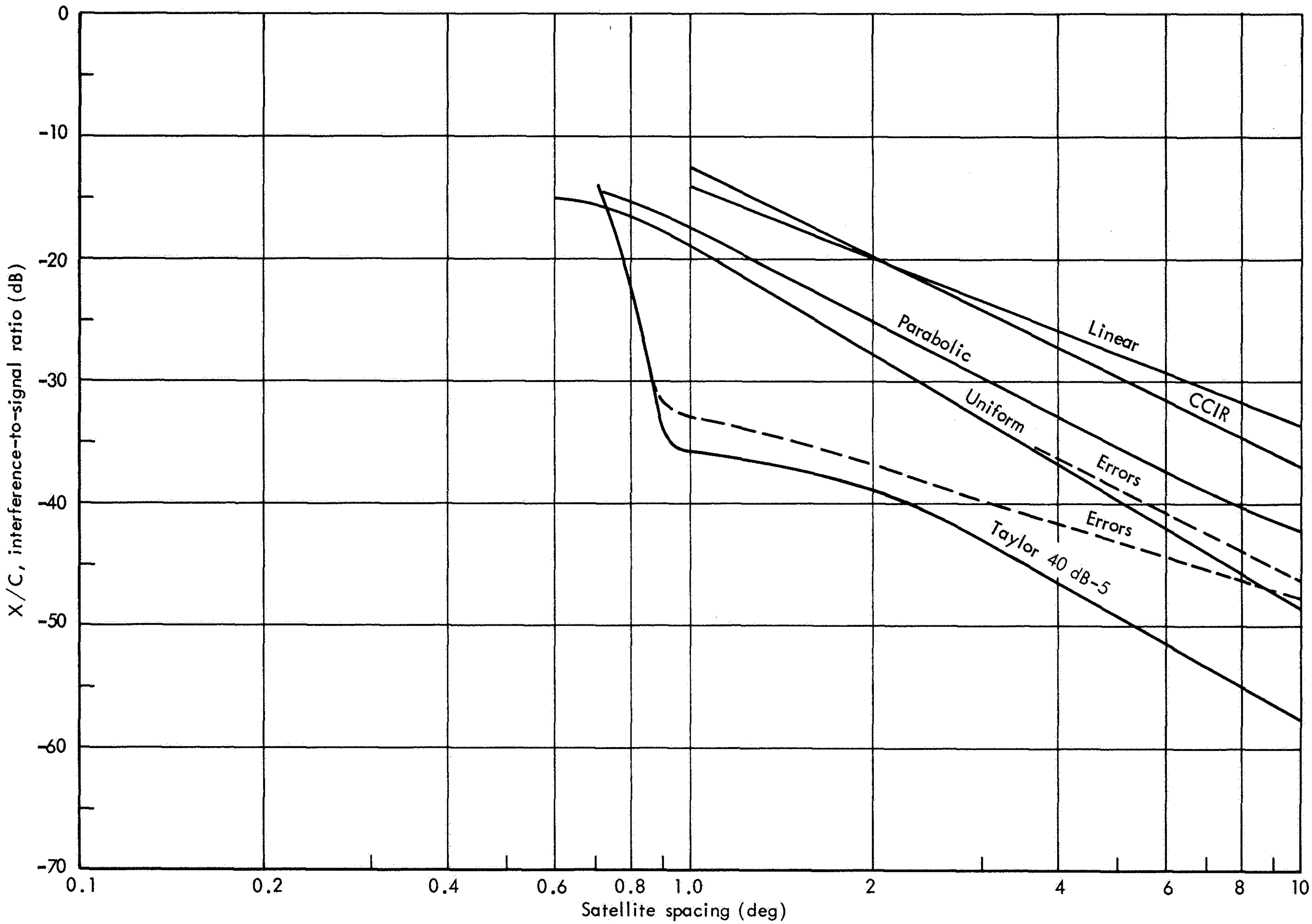


Fig.A-6—X/C versus satellite spacing for antennas with $D/\lambda = 100$

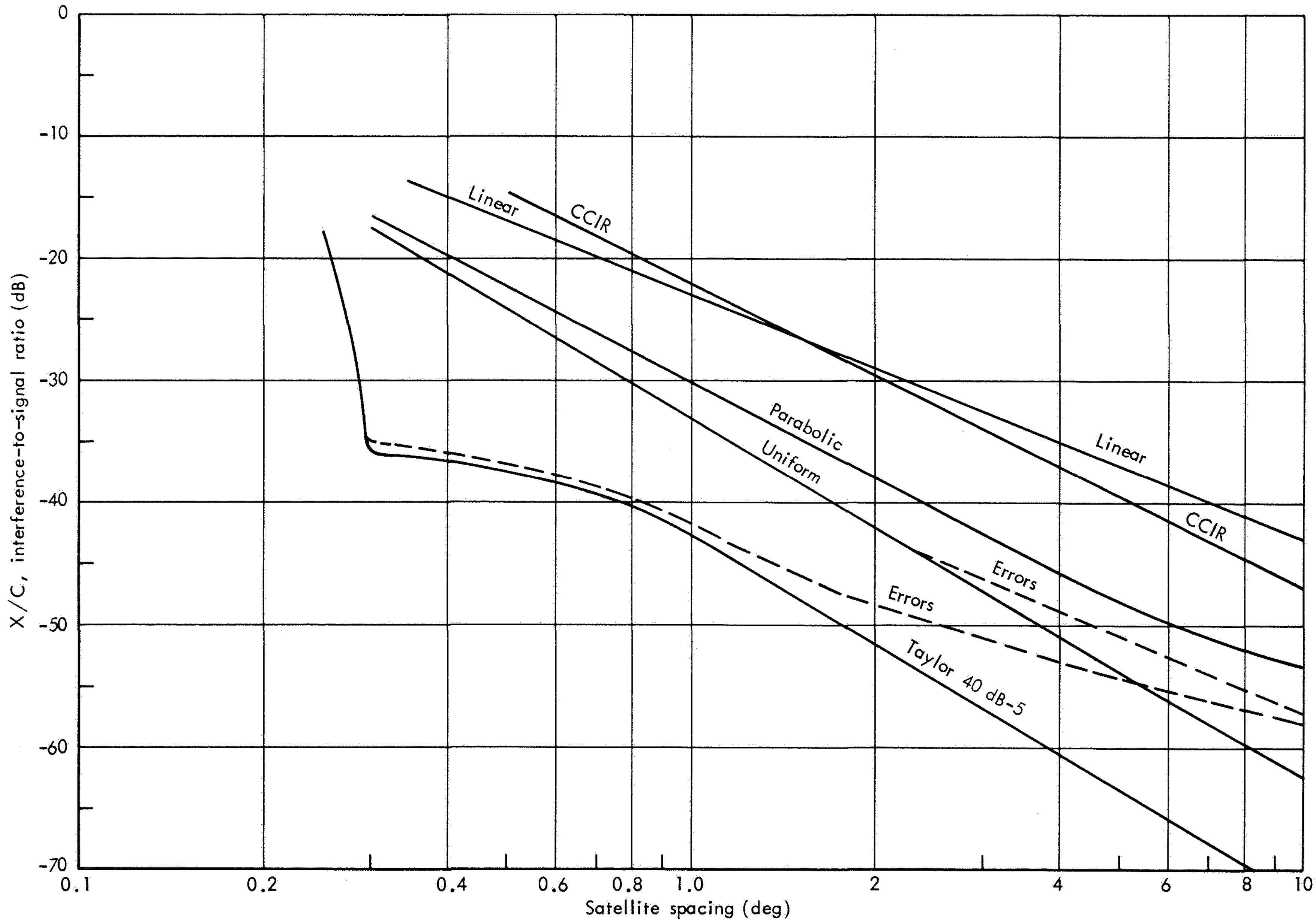


Fig. A-7—X/C versus satellite spacing for antennas with $D/\lambda = 300$

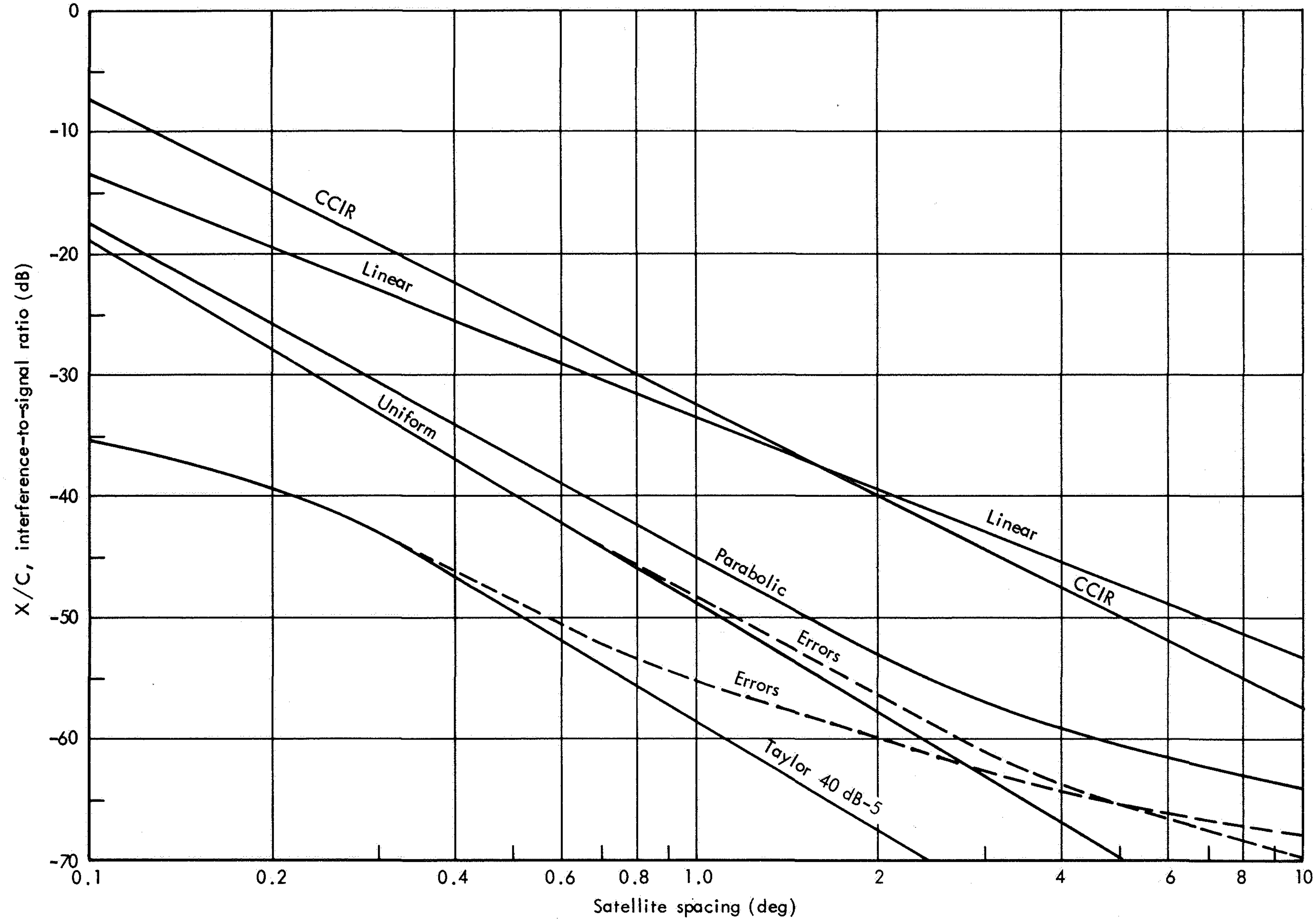
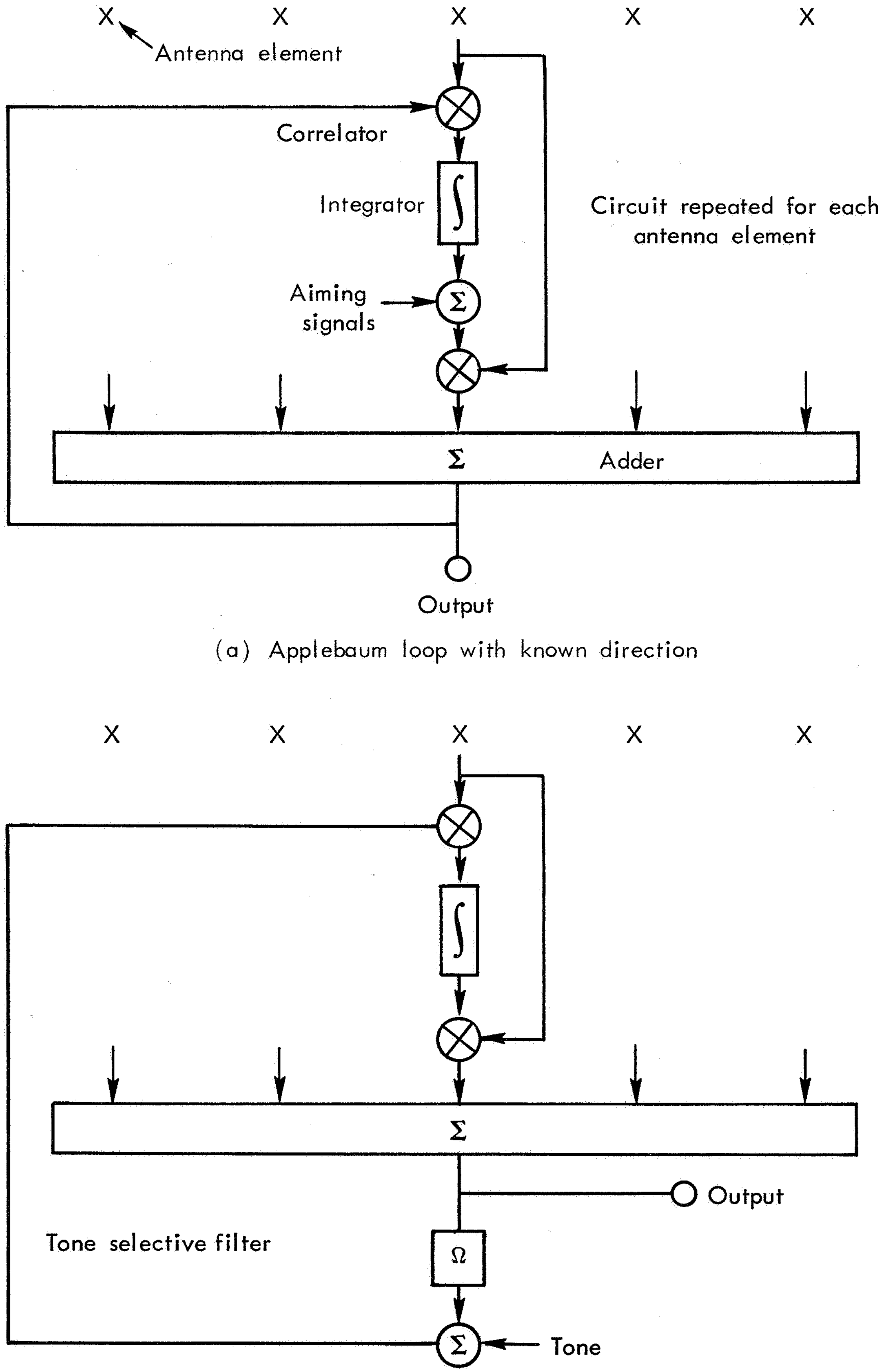


Fig.A-8— X/C versus satellite spacing for antennas with $D/\lambda = 1000$



(a) Applebaum loop with known direction

(b) Modified Widrow loop using tone code

Fig.A-9—Several schematic adaptive-array loops

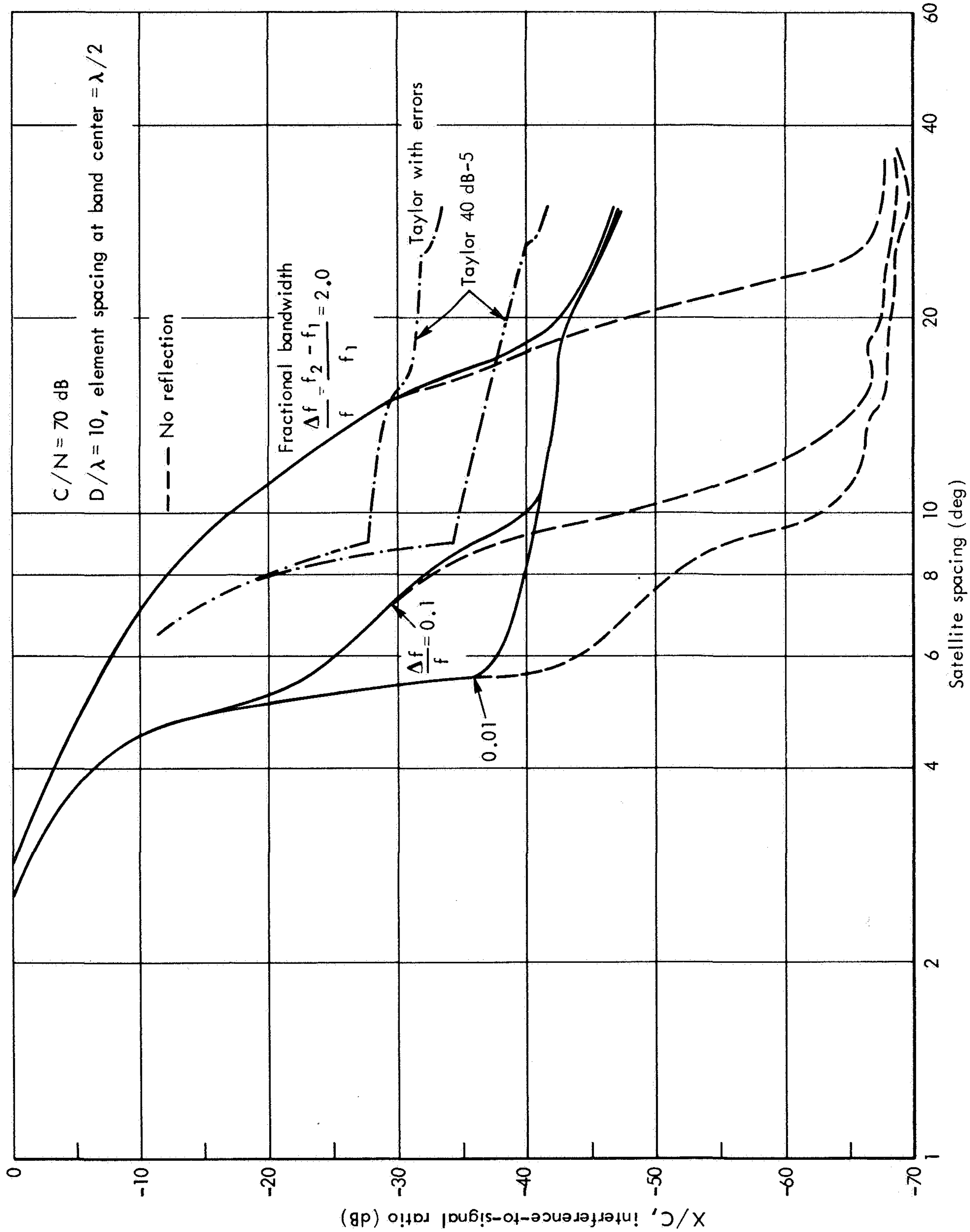


Fig. A-10 — X/C versus satellite spacing for adaptive arrays uniform illumination in the other dimension. The dashed lines are for a linear array without reflection. For comparison, the performance of the Taylor 40 dB-5 with a $D/\lambda = 10$ is also shown. The narrower bandwidth adaptive-array cases are considerably better, as would be expected.

Many other cases were examined, and it was found that the no-reflection antenna patterns did not vary more than a few dB with wide changes in earth-station latitude, and that the curves for values of D/λ from 5 to 100 were very close to the no-reflection curves in Fig. A-10 when the spacing scale was normalized by multiplying by 10 λ/D. Thus these curves indicate what could be achieved over a wide range of D/λ values.

The results in Fig. A-10 assumed that the desired satellite was in a direction perpendicular to the array. Electronically steering to off-axis satellites would cause a degradation due to effective antenna shortening. Estimates of the degradation can be made by assuming that the effective D/λ is $(D/\lambda)\cos\theta$, where θ is the off-axis angle of the desired satellite.

A study of the behavior of these adaptive arrays indicates that when the number of satellites visible to the antenna is less than the number of elements, antenna weights are fairly uniform, giving a narrow main beam with the antenna pattern adjusted to place the interfering satellites in pattern nulls. For wide bandwidth cases, where the interfering satellites are effectively spread out in angle, the interference cannot be put into nulls, and a more Taylor-like pattern results in which near-in sidelobe levels are reduced at the expense of some main-beam broadening to obtain the best compromise. Optimum element weights obtained from this type of program can be used as a design goal for fixed-array antennas.

Antenna errors and possibly reflections limit average sidelobe levels of conventional antennas to about 10 dB below isotropic. Adaptive arrays can achieve much greater rejection ratios against discrete interference sources than is implied by this empirical rule, as indicated in Fig. A-10. However, the presence of reflections from surrounding objects or a rough earth does impose a need for more gain, i.e., more elements. Objects causing reflections that are not in the main beam are treated as additional interfering satellites and are discriminated against without much loss in C/X. Diffuse or Lambert scattering, however, produces a distributed return that is difficult to suppress with small numbers of elements. In the case of a line array, some of this energy falls in the pancake-like main beam, causing severe degradation. The reflection calculations of Fig. A-10 were made for a square array of $D/\lambda = 10$, adaptive in one dimension and having a flat nonadaptive illumination pattern in the other. The break-away point is primarily dependent on the length and illumination function of the array in the direction perpendicular to that of the adaption.

The time allowed for the network to adapt will determine how closely the optimum is approached. Widrow[4] shows that for a particular adaptive algorithm the least-square error between actual and optimal signal output is proportional to the number of antenna adjustments and inversely proportional to the adaptive time constant. A general computer simulation model constructed by W. Doyle was used to compare the adaptive properties of the circuit of Fig. A-9a with that of Fig. A-9b of Widrow. The results were quite similar, both indicating that when adaption takes place for a ten-element (20-adjustment) array with the desired signal absent, the process approaches 1 or 2 dB of the optimum in a time required for about 75 independent samples. If the time for an independent sample is assumed to be the reciprocal of the bandwidth, a 1-MHz bandwidth would require about 75 μsec to adapt this well.

The analysis and simulations of Doyle also point out the very major difference between adapting when the desired signal is present and when it is absent. If one adapts as in Widrow using a signal known at the receiver, or as in Applebaum to a given angular direction without a signal in that direction, then the adaptive times given above and in Widrow apply. If, however, in either of these circuits a desired signal is present that is unknown at the receiver (the usual case in practice), then the circuits approach ideal adaption at a very much slower rate. The desired signal enters the adaptive loop as an error source with zero mean, requiring very strong smoothing, i.e., a long time constant to reduce its value. If, as it now seems, this is a fundamental limitation, then for the fairly fast adaptive times that might be required for mobile adaptive arrays or for rapidly moving interference sources, coding may be required on the transmitted signal to allow faster adaption.

There are several potential applications for which adaptive arrays appear especially attractive. For microwave relays where fading is the limiting factor in system performance, it should be useful to adaptively couple two antennas in height diversity so as to null out one multipath component and limit the depth of fade, thus improving system performance by a corresponding amount.

For earth terminals requiring links to many different satellites simultaneously, an adaptive array with a given total aperture could provide n such links with better interference rejection than n conventional reflectors, each with the same total aperture. When n exceeds some small number, this could probably be achieved at less cost than with n conventional antennas.

For small nearly omnidirectional antennas where size or cost limits the number of elements that can be used (e.g., mobile or home receivers), and where strong rejection of a small number of interfering signals is needed (e.g., reflected multipath component or an independent signal at a quite different polarization or direction), adaptive coupling of these antennas may prove to be a desirable means of achieving the needed X/C.

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3. Applebaum, S. P., Adaptive Arrays, Special Projects Laboratory, Syracuse University Research Corporation, SPL 709, June 1966.
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Appendix B

EARTH-TO-SATELLITE GEOMETRY FACTORS

THE EFFECTS OF EARTH-STATION LOCATION

For any given distribution of synchronous satellites, the differences in the ratio of wanted to unwanted signal C/X at various earth-station locations are due primarily to the variations in the angular separations between a desired satellite and its nearest interfering neighbors. The angular separations are determined largely by the distance from the earth station to the desired satellite, a distance fully prescribed by specifying the satellite elevation above the horizon. Variations of C/X are thus determined primarily by apparent elevation of the satellite above the horizon regardless of the earth-station location.

The location of an equatorial satellite relative to an earth station is defined by the earth-station latitude and the satellite longitude relative to the earth station. With these coordinates, the satellite elevation above the horizon can be obtained from Fig. B-1. The angular separation factor f of Fig. B-2 relates the observed angular spacing to the longitudinal spacing and is principally dependent on the elevation angle. There is only a very weak dependence upon relative longitude as shown. For an antenna with a power gain pattern of characteristic n, $G \propto u^{-n}$, where $u = (\pi D/\lambda \sin \theta)$ and θ is the angle from the axis of the main beam, C/X will be n·f that obtained with an antenna at the center of the earth. For example, from an earth station at 45^{o} latitude viewing a satellite on the same meridian, the elevation angle is 38.2^{o} and f = 0.46. If C/X is expressed in dB, the general relationship for C/X, relative to that appearing in the reference geometry (latitude $\varphi = 45^{o}$, relative longitude L = 0), is:

$$(C/X)_{\varphi,L} = (C/X)_{\varphi=45^{o},L=0} + n\cdot [f - 0.46]$$

The foregoing method of estimating the effects of various alternative geometries upon C/X has been checked against more detailed calculations

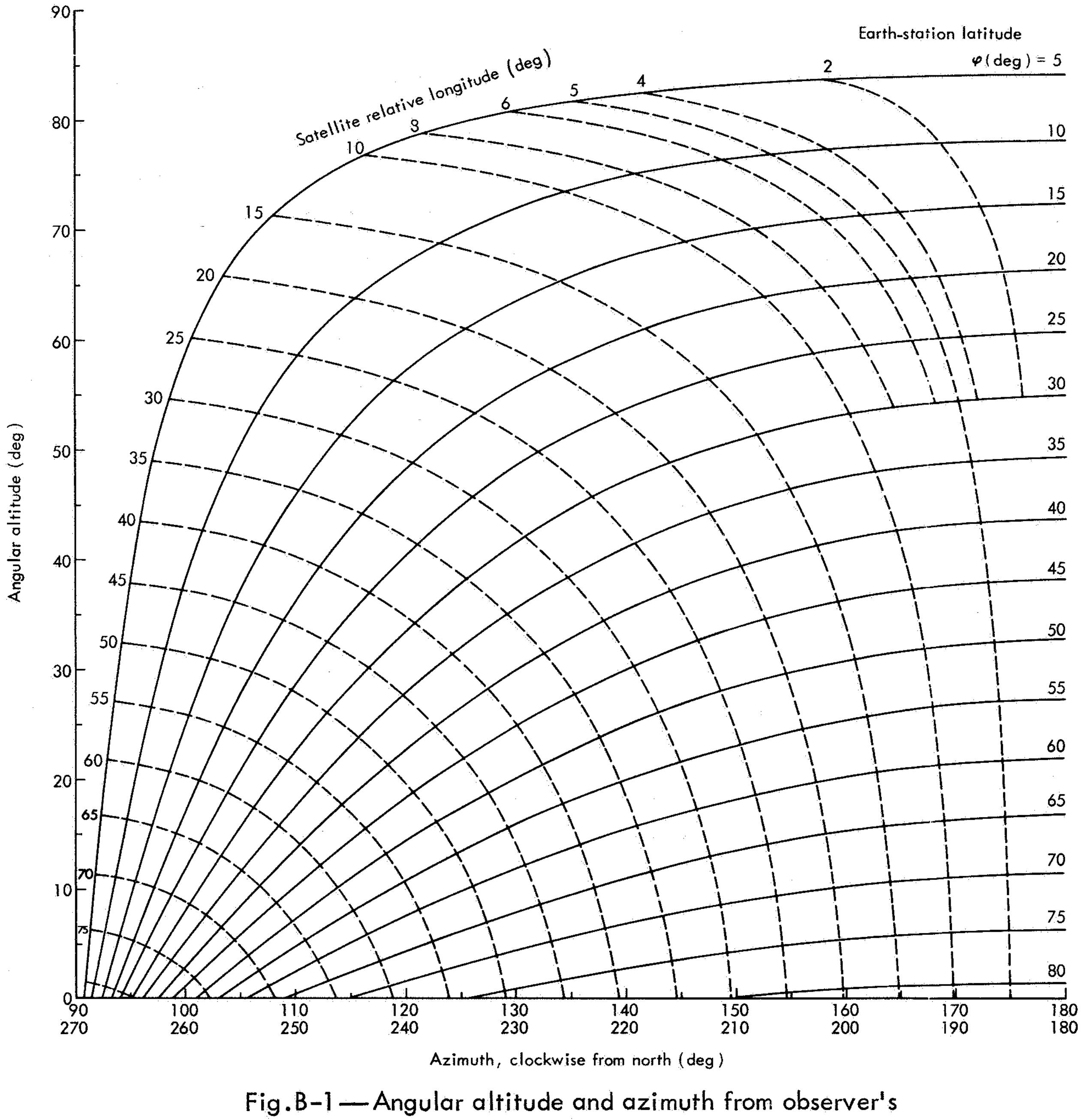


Fig. B-1 — Angular altitude and azimuth from observer's latitude and relative longitude

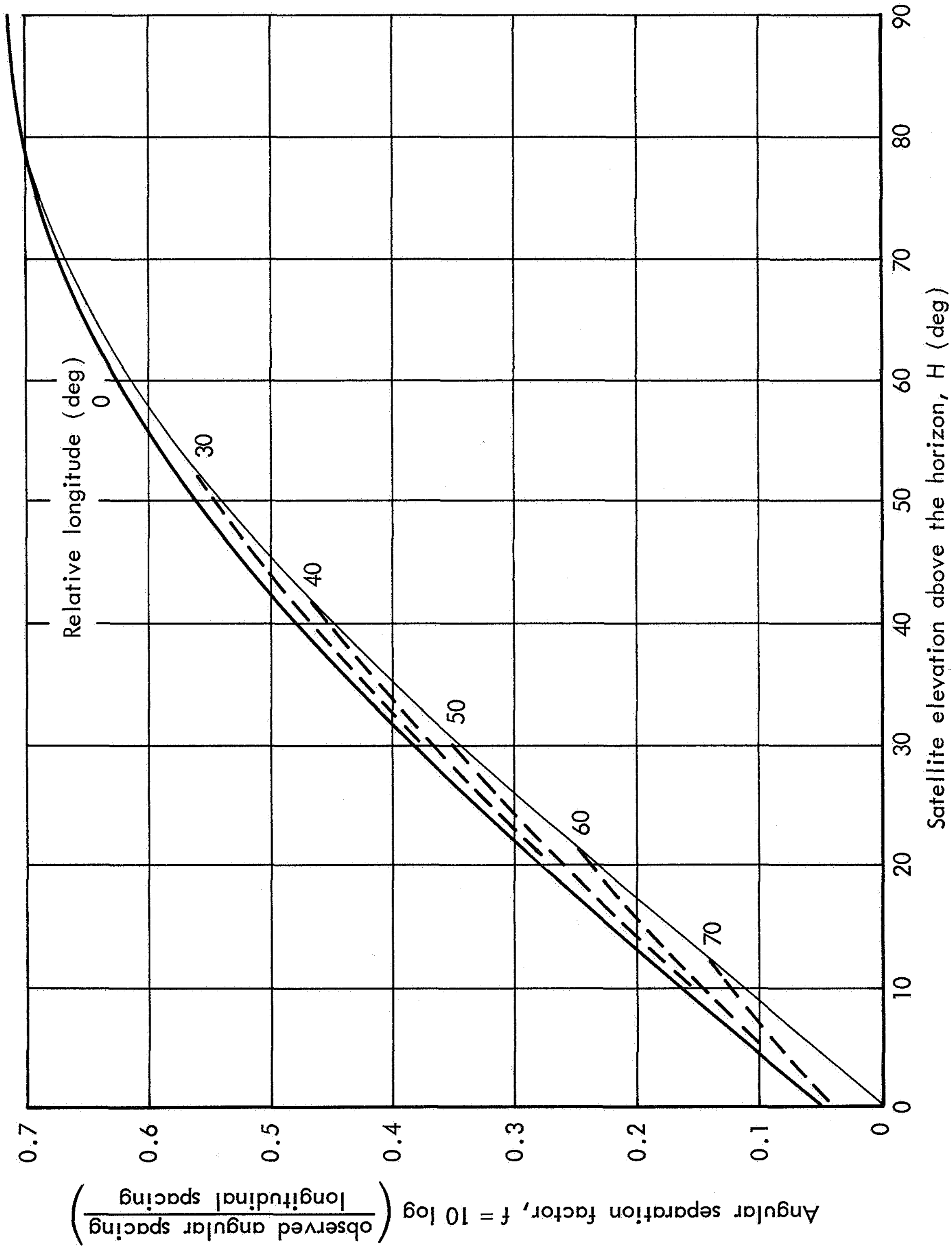


Fig. B-2—Angular separation factor versus elevation and relative longitude where all the satellites are considered and the distribution of satellites is cut off by the horizon. Three antenna types are considered. The first two cases are based upon $G \propto u^{-n}$, with n = 2.5 and 3 respectively. The third case is termed the Taylor 40 dB-5 design. The pattern $G \propto u^{-3}$ corresponds to the theoretical envelope for a uniformly illuminated aperture, whereas some antennas (including the CCIR model) are better described with n = 2.5. In the Taylor 40 dB-5 design, the peaks of the first five sidelobes are uniformly down 40 dB from the mainlobe with the more remote sidelobes decaying as u^{-3}. In the example treated below, the transition from the flat region to the u^{-3} region is taken at $\theta = 20^{o}$, and it is further assumed that the nearest interfering satellites appear in the flat region, never in the mainlobe.

Satellites are assumed to be uniformly distributed in longitude, and the wanted power C_i received from the i^{th} satellite is

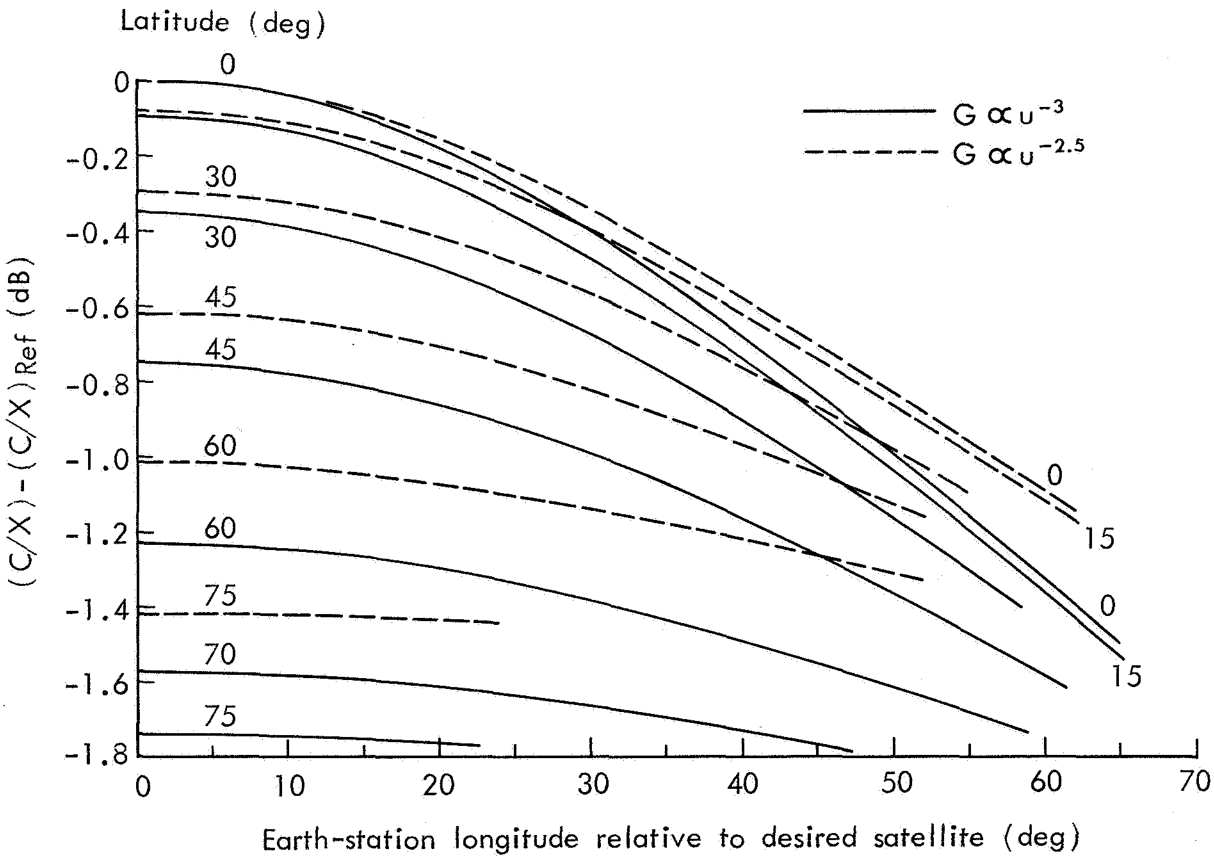
$$C_i \propto (G_{i,i})/d_i^2$$

and the interference from other visible satellites (summing over the visible satellites) is

$$X_i \propto \sum_{j \neq i} (G_{i,j})/d_j^2$$

where $G_{i,j}$ is the antenna power response to the j^{th} interfering satellite when the antenna is pointed in the direction of the i^{th} satellite, and d_j is the distance to the j^{th} satellite. The ratio C_i/X_i, calculated for a particular earth-station latitude, satellite relative longitude, and antenna characteristic, is divided by the same ratio at the subsatellite point on the earth's surface. This normalization differs from that used in constructing Fig. B-2 but the variations are the relevant information; no absolute levels are implied in either construction.

The results for the three cases are shown in Fig. B-3. The zenith situation provides the largest angular separation between the desired satellite and the interfering neighbors and the largest values of C/X. A secondary effect arises near the horizon where the dominant nearest neighbor interfering satellites may be obscured by the horizon. The curves of



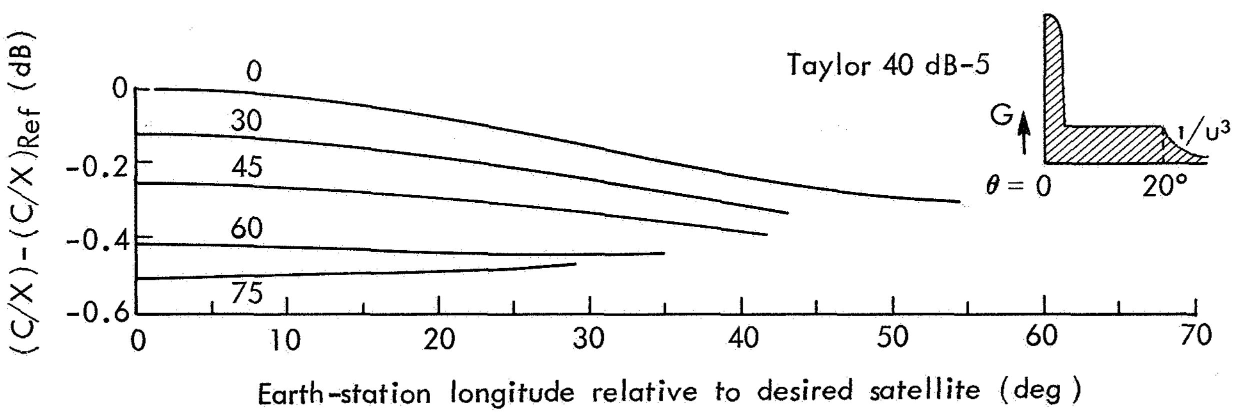


Fig. B-3—Variation of C/X with earth-station location

Fig. B-3 were calculated assuming a satellite spacing of 5^{o} and they have not been continued into the region where horizon cutoff becomes a dominant feature.

Figures B-1 and B-2 can be used with a power-law antenna characteristic to quickly derive C/X variations with earth-station location similar to that shown in Fig. B-3. Caution is advocated, however, in applying this technique in regions where horizon cutoff may become important.

ANTENNA ALIGNMENT AND MUTUAL INTERFERENCE BETWEEN TERRESTRIAL STATIONS AND SATELLITES

If synchronous near-equatorial satellites are densely distributed in longitude, they will appear to an observer in the northern hemisphere to occupy a rainbow-like strip in the southern sky. The width of the strip will depend upon the satellite orbital inclination, and if the strip is regarded as structureless its appearance will be independent of the observer's longitude. Figure B-4 illustrates the appearance of a dense group of satellites having 5^{o} orbital inclination when viewed from 45^{o} N latitude. The horizontal line at H = 0 represents the horizon; the group appearance for more northerly observers may be qualitatively approximated by shifting the horizon line upward approximately 1^{o} for each incremental degree of observer's latitude.

The boundary arcs of Fig. B-4 represent the limits in altitude and azimuth corresponding to the extreme declinations for an orbital inclination of 5^{o}; the connecting line segments illustrate the sequence of altitude-azimuth coordinates occupied by a satellite in its daily declination cycle. The spacing between successive segments corresponds to 5^{o} intervals in the longitudinal spacing of those satellites used to illustrate the declination cycle.

If the relative alignment of terrestrial microwave relay and satellite relay antennas is controlled so that either or both do not point at the other, the mutual interference potential could be significantly reduced. If the microwave relay sites are chosen so that the axis of none of their antennas points within some interference angle (B, in degrees) of a potential synchronous orbit location, the interference potential could be greatly reduced. If, in addition, the satellite antennas are

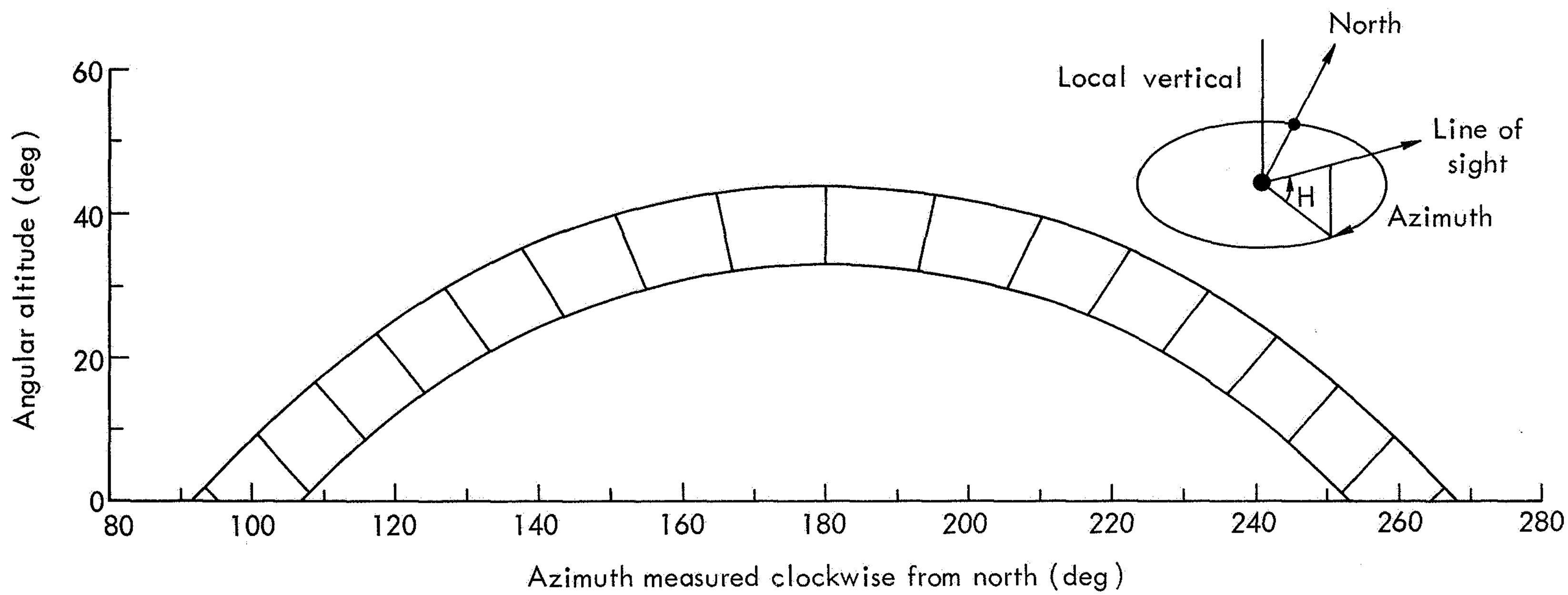


Fig. B-4—A dense distribution of synchronous satellites viewed from 45° N latitude sufficiently directive and are pointed so as to significantly reduce their gain toward the earth's horizon, the effective interference horizon for the microwave relays could be raised by M degrees above the physical horizon. Figure B-5 shows a set of azimuth circles centered on microwave relay stations at various latitudes. Three types of azimuthal regions appear: (a) an unshaded region free of mutual interference, (b) a stippled region of mutual interference, and (c) a striped region which is free from interference because satellites are beyond the effective horizon. Figures B-6 through B-9 define boundary lines between such regions, which are identified (a), (b), and (c) for convenience. The computations assume that the microwave relay antennas are pointed perpendicular to the vertical, and the nature of the system requires mirror symmetry about both the north-south and east-west axes. Examination of the figures indicates that the techniques for controlling the antenna alignments may be useful for low-to-mid-latitudes where relatively small bands of east-west microwave relay azimuths need to be excluded. However, above latitudes of about 60° these techniques may require such large north-south azimuth exclusion that sharing the same frequencies between microwave relays and satellite relays at these latitudes may not be feasible.

The comparison of cases which differ only in orbital inclination shows that a reduction in usable azimuths attends an increase in orbital inclination.

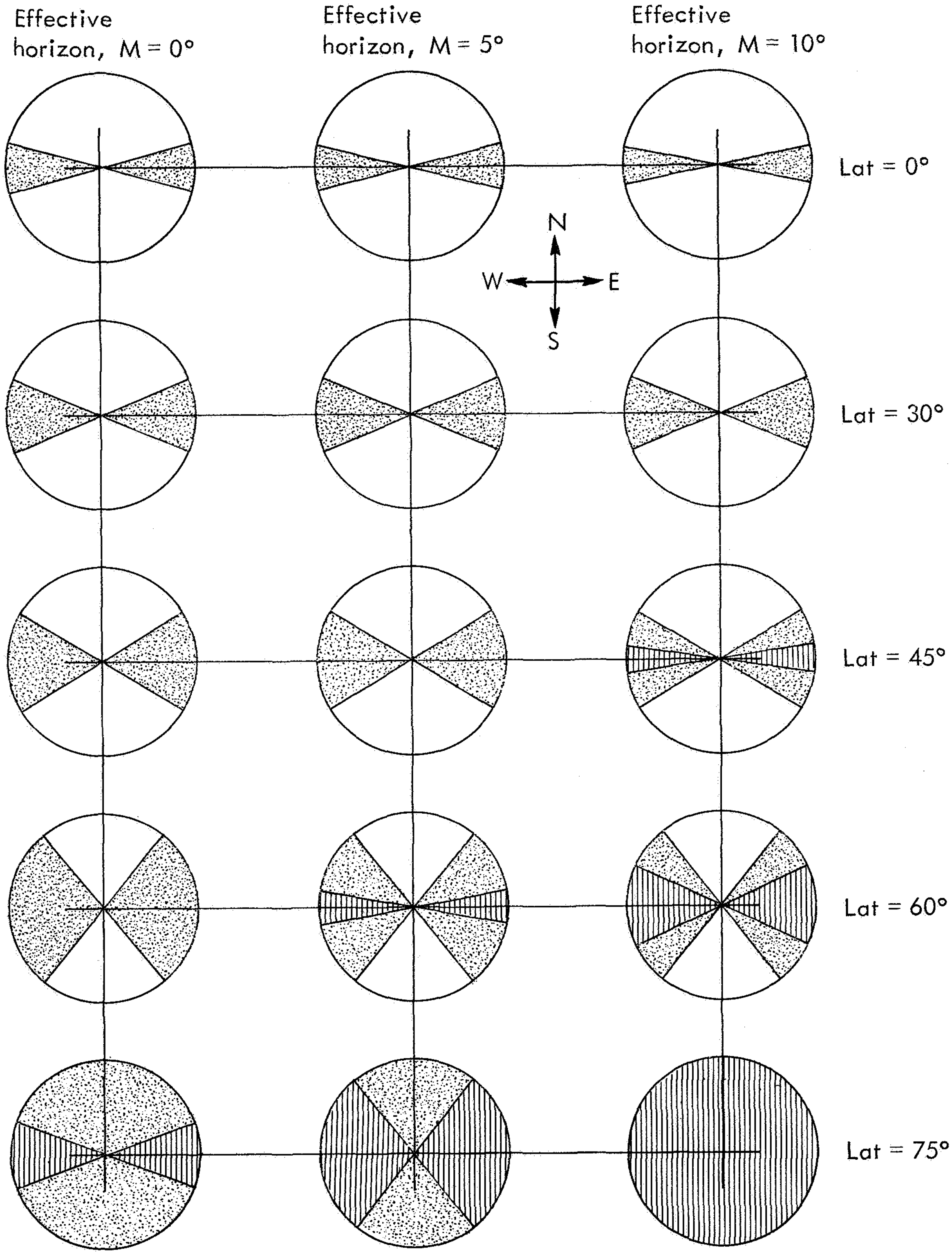


Fig. B-5—Azimuthal domains of interference at microwave relay stations
(Inclination = 0°, B = 15°)

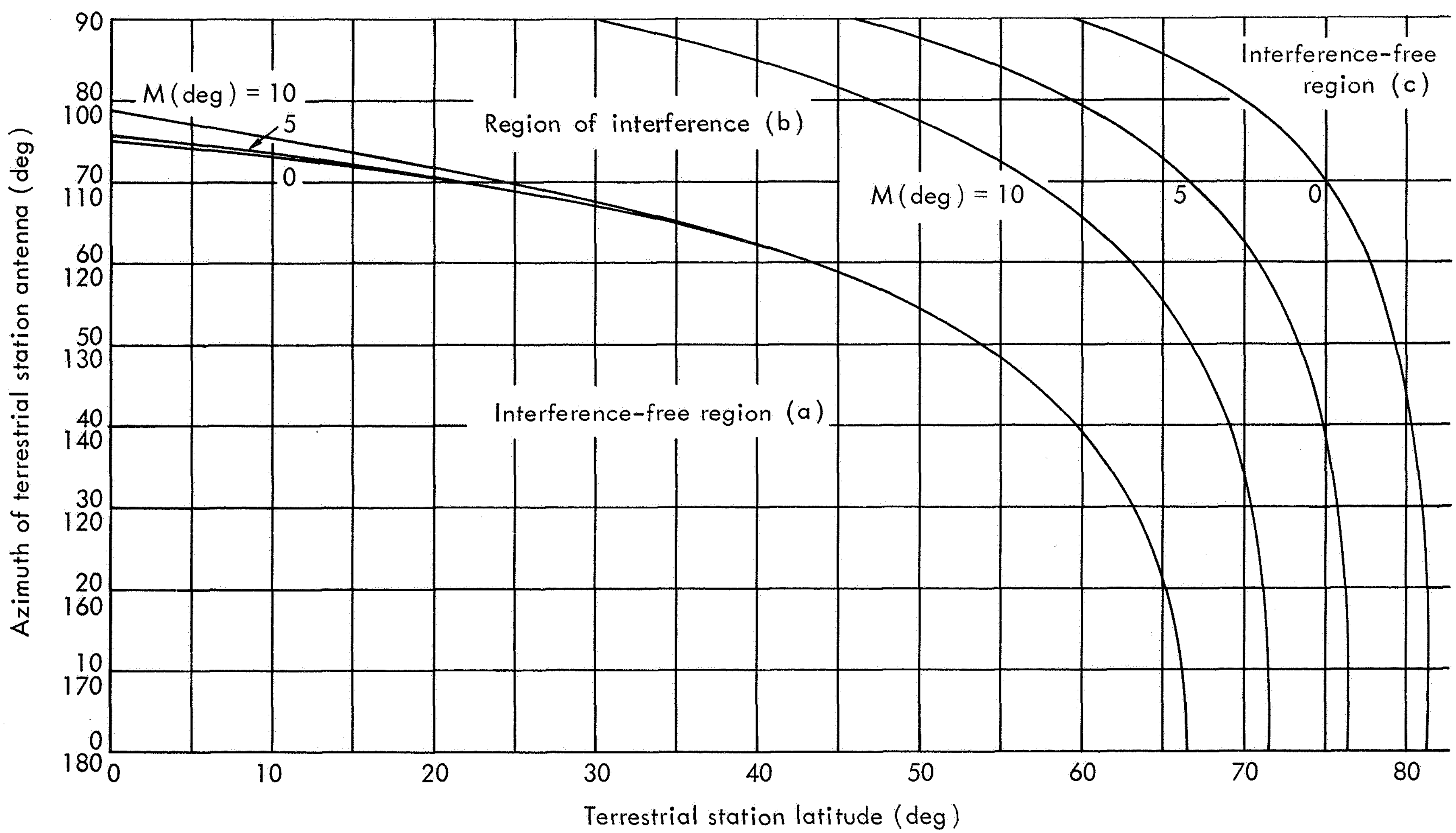


Fig. B-6—Interference regions for B = 15° as a function of latitude and M (Inclination = 0°)

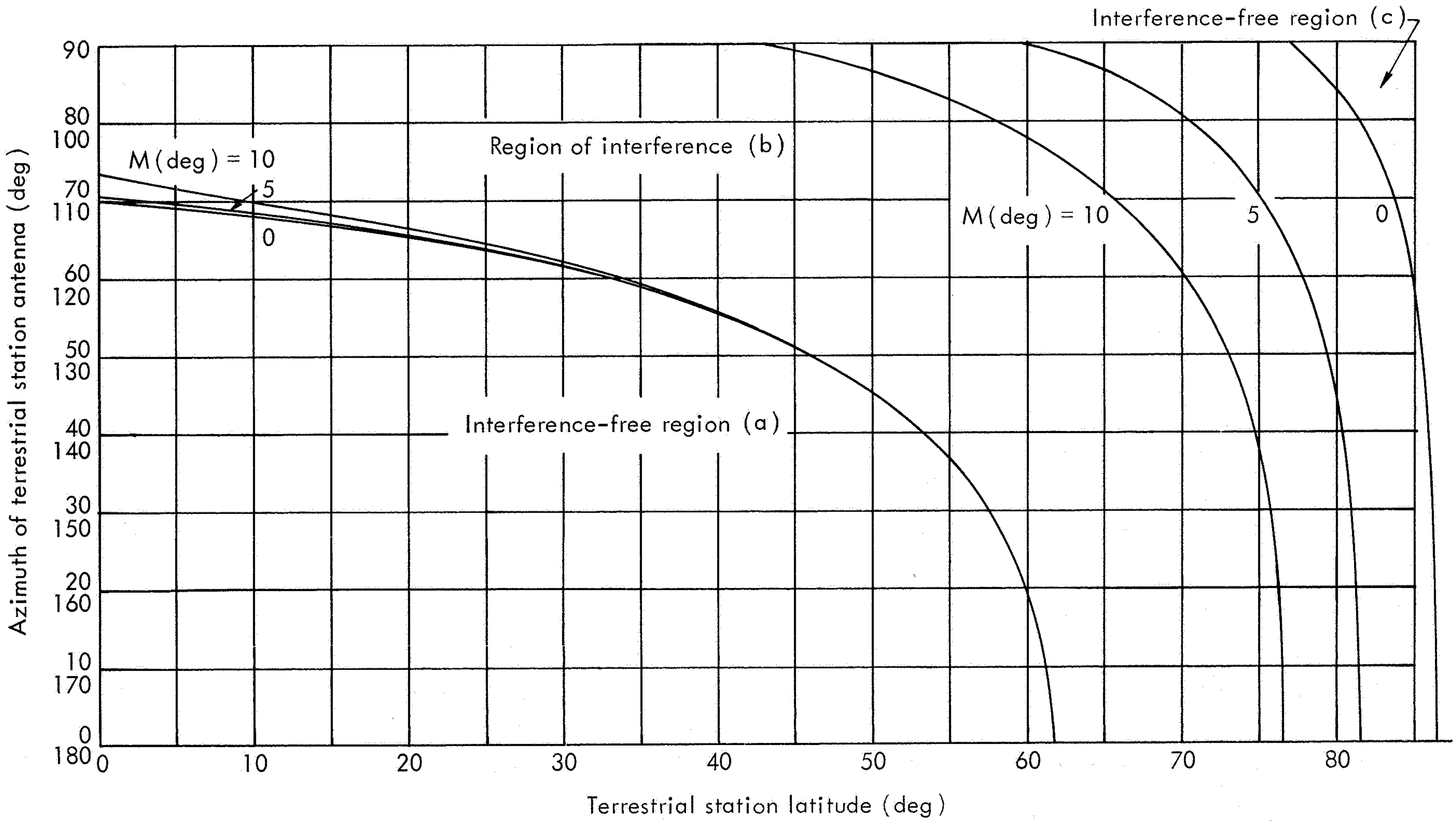


Fig. B-7 — Interference regions for B=15° as a function of latitude and M (Inclination = 5°)

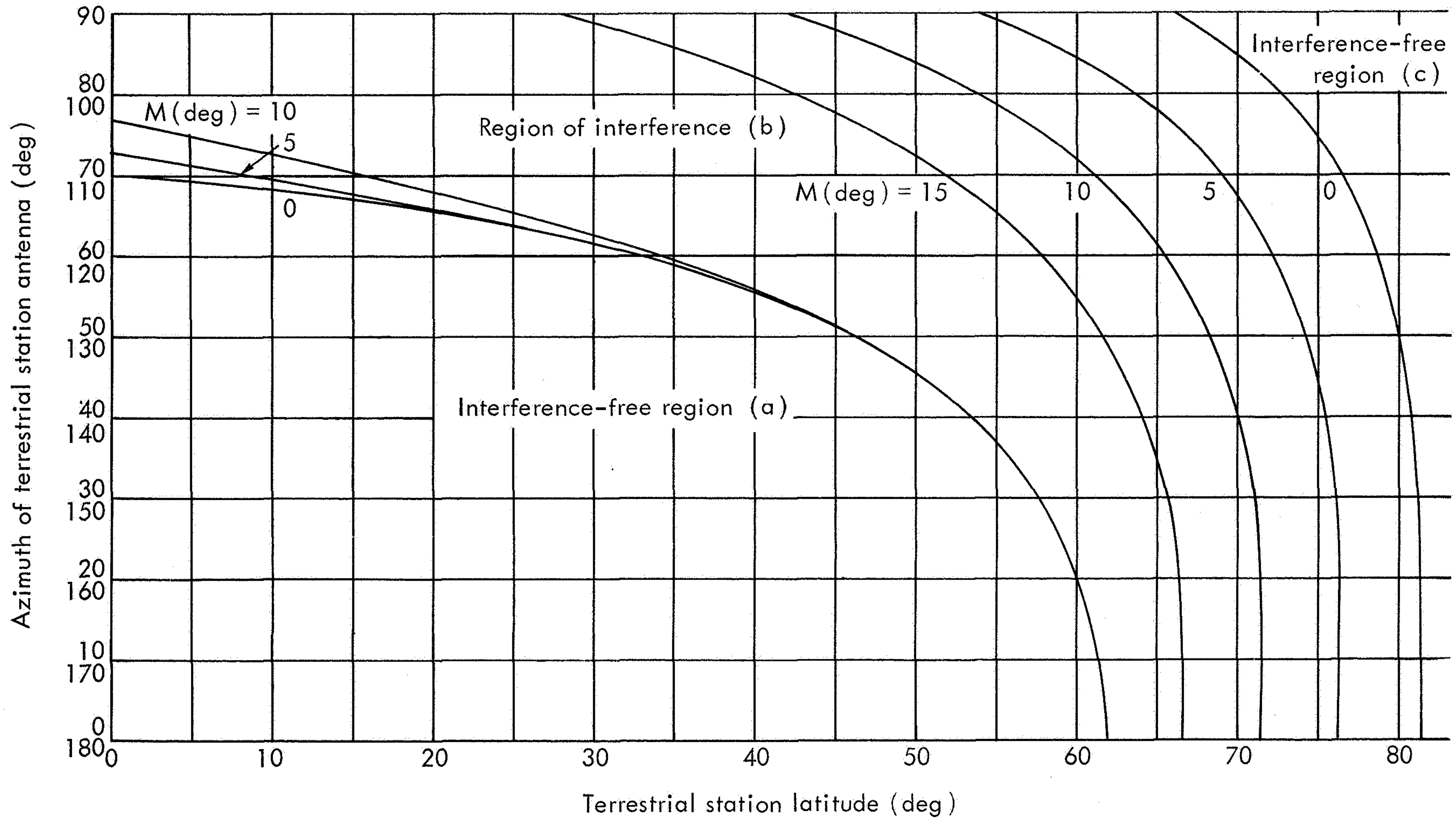


Fig. B-8 — Interference regions for B=20° as a function of latitude and M (Inclination = 0°)

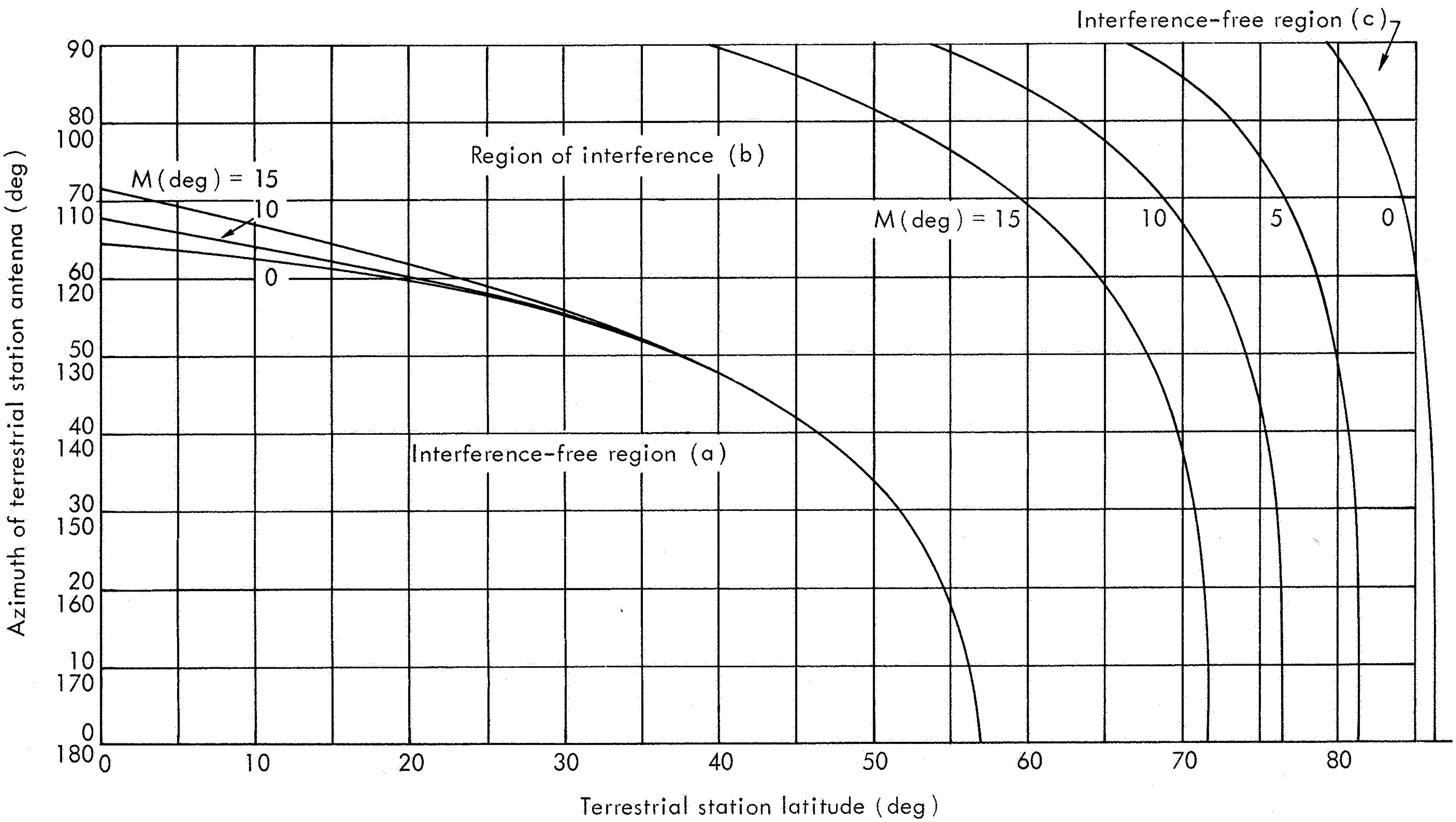


Fig. B-9 — Interference regions for B = 20° as a function of latitude and M (Inclination = 5°)

Appendix C

PROPAGATION FACTORS AND PRECIPITATION SCATTER INTERFERENCE

PROPAGATION FACTORS

Both earth-based microwave radio-relay systems and satellite communication systems have been assigned common frequencies in the lower end of the centimeter region of the spectrum because of a common need for wide bands suitable for line-of-sight propagation. While it would be desirable to use higher microwave and millimeter frequencies to avoid interference from overcrowding of the frequency spectrum and to obtain larger bandwidths, this has so far been avoided because of degraded propagation from atmospheric effects that increase rapidly with frequency.

In a clear atmosphere, the atmospheric propagation loss and noise are relatively small for frequencies less than about 10 GHz, but increase rapidly for higher frequencies, primarily due to water vapor and oxygen. Although these effects are significant, the attenuation through a typical atmosphere, as illustrated in Fig. C-1,[1,2] does not rule out the use of the spectrum above 10 GHz for either terrestrial radio-relay or satellite communications.

Rather, it is attenuation by precipitation, primarily rain, that has restricted the use of these higher frequencies in the atmosphere. The attenuation to be expected from clouds is simply proportional to their liquid water content in the frequency range of interest. This attenuation, as computed by Ryde and Ryde,[3-5] is shown in Fig. C-2 for temperatures of 0^{o} and 20^{o} C. Use of these curves requires knowledge of the water density and thickness of the cloud, both of which may vary widely.

Attenuation by rain, on the other hand, is a complex process involving both absorption and scattering. Nevertheless, this attenuation can be calculated for a specified density and size distribution[3-8] of drops, as shown in Fig. C-3.

The principal practical problem is to determine the expected spatial distribution of rain and its statistics for any locality in order to plan diversity that will give the desired system performance. The drop size

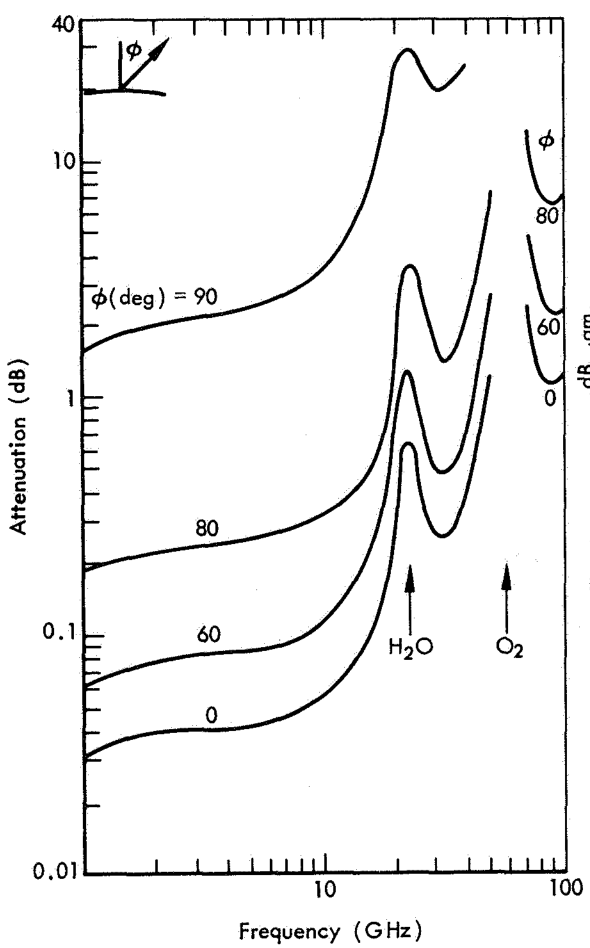


Fig. C-1 — One-way attenuation through the total atmosphere (O_2 and H_2O vapor)

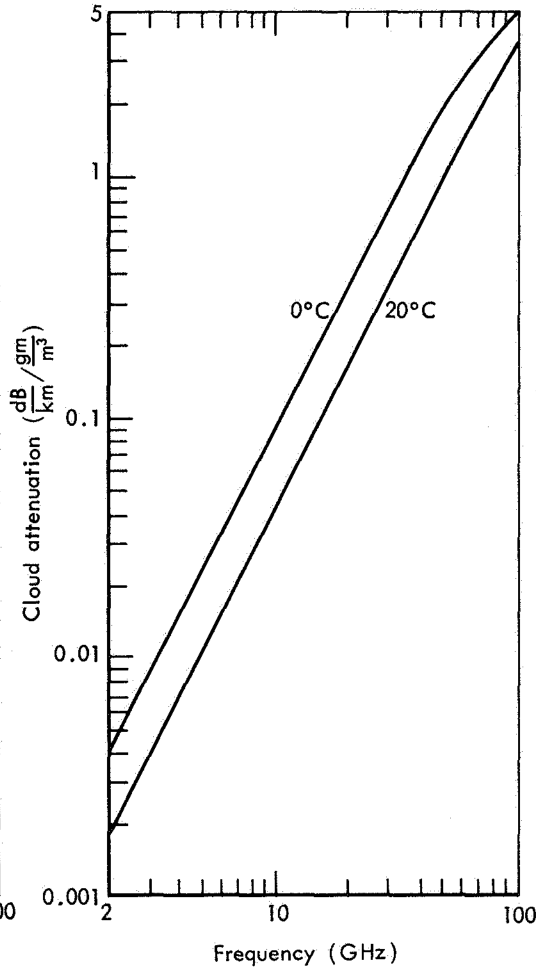


Fig. C-2 — Cloud (fog) attenuation coefficients

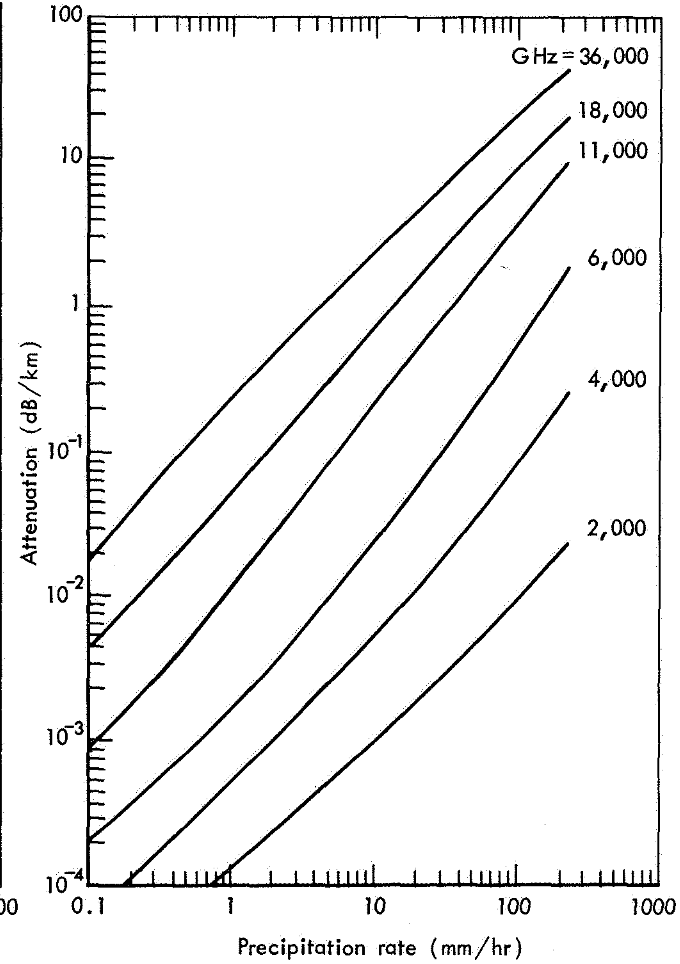


Fig. C-3 — Theoretical rainfall attenuation distribution of Laws and Parsons[9] is generally accepted as a reasonable basis for the prediction of rain attenuation.

Experience with rain at low altitudes, where rain gauges on the ground can be used to determine rain rates,[10] has demonstrated that very high rain intensities (e.g., > 100 mm/hr) usually occur in cells of a few km characteristic horizontal dimensions. If similar spatial distributions of intensities and characteristic dimensions exist for high-altitude storms, there is promise that systems exploiting space diversity could greatly alleviate problems of attenuation at the higher frequencies.

However, the great demand for spectrum occasioned by the rapid proliferation of radiating systems will probably lead to intense frequency sharing by microwave relays, communication satellites, and other systems. In this environment, and particularly at the higher frequencies above 10 GHz, precipitation scatter interference may become an important compatibility consideration, and it is important to know how it may constrain system design parameters.

PRECIPITATION SCATTER INTERFERENCE

Interference from precipitation scatter may become important in two principal modes when microwave relays and communication satellites share common frequencies:

- o Interference scattered from terrestrial microwave relays to satellite earth-station receivers.
- o Interference scattered from earth stations to microwave relay receivers.

At present, both the 4- and 6-GHz bands are used for terrestrial microwave relays and for satellite communications, and the first mode can occur in the 4-GHz band (satellite down link) and the second in the 6-GHz band (satellite up link). As other bands are shared by these services in the future, both modes may arise, depending on which frequencies are used for the satellite up and down links.

The computations described here provide rough estimates of the radiated powers and the constraints on other parameters which will permit compatible frequency sharing between microwave relays and communication satellites with respect to precipitation scatter interference.

Precipitation Scatter Model

Since it has been shown that nonintersecting beams are not likely to produce harmful interference from precipitation scatter,(11) the only interference to be considered is that between the intersecting main beams of an earth-station antenna of a synchronous satellite communication system and a terrestrial microwave relay antenna. The geometry of the scattering problem is shown in Fig. C-4.

An earth-station antenna with diameter D_e is pointed at a fixed azimuth with elevation angle ϵ_e. A terrestrial station with antenna diameter D_t and elevation angle ϵ_t is located at a distance d_{et} from the earth station with an azimuth such that its beam axis intersects the earth-station beam axis. Using the 4/3 earth radius approximation to allow for refraction effects, the lengths of the propagation paths from the earth station to the beam intersection, d_{xe}, and from the terrestrial station to the intersection, d_{xt}, can be computed.

These path lengths are combined with two different precipitation models in order to calculate the extremes of carrier-to-interference ratios resulting from precipitation scatter interference.

Analysis

The analysis first considers the mode involving earth-station scatter interference at the terrestrial-station receiver. For the large parabolic dish antennas presently used for earth stations, the Fresnel region extends several kilometers along the axis from the antenna. For example, for an 85-ft parabolic dish (25.9 m) operating at 4 GHz (wavelength λ = 7.5 cm), the Fresnel region extends a distance of $2\ D_e^2/\lambda$ = 17.9 km. Since most rainstorms, including thunderstorms, seldom reach a height of 20 km, it is assumed for the purpose of this analysis that all of the power transmitted by the earth station passes through the rain cloud in a collimated beam with diameter D_e.* If the rain cloud encloses the entire volume of intersection

*Although this assumption greatly simplifies the analysis, it does not restrict the applicability of the results since the carrier-to-interference ratio at the terrestrial station can be shown to be relatively independent of the earth-station antenna gain for gains greater than $\sim$ 20 dB.

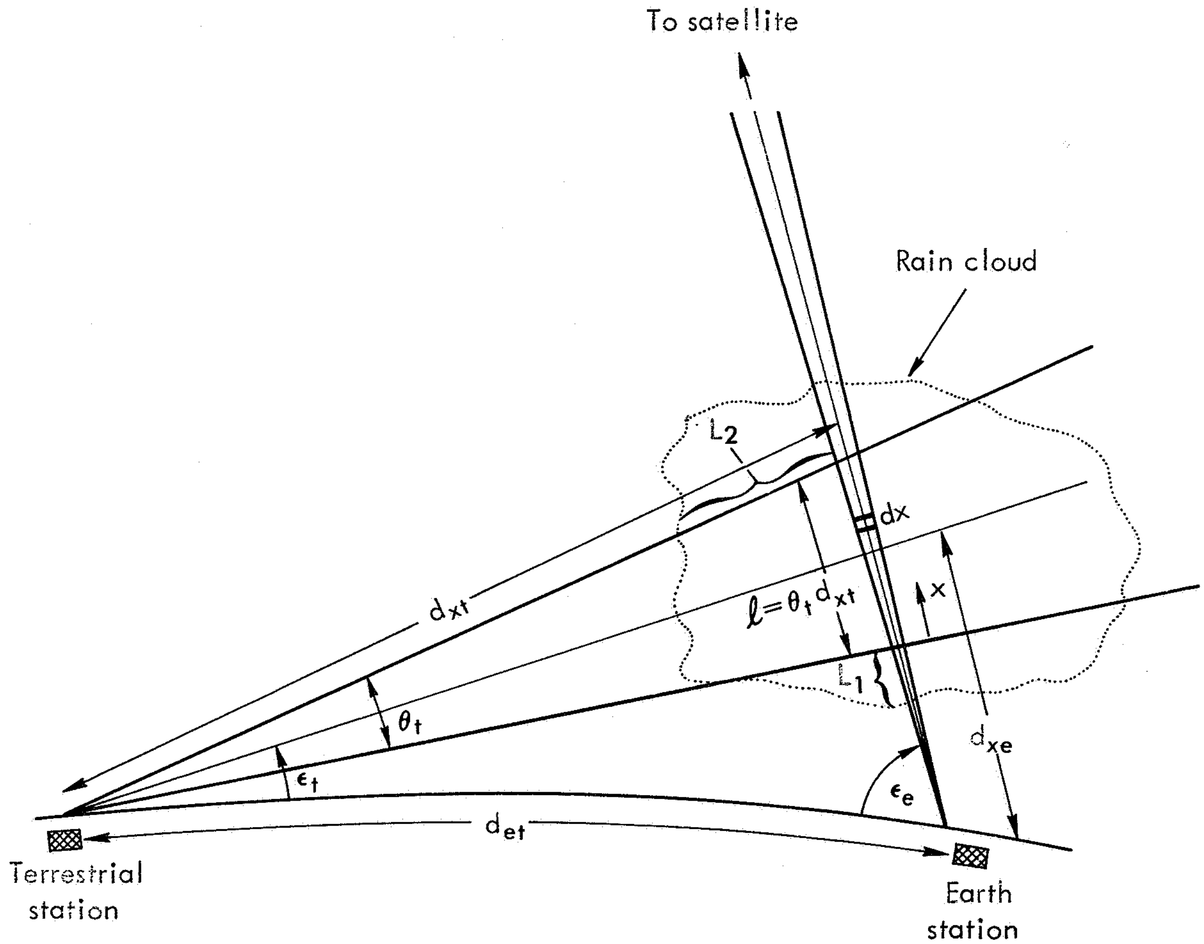


Fig. C-4 — Geometry of scattering problem of the two antenna beams, as shown in Fig. C-4, then the region which contributes to precipitation scatter can be considered as limited to this common volume.

Furthermore, since the terrestrial-station antenna beamwidth is usually much wider than that of the earth station at the intersection and since the earth-station antenna elevation ϵ_e is usually much greater than ϵ_t, this common volume can be approximated* by a cylinder of diameter D_e and length $\ell = \theta_t d_{xt}$, where $\theta_t = \lambda/D_t$ is the beamwidth of the terrestrial-station antenna.

In this case, the flux density illuminating the common volume is, to a good approximation,

$$P_\sigma = \frac{P_e}{A_e} e^{-\beta_{ex} L_1} \qquad \text{(C-1)}$$

and the amount of this flux which is scattered from an element of unit cross section and length dx at depth x in the common volume is

$$P_\sigma e^{-\beta_{ex} x} \beta_{sc} dx$$

In these expressions, P_e is the power transmitted by the earth station, A_e is the effective aperture of the earth-station antenna, L_1 is the length of precipitation the beam passes through before reaching the common volume, and β_{ex} and β_{sc} are the extinction and scattering coefficients defined in Ref. 7.** Although the present analysis assumes isotropic scattering,

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*See footnote on previous page.

**In polydispersed rain, the values of the extinction and absorption cross sections per unit volume, β_{ex} and β_{ab}, are determined by an integral of the form

$$\beta[m,n(r)] = \pi \int r^2 n(r)\, K(m,r) dr$$

and the scattering coefficient β_{sc} is given by their difference. Here r is the geometrical radius of a raindrop, n(r) is the raindrop size distribution function, m is the complex index of refraction of the rain, and K(m,r) is a measure of the efficiency of a drop in extinction. Series expansions for K_{sc} and K_{ab} may be found in Refs. 7 and 12.

the anisotropy of the scattering can be accounted for by including the proper directional factor in the previous expression as discussed in Refs. 7 and 11. This refinement was not felt to be justified in the approximate treatment given here.

The total power scattered from the common volume is

$$P_{sc} = \int_0^{\ell} P_{\sigma} e^{-\beta_{ex} x} \beta_{sc} A_e dx$$

$$= \frac{P_{\sigma} \beta_{sc} (1 - e^{-\beta_{ex} \ell}) A_e}{\beta_{ex}} \qquad \text{(C-2)}$$

Due to the assumed isotropy of scattering, the power per unit solid angle scattered toward the terrestrial station is $P_{sc}/4\pi$. If the effective aperture of the terrestrial-station antenna is A_t, it subtends a solid angle A_t/d_{xt}^2, and so the scattered power received by the terrestrial-station antenna is

$$X_{xt} = \left(P_{sc} e^{-\beta_{ex} L_2}\right) \frac{{}_tG \lambda^2}{16\pi^2 d_{xt}^2} \qquad \text{(C-3)}$$

where L_2 is the length of precipitation along the beam axis between the common volume and the terrestrial station, and ${}_tG = 4\pi A_t/\lambda^2$ is the gain of the terrestrial-station antenna.

The desired carrier signal received at a terrestrial station from a neighboring terrestrial station is given as

$$C_{tt} = \frac{P_t \cdot {}_tG^2}{16\pi^2 d_{tt}^2} \lambda^2 e^{-\beta_{ex} L_4} \qquad \text{(C-4)}$$

where P_t is the terrestrial-station power, d_{tt} is the line-of-sight path between the two terrestrial stations, and L_4 is the portion of d_{tt} filled with the rain, so that $e^{-\beta_{ex} L_4}$ accounts for the rain attenuation of the wanted signal. Both terrestrial stations are assumed to use identical antennas.

Combining the previous expressions, the carrier-to-interference ratio at the terrestrial station can be expressed as

$$\frac{C_{tt}}{X_{xt}} \equiv \left(\frac{C}{X}\right)_t = \left(\frac{P_t \cdot {}_tG}{P_e}\right)\left(\frac{d_{xt}}{d_{tt}}\right)^2 \cdot \frac{e^{-\beta_{ex}(L_4-L_1-L_2)}}{\left(\frac{\beta_{sc}}{\beta_{ex}}\right)\left(1-e^{-\beta_{ex}\ell}\right)} \tag{C-5}$$

On the basis of reciprocity, it can be shown that the interference in the mode of a terrestrial microwave relay interfering with a satellite earth station is given by an equation of the same form as Eq. (C-3). If the aperture A_e and the gain ${}_eG$ of the earth station are related by ${}_eG = 4\pi A_e/\lambda^2$, then the desired carrier signal received at the earth station from a synchronous satellite is given as

$$C_{se} = \frac{P_s \cdot {}_sG \; {}_eG}{16\pi^2 d_{se}^2} \lambda^2 e^{-\beta_{ex}L_3} \tag{C-6}$$

where $P_s \cdot {}_sG$ is the equivalent isotropically radiated power of the satellite; d_{se} is the distance from the satellite to the earth station; L_3 is the length along the path d_{se} filled with rain; and $e^{-\beta_{ex}L_3}$ accounts for the rain attenuation of the signal.

In this case the carrier-to-interference ratio at the earth station is

$$\frac{C_{se}}{X_{xe}} \equiv \left(\frac{C}{X}\right)_e = \left(\frac{P_s \cdot {}_sG}{P_t}\right)\left(\frac{d_{xt}}{d_{se}}\right)^2\left(\frac{{}_eG}{{}_tG}\right) \frac{e^{-\beta_{ex}(L_3-L_1-L_2)}}{\left(\frac{\beta_{sc}}{\beta_{ex}}\right)\left(1-e^{-\beta_{ex}\ell}\right)} \tag{C-7}$$

<u>Computations and Results</u>

The extinction coefficient β_{ex}, the absorption coefficient β_{ab}, and the albedo $A \equiv (\beta_{sc}/\beta_{ex})$ have been computed by Deirmendjian(7) using the Mie theory(13) of single scattering, for polydispersed rain irradiated by microwaves of various frequencies. The results of these computations, which correspond to a moderate rain rate of 7.7 mm/hr, are presented in Fig. C-5. Using linear extrapolation to obtain the extinction coefficients for rain rates above 7.7 mm/hr and assuming the albedo to remain approximately constant with increasing rain rate, Eqs. (C-5) and (C-7) were used to compute the values of satellite EIRP and earth-station EIRP required to permit operation in the presence of precipitation scatter interference.

For the case of a terrestrial radio relay interfering with an earth station receiving from a synchronous satellite, Eq. (C-7) has been evaluated at 4 GHz with D_t = 2.34 m, D_e = 25.9 m, and at 11 GHz with D_t = 1.53 m, D_e = 9.14 m, and the results are shown in Figs. C-6 and C-7 in the form of satellite EIRP required to provide a given $(C/X)_e$ for a given rain rate, for several values of separation d_{et} between the interfering source and the receiver. For both frequencies the terrestrial station is assumed to have a 1-W transmitter, which results in the values of $(EIRP)_s$ shown on the figures, and in both cases it is assumed that $\epsilon_e = 30^o$, $\epsilon_t = 2^o$, d_{se} = 35,800 km, and $L_4 = (\ell + L_1)$.

The solid curves of Figs. C-6 and C-7 correspond to the case where only the common volume of the antenna beams is filled with rain, and the dashed curves correspond to the case where the entire volume containing the beams is filled with rain. Although the solid curves represent an extremely unlikely physical situation, they give the upper bound on the satellite EIRP required when this scatter interference dominates the unwanted signals. The solid and dashed curves together can be used to evaluate the required satellite EIRP in the more realistic case where it may be raining very hard in the common volume of the antenna beams and at a lesser rate over the rest of the volume containing the beams by simply subtracting the difference between the solid and dashed curves at the lower rain rate from the value of the solid curve at the higher rain rate. For example, consider the case at 4 GHz of a required $(C/X)_e$ of 30 dB and a separation d_{et} of 100 km. If there is a continuous rain rate of 15 mm/hr everywhere except for a higher rate in the common volume, the scatter interference effects can be estimated as follows: The difference between the solid and dashed curves at 15 mm/hr is $\sim$ 6.5 dB, so at 100 mm/hr in the common volume the satellite EIRP required is 37 dBw (43.5 - 6.5 dBw).

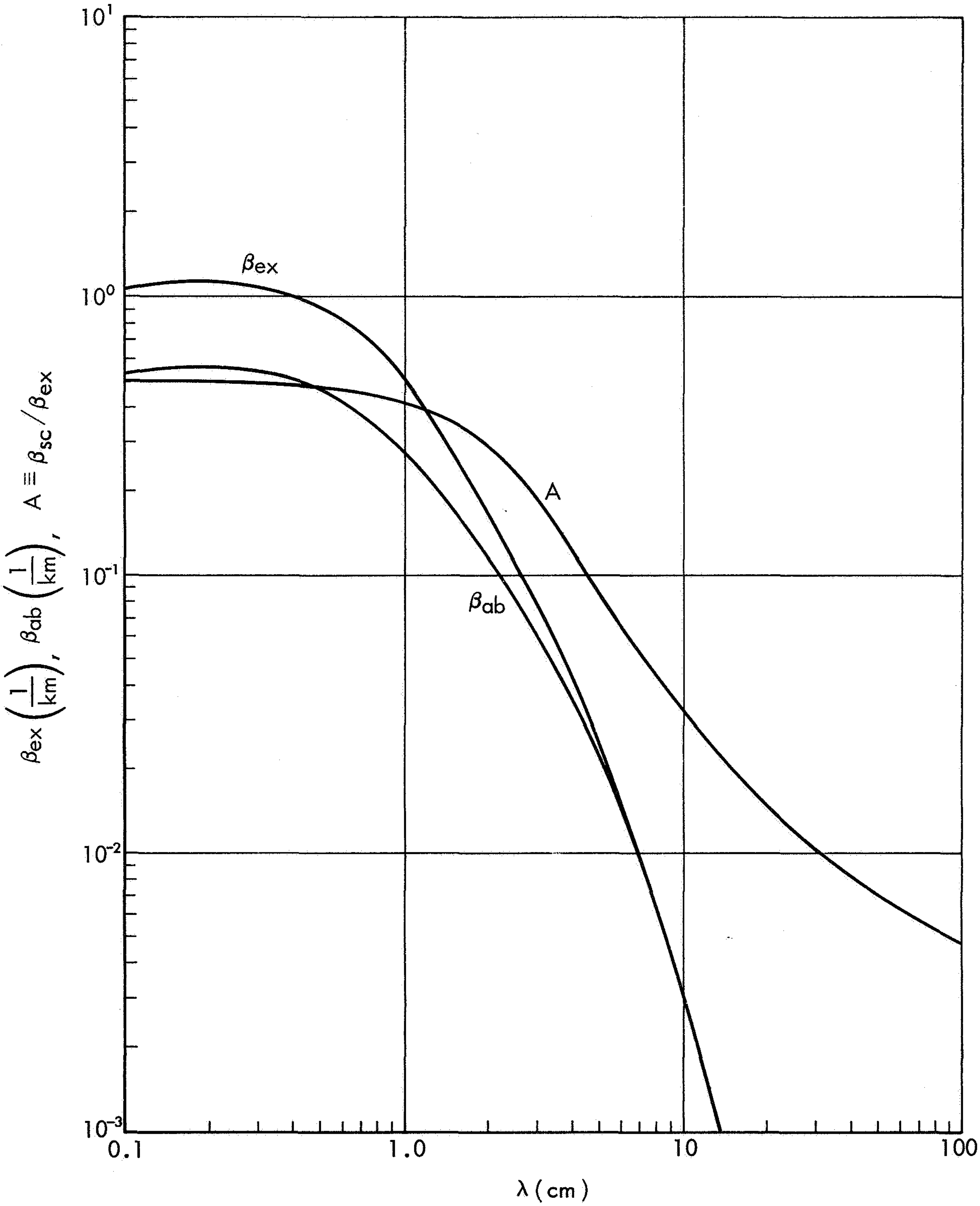

Fig. C-5— Microwave extinction coefficients and albedo for polydisperse precipitation

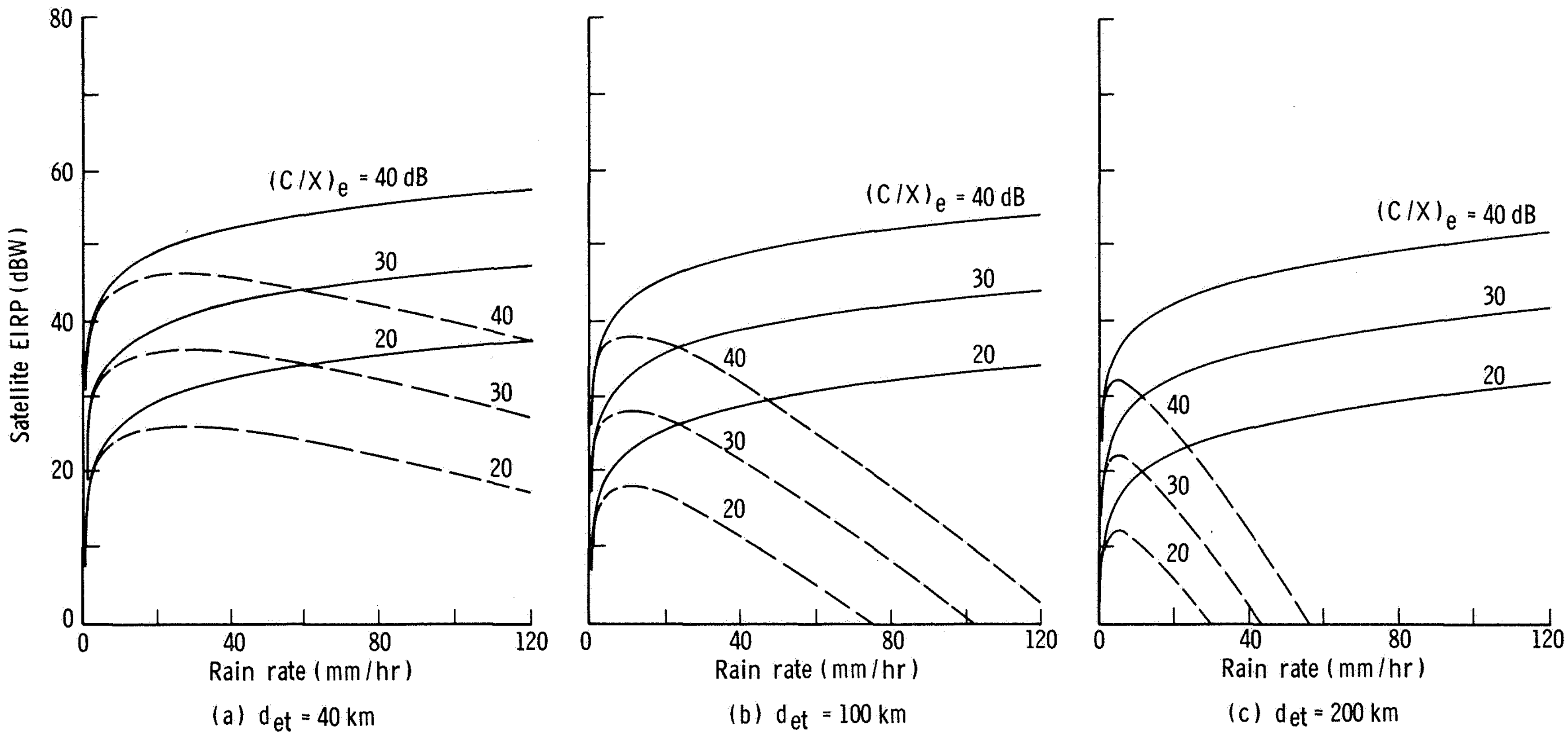

Fig. C-6—Scatter interference, microwave relay to earth station

Frequency = 4 GHz
D_e = 25.9 m, D_t = 2.34 m, P_t = 1W, EIRP in bandwidth of P_t

——— Rain in common volume only
— — — Rain in entire volume

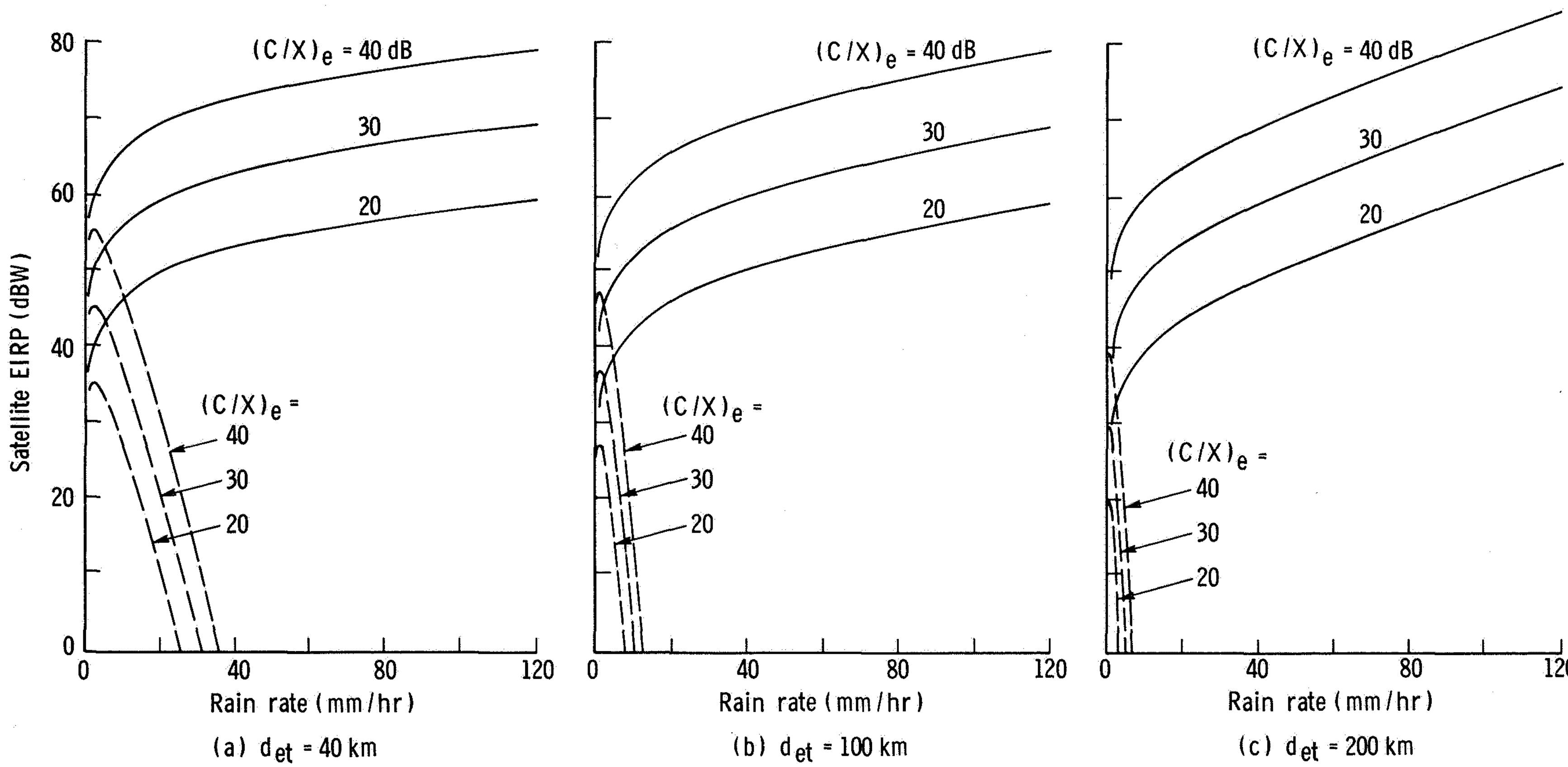

Fig. C-7—Scatter interference, microwave relay to earth station

Frequency = 11 GHz
D_e = 9.14 m, D_t = 1.53 m, P_t = 1W, EIRP in bandwidth of P_t

——— Rain in common volume only
– – – Rain in entire volume

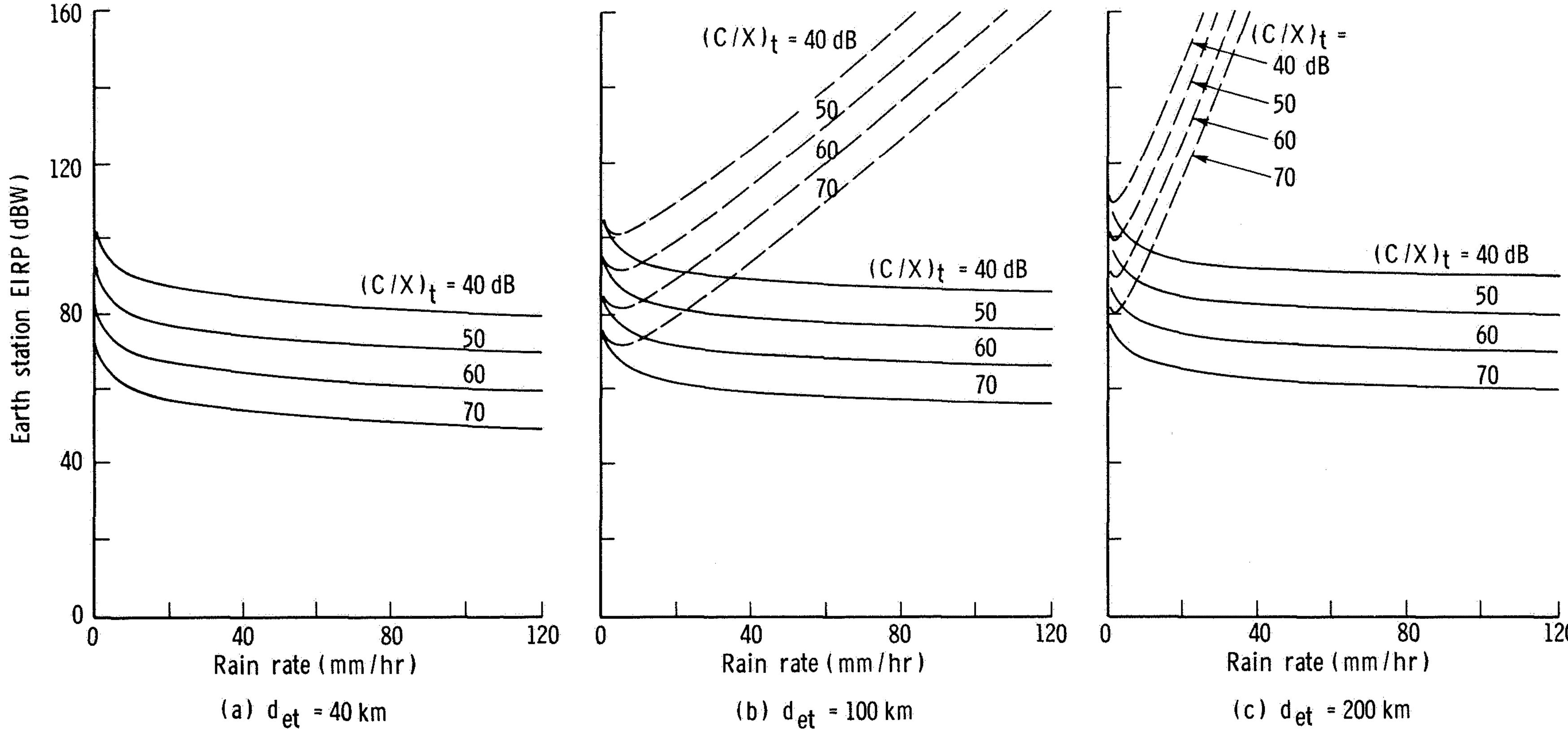

Fig. C-8—Scatter interference, earth station to microwave relay

Frequency = 6 GHz

D_e = 25.9 m, D_t = 1.83 m, P_t = 1W, EIRP in bandwidth of P_t

——— Rain in common volume only

– – – Rain in entire volume

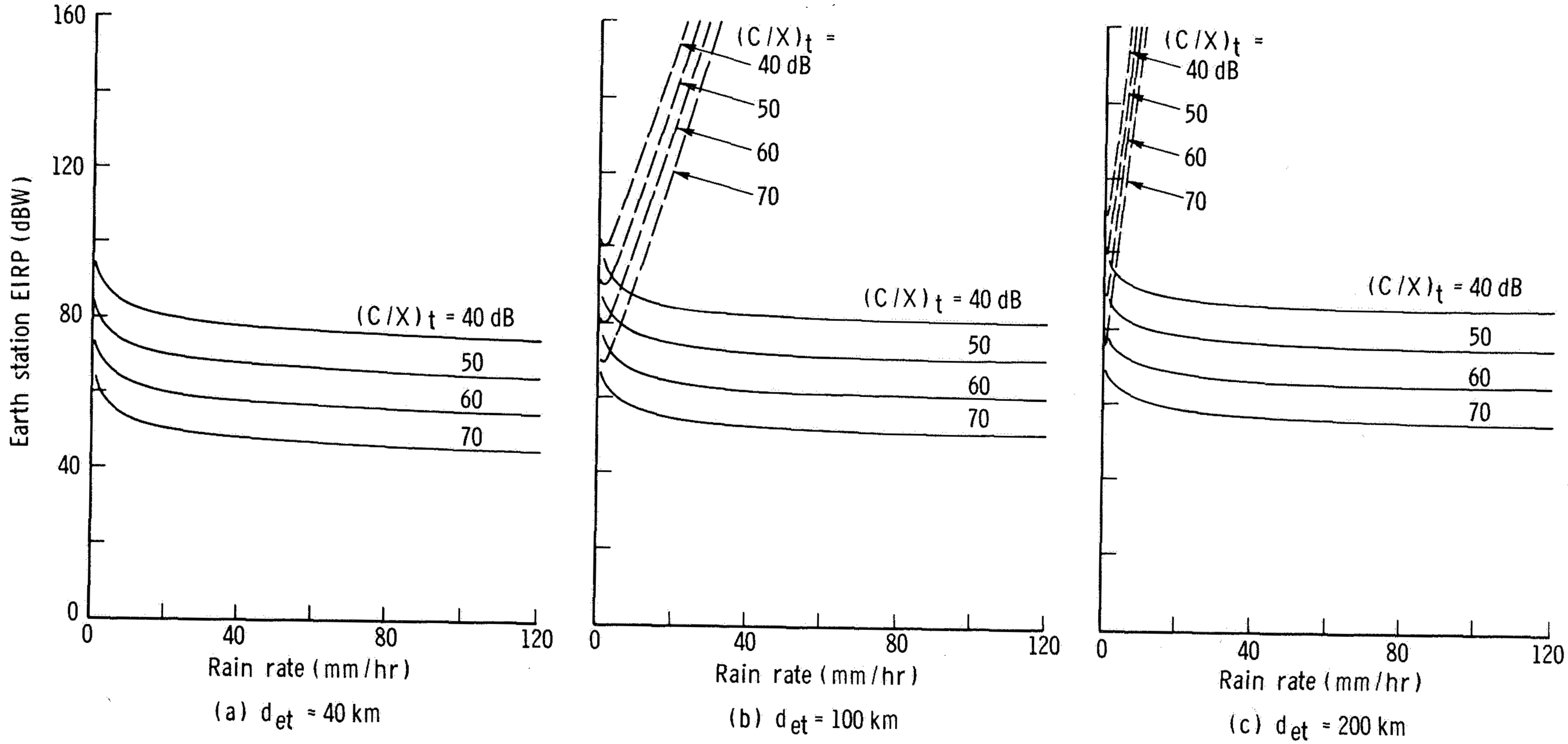

Fig. C-9—Scatter interference, earth station to microwave relay

Frequency = 11 GHz
D_e = 9.14 m, D_t = 1.53 m, P_t = 1W, EIRP in bandwidth of P_t
——— Rain in common volume only
— — — Rain in entire volume

For the case of transmitting earth station interfering with a terrestrial radio-relay receiver, Eq. (C-5) has been evaluated at 6 GHz with D_t = 1.83 m, D_e = 25.9 m and at 11 GHz with D_t = 1.53 m, D_e = 9.14 m, and the results are shown in Figs. C-8 and C-9 in the form of limitations on earth-station EIRP for several values of separation d_{et} between the earth station and the terrestrial station. For both frequencies the terrestrial stations are assumed to have 1-W transmitters, to be spaced (d_{tt}) 40 km apart, and to have $\epsilon_e = 30^o$, and $\epsilon_t = 2^o$.

Again the solid curves of Figs. C-8 and C-9 correspond to the case where only the common volume of the antenna beams is filled with rain, and the dashed curves correspond to the case where the entire volume containing the beams is filled with rain. For a separation d_{et} of 40 km, the solid and dashed curves coincide since the wanted and unwanted signals are attenuated equally. These curves can be used to evaluate the permitted earth-station EIRP in the more realistic situation where it may be raining harder in the common volume of the antenna beams than everywhere else by simply <u>adding</u> the difference between the dashed and solid curves at the lower rain rate to the value of the solid curve at the higher rain rate.

<u>Discussion</u>

Figures C-6 through C-9 show that precipitation scatter interference can be deleterious even for large separations d_{et} between the interfering source and the receiver. The results are presented for various values of d_{et} with the assumption that the common volume is on-path between the two stations. The results for other geometries can be determined as needed.

Also, the calculations presented here are all based on a model of <u>continuous</u> rain. Although continuous rain does in fact fall over large areas and may be considered an extreme representative of rain type, the other extreme is <u>convective</u> rain, which is typically represented by thundershowers during the summer months. In a convective rainstorm, there are heavy small-area concentrations of rain (cells) which shift around within a wide-area storm. The curves of Figs. C-6 to C-9 are especially useful in that they may not only be used to evaluate the bounds on satellite and earth-station EIRP for continuous rain, but may also be used for the case of convective rain with a high-intensity cell in the common volume.

While the theoretical results presented here show the relations between the pertinent parameters involved in precipitation scatter interference, their use requires a knowledge of the time and spatial distributions of precipitation for the locations involved.

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Appendix D

FREQUENCY DEPENDENCE OF LOW-NOISE AND OUTPUT DEVICES

MICROWAVE RECEIVER SENSITIVITY AND COST*

Based on published research on microwave low-noise devices and manufacturers' data on currently available models, Fig. D-1 compares their bounds in the noise-figure-frequency domain. The figure includes only the more common low-noise devices. Thus resistive diode mixers and traveling-wave masers are included, but parametric frequency converters and cavity masers are excluded. As an example of the significance of the limits or bounds of Fig. D-1, consider the tunnel diode amplifier. Few tunnel diode amplifiers would lie outside the upper bound indicated except, perhaps, for early experimental designs in which noise-figure performance was traded for some other parameter such as power output, bandwidth, or lower cost (as in hybrid integrated circuits). The lower bound lies below currently achievable performance even in the laboratory, and thus represents performance which may be achieved by selected devices under optimum conditions over the next few years.

Two of the low-noise devices which could have widespread utilization over the next decade are the tunnel diode amplifier and the uncooled parametric amplifier. Limitations in the performance of transistors around 10 GHz and above, evident in Fig. D-1, will probably lead to the utilization of the tunnel diode at higher frequencies. The crossover between the two has been steadily moving upwards from around 100 MHz a decade ago to about 6000 MHz today. Where the contributions to system noise figure other than from the receiver itself are sufficiently low, the use of an uncooled parametric amplifier may be worth the increased cost and complexity. Below 10 GHz, wider use of the parametric amplifier would displace the transistor amplifier; at higher frequencies, mixers and low-noise traveling-wave tubes may tend to be displaced in some applications by the parametric amplifier.

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*This material is largely abstracted from a survey of low-noise device technology by N. Feldman which was made during the past year. It covered manufacturers' data sheets, approximately one hundred articles in the letters in the Proc. IEEE, IEEE transactions, the microwave journal, Microwaves, and records of various symposia.

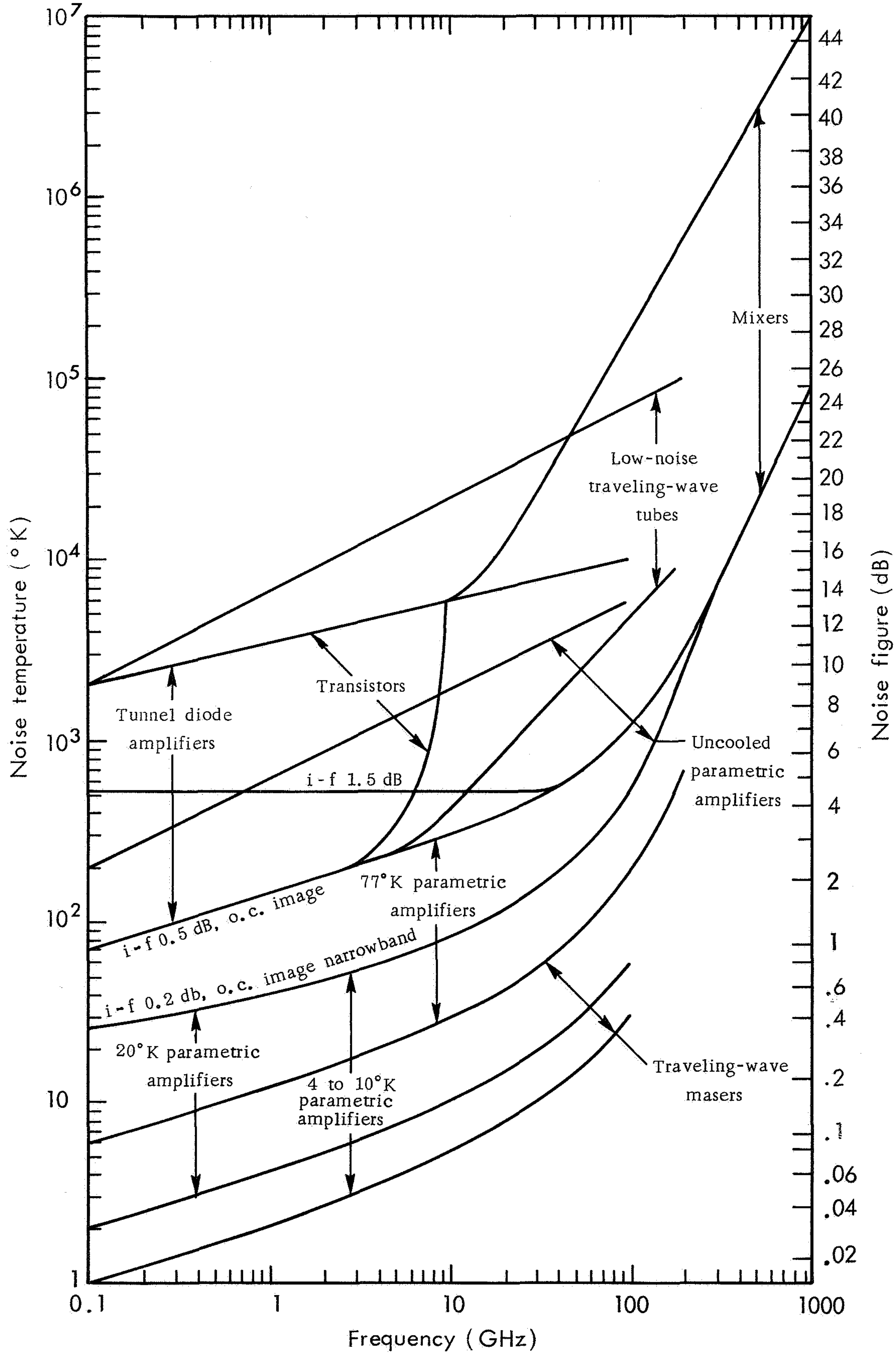

Fig.D-1—Comparison of noise-figure and frequency ranges of low-noise devices

Cooled receivers, such as the cooled parametric amplifier and the traveling-wave maser, are likely to remain relatively expensive for some time to come. The number of applications which can benefit from their use is restricted; i.e., the improvement in system noise figure over a good uncooled parametric amplifier may not be significant in many cases. In some commerical and military airborne communication and in tactical military communication equipment at microwave frequencies, cooled receiver devices may become important. For applications in the 10 to 100 GHz region, where much new system development is likely to occur in the next decade, the tunnel diode amplifier and the uncooled parametric amplifier will be particularly important.

One of the most important parameters in the price of these devices is the cumulative quantity produced. Figure D-2 illustrates price projections versus quantity produced; i.e., it is a curve which reflects both learning and economies of scale for various categories of tunnel diode amplifier and uncooled parametric amplifier.

The extreme prices on each item roughly indicate the costs of differences in performance. The slope of the curves is 80 percent, where slope is defined as the ratio of prices for a factor of two increase in the cumulative quantity produced. Electronic equipment cost does not often decline as far as theoretically indicated in Fig. D-2 because of limited production due to rapid changes in the state of the art, frequent design changes, the large numbers of variants of a given design, etc.

The frequency dependence of the cost of microwave devices is difficult to establish due to the large number of variables involved. Examination of a variety of microwave device price data suggests that when quantity effects are eliminated, price varies slowly with frequency. An exponent of roughly 0.3, i.e., $(\text{frequency})^{0.3}$ could be obtained by various subjective methods. On this basis, price doubles for each decade increase in frequency. These prices do not correspond to the same performance, e.g., to a constant noise figure, but rather to typical performance at each frequency. In the case of tunnel diode amplifier noise figure, for example, the prices correspond to values roughly midway between the upper and lower bounds of tunnel diode amplifier noise figure shown in Fig. D-1.

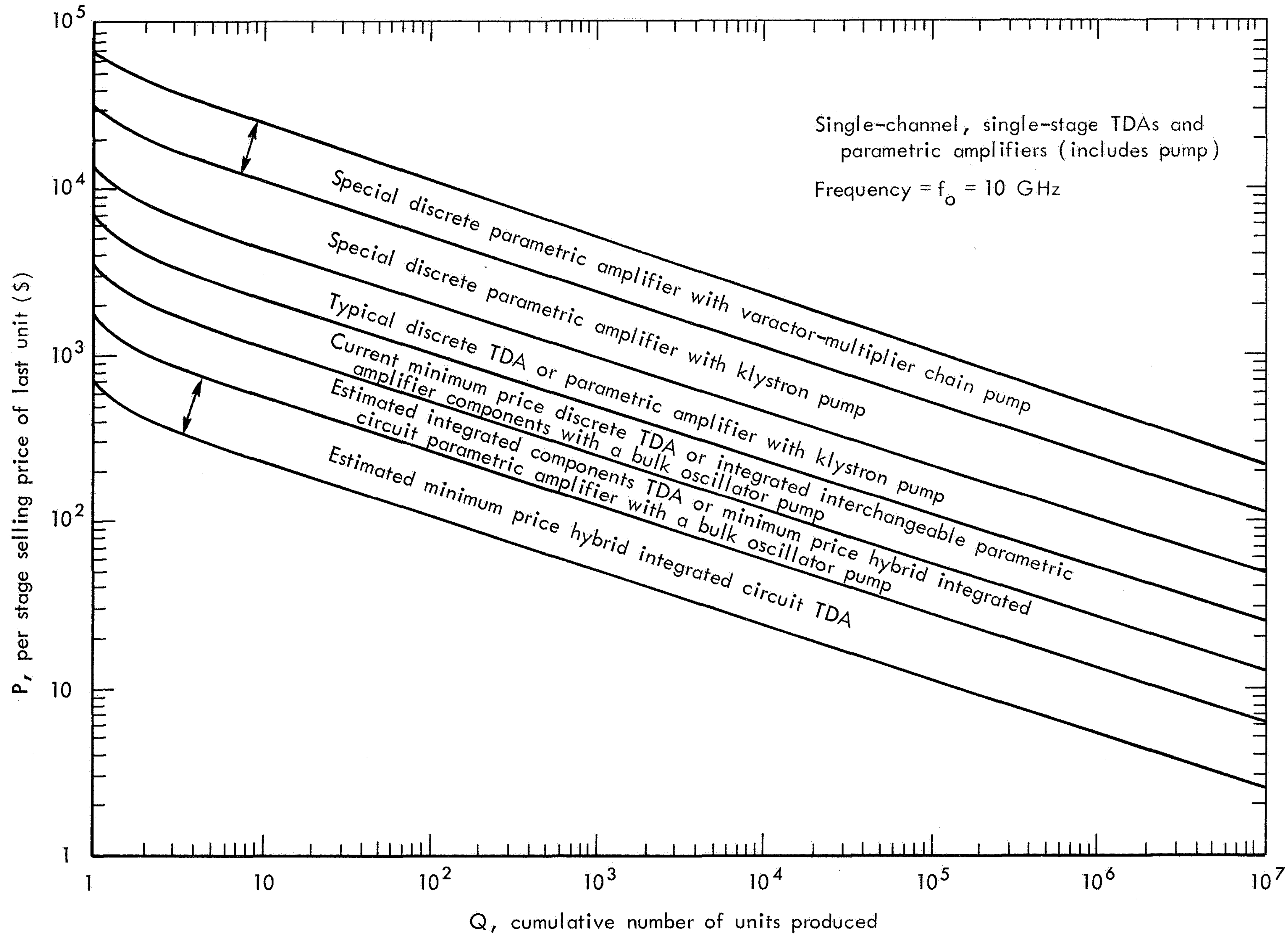

Fig.D-2—Tunnel diode and uncooled parametric amplifier price

The obvious conclusion to be drawn from Fig. D-1 is that noise-figure performance degrades only gradually over the 1 to 100 GHz region. Figure D-2 indicates that some low-noise devices classified today as sophisticated and costly may become commonplace and inexpensive. If the $(\text{frequency})^{0.3}$ dependence of price proves to be approximately correct, then factors other than availability, performance, or cost will determine the choice of frequency for advanced communication systems in the 1 to 100 GHz range.

TRANSMITTER POWER OUTPUT AND EFFICIENCY*

Power output and efficiency for cw Gunn, IMPATT (impact avalanche transit time) and LSA (limited space-charge accumulation) devices are shown in Fig. D-3. The solid lines are smoothed envelopes enclosing as upper bounds the published data through July 1968. Left skirts pass through two data points, the flat tops pass through the highest published power level, and the right skirts all have $1/f^2$ slopes drawn through the point of maximum Pf^2 product (where P is the power output and f the frequency). Power thus has been generated over the entire frequency range covered by these solid curves. Readily available devices and devices designed for wide tuning range or low noise level have power outputs 10 to 20 dB below these values. The dashed lines represent similar envelopes enclosing the most optimistic published estimates of the past two years with the estimates arbitrarily extended down to 1 GHz and up to 1000 GHz. The left half of the LSA curve has frequently appeared in the literature; it is based on a theoretically calculated limit for copper heat sinks on opposite faces.

R&D on various ways to modify the structure of these three devices and to cool them makes these estimates subject to drastic revision. It also is questionable whether a crossover in Gunn and LSA performance will ever exist in either the power or efficiency domain as shown in Fig. D-3, since it now appears that the much more recent LSA devices will be superior in performance.

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*This material is based on a recent survey by N. Feldman of over 150 journal articles and over 50 advertisements and manufacturers' data sheets.

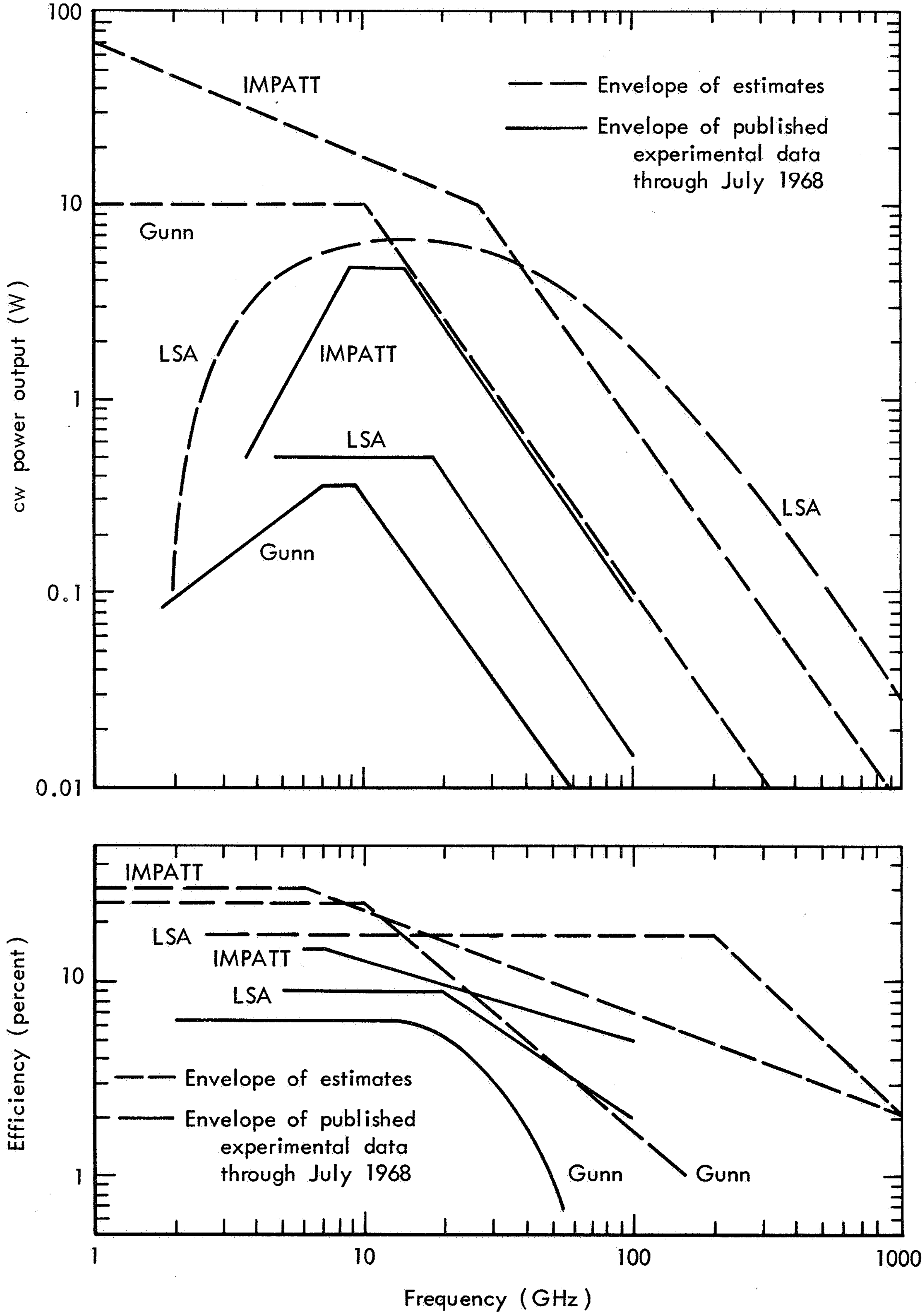

Fig. D-3—cw Gunn, IMPATT, and LSA power output and efficiency

The power output and efficiency of transistors and varactor multipliers are shown in Fig. D-4. The dashed varactor-multiplier curve was specifically generated for this report by calculations using the best performance data for transistor collector efficiency and stage gain and for varactor doubler conversion efficiency and interstage coupling. Although varactor multipliers could achieve high efficiencies in narrow bandwidth circuits with selected devices as indicated by the projected curve, the cost is too high for most applications.

Transistor amplifiers below 10 GHz may achieve efficiencies roughly comparable to those projected for the varactor multipliers within the next few years. Above roughly 10 GHz, LSA and IMPATT devices will both offer better power output and efficiency than the varactor-multiplier chains. Thus above 10 GHz, the varactor multiplier may be limited to special applications, e.g., those requiring exceptionally low noise levels.

Despite the currently low cw power output of 0.1 to 10 W for IMPATT, Gunn, and LSA devices around 10 GHz, they already challenge varactor-multiplier chains and have eclipsed the tunnel diode oscillator. The tunnel diode amplifier is roughly 10 dB lower in saturated power output than the oscillator version. The parametric amplifier output power at the 1-dB gain compression point is -10 to -25 dBm for typical amplifiers from 0.3 to 35 GHz; the corresponding saturated power output varies from +5 to -15 dBm. Pump power ranges from 500 to 10 mW. The parametric amplifier is thus an inefficient, low-power output amplifier, albeit a highly effective low-noise one. The parametric frequency converter has a power output capability comparable to the varactor multiplier and a pump-to-output signal conversion loss as low as 2 to 3 dB. It is thus likely to have fairly widespread use, as its efficiency would be 2 to 3 dB below that of the solid-state source supplying the pump power.

The LSA amplifying mode cannot exist under stable conditions; thus an LSA device can amplify only with injection locking. IMPATT and Gunn devices can operate either as injection locked amplifiers or as stable linear reflection amplifiers. Gunn amplifiers are likely to be utilized primarily in the injection locked mode for most applications to maximize

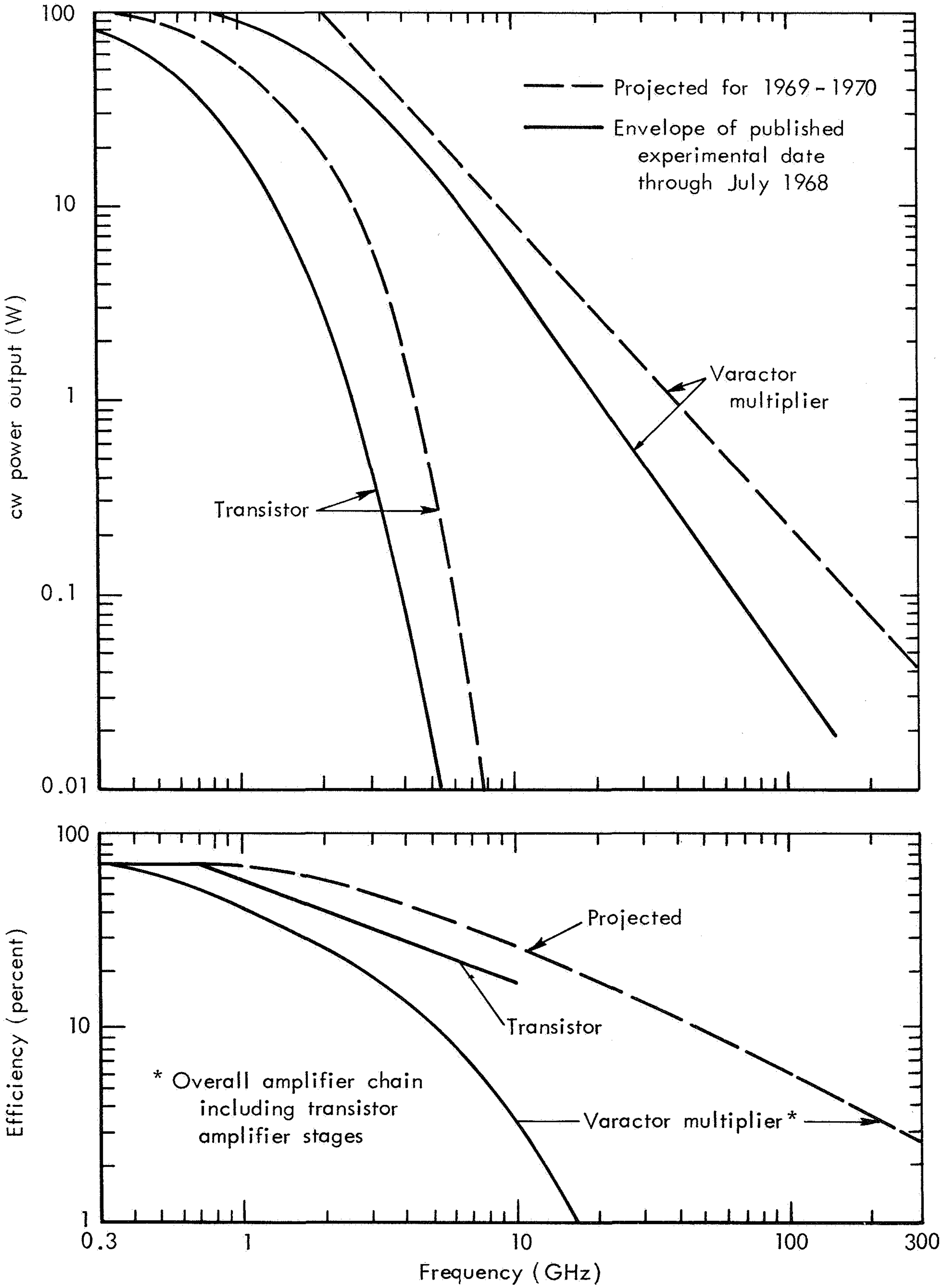

Fig.D-4—cw transistor and varactor-multiplier power output and efficiency power output and efficiency. This may not be the case for IMPATT amplifiers. The stable IMPATT amplifier configuration is somewhat less efficient but has a much broader bandwidth than the injection locked oscillator version. Injection locking provides gains of 5 to 30 dB as the locking range or frequency offset decreases from 100 to 5 MHz.

Recently, promising techniques have been developed for arraying devices on a transverse electromagnetic field transmission line, for paralleling devices with high isolation and low combining loss, for producing series stacks, and for producing series-parallel arrays on a single chip. Hybrid combining can provide 2 to 64 times the output of a single device with little loss in efficiency. Thus solid-state devices may displace some traveling-wave tubes (TWTs) within a few years.

The best long-lived amplifier for current space use where output power level or bandwidth requirements preclude the use of solid-state devices is the TWT. Tubes of this type in the range of 2 to 8 GHz and 1 to 35 W output are available, and the development of similar tubes at power levels in the same range for frequencies up through 100 GHz should be straightforward. For existing high-power or high-frequency tubes, end of life (wearout) is roughly a factor of 10 worse than for the typical 10-W S-band space TWT, and reliability is worse by factors of 20 to 50.

By hybrid or space combining, low-power space TWTs could make available kilowatts of microwave power for reliable space use. Overall TWT amplifier efficiencies, including the power supply, fall from a typical value of about 25 to 35 percent at 2 GHz to about one-half that or 12 to 17 percent at 100 GHz.