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APPLICATION OF THE TECHNICAL SPECTRUM EFFICIENCY FACTOR (TSEF) TO THE FIXED SERVICE IN THREE FREQUENCY BANDS

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ABSTRACT

The Technical Spectrum Efficiency Factor (TSEF) concept was developed in the Technical Subcommittee (TSC) of the Interdepartment Radio Advisory Committee (IRAC) and was applied, by NTIA, to government fixed telecommunications systems in the 7 - 8 GHz bands. Both computer automated and manual procedures were employed.

The Technical Spectrum Efficiency Factor (TSEF) evaluates the technical spectrum efficiency of a spectrum-using system by comparing the amount of the spectrum it uses to that used by the most spectrum-efficient system that could be procured to accomplish the same mission. The "most spectrum-efficient" system is called "the reference system." An extensive catalog of spectrum conservation techniques is also included.

KEY WORDS

Technical Spectrum Efficiency

Spectrum Conservation Techniques

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SECTION 1

INTRODUCTION

BACKGROUND

The National Telecommunications and Information Administration (NTIA) is responsible for managing the Federal Government's use of the radio frequency spectrum. Part of NTIA's responsibility is to "...establish policies concerning spectrum assignment, allocation and use, and provide the various departments and agencies with guidance to assure that their conduct of telecommunications activities is consistent with these policies" [NTIA, 1985]. In support of these requirements, NTIA has undertaken a number of spectrum resource assessments. The objectives of these studies are to assess spectrum utilization, identify existing and/or potential compatibility problems among the telecommunications systems that belong to various departments and agencies, provide recommendations for resolving any compatibility conflicts that may exist in the use of the radio frequency spectrum, recommend changes to increase spectrum efficiency and improve spectrum management procedures, and help establish a long-range plan for spectrum use.

To ensure efficient and effective use of the spectrum, Executive Order 12046 and Department of Commerce Order 10-10 direct NTIA to develop, in cooperation with the Federal Communications Commission (FCC), a long-range plan for spectrum use. As part of this planning effort, the following task was issued.

"Determine a method that can be used in the evaluation of spectrum efficiency and thus will provide technical support in evaluating efficient use of the radio spectrum. Specifically, this will require the following developmental tasks:

1. a quantitative definition of the terms relating to "efficient spectrum use"

2. a method of application and general formula for evaluating technical spectrum efficiency, which should initially be applied to the fixed service in selected bands from 947 MHz to 40 GHz

3. the development of a computer model for evaluation of the specific application examined in item 2

4. an examination of typical outputs from the computer model for the specific application examined in item 3."

These tasks were accomplished jointly by the Technical Subcommittee (TSC) of the Interdepartment Radio Advisory Committee (IRAC) and the Spectrum Engineering and Analysis Division (SEAD) of NTIA. This joint effort was done in a way that the methods and procedures developed and applied for evaluating government use of the radio spectrum would have the consensus of the Federal Government agencies. Specifically, the TSC of the IRAC accomplished tasks 1, 2, and 3, while SEAD accomplished task 4 that is the subject of this report.

TECHNICAL SPECTRUM EFFICIENCY FACTOR (TSEF) CONCEPT

The Technical Spectrum Efficiency Factor (TSEF) evaluates the technical spectrum efficiency of a spectrum-using system by comparing the amount of the spectrum it uses to the spectrum used by the most spectrum-efficient system that could be procured to accomplish the same mission. The "most spectrum-efficient" system is called "the reference system." It must be a practical state-of-the-art system that can be procured.

The TSEF is a dimensionless number that typically will be between 0 and 1 for most cases. It gives a relative measure of how close a system is to being spectrum-efficient for its particular mission. This comparison is valid only if both the actual and the reference system are adequately described. The critical step in making a TSEF calculation is specifying the reference system. Specification of a reference system that accomplishes the same mission as the evaluated system and uses minimum spectrum requires the combined experience of equipment/system designers and frequency managers.

The mission statement must include all aspects of the mission that affect spectrum resource requirements. One aspect is the communications requirement. This should include, for example, the type of information, the required information transfer rate, the error rate, for digital systems, of the signal-to-noise-plus-interference ratio, the reliability or time availability, and the required accessibility. All the details of the mission statement should shape the parameters of the reference system.

OBJECTIVE

The objective of this study was to apply and evaluate the TSEF methodology and computer model as tools for estimating efficient spectrum use.

APPROACH

To accomplish the objectives of this study, the following approach was taken.

1. The automated TSEF Model was applied to three bands in the 7125-8500 MHz frequency range using a limited number of system parameters obtained from a subfile of the Government Master File (GMF).

2. A more detailed TSEF calculation was made, using manual procedures for selected major networks in the 7125-8500 MHz GMF subfile. This was done by comparing each present system with the newer replacement system using the TSEF formula and the following methodology.

   a. The baseband characteristics for the existing system and the newer system were obtained.

   b. The TSEF was calculated manually using the new system as the reference system. (The automated model and the GMF were not used.)

3. The utility of the TSEF concept and the computer model were evaluated to determine if changes to the model and other possible approaches would improve the calculation of technical spectrum efficiency from a frequency management perspective.

SECTION 2

CONCLUSIONS AND RECOMMENDATIONS

GENERAL

The TSEF is a method to aid in the evaluation of spectrum efficiency. It evaluates the technical spectrum efficiency of a spectrum-using system by comparing the amount of the spectrum-space resource used by the system to the resource used by the most spectrum-efficient off-the-shelf system that could be procured to accomplish the same mission. The "most spectrum-efficient" system is called "the reference system." It is further constrained to be a practical state-of-the-art system that is procurable. The TSEF is a dimensionless number that will be between 0 and 1 for most cases. This means that TSEFs for quite different types of systems can be compared. The TSEF gives a relative measure of how close a system is to being spectrum-efficient for its particular mission. This comparison is valid only if both the actual and the reference systems are adequately described. This report demonstrates the TSEF concept. The 7125-8500 MHz frequency range was used in the demonstration of the TSEF because GMF records for this band are more complete than those of lesser used frequency bands.

CONCLUSIONS

1. The TSEF equation and methodology is a technically sound and useful concept that has both important applications and key limitations.

   a. The plots of TSEF for the three frequency bands within the 7125-8500 MHz range shown in Section 4 indicate that the TSEF model (even in the simple form employed here) provides TSEF values that indicate sensitivity to system characteristics. The TSEF concept and model thus have the potential for evaluating systems for technical spectrum efficiency when adequate system characteristics are provided.

b. Applying the TSEF methodology manually (Section 5), using more detailed evaluated system and reference system parameters, yields more realistic values of TSEF than the automated TSEF model.

c. Application of the TSEF concept to bands in the 7125-8500 MHz frequency range showed several ways to support NTIA's spectrum management functions.

   1. The TSEF concept can be used to assign a numerical TSEF value (figure-of-merit) to government telecommunications systems.

   2. The TSEF concept may be used in the system review process to compare technical efficiency aspects of alternate system designs where each has the same communications requirements.

d. The communication requirement of a system must include message related constraints relative to time sharing, space sharing, signal quality criteria, interference criteria, and security.

e. APPENDIX A describes other technical factors and concepts that promote spectrum efficiency. These are, for example, extending the TSEF to determine band efficiency and applying allocation and usage rules to avoid band congestion.

f. APPENDIX D describes the calculation of band efficiency as an extension of the TSEF concept.

2. The following are a number of key limitations that were identified in using an automated TSEF model with the GMF as a data base for both the actual and the reference system to evaluate fixed systems in the 7125-8500 MHz bands.

   a. A reference system must be established for each government system in the GMF that is evaluated by the TSEF model. This requires that a large number of reference systems be established prior to executing the TSEF model.

   b. The specification of a reference system that accomplishes the same mission as the evaluated system and uses minimum spectrum resources requires specific knowledge of the system's communications and performance requirements. Therefore, a realistic effort to establish reference systems for an automated application of the TSEF necessitates making assumptions about the systems. This will have a significant effect on the accuracy of TSEF calculation.

   c. The GMF does not contain the information required to evaluate the fixed systems for technical spectrum efficiency. For example, the GMF does not contain adequate technical information, data on communications requirements of systems, or the percentage of time the systems are operated.

RECOMMENDATIONS

The following are NTIA staff recommendations based on the technical findings contained in this report. Any action to implement these recommendations will be accomplished under separate correspondence by modification of established rules, regulations, and procedures.

It is recommended that the following be done.

1. NTIA should develop a methodology for specifying reference systems for the fixed service, which defines the essential features and communications' requirements and missions.

2. NTIA should, upon successful completion of 1, develop a procedure for evaluating alternate system designs to accomplish the same mission through the NTIA spectrum management process (e.g., systems review process).

3. NTIA should continue to explore other approaches for determining technical spectrum efficiency of government telecommunications systems.

4. NTIA should evaluate the overall effectiveness and the potential uses of the TSEF concept after completing the mobile TSEF.

SECTION 3

COMPUTER MODEL FOR EVALUATING THE TSEF

GENERAL

To compute the TSEF for the system being evaluated, the time-bandwidth-space (TBS) that a station denies to other potential users must be calculated. This calculation requires information on the system's characteristics and environment or transmission path, and a propagation model.

The specification of a reference system is a critical part of the TSEF calculation. APPENDIX A contains a detailed discussion of major telecommunications systems' design factors that affect spectrum utilization efficiency and that need to be addressed when specifying a reference system. The reference system is to perform the same mission as the evaluation system (i.e., satisfy the same mission-dependent communications requirement). The communications requirement should be stated in terms of the output required to perform the mission. For example, in the fixed service, it might be something like providing 24 voice circuits between points A and B with a signal-to-noise ratio (S/N) greater than 25 dB, with 99% reliability.

The TSEF is defined by Equation 3-1.

$$TSEF = (T_r \times B_r \times S_r)/(T_s \times B_s \times S_s) \qquad (3\text{-}1)$$

where:

T_r = the time the reference system denies to others

B_r = the bandwidth the reference system denies to others

S_r = the physical space the reference system denies to others

T_s = the time the evaluated system denies to others

B_s = the bandwidth the evaluated system denies to others

S_s = the physical space the evaluated system denies to others

The reference system is a practical, state-of-the-art system that accomplishes the same mission as the evaluated system. The most relevant aspect of a mission, with respect to technical spectrum efficiency, is the communications requirement. This is stated in terms of the necessary information-transfer rate, reliability, accessibility, etc. The reference system should be procurable and should use the minimum TBS product.

The reference system is designed to satisfy the requirement, while denying the minimum TBS product to other potential users. Once the reference system is specified, its TBS product is calculated using the same procedures and models used to calculate TBS for the evaluated system. The ratio of the two products is the TSEF.

DESCRIPTION OF THE TSEF MODEL

There is a tradeoff between the accuracy and completeness of the TSEF calculation for each station. For example, in the 7125-8500 MHz frequency range where the TSEF model was demonstrated, there are nearly 7000 assignments to the fixed service. It was decided that the statistical distribution of the TSEF for all stations in the band would give an indication of how efficiently they used the band. The TSEF computer model was developed to automatically evaluate TSEF for a large number of stations in the 7125-8500 MHz frequency range. The model's input parameters were directly obtained from the GMF.

The completeness and accuracy of the subsequent calculations were limited by the decision to use the GMF. In particular, the GMF does not contain the communications requirement for an assignment; however, it does contain the necessary bandwidth and an operation time code. It was assumed that these two parameters were an accurate reflection of the time and bandwidth needed for the station, and that the same time and bandwidth would be needed by the reference system. It was also assumed by the TSC that the time and bandwidth factors in Equation 3-1 cancel out for all the stations considered, and thus, the model calculated the TSEF for all stations using only the ratio of the physical space denied.

For the terrestrial fixed service, the area denied by a transmitter to another potential user is the area where the receiver of the potential user's system experiences unacceptable interference. Determination of this area depends on the characteristics of the victim receiver, the communications requirement, and the type and pointing angle of its antenna. Much of this information is unknown. A calculation of the TSEF should not be dependent on specific characteristics of a system that happens to be in the area. For the TSEF calculations in this report, the boundary of the denied area was defined by the power level received by a hypothetical victim receiver with an isotropic, lossless antenna at a specified height. The interfering power level and the victim antenna height are input to the model and can be varied to determine the sensitivity of the TSEF to their values.

The distance from the transmitting antenna to the boundary of the denied area depends on the transmitting antenna height, its directional gain, transmitter power, and transmission loss as a function of distance. The way these factors are used in the model are described in the following sections. More details can be found in APPENDIX C.

TERRAIN, PATH PROFILE, AND PROPAGATION LOSS MODEL

The GMF contains the height and the elevation of an antenna above the site for both the transmitter and receiver of a terrestrial fixed service station. With this information and a complete description of the terrain and structures in the vicinity of the link, the propagation loss can be calculated. At frequencies in the region of 8 GHz, the signal power level may be greatly attenuated by a terrain feature or the curve of the earth between the transmitter and the observation point. The signal power decreases rapidly with distance for distances beyond which there is first Fresnel Zone clearance of all terrain [ITT, 1972].

The TSEF computer model incorporates a smooth-earth propagation mode [Frazier, 1963]. Since microwave links are designed such that free-space transmission loss applies for distances short enough that the first Fresnel Zone is clear of obstacles, the free-space transmission loss was used for link distances having first Fresnel Zone clearance. To calculate loss at greater distances, a smooth-earth propagation model was used. This model produces a smooth, monotonic-loss curve, using a distance that is convenient for use in the TSEF model.

The actual transmission path is represented by a smooth earth path with antenna heights of sufficient height to clear the first Fresnal Zone. Thus, it was assumed that the antenna heights for the evaluated link were engineered so that the horizon (the highest terrain feature along the path) was just cleared by the first Fresnel Zone. This is a reasonable assumption because it is the basis upon which fixed microwave links are designed to avoid unacceptable link outages. Using the site elevations and structural antenna heights from the GMF, the model computed equivalent effective antenna heights that provide Fresnel Zone clearance over a smooth spherical earth for a path of the same length, while maintaining the real antenna height difference. These equivalent antenna heights were used in the smooth-earth propagation-loss calculation.

ANTENNA DIRECTIONAL GAIN

To find the distance to the denial boundary in a specific direction, the power radiated in that direction must be known. The GMF contains the antenna type and mainbeam gain for each assignment, but not the antenna pattern. For the fixed-service-TSEF model, the antenna pattern was approximated by a three-section, two-dimensional pattern, as illustrated in Figure 3-1.

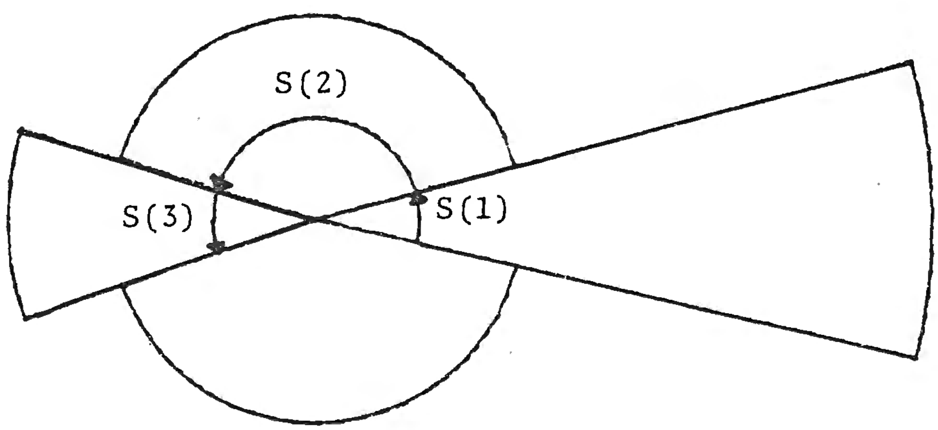


Figure 3-1. A three-sector approximation of a directional antenna pattern.

The sectors of the pattern represent the antenna mainbeam (1), sidelobes (2), and backlobe (3). Within each sector, the directional gain of the antenna was assumed to be constant, as a function of the angle.

The sector angle for the mainbeam (the beamwidth) is denoted by S(1) and is computed by Equation 3-2, using an approximation from a CCIR Report [CCIR, 1982]:

$$S(1) = 170.3 \times 10^{-G/20} \qquad \text{(degrees)} \qquad (3\text{-}2)$$

where:

G = the gain taken from the GMF.

The sector angles and the gains for the sidelobes and backlobe were modeled using manufacturer's antenna-pattern-envelope data for antennas designed for the 7125-8500 MHz band. Details of the modeling are given in APPENDIX C. The result is that the backlobe sector angle (S(3)) is given by Equation 3-3.

$$S(3) = 20 \qquad \text{(degrees)} \qquad (3\text{-}3)$$

The backlobe gain is -8 dB. The sidelobe sector angle (S(2)) is calculated by Equation 3-4.

$$S(2) = 360 - S(1) - S(3) \qquad \text{(degrees)} \qquad (3\text{-}4)$$

The sidelobe gain (G(3)) is given by Equation 3-5.

$$G(2) = 0.55\ G - 12 \qquad \text{(dBi)} \qquad (3\text{-}5)$$

TRANSMISSION LOSS

The transmission loss that must be sustained to reduce the incident field to a specified interference power (P_i) is determined by Equation 3-6.

$$L(n) = P_t + G(n) - P_i \qquad \text{(dB)} \qquad (3\text{-}6)$$

where n = 1. 2. 3... L(n), in dB, is the loss to the boundary in sector n, and P_t is the transmitter power, in dBW and G(n) and P_i are previously defined. The distance (d(n)) to the boundary of sector n is found by inverting L(n), using the propagation loss model. Then, the area the transmitter denies in sector n (A (n)) is determined using Equation 3-7.

$$A(n) = d(n)^2 \pi S(n)/360 \qquad (3\text{-}7)$$

The total area denied is the sum of the sector areas, as in Equation 3-8.

$$\text{Area} = A(1) + 2\,A(2) + A(3) \qquad (3\text{-}8)$$

PASSIVE REFLECTORS AND FREQUENCY DIVERSITY OR REDUNDANCY

Some links in the fixed service contain passive reflectors that are used to redirect the power in a desired direction when a direct path is not available because of some obstacle. For example, if a transmitter in a canyon transmits towards a reflector on top of the wall of the canyon, the reflector directs the beam towards the desired receiver. The scattering properties of a passive reflector cannot be computed unless the dimensions and curvature (if any) of the reflector are known, and this information is not in the GMF. Therefore, the area denied on links containing passive reflectors can only be modeled approximately. The approach used was to treat the reflector as a 1.2 meter (four-foot) dish antenna and compute the area denied by the link from the transmitter to the reflector and the area denied by the link from the reflector to the receiver separately. Then, add these two areas to get the total denied area. A more realistic way to model passive reflectors is needed.

Some links in the fixed service use frequency diversity or frequency redundancy to achieve the desired reliability. The same reliability sometimes can be achieved with space diversity that may deny a smaller frequency-area product [Lenkurt, 1975]. The use of frequency diversity is noted in the "Supplementary Details" field in the GMF, which does not have a fixed format and thus makes the information difficult to retrieve. In addition, the decision as to whether frequency diversity could reasonably be replaced by space diversity on a particular link depends on information not available in the GMF and on judgments that are difficult to automate. Thus, it was not possible to include frequency diversity in the TSEF calculation.

THE REFERENCE SYSTEM

As discussed earlier, the reference system is to be a spectrum-efficient system that performs the same communications mission as the evaluation system. Since the communications mission is not included in the GMF, reference systems were specified for each fixed system in the GMF by assuming that systems had not been over-designed in certain respects. In particular, the bandwidth and time for the fixed systems in the GMF were assumed to be the bandwidth and time necessary for both the GMF system and the respective reference system. The reference system used the same antenna locations and heights as the evaluated system, and the power radiated in the mainbeam was assumed to be the minimum power needed to provide the required system reliability.

The TSC decided that the reference antenna should be a 3.2 meter (10 ft) high-performance dish having a modeled gain of 45 dBi. The transmitter effective radiated power (ERP) for the reference station was computed by Equation 3-9.

$$\text{Reference power} = P_t + G(1) - 45 \quad \text{(dB)} \qquad (3\text{-}9)$$

THE "UNAVAILABLE INFORMATION" FACTOR

Several items of information necessary for a complete and accurate calculation of the TSEF were not in the GMF. These include (at least) the communications requirement of the system, the emission spectrum of the transmitter, the selectivity of the receiver, and the dimensions of any passive reflectors used. As a result, the number calculated by the model differed from the true TSEF by an unknown factor (U). The output of the program is thus denoted as U x TSEF.

RETRIEVAL OF INPUT VALUES FROM GMF

A computer program to calculate TSEF automatically obtains input data from a computer file of fixed stations (systems) in the GMF.

The first step is to retrieve a working file of fixed stations from the GMF that contains only those within a specified frequency range, located in a specified geographic region, and of a specified station class. This selection excludes any entries that do not include the minimum necessary data to compute the desired calculations. For each station, the minimum necessary information in the station file includes:

assignment number
station class
frequency
transmitter power
latitude
longitude
site elevation
antenna height
antenna gain for the transmitter
receiver
reflector.

COMPUTATION OF AREA DENIED BY THE GMF STATION

The boundary of the denied area will be at a distance where the power incident on a lossless isotropic receiving antenna at a height (H_i), in meters, is P_i, in dBm. H_i and P_i are constants that are manually input. The ratio of the area denied by the reference station to the area denied by the evaluated station is computed for each station in the station file and stored.

COMPUTATION OF THE AREA DENIED BY THE REFERENCE STATION

The equivalent isotropically radiated power (EIRP) of the reference station is determined using Equation 3-9. The reference station antenna sector angles and directional gains are determined by Equations 3-2 and 3-5, using G = 45 dB. The equivalent antenna heights are obtained using the procedure described in APPENDIX C.

The distance represented by the sector boundary loss is determined using a propagation model and the equivalent antenna heights. The sector area is computed using Equation 3-7 and the total area for this link is computed using Equation 3-8.

If the link includes a passive reflector, the steps above apply to the link from the transmitter to the reflector. Then, the steps for the link from the reflector to the receiver are repeated, except that the "transmitter power" of the reflector is the power incident on the reflector. The total area denied is then the sum of the areas of the two legs of the link.

THE AREA DENIED BY THE EVALUATED STATION

The area denied by the evaluated station is computed using the same procedure as for the reference system, except the transmitter power and antenna gain are obtained from the GMF, and the passive reflector (if present) is replaced by a four-foot dish instead of the reference antenna.

SECTION 4

APPLICATION OF THE TSEF CONCEPT

GENERAL

In addition to discussing the application of the TSEF computer model to the three frequency bands in the 7125-8500 MHz range, this section discusses the manual application of the TSEF concept to specific systems in this same frequency range. The manual application considers two selected major systems that are currently being used in the 7125-8500 MHz bands. This manual method uses assumed reference systems that include the emission bandwidth that is not considered in the automated computer model.

APPLICATION OF TSEF COMPUTER MODEL TO THE 7125-7750 MHz BAND SYSTEMS (BAND I)

The TSEF computer model described in Section 3 was used to generate the TSEF values plotted on the Figure 4-1 histogram of TSEF values for the GMF assignments in the 7125-7750 MHz band. The abscissa is the numerical value of the TSEF as computed by the program. The ordinate gives the number of GMF systems in which the TSEF falls within each range of numerical values. In the expression U*TSEF, the factor U provides a factor that can be used to compensate for "unavailable" information. Unavailable information could be, for example, identifying the uncertainty associated with the ratio of the reference system bandwidth to the bandwidth of the system being evaluated. The U factor was not assigned a value for the execution of the computer model, but it is taken into consideration in the manual calculation later in this section.

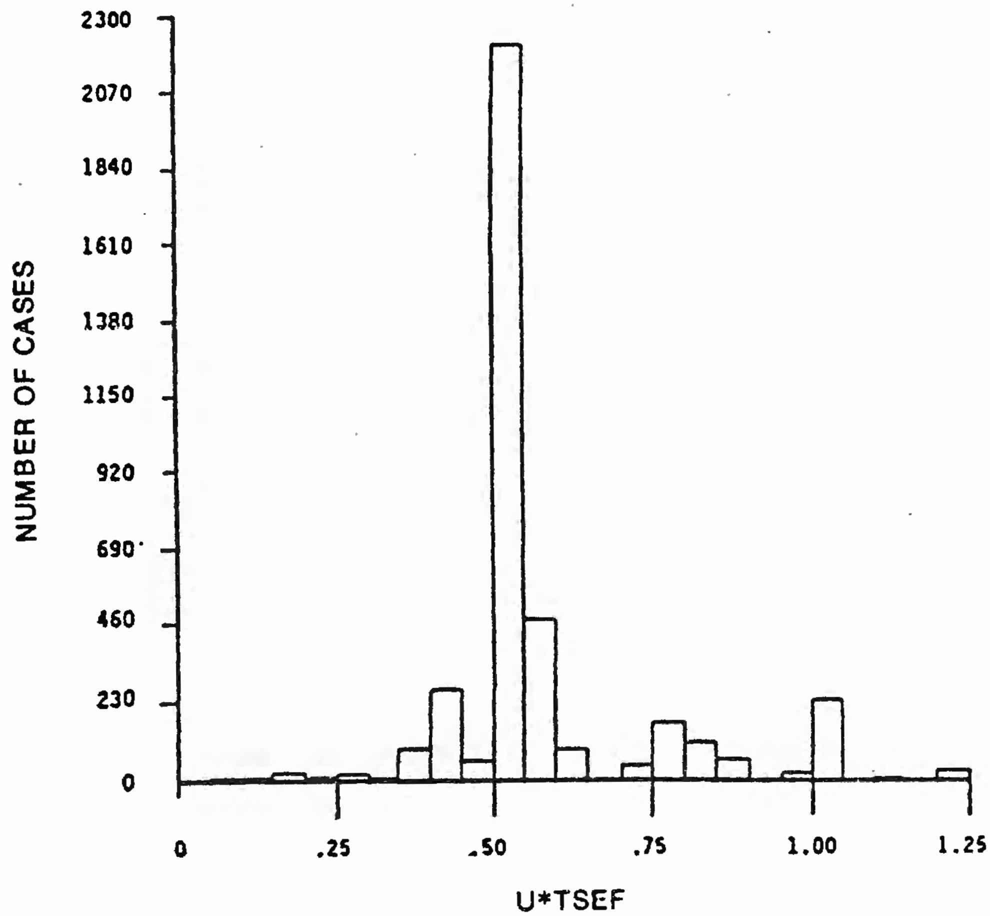


Figure 4-1. Histogram of TSEF values in the 7125-8500 MHz band from the TSEF Computer Model

The TSEF computer model calculated the TSEF for 3963 GMF assignments in the 7125-7750 MHz band. As Figure 4-1 shows, the TSEF histogram has a peak between 0.50 and 0.55 that includes over half of the systems. Due to the make up of the TSEF computer model and the GMF data used in the TSEF calculations, the histogram actually reflects the distribution of antenna gains for the systems in the band. As a result, the peak of the histogram indicates the relatively large number of transmitter antennas in the band that have the same gain of 39 dBi. Ninety, fifty, and ten percent of the TSEF values are greater than or equal to 0.412, 0.507, and 0.807 respectively. The mean value of the TSEF for all the fixed systems in the 7125-7750 MHz band is 0.57

Figure 4-2 is a histogram showing the TSEF values for the GMF assignments in the 7750-8025 MHz band. The format of this histogram is the same as that described in the preceding subsection that discusses Figure 4-1.

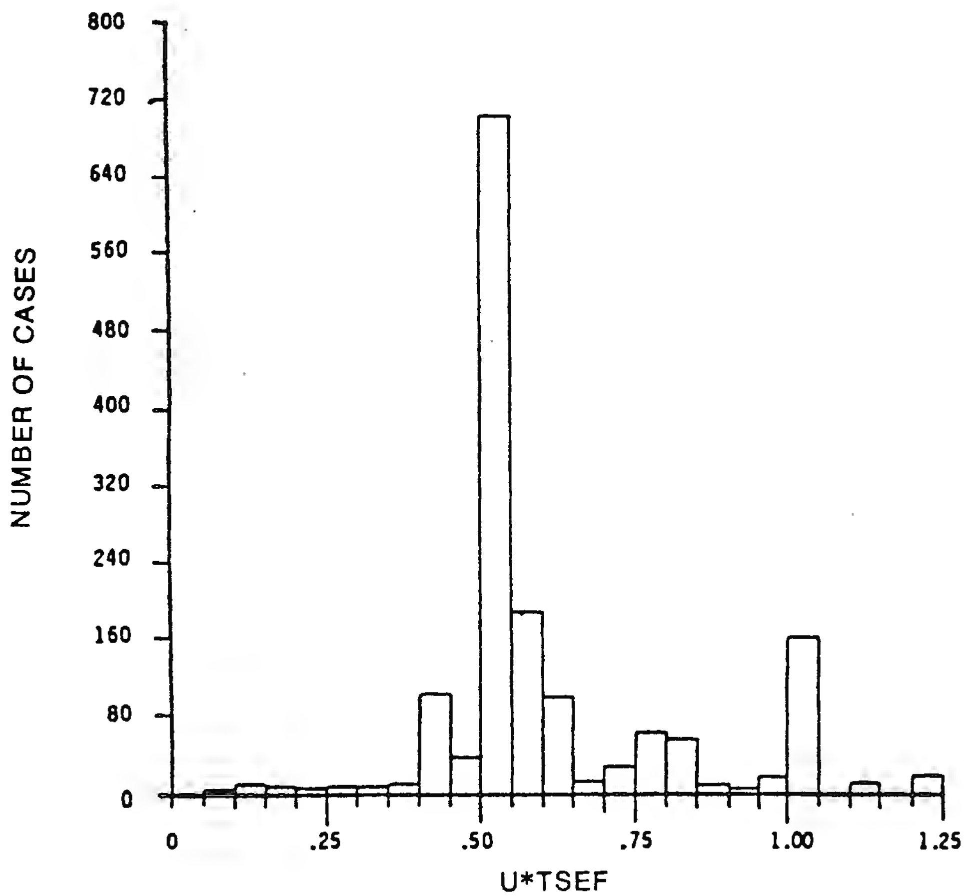


Figure 4-2. Histogram of TSEF values in the 7750-8025 MHz band from the TSEF computer model.

The model computed the TSEF for 1074 GMF assignments in the 7750-8025 MHz band, and, as Figure 4-2 shows, the TSEF was between 0.50 and 0.55 in nearly half of the cases. Again, as in the 7125-8025 MHz band, this is an indication of the relatively large number of transmitter antennas that have a gain of 39 dBi. Ninety, fifty, and ten percent of the TSEF values are greater than, or equal to, 0.412, 0.507, and 1.000 respectively. The mean value of the TSEF for all the fixed systems in the 7750-8025 MHz band is 0.61.

Figure 4-3 is a histogram produced by the computer model for this band. The format of the histogram is the same as the format for the histograms in Figures 4-1 and 4-2.

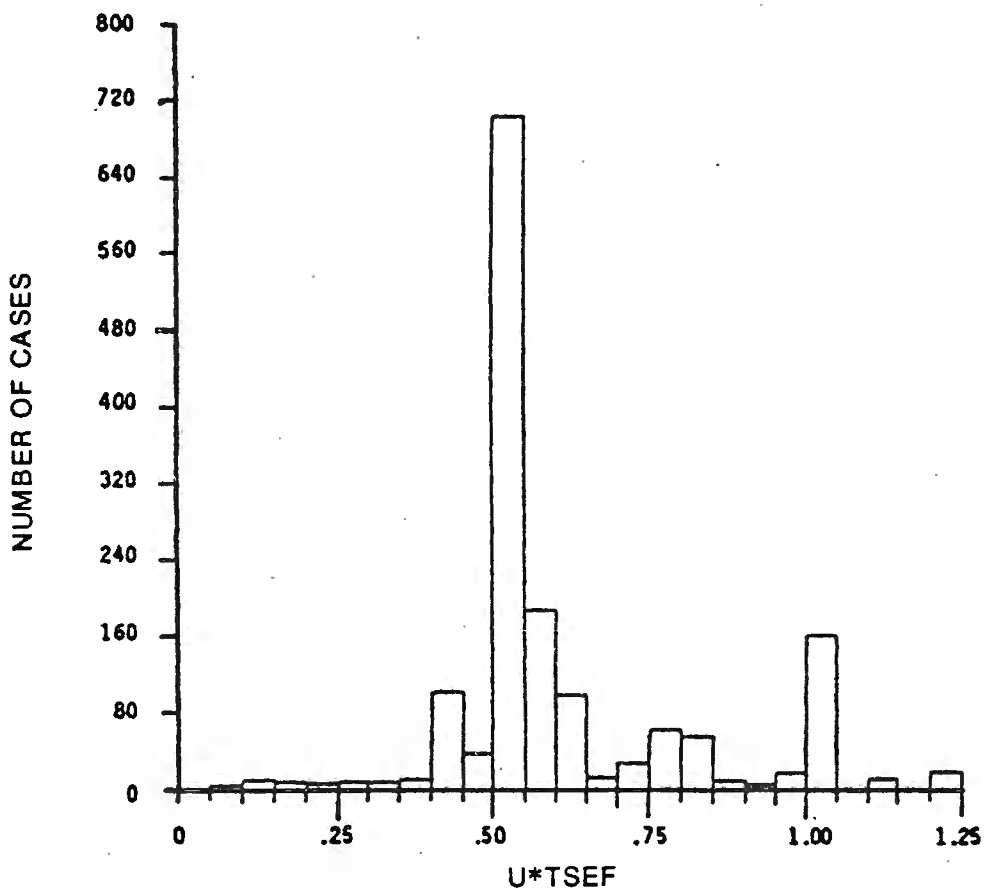


Figure 4-3. Histogram of TSEF values in the 8025-8500 MHz band.

The computer model calculated the TSEF for 1537 GMF assignments in this band. As Figure 4-3 shows, the largest concentration of TSEF values is between 0.50 and 0.55, as is true for the other two bands that were discussed earlier in this section. Again, this is an indication of a relatively large number of transmitter antennas that have a gain of 39 dBi. The mean value of the TSEF for all the fixed systems in the 8025-8500 MHz band in Figure 4-3 is 0.61. Ninety, fifty, and ten percent of the TSEF values are greater than or equal to 0.453, 0.507, and 1.000, respectively.

SUMMARY OF APPLICATION OF TSEF COMPUTER MODEL TO GMF SYSTEMS IN THE 7125-8500 MHz BANDS

The three TSEF histograms, Figures 4-1, 4-2, 4-3, generated by the TSEF computer model and GMF data for the fixed systems in the 7125-8500 MHz bands, reflect the dominance of the transmitter antenna gain value in calculating the TSEF in this particular application. This antenna gain dominance was expected for three reasons. First, the spectrum denied, which is inferred by the ratio of bandwidths (B_r/B_s), as in Equation 3-1, was assigned the value of one (unity) since the minimum required bandwidth of the actual system was not available through the GMF. Second, the ratio of the on-air time denied, which is inferred by the ratio T_r/T_s, as in Equation 3-1, was assigned the value of one (unity) since information concerning usage was not available for the actual system through the GMF. Third, the physical space denied, which is given by the ratio S_r/S_s, as in Equation 3-1, is calculated in the TSEF computer model. This calculation is based on the actual systems from the GMF, using the data base parameters of transmitter and receiver locations, site elevations, transmitter antenna height, antenna gain, and transmitter power. For the reference systems, the EIRP is assumed equal to the EIRP of the actual system's from the GMF. For this application of the TSEF, the only parameter selected from the GMF for the actual system that significantly contrasts with the reference system's similar parameters is the antenna gain value that dominates the calculation of physical space denied (S_r/S_s) and thus, dominates the TSEF calculation and overall distribution of TSEF values.

However, one observation may prove useful. For these histograms, the peaks indicate a dominance of the antenna gain parameter in the TSEF model, which also indicates the dominance of a particular system in the band. This dominance would have shown up (perhaps at a different value) if the emission bandwidth also had been included. The peaks serve to identify systems that must be efficient if the band is to be used efficiently. Although this was obvious for this band before the TSEF model was employed, it may not be in other bands. As such, the TSEF computer application could possibly have a useful future role in identifying those systems in all the bands where technical efficiency has a large impact on the bands.

APPLICATION OF TSEF FORMULA USING THE MANUAL METHOD

This subsection discusses the manual application of the TSEF concept to specified systems in the 7125-8500 MHz range. As mentioned earlier, the analysis method compares the existing (actual fixed system) system to the newer (designated reference system) system for each agency. The fixed microwave systems chosen for the analysis are the FAA RML-4 system and the DOE Bonneville System. The manual method considers factors, mainly bandwidth, that are not considered in the automated computer model. Unlike the TSEF computer model that calculates the reference system power, the TSEF manual calculations use reference system power as an input. Reference system antenna gain, a fixed value in the computer model, is an input parameter for the manual calculations. Thus, the manual calculation of the TSEF would conceptually provide the more detailed and more realistic evaluation of technical spectrum efficiency.

SYSTEMS USED IN MANUAL APPLICATION

This section briefly describes the fixed systems evaluated using the TSEF calculation and the assumed reference systems for the TSEF manual demonstration. The systems evaluated are real systems in the 7125-8500 MHz frequency range. It is important to note that this is a feasibility study yielding results that are useful only to evaluate the TSEF applications and not necessarily the actual systems. These systems have been selected for a manual calculation of the TSEF because they dominate this frequency band. This manual analysis of the TSEF procedure actually compares the older, or existing system, and the newer, or upgraded system, being introduced by each agency, <u>where the newer system is defined here as the reference system.</u> The reference systems used in the manual application do not strictly fulfill all the requirements of a reference system as defined in APPENDIX C. They do, however, include bandwidth, which is not considered in the automated TSEF model. A discussion of the reference system parameters for the FAA and DOE systems follows the description of the major systems.

DESCRIPTION OF DOE BONNEVILLE FIXED SYSTEM (OLDER SYSTEM)

The DOE Bonneville microwave system consists of predominantly 600-channel FDM/FM commercial off-the-shelf equipment. It requires high reliability due to the critical function it serves. This is accomplished by the design of fade margins of 30 to 40 dB, using diversity schemes with typical diversity spacings of five percent. Technical parameters for the existing DOE Bonneville System are given in TABLE 4-1.

DESCRIPTION OF DOE BONNEVILLE SYSTEM (NEWER SYSTEM)

As mentioned previously in this section, the manual analysis procedure compares older and newer systems used by the agency where the newer system is assumed to be the reference system. An important difference in the Bonneville older and newer systems is the frequency of the continuity pilot. This is the key factor in determining the RF bandwidth of the system. This parameter and the other parameters chosen for the reference system are given in TABLE 4-1. The antenna gain and transmitter power output were obtained from a computer-generated distribution of the values of these parameters for Bonneville frequency assignments in the GMF.

TABLE 4-1

PARAMETERS FOR MANUAL APPLICATION OF TSEF TO THE DOE BONNEVILLE SYSTEMS

	Actual System Existing Older System	Reference System Newer Replacement System
Modulation Type	FDM/FM	FDM/FM
Number of Channels	600	600
Baseband Bandwidth	2.54 MHz	2.54 MHz
Deviation per Channel	200 kHz	200 kHz
Continuity Pilot Location	8.5 MHz	3.2 MHz

FAA RML-6 SYSTEM DESCRIPTION

The FAA RML-6 Radar Microwave Link System (RML) accounts for some of the fixed service frequency assignments in the 7125-8500 MHz bands. These RML systems transmit radar data between remote radars and the associated control centers. Microwave links serving this purpose are generally long and consist of more than one hop. Although the RML-6 radar microwave link systems were used in this TSEF demonstration, other RML systems in the GMF, which are very similar to the RML-6, are also considered to be the RML-6 system for purposes of this report.

The RML-6 transmits information from the radar site to the indicator site through a single RF channel. A single channel also serves to transmit information from the indicator site to the radar site. The RML-6 has two spare channels available, one for transmission of information in each direction. Parameters for the RML-6 system are given in TABLE 4-2.

TABLE 4-2

SYSTEM PARAMETERS FOR THE FAA RML-6 MICROWAVE LINK	
Frequency Range	7125-8400 MHz
Modulation	FM
Power Output	8 dBW
RF Bandwidth	45 MHz
Deviation Ratio	0.25
Peak Frequency Deviation	± 4 MHz
Baseband Bandwidth	16 MHz
Receiver Noise Figure	10 dB
Antenna Type	Parabolic Dish Direct
Antenna Polarization	Horizontal or Vertical
Antenna Gain	40 dBi
Antenna Diameter	1.9 meters (6 feet)

FAA REFERENCE SYSTEM DESCRIPTION

TABLE 4-3 gives parameters for the Radio Communications Link (RCL) reference system, which are used to transmit information between radar sites and the control centers. The table lists two values of transmitter power output, 0 dBW and 7 dBW. The smaller value is to be the standard value, while the larger value is to be used where more propagation loss conditions exist. The standard antenna diameter is 1.8 meters (6 feet), but there are options to use an antenna that has a diameter of 2.4 meters (8 feet) or 3.0 meters (10 feet).

TABLE 4-3

SYSTEM PARAMETERS FOR THE RCL MICROWAVE LINK	
Frequency Range	7125-8400 MHz
Modulation	FDM/FM
Power Output per RF Channel	0 dBW or 7 dBW
Bandwidth per RF Channel	20 MHz
Number of RF Channels	2 (each way)
Receiver Noise Figure	6 dB or better
Receiver IF	70 MHz
Antenna Type	Parabolic
Antenna Polarization	Vertical or Horizontal
Antenna Diameter	1.8 meter (6 feet)
Antenna Gain (Max)	41 dBi @ 55% efficiency

MANUAL APPLICATION OF THE TSEF TO THE DOE BONNEVILLE SYSTEM

The following provides a comparison of older and newer radio-relay systems in the DOE Bonneville System network. TABLE 4-1 lists the relevant parameters of the existing system or evaluated system and TABLE 4-2 lists the relevant parameters of the newer system or designated reference system. Since it is assumed that the two transmitters are being compared over the same physical path, and since the modulation types are the same, it is also assumed that the transmitted power and antenna gains are the same for the two systems. Thus, the area denied is the same for the two systems and does not enter into the calculation of the TSEF for these systems. Under these conditions, the TSEF is equal to the ratio of the necessary bandwidths as seen in Equation 4-1.

$$TSEF = B_r/B_e \quad (4\text{-}1)$$

where B_r = the necessary bandwidth of the reference system

B_e = the necessary bandwidth of the evaluated system

The necessary bandwidth of each system is found using the procedures given in the NTIA Manual [NTIA, 1985] and the ITU Radio Regulations [ITU, 1982]. From these sources, it was found that the necessary bandwidths of the actual (newer) and the reference (older) system are 11.64 and 17.0 MHz, respectively. Thus, the TSEF, for the existing system is:

$$\begin{aligned} \text{TSEF} &= 11.64/17.0 \\ &= 0.68 \end{aligned}$$

It is seen in Figures 4-4 and 4-5[a] that the difference in necessary bandwidths between the older and newer system is due entirely to a difference in location of the continuity pilot frequencies of the two systems. Figure 4-4 shows the measured spectrum for the Bonneville reference system that has a pilot frequency of 3.2 MHz from the carrier, while Figure 4-5 shows the spectrum for the actual older Bonneville system that has a pilot frequency of 8.5 MHz from the carrier. The extra spectral spreading from the pilot being at 8.5 MHz separation is evident in a comparison of the two figures.

MANUAL APPLICATION OF THE TSEF TO THE FAA RML-6 SYSTEM

The following discussion provides a comparison of older and newer radio-relay systems in the FAA fixed microwave network. In the calculation, the FAA RML-6 system was the older system evaluated and the newer FAA RCL system was the reference. The existing radar microwave link systems were the older systems in this analysis.

a. The NTIA Radio Spectrum Measurement System (RSMS) was used to measure the spectra of each of the types of systems shown in the figures.

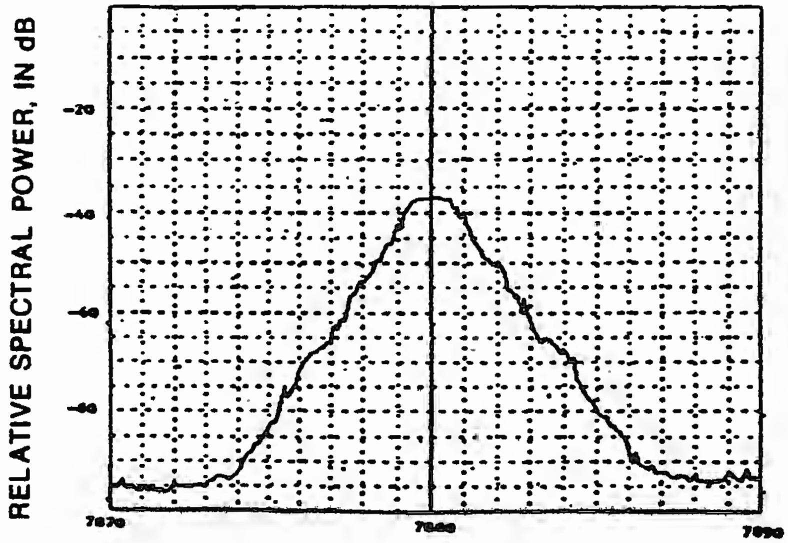


Figure 4-4. Emission spectrum of the DOE Bonneville reference system (new) (f_o = 7880 MHz, radiated).

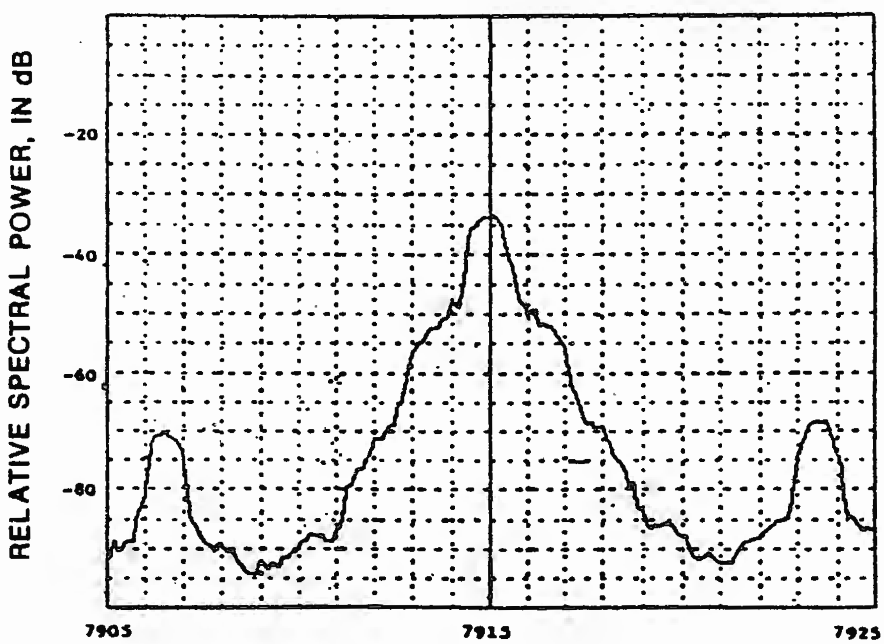


Figure 4-5. Emission spectrum of DOE Bonneville actual (existing) system (f_o = 7915 MHz, radiated).

The TSEF of the RML-6 was evaluated using the RCL reference system, since it is likely that the RCL system will replace the RML system. Parameters for these microwave links are given in TABLES 4-2 and 4-3 for the RML-6 and RCL systems, respectively. For convenience, TABLE 4-6 presents parameters used directly in the determination of the TSEF for the RML-6. As indicated by TABLE 4-6, both the reference system and the evaluated system use frequency diversity so, this factor cancels out. Since both systems use a 1.8 meter antenna, the antenna patterns for the two systems are essentially the same and thus, cancel out. Based on the above and assuming that the RML-6 and RCL systems have equal time availability for purposes of computing an upper bound on the TSEF of the RML-6 system, the area denied by the RML-6 and RCL transmitter was assumed to be linearly proportional to the RF power output.

TABLE 4-6

SYSTEM PARAMETERS FOR THE FAA RML-6 SYSTEM
AND THE RCL REFERENCE FAA RCL SYSTEM

PARAMETER	RML-6	RCL
Antenna Diameter	1.8 meter (6 feet)	1.8 meter (6 feet)
RF Power Output	8 dBW	0 dBW or 7 dBW
RF Bandwidth	45 MHz	20 MHz
Frequency Diversity	Yes	Yes

The TSEF for the RML-6 system may now be calculated based on the parameters in Table 4-6, the free-space propagation model, the transmitter power and antenna gain, the effective height of the microwave tower, and the interference boundary conditions, as defined in Section 3. Based on these conditions and assumptions, the usual ratio of denied areas becomes a ratio of transmitter powers and necessary bandwidths and the formula for the TSEF reduces to Equation 4-5.

$$\text{TSEF} = \frac{P_r \ B_r}{P_s \ B_s} \tag{4-5}$$

where P_r = the RF power output of the reference system

P_s = the RF power output of the evaluated system

B_r = the necessary bandwidth of the reference system

B_s = the necessary bandwidth of the evaluated system.

The TSEF is computed for the RML-6 system using the parameters from TABLE 4-6 and Equation 4-6 and first assuming OdBw (1.0 watt) as the transmitter power for the reference system.

$$\text{TSEF} = \frac{P_r \; B_r}{P_s \; B_s}$$

$$= \frac{(1.0 \text{ Watt}) \; (20 \text{ MHz})}{(6.3 \text{ Watt}) \; (45 \text{ MHz})} \qquad (4\text{-}6)$$

$$= 0.07$$

Then, the TSEF is calculated assuming the reference system transmitter power of 7 dBW (5.0 watts).

$$\text{TSEF} = \frac{(5.0 \text{ watts})(20 \text{ MHz})}{(6.3 \text{ watts})(45 \text{ MHz})} = 0.35$$

It is evident that, under these assumptions, the value of TSEF for the RML-6 is highly dependent on the transmitter power employed by the RCL reference system.

SUMMARY OF MANUAL METHOD

The manual application of the TSEF focused on specific government telcommunications systems, rather than a conglomerate of systems, as in the computer application of the TSEF. This allowed the use of more details of the specific systems in the TSEF calculations. In both the DOE and FAA systems, the actual transmitter power and emission bandwidths were used in the calculation. Since for both systems, the operating time periods used and the antenna characteristics were the same, these factors did not need to be included in the TSEF calculation. In the manual calculations, the reference system was assumed to be the new replacement system for the actual FAA and DOE systems. In this way, the TSEF calculation indicates whether the new system is more technically spectrum efficient then the older systems being replaced. Based on the limited number of parameters used in the TSEF calculations, both the DOE and the FAA next generation (newer) systems are more technically spectrum efficient.

SECTION 5

EVALUATION OF THE TSEF CONCEPT

GENERAL

This section discusses the uses of the TSEF concept (defined in Section 1) and the scope of its utility. It also discusses the TSEF computer model that calculates the TSEF for the fixed service, along with possible modifications to the model. Recognizing that there is no single acknowledged method for measuring spectrum efficiency, this section discusses possible modifications and additions to the concept of the TSEF, as well as alternative approaches to identifying technical spectrum efficiency.

TECHNICAL SPECTRAL EFFICIENCY, BAND CONGESTION, AND BAND EFFICIENCY

Before discussing possible modifications, additions, or alternatives to the concept of the TSEF, it is necessary to put the TSEF in perspective. This is accomplished by identifying three major spectrum-related areas of investigation that have been discussed within the spectrum management community. Technical spectrum efficiency, frequency band congestion, and frequency band efficiency are three such topics. These are different spectrum use measures that could require separate technical investigation and development. The TSEF is one possible indicator of the technical spectrum efficiency of a system and could be used as one of a set of indicators to define, more completely, the technical spectrum efficiency of a system and a frequency band. One definition of band efficiency, as an extension of the technical spectrum efficiency concept, is given in APPENDIX D.

UTILITY OF THE TSEF

The TSEF concept can be used to evaluate the technical spectrum efficiency of current or planned systems. The evaluation of current systems might show that a significant increase in technical spectrum efficiency is possible with new technology. When new systems are being planned, the TSEF can be used as a figure-of-merit for rating the technical spectrum efficiency of various candidate systems or components. It can be used in parametric tradeoff studies, finding the combination of controllable parameters that lead to the design that uses the fewest spectrum resources. The spectrum efficiency of a particular design, shown by a decrease in the TSEF value, could then be compared with other factors, such as the costs of the designs, to make final procurement decisions.

The TSEF of proposed system upgrades could be compared with the TSEF of current systems to determine the degree of improvement in technical spectrum efficiency. Because factors such as required bandwidth and area denied often tradeoff with each other, counter-intuitive results may be discovered (e.g., a saving in bandwidth caused by a change of modulation may actually increase the area denied by increasing the susceptibility to interference).

It is possible that the TSEF can be extended to compare the time-bandwidth-area (TBA) product of telecommunications equipment. These TBA products can be used by frequency managers and regulators to indicate which services offer the greatest opportunity for increased spectrum efficiency through tightened equipment standards or changes in system configurations.

EVALUATION OF THE TSEF COMPUTER MODEL

Section 3 of this report describes a TSEF computer model for evaluating the TSEF for government stations in the fixed services. The computer model includes calculation of transmitter denial area, including the effects of the transmitter antenna height and gain pattern. Effects of the transmitter emission bandwidth (spectrum denial) or of time denial are not included in the computer model since acceptable values for these parameters were not consistently available from the computer data base input parameter file.

Since the GMF contains the antenna gain, with no information about the antenna pattern, the TSEF model used an idealized sector antenna pattern that was based on the antenna gain alone. Since the dimensions of the reflector are not included in the GMF, when periscope antennas or passive reflectors are used, modeling of such reflectors in the TSEF model is limited.

Although frequency redundancy or diversity affects the bandwidth factor in the TSEF, the model for the fixed service does not consider frequency diversity due to GMF limitations.

Since information about the emission spectra of the transmitters or the selectivity characteristics of the receivers is not available in the GMF, only area denied was computed. It is not known how much this limitation affected the accuracy of the TSEF calculation.

Because the communications requirement of the station is not included in the GMF, the reference system could not be accurately specified in the computer model. (The reference system is supposed to be the system that uses the smallest TBA product to accomplish the same mission.) For the model described in Section 3, the mission is inferred from certain characteristics of the station class listed in the GMF.

In its application to the fixed services, the TSEF has important strengths and weaknesses when used as an indication of the technical spectrum efficiency of a system. The TSEF definition enables the technical spectrum efficiency of a system to be expressed as a single number. Use of a single number facilitates a quantitative comparison of different systems. However, sole use of a single number provides no information on the individual parameters that determined the number. Potential problems arise when consideration is given to including effects between the fixed service and other services in the calculation of the TSEF.

ALTERNATE APPROACHES TO EVALUATING SPECTRUM EFFICIENCY

Several alternate approaches for evaluating spectrum efficiency may be pursued either individually or in parallel. One approach is to improve the automated model that uses the GMF as the database. The GMF however, is extremely limited as a technical database. In fact, it is not designed to be a technical database. Of the parameters that are being used from the GMF for the automated TSEF model, at least two are difficult to use in the computer

TSEF model and the manual TSEF calculations. The GMF gives the authorized bandwidth whereas, the TSEF requires the necessary bandwidth. Although the GMF is probably the best available database for this task, the bandwidth provided is inappropriate for the TSEF calculation. Another difficulty concerns the propagation model. The propagation model requires, as an input, the height above ground of the microwave tower. However, the antenna heights in the GMF are not always the tower height above ground. Furthermore, the GMF antenna heights are not clearly defined, which makes this field difficult to interpret. For example, some entries appear to include the height of the building on which the tower is located, while others do not. In addition, the GMF does not include the communications requirement of the mission, which is essential for computing the TSEF

One approach to preparing inputs for the TSEF model is to expand the GMF to include additional system parameters that would be supplied by the agencies. It is anticipated that such an approach would entail a very considerable amount of agency and NTIA resources for the computer model and the expanded database developement.

The approach chosen for the manual analysis of systems is the comparison of the older and newer systems. This analysis is presented in Section 4. In the analysis of the FAA RML-6, it is necessary to evaluate several assignments, or prospective assignments, as a group. An example of such a grouping is provided in the analysis of a RML-6 backbone system. Even the limited analysis of the RML-6 in Section 4 required a considerable amount of background investigation.

Manual approaches, such as those used in Section 4, provided the possibility of doing a more thorough evaluation of the TSEF for a smaller number of systems in a given amount of time. These types of studies do not necessarily involve large databases, but still require a sizable expenditure of manpower. The approach used in Section 4 may provide insight into techniques that could be used to produce spectrally efficient systems; however, applying such methods to existing systems may not directly contribute to spectrum efficiency, but may provide the bans for future planning of systems.

Manual approaches to the calculation of the TSEF may also be used to evaluate the spectrum efficiency of new systems being developed in a band. This was referred to earlier as the spectrum conservation approach. This approach has two major advantages. The first is that the evaluation could take place in time to influence the development of the system. The second is that a database would not be necessary, since the evaluation would be on a case by case basis by the appropriate analysis. A variation of this approach would evaluate the technical spectrum efficiency without the use of a reference system by comparing the TBA of competing system designs. A combination of these two approaches is also possible. It may be that a subjective analysis would be necessary where the reference system is not use.

BAND CONGESTION

Finally, an analysis could be based on spectrum congestion in a band. One method of analyzing spectrum congestion would be to use the GMF to plot the number of assignments as a function of the assigned frequency. Another possible approach for measuring spectrum congestion is given in an NTIA report [Hinkle, 1975] where it is referred to as spectral crowding and is defined as follows.

$$M = \frac{1}{A\ B} \sum_{j=1}^{n} A_j \ B_j$$

where,

M = the measure of spectrum congestion

A_j = area required to operate the j'th system without having mutual interference to another system

B_j = frequency separation required to operate the j'th system without having mutual interference to another system

A = the total area outlined by a summation of the individual A_j areas, excluding their overlapping areas

B = the total allocation bandwidth

APPENDIX A

SPECTRUM CONSERVATION TECHNIQUES

INTRODUCTION

Several radio frequency bands have been identified, within the United States, by the National Telecommunications and Information Administration (NTIA), the Federal Communications Commission (FCC), and internationally by the World Administration Radio Conference (WARC) as being heavily used. Because of the concern engendered over not being able to provide spectrum to satisfy future telecommunications requirements, the spectrum management community in recent years has developed interest in spectrum conservation techniques, which will lead to more efficient utilization of spectrum. The increased interest in this subject is evident in the fact that the first conference, devoted entirely to radio spectrum conservation techniques, was held in London, England, in 1980, followed by a second conference in 1983.

Long-range planning is necessary to ensure that spectrum resources will be available to accommodate future telecommunications service requirements. Potentially congested bands need to be identified with sufficient lead time for economical technology to be developed, and procedures/regulations developed that encourage users to deploy spectrum efficient telecommunication systems before a spectrum crisis occurs. That is, by the time more efficient spectrum utilization techniques become economically available, there is such a large investment in the deployed equipment that it is practically impossible to take advantage of equipment/techniques that conserve spectrum space.

The subject of efficient spectrum utilization and spectrum conservation techniques generally falls into two categories:

1. those associated with the design of a new system involving the choice of techniques that will optimize the spectrum utilization characteristics of the new system in the electromagnetic environment where it must operate

2. those associated with the mutual operation of the new system with other systems in the electromagnetic environment and the optimization of the design parameters of the new system so any potential interactions that could degrade the performance of the new system or other systems are minimized

The second category is generally referred to as the electromagnetic compatibility (EMC) problem and must be taken into consideration in evaluating the spectrum efficiency of any system.

One important general principal of spectrum utilization efficiency is that any energy radiated should be kept to a minimum consistent with providing an adequate grade of service. Such minimization should be effective in the time, frequency, and space domains. The major telecommunications system design factors that affect spectrum utilization efficiency are modulation, transmitter output device, waveguide components and related devices, antennas, and receiver signal processing. The following is a discussion of these design factors and their importance in relation to efficient spectrum utilization and the related tradeoffs in implementing spectrum conservation techniques.

MODULATION

The evaluation of spectrum-efficient modulation schemes is very complex. In general, there are tradeoffs in the occupied bandwidth, required signal-to-noise ratio (S/N) (for analog modulated systems), required bit-error rate (BER) (for digital modulated systems), and the specified performance and emission spectrum sideband radiated energy. For each modulation approach, all of these factors can be exchanged. For example, one can increase the S/N, if one is willing to decrease the occupied bandwidth and vice versa. This tradeoff is expressed by the Shannon-Hartley Law:

$$C = B \log_2 \left(1 + \frac{S}{N}\right) \qquad \text{(A-1)}$$

where C is the channel capacity or the maximum rate of message transmission, in b/s, and B is the bandwidth of the channel, in Hz. In any given telecommunications requirement, the effect on spectrum utilization in terms of greater bandwidth and less radiated power or vice versa must be addressed. In some cases, greater spatial denial may be more acceptable than a greater emission bandwidth from a frequency assignment viewpoint.

Narrowband analog techniques for land mobile operation are being investigated by NTIA [Shelton, 1984] for assessment of their spectrum efficiency factors. In an overall comparison of amplitude compandored single sideband (ACSSB), 12.5 kHz narrowband FM (NBFM), and 25 kHz FM, the NTIA study found ACSSB to be as high as 2.5 times as spectrum efficient as 25 kHz FM, and NBFM as high as 1.8 times as spectrum efficient as 25 kHz FM.

Analog Modulation

As a rule, analog modulated voice circuits require less bandwidth than their digitally encoded counterparts. A speech signal requires a nominal 4 kHz of analog bandwidth in the transmission channel, if the signal is sent in analog form. Since current PCM practice converts the analog speech signal into a 64-kb/s digital pulse stream called a voice channel (voice circuit), it appears that a bandwidth of 32 kHz is required to transmit speech using PCM. Until recently, only FM transmission has been used in microwave radio systems, despite its relatively large bandwidth requirement. New microwave systems are being designed for SSBAM operation, resulting in a substantial improvement in frequency-spectrum utilization. An SSB signal can be transmitted in a bandwidth equal to the highest signal frequency. However, by Carson's rule, the FM system requires approximately twice the sum of the peak frequency deviation and the highest signal frequency. The frequency deviation is selected to bring the thermal and intermodulation noise into balance. This optimum frequency deviation is about 4-MHz peak on terrestrial systems and contributes significantly to the necessary bandwidth.

The "capture" characteristic of an FM signal, in which the strong signal captures the receiver, is inherent in FM systems. The threshold carrier-to-noise ratio (C/N), below which the output S/N ratio deteriorates significantly, is approximately 10 to 15 dB. This can be improved and extended by using the threshold extension receiver. A threshold improvement in the range of 3 to 7 dB can be achieved. Because of the noise characteristic, a substantial improvement in the postdetection S/N ratio can be obtained with pre-emphasis and de-emphasis techniques. In practical FM systems, the input high-frequency signal components that are normally low are pre-emphasized at the transmitter, and the output high-frequency noisy components that are high are de-emphasized at the receiver. By using properly matched pre-emphasis and de-emphasis filters, an improvement in $(S/N)_o$ in the range of 10 to 15 dB can be obtained.

TABLE A-1 shows a comparison of performance of the FM and SSB systems [Bell Laboratories, 1982]. In this table, S/N_{SSB} is defined as the ratio of the peak transmitted power to the noise spectral density; f_m is the highest signal frequency, and f_d is the peak frequency deviation. Without pre-emphasis, the relative performance of the FM and SSB systems, with respect to S/N, is 0.5 when $f_d = f_m$, as shown in TABLE A-1. With pre-emphasis, this ratio can be brought to unity when $f_d = f_m$. For high-index modulation, where $f_d > f_m$, the FM system can be made to have an S/N advantage over the SSB, but at the expense of much greater bandwidth.

TABLE A-1

PERFORMANCE COMPARISON

Modulation Scheme	Necessary Bandwidth	Postdetection Signal-to-Noise Ratio
FM	$2\ (f_d + f_m)$	$(f_d^2/2\ f_m^2)\ (S/N_{SSB})$
SSBAM	f_m	S/N_{SSB}

Digital Modulation

An important aspect of digital radio transmission is the type of modulation used, relative to its efficiency in spectrum utilization and its sensitivity to noise and interference. Generally, in the 7/8 GHz band, the digital data signal is modulated onto an IF (70 MHz) carrier using the appropriate modulation techniques. Then, the RF subsystem upconverts the modulated signal to the 7/8 GHz transmission frequency. The frequently used digital modulation techniques are quadrature amplitude modulation (QAM), quadrature partial response (QPR), frequency shift keying (FSK), and phase shift keying (PSK). Functional descriptions of these modulation schemes, which are appropriate for digital radio applications, are provided. The performance analysis of these modulation schemes is summarized by means of a table. Based on the Shannon limit, it is possible to find a modulation scheme that requires the lowest average S/N to achieve the same b/s/Hz efficiency [Feher, 1981].

In the Nyquist theorems on necessary bandwidth, it is possible to transmit the symbol rate (f_s) independent symbols per second in a channel (low-pass filter), having a bandwidth of only $B_s = (f_s/2)$ Hz. To transmit f_b bits per second in a baseband channel, the necessary bandwidth (B_b) equals ($f_b/2$) Hz in a binary transmission system. In M-ary transmission systems, each transmitted symbol contains n information bits, where $n = (\log_2 M)$. The f_s is given by $f_s = (f_b/n)$ bits/s, B_s is only $(f_s/2) = (f_b/2\ n)$ Hz. The spectrum efficiency of baseband transmission systems is defined to be the ratio of the bit rate to the necessary bandwidth, in units of bits/second/Hertz, (b/s/Hz). It is a convenient and frequently employed characteristic of all digital transmission systems. When the bit rate is the same for all systems, the binary transmission system provides bandwidth efficiency of 2 bits/s/Hz, the M-ary transmission system has (2 n) bits/s/Hz baseband spectral efficiency.

Since double-sideband amplitude modulation (DSB AM) involves the transmission of a redundant sideband, single-sideband amplitude modulation (SSB AM) is used because of 50% bandwidth reduction. The single-sided Nyquist bandwidth equals one-half of the symbol rate, and the double-sided Nyquist bandwidth equals the symbol rate. When the bit rate is the same for both systems, the binary transmission system using single-sideband amplitude modulation has 2 b/s/Hz baseband spectral efficiency, and the binary transmission system using double-sideband amplitude modulations provides bandwidth efficiency of 1 b/s/Hz. Often, double-sideband amplitude modulation is used because digitally modulated carrier signals can be expressed in double sideband-suppressed carrier amplitude modulation (DSB-SC AM) quadrature notation.

If dual polarization waves are used, one can double the rate of transmission of information. The spectral overlapping can be achieved by the use of both V polarization and H polarization. Theoretically, the M-ary transmission system provides bandwidth efficiency of n b/s/Hz on a single polarization or (2 n) b/s/Hz if dual polarization is used [Yamamoto, 1981].

Description Of Digital Modulation Schemes

There are three basic modulation techniques: amplitude modulation (AM), phase modulation (PM), and frequency modulation (FM). All of the digital modulation techniques can be conceptualized in terms of intermediate frequency sinusoids (carriers) modulated by low frequency (baseband) signals that convey the digital information. Demodulation is the inverse process of the modulation. Coherent demodulation requires a reference signal transmitted from the modulator to the demodulator. Noncoherent demodulation does not require a reference signal.

Amplitude Modulation (AM) Techniques

The simplest digital AM technique is double-sideband amplitude modulation (DSB AM). Since the carrier conveys no information, efficiency can be improved by the use of double sideband-suppressed carrier (DSB-SC) AM. This process preserves the characteristics of the shape of the power spectrum of the baseband signal. Both DSB techniques involve the transmission of a redundant sideband. The occupied bandwidth can be reduced by a factor of two by using single sideband (SSB) amplitude modulation. However, the digitally modulated carrier signals (including QPR, QAM) are expressed in double sideband-suppressed carrier (DSB-SC) quadrature component notation.

Quadrature Amplitude Modulation (QAM) Techniques

Quadrature Amplitude Modulation (QAM) is another AM alternative. This technique involves summing two DSB-SC AM signals 90° apart in phase. The modulated signals consist of two components, in-phase I and quadrature Q. 16-QAM is a combined amplitude and phase modulation technique, this scheme conveys 4 bits of information per symbol with 16 possible signal states. A 16-QAM modulator block diagram is illustrated in Figure A-1 [Feher, 1981]. The f_b rate binary signals are divided into two independent binary data signals, each having a rate of $(f_b/2)$ bits/s. The following 2-to-4-level baseband converter converts these $(f_b/2)$ bits/s data streams into four-level signals having a symbol rate of $f_s = (f_b/4)$ symbols/second. These four-level signals are spectrum-shaped by a premodulation low-pass filter that causes no intersymbol interference at the sampling instants. Both in-phase I and quadrature Q carriers (70 MHz) are then amplitude modulated with the spectrum-shaped signals. By summing these two double sideband-suppressed carrier amplitude-modulated signals 90° apart in phase, 16-QAM is obtained.

64-QAM is an extension of 16-QAM. This is achieved by doubling the number of levels from 4 to 8. The modulator block diagram for 64-QAM is identical to that of 16-QAM as seen in Figure A-1. The basic difference for 64-QAM is a 2-to-8-level baseband converter. Another method of implementing 64-QAM is the nonlinear amplified 64-state quadrature amplitude modulation (NLA 64-state QAM) [Hill, 1983]. NLA 64-state QAM employs a modulation technique where three QPSK modulators operate in parallel. The 64-QAM signal is generated by summing three QPSK RF signals of relative powers 0, -6, and -12 dB with an equal weight combiner [Demarest, 1984]. Quadrature phase shift keying (QPSK) signals could be accomplished by using 4 phases in a PSK system.

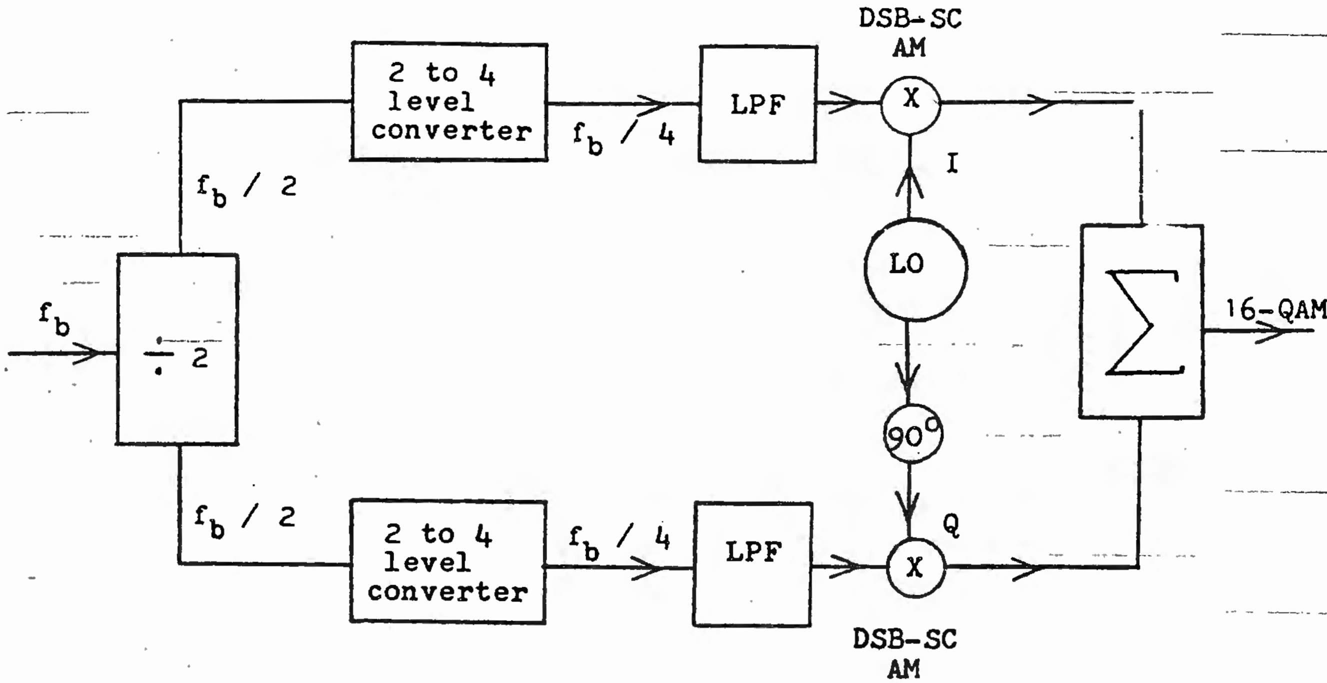


Figure A-1. 16-QAM modulation block diagram.

Phase Modulation (PM) Techniques

The most simple approach is binary phase shift keying (BPSK), in which the carrier phase is shifted by 0 or 180 degrees. A constant-amplitude carrier is utilized over the pulse interval. All pulses designating logic 1 have 0° phase, and all pulses designating logic 0 have 180° phase.

Quadrature, or four-phase PSK (QPSK) encodes two bits at a time into one of four possible carrier phases spaced 90° apart. The QPSK signal consists of in-phase I and quadrature Q components. During each two bit time interval, the I carrier is BPSK signal modulated by one bit and the Q carrier is modulated by the other bit [Feher, 1981]. The modulated I channel and the modulated Q channel BPSK signals are orthogonal because there is a 90° shift in the I and Q modulated paths. The combined signal can take on any one of four possible phases, and abrupt phase transitions of 90°, 180°, or 270° can occur. When a 225° vector has been encoded to represent the 00 logic state, then the 135°, 45°, and 315° vectors represent the 01, 11, 10 logic states, respectively.

The classical 8-phase PSK signal can be obtained by amplitude modulations in quadrature. A modern approach in the system design reports that 8-PSK signal may be generated by a set of phase shifters. The classical 8-PSK modulator block diagram is illustrated in Figure A-2 [Feher, 1981]. The f_b b/s data is split into three independent binary data signals, each having a rate of $(f_b/3)$ b/s. The 2-to-4-level converter provides one of the four possible levels of a polar baseband signal at a and at b. The logic state of symbol A determines whether the positive or negative signal level should be present at a. The logic state of symbol C determines whether the larger or smaller signal level should be present at a. If C = 1, then the signal level at a is larger than that at b. If C = 0, then the signal level at a is smaller than that at b. The four-level polar baseband signals at a and at b are amplitude-modulated with the in-phase I and quadrature Q carriers (70 MHz). By combining these two double sideband-suppressed carrier amplitude-modulated signals 90° apart in phase, the 8-phase PSK signal is obtained. The eight equiprobable phase states are 22.5°, 67.5°, 112.5°, 157.5°, 202.5°, 247°, 292.5°, and 337.5° vectors in the signal-state space diagram.

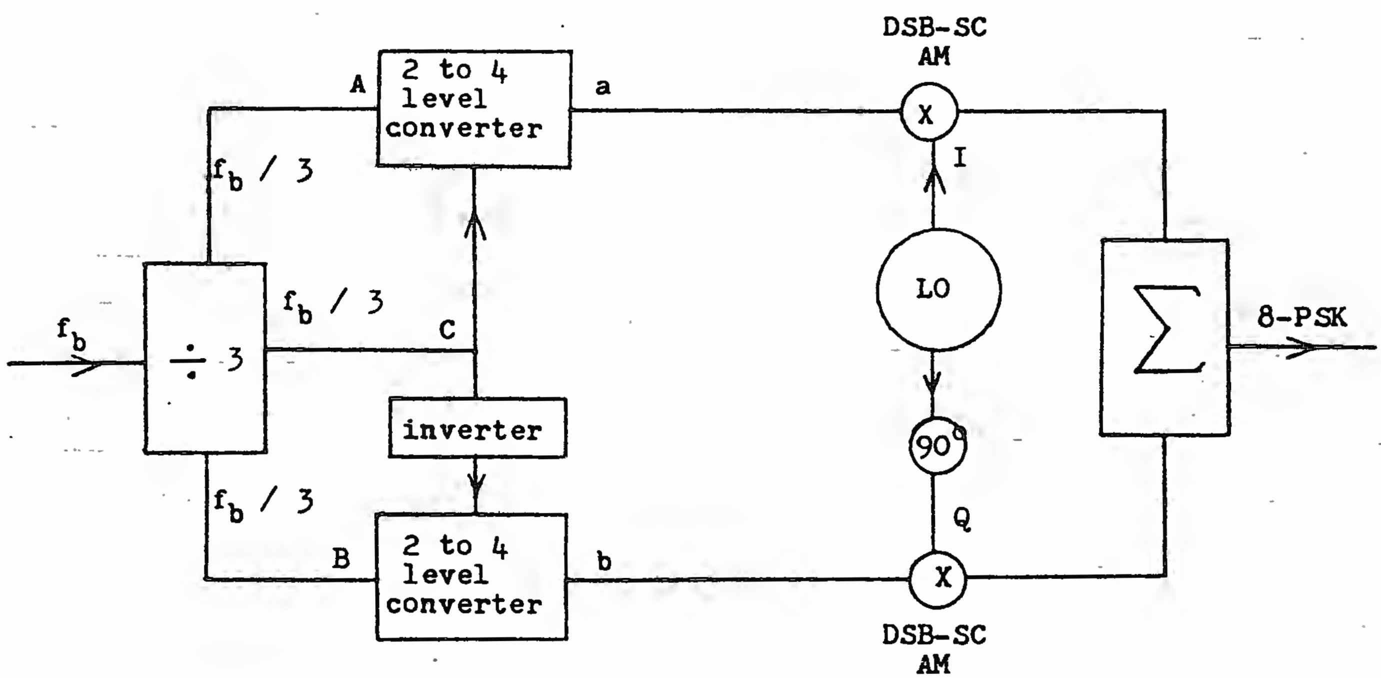


Figure A-2. 8-PSK modulator block diagram.

Frequency Modulation (FM) Techniques

In binary frequency shift keying (2-FSK), the carrier (70 MHz) signal is shifted back and forth between two distinct, predetermined frequencies. For binary data signals, there are two logic levels that determine a mark frequency (f_m), and a space frequency (f_s), separated from the center frequency (f_c) by a deviation frequency (f_d), and from each other by ($2\ f_d$) [Ehrman, 1979]. The frequency deviation parameter or modulation index, m, is defined to be the ratio of frequency spacing ($2\ f_d$) to baud rate R; $m = (2\ f_d) / R$. The spectrum of four-level frequency shift keying (4-FSK) with a 0.25 modulation index is similar to that of 2-FSK with a 0.5 modulation index, allowing a doubling of the data rate for the same modulation bandwidth. In the M-ary signaling, there are M logic levels that determine M frequencies, again separated by ($2\ f_d$) and centered about f_c. Thus, for a 3-level partial response modulation scheme, the three frequencies are ($f_c - 2\ f_d$), f_c, ($f_c + 2\ f_d$). Since phase discontinuity restrictions apply to frequency shift keying, a modern approach in the design of FSK is employed to avoid the serious phase discontinuity.

3-Level Partial Response (PR) Modulation Techniques

Adjacent pulses may interfere with one another, this is known as intersymbol interference. A 3-level partial response modulation technique deliberately introduces a limited amount of intersymbol interference (ISI) over a correlation span of one bit [Feher, 1981]. The net result is reshaping of binary pulse trains. When there is overlap between this bit and the last bit at the sampling instant, the output of the partial response technique is the algebraic sum of these two adjacent data bits. If the individual bit waveform has unit amplitude at the sampling instant, there, will be three, rather than two, distinct amplitude levels: +2, 0, -2, depending on the relative polarity of adjacent data bits. A 3-level partial response modulation technique is employed in digital radio systems in conjunction with FM modulation techniques.

Quadrature Partial Response (QPR) Techniques

Quadrature partial response modulation is the addition of two 3-level partial response signals in quadrature. Each 3-level partial response signal has three states due to the reshaping that results from its adjacent bit correlation. When the in-phase and quadrature signals are combined, the result is nine possible states. With QAM, two binary baseband data signals can be transmitted through the use of two DSB-SC AM carrier signals in quadrature phase. It is possible to convert the modulator from a binary signal to a 3-level partial response signal. This simple change of the baseband signal modifies the modulation from binary QAM to quadrature partial response. The implementation of quadrature partial response is shown in Figure A-3. The conversion filter delays the input by one bit and adds algebraically undelayed and delayed bit waveforms.

In AM-PSK modulation, the carrier is amplitude modulated and phase modulated in a binary manner [Feher, 1981]. The quadrature partial response signal can be considered as two AM-PSK 3-level partial response signals, phase modulated in quadrature. Each such signal has three states. The state at the tips of the vectors, 180° apart in phase, represents logic 0, and the state at the center (absence of carrier) represents logic 1.

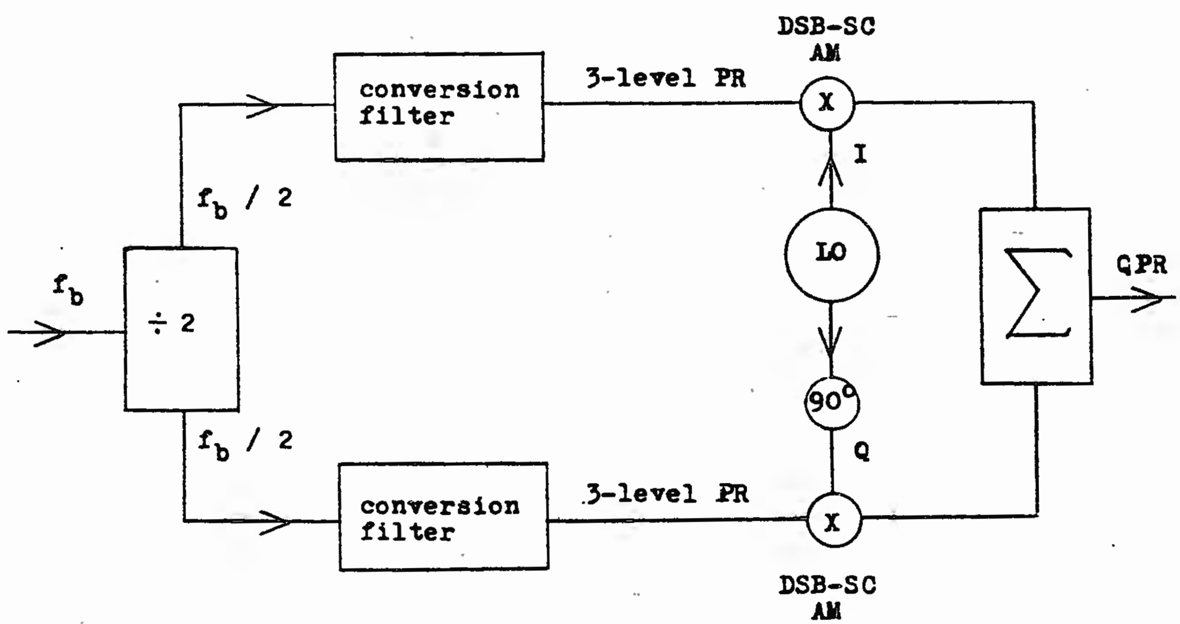


Figure A-3. Quadrature partial response modulator block diagram.

The spectral efficiency of a given modulation scheme requires the knowledge of power spectral density (PSD) and the measurement of necessary bandwidth. The power spectral density of higher order phase shift keying, such as 4-PSK, will take the same form as BPSK, but with the frequency variable rescaled on the frequency axis to reflect higher bandwidth efficiency. The necessary bandwidth is defined to be Nyquist bandwidth. CCIR Report 378-4 [CCIR, 1982a] lists the Nyquist bandwidth for various modulation schemes.

The theoretical maximum limits of these modulation schemes are summarized in TABLE A-2. The spectral efficiency is specified in the double-sided Nyquist bandwidth that equals the symbol rate. Single polarization has been assumed. By increasing the number of transmission levels (M), a higher spectrum efficiency is achieved. M-ary signaling could be accomplished by using M frequencies in FSK system or M phases in PSK system.

TABLE A-2 indicates performance in terms of the theoretical C/N requirement to obtain a specified probability of error (P_e). The C/N term represents the mean-carrier power to mean-noise power ratio specified in the double-sided Nyquist bandwidth that equals the symbol rate. Ideal roll-off filtering has been assumed. When the system is in an additive white Gaussian noise (AWGN) environment with no intersymbol or external interference, then the error is caused by AWGN. In this table, the value of C/N is estimated [Feher, 1981], [Hill, 1983]. This approximation for C/N is accurate to within 1 dB. It can be seen that a $P_e = 10^{-8}$ performance requires a higher C/N ratio than for a $P_e = 10^{-4}$. By increasing the number of transmission levels (M), a higher C/N ratio is required to attain the same P_e. When the C/N ratio is limited, a higher number of modulation is used, since this provides a tradeoff of quality signaling for bandwidth efficiency.

TABLE A-2

COMPARISON OF COMMON DIGITAL MODULATION TYPES

System	Variant	Spectrum Efficiency (b/s/Hz)	C/N (dB) for $P_e = 10^{-4}$	C/N (dB) for $P_e = 10^{-8}$
Amplitude	SSB, SC	2		
Modulation	DSB, SC	1		
Phase	BPSK	1	8.5	12
Modulation	QPSK	2	12	15
	8-PSK	3	17	20.5
Frequency	2-FSK	1		
Modulation	4-FSK	2		
Other	3-level PR	1		
Modulation	QPR	2	15	18.3
	16-QAM	4	19.5	23
	64-QAM	6	25.3	27

TABLE A-3 details the spectrum efficiency requiement for reported commercial digital radio systems at 7 GHz and 8 GHz. In this table, the figures are based on a manufacturer's technical data sheet, and the b/s/Hz (RF bandwidth) are assumed. The equipment manufacturers include Rockwell International (Collins), Northern Telecom, Harris (Farinon), and TRW.

TABLE A-3

TECHNICAL DATA FOR DIGITAL RADIO SYSTEM				
Type Number	Modulation Technique	Data Rate (Mb/s)	Emission Bandwidth (MHz)	Spectrum Efficiency (b/s/Hz)
MDR-8-5N	8-PSK	45.501	20	2.26
DM8-4A-45	16-QAM	44.736	20	2.24
AN/FRC-171(V) 10	QPR	26.112	14	1.9
AN/FRC-171(V)	QPSK	--	--	1.0
AN/FRC-162	3-level PR	--	--	--

Summary

Digital modulation requires more frequency spectrum than does analog modulation; however, it has the advantage of better performance (i.e., more accurate information transmission).

To make use of digital modulation while simultaneously conserving frequency spectrum, multiple level digital modulation techniques, such as 64-QAM, have been devised. In 1980, 16-QAM was the state-of-the-art, in 1985, it was 64-QAM. The latest goal is 256-QAM. Ultimately, system noise is the limiting factor in the maximum number of multilevels attainable.

TRANSMITTER OUTPUT DEVICES

Unwanted emissions, such as harmonics and a spurious noise floor are sources of communication interference. The transmitter of a communications system should be designed to minimize the power levels of all unwanted emissions or at least to reduce their power levels to a required number of decibels below that of the carrier. By minimizing unwanted emission generation, spatial denial can be reduced.

Typical unwanted frequency power level values of various candidate 7/8 GHz transmitter output devices are given in this discussion.

Transmitter Output Device Candidates

The candidate 7/8 GHz transmitter output devices are as follows.

I. Vacuum Tube Category
   A. Klystrons
   B. Traveling-Wave Tubes (TWT)

II. Semiconductor Category
   A. Gallium Arsenide Field Effect Transistors (G_aA_s FET)

Klystrons

Klystrons have the following harmonic power levels below carrier power.

2nd Harmonic	:	-20 dbc
3rd Harmonic	:	-25 dbc
4th Harmonic	:	-35 dbc

Klystrons have spurious noise power, measured across a 1 MHz bandwidth, of -90 dbc. That is, the spurious noise power is 90 decibels below the desired carrier power (dbc).

Traveling-Wave Tubes (TWT)

Traveling-wave tubes have the following harmonic power levels below carrier power.

2nd Harmonic	:	-15 dbc
3rd Harmonic	:	-25 dbc
4th Harmonic	:	-30 dbc

Traveling-wave tubes have spurious noise power, measured across the 1 MHz bandwidth, of -90 dbc. That is, the spurious noise power is 90 decibels below the desired carrier power.

Gallium Arnsenide Field-Effect Transistors (G_aA_s FET)

The G_aA_s FET has a harmonic power level of >-30 dbc, if it operates over a wide range of input signals. It improves to >-45 dbc, if its operation stays within the linear range of its characteristic. That is, for the above two cases, the harmonic power is at least 30 or 45 decibels below the desired carrier power, respectively.

Summary

In general, all transmitting output devices used in LOS microwave systems have inherently high harmonic output levels and require the use of transmitter filters to suppress the harmonic frequencies.

WAVEGUIDE COMPONENTS

A variety of hardware devices (filters) have been developed to minimize potential interference and thus, use the spectrum more efficiently.

Filters

Filters are used to reduce the amount of unwanted energy radiated or received. Highpass, lowpass, and bandpass filters are available for both transmitter and receiver applications.

The effects of transmitted harmonics, particularly from high power systems, pose serious problems to other equipments. TABLE A-4 has transmitter filter information. It lists various types and classifications of filters with their respective applicable frequency range.

TABLE A-4

TRANSMITTER-FILTER CHARACTERISTICS

Filter type	Classification	Applicable frequency range	Power capabilities	Passband loss, db	Stop-band attenuation, db	Tuning	Spurious
Microwave-reflective:							
Coupled resonators	Bandpass	Above 1 GHz	20% of waveguide	0.1	40	Mechanical	Many
SRI three-cavity resonator	Bandpass	Above 1 GHz	1-9 MW	0.3	30-60	Mechanical	Many
Serrated side waveguide	Bandpass	Above 1 GHz	Less than waveguide	0.1	40	Fixed	Few
Waffle iron	Bandpass	Above 1 GHz	1-4 MW	0.1	>40	Fixed	Few
Stripline	Band reject	200-1,000 MHz	50 MW	0.1	>50	Fixed	Few
Microwave dissipation:							
General leaky wall	Bandpass	Above 1 GHz	80% of waveguide	0.25	40	Fixed	Few
Straight wall	Low pass	Above 1 GHz	80% of waveguide	0.5	10-30	Fixed	Few
Offset wall	Low pass	Above 1 GHz	80% of waveguide	0.5	20-40	Fixed	Few
Serpentine wall	Low pass	Above 1 GHz	80% of waveguide	2	5-55	Fixed	Many
Circular leaky wall	Low pass	Above 1 GHz	80% of waveguide	0.5	25-45	Fixed	Few
Circular side waveguide	Low pass	Above 1 GHz	10-20 kW	0.5	>18	Fixed	Few
Coaxial leaky wave	Low pass	1-3 GHz	140-280 kW	0.3	15-40	Fixed	Few
General directional coupler	Low pass	Above 1 GHz	80% of waveguide	0.3	30	Fixed	Few
Transfer coupler	Low pass	Above 1 GHz	80% of waveguide	0.2	6	Fixed	Few
Ferrite filters	Band reject	Above 1 GHz	Low	0.1	>30	Fixed or electrical	Few
Leaky wall wire grid	Mode filter	Above 1 GHz	80% of waveguide	TE_{10} mode 0.1	TE_{20}, TE_{30} 30	Fixed	
Periodic filter	Bandpass	Above 1 GHz	500 kW peak	1.4	>50	Fixed	None

Note, from TABLE A-4, that each filter type has a corresponding power capabilities range. The dissipation filter types have the advantage of being able to handle more power than the reflective filter types. Also note the wide difference in the values shown in the power capabilities column.

Receiver spurious responses and intermodulation can be mitigated by the use of waveguide filters. TABLE A-5 has receiver filter information. It lists various techniques and classifications of filters with their respective applicable frequency range. Filters that are described as "1-30 GHz" (waveguide cavity resonator) would be applicable candidates for point-to-point microwave systems.

TABLE A-5

RECEIVER-FILTER CHARACTERISTICS

Technique	Classification	Frequency range	Fractional bandwidth, %	Passband insertion loss, db	Stop-band attenuation, db	Tuning	Spurious transmission characteristics
LC resonant circuits	Bandpass or band reject	Up to ~200 MHz	2	1-2	>30	Wide range	Medium
Spiral inductance tuner	Bandpass	200-400 MHz	1-2	~2	>30	Wide range	Medium
Helical resonator	Bandpass	200-400 MHz	0.4	1.5	>50	5:1 range	Medium
Butterfly resonator	Bandpass	100-1,000 MHz	0.5	~2	>40	5:1 range	Poor
Butterfly resonator	Band reject	100-1,000 MHz	0.2	1-2	26-44	2:1 range	Poor
Bridged-T filter	Band reject	To 1,000 MHz	0.01	~3	60-100	Difficult and critical	Good
Cavity resonator	Bandpass	200-3,000 MHz	0.2	1-2	>50	....	Medium
Coaxial cavity resonator	Bandpass	100-3,000 MHz	0.2	~1.5	~60	....	Good
Ferrite resonator	Bandpass	2-7 GHz	0.2	~2	~40	Electrical	Medium
Ferrite resonator	Band reject	2-7 GHz	0.5	~0.5	10-15	Electrical	Medium
Backward-wave amplifier	Bandpass	2-4 GHz	0.5	-20	20	Electrical	Poor
Modified hybrid	Band reject	500-4,000 MHz	......	~2	13-40	Less than an octave	Medium
Waveguide cavity resonator	Bandpass	1-30 GHz	0.2	~0.5	~60	....	Good
Waveguide cavity resonator	Band reject	1-30 GHz	~0.1	~0.3	~60	....	Good
Diode up-converter	Bandpass	Broad	......	-12	40	Electrical	Excellent

Note, in TABLE A-5, power is not a major concern in receiver filters, since receivers operate on low power signals.

Filter tuning (see TABLES A-4 and A-5) can either be fixed (i.e., non-tunable) or variable. For the latter, either electrical or mechanical tuning can be used.

A typical commercially available microwave lowpass filter characteristic is shown in Figure A-4. Attenuation, in dB, versus normalized frequency is shown. The loss spec is the same as insertion loss. Its the loss that exists over the desired frequency range. The rejection spec is the required/specified loss at a certain frequency in the undesired frequency range, hence it defines the sharpness of the rolloff of the filter. The measurement limit describes the limitations of the attenuation test measuring apparatus.

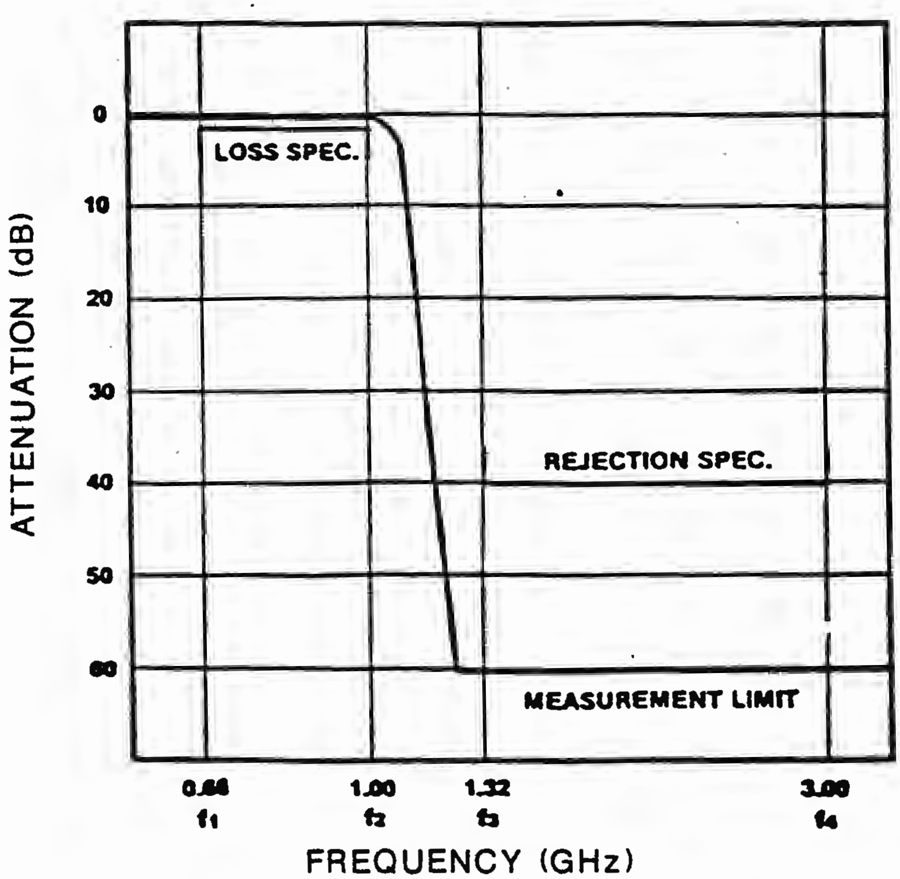


Figure A-4. Lowpass filter characteristics.

ANTENNAS

Spatial denial is a key factor that must be taken into consideration when addressing spectrum conservation. One of the major telecommunications system components that contributes to spatial denial of frequency re-use is the system antenna.

In recent years, significant advances in technology in the antenna design areas of polarization discrimination and sidelobe reductions have resulted in capabilities for using use the spectrum more efficiently for point-to-point microwave telecommunications.

Frequency re-use can be achieved through the use of antenna-design spectrum-conservation techniques. The following is a discussion of the principal antenna design factors that determine spatial denial and frequency re-use.

Spatial Denial

For a fixed beamwidth, spatial denial is minimized if sidelobe generation is minimized. The antenna radiation pattern (and therefore sidelobe distribution) is mainly a function of the following.

1. Generic Antenna Type
2. Method of Feed
3. Aperture Blocking
4. Random Errors

Generic Antenna Types

The five antenna types discussed in this section including their respective feed techniques are shown in Figure A-5:

1. Cassegrain Reflector
2. Prime Focus Reflector
3. Offset-Fed Reflector

4. Conical Horn Reflector
5. Cassegrain Horn Reflector

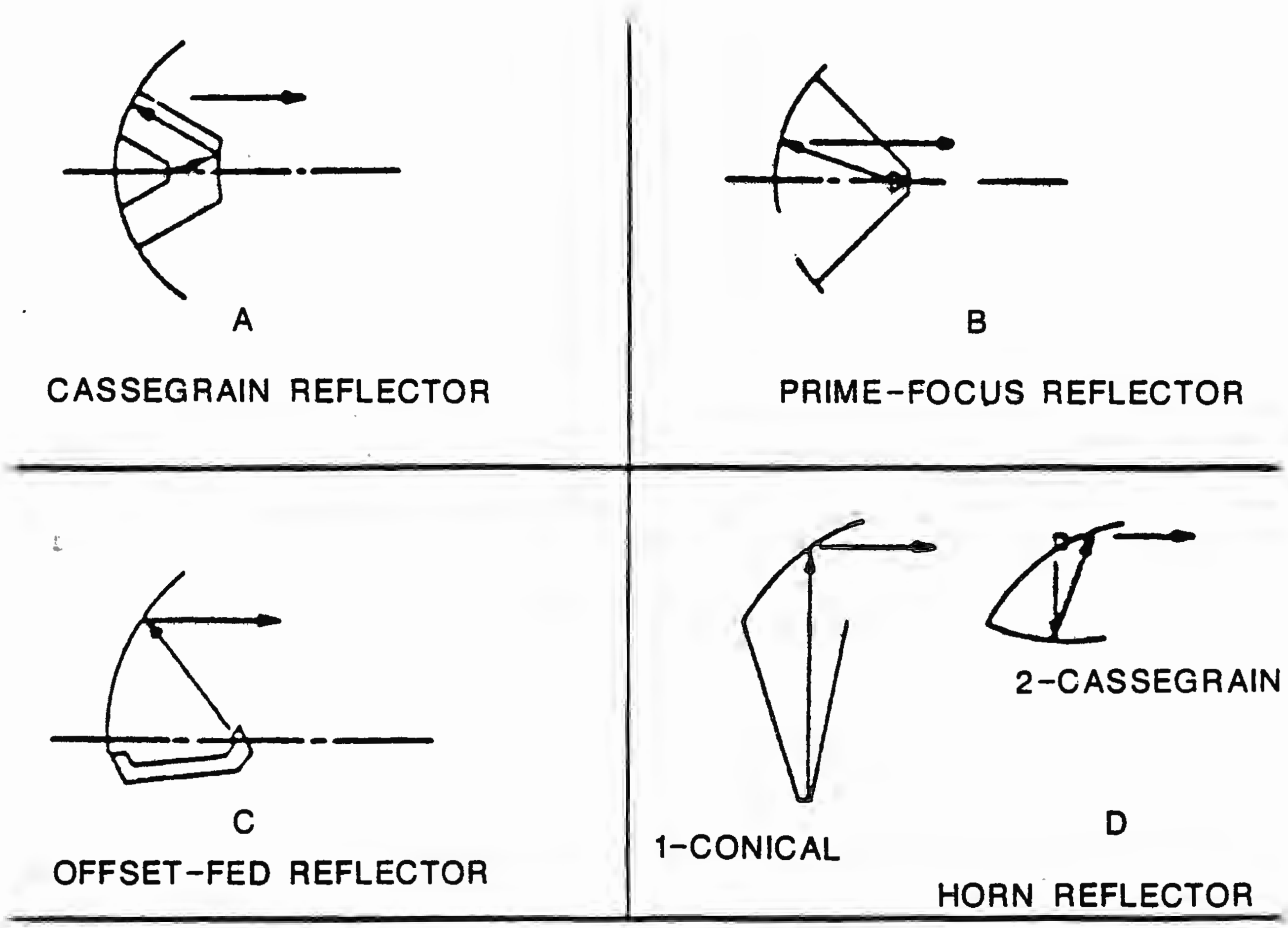


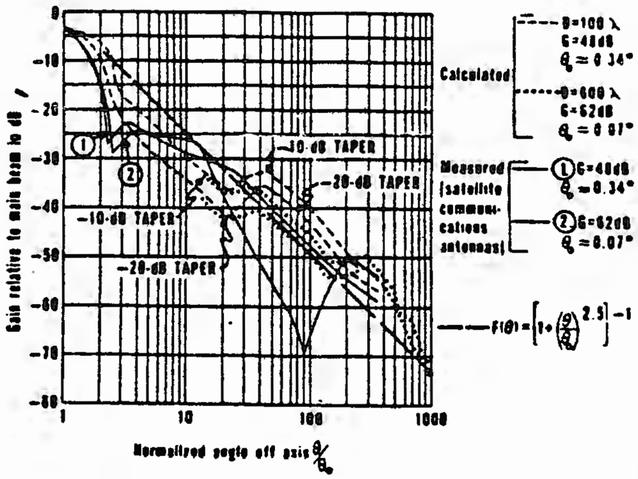
Figure A-5. Generic antenna types.

Figures A-6 through A-10 show antenna gain, relative to mainbeam, versus normalized angle off-axis for the Cassegrain-fed reflector, prime-focus-fed parabolic reflector, offset-fed parabolic reflector, conical horn reflector, and Cassegrain horn reflector antenna, respectively. In each figure, the independent variable (i.e., horizontal axis variable) has been normalized in terms of half beamwidth. That is Θ/Θ_o is the ratio of angle off-axis (Θ) and 1/2 times the -3 dB angle ($\Theta_o = 1/2\ \Theta_{3dB}$).

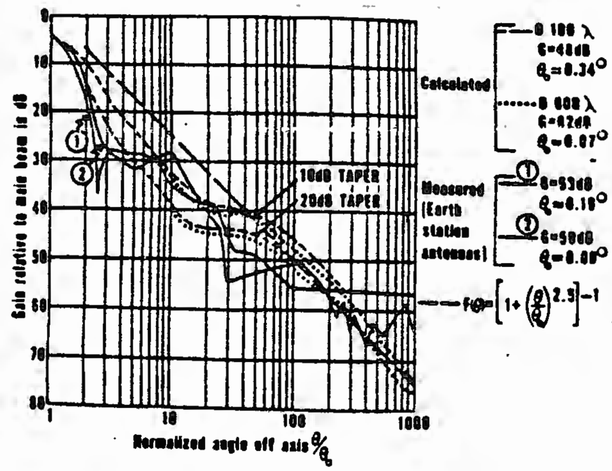
Figure A-6, representing the Cassegrain antenna pattern characteristics, has calculated envelope patterns for 100λ and 600λ diameter reflectors with -10 dB and -20 dB edge illumination. Aperture blockage and primary pattern spillover are included. Two measured antennas are plotted and identified on the graph. The measured patterns are precise plots to $\Theta/\Theta_o = 3.3$ (peak of first sidelobe), and only pattern envelopes of peaks are plotted beyond. An estimate for the amount of isolation from interference can be determined from these curves.

Figure A-7 represents the characteristics of prime-focus-fed parabolic reflectors with the same type of data as Figure A-6. An estimate of isolation from interference can be obtained from this figure.

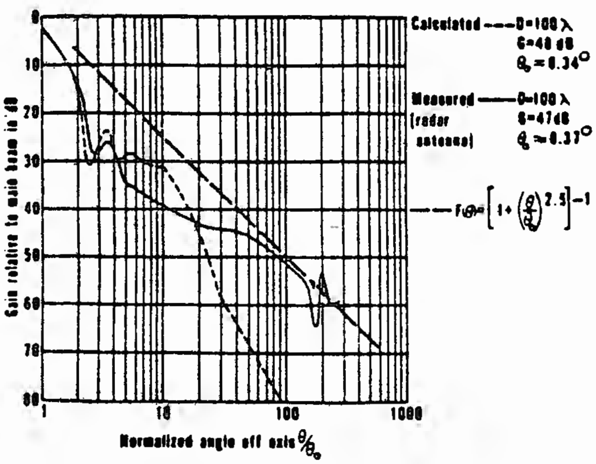
Figures A-8 through A-10 represent characteristics (calculated and measured) for the offset-fed reflector, conical horn reflector, and Cassegrain horn reflector types, respectively. The available data on these three figures are not as extensive as for the Cassegrain and prime-focus-fed parabolas, but these figures show how these antennas compare with respect to effectiveness in isolation from interference. The improvement over the Cassegrain and prime-focus center-fed reflector systems is basically due to the fact that blockage obstacles are practically nonexistent in the generic antenna types of Figures A-8 through A-10.



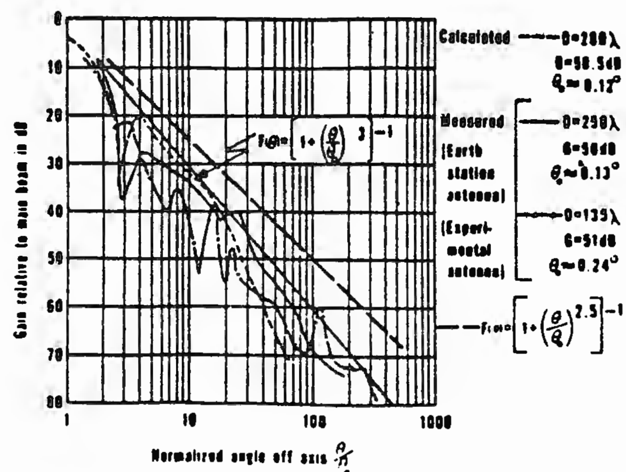
A-6. CASSEGRAIN-FED REFLECTOR



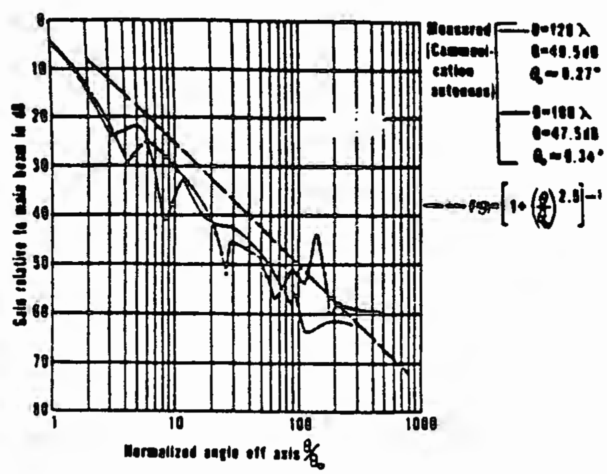
A-7. PRIME-FOCUS-FED PARABOLIC REFLECTOR



A-8. OFFSET-FED PARABOLIC REFLECTOR



A-9. HORN REFLECTOR (CONICAL) ANTENNA



A-10. HORN REFLECTOR (CASSEGRAIN) ANTENNA

TABLE A-6 is a comparative presentation of the generic antenna types. The data in TABLE A-6 is plotted in Figure A-11. A relative figure of merit can be calculated between any two antenna types as follows:

1. consider two different types of antennas of the same electrical size - for example, choose the Cassegrain and Cassegrain horn 100λ, reflector types, each with a reflector diameter of 100λ

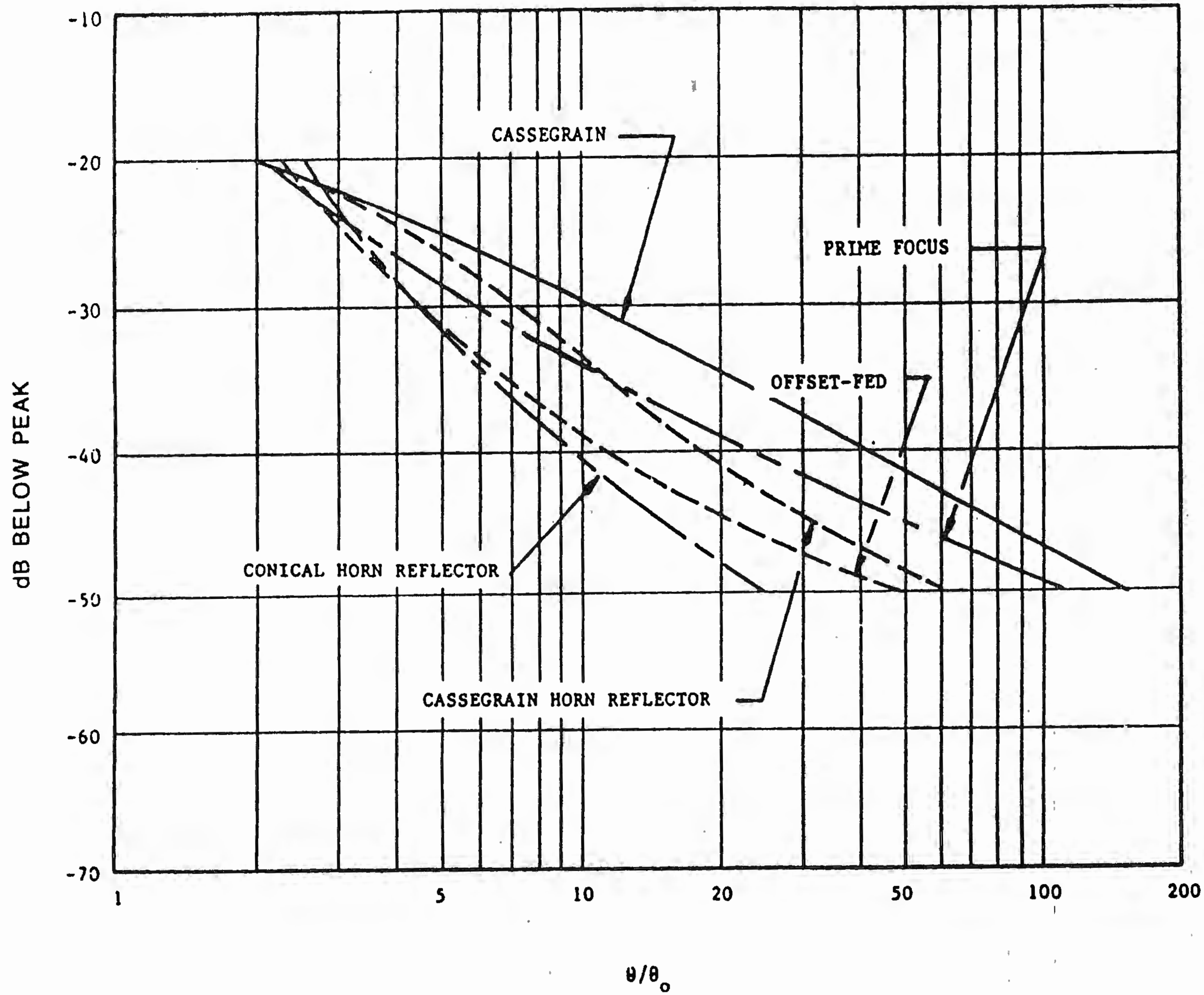
2. select a dB below peak for each type - for example, for the above, choosing -40 dB below peak gives a Θ/Θ_o value of 40 for the Cassegrain and a Θ/Θ_o value of 18 for the Cassegrain horn type

3. use a relative figure of merit of the Cassegrain horn type over the Cassegrain type is ≈ 2.2. - that is, the Cassegrain horn type achieves a -40 dB isolation capability at less than half the off-axis angle of the cassegrain type.

TABLE A-6

RADIATION PATTERN ENVELOPE COMPARISON

dB Below Peak	Cassegrain Θ/Θ_o 100λ	Prime Focus Θ/Θ_o 100λ	Offset Fed Θ/Θ_o 100λ	Conical Horn Reflector Θ/Θ_o 100λ	Cassegrain Horn Reflector Θ/Θ_o 100λ	Relative figure of Merit Average Values
-20	2	2.1	2.25	2.6	2	A/D = 1
-30	10	6	4.4	4.6	8	A/D ≈ 2
-40	40	25	11	9.6	18	A/D ≈ 3
-50	160	110	60	25	60	A/D = 4



A-11. RADIATION PATTERN ENVELOPE COMPARISON

Note: The tabulated figures of merit in TABLE A-6 represent an average; however, by using the method just described, the relative figure of merit for any antenna type pair can be calculated.

Shrouded Antennas

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

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

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

Frequency Re-use

Antenna polarization discrimination is an effective and economical means to double the channel capacity of the fixed (point-to-point) service. The method consists of transmitting one wideband channel on vertical polarization and a second channel, occupying the same spectrum, on horizontal polarization. Alternatively, right- and left-hand circular polarization may be used. The discrimination obtainable between two cross-polarized beams depends on two geographic features: the climate (which determines the rain statistics) and the relative locations of the areas served by the beams.

Depolarization caused by rain is an important effect, both with linear and with circular polarization. It is possible to correct for the differential phase shift component of rain-induced cross-polarization and location-related cross-polarization by use of a mechanical (rotatable polarization) and electronic component (variable differential) phase shifter. Also, receiver adaptive cancellation techniques can be used. In the mainbeam, up to the -1 dB region, typical antennas can provide in excess of 30 dB of discrimination for circular polarization and 35 dB for linear polarization. New grid-type antennas recently developed have a cross-polarization discrimination in excess of 40 dB; however, grid-type antennas can only be used at frequencies below 3 GHz.

SIGNAL PROCESSING

In a LOS microwave communication system, signal processing is done at the transmitter and receiver terminals. Signal processing consists of performing electrical operations on a signal to produce certain desired characteristics. Signal processing can affect such parameters as amplitude, frequency, phase, level, reliability, etc. [Freeman, R.L., 1981].

The use of signal processing techniques can improve the processing gain of a system, permitting lower transmitter output power levels for specified receiver output performance criteria. Thus, through the use of signal processing techniques, the transmitter output power can be lowered, which reduces the spatial denial of a system. However, it should be noted that signal processing techniques are used by the microwave link designers to improve link reliability and are not generally considered for the purpose of spectrum efficiency.

Companding

Companding consists of two functions, compressing and expanding. Compressing is done at the transmitter terminal and expanding at the receiver terminal. Figure A-12 shows a functional analogy of a compandor, showing some typical numbers.

A compression of the intensity range of speech signals at the input signals to the input circuit of a communication channel is done by imparting more gain to weak signals than to strong signals. At the far-end output of the communication circuit, the expandor performs the reverse function. It restores the intensity of the signal to its original dynamic range.

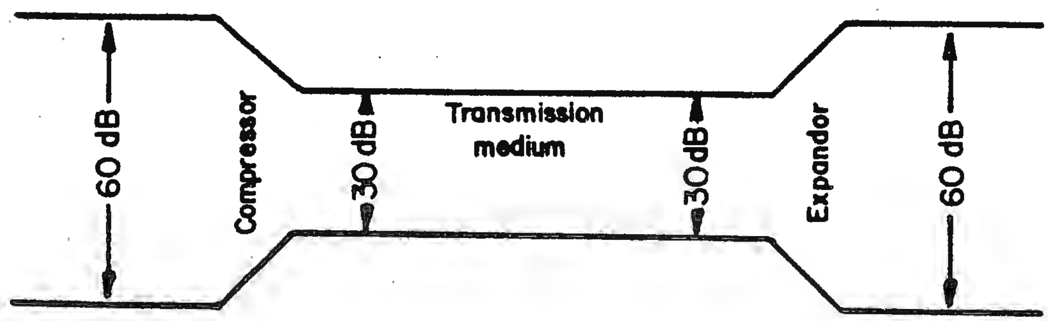


Figure A-12. Functional analogy of a compandor.

There are three advantages of compandors.

1. They tend to improve the signal-to-noise ratio on noisy speech circuits.

2. They limit the dynamic power range of voice signals reducing the chances of overload of carrier systems.

3. They reduce the possibility of crosstalk.

The signal-to-noise advantage of a compandor varies with the multichannel loading factor of FDM equipment and thus, depends on the voice level into the FDM channel modulation equipment. At best, an advantage of 20 dB may be attained on low-level signals.

Pre-emphasis and De-emphasis

Pre-emphasis is done at an FM transmitter and de-emphasis at the FM receiver to have an essentially constant signal-to-noise ratio across the entire baseband at the receiver output. Pre-emphasis is accomplished by increasing the peak deviation during the FM modulation process for the higher baseband frequencies. This increase of peak frequency deviation is done in accordance with a curve designed to effect compensation for the ramplike noise at the FM receiver output.

The pre-emphasis characteristic is achieved by applying the modulating baseband frequencies to a passive network that "forms" the input signal. At the receiver end, after demodulation, the baseband signal is applied to a de-emphasis network that restores the baseband frequencies to their original amplitude configuration. Figure A-13 shows a simplified block diagram of an FM radiolink system, indicating the effects of pre-emphasis and de-emphasis on the signal and thermal noise. The use of pre-emphasis and de-emphasis can improve the signal-to-noise ratio of a system by approximately 4 to 7 dB. This is equivalent to a reduction of 4 to 7 dB in transmitted carrier power; thus, it can be equated to a reduction in spatial denial.

Diversity Transmission and Reception

Diversity transmission and reception is based on the fact that radio signals arriving at a point over separate paths may have noncorrelated signal levels. That is, at one instant of time, a signal on one path may be in a condition of fade, while the identical transmitted signal on another path may not.

Diversity transmission separates a transmitted signal on one or more basis such as:

1. frequency (more than one carrier frequency is used)
2. space (includes antenna polarization)
3. time (a time delay of two identical signals on parallel paths)

4. path (signals transmitted on geographically separate paths)

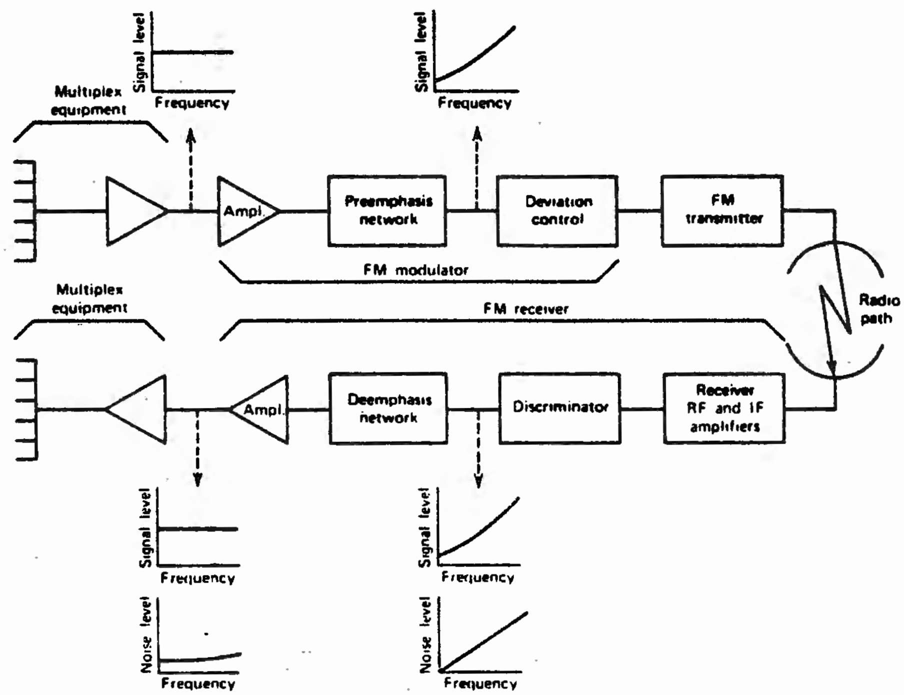


Figure A-13. Simplified block diagram of an FM radiolink system showing the effects of pre-emphasis and de-emphasis on the signal and thermal noise.

The most common forms of diversity are those of frequency and space. A frequency diversity system utilizes the phenomenon that the period of fading differs for carrier frequencies separated by 2 to 5%. Such a system employs two transmitters and two receivers, with each pair tuned to a different frequency (usually 2 to 3% separation, since the frequency band allocations are limited). If the fading period at one frequency extends for a period of time, the same signal on the other frequency will be received at a higher level, with the resultant improvement in system reliability.

As far as system reliability is concerned, frequency diversity provides a separate path, complete and independent, and consequently, one whole order of reliability has been added. However, its disadvantage is lower spectrum utilization efficiency, since additional frequencies are used without carrying additional information. This is a major drawback for frequency assignments in highly developed areas where the spectrum is congested.

The major advantage of space diversity is that no additional frequency assignment is required. In a space diversity system, if two or more receiving antennas are spaced many wavelengths apart (in the vertical plane), it has been observed that multipath fading will not occur simultaneously at both antennas. Sufficient output is almost always available from one of the antennas to provide a useful signal to the receiver diversity system. The use of two antennas at different heights provides a means of compensating, to a certain degree, for changes in electrical path differences between direct and reflected rays by favoring the stronger signal in the diversity combiner.

The receiver antenna separation required for optimum operation of a space diversity system may be calculated using the following formula.

$$S = \frac{3\lambda R}{L} \qquad \text{(A-2)}$$

where S = Separation (m)

R = Effective earth radius (m)

λ = Wavelength (m)

L = Path length (m)

Any spacing between 100 and 200λ is usually found to be satisfactory. The goal in space diversity is to make the separation of diversity antennas such that the reflected wave travels a half-wavelength further than the normal path.

Figure A-14a is a simplified receiver block diagram of a frequency diversity system and Figure A-19b is one of a space diversity system.

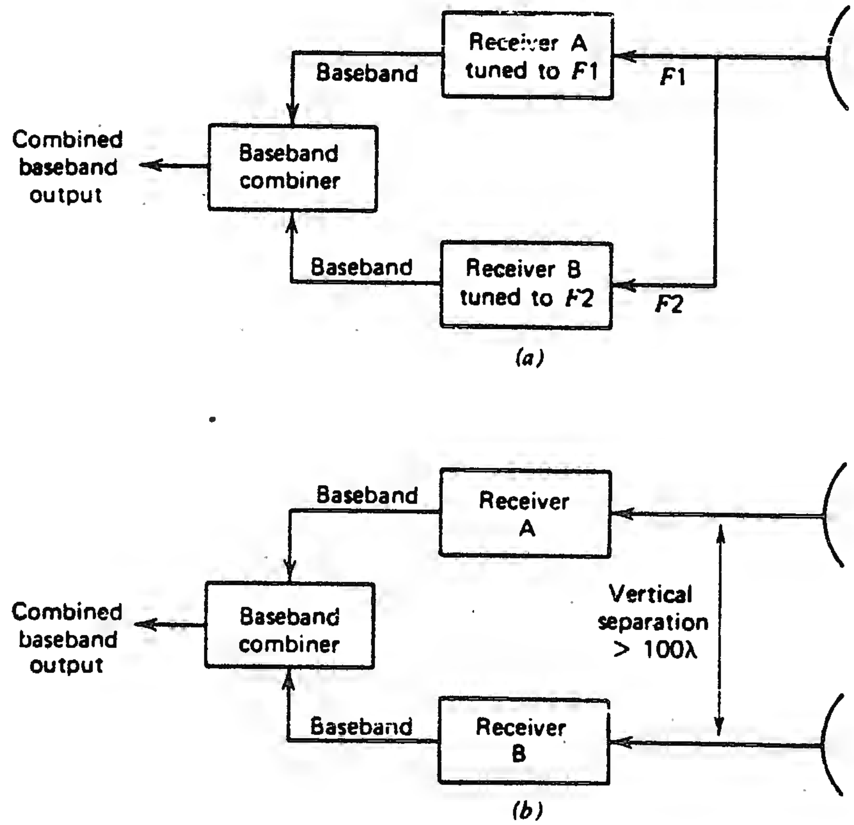


Figure A-14a. Simplified block diagram of a frequency diversity configuration $F_1 < F_2 + 0.02F_2$. Figure A-14b. Simplified block diagram of a space diversity configuration.

Figure A-15 shows approximate interference fading for nondiversity versus frequency diversity systems for various percentages of frequency separation.

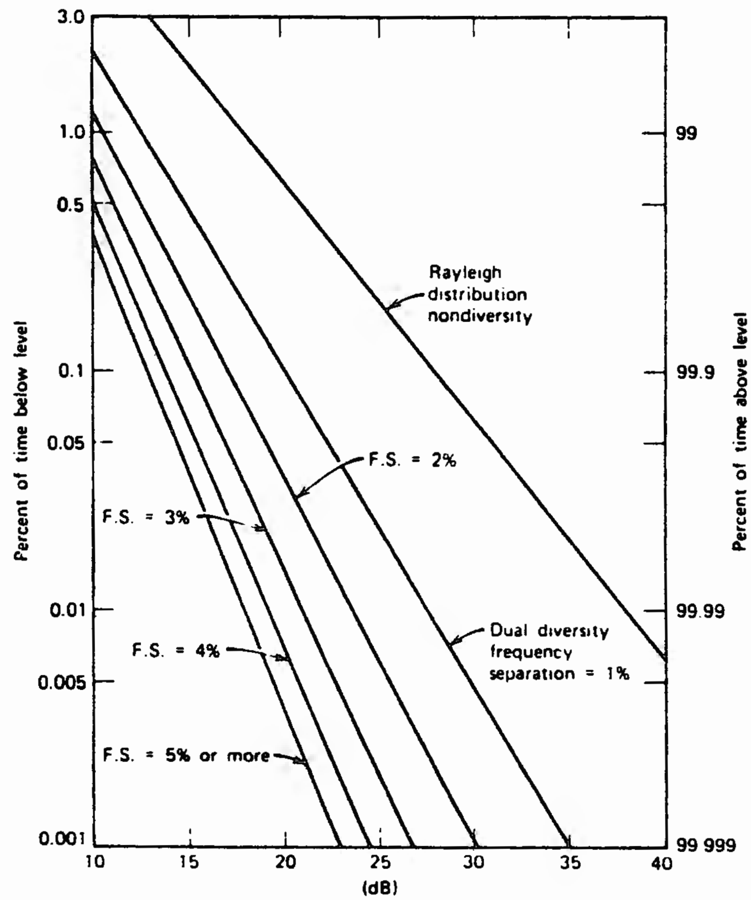


Figure A-15 Approximate interference fading distribution for a nondiversity system with Rayleigh fading versus frequency diversity systems for various percentages of frequency separation.

Diversity Combiners

A diversity combiner combines signals from two or more diversity paths. Combining is traditionally broken down into two major categories:

1. predetection

2. postdetection.

The classification is made according to where in the reception process the combining takes place. Predetection combining takes place in the IF. However, at least one system (manufactured by STC, U.K.) performs combining at RF. With the second type, combining is carried out at baseband (i.e., after detection).

For predetection combining, phase control circuitry is required unless some form of path selection is used. Figures A-16a and A-16b show simplified functional block diagrams of radiolink receiving systems using predetection and postdetection combiners.

Types of Combiners

Three types of combiners find common application in radiolink diversity systems. These are:

1. selection combiner
2. equal-gain combiner
3. maximal-ratio combiner (ratio squared).

The selection combiner uses but one receiver at a time. The output signal-to-noise ratio is equal to the input signal-to-noise ratio from the receiver selected for use at the time.

The equal-gain combiner adds the diversity receiver outputs. The output signal-to-noise ratio of the combiner is:

$$\frac{S_0}{N_0} = \frac{S_1 + S_2}{2N} \qquad \text{(A-3)}$$

where N = receiver noise.

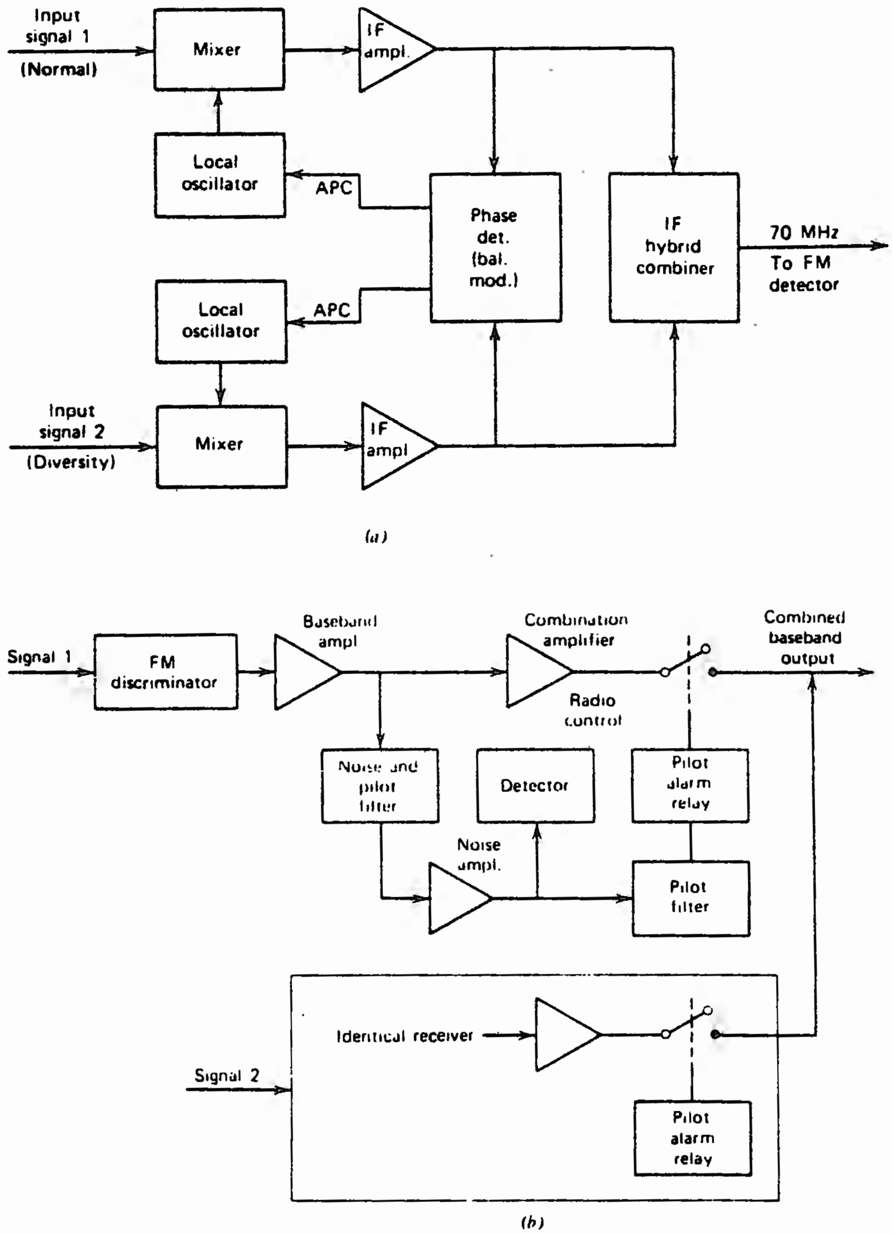


Figure A-16a. Predetection combiner (APC=automatic phase control). A-16b. Postdetection combiner (maximum ratio squared).

The maximal-ratio combiner uses a relative gain change between the output signals in use. For example, let it be assumed that the stronger signal has unity output and the weaker signal has an output proportional to gain (G). If signal gain is adjusted to be proportional to the ratio of input signals:

$$G = \frac{S_2}{S_1} \qquad \text{(A-4)}$$

then,

$$\left(\frac{S_0}{N_0}\right)^2 = \left(\frac{S_1}{N}\right)^2 + \left(\frac{S_2}{N}\right)^2 \qquad \text{(A-5)}$$

where N = receiver noise.

For the signal-to-noise ratio equation of the latter two combiners, the following are assumed.

1. All receivers have equal gain.
2. Signals add linearly; noise adds on an RMS basis.
3. Noise is random.
4. All receivers have equal noise outputs (N).
5. The output (from the combiner) signal-to-noise ratio S_0/N_0 is a constant.

Figure A-17 shows graphically a comparison of the three types of combiners. The expected gain in output signal-to-noise ratio for various orders of diversity is shown in this figure. The order of diversity refers to the number of independent diversity paths. If space or frequency diversity alone were used, two orders of diversity would be the case. If space and frequency diversity were used, four orders of diversity would exist.

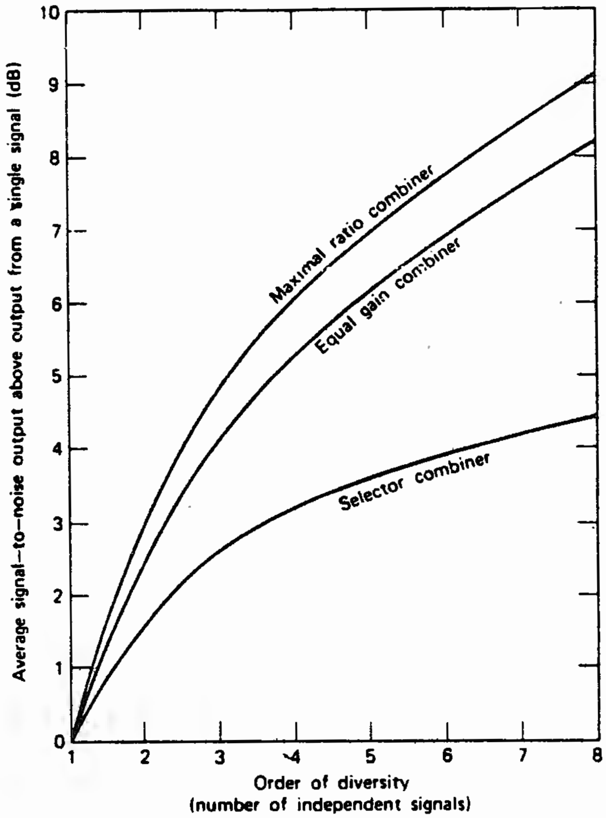


Figure A-17. Signal-to-noise ratio improvement in a diversity system for various orders of diversity.

Efficiency of diversity depends on the correlation of fading of the independent diversity paths. If the correlation coefficient is zero (i.e., there is no relationship in fading for one path to another), maximum diversity enhancement can be expected. The efficiency of a diversity system drops by half with a correlation coefficient of 0.8. Nearly full efficiency can be expected with a correlation coefficient of 0.3.

From Figure A-17, S/N improvement, using 2nd order diversity and a maximal ratio combiner, is 3; using 2nd order diversity and an equal gain combiner gives an S/N improvement of 2.5; using 2nd order diversity and a selector combiner gives an S/N improvement of 1.5.

Polarization Diversity

Polarization diversity consists of transmitting signals in the horizontal plane from one antenna and in the vertical plane from a second antenna. On the receiving end, two antennas are used, each having dual polarized feed horns for receiving signals in both planes of polarization. Polarization diversity is usually used in conjunction with conventional space diversity. This produces four signal paths that are relatively independent.

Polarization diversity can improve the system signal-to-noise ratio by approximately 20 to 30 dB.

Adaptive Equalizers

There are two types of adaptive equalizers used at the IF stage in receivers. They are known as frequency domain and time domain adaptive equalizers that effect the delay and phase distortions in receivers. Adaptive equalizers make a radio receiver less susceptible to fading. An approximate 4 to 6 dB improvement in the dispersive fade margin can be achieved with these equalizers in 64-QAM receivers. The major drawback of adaptive equalizers is their expense.

FM on Reducing Processing Techniques

In the transmission of data, an important design goal is error rate reduction. Various processing techniques such as redundant information transfer, use of parity, and use of supervisory signals are used to minimize error occurrences. The tradeoff in each case is reduced spectrum efficiency. If redundant transmission uses an additional modulated carrier, then the required spectrum is doubled. If a parity technique is used, additional overhead bits are required to be transmitted, increasing the required spectrum. If a separate supervisory signal (such as a pilot carrier) is transmitted to describe each channel's status (operational; non-operational, etc.), spectrum again is increased. Hence, in any considered error reducing scheme, the additional required frequency spectrum should be calculated and weighted against the error reduction before a firm design commitment is made.

Coding and Coding Gain

Forward error correction (FEC) is a method of optimizing a digital radio link, particularly on systems that are power limited. This is reflected by the coding gain, described later in this discussion. FEC utilizes a type of coding that permits a limited number of errors to be corrected at the receiving end by means of special coding and software (or hardware) implemented at both ends of a circuit. There are two broad classes of codes: block codes and convolutional codes.

With block-coding techniques, each group of K consecutive information bits is encoded into a group of N symbols for transmission over the channel. Normally, the K information bits are located at the beginning of the N-symbol block code, and the last N-K symbols correspond to the parity check bits formed by taking the modulo-2 sum of certain sets of K information bits. Block codes containing this property are referred to as systematic block codes. The encoded symbols for the $(K + 1)^{th}$ bit and beyond are completely independent of the symbols generated for the first K information bits and hence, cannot be used to help decode the first group of K information bits at the far-end receiver. This essentially says that blocks are independent entities, and one block has no enhancement capability on another.

Because N symbols are used to represent K bits, the code rate R of such a block code is K/N bits per symbol, or:

$$R = \frac{K}{N} \tag{A-6}$$

For instance, an encoder structure could be (7, 4) meaning N = 7 and K = 4. The information bits are stored in K = 4 storage devices, and the device is made to shift N = 7 times. The first K symbols of the block output are information symbols and the last N - K symbols are a set of check symbols that form the whole N - K symbol word. A block code may be identified with the notation (N, K, t), where t corresponds to the number of errors in a block of N symbols that the code will correct.

A convolution encoder is a linear finite-state machine consisting of a K - stage shift register and M linear algebraic function generators. The input data, which is usually, but not necessarily always, binary, is shifted along the register b bits at a time. Figure A-18 is an example of a convolutional encoder. If there were a five-stage shift register where the input data was shifted along 1 bit at a time and there were three modulo-2 adders (i.e., M = 3), using viterbi notation, the code would be described as a 5, 3, 1 convolutional code (i.e., K = 5, M = 3, b = 1).

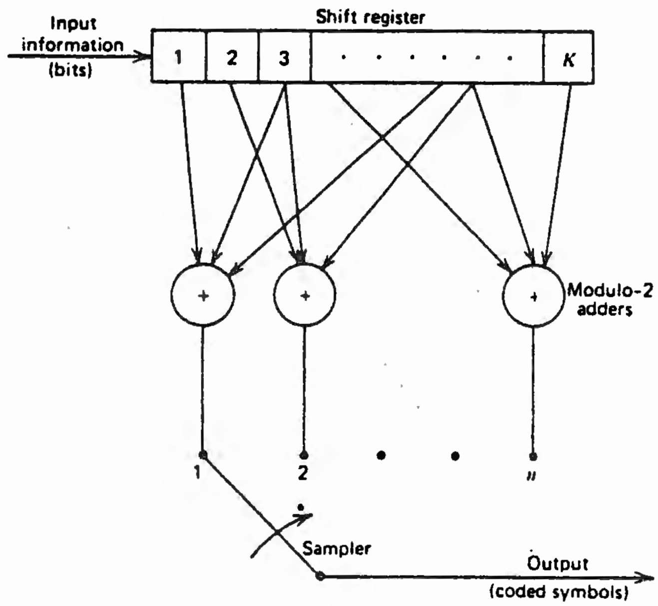


Figure A-18. Convolutional encoder, b = 1.

In Figure A-18, information bits are shifted to the right, 1 bit at a time (b = 1), through the K - stage shift register as new information bits enter from the left. Bits out of the last stage are discarded. The bits are shifted, one position each T seconds, where 1/T is the information rate in bits per second. The modulo-2 adders are used to form the output coded symbols, each of which is a binary function of a particular subset of the information bits in the shift register. The output coded symbols can be seen to depend on a sequence of K information bits, and thus, K is defined as the constraint length.

If 1 bit at a time is fed to the encoder (i.e., b = 1), each coded symbol carries an average of 1/M information bits, and the code is said to have a rate of 1/M. For the more generalized case, where b = 1, the rate is expressed:

$$R = b/M \qquad \text{(A-7)}$$

where b = number of bits shifted into the register at a time.

In Figure A-18, when the first modulo-2 adder is replaced by a direct connection to the first stage of the shift register, the first symbol becomes a replica of the information bit. Such an encoder is called a systematic convolutional encoder, as shown in Figure A-19

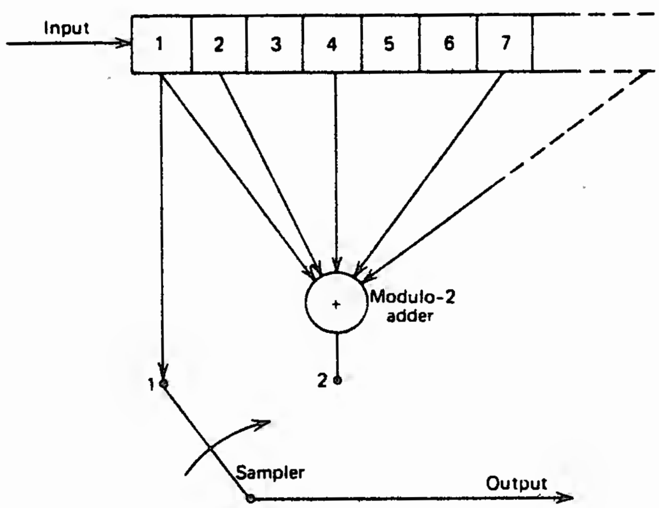


Figure A-19. Systematic convolutional encoder.

A convolutional code can be thought of as forming a tree structure as shown in Figure A-20.

At each node (branch point), the information bit determines which direction (i.e., which branch) will be taken following the convention "up" for 1 and "down" for 0. The M digits occurring on the branch selected correspond to the output symbols. A particular sequence of information bits then describes a particular path through the tree.

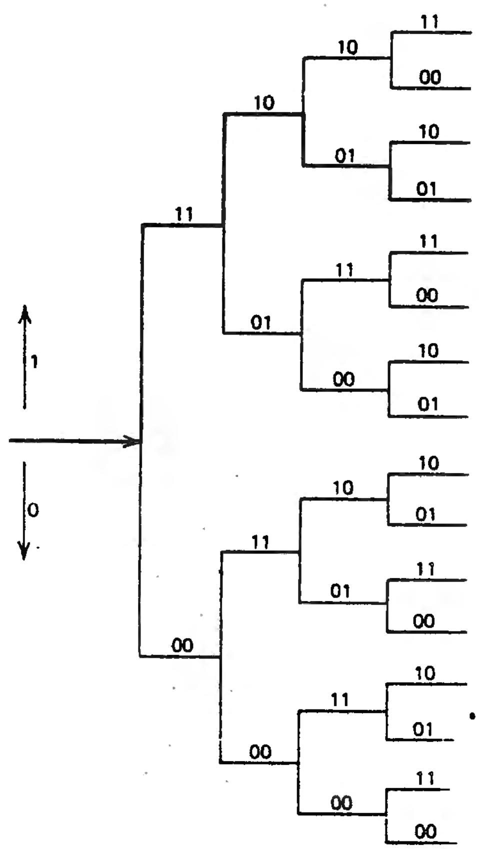


Figure A-20. A code tree.

Decoding algorithms for block and conventional codes are quite different. Because a block code has a formal structure, advantage can be taken of the known structural properties of the words or the algebraic nature of the constraints among the symbols used to represent an information sequence. An example of a class of powerful block codes with well defined decoding algorithms is the Bose-Chaudhuri-Hocquenghem (BCH) codes.

The decoding of convolutional codes usually is carried out by means of the probabilistic techniques, such as the sequential decoding algorithm by Wozencraft and the maximum likelihood technique by Viterbi (i.e., the Viterbi algorithm). Such techniques depend on the ability to home in on the correct sequence by designing efficient search procedures that discard unlikely sequences quickly. The sequential decoder differs from most other types of decoders in that when it finds itself on a wrong path in the tree, it has the ability to search back and forth, changing previously decoded information bits until it finds the correct tree path. The frequency with which the decoder has to search back and the depth of these backward searches is dependent on the value of the channel BER.

An important property of a sequential decoder is that if the specified constraint length is large enough, the probability that the decoder will make an error, approaches zero (i.e., a BER better than 1×10^{-9}). One cause of error is overflow, being defined as a situation in which the decoder is unable to perform the necessary number of computations in the performance of the tree search. If a computation is defined as a complete examination of a path through the decoding tree, a decoder has a limit on the number of computations it can make per unit time. The number of searches and computations is a function of the number of errors arriving at the decoder input, and the number of computations that must be made to decode one information bit is a random variable. An important parameter for a decoder is the average number of computations per decoded information bit. As long as the probability of bit error is not too high, the chances of decoder overflow will be low, and satisfactory performance will result.

For the previous discussion, it has been assumed that the output of a demodulator has been a hard decision, either a mark or a space, as an input to the decoder. If these decisions were soft decisions instead of hard decisions, additional improvement in error performance on the order of several decibels can be achieved. A "soft" decision means that the output of a demodulator is quantized into four or eight levels (2 or 3 bit quantization, respectively), and then certain decoding algorithms can use this additional information to improve the output BER. Sequential and Viterbi decoding algorithms can use this soft decision information effectively, giving them an advantage over algebraic decoding techniques that are not designed to handle the additional information provided by the soft decision.

The soft decision level of quantization is indicated conventionally by the letter Q that indicates the number of bits in the quantized decision sample. If Q = 1, a hard decision demodulator output is being dealt with; Q = 2 indicates a quantization level of 4; Q = 3, a level of 8; and so on.

TABLE A-7 is a summary of E_b/N_o requirements of several coded communication systems using BPSK modulation, showing coding gain capabilities. Coding gain is also a function of the modulation type selected for a particular digital radio link or system. TABLE A-8 illustrates this point.

Coding gain is strongly dependent on the system external to the code. It depends on the form of PSK (or other modulation type), interleaving and channel fading. Consequently, one must be extremely cautions in using the term coding gain and in applying coding gain to systems other than that for which it was calculated or measured. The significant decrease in encoding gain when DPSK is used rather than CPSK, and the extraordinary improvement when coding is applied to a rapidly fading channel are particularly noteworthy [Shaft, 1975].

TABLE A-7

CODING GAIN FOR SEVERAL CODED COMMUNICATION SYSTEMS, BPSK MODULATION, BER 10^{-5}

Quantization Coding Type	Level	Coding Gain[a] (dB)
Convolutional, K=7, R=1/2, Viterbi decoded	1	3.1
Convolutional, K=7, R=1/2, Viterbi decoded	3	5.2
Convolutional, K=7, R=1/3, Viterbi decoded	1	3.6
Convolutional, K=7, R=1/3, Viterbi decoded	3	5.5
Convolutional, K=9, R=1/4, Viterbi decoded	1	2.4
Convolutional, K=9, R=1/4, Viterbi decoded	3	4.3
Convolutional, K=24, R=1/2, sequential decoded, 20 kbps, 1000-bit blocks	1	4.2
Convolutional, K=24, R=1/2, sequential decoded, 20 kbps, 1000-bit blocks	3	6.2
Block Golay (24, 12)	3	4.0
Block Golay (24, 12)	1	2.1
Block Golay (127, 92)	1	3.3
Block BCH (127, 64)	1	3.5
Block BCH (127, 36)	1	2.3
Block Hamming (7, 4)	1	0.6
Block Hamming (15, 11)	1	1.3
Block Hamming (31, 26)	1	1.6

a. 9.6 dB required for the uncoded system and only additive white Gaussian noise conditions assumed.

TABLE A-8

CODING GAINS FOR SEVERAL MODULATION TYPES[a]			
Modulation	Quantization Level (bits)	E_b/N_o (dB) (for BER=$1X10^{-5}$)	Coding Gain (dB)
Coherent BPSK/QPSK	3	4.4	5.2
Coherent BPSK/QPSK	2	4.8	4.8
Coherent BPSK/QPSK	1	6.5	3.1
Octal PSK[b]		19.3	3.7
DBPSK[b]		36.7	3.6
DBPSK[b]		18.2	2.1
DE-QPSK[b]		19.0	3.0
Noncoherently demodulated binary FSK	1	11.2	2.1

Processing Gain Summary

TABLE A-9 is a summary of the processing gain that can be realized for each of the signal processing techniques discussed in this report.

---

a. In all cases, convolutional coding, K=7, R=½, Viterbi decoded.

b. Interleaving/deinterleaving assumed.

TABLE A-9

PROCESSING GAIN SUMMARY	
Signal Processing Technique (dB)	Signal Processing Gain (dBi)
Companding	10 to 20[a]
Preemphasis/ Deemphasis	4 to 7[b]
Frequency and/or Space Diversity	1.5 to 6[c]
Polarization Diversity	1.5 to 6[c]
Adaptive Equalization	4 to 6[d]
Coding	0.6 to 6.2[e]

---

a. Compandor S/N ratio advantage varies with the multichannel loading factor of FDM equipment and thus, depends on voice level into the FDM channel modulation equipment.

b. This is based on a single voice channel.

c. The 6 dBi gain corresponds to using 4th order diversity (i.e., using space and frequency diversity) and a maximal ratio combiner. The 1.5 dBi gain corresponds to using 2nd order diversity (i.e., using either space or frequency diversity) and a selector combiner.

d. This 4 to 6 dBi range is the improvement that can be achieved using adaptive equalization in 64-QAM receivers.

e. 6.2 dBi gain is based on using a convolutional code, with K=24, R=½. 0.6 dBi gain is based on using a block hamming (7, 4) code type.

Summary

This section has provided an overview of the major system design factors that affect the spectrum efficiency of a LOS microwave system. To evaluate the TSEF, a reference system must be identified. Therefore, a figure of merit of spectrum efficiency that considers these major design factors must be developed to determine appropriate reference systems. The figure of merit of spectrum efficiency would be derived by assigning a corresponding weight to each of the major system design factors discussed in this section. Then, within each major design factor category, each component/circuitry could be ranked in order by spectrum efficiency.

The TSEF concept could be used to develop the weighting and ranking values of each major design factor. A set of assumptions may have to be made at the outset to develop a figure of merit.

The development of a figure of merit of spectrum efficiency would provide a quantitative measure of spectrum efficiency of a system (i.e., the system is efficient, moderately efficient, or inefficient). This information could also be used in the NTIA system review process to assess the spectrum efficiency of a new system.

APPENDIX B

SYNOPSIS OF THE 7125-8500 MHz FREQUENCY RANGE

GENERAL

Equipment and systems in the 7125-8500 MHz frequency range were used to study the feasibility of the TSEF computer model. The 7125-8500 MHz frequency range is allocated to the fixed, mobile, and various space services on a primary basis. For convenience, the 7125-8500 MHz frequency range was grouped into three bands as follows:

Band 1:	7125 - 7750 MHz
Band 2:	7750 - 8025 MHz
Band 3:	8025 - 8500 MHz.

The 7125-8500 MHz band is used to support high-capacity government fixed microwave links (typically 600 voice channels or several video links). A large portion of this band is also used for military space radiocommunications in the fixed-satellite and mobile-satellite services.

ALLOCATION AND ASSIGNMENT DISTRIBUTION

The US National and ITU tables of frequency allocations for the entire 7125-8500 MHz, band along with the footnotes, for all the services are shown in TABLE B-1. The assignment distribution among federal agencies for the fixed service for each of the three bands (as identified previously) is given in TABLE B-2. Many of the point-to-point microwave links are used to transmit information from the remote enroute radar sites to control centers. Other point-to-point microwave links are used to support power distribution systems.

TABLE B-1

THE US NATIONAL AND ITU TABLES OF FREQUENCY ALLOCATIONS FOR THE 7125-8500 MHz BAND

(Page 1 of 3)

INTERNATIONAL			UNITED STATES			
Region 1 MHz	*Region 2 MHz*	*Region 3 MHz*	*Band MHz* *1*	*National Provisions* *2*	*Government Allocation* *3*	*Non-Government Allocation* *4*
7075–7250 FIXED MOBILE 809 810 811			7075–7125	809		FIXED MOBILE NG118
			7125–7190	US252 809	FIXED G116	
			7190–7235	809	FIXED SPACE RESEARCH (Earth-to-space)	
			7235–7250	809	FIXED	
7250–7300 FIXED FIXED-SATELLITE (Space-to-Earth) MOBILE 812			7250–7300		FIXED-SATELLITE (Space-to-Earth) MOBILE-SATELLITE (Space-to-Earth) Fixed G117	
7300–7450 FIXED FIXED-SATELLITE (Space-to-Earth) MOBILE except aeronautical mobile 812			7300–7450		FIXED FIXED-SATELLITE (Space-to-Earth) Mobile-Satellite (Space-to-Earth) G117	
7450–7550 FIXED FIXED-SATELLITE (Space-to-Earth) METEOROLOGICAL-SATELLITE (Space-to-Earth) MOBILE except aeronautical mobile			7450–7550		FIXED FIXED-SATELLITE (Space-to-Earth) METEOROLOGICAL-SATELLITE (Space-to-Earth) Mobile-Satellite (Space-to-Earth) G104 G117	
7550–7750 FIXED FIXED-SATELLITE (Space-to-Earth) MOBILE except aeronautical mobile			7550–7750		FIXED FIXED-SATELLITE (Space-to-Earth) Mobile-Satellite (Space-to-Earth) G117	
7750–7900 FIXED MOBILE except aeronautical mobile			7750–7900		FIXED	
7900–7975 FIXED FIXED-SATELLITE (Earth-to-space) MOBILE 812			7900–8025		Fixed FIXED-SATELLITE (Earth-to-space) MOBILE-SATELLITE (Earth-to-space) G117	
7975–8025 FIXED FIXED-SATELLITE (Earth-to-space) MOBILE 812						

TABLE B-1

THE US NATIONAL AND ITU TABLES OF FREQUENCY ALLOCATIONS FOR THE 7125-8500 MHz BAND

(Page 2 of 3)

INTERNATIONAL			UNITED STATES			
Region 1 MHz	*Region 2 MHz*	*Region 3 MHz*	*Band MHz* 1	*National Provisions* 2	*Government Allocation* 3	*Non-Government Allocation* 4
8025–8175 FIXED FIXED-SATELLITE (Earth-to-space) MOBILE Earth Exploration-Satellite (Space-to-Earth) 813 815	8025–8175 EARTH EXPLORATION-SATELLITE (Space-to-Earth) FIXED FIXED-SATELLITE (Earth-to-space) MOBILE 814	8025–8175 FIXED FIXED-SATELLITE (Earth-to-space) MOBILE Earth Exploration-Satellite (Space-to-Earth) 813 815	8025–8175	US258	EARTH EXPLORATION-SATELLITE (Space-to-Earth) FIXED FIXED-SATELLITE (Earth-to-space) Mobile-Satellite (Earth-to-space) (No Airborne Transmission) G117	
8175–8215 FIXED FIXED-SATELLITE (Earth-to-space) METEOROLOGICAL SATELLITE (Earth-to-space) MOBILE Earth Exploration-Satellite (Space-to-Earth) 813 815	8175–8215 EARTH EXPLORATION-SATELLITE (Space-to-Earth) FIXED FIXED-SATELLITE (Earth-to-space) METEOROLOGICAL-SATELLITE (Earth-to-space) MOBILE 814	8175–8215 FIXED FIXED-SATELLITE (Earth-to-space) METEOROLOGICAL SATELLITE (Earth-to-space) MOBILE Earth Exploration-Satellite (Space-to-Earth) 813 815	8175–8215	US258	EARTH EXPLORATION-SATELLITE (Space-to-Earth) FIXED FIXED-SATELLITE (Earth-to-space) METEOROLOGICAL SATELLITE (Earth-to-space) Mobile-Satellite (Earth-to-Space) (No Airborne Transmissions) G104 G117	
8215–8400 FIXED FIXED-SATELLITE (Earth-to-space) MOBILE Earth Exploration-Satellite (Space-to-Earth) 813 815	8215–8400 EARTH EXPLORATION-SATELLITE (Space-to-Earth) FIXED FIXED-SATELLITE (Earth-to-space) MOBILE 814	8215–8400 FIXED FIXED-SATELLITE (Earth-to-space) MOBILE Earth Exploration Satellite (Space-to-Earth) 813 815	8215–8400	US 258	EARTH EXPLORATION-SATELLITE (Space-to-Earth) FIXED FIXED-SATELLITE (Earth-to-space) Mobile-Satellite (Earth-to-space) (No Airborne Transmissions) G117	
8400–8500	FIXED MOBILE except aeronautical mobile SPACE RESEARCH (Space-to-Earth) 816 817		8400–8450		FIXED SPACE RESEARCH (Space-to-Earth) (Deep Space Only)	
	818		8450–8500		FIXED SPACE RESEARCH (Space-to-Earth)	SPACE RESEARCH (Space-to-Earth)

TABLE B-1

THE US NATIONAL AND ITU TABLES OF FREQUENCY ALLOCATIONS FOR THE 7125-8500 MHz BAND

(Page 3 of 3)

INTERNATIONAL

809—In the band 6 425-7 075 MHz, passive microwave sensor measurements are carried out over the oceans. In the band 7 075-7 250 MHz, passive microwave sensor measurements are carried out. Administrations should bear in mind the needs of the earth exploration-satellite (passive) and space research (passive) services in their future planning of this band.

810—Subject to agreement obtained under the procedure set forth in Article 14, in Region 2, the band 7 125-7 155 MHz may be used for Earth-to-space transmissions in the space operation service.

811—Subject to agreement obtained under the procedure set forth in Article 14, the band 7 145-7 235 MHz may be used for Earth-to-space transmissions in the space research service. The use of the band 7 145-7 190 MHz is restricted to deep space; no emissions to deep space shall be effected in the band 7 190-7 235 MHz.

812—The bands 7 250-7 375 MHz (space-to-Earth) and 7 900-8 025 MHz (Earth-to-space) may also be used by the mobile-satellite service. The use of these bands by this service shall be subject to agreement obtained under the procedure set forth in Article 14.

813—In the band 8 025-8 400 MHz, the power flux-density limits specified in No. 2570 shall apply in Regions 1 and 3 to the earth exploration-satellite service.

814—In Region 2, aircraft stations are not permitted to transmit in the band 8 025-8 400 MHz.

815—Subject to agreement obtained under the procedure set forth in Article 14, the band 8 025-8 400 MHz may be used for the earth exploration-satellite service (space-to-Earth) in Bangladesh, Benin, Cameroon, China, the Central African Republic, the Ivory Coast, Egypt, France, Guinea, Upper Volta, India, Iran, Israel, Italy, Japan, Kenya, Libya, Mali, Niger, Pakistan, Senegal, Somalia, Sudan, Sweden, Tanzania, Zaire and Zambia, on a primary basis.

816—In the space research service, the use of the band 8 400-8 450 MHz is limited to deep space.

817—Different category of service: in Belgium, Israel, Luxembourg, Malaysia, Singapore and Sri Lanka, the allocation of the band 8 400-8 500 MHz to the space research service is on a secondary basis (see No. 424).

818—Alternative allocation: in the United Kingdom, the band 8 400-8 500 MHz is allocated to the radiolocation and space research services on a primary basis.

UNITED STATES

US252—The bands 2110-2120 and 7145-7190 MHz, 34.2-34.7 GHz are also allocated for earth-to-space transmissions in the Space Research Service, limited to deep space communications at Goldstone, California.

US258—In the band 8025-8400 MHz, the non-Government earth exploration-satellite service (space-to-earth) is allocated on a primary basis. Authorizations are subject to a case-by-case electromagnetic compatibility analysis.

GOVERNMENT

G104—In the bands 7450-7550 and 8175-8215 MHz, it is agreed that although the military space radio communication systems, which include earth stations near the proposed meteorological-satellite installations will precede the meteorological-satellite installations, engineering adjustments to either the military or the meteorological-satellite systems or both will be made as mutually required to assure compatible operations of the systems concerned.

G116—The band 7125-7155 MHz is also allocated for Earth-to-space transmission in the Space Operations Service at a limited number of sites (not to exceed two), subject to established coordination procedures.

G117—In the bands 7250-7750 and 7900-8400 MHz and 20.2-21.2, 30-31, 39.5-40.5, 43.5-45.5 and 50.4-51.4 GHz the Government fixed-satellite and mobile-satellite services are limited to military systems.

TABLE B-2

DISTRIBUTION OF FIXED STATIONS IN THREE SUBBANDS AND THE NUMBER OF OCCURRENCES IN GMF

AGENCY[a]	7125-7750 MHz	7750-8025 MHz	8025-8500 MHz	7125-8500 MHz
FAA	2313	512	776	3601
DOE	496	162	257	915
N	355	117	155	627
AF	392	113	118	623
AR	191	113	147	451
TVA	134	34	78	246
CG	48	19	26	93
NASA	26	12	23	61
C	25	14	16	25
J	25	8	19	52
USIA	52	0	0	52
VA	27	11	0	38
I	17	6	1	24
Other	20	9	7	36
Total	4121	1130	1630	6874

a. Abbreviations are the same as those in ANNEX G of the NTIA Manual.

APPENDIX C

CALCULATION OF THE TSEF FOR FIXED TERRESTRIAL SERVICES

GENERAL

The technical spectrum efficiency factor (TSEF) is defined in Section 3 as:

$$TSEF = (T_r \times B_r \times S_r)/(T_s \times B_s \times S_s) \qquad (C-1)$$

where:

T_r = the time the reference system denies to others

B_r = the bandwidth the reference system denies to others

S_r = the physical space the reference system denies to others

T_s, B_s, S_s = the time, bandwidth, and physical space that the evaluated system denies to others, respectively.

All that is needed to calculate the TSEF is a method of computing the product of the time, bandwidth, and physical space denied to other potential users. The same method can be applied to both the system being evaluated and the reference system, and then the ratio in Equation C-1 can be computed. In general, both the transmitter and the receiver of a link deny spectrum space to other users; only transmitter denial is computed in this report. APPENDIX D provides a discussion and examples of combining transmitter and receiver denial to calculate band efficiency.

General formulas for computing the TBS product have been derived [Berry, 1977]. The most accurate of these requires both the emission spectrum of the transmitter and the admission function of the victim receiver, as well as their relative locations and a transmission loss model. The program used for this report, which computes the TSEF for the transmitter only, is based on Berry's "uniform denial metric" that assumes an idealized victim receiver with an isotropic, loss-free antenna and a perfect narrow selectivity function.

To further simplify the calculation, the transmitter emission spectrum is approximated (or bounded) by a step-function. Then the TBS product is:

$$TBS = T \sum_{i} B_i A_i(B_i) \tag{C-2}$$

where

T = the time that the transmitter is emitting

B_i = the increment in bandwidth

$A_i(B_i)$ = the area denied in bandwidth B_i.

In simplified calculations, B_1 is the intended emission bandwidth, and the B_2, B_3, etc., are sidebands or unwanted emissions. However, the B_i can be of any desired width. The area denied in each bandwidth increment depends on the power of the emission in that increment.

COMPUTING THE AREA DENIED

For the purposes of this report, the area denied by a transmitter is the area within which the power available at the terminals of an isotropic, loss-free antenna at a specified height (H_i) is greater than a specified value (P_i).

$$P_R = P_T + G_T(a) + G_R(b) - L - FDR, \quad \text{in dBW,} \tag{C-3}$$

where:

P_T = the transmitter's power in dBW

$G_T(a)$ = the power gain of the transmitter antenna at angle a

a = the angle between the transmitter antenna boresight and the line connecting the transmitter and receiver

$G_R(b)$ = the power gain of the receiver antenna at angle b

b = the angle between the receiver boresight and the line connecting the transmitter and receiver

L = the basic transmission loss in dB, depending on radio frequency, antenna heights, path characteristics, etc.

FDR = the frequency-dependent rejection of the receiver (tuned to its frequency) to the emission spectrum of the transmitter.

When the transmitter and receiver are the terminals of a fixed link, it is assumed that a = b = 0, and that FDR = 0. When the transmitter is a potential interferer to the receiver, the interference power depends on the FDR and on the angles a and b, as well as on all other terms. For the assumption that the probe receiver has an isotropic antenna, $G_T(b) = 0$.

The power level that will cause interference to the receiver depends on the receiver noise figure and the required signal-to-interference margin for the modulation type or emission characteristics of both the receiver and transmitter. For the calculations in this report, the power level that causes interference, denoted P_i = -70 dBm, is an input parameter. In a particular case, it would be determined in a separate analysis based on the system's characteristics and communications requirement. For the calculations in this report, P_i was set to a constant agreed to by TSC Working Group 13.

Calculation of the area denied is illustrated in Figure C-1. The boundary of the area is at a distance (d_b) from the transmitter (T). The d_b is the distance at which the transmitted power has decreased to the interference power (P_i). This distance depends on the power emitted in the direction (the directional antenna pattern) and on the transmission loss.

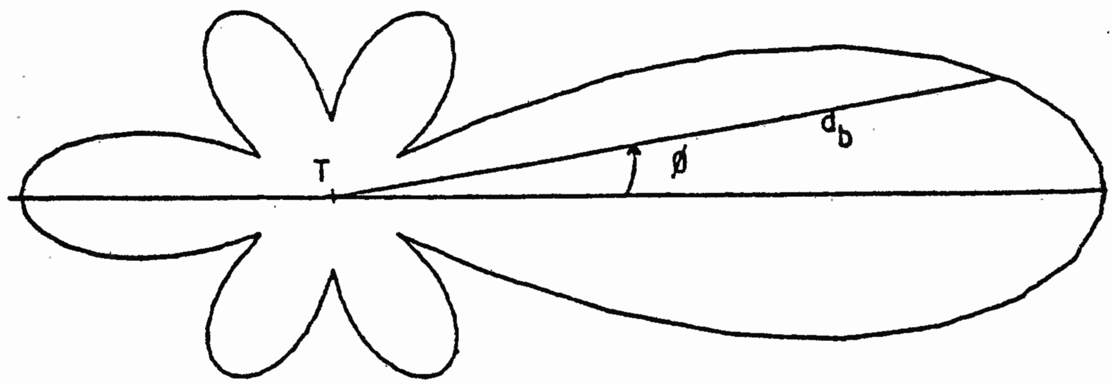


Figure C-1. Plan view of generic directional antenna pattern. The area denied to other users will have the same general shape as the antenna pattern.

The antenna pattern and transmission loss models used are discussed in the next two sections.

Modeling The Antenna Patterns

Microwave line-of-sight links usually use directional large aperture antennas. The directional pattern of such an antenna looks something like the solid line shown in Figure C-2. There is a narrow mainbeam containing most of the radiated power, many sidelobes containing much less power, and back-lobes, in the opposite direction from the mainbeam. The number, width, and relative power in the sidelobes depends mainly on the aperture size of wavelengths, the type of aperture, and the type of feed; but the details of the lobes are influenced by many other factors, such as mounting, surface tolerances, etc. For interference calculations, the detailed structure of the lobing is not useful. Instead, the envelope of the pattern, shown by the dashed line, is used to represent the power gain of the antenna as a function of direction.

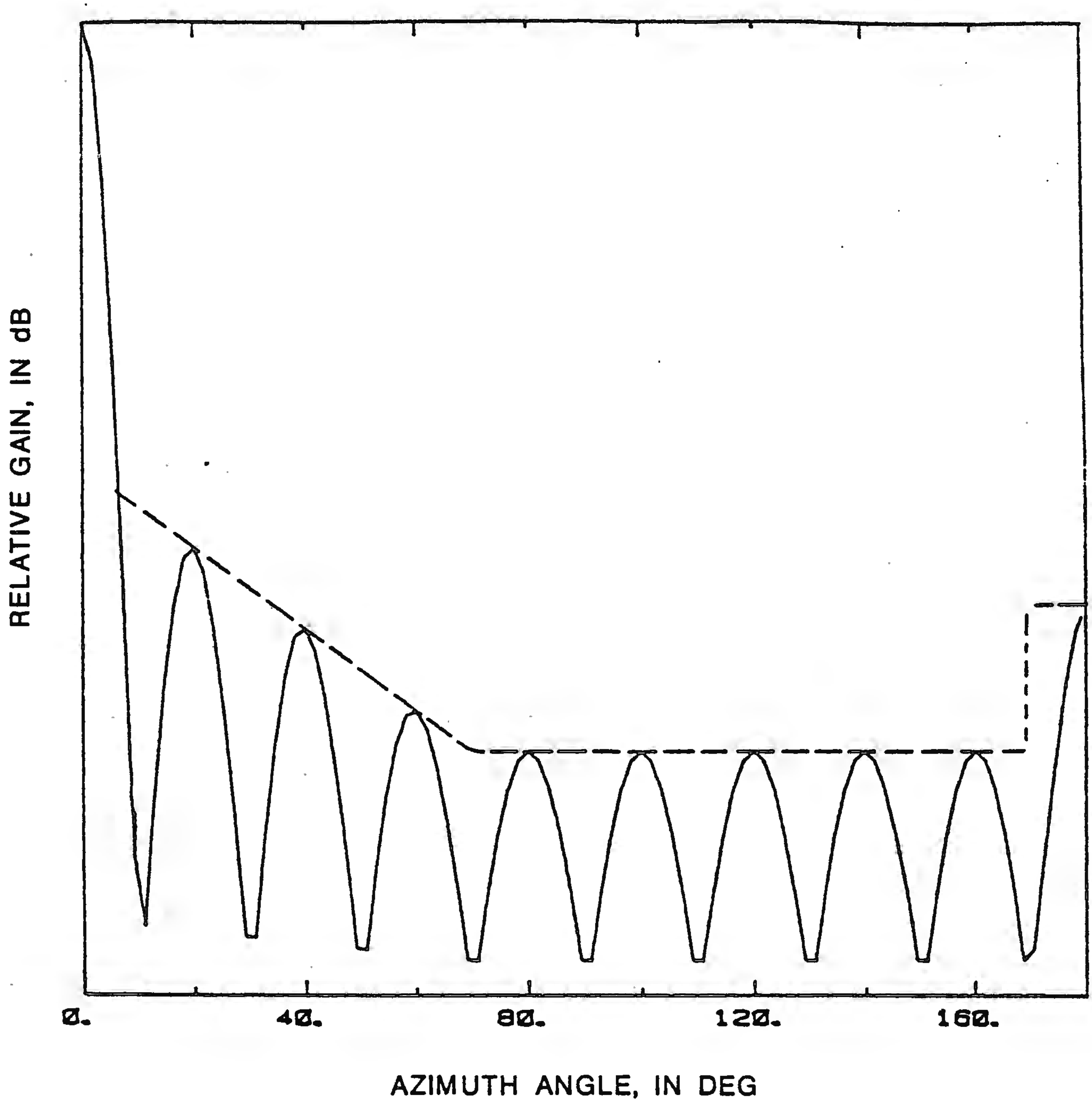


Figure C-2. Directional pattern of typical directional antenna, showing mainbeam, sidelobes, and backlobe. The dashed line is the envelope of the lobes.

For the present study, the only information available about the antennas was the mainbeam gain and the antenna type, so further simplification was necessary. First, the envelope of the antenna pattern (as illustrated by the dashed line in Figure C-2 was approximated by three sectors--the mainbeam sector, the sidelobe sector, and the backlobe sector. Within each sector, the power gain was assumed to be constant. The angular extent of each sector and the relative power gain of the side- and backlobes were determined from the mainbeam gain as described next.

Equation 6 in Annex I of CCIR Report 614 [CCIR, 1982b] can be rewritten as:

$$G(0) = 7.27 + 10 \log (D/\lambda)^2, \quad \text{(dB)} \tag{C-4}$$

where D/λ is the aperture size in wavelengths and $G(0)$ is the mainbeam gain. From the same annex,

$$\phi_o = 70/(D/\lambda)^2, \quad \text{(dB)} \tag{C-5}$$

where ϕ_o is the beamwidth of the mainlobe.

Solving Equation C-4 for D/λ and substituting this value into Equation C-5 yields:

$$\phi_o = 170.3\ (10^{-G(0)/20}). \tag{C-6}$$

Parameters for the side- and backlobes of antennas in the 7/8 GHz band were derived from a number of measured radiation pattern envelopes obtained from a commercial antenna supplier. This is because the TSEF computer model was initially tested using systems in the 7/8 GHz band. For application to other bands, appropriate antenna parameters must be determined. The patterns were for co- and cross-polarization coupling of vertically and horizontally polarized antennas. Horizontal copolarization was used for this study.

Aperture sizes ranged from 1.2 m (4 ft) to 4.6 m (15 ft). Figure C-3 shows the patterns used.

For these antennas, the backlobe was 20 degrees wide and had a gain of -8 dBi. The sidelobe sector is then 360 degrees minus the main beam and back lobe sectors.

The power level of the envelope of the sidelobes varies over the sidelobe sector, as shown in Figure C-3. For this study, the critical value is the area denied by the sidelobe sector. The power level computed for the sidelobe sector was the one that would deny the same area denied by the true envelope assuming free-space propagation loss. Free-space propagation loss (L) is given by:

$$L = C + 10 \log d^2, \quad \text{(dB)} \tag{C-7}$$

where C depends on frequency and the units of d. Its value is not necessary for this analysis. The boundary of the interference area is at the distance that $P_R = P_i$ in Equation C-3 when $G_R(b) = 0$, and FDR = 0. The value that is being sought is G(2), the gain in the sidelobe region, so G(2) is substituted for $G_T(a)$ in Equation C-3. Making these substitutions, and using Equation C-7:

$$P_i = P_T + G(2) - C - 10 \log (d^2) \tag{C-8}$$

or

$$\log d^2 = 0.1(P_T - P_i - C) + G(2)/10 \tag{C-9}$$

Taking the antilog yields:

$$d^2 = B(10^{G(2)/10}) \tag{C-10}$$

where B is the antilog of $0.1(P_T - P_i - C)$

The area of a sector of a circle is:

$$A(s) = (s/360)\ \pi d^2 \tag{C-11}$$

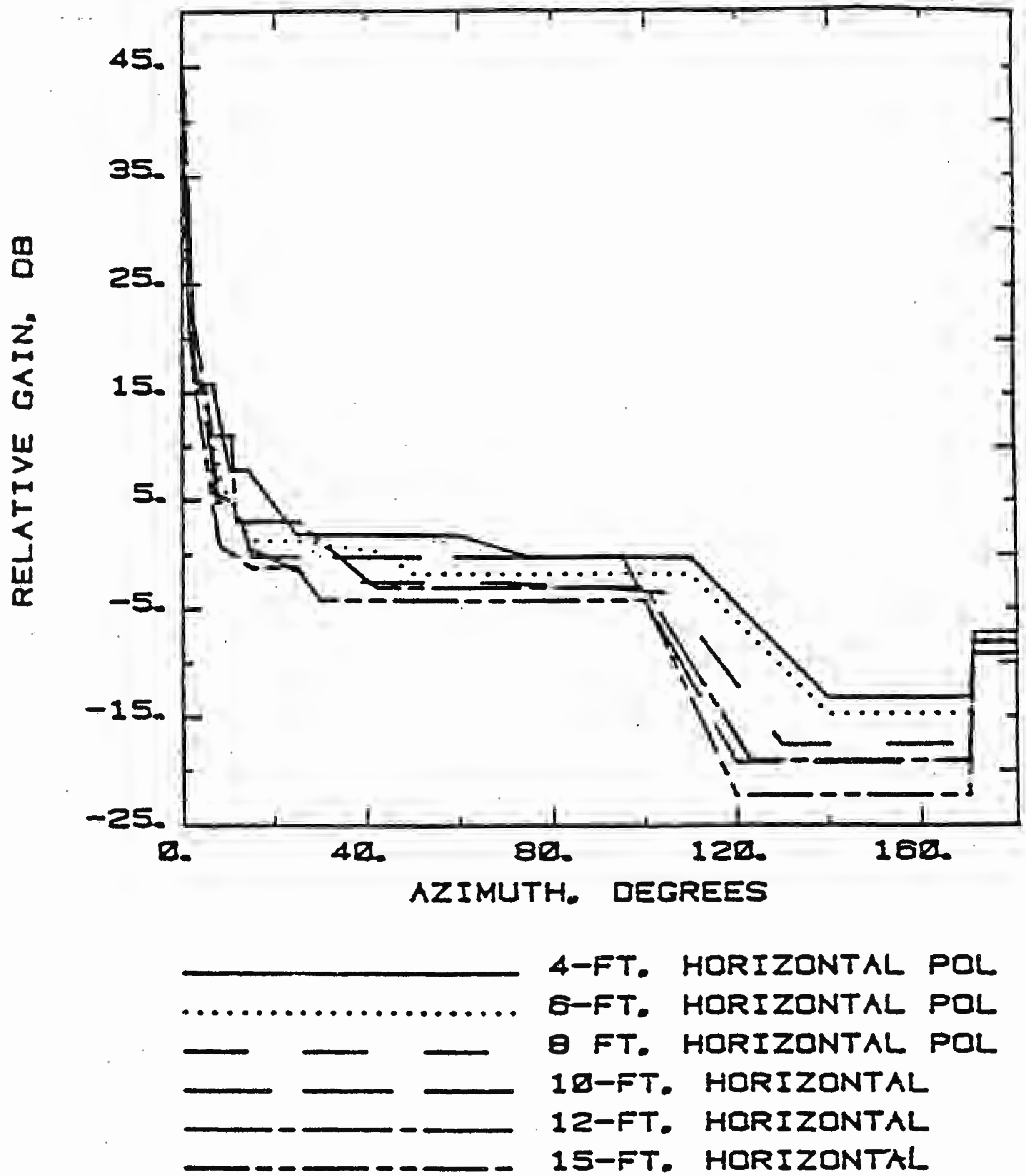


Figure C-3. Actual envelopes of antenna patterns modeled for the calculations in this report.

where s is the sector angle in degrees, and d is the radius of the circle. The area A(s) is computed for the sidelobe envelope of an antenna by numerial integration, so it is known. Then Equation C-10 is substituted into Equation C-11 to get:

$$A(s) = K(10^{G(2)/10}) \qquad \text{(C-12)}$$

where $K = B\pi(s/360)$. Finally, solving Equation C-12 for G(2) yields:

$$G(2) = 10 \log [A(s)/K] \qquad \text{(C-13)}$$

Equation C-13 was evaluated numerically for measured antenna patterns for antenna diameters from 1.2 m to 4.6 m (4 to 15 feet) in the 7/8 GHz band (shown in Figure C-3). It was found that G(2) could be approximated by:

$$G(2) = 0.45\ G(0) - 12, \quad \text{(dB)} \qquad \text{(C-14)}$$

where G(0) is the mainbeam gain. The error in Equation C-14 is less than 0.5 dB.

Modeling Propagation Loss

Most terrestrial fixed systems at frequencies above 947 MHz are composed of LOS links. Station locations and antenna heights are carefully engineered using local site surveys to ensure that there is first Fresnel Zone clearance of the radio horizon (often the highest terrain feature) between transmitter and receiver. Although there are digital data bases that provide terrain elevations at a grid of points for the U.S. The resolution (grid spacing) and accuracy of the elevation data are too coarse for complete engineering of LOS links. It is unlikely that the highest local terrain feature coincides with a grid point, and the error in the terrain height is large compared to a Fresnel Zone radius. So, digital terrain data bases do not define LOS path profiles with sufficient accuracy for the TSEF calculation.

Because digital terrain data are not sufficiently accurate, a propagation loss model that uses the terrain data is not indicated. In addition, the computer time required for such models to make a large number of calculations might be excessive. It was, therefore, decided to use a simple, fast propagation loss model that computes the loss over a smooth, spherical earth. The model selected was the Integrated Propagation System (IPS), [Frazier, 1963].

Experience shows that for microwave links, free-space transmission loss is accurate for distances short enough that the first Fresnel Zone is clear of obstacles. So, for these calculations, free-space transmission loss was used for distances from the transmitter out to a distance at which the first Fresnel Zone was first obstructed. At greater distances, the larger of free-space loss and the loss computed for propagation over a smooth spherical earth was used. This procedure produces a smooth, monotonic loss curve without lobing, as illustrated in Figure C-4. The solid line is the value used, the dashed line shows the transmission loss for a smooth curved earth, including lobing, and the dotted line shows free-space transmission loss.

Some aspects of local terrain must be included in the calculations, because engineers take advantage of them. In particular, antennas are likely to be located on local high spots to increase the possible distance between stations in the network. Site elevations and antenna tower heights combine to raise the antenna to the necessary height, and both data are available in the GMF. A procedure was devised to take advantage of this information while using a smooth-earth propagation model.

The procedure was based on the assumption that total antenna heights (site elevation plus tower height) were engineered so that the horizon of any link was just cleared by the first Fresnel Zone. This assumption is reasonable because lower total heights would result in unacceptable link performance, and because higher antenna towers would be unacceptably expensive. Using the site elevations and antenna heights from the station file and the path length, equivalent antenna heights with the same height differential were computed for a smooth spherical earth. That is, heights that would provide first Fresnel Zone clearance over a smooth earth, while maintaining the real height difference between the transmitting and receiving antennas, were calculated and used as the equivalent heights. Although the earth's curvature might not have the same shape as the actual terrain obstacle

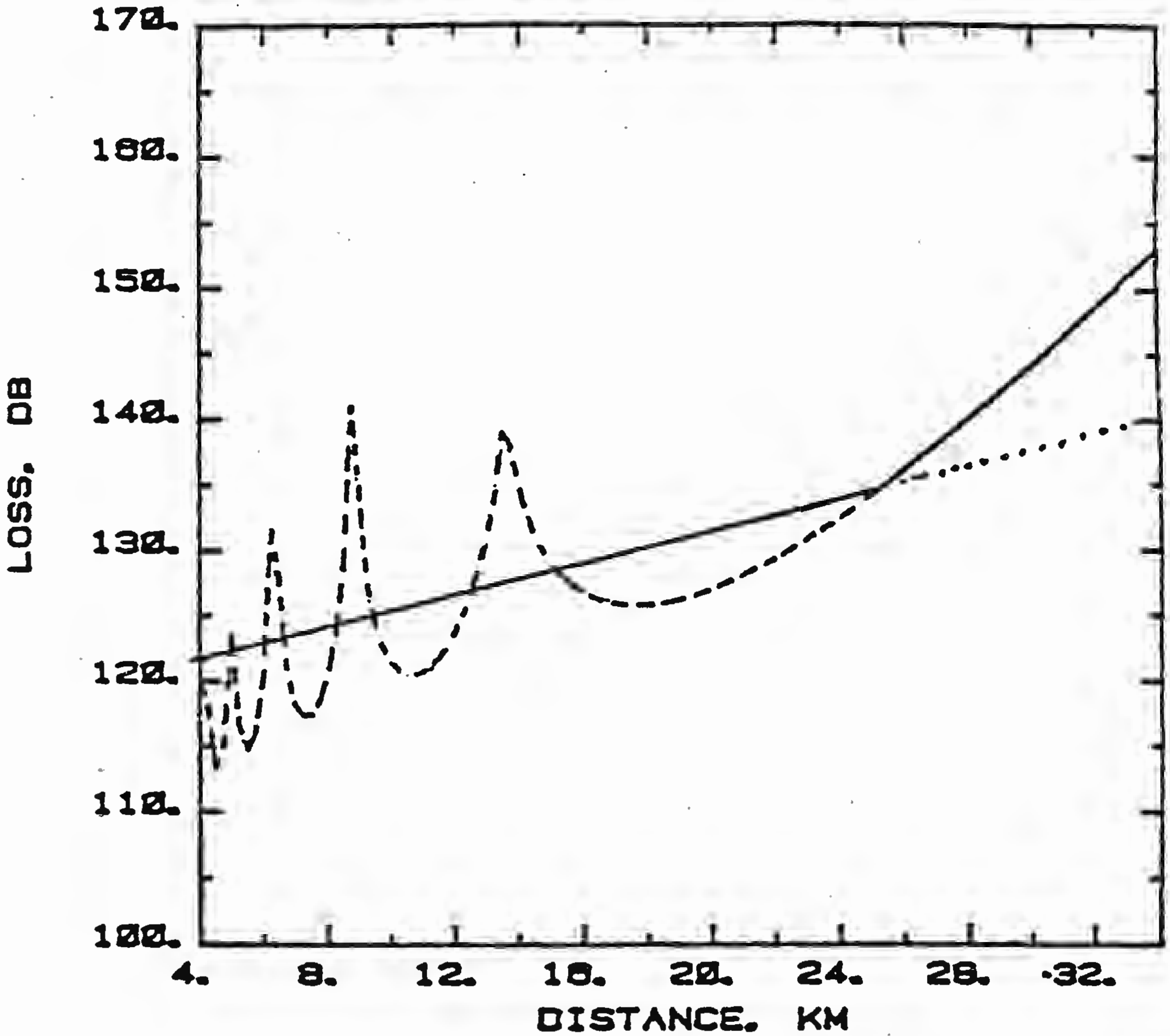


Figure C-4. Transmission loss as a function of distance over a smooth curved earth at 7 GHz. The dashed line is the loss over a smooth curved earth; the dotted line is free-space loss; and the solid line is the loss used in the calculations discussed in the text.

in the path, transmission loss increases so rapidly beyond the horizon in this frequency range that the effective interference computed this way would be very similar to the actual interference.

Computing The Sector Areas

The antenna model, the transmission loss model, and the input parameters P_i and H_i are used to compute the area denied within the emission channel of the station. If information about out-of-channel emissions and unwanted receiver responses is available, the area denied by these features may also be computed.

First, assuming that $P_R = P_i$ and that $G_R(b) = 0$, Equation C-3 is solved for L:

$$L = L(H_T, H_i, d) = P_T - P_i + G_T(a) - FDR, \qquad \text{(C-15)}$$

where

H_T = the transmitter antenna height

H_i = the (input) height of the interference boundary (the height of the antenna of the victim probe receiver),

d = the distance to the boundary

FDR = 0 for in-channel denial.

$G_T(a)$ is found for each sector of the antenna model and the transmission loss model is inverted to find the distance (d) at which the loss equals the right side of Equation C-15. The area of a sector is then found using Equation C-11 where the width of the sector angle is provided by the antenna model. The total area denied is the sum of the areas in the three sectors.

LINKS CONTAINING REFLECTORS

Many fixed-service links in the government bands contain passive reflectors. These reflectors may be in the nearfield of the primary antenna (as in periscope antennas), or in the farfield to redirect the energy when a direct path between nodes is not possible. These reflectors may scatter some of the incident energy causing local interference, or they may redirect interference (perhaps from airborne sources) into the victim receiver. If they do either of these things, they contribute to the denial area of a system, and hence, lower the TSEF. Ideally, these effects should be included in the TSEF calculation.

However, the characteristics of reflectors can be computed only if their dimensions and curvature are known. (Even if these data are known, the calculation is difficult, and not well established.) Computer-readable data files such as the GMF do not include the dimensions of reflectors used in the government fixed service, so such calculations were not possible for the model described in this report. Instead, it was agreed that passive reflectors would be replaced by a 1.2 m (4-ft) dish for the purposes of computing the area denied by a link. No theoretical justification for the substitution was developed, but the substitution probably biases the calculation in the correct direction.

SENSITIVITY OF THE TSEF TO SELECTED INPUT PARAMETERS

The area denied by a transmitter depends, in principle, on two input parameters determined by the user: the interference power (P_i) and the antenna height of the victim receiver (H_i). To test the sensitivity of the TSEF to the values selected, they were varied over a plausible range of values for selected cases. A typical result is shown in Figure C-5, where TSEF is plotted as a function of P_i for various heights (H_i), and the other system characteristics shown in the figure caption.

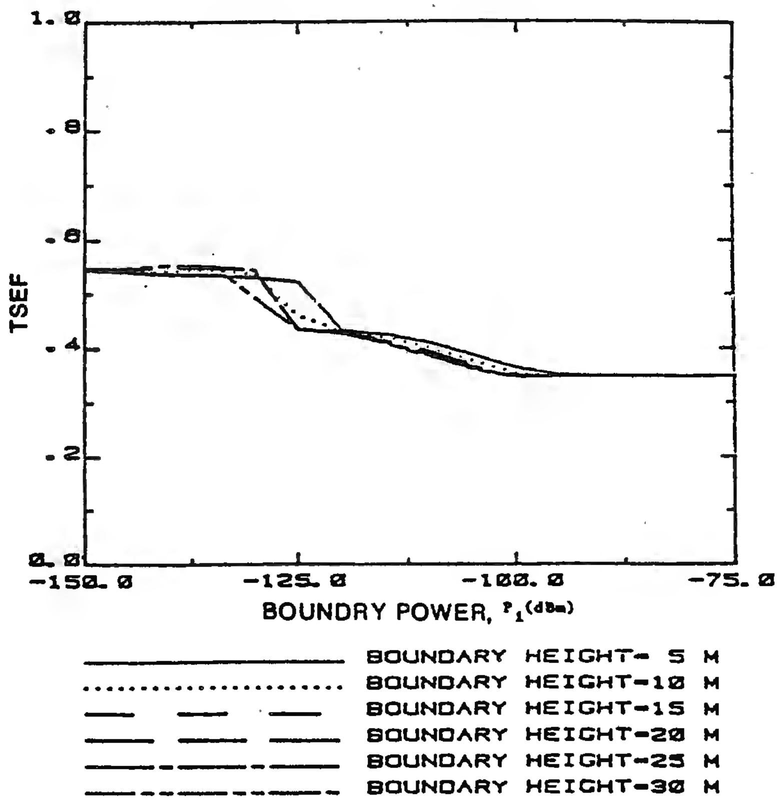


Figure C-5. The sensitivity of the TSEF of the selected value of power at the boundary for various heights of the boundary of the denied area. For this case, the reference antenna is a 12-ft dish and the system antenna is a 4-ft dish. The transmitter antenna height is 10 m; its power is 1 W.

The curves in Figure C-5 are characterized by two plateaus and a transition region between them. Inspection of intermediate printout from the computer program shows that the area denied for the small P_i on the left is almost entirely in the sidelobe sector of the transmitter antenna. The area denied for the large values of P_i on the right is almost entirely in the mainbeam of the antenna. In the transition region (intermediate values of P_i), both sectors are contributing, but their relative importance is changing. The height of the interference boundary (H_i) affects the location of the transition region, but the effect is small.

Changing the transmitter power does not affect the shape of the curves, but merely shifts them left or right, on Figure C-5, to a distance corresponding to the change in transmitter power. Increasing the diameter of the system antenna primarily raises the curves (higher TSEF), changes the shape and location of the transition region a little, and changes the relative heights of the plateaus.

CONCLUDING REMARKS

The analytial basis for the model used for the calculations in the main body of this report is given in this appendix. The analysis admits of more complexity and completeness than was implemented in the computer model, because the GMF and other computer-readable data bases do not contain all the information needed to compute all terms in the analytical model.

APPENDIX D[a]

BAND EFFICIENCY

by

Stanley I. Cohn

GENERAL

When a number of systems operate within a band in a specific geographic area, the area denied by one system may overlap the area denied by another system. These common denial areas will increase the efficiency of the band use over that which would be determined from considering the area denied by individual systems.

For instance, when a number of UHF-TV channels are operating in a given market, the denied area for a taboo channel resulting from one operating station, may overlap a denied area caused by another operating station. Also, for operations involving high and low channels certain taboos are not in the UHF-TV band. Because of these factors, the efficiency of the band would be higher than for a single operating channel. The introduction of low-power stations (LPTV) at UHF would also tend to increase the efficiency of band use since the taboos are smaller for these stations and they would "fill in" some areas in which high-power stations are precluded. The present use of LMR systems in the UHF-TV band on Channels 14 to 20 in some geographic areas increases the efficiency of band usage for similar reasons. Decreases in efficiency can occur if larger-than-taboo distances are used in an area.

In similar manner, the band efficiency in the LMR VHF high band would also differ from that of a single system because of the common denial areas, locations of stations, powers and antennas heights involved.

To illustrate the effect of common denial areas for broadcasting consider the following example. For the purpose of this example, assume that

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a. This work was done under Purchase Order No. 43 SANT503717, Methodology for Determining Spectrum Efficiency, Sachs/Freeman Assocs. Inc.

UHF channels, 30, 36, 42, 48, and 54 have co-located antennas in a city. Based on part 73.610 of the FCC rules, certain distance separations must be maintained between these stations and other stations operating on various channels. Table D-1 shows the denied channels, the reasons for denial, the distance separation required and the area corresponding to the distance separation. In some cases, a channel is denied for several reasons. For these cases the largest distance separation is used since it encompasses the maximum denial due to the various reasons. In this manner, the area denied for a specific channel is only considered once (at maximum area for any other interference conditions) when overlap of denial areas occurs. Note that a denial area for Channel 38 is not included since this channel is now used for radio astronomy and no TV transmitter or receiver emissions are present in this channel from the existing station.

Assuming a reference system with a 10 dB front-to-back receiving antenna gain at fringe areas, no other than cochannel restrictions and a 1,000 transmitting antenna height; cochannel separation distance would be approximately 100 miles which is an area of 31,416 square miles denied. Then

$$\text{Band Eff.} = \frac{6 \times T \times 31{,}416 \times 5}{6 \times T \times 911{,}770} = .172$$

The band efficiency is somewhat greater than the efficiency of a single station as shown earlier. This is due to the overlap of certain denial areas as mentioned earlier.

To illustrate the effect of common denial areas for LMR, consider the following example. For the purpose this example assume three VHF LMR co-located stations are assumed to be A = f, B = f + 30 kHz and C = f + 60 kHz. The frequencies of the two allocated stations at 35 miles distance are D = f + 15 kHz and E = f + 45 kHz. The coordination requirements of Part 90 of the FCC Rules and Regulations (75 miles cochannel and 35 miles for 15 kHz adjacent channels) are satisfied by this arrangement.

The two center adjacent channels at the three co-located stations (f + 15 kHz and f = 45 kHz) are denied for a radius of 35 miles by each of two

TABLE D-1

DENIAL DUE TO
CHANNELS 30, 36, 42, 48, & 54 ON SAME TOWER

(Page 1 of 2)

CHANNEL DENIED	REASON FOR DENIAL Type of Inteference With Channel:*					MAX DIST DENIED (Miles)	AREA DENIED (Sq. Miles)
	30	36	42	48	54		
15	P					75	17,671
16	S					60	11,310
21		P				75	17,671
22	IF	S				60	11,310
23	LO					60	11,310
25	I					20	1,257
26	I					20	1,257
27	I		P			75	17,671
28	I	IF	S			60	11,310
29	A	LO				60	11,310
30	C					175	96,211
31	A	I				55	9,503
32	I	I				20	1,257
33	I	I		P		75	17,671
34	I	I	IF	S		60	11,310
35	I	A	LO			60	11,310
36		C				175	96,211
37	LO	A	I			60	11,310
38**	IF	I	I			0	0
39		I	I		P	75	17,671
40		I	I	IF	S	60	11,310
41		I	A	LO		60	11,310
42			C			175	96,211
43		LO	A	I		60	11,310
44	S	IF	I	I		60	11,310
45	P		I	I		75	17,671

TABLE D-1

DENIAL DUE TO
CHANNELS 30, 36, 42, 48, & 54 ON SAME TOWER

(Page 2 of 2)

CHANNEL DENIED	REASON FOR DENIAL Type of Interference With channel:*					MAX DIST DENIED (Miles)	AREA DENIED (Sq. Miles)
	30	36	42	48	54		
46			I	I	IF	20	1,257
47			I	A	LO	60	11,310
48				C		175	96,211
49			LO	A	I	60	11,310
50		S	IF	I	I	60	11,310
51		P		I	I	60	11,310
52				I	I	20	1,257
53				I	A	55	9,503
54					CO	175	96,211
55				LO	A	60	11,310
56			S	IF	I	60	11,310
57			P		I	75	17,671
58					I	20	1,257
59					I	20	1,257
61					LO	60	11,310
62				S	IF	60	11,310
63				P		75	17,671
68					P	75	17,671
69					S	60	11,310
							TOTAL 911,770

* C = Cochannel, A = Adjacent Channel, I = Intermodulation
IF = IF Beat, LO = Local Oscillator, P = Picture Image, & S = Sound Image.

** Channel 38 is Denied Because of Radio Astronomy.

stations (A and B have a common adjacent channel denial area for f + 15 kHz, while B and C have a common denial area at f + 30 kHz. Further, the cochannel denial area for D completely encompasses the adjacent channel denial area of A and B. A similar situation exists for E with regard to B and C and for B, with regard to D and E. The cohannel denial for A encompasses the lower adjacent channel denial area of D, while the cochannel denial area of C completely encompasses the upper adjacent channel denial area of E.

As a result of these common denial areas the only excess spectrum denied is that due to the lower adjacent channel of A (f - 15 kHz) and the upper adjacent channel of C (f = 75 kHz). As indicated in the LMR example in Section II, the cochannel denial area is 17,671 square miles for each channel, while the adjacent channel denial area is 3,858 square miles for each adjacent channel.

In this specific example of LMR band efficiency there are five cochannel denial areas and only two adjacent channel denial areas. Thus the band efficiency would be:

$$\text{Band Eff.} = \frac{15 \times T \times 5 \times 17{,}671}{15 \times T \times (5 \times 17{,}671 + 2 \times 3848)} = .920$$

The band efficiency for LMR systems is therefore, significantly increased over that of a single system due to common denial areas.

COMPUTATION OF DENIED AREAS

In the previously discussed examples, omni directional antennas were used and when common denial areas were involved the denial area of one system encompassed that of another system. In computation of band efficiency, such situation will not always occur. Common and irregular denial areas in such general situations can occur frequently, since a large number of systems and considerable possibilities of having overlapping denial areas will exist. Since the denial area for a band would be computed for each frequency increment for both existing and reference systems, the computation can be quite lengthy.

The use of a matrix representation of areas is recommended to simplify the computation process. This would involve subdividing the area under consideration into square sub-areas of sufficiently small size to provide for reasonable approximation to denial area. In the case of omni directional systems in order to simplify computations, a square of area equal to the circular area could be used as an approximation as shown in Figure D-1. Similarly, Figure D-2 shows the two-level antenna pattern approximation for a directional antenna. A "1" in each of the small squares indicates denial to another system.

In order to determine the denial area for a particular frequency increment (or channel), first determine the denial matrix with the system placed at the center of the area. Then move the center of the matrix to the coordinates at which that system is located. Figure D-3 shows this process for a system located at coordinates (m, n). The relative locations of each matrix element are then changed by adding (m, n) to the element locations of the matrix at the center of the area.

This step is repeated for each system that can produce area denial for the frequency increment under consideration ad the numbers appearing in each element are added.

Figure D-4 shows the results of this process for three omni directional systems, while Figure D-5 depicts the denial area for two directional systems. The numbers in the elements indicated the number of systems causing denial to the element. To determine the denied area, the number of elements with a value of "1" or greater are added and multiplied by the area of the element. For example, in Figure D-4 there are 216 elements which are denied. If each element represents one square mile that total area denied would be 216 square miles. Similarly, in Figure D-5, 82 square miles are denied.

After determining the area denied for each frequency increment, the results are added to produce the total denial in term of bandwidth - area product, or:

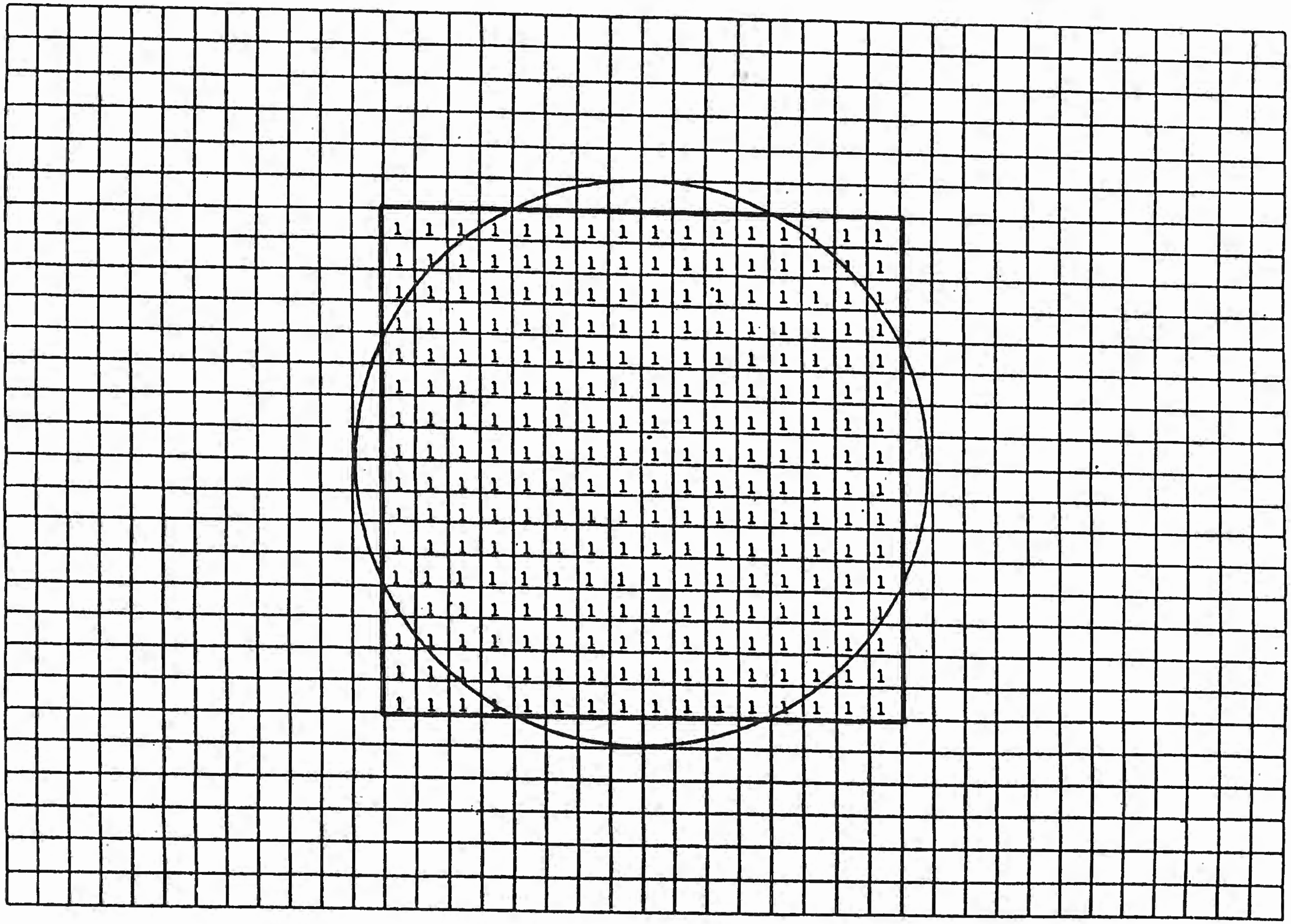
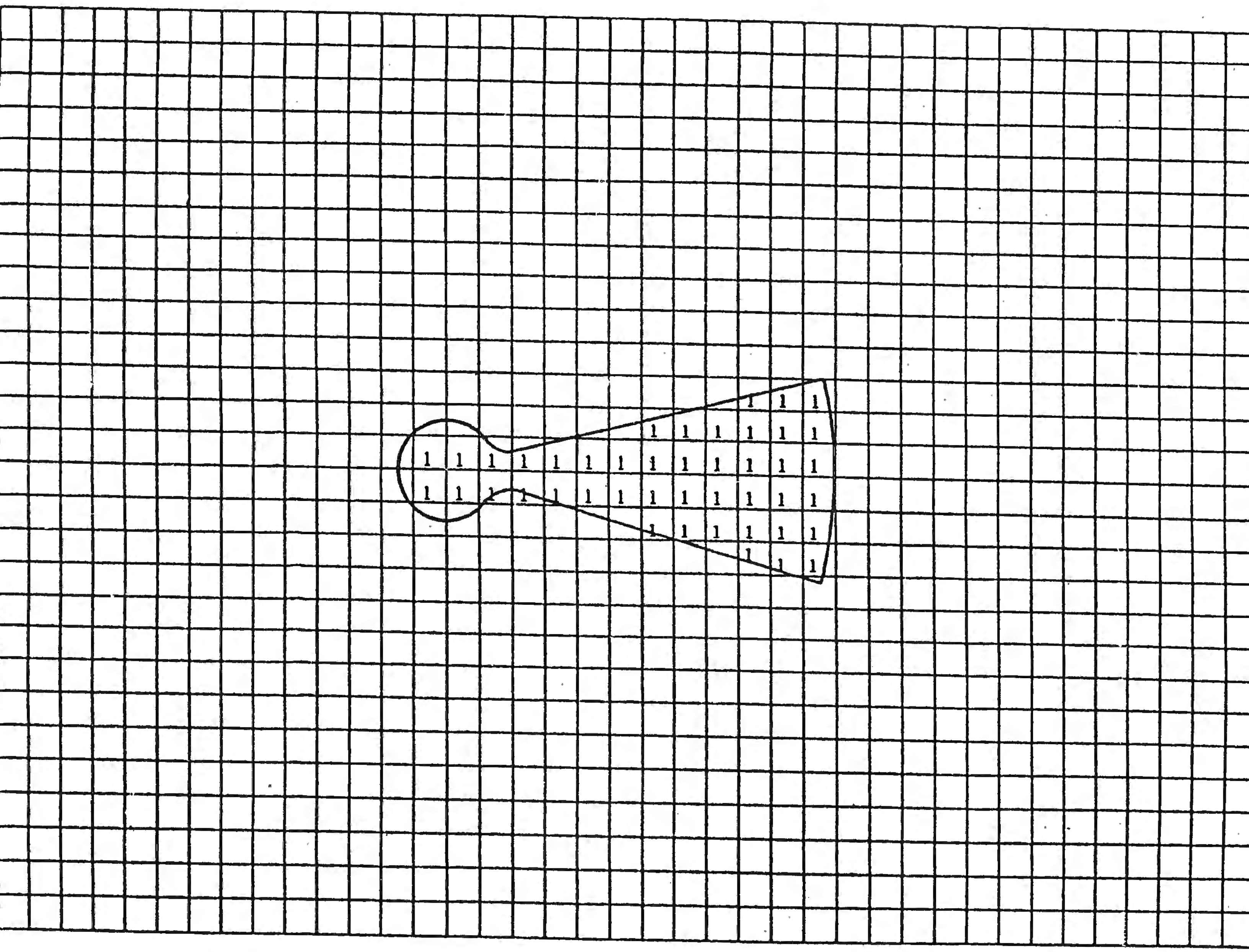


Figure D-1. Square approximation to OMNI directional denial area.



gure D-2. Two-level directional antenna approximation.

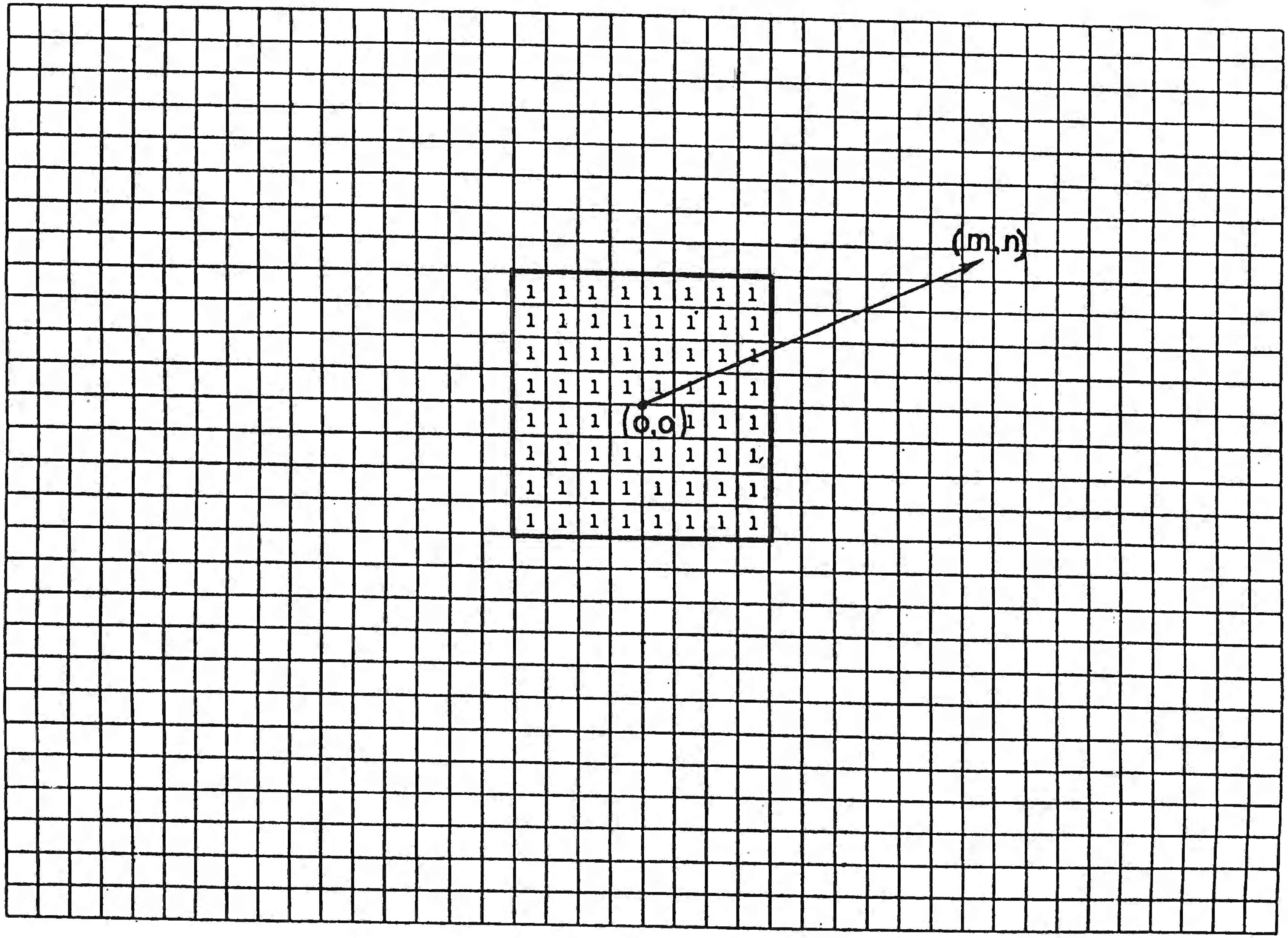


Figure D-3. Denial matrix for a system at (m, n).

$$\text{Total Systems Denial} = D_s = \sum_{i=1}^{n} \Delta f_i \; S_{si} \qquad \text{(D-1)}$$

If Δfrepresents a uniform increment then:

$$D_s = \sum_{i=1}^{n} \Delta f \cdot \sum_{i=1}^{n} S_{si} \qquad \text{(D-2)}$$

$$D_s = B_T \cdot \sum_{i=1}^{n} S_{si} \qquad \text{(D-3)}$$

where B_T is the width of the total band under consideration.

The entire process is then repeated with reference systems substituted for the actual systems at each location for each frequency increment. The results of this are:

$$\text{Total Reference Denial} = D_r = B_r \sum_{i=1}^{n} \quad S_{ri} \qquad \text{(D-4)}$$

The band efficiency is then determined by dividing D_r by D_s or:

$$\text{Band Eff.} = \frac{D_r}{D_s} = \frac{\sum_{i=1}^{n} S_{ri}}{\sum_{i=1}^{n} S_{si}} \qquad \text{(D-5)}$$

TIME DENIAL

The previous discussions have considered that all systems operate at all times. This may not be the case for practical operational systems. The usage time can be factored into the previously discussed calculations. However, it is important to note that the time factor may not be represented by time that the transmitter is emitting energy or by the time a desired signal is detected by a receiver.

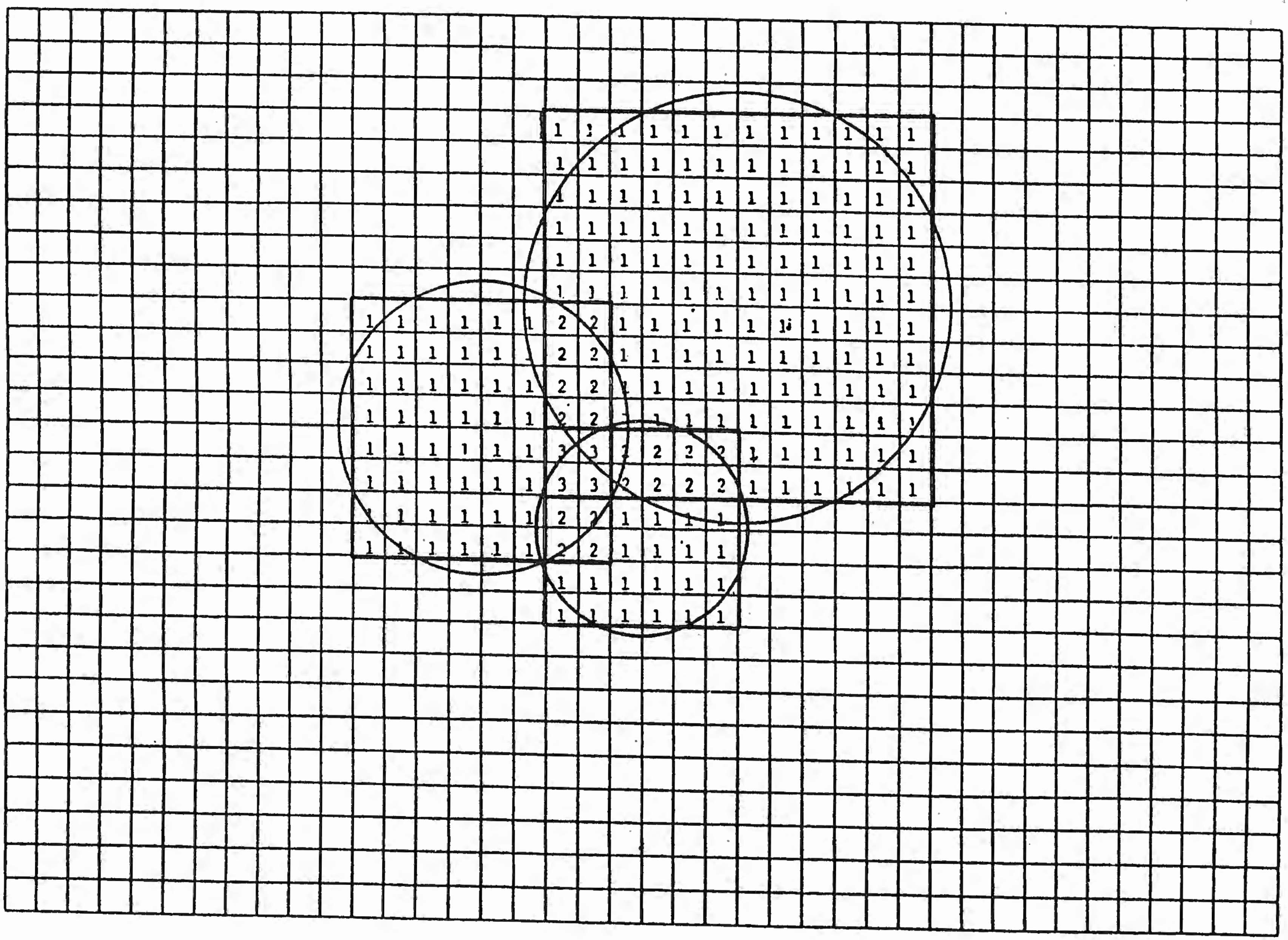


Figure D-4. Denial area for three OMNI directional system.

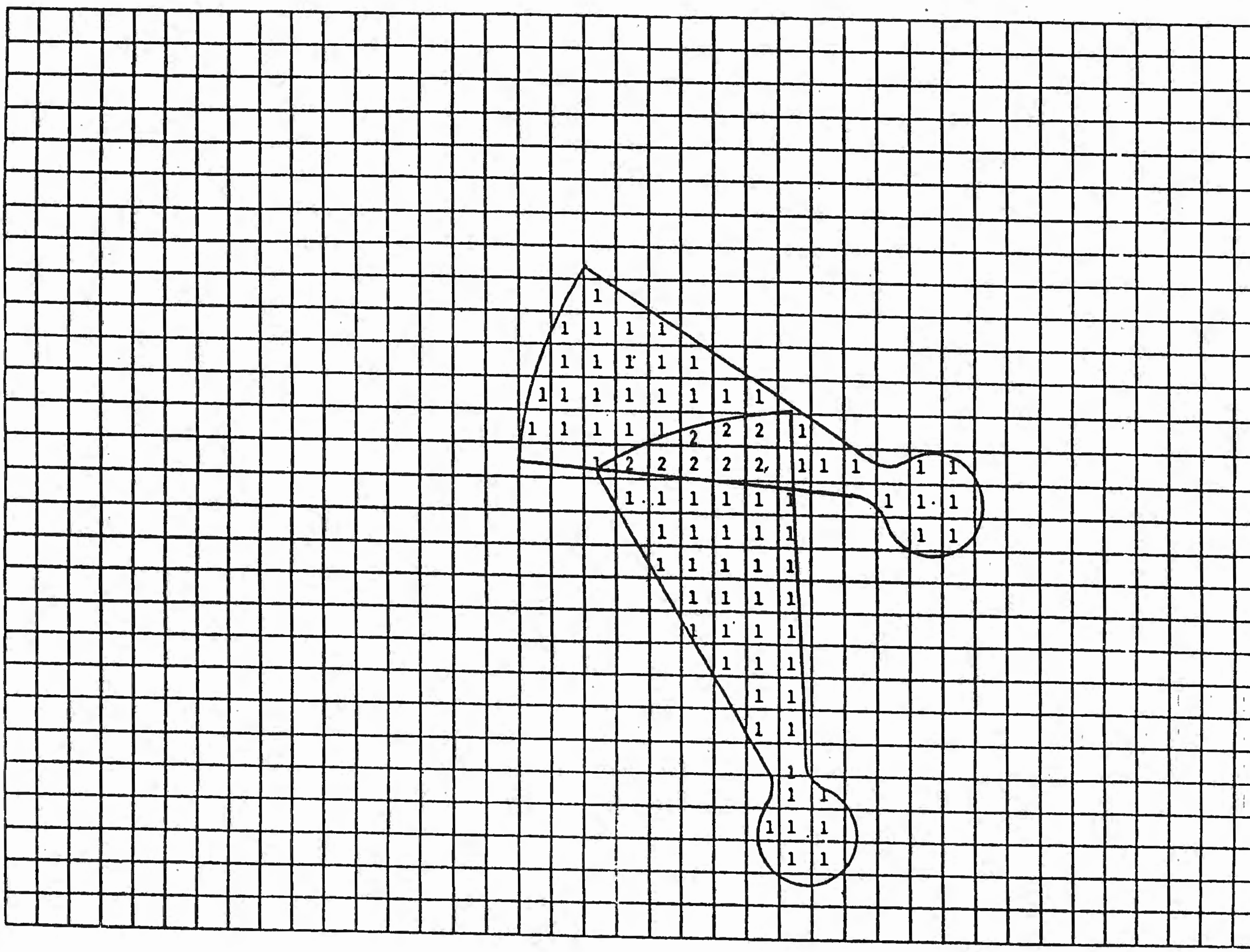


Figure D-5. Denial area for two directional system.

For example, an emergency communication system may have on-going transmissions for only a small portion of the time, but someone may be listening at all times. Under such conditions the system can be considered to be used at all times. A radio astronomy system does not transmit, but its receiver may be on at all times. In certain situations, such as land mobile radio, a system is designed to account for peak traffic in a "busy hour" and, if fully loaded during the "busy hour", cannot share with other systems. Sharing may be possible at times other than the busy hour. These factors must, therefore, be considered in determining the time factor and they will differ from system to system.

Where systems are not considered to be in operation at all times a time factor can be associated with the denial area at a specific frequency increment. For example, if a non-emergency LMR system has a maximum saturation loading of 100 mobile units and is only using 40 units, the time factor would be 0.4. Under such conditions, time sharing with another system is possible.

To account for this in the previously discussed matrix addition process, each matrix element for each frequency increment would be assigned a time factor of denial rather than complete denial. For example, if a system is being used 40% of the time a factor of 0.4, instead of "1" would be assigned for the cochannel denial matrix elements and for adjacent channel matrix elements associated with that system. Another system on an adjacent frequency might have a time factor of 0.3. In common denial areas, the matrix elements would be assigned a value of 0.3 + 0.4 = 0.7 instead of "1".

In determining the denial for a given frequency increment, the values in the elements and it would then be totaled, with values less than "1" kept and values greater than "1" being assigned as "1" (fully loaded). To illustrate, if each of the systems shown in Figure D-5 had a time factor of 0.6, the elements with "1" values would be replaced by elements of value 0.6 and the "2" values replaced by "1". The total area-time product would then be 52.4, rather than 82 that would occur if each system was on at all times.

The procedure for determining band efficiency would then be the same as previously discussed except that the time factor values would be used.

CONSIDERATION OF MIXED SERVICES IN A BAND

A number of bands are allocated to more than one Primary Radio Service as well as allowing for other secondary and permitted Radio Services. In such cases, the frequency distance separations for interservice operations can be different than those for intraservice operation. The denial areas will, therefore, be different. As a consequence, the band efficiency will depend on which Radio Service the "new" system being added to the environment represents.

As an illustrative example, assume that a particular band is allocated on a primary basis to both the Mobile and Broadcasting Radio Services. As indicated previously, the cochannel transmitter separation for UHF TV Broadcasting is 175 miles and for LMR is 75 miles. Assume, for the purpose of this example, that the transmitter separations between Broadcast and LMR transmitters is 100 miles. In the area of concern, only Broadcasting is presently used. It is clear that the area denied by existing systems will be different if one considers the denial to LMR systems than it would be if they consider the denial to Broadcast systems.

Thus, the band efficiency, when mixed services are involved, must be specified in terms of the service which is being introduced. It should be noted that the results of band efficiency studies, when comparing the introduction of various services in a mixed band, can provide very useful information for spectrum planning purposes concerning future flexibility of the band being investigated.

The matrix addition procedure for determining band efficiency is readily adaptable to a mixed service situation. With the exception of specifying the type of service being introduced, the procedure would be exactly the same for mixed service bands as it would for single-service bands.

A DEFINITION OF BAND EFFICIENCY FACTOR

Based on the above discussions, a general definition of a Band Efficiency Factor can be developed. Consider that a number of existing systems tuned to various frequencies, at various locations, with varying time occupany and representing various types of Radio Services can exist in a band for a defined geographic area. Each existing system uses a certain bandwidth x spatial volume x time to the placement of a new system. The Band Efficiency Factor (BEF) would be:

$$BEF = \frac{\text{(spectrum resource used by all reference systems)}}{\text{(spectrum resource used by all existing systems)}} = \frac{SRU_r}{SRU_s} \qquad \text{(D-6)}$$

or

$$BEF = \frac{\Sigma B_{ri} S_{ri} T_{ri}}{\Sigma B_{si} S_{si} T_{si}} \qquad \text{(D-7)}$$

where the summation in the denominator covers all existing systems and the summation in the numerator covers all reference systems used to replace existing systems.

Since two or more systems may have overlapping denial of spectrum resource to a new system, it is necessary to subdivide the spectrum into bandwidth x physical space x time increments. The "kth" increment of spectrum resource used would be:

$$SRU_k = \Delta f_k \cdot \Delta S_k \cdot \Delta t_k \qquad \text{(D-8)}$$

The spectrum resources used by all systems would be:

$$SRU = \Sigma\, SRU_k = \Sigma\, \Delta f_k \cdot \Delta S_k \cdot \Delta t_k \qquad \text{(D-9)}$$

where the summation is taken over the total number of increments.

If the total width of the band being examined is B_T and the frequency increment $\Delta f$is uniform and equal to or less than the channel width of the system with the smallest channel width, then:

$$B_T = \sum_{i=1}^{m} \Delta f_{ki} \qquad \text{(D-10)}$$

where m is the total number of frequency increments used.

Equation (D-9) then becomes

$$SRU = B_T \sum_{j=1}^{n} \Delta S_j \, \Delta t_j \qquad \text{(D-11)}$$

where n equals the total number of spatial x time increments used.

As discussed previously, the physical space, geographic area (or volume) can be subdivided in sub-areas (or sub-volumes). Each of these would have an associated time factor obtained by adding the time factors for each system denying use of the sub-area (or sub-volume). Values of "1" or less would be directly used, while values greater than "1" (complete denial) would be replaced by "1". For each frequency increment the bandwidth x physical space x time product would then be:

$$SRU_i = \Delta f_i \sum_{j=1}^{p} \Delta S_{ji} \, \Delta t_{ji} \qquad \text{(D-12)}$$

where p equal the total number of sub-areas (or sub-volumes) in the area (or volume) being considered.

The total spectrum resource would be found by:

$$SRU = \sum_{i=1}^{mp} SRU_i = \sum_{i=1}^{m} \Delta fi \sum_{j=1}^{p} \Delta S_{ji} \, \Delta t_{ji} \qquad \text{(D-13)}$$

Substitution equation (D-10) in (D-13) results in:

$$SRU = B \sum_{i=1}^{m} \sum_{j=1}^{p} \Delta S_{ji} \, \Delta t_{ji} \tag{D-14}$$

Substituting equation (D-14) in equation D-6) for the reference systems and for the existing systems respectively results in:

$$BEF = \frac{\sum_{i=1}^{M} \sum_{j=1}^{p} \Delta S_{jir} \, \Delta t_{jir}}{\sum_{i=1}^{p} \sum_{j=1}^{p} \Delta S_{jis} \, \Delta t_{jis}} \tag{D-15}$$

where m is the total number of frequency increments used and p is the total number of sub-areas (or sub-volumes).

Equation (D-15) provides a general method of determining the Band Efficiency Factor (BEF) for mixed Radio Services and for systems with varying time occupancy.

As noted previously, the BEF in a mixed Radio Service band depends on the Radio Service being introduced in the band. This can provide a very useful planning technique since it aids in determining future flexibility for use of a band by various Radio Services.

The matrix addition method is discussed previously is well suited to computer solution of Equation (D-15). For terrestrial systems, where only the area is of interest, a two-dimensional matrix would be used. In other systems, where physical volume is of interest, a three dimensional matrix would be used. In the case of the geostationary orbit, the calculation reduces to a one dimensional matrix; the orbital spacing.

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15. ABSTRACT *(A 200-word or less factual summary of most significant information. If document includes a significant bibliography or literature survey, mention it here.)*

The Technical Spectrum Efficiency Factor (TSEF) concept was developed in the Technical Subcommittee (TSC) of the Interdepartment Radio Advisory Committee (IRAC) and was applied, by NTIA, to government fixed telecommunications systems in the 7 - 8 GHz bands. Both computer automated and manual procedures were employed.

The Technical Spectrum Efficiency Factor (TSEF) evaluates the technical spectrum efficiency of a spectrum-using system by comparing the amount of the spectrum it uses to that used by the most spectrum-efficient system that could be procured to accomplish the same mission. The "most spectrum-efficient" system is called "the reference system." An extensive catalog of spectrum conservation techniques is also included.

16 Key Words *(Alphabetical order, separated by semicolons)*

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