TELECOMMUNICATIONS INDUSTRY ASSOCIATION

A REPORT ON TECHNOLOGY INDEPENDENT METHODOLOGY FOR THE MODELING, SIMULATION AND EMPIRICAL VERIFICATION OF WIRELESS COMMUNICATIONS SYSTEM PERFORMANCE IN NOISE AND INTERFERENCE LIMITED SYSTEMS OPERATING ON FREQUENCIES BETWEEN 30 AND 1500 MHz

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TIA TR8 Working Group 8.8
Technology Compatibility

For:

TR-8 Mobile and Personal Private Radio Standards Committee
TIA Mobile and Personal Communications

In Conjunction with the:

Institute of Electrical and Electronics Engineers
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Rubinstein added updates and corrections to the Okumura/Hata/Davidson Model in Sections 5.1.2, 5.1.2.1, and 5.1.2.2.

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Figures 2, 3, and 4 were modified to fit on the page (no content was changed). Section 5.3 was modified to reflect the FCC filing requirements of 2-10 miles (3-16 km). Rubinstein added reference [20] Hufford, to Section 8.0, and included references to Hufford in 5.6. Rubinstein modified the definition of HAAT in 7.1. Section 7.2 was modified to include EDACS®, change dBu to dBμ, and correct ITU-R and ITU-T. Stone clarified the use of mean power value in 6.8.1, 6.8.4, and 6.8.5. Rubinstein and Olson modified Appendix-E to add new E.3 and E.4. New Sections 3.4.3 and 3.4.4 added by Olson, along with definitions in Section 7.1 for Tile Reliability and Tile Reliability Margin.

20 May 1997

- Version 20 created. Sections 3.6.2.3.1 and 3.6.2.3.2 replaced with revised Sections by Olson. Section 3.6.2.3.3 deleted by Olson, and Section 3.6.2.3.4 renumbered as 3.6.2.3.3. Sections 6.0, 6.2, 6.3.2, 6.4.3, 6.5.3, 6.5.4, and 6.8.1 rewritten and revised by Hofmeister. Bullets in Section 6.8.2 indented by Hofmeister. Section 5.0, 5.2, 5.2.1, 5.2.2, 5.2.2.1, 5.2.2.2, 5.2.2.3, 5.2.3, 5.2.3.1, 5.2.3.2, and 5.2.3.3 rewritten/added by Anderson. Equations 27 through 39 renumbered as Equations 46 through 58 due to the additional equations added by Anderson in Section 5.0 and subsections thereof. All occurrences of EDACS throughout the document annotated with ®. Errata sheet for Sections 5.3.1, 5.3.1.1, and 5.3.1.2 from Rubinstein for Version 19 incorporated into main body of document. An introductory sentence to Appendix-E was added by Olson on Page 2 preceeding Section 2.0. Olson modified Section 3.4.2, last sentence, to include "at a minimum". Figure 2 was replaced with a new Figure 2 by Olson - no changes were made, the graphics of the figure were upgraded. The bullets on pages 9-10 were re-ordered and modified by Olson. All occurrences of IIP^3, OIP^3, and IP^3 were changed to IIP^3, OIP^3, and IP^3. Section 5.2, fourth paragraph, first sentence was replaced by Ericsson. Section 5.2.2, second paragraph, was corrected by Rubinstein to reflect 8,451,000 metes. Section 5.2.2.1, second to last paragraph, "angles of incident" was changed to "angles of incidence" by Olson. The path numbers in Section 5.6 were modified by Olson to delete one incorrect number and delete one redundant number. Section 5.9.2.2 "mean = O" was changed to "mean = 0". Section 6.0, fourth paragraph, third sentence was deleted as agreed to by Working Group in Phoenix. Section 6.3.2, last sentence was deleted as agreed to by Working Group in Phoenix. Section 6.8.1, last sentence, the word "approximately" was inserted preceding 2 dB by Olson. Section 6.8.3, last paragraph, last sentence, the words "simulated test" were inserted by Olson preceding "reference sensitivity". Table B-3, the last sentence in the text following the paragraph was deleted by Olson and Stone. Section C.5 was modified by Olson to replace "OHD" in C.5.1 with "propagation". Section C.5.2 was deleted, and the subsequent subsections renumbered by Olson. Section C.5.3, L_3 was changed by Olson to L_2. Section C.5.4 was changed by Olson to delete $L_{3,}$ and the values of the equation changed to "136 + " to reflect this. Section C.6.4.3 was modified by Olson to include a complete square root sign in the equation.

ACKNOWLEDGMENTS

It is believed that this document represents a watershed event in the wireless communications arena. For the first time a quantitative and systematic approach to addressing the challenges posed by technology evolution in a wireless communications environment is presented.

While this work is indeed the product of many individuals technical contributions, the genesis of the document is the brainchild of Mr. Carl B. "Bernie" Olson. Mr. Olson's formative contributions are therefore rightly noted. In addition, substantive contributions have been made by: Dr. Harry Anderson of EDX Engineering; Mr. Dominic Arcuri of Ericsson; Mr. John Oblak of the E.F. Johnson Company; Mr. Brad Hiben of Motorola; Mr. Tom Rubinstein of Motorola; Mr. Casey Hill of Motorola; Mr. Al Wieczorek of Motorola; and Ms. Judith F. Furie of INS/CECOM.

A special acknowledgment is made to the United States Immigration and Naturalization Service, and in particular to Mr. David G. Butler, Chief of the Headquarters Radio System Section, for providing staff and resources that were instrumental in the formulation, editing, and production of this document.

Mr. David Brown
Dr. Gregory M. Stone
Co-Chairmen, WG 8.8

Washington, D.C.
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PREFACE

This Version 20 release is intended to support the engineering, design, and spectrum management of wireless systems operating between 30 and 1500 MHz employing analog and digital voice or integrated digital voice and data teleservices.

Version 20 is not intended for use with packet or circuit switched data only teleservices.

A future release will incorporate those parameters and procedures applicable to packet and circuit switched data only teleservices.

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1.0 TIA TR8 WG8.8 Technology Compatibility Committee Charter and Mission

TIA's Land Mobile Radio Section TR8 WG 8.8 Technology Compatibility Committee is working under a charter and mission statement to address the following technical challenges:

- Accommodating the insertion of bandwidth efficient narrowband technologies likely to be deployed as a result of the Commissions "Spectrum Refarming" efforts;

- Assessing and quantifying the impact of new narrowband/bandwidth efficient digital and analog technologies on existing analog and digital technologies;

- Assessing and quantifying the impact of existing analog and digital technologies on new narrowband/bandwidth efficient digital and analog technologies; and

- Addressing migration and spectrum management issues involved in the transition to narrowband/bandwidth efficient digital and analog technologies. This will include developing solutions to the spectrum management and frequency coordination issues resulting from the narrowbanding of existing spectrum considering: channel spacing from 30 and 25 kHz to 15, 12.5, 7.5, 6.25, and 5 kHz.

To accomplish these objectives, the WG8.8 Committee has joined forces with the Institute of Electrical and Electronic Engineers (IEEE) Vehicular Technology Society's (VTS) Propagation Committee. The IEEE Propagation Committee's contribution to this technology compatibility effort is in the area of supporting development and adoption of standard two dimensional (2D) and three dimensional (3D) electromagnetic wave propagation models, a diffraction model, and reports relating to the selection of terrain and land use data bases. This propagation related effort will be generalizable to the electromagnetic wave propagation modeling and simulation of both current and future land mobile wireless systems.

1.1 Responsiveness to User Requirements

The Committee also has been particularly responsive to specific requests from two particular user organizations: the Association of Public Safety Communications Officials, International (APCO) and the Land Mobile Communications Council (LMCC).

On the 21st of July 1995, APCO Automated Frequency Coordination Inc., requested technical assistance from the Committee in facilitating the accommodation of advanced technologies in a post refarming environment. APCO, among others, specifically requested that the Committee establish a standardized methodology for the modeling and simulation of narrowband/bandwidth efficient technologies operating in a post “refarming” environment as applicable to Spectrum Management. Subsequently, on 20 November 1995, the LMCC requested the Committee's efforts be expanded to address recommendations for a "Licensee Data Set" and for a methodology to determine "Service Areas" for existing licensees.

In response to these requests of the user community, a substantive evolution of this Committee work product has occurred. For example, Appendix-B to this document contains a recommended set of data elements for automated modeling, simulation, and spectrum management of wireless communications systems. This technical appendix addresses one of the LMCC requirements for a "Data Set" for post refarming spectrum management.

Likewise, Appendix-C serves to provide a hypothetical information flow in a simplified explanation of the spectrum management/frequency coordination process employing the specific reports and recommendations contained herein.

Appendix-D contains a methodology for establishing service areas for existing licenses, in response to LMCC's request.

Appendix-E contains a work sheet for selecting various optional user choices.

2.0 Document Introduction & Scope

In satisfaction of TIA's commitment to the spectrum refarming effort and in response to a request from APCO Automated Frequency Coordination, Inc., for post refarming technical support, the Compatibility Committee's effort has focused on the following:

- Establishment of standardized methodology for modeling and simulating narrowband/bandwidth efficient technologies operating in a post "Refarming" environment;

- Establishment of a standardized methodology for empirically confirming the performance of narrowband/bandwidth efficient systems operating in a post "Refarming" environment; and

- Aggregating the modeling, simulation and empirical performance verification reports into a unified "Spectrum Management Tool Kit" which may be employed by frequency coordinators, systems engineers and system operators.

This document entitled, "A Report on Technology Independent Methodology for the Modeling, Simulation, and Empirical Verification of Wireless Communications System Performance in Noise and Interference Limited Systems Operating On Frequencies Between 30 And 1500 MHz," serves as a report to define the compatibility criteria of the various different modulation types using terms consistent with overall TIA, IEEE, and ITU land mobile efforts.

The expressed purpose of this Committee document is to define and advance a scientifically sound standardized methodology for addressing technology compatibility. This document provides a formal structure and quantitative technical parameters from which automated design and spectrum management tools can be developed based on proposed configurations that may temporarily exist during a migration process or for longer term solutions for systems that have different technologies.

As wireless communications systems evolve, the complexity in determining compatibility between different types of modulation, different operational geographic areas, and application usage increases.

Spectrum managers, system designers and system maintainers have a common interest in utilizing the most accurate and repeatable modeling and simulation capabilities to determine likely wireless communication system performance. With increasing market competition both in terms of modulation techniques offered and in the number of entities involved in wireless communications systems a standardized approach and methodology is needed in the modeling and simulation of wireless communications system performance considering both analog and digital practices at all frequency bands of interest.

In addition, subsequent to wireless communications system implementation, validity or acceptance testing is often an issue subject to much debate and uncertainty. Long after a system is in place and optimized, future interference dispute resolution demands application of a unified quantitative methodology for assessing system performance and interference.

This document also provides a standardized definition and methodology to a process for determining when various wireless communications configurations are compatible. The document contains performance recommendations for public safety and non-public safety type systems that should be used in the modeling and simulation of these systems. This document also satisfies the requirement for a standardized empirical measurement methodology that will be useful for routine proof-of-performance and acceptance testing and in dispute resolution of interference cases that are likely to emerge in the future.

To provide this utility requires that various performance criteria be defined for the different modulations and their specific implementations by specific manufacturers. Furthermore, sufficient reference information will be provided so that software applications can developed and employed to determine if the desired system performance can be realized.

Wireless system performance will be modeled and simulated with the effects of single or multiple potential distortion sources taken into account. These sources include:

- Co-channel users
- Adjacent channel users
- Internal noise sources
- External noise sources
- Equipment non-linearities

- Transmission path geometry
- Delay spread and differential signal phase

Predictions of system performance will be based on the desired RF carrier verses the combined effects of single or multiple performance degrading sources. Performance will be based on a faded environment to more accurately simulate actual usage and will consider both signal magnitude and phase attributes.

It is anticipated that this document will serve as the standard reference for developers and suppliers of wireless communications system design, modeling, simulation and spectrum management software and automated tools.

As the concepts, parameters and methodologies advanced here represent the collective work product of thousands of successful systems worldwide, it is envisioned that future wireless systems that employ this report in the design , modeling, simulation and implementation processes will benefit from consistent performance as designed. Furthermore, spectrum management based upon the same precepts and standard will not only be "consistent" with the designs submitted, but will be more accurate and more flexible accommodating each unique set of conditions rather than relying upon generalized tables and "rules-of-thumb".

Since the migration from the analog world of today to the digital future will be gradual, we anticipate additions to the collective knowledge base. Therefore, on a regular basis, initially on an annual basis, this document will be revised based upon the receipt of relevant additions and/or corrections. Updates will also be issued that reflect refinements as requested by the body of systems designers, and spectrum managers who will ultimately be the users of this report.

3.0 Wireless System Technical Performance Definition and Criteria

The complete definition of the user requirements eventually evolves into the set of conformance requirements. Based on a knowledge of what the User Requirements are and how the conformance testing will be conducted, iterative predictions can be made to arrive at a final design. The following factors should be defined before this process can be accomplished.

3.1 Service Area

This is the users operational area within which a radio system should:

- Provide the specified Channel Performance in the defined area
- Provide the specified CPC Reliability in the defined area

The Service Area Reliability is the computed average of all the individual reliabilities calculated at the data base locations as predicted by the propagation model. These locations shall be uniformly distributed across the Service Area. The Service Area shall be defined in geographic terms.

3.2 Channel Performance Criterion (CPC)

The CPC is the specified minimum design performance level in a faded channel. Its value will be dependent upon ratios of the desired signal to that of the other distortion mechanisms which exist within the service area. It will be defined as a minimum Rayleigh faded carrier magnitude to the sum of all the appropriate static or faded distortion sources, $C_f/(\Sigma I+\Sigma N)$. This Faded Reference Sensitivity will require an absolute power reference, and for digital systems an absolute value in terms of a delay spread performance factor which addresses the decrease in sensitivity which occurs at some given delay spread parameter, after which critical delay spread is achieved. This is provided via the Reference Sensitivity, a static desired carrier-to-noise ratio, C_s/N, for bench testing which provides the absolute power requirement for the C_s/N criterion, for example, 5% static BER at 55 μs delay spread for a given digital modulation or 12 dB SINAD for analog frequency modulation. The delay spread test is with standard input signal level. Table 5 of Appendix-A contains a tabulation of common modulations for projected CPCs.

The Faded Reference Sensitivity may be for a lower CPC than specified by the User. The appropriate design faded sensitivity for the required CPC shall be used. It will be based on the required $C_f/(\Sigma I+\Sigma N)$ for the signal quality baseline required for the particular radio service.

3.3 CPC Reliability

Reliability is the probability that the required CPC will exist at a specified location. It is computed by predicting the mean signal level at a point and determining the margin between the mean and the prediction. The magnitude of this margin determines the probability of achieving the signal level required to produce the CPC.

3.4 CPC Reliability Design Targets

The reliability of wireless communications over a prescribed area is often a issue that is misunderstood. Standardized definitions that are universally applicable are necessary and are presented in the following:

3.4.1 Contour Reliability

The concept of Contour Reliability is a method of specifying both a minimum CPC and a minimum probability of achieving that requirement. The locus of points that meet these criteria would form a contour. Ideally that contour would follow the boundaries of the User's Service Area.

A regulatory contour reliability represents a specific case where the prediction model uses a single "height above the average terrain" value along each radio propagation path, radial between the site and a predicted point, such that predicted signal levels will only decrease with increased distance from the site. This is unrealistic but useful in administration of frequency reuse as it eliminates the randomness of predicted signal levels due to terrain variations, producing a "single unambiguous predicted location" along each radial that provides the specified field strength. The contour is then the locus of those points. Note that the signal strength may denote some specific CPC, but not necessarily. Historically, reuse coordination is based on a non overlapping of contours. The existing systems desired (C) signal contour at some reliability, typically 50%, cannot be overlapped by the proposed new co-channel carrier (I) at a specific reliability, typically 10%.

A formula for converting contour reliability into area reliability from Reudink [1], page 127 is:

$$F_u = \frac{1}{2} \bullet \left(1 + erf(a) + \exp\left(\frac{2 \bullet a \bullet b + 1}{b^2} \right) \bullet \left(1 - erf\, \frac{(a \bullet b + 1)}{b} \right) \right) \qquad \text{[Eq. 1]}$$

At the contour,

$$Px_0 = \frac{1}{2}\left(erfc(a)\right) \qquad \text{[Eq. 2]}$$

where

$$a = \frac{x_o - \alpha}{\sigma \bullet \sqrt{2}} = \frac{Z}{\sqrt{2}}$$
$$b = 10 \bullet n \bullet \left(\frac{\log_{10} e}{\sigma \bullet \sqrt{2}} \right)$$

and,

α is a constant
x_o is the threshold
σ is the log normal standard deviation
n is the power loss value for r^{-n}
F_u is the fractional useful service area probability
Px_o is the fractional probability of x_o at the contour
Z is the standard deviate unit for the fractional reliability at the contour

The resultant solution is based on a uniform power loss exponent and a homogenous environmental loss (smooth earth). Although it doesn't include the effects of terrain, it provides a reasonable first order estimate.

3.4.2 CPC Service Area Reliability

Since contour reliability is a frequently specified user requirement, its conversion to Area reliability is very important as confirmation testing (Section 6.0) is based on the Area reliability, ***not*** on the contour reliability. Note however that the area being defined is that of a bounding contour, not of an irregular Service Area. The design process will produce an area reliability where, at a minimum, the contour reliability is provided throughout the service area.

Figure 1 shows a conversion chart between contour reliability and Area reliability for a constant power loss exponent of 3.5, and three different values of σ.

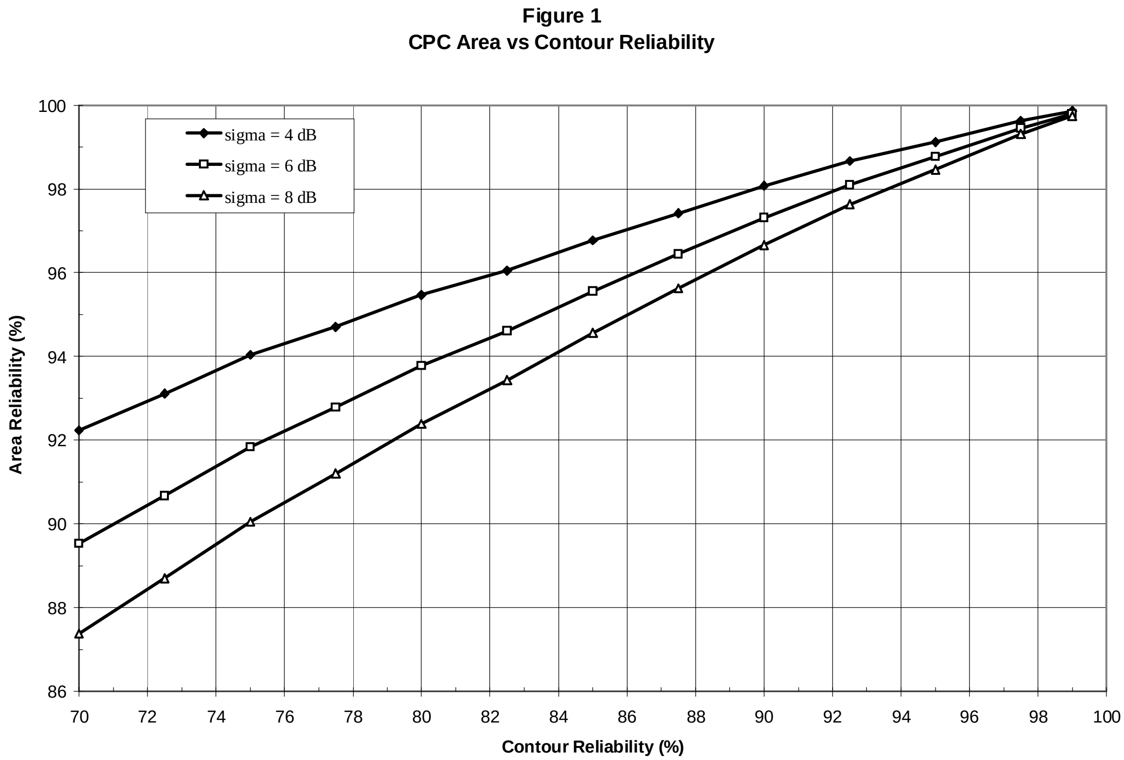

Figure 1. CPC Area vs. Contour Reliability

This seemingly confusing criterion results from two different definitions for Contour reliability. The regulatory definition is not useful in designing a system as the contours do not actually exist. They were developed as an aid for frequency reuse coordination. The definition used in this document is to develop a design target for predicting a CPC.

3.4.3 Tile Reliability Margin

The margin, in dB, provided to create a minimum acceptable probability of achieving the required CPC. This margin is used to determine whether a specific tile will be considered as being a pass or fail, which is used in the calculation of the CPC Area Reliability. This tile reliability margin will be less than the Area Reliability. For example, if the minimum acceptable probability for a tile is 90% probability of achieving the CPC target value, the tile reliability margin, from Section 5.8, would be 8.2 dB. This would produce a CPC Area Reliability, from Figure 1, of approximately 97.5%. The exact value would be subject to an actual prediction rather than the use of the simple model of Figure 1.

3.4.4 Tile Reliability

CPC Area Reliability requires that within a user specified percentage of tiles (sectors) bounded by the service area, the predicted mean signal must exceed the CPC design target by an amount equal to or greater than the Tile Reliability Margin. Thus, the predicted mean signal within each tile includes a tile reliability margin, over and above the specified CPC threshold signal level, to predict that specific level of performance. Any location that has a predicted signal level producing a tile reliability margin less than specified is treated as failing, and is represented on the map accordingly.

The number of tiles which contain a tile margin equal to or greater than that specified above the CPC requirement, divided by the total numbers of tiles, directly predicts the CPC Area Reliability.

3.5 Margins for CPC

Different CPCs, such as those for digital data, may require additional margins above the "standard faded sensitivity". These margins should be used to increase the predicted signal levels and to compensate for the aggregated delay spread so as to achieve the appropriate $C/(\Sigma I+\Sigma N)$ required to provide the CPC.

3.5.1 CPC Subjective Criterion

SINAD equivalent intelligibility, Mean Opinion Scores (MOS) and Circuit Merit have been frequently used to define a Channel Performance Criterion (CPC). A new term, Delivered Audio Quality (DAQ), was developed to facilitate mapping of analog and digital system performance to Circuit Merit and SINAD equivalent intelligibility. DAQ and its SINAD equivalent intelligibility define a subjective evaluation using understandability, minimizing repetition and degradation due to noise to establish high scores. For the purposes of this report, DAQ values are defined in terms of SINAD equivalent intelligibility. These are shown in Table 1 in Appendix-A. Table 1 sets out the approximate equivalency between DAQ and SINAD. Recommendations for public safety, and non-public safety, are provided in Section 3.6.2.2 that follows and D.2.10. In digital systems, the noise factor is greatly diminished and the understandability becomes the predominant factor. The final conversion is what will be defined as the CPC.

The goal of DAQ is to determine what mean $C/(I+N)$ is required to produce a subjective audio quality metric under Rayleigh multipath fading. The reference is to FM analog radio SINAD equivalent intelligibility. That is a static analog measurement so the Table 1 description has been provided to provide a cross-reference.

The requirement for 20 dBS equivalency produces a DAQ of approximately 3.4. This value can then be used for linear interpolation of the existing criteria. CPC requirements would normally specify either a 3 or 3.4 DAQ at the boundary of a protected service area. Note that regulatory limitations may preclude providing this level of CPC for portable in-building coverage.

Noise/Distortion is intended to represent Analog/Digital configurations, where Noise is the predominant factor for degrading Analog DAQ, while Distortion and vocoder artifacts represent the predominant factor for degrading Digital DAQ. Repetition represents the requirement due to low intelligibility.

These values are subjective and will have variability amongst individuals as well as configurations of equipment and distractions such as background noise. They are intended to represent the mean opinion scores of a group of individuals, thus providing a goal for evaluation. It is recommended that samples of each criterion be provided to calibrate user expectations.

Figure 2 shows the various factors that must be included to make a prediction for a specific CPC.

- The Thermal Noise Threshold is the noise contribution of the receiver due to thermal noise. It can be calculated using Boltzmann's constant and an assumed room temperature of 290°K, correcting for the receiver's Equivalent Noise Bandwidth (ENBW) and Noise Figure. This is:

  Thermal Noise Threshold (dBm) = -174 dBm + 10 log (ENBW) + NF_{dB} [Eq. 3]

  Where ENBW is in Hz.

  This then defines the Noise used in all subsequent tests.

- The Static Threshold is the Reference Sensitivity of the receiver. It shall have a static carrier to noise (C_s/N) value for a static performance test, relative to the Thermal Noise Threshold and can be expressed as an absolute power level in dBm or μV's across 50Ω.

- The Faded Threshold differs slightly in definition from the Faded Reference Sensitivity as it is for a faded performance criterion. In the specific case of C4FM, the Faded Reference Sensitivity is for the standard BER (5%) per clause 2.1.5.1 of TIA TSB102.CAAA. The Faded Threshold is for a BER that provides for the specific design CPC. A faded carrier to noise (C_f/N) value must be available for this performance level. This C_f/N value will be evaluated as being a $C_f/(\Sigma I+\Sigma N)$.

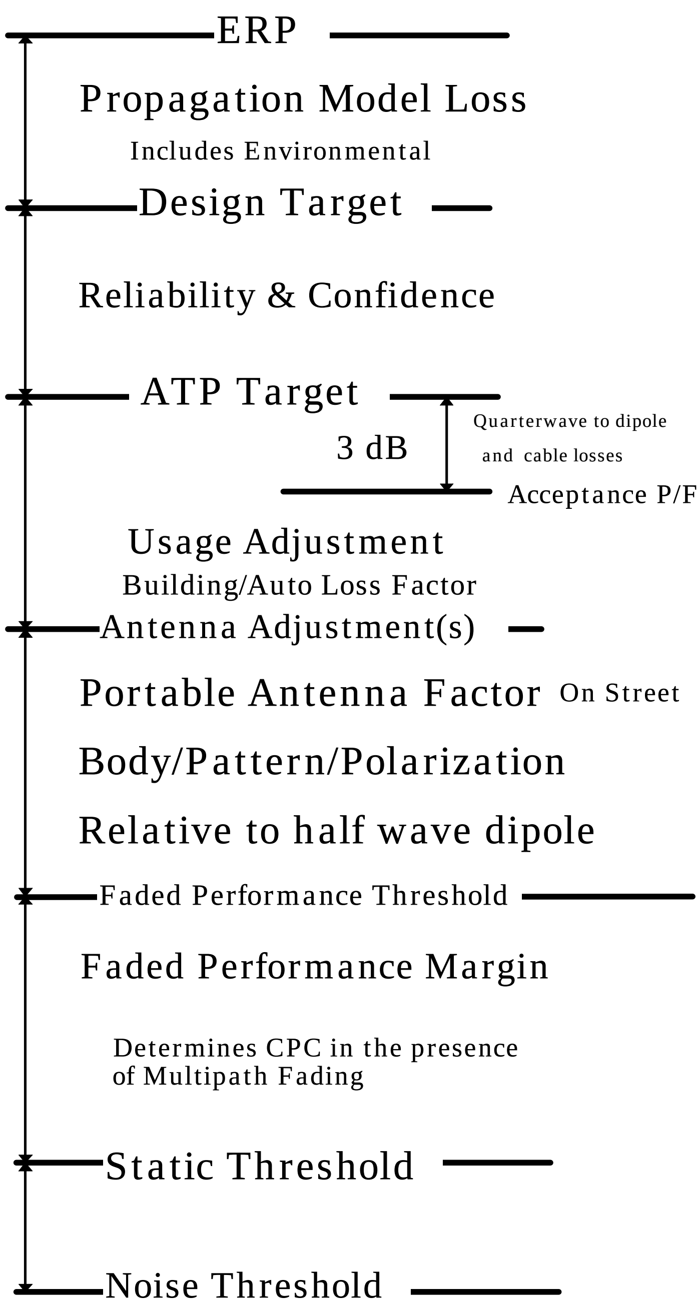

Figure 2. Prediction Factors

- The Adjustment for "Antenna" represents the antenna efficiency of the configuration being designed for. It shall represent the mean losses for that antenna configuration relative to a vertically polarized $\lambda/2$ dipole. For portables it shall include body absorption, polarization effects, and pattern variations for the average of a large number of potential users. For mobiles, it shall include losses for pattern variation for the mounting location on the vehicle and coaxial cable.

  Mobile type unit antenna height corrections shall also be included under this definition. The formulas from Hata [18] are to be employed (Sections 5.1.2.1 and 5.1.2.2).

- User adjustment is for specific usage as necessary for determining portable reliability when operating in a vehicle or in a building with specified penetration loss(es).

- The Acceptance Test Plan (ATP) Target for a hypothetical system is then the absolute power defined by the Static Threshold minus the difference between C_f/N - C_s/N plus the antenna adjustment and any usage adjustment required. For example if the static threshold is -116 dBm (C_s/N = 7 dB (arbitrary)), and -108 dBm is the Faded Threshold (C_f/N = 15 dB), the Fading Margin is 8 dB. This may not be enough for the specific CPC required. If the C_f/N for the desired performance level is 17 dB, then the fading margin is 10 dB, and the Faded Threshold becomes -106 dBm. If the portable antenna has a mean gain of -10 dBd and building losses of 12 dB are required then the average power for the design at street level should be 22 dB greater than -106 dBm (-84 dBm) for this example configuration. Table 5 in Appendix-A provides the *projected* CPC requirements for DAQ 3, 3.4, and 4.

  This establishes the average power which should be measured by a test receiver that has been calibrated to offset its test antenna configuration and cable losses. For example, if the design was for a portable system and the test receiver is using a $\lambda/4$ center mounted antenna with 2 dB of cable loss then a correction factor of -1 dB is applied for the antenna to reference it back to a $\lambda/2$ dipole plus an additional -2 dB for cable loss, -3 dB which would modify the pass/fail criterion from -84 dBm to -87 dBm.

- The Design Target includes the necessary margins to provide for the location variability to achieve the design reliability and a "confidence factor" so that average measured values will produce the CPC. For example, if the desired minimum probability of achieving the CPC is 90%, and a design actually produces such a condition, 50% of the tests would produce results greater than the 90% value and 50% would produce results less than the 90% value. A minor incremental increase in the design would allow the 90% design objective to be achieved. The necessary correction factor varies with the system parameters as indicated in Section 5.8.

- The final element in the prediction involves the actual propagation model, which predicts the mean loss from the transmitter site to a specific predicted location at some probability. The specific electromagnetic wave propagation model selected is critical as the system design, simulation, and modeling accuracy versus system performance will be dependent upon the validity and universality of the selected model. Section 5.0 contains the recommended models and methodology. Section 3.6.2.2 recommends when to use them. The completion of a specified ATP, where close agreement between predicted and measured values is achieved, essentially validates the specific models used. It is recommended that the specific models be employed for system coverage and for frequency reuse and interference predictions to assure consistency and long term validity.

3.6 Parametric Values

The data provided in Table 5 of Appendix-A were voluntarily provided by the manufacturers as "*projected*" values for system design and spectrum management. Publication of these data does not imply that either the manufacturers or TIA guarantees the conformance of any individual piece of equipment to the values provided. Users of these parametric values should validate these values with their supplier(s) to ensure applicability.

3.6.1 BER vs. E_b/N_o

The measurement of E_b/N_o vs. BER for both static and faded conditions is commonly made. For conventional technology implementations, this can be converted to static and faded C/N values with the following equation:

$$\frac{C}{N} = \frac{E_b}{N_o} + 10\,Log\,\frac{[BitRate(Hz)]}{[ENBW(Hz)]} \qquad \text{[Eq. 4]}$$

The ENBW for a known receiver can be used, or a value may be selected from standard receiver bandwidths, to determine faded C/N values for various CPCs. Table 3 in Appendix-A includes the ENBW for various configurations.

From the known static sensitivity and its C_s/N, the value of N, the Thermal Noise floor can be calculated. Based on N and the requirement for $C_f/(\Sigma I+\Sigma N)$ from the faded reference sensitivity for a specified CPC, the absolute value of the average power required is known if the various values of I are also known. The coverage prediction model will predict the value of I.

For example, if E_b/N_o for the reference sensitivity is 5.4 dB for a C4FM receiver (ENBW = 5.76 kHz, IMBE vocoder) at -116 dBm then the C_s/N = 5.4 + 10 Log 9,600/5,760 = 7.6 dB. The calculated Inferred Noise Floor is then -123.6 dBm. From TSB102.CAAB the faded reference sensitivity limit is -108 dBm. This implies a C_f/N = 15.6 dB for 5% BER. If the specified CPC (DAQ = 4) requires 1% BER, then the C_f/N would be appropriately increased by its appropriate value, e.g., 15.6 dB to 21.2 dB. [These numbers are based on the specified minimum performance as listed in TSB102.CAAB clauses 3.1.4 and 3.1.5. The increase for improving 5% BER to 1% BER is from Table 5 of Appendix-A.] Thus the mean power level to provide this performance would be -123.6 + 21.2 = -102.4 dBm.

In a Noise Limited System, the C/N of -102.4 dBm would be the faded performance threshold. In an Interference Limited system, the requirement for $C/(\Sigma I+\Sigma N)$ where ΣI's is, for example, >>N, would require that the design C be 21.2 dB higher for the minimum probability required to provide the CPC at the worst case location. The computer simulations recommended can accurately predict this probability.

3.6.2 Co-Channel Rejection

Different modulation types and implementations require different co-channel protection ratios. The significance of Co-Channel Rejection goes beyond operation in co-channel interference: as measured per TIA TSB102.CAAA, Co-Channel Rejection is equivalent to the static IF carrier-to-noise ratio (C_s/N) required to obtain the sensitivity criterion of the receiver under test. Therefore, a receiver's Co-Channel Rejection number can be used to determine a receiver's IF filter noise floor. This is done using the formula:

$$\text{Noise Floor} = \text{Reference Sensitivity} - C_s/N \qquad \text{[Eq. 5]}$$

The receiver noise floor will be used in the interference model presented in the sections to follow.

Column 2 of Table 5 in Appendix-A gives Co-Channel Rejection values, i.e., static sensitivity in terms of IF carrier-to-noise ratio for the reference sensitivity listed, for many current modulation types.

3.6.2.1 Channel Performance Criterion

Criteria for channel performance are listed in Table 5 of Appendix-A.

3.6.2.2 Propagation Modeling and Simulation Reliability

For public safety agencies, it is recommended that the CPC be applied to 97% of the prescribed area of operation in the presence of noise and interference. Law enforcement and public safety systems should be designed to support the lowest effective radiated power subscriber set intended for primary usage. In most instances this will necessitate systems be designed to support handheld/portable operation. In these instances it is recommended the lowest practicable power level mobile/vehicular radio be assumed. If direct unit-to-unit communications are a primary operational modality, it is recommended that per-channel power control be used, where available, to minimize system imbalance and interference potential. Special consideration of this modality is required as unit-to-adjacent channel unit interference potential is increased.

For Land Mobile Radio (LMR) systems other than public safety, it is recommended that the CPC be applied to 90% of the prescribed area of operation in the presence of noise and interference. Non-public safety systems should be designed to support the typical effective radiated power subscriber set intended for primary usage. In most instances this will necessitate systems be designed to support mobile/vehicular operation. Handheld/portable operations are often secondary. In all instances it is recommended the lowest practicable power level mobile/vehicular radio be assumed. If direct unit-to-unit communications are a primary operational modality, it is recommended per channel power control be used, where available to minimize system imbalance and interference potential. Special consideration of this modality is required as unit-to-adjacent channel unit interference potential is increased. LMR systems that make primary use of handheld/portables are advised prohibit mobile station operation at power levels significantly greater than the design level used for handheld/portable usage.

3.6.2.3 Protected Service Area (PSA)

To determine suitability for assigning channels, a determination of whether the user can qualify for a Protected Service Area (PSA) is required. If the user does not qualify, then it is assumed that sharing will occur. The next requirement is whether the user can monitor the channel before transmitting so as to prevent interfering with current usage. An example of a simple weighted ordering process to select from candidate channels is provided later.

3.6.2.3.1 Proposed System Is PSA

1. Based on the Service Area defined and the appropriate licensing rules, limit the evaluation area to include only those interfering systems which can have a direct impact on the applicant's PSA.
2. Eliminate candidate channels with overlapping co-channel operational service areas.
3. Re-evaluate the remaining candidate channels by quickly evaluating potential signal(s) overlapping service areas using the following simplified prediction method: Use the recommended models, procedures, and ERP adjustments for Adjacent Channel Coupled Power in a "coarse" mode to reduce the number of candidate channels for later detailed evaluation.
4. From the remaining candidate channels, start by calculating the Service Area CPC Reliability of the PSA under evaluation due to noise and all interference sources (co- and adjacent channel interference from PSAs and non-PSAs) using the "fine" mode.
5. When a candidate channel has been identified as meeting the licensee's requirements, an evaluation of the incumbent channels due to the applicant should be made to determine the interference impact to incumbents.
6. If Step 5 produces a successful assignment, the process is complete. Alternatively, it can be continued to evaluate the remaining candidate channels,

looking for an optimal solution. It is anticipated that this alternative solution may involve higher fees due to the greater time and resources required.

3.6.2.3.2 Proposed System Is Not PSA

In this scenario, adjacent channels are assumed to not be capable of being monitored before transmitting. Co-channels may be monitored if they use similar type modulation.

The assignment of a non-PSA frequency assumes that, at some time, sharing will occur. Therefore, there is no optimal solution, and any immediate solution may change in the future. Numerous tradeoffs and coordinator judgment will be required out of necessity. For that reason, this section will identify some of the factors that could potentially rank candidate channels for a recommendation. Weighting factors and the way they are applied are not specified. A similar coverage evaluation process as defined in Section 3.6.2.3.1, in conjunction with the judgmental factors, should be applied.

1. Based on the Service Area defined and the appropriate licensing rules, limit the evaluation area to include only those interfering systems which can have a direct impact on the applicant's Service Area.
2. Eliminate candidate channels using the following judgmental factors:
   - Number of licensees
   - Simplex base-to-base interference potential, point-to-point path
   - Number of units shown for each incumbent
   - Overlap of service areas
   - Similar size of co-channel service areas
   - Potential for adjacent channel interference due to overlapping service areas, potential of the near/far problem
   - Potential for adjacent channel interference due to signals overlapping service areas
   - Common or nearby site compatibility
   - Time of day utilization
   - Competition, same type of business
   - Ability to monitor before transmitting
   - Compatibility of modulation to allow monitoring of "over the air audio"
   - Use of encryption
   - Use of trunking
     - ♦ Dedicated control channel
     - ♦ Non-dedicated control channel
3. Re-evaluate the remaining candidate channels by quickly evaluating potential signal(s) overlapping service areas using a simplified prediction method. This method should use the recommended models, procedures, and ERP adjustments for Adjacent Channel Coupled Power in a "coarse" mode to reduce the number of candidate channels for later detailed evaluation..
4. From the remaining candidate channels, start by calculating the Service Area CPC Reliability of the non-PSA under evaluation due to noise and all

interference sources (co- and adjacent channel interference from PSAs and non-PSAs).

5. When a candidate channel has been identified as meeting the licensee's requirements, an evaluation of the incumbent channels due to the applicant should be made to determine the interference impact to incumbents.
6. The judgmental factors of Step 2 should be re-examined for applicability.
7. If Step 5 produces a successful assignment, the process is complete. Alternatively, the process can be continued to evaluate the remaining candidate channels, looking for an optimal solution. It is anticipated that this alternative solution may involve higher fees due to the greater time and resources required.

3.6.2.3.3 Example of Ordering

Consider a case with four successful candidates. Each has two co-channel PSAs and three have adjacent channel PSAs. Refer to Table 6 in Appendix-A for the example.

3.6.3 Interference Prediction

It is assumed that for any modulation combination, it is valid to treat adjacent channel interference as additional noise power that enters a receiver's IF filter. Interference between different modulation types may be calculated based on the power spectrum of the given transmitter modulation and the IF filter selectivity and IF carrier-to-noise ratio required to obtain the specified CPC in a Rayleigh faded channel. The $C_f/(I+N)$ then becomes a predictor of CPC.

The $C_f/(I+N)$ required for the "victim" system to meet its required CPC must be known in order to determine an interference level. The subscript "f" indicates that the carrier-to-noise ratio is determined for Rayleigh *faded* conditions. When performing interference calculations, it is important to use faded carrier-to-noise values since faded conditions more accurately represent the field environment.

Columns 3-5 of Table 5 in Appendix-A list projected CPC requirements for mainstream modulation techniques at various DAQ levels in faded conditions. For digital modulations, bit error rates associated with each CPC are given. These may be used to determine if a given $C_f/(I+N)$ exists in an actual field test application. Static reference sensitivity (C_s/N) also is given. This value can be used to determine the receiver noise floor for interference modeling. A particular manufacturer's implementation may vary from these values somewhat, but the variation is expected to be small.

A key factor in determining adjacent channel interference is the IF selectivity of the victim receiver. There is potentially wide variation in IF selectivity between manufacturers, but definition of a standard IF selectivity is helpful in defining a reproducible test. A set of prototype IF filters is given in Table 3. The filter implementations used here were selected for their ability to compactly define an explicit and reasonable implementation, not to suggest an optimum implementation for a given modulation type. Formulas also are provided for use in simulations in Table 4 of Appendix-A.

Receiver local oscillator noise also is a factor in interference. Since this is a function of receiver design, and performance may vary greatly between various implementations, and since the type of interference does not affect co- or adjacent channel performance, this factor will not be considered in the analysis. It is understood, however, that a certain noise floor due to local oscillator noise will exist.

Transmitter spectra will be modeled using measured spectrum power densities (SPDs). The SPDs are measured according to the procedures given in Section 6.6. Some are represented in tabular form in Appendix-C. The SPDs in Appendix-C are given in terms of Watts and normalized to a total transmit power of 1 Watt.

4.0 Noise

4.1 Environmental RF Noise

To determine effective receiver sensitivity, it is essential that the level of environmental noise be known. It should first be pointed out that it is seldom necessary to measure environmental noise in a mobile environment at frequencies higher than 400 MHz because it is rare for the total environmental noise to exceed kT_0b. A major exception to the foregoing statement is frequencies near 821 MHz in which the mobile can experience noise generated by non-wireline cell sites. The foregoing advice is summarized in Table 9.

4.2 Historical RF Noise Data

Noise measurements have been conducted by many researchers. One representative noise survey was that of Spaulding and Disney [9]. Their work resulted in the following RF noise equation:

$$N_r = 52 - 29.5 \log_{10} f_{MHz} \quad \text{dB} \qquad (\text{Relative to } kT_0b) \qquad [\text{Eq. 6}]$$

Where N_r is the "quiet rural" noise level relative to kT_0b. They also arrived at the following corrections for environments other than "quiet rural" should be added to N_r:

Rural: 15 dB   Residential: 18 dB   Business: 25 dB

The total cannot be less than 0 dB (relative to kT_0b).

Environmental noise is highly variable even within the same environment and the only certain means of determining the level of environmental noise (and thus the effective sensitivity) is to conduct a noise measurement program.

4.3 RF Noise Measurement Methodology

4.3.1 Receiver Selection

By far, the best tool for making a noise measurement is a receiver designed specifically for that purpose, such as an Rohde & Schwartz RSVS (example only). This type of receiver has numerous advantages, and two disadvantages when compared to a communications receiver:

- A specialized measurement receiver is expensive.
- The measurement bandwidth is somewhat inflexible.

This last may not be much of a disadvantage, since the noise spectral power density can easily be calculated and the noise power in any given ENBW can be calculated from that.

A communications receiver can also be used for making noise measurements. Although they do not have the many features provided by a measuring receiver, they are adequate for the job when properly applied and do have a small number of advantages over measuring receivers, including low cost and having the exact bandwidth that is needed for the given application.

If a communications receiver is to be used, consideration should be given to adding a low noise preamplifier to increase the measurable range at the low end. Otherwise, noise that is below the measurement threshold but may still contribute to degradation will be ignored. Care should be exercised such that intermodulation products can be produced, distorting the measurements.

4.3.2 Antenna Selection

Since noise originates from all directions, an argument can be made for measuring noise by using an antenna that is sensitive in all directions; i.e., one with an isotropic pattern. In the real world, however, specific types of antennas are used in land mobile communications and they typically have a great deal of vertical directivity. To match the results to the hardware that a user will be using in the real world, the measurements should be taken with the type of antenna that will be used by the typical user.

Radio frequency noise is frequently expressed in terms of dB above the noise floor (kT_0b) or in terms of spectral power density (in units such as dBm/kHz). Using such terms rather than the received signal level has the advantage of making the measurement "portable" to receivers with any noise bandwidth. To do so, of course, it is necessary to know the following in addition to the received signal level: (a) the gain or loss of the antenna system (including cable and connector losses), and (b) the measuring receiver's ENBW.

4.3.3 RF Noise Measurement in a Mobile Environment

A typical receiver's sensitivity can be stated in terms of a carrier to noise value; e.g., a particular receiver may require a 7 dB C_s/N to produce the static reference sensitivity. Knowing the noise power at the frequency of interest at a given location and the values from Table 5 of Appendix-A allows, the user to calculate the receiver's sensitivity for the desired CPC in that environment.

A standard communications receiver can be used for the noise measurement. If the receiver's Received Signal Strength Indicator (RSSI) bus is considerably more sensitive than the sensitivity corresponding to the desired CPC, a preamplifier will not be necessary to extend the measurable range; otherwise, a low noise preamplifier must be connected between the antenna and the receiver. The receiver must then be precalibrated. Connect a signal generator to the input of the preamplifier (or the receiver if no preamplifier is used). In the low signal range, this calibration should be done in 1 decibel intervals. Each calibration point should be repeated many (≥ 30) times to ensure a valid reading. All of this may be automated by a data acquisition device/system.

The actual readings are taken by driving around the evaluation area using a test setup to take readings in an automated fashion. A typical test setup would consist of the antenna and receiver, a notebook computer, and an analog-to-digital (A/D) converter on a PCMCIA card. A more fully automated system could include Global Positioning Satellite (GPS) or Differential Global Positioning Satellite (DGPS) data to eliminate user interface for location information.

A computer program can be written to take the necessary readings subtract the effects of the antenna system, compare the results to the calibration curve, and note the results corresponding to a given location. This will give a noise power value, typically in dBm. To arrive at the noise level relative to kT_0b, one must know the Equivalent noise bandwidth. Knowing that, one merely subtracts kT_0b from the (already determined) noise power.[1]

After taking the data, the user can then establish noise contours for the area of interest. Using this information, it is possible to, knowing the receiver's C_s/N performance for a given CPC, establish the receiver's effective sensitivity on a geographic basis.

4.3.4 Fixed RF Noise Measurement

An entirely different approach is taken to doing site noise measurements. Connect a coaxial switch so that one pole is connected to a simulation of the proposed antenna system, and the other pole is connected to a matched coaxial load. The moving contact is connected via an isolated RF coupler (such as a directional coupler) to a receiver similar to the one that will be used in the proposed system. Switch the coaxial switch so that the load is connected in. Connect a (1 kHz 60% system deviation) modulated RF signal generator to the isolated port of the coupler. Increase the RF level of the RF signal generator until the SINAD and/or BER produced by the receiver approaches the value that corresponds to the desired CPC. Note the RF level. Next, switch the coaxial switch to the antenna system. Increase the RF level until the SINAD reading again reaches the desired level. Note the RF level. The difference in levels is the amount by which the specified sensitivity must be increased to arrive at the effective sensitivity. It should be noted that it is very advisable to make this measurement at several times throughout the workday to account for variations in the use of the RF sources on the site.

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[1] For ease of calculation, it should be noted that the value of kT_0 is -144 dBm. Note that to use this number, bandwidth must be expressed in kHz.

The method discussed in the previous paragraph is identical to that discussed in Section 6.8.5.1. See that section for a more detailed discussion.

The noise power can be ascertained from this measurement by knowing the required C_s/N for the target CPC. (See Section 3.6.2. and Table 5 of Appendix-A) Using the (previously calculated) effective sensitivity and subtracting out the required C_s/N, yields the received noise power. Knowing the receiver's ENBW, it is a simple matter to calculate the noise relative to kT_0b merely by subtracting kT_0b (in dB units) from the received noise power (in dB units).

4.4 Symbolic RF Noise Modeling and Simulation Methodology

4.4.1 Receiver/Multicoupler Interference

Receiver intermodulation effects are rarely considered in system interference. When tower mounted amplifiers and/or amplified receiver multicouplers are used they can dramatically increase the link margins, but introduce intermodulation which is detrimental.

The amount of gain provided has a direct impact on the overall noise figure of the cascaded combination of elements and on the intermodulation performance. As linear systems come into existence an increased awareness of the tradeoffs is necessary to more accurately calculate the effect. Adding gain without determining its overall effect. on the system performance and interference potential should not be tolerated.

Some base stations specify the performance sensitivity at the input to the receiver multicoupler. Most base stations receiver noise figures fall between 9 and 12 dB, with a typical design noise figure of 10 dB. The overall receiver multicoupler scheme has a composite noise figure of between 5 and 7 dB, with 6 dB being a typical design value. With a true noise figure of 4 dB, 25 dB of gain, followed by 16 dB of splitting loss and one dB of cable loss, the resulting noise figure of the cascaded chain can be calculated using the formula:

$$NF_c = NF_1 + [NF_2 - 1]/G_1 + [NF_3 - 1]/[G_1 \bullet G_2] \qquad \text{[Eq. 7]}$$

where:

NF is the Noise Factor (numeric)
G is the Gain of an Amplifier (numeric)

NF_1 = 4.0 dB = **2.5**. G_1 = 25 dB = **316**
NF_2 = 17 dB = **50** G_2 = -17 dB = **0.02**
NF_3 = 10 dB = **10**
NF_c = 2.5 + [50 -1]/316 + [10 - 1/[316 • 0.02] = 4.08 = 6.1 dB

From this example, the overall noise figure of the combination is improved over the base station receiver by itself but degraded from the noise figure of the multicoupler's amplifier. By increasing the gain of the amplifier, and reducing the loss in the splitter, the cascaded noise figure trends toward the noise figure of the multicoupler. However, all the excess gain tends to increase the level of intermodulation products for components down stream. With linear systems, a specification that limits the amount of "excess gain" that can be introduced prior to the base receiver may be necessary to keep the entire system operating within a linear region.

To determine the absolute power level of the intermodulation products requires the use of the Third Order Intercept point (IP^3). Considerable confusion exists around the IP^3 due to manufacturers specmanship. Most manufacturers use the Output Third Order Intercept Point (OIP^3) as it produces a higher number. Reducing the manufacturers OIP^3 by the gain of the amplifier calculates the Input Third Order Intercept Point (IIP^3). This is more useful as one can now determine the intermodulation products with respect to the desired carrier and design noise threshold, adjusting absolute levels by selecting gain and loss elements.

4.4.2 Intermodulation

A receiver with an 80 dB Intermodulation Rejection (IMR) has an IIP^3 in the 0 to +5 dBm range. To measure the IMR, start with the static sensitivity criterion, such as 12 dB SINAD, C_s/N = 5 dB for an analog FM radio with 25 kHz channel spacing. The desired is increased by 3 dB and two interfering signals are injected. One is the adjacent channel and the other is the alternate channel. In this case, 2 times the adjacent channel, minus the alternate channel will create a product that falls back on the same frequency as the desired. The two signals are increased at the same level until the 12 dB SINAD performance specification is again reached. The difference between the intermodulating signals and the original reference is the IMR of the receiver.

In Figure 3, if the IMR specification is 80 dB, and the 12 dB SINAD is -119 dBm, (0.25 µV), the following test would be conducted. Inject -119 dBm and measure 12 dB SINAD.

The inferred design noise threshold would be -124 dBm. Increase the desired signal level to -116 dBm, 3 dB boost. Inject the adjacent and alternate channels, increasing them until 12 dB SINAD is once again obtained. With a receiver of 80 dB IMR, the adjacent and alternate channels should be 80 dB above the 12 dBS, -39 dBm. This once again produces a C_s/N of 5 dB, 12 dBS, comprised of the -124 dBm design thermal noise and another -124 dBm noise equivalent from the interference from the IMR. The combined noise sources equal -121 dBm versus the desired signal at -116 dBm. Figure 3 illustrates a graphical solution for the IIP^3 of +3.5 dBm. Two slopes are constructed. A 1:1 relationship from the design noise threshold and a 3:1 slope for the third order products offset by (80 + 5) 85 dB at the design noise threshold. A formula for this relationship is:

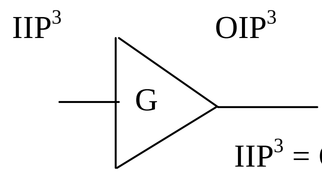
$$IMR = 2/3\ (IIP^3 - Sens) - 1/3\ (C/N\ @\ Sens) \qquad \text{[Eq. 8]}$$

In this example, sensitivity for 12 dB SINAD was -119 dBm with a C/N of 5 dB. If the IMR is 80 dB, the IIP^3 is = +3.5 dBm.

The preceding calculation was for a single receiver. When cascaded with a receiver multicoupler the process becomes more complex. The IIP^3 of the receiver must be found to determine the interaction with the parameters of the receiver multicoupler chain.

Receiver multicoupler manufacturers typically use the OIP^3 for their specification. Knowing the gain of the amplifier and the splitting losses one can calculate the impact on the desired and undesired portions. This will also highlight the case of when there are two amplifiers in the multicoupler chain and the gain inserted to lower the cascaded effective noise figure reduces IMR performance too much. Tower top amplifiers normally involve three amplifiers, the tower top amp, a distribution amplifier and the actual receiver.

An example will illustrate the issues. Consider the previously described base station configuration with a receiver multicoupler. The parameters and lineup are shown in Figure 4. The noise figure is calculated to be 9.2 dB, based on 12 dBS = -119 dBm, C/N = 5 dB and the ENBW = 12 kHz.



$$IIP^3 = OIP^3 - G \text{ dB}$$

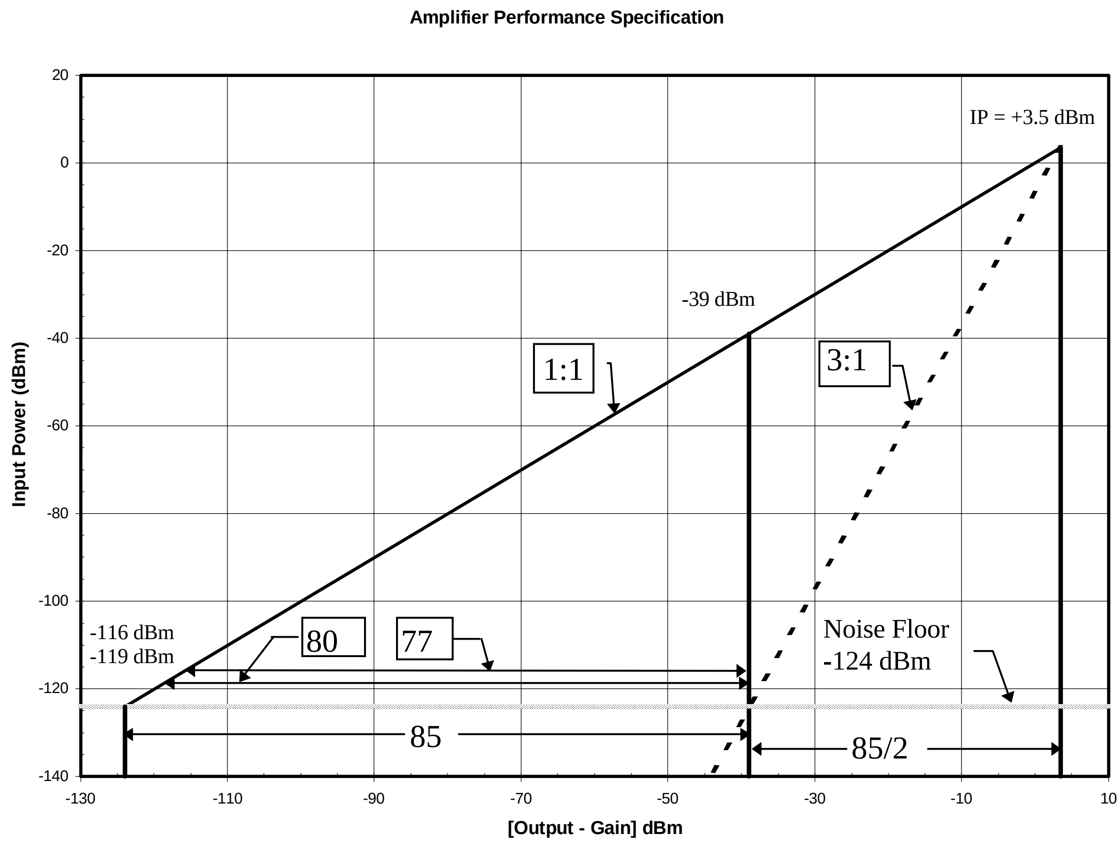
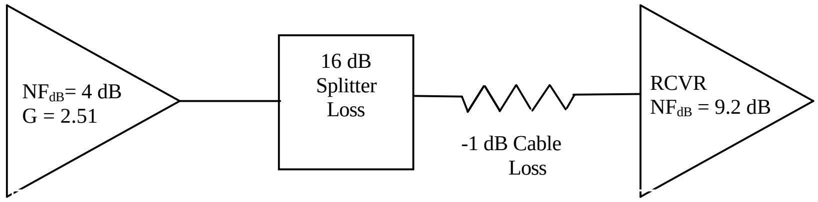


Figure 3. Amplifier Performance Specifications

The receiver multicoupler has 25 dB of gain and 17 dB of losses prior to the receiver's antenna port. The OIP^3 is given as +34 dBm. By subtracting the gain we calculate an IIP^3 of +9 dBm.



NF_1 = 4 dB = 2.51
NF_2 = 17 dB = 50
NF_3 = 9.2 dB = 8.32

$G1$ = 25 dB = 316
$G2$ = -17 dB = 0.02

$$NFc = 2.51 + \frac{(50-1)}{316} + \frac{(8.32-1)}{(316)(0.02)}$$

$$NFc = 2.51 + 0.16 + 1.16 = 3.83 = 5.83 \text{ dB}$$

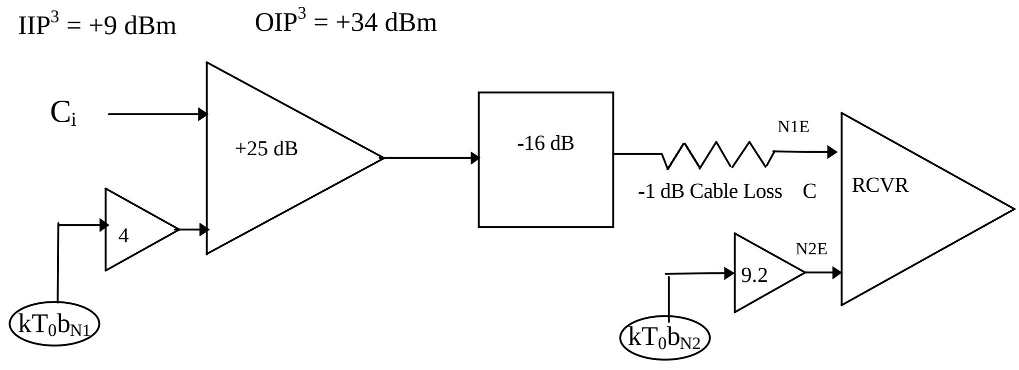
NF_{imp} = 9.2 - 5.83 = 3.37 dB

Figure 4. Noise Figure Calculation

The traditional cascaded noise figure approach calculates an effective noise figure at the input of the multicoupler of 5.83 dB, indicating a 3.37 dB improvement in the noise figure for the combination.

4.4.3 The Symbolic Method

Symbolically all active devices are shown, in Figure 5, as a single amplifier with some known amount of gain. Inputs to the amplifier include another amplifier which has the gain of the device's noise figure which is fed from a noise source equal to the kTb value of the actual receiver. Following the flow from the first amplifier, the noise source is amplified and attenuated until it arrives at the input of the final receiver. In this case the accumulated noise power is -121.2 dBm. The receiver has its own noise source which is -124.0 dBm. The sum of these two noise sources is -119.37 dBm. To achieve a C/N of 5 dB requires that the C be -114.37 dBm. To achieve that power with the gain and losses would require a -122.37 dBm signal at the input to the first amplifier. The receivers sensitivity by itself for a C/N of 5 dB is -119 dBm so the improvement of the combination is -119 - (-122.37) = 3.37 dB, the same as calculated by the cascaded noise figure formula.



N1E = -133.2
+4.0
+25.0
<u>-17.0</u>
-121.2 dBm

N2E = -133.2
<u>+9.2</u>
-124.0 dBm

IIP^3 = +3.5 dBm
12 dBS = -119 dBm
ENBW = 12 kHz
C_s/N @ 12 dBS = 5 dB
NF_{dB} = 9.2 dB
kT_0b = -133.2 dBm

Σ(N1E + N2E) = -119.37 dBm
Ci = [-119.37 + 5] -25 +16 +1 = -122.37 dBm
Performance Improvement = -119.0 - (-122.37) = **3.37 dB**

Figure 5. Symbolic Method

This approach allows evaluating the effect of system IMR noise power. Equations 11 and 12 can be used to calculate either a relative or absolute power level for the third order product. First an equivalent signal power level must be calculated to use in this evaluation. For the classic IMR case as measured by the EIA, the equivalent signal power C_i, is:

$$C_i = \frac{2(\text{Adjacent Channel Power}) + \text{Alternate Channel Power}}{3} \qquad \text{[Eq. 9]}$$

For the EIA test, both the adjacent and alternate channels are held at the same power level. However in the field, users frequently must deal with IMR where the frequency relationships aren't that close and are unequal in power. In these cases the equivalent power to use for C_i would be to consider only the worst case which would be where the two signals have different average powers. It is also assumed that the mixer remains constant and that no additional selectivity is available. In this case:

$$C_i = \frac{2(\text{Highest Channel Power}) + \text{Lowest Channel Power}}{3} \qquad \text{[Eq. 10]}$$

An application with specific frequencies, calculates the interfering carrier levels and the intermodulation power that will result for a specific design or problem evaluation.

At the input of an amplifier:

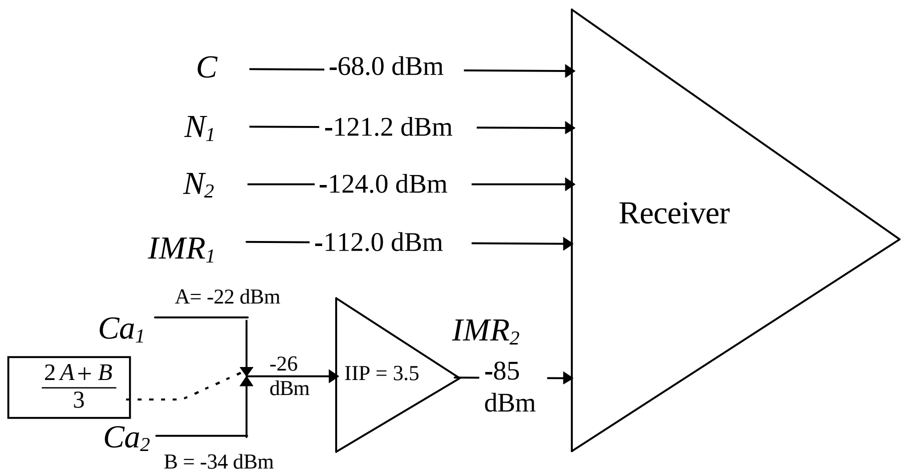
Relative IMR = 2 (IIP^3 - C_i), where C_i = Equivalent interferer. [Eq. 11]

Absolute IM Level = C_i - Relative IMR. [Eq. 12]

In most cases system designers will be interested in the level of the IM and will then follow it through the chain of amplifiers and loss elements until it arrives at the input of the last amplifier stage. At the final stage, the individual carriers also will be present and will once again produce IMR. The total noise would then be the sum of the individual noise sources and the individual IMRs, C/Σ (N + IMR). Continuing with the example, consider the following case.

The Adjacent channel power, C_{a1}, at the input to our multicoupler amplifier is -30 dBm, and the Alternate channel, C_{a2}, is -42 dBm. This is the classic 2A-B IM case. From [Eq. 10]:

C_i = [2(-30) + (-42)] / 3 = -34 dBm [Eq. 13]



$$\Sigma Ns = N_1 + N_2 + IMR_1 + IMR_2 = -85 dBm$$

$$\frac{C}{\Sigma(N + IMR)} \equiv 17 dB, \therefore C = -68 dBm$$

Figure 6. Multicoupler IMR Performance Example

The IIP^3 of the first amplifier is +9 dBm. The absolute IMR at the input of the receiver is calculated to be -34 dBm -2(43) + 25 -17 = -112 dBm. The individual C_{a1} and C_{a2} would be amplified (25 - 17) = 8 dB to -22 dBm and -34 dBm respectively. Their C_i is now -26 dBm. Thus the absolute IMR , using equations 10 and 11, introduced by the receiver itself is -85 dBm. Now there are five different inputs to the final receiver that impact its performance; the desired C, the four noise sources that must be overcome, $N_1 + N_2 + IMR_1 + IMR_2$. In this example, the IMR due to the high adjacent and alternate channels are controlling. To achieve a desired C = 17 dB above the composite noise generators requires that the signal at the input of the receiver would have to be -68 dBm to achieve our CPC for 25 kHz analog FM performance of DAQ = 3. As shown from this example, additional amplifiers in the "gain chain" can amplify high interfering signals to such a high level that IMR in unavoidable. Proper addition of attenuators is necessary to optimize the sensitivity verses IMR performance.

It is important to remember that there is a probability consideration that has to be included, and that the type of interference must also be considered. For example, if the interfering adjacent channel had the same CTCSS code, a receiver would open whenever the interference was present and no desired carrier was present. This would dramatically impact the users perception of the amount of interference.

4.4.4 Non-Coherent Power Addition Discussion

When adding powers, the values must be in some form of watts before they are added. In microwave systems the picowatt is commonly used. To add the powers, it is not necessary to convert them to a specific watt level, milliwatts, microwatts, or picowatts. As long as they all are at the same pseudowatt level they can be added and converted back and forth to the nonlinear form of decibels.

The following simple method may be used to combine powers in the decibel form. It only requires taking the dB difference of two powers and looking up in Figure 7 or Table 10 of Appendix-A a value to add to the higher power. For example, if a -113 dBm and -108 dBm are to be combined, the difference is 5 dB which from Table 10 indicates that +1.2 dB must be added to the -108 dBm for a composite -106.8. For cases with more than two power levels, the process can be repeated multiple times. P1 and P2 can be combined to Pc which can then be combined with P3 for the average power of all three.

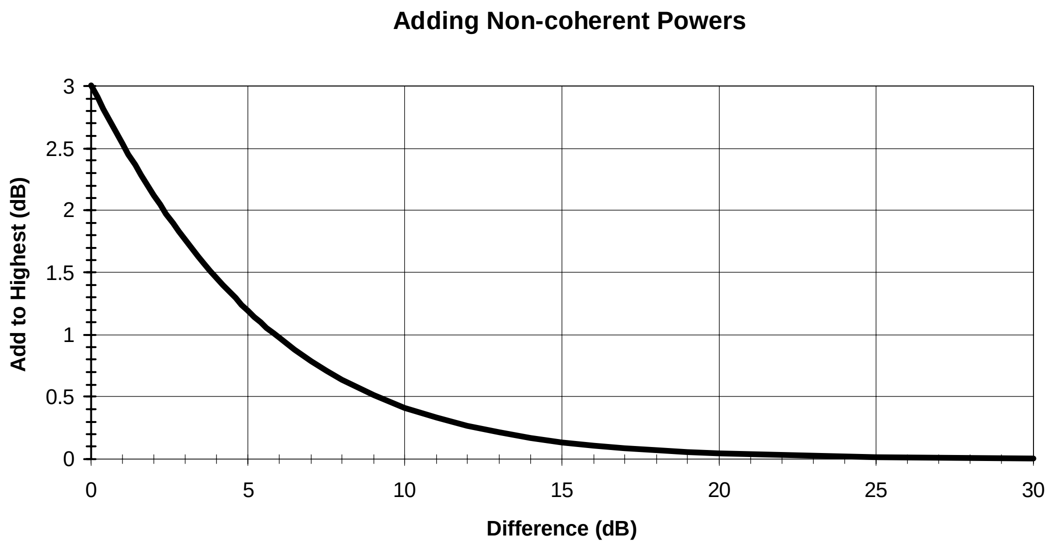


Figure 7. Adding Non-Coherent Powers

5.0 Electromagnetic Wave Propagation Prediction Standard Model

For studies involving spectrum management, two types of propagation models have been identified as appropriate. The first is a simple empirically-based model described below as the "Okumura/Hata/Davidson" model, which provides rapid calculation of path loss for line of sight conditions using terrain and land usage data.

The second model is a physical rather than empirical model, which explicitly takes into account terrain and ground clutter features present along with the great circle path from the transmitter to the receiver. It is described below as the "Anderson 2D" model. It provides more accurate path loss predictions than the "Okumura/Hata/Davidson" model under non line of sight conditions. Based on extensive comparisons with measurement data, this model produced the best overall results when compared to several other models that were evaluated. The "Anderson 2D" model is therefore recommended as the standard for frequency coordination of systems requiring a "Protected Service Area" (PSA), or other conditions where a detailed assessment of interference is desired. This process is contained in Section 3.6.2.3.1. For non-PSA systems, the rapid calculation method contained in Section 3.6.2.3.2 using the "Okumura/Hata/Davidson" model is recommended.

5.1 The OKUMURA Model

The OKUMURA model [14] is an empirical model. The results were published as curves which contain various correction factors for predicting the average power levels. When used in this section and associated subsections, the term "HAAT" refers to the HAAT in the direction of the radial under consideration, ***not*** to the overall site HAAT.

5.1.1 Hata Conversion

Hata converted the OKUMURA model for computer use [18]. He developed a series of formulas that provide OKUMURA predictions, but limited their applicability to:

- Range from Base, 1 - 20 km
- Frequency Range, 150 - 1500 MHz
- Base HAAT, 20 - 200 meters

5.1.2 Davidson Extension

Davidson has added correction factors to extend Hata's formulas back to the full range of OKUMURA and has extended the applicable distance to 300 km. This covers the following parameters:

- Frequency Range, 30 - 1500 MHz
- Base HAAT, 20 - 2500 Meters
- Range from Base, 1 to 300 km

Use the larger (greater loss) of PL or PL2 as calculated by either one of the subroutines below.

5.1.2.1 Sample OKUMURA/HATA/DAVIDSON Program - Metric

```
C SUBROUTINE TO COMPUTE THE HATA PATH LOSS FROM OKUMURA MODIFIED BY
C DAVIDSON, METRIC VERSION 2.1     10/21/96
C ASSUMES THAT THE MOBILE HEIGHT FOR MEDIUM SMALL CITY IS SUBURBAN-
C QUASI OPEN-OPEN AND FOR LARGE CITY IS URBAN
C ***************************************************************
C INPUT TO THE SUBROUTINE
C FREQ ..... FREQUENCY IN MHZ
C HEIGHT ... BASE HEIGHT ABOVE AVERAGE TERRAIN (HAAT) IN METERS
C HIMOB .... MOBILE HEIGHT IN METERS
C RANGE .... DISTANCE BETWEEN TRANSMITTER AND RECEIVER IN km
C ENVIOR ... THE ENVIRONMENT OF THE MOBILE, CHOICE OF 4
C            1) URBAN
C            2) SUBURBAN
C            3) QUASI OPEN
C            4) OPEN
C **************************************************************
C OUTPUT OF THE SUBROUTINE
C PL .... HATA/DAVIDSON PATH LOSS BETWEEN ISOTROPIC POINT SOURCES IN DBI
C PL2 ... FREE SPACE PATH LOSS BETWEEN ISOTROPIC POINT SOURCES IN DBI
C **************************************************************
C
      SUBROUTINE LOSS (FREQ,HEIGHT,HIMOB,RANGE,ENVIOR,PL,PL2)
      CHARACTER ENVIOR*8
C FIRST COMPUTE HATA URBAN
      PL=69.55+26.16*ALOG10(FREQ)-13.82*ALOG10(HEIGHT)+
     + (44.9-6.55*ALOG10(HEIGHT))*ALOG10(RANGE)
C SUBTRACT HATA CORRECTION FOR MOBILE HEIGHT, URBAN = LARGE CITY ELSE
C OTHER ONE.   USE 300 MHz AS THE FREQUENCY BREAK POINT.
      IF(ENVIOR.EQ.'URBAN') THEN
        IF(FREQ.GT.300) THEN
          PL=PL-(3.2*(ALOG10(11.75*HIMOB))**2-4.97)
        ELSE
```

```
          PL=PL-(8.29*(ALOG10(1.54*HIMOB))**2-1.1)
        ENDIF
      ELSE
        PL=PL-(1.1*ALOG10(FREQ)-0.7)*HIMOB+
     +   (1.56*ALOG10(FREQ)-0.8)
      ENDIF
C SUBTRACT HATA CORRECTION FOR OTHER ENVIRONMENTS
      IF (ENVIOR.EQ.'SUBURBAN') PL=PL-5.4-2*(ALOG10(FREQ/28)**2)
      IF (ENVIOR.EQ.'OPEN'.OR.ENVIOR.EQ.'QUASI O')
     +    PL=PL-40.94+18.33*ALOG10(FREQ)-4.78*(ALOG10(FREQ)**2)
      IF (ENVIOR.EQ.'QUASI O') PL=PL+5
C NOW EXTEND IT IF YOU ARE OVER THE RANGE LIMIT OR BASE HEIGHT LIMIT
       R1=20
       R2=64.38
C FOR ALL RANGES GREATER THAN 20 km ADD A FACTOR
      IF (RANGE.GT.R1) THEN
        PL=PL+(0.5+0.15*ALOG10(HEIGHT/121.92))*(RANGE-R1)*0.62137
      ENDIF
C FOR ALL RANGES GREATER THAN 64.38 km SUBTRACT A FACTOR
      IF (RANGE.GT.R2) PL=PL-0.174*(RANGE-R2)
C FOR ALL BASE HEIGHTS GREATER THAN 300 M SUBTRACT A FACTOR
      IF (HEIGHT.GT.300) THEN
        PL=PL-0.00784*ABS(ALOG10(9.98/RANGE))*(HEIGHT-300)
      ENDIF
C MAKE THE EQUATIONS THAT WORK FOR 1500 MHz GO DOWN TO 30 MHz
      PL=PL-(FREQ/250)*ALOG10(1500/FREQ)
      R3=40.238
      IF(RANGE.GT.R3) PL=PL-0.112*ALOG10(1500/FREQ)*(RANGE-R3)
C COMPUTE FREE SPACE PATH LOSS IN DBI
      PL2=32.5+20*ALOG10(FREQ)+20*ALOG10(RANGE)
      RETURN
      END
```

5.1.2.2 Sample OKUMURA/HATA/DAVIDSON Program - English

```
C SUBROUTINE TO COMPUTE THE HATA PATH LOSS FROM OKUMURA MODIFIED BY
C DAVIDSON, ENGLISH VERSION 1.2   10/21/96
C ASSUMES THAT THE MOBILE HEIGHT FOR MEDIUM SMALL CITY IS SUBURBAN-
C QUASI OPEN-OPEN AND FOR LARGE CITY IS URBAN
C **************************************************************
C INPUT TO THE SUBROUTINE
C FREQ ..... FREQUENCY IN MHZ
C HEIGHT ... BASE HEIGHT ABOVE AVERAGE TERRAIN (HAAT) IN FEET
C HIMOB .... MOBILE HEIGHT IN FEET
C RANGE .... DISTANCE BETWEEN TRANSMITTER AND RECEIVER IN MILES
C ENVIOR ... THE ENVIRONMENT OF THE MOBILE, CHOICE OF 4
C            1) URBAN
C            2) SUBURBAN
C            3) QUASI OPEN
C            4) OPEN
C *************************************************************
C OUTPUT OF THE SUBROUTINE
C PL .... HATA/DAVIDSON PATH LOSS BETWEEN ISOTROPIC POINT SOURCES IN DBI
C PL2 ... FREE SPACE PATH LOSS BETWEEN ISOTROPIC POINT SOURCES IN DBI
C *************************************************************
C
      SUBROUTINE LOSS (FREQ,HEIGHT,HIMOB,RANGE,ENVIOR,PL,PL2)
      CHARACTER ENVIOR*8
C FIRST COMPUTE HATA URBAN
C EQUATIONS FROM HATA HAVE BEEN CONVERTED TO ENGLISH UNITS
      PL=86.65+26.16*ALOG10(FREQ)-15.17*ALOG10(HEIGHT)+
```

```
     + (48.28-6.55*ALOG10(HEIGHT))*ALOG10(RANGE)
C SUBTRACT HATA CORRECTION FOR MOBILE HEIGHT, URBAN = LARGE CITY ELSE
C OTHER ONE.   USE 300 MHz AS THE FREQUENCY BREAK POINT.
      IF(ENVIOR.EQ.'URBAN') THEN
        IF(FREQ.GT.300) THEN
          PL=PL-(3.2*(ALOG10(11.75*HIMOB*0.3048))**2-4.97)
        ELSE
          PL=PL-(8.29*(ALOG10(1.54*HIMOB*0.3048))**2-1.1)
        ENDIF
      ELSE
        PL=PL-(1.1*ALOG10(FREQ)-0.7)*HIMOB*0.3048+
     +   (1.56*ALOG10(FREQ)-0.8)
      ENDIF
C SUBTRACT HATA CORRECTION FOR OTHER ENVIRONMENTS
      IF (ENVIOR.EQ.'SUBURBAN') PL=PL-5.4-2*(ALOG10(FREQ/28)**2)
      IF (ENVIOR.EQ.'OPEN'.OR.ENVIOR.EQ.'QUASI O')
     +    PL=PL-40.94+18.33*ALOG10(FREQ)-4.78*(ALOG10(FREQ)**2)
      IF (ENVIOR.EQ.'QUASI O') PL=PL+5
C NOW EXTEND IT IF YOU ARE OVER THE RANGE LIMIT OR BASE HEIGHT LIMIT
      R1=12.4
      R2=40.0
C FOR ALL RANGES GREATER THAN 12.4 MILES (20 km) ADD A FACTOR
      IF (RANGE.GT.R1) THEN
        PL=PL+(0.5+0.15*ALOG10(HEIGHT/400))*(RANGE-R1)
      ENDIF
C FOR ALL RANGES GREATER THAN 40 MILES (64.38 km) SUBTRACT A FACTOR
      IF (RANGE.GT.R2) PL=PL-0.28*(RANGE-R2)
C FOR ALL BASE HEIGHTS GREATER THAN 984 FEET (300 M) SUBTRACT A FACTOR
      IF (HEIGHT.GT.984) THEN
        PL=PL-4.7*ABS(ALOG10(6.2/RANGE))*(HEIGHT-984)/1968
      ENDIF
C MAKE THE EQUATIONS THAT WORK FOR 1500 MHz GO DOWN TO 30 MHz
      PL=PL-(FREQ/250)*ALOG10(1500/FREQ)
      IF(RANGE.GT.25) PL=PL-0.18*ALOG10(1500/FREQ)*(RANGE-25)
C COMPUTE FREE SPACE PATH LOSS IN DBI
      PL2=36.6+20*ALOG10(FREQ)+20*ALOG10(RANGE)
      RETURN
      END
```

5.2 Anderson 2D Model

The Anderson 2D model is a comprehensive point-to-point radio propagation model for predicting field strength and path loss in the frequency range of 30 MHz to 60 GHz. This model draws upon techniques which have been successfully used for many years, such as those described in NBS Technical Note 101 [1], and improves upon them by making use of widely available terrain elevation and local land use (ground cover) databases. As described in Section 5.5, this model can also be extended to provide for the first time 3D modeling of reflections from terrain features which are not along the great circle path between the transmitter and the receiver. Such reflections result in multipath and time-dispersed signal energy at the receiver. Such an extension is important for prediction the performance of certain digital systems where time-dispersed reflections are a primarily cause of irreducible data errors due to inter-symbol interference (ISI).

The model specification is divided into several sections which describe its various components. Section 5.2.1 is a basic model outline which describes how the components fit together. Sections 5.2.2 and 5.2.3 define the model for the line-of-sight (LOS) and non-line-of-sight (NLOS), respectively. Section 5.4 discusses the local clutter attenuation and the uses of the land use/ land cover database to incorporate this attenuation. The Anderson 2D model also includes a troposcatter mode for long-range over the horizon path loss prediction, and atmospheric absorption loss which is relevant at frequencies above 10 GHz. For the systems for which this Report is intended, the troposcatter mode and atmospheric absorption loss are not applicable and will not be described here.

The level of detail in this specification is in keeping with scientific standards. Equations and specific information are provided such that knowledgeable researchers in the field can replicate the model in computer code and reproduce the model results. However, no computer code or pseudo code is provided here since approaches to implementation can vary widely.

The Anderson 2D model is supported for administrative purposes in spectrum management and regulation. As such it has been designed to take into account the more important elements of propagation prediction while still remaining simple enough so that computer implementation is straightforward and the model can be broadly applied.

An important objective in designing the model described in this document was to make it simple and thereby accessible. In keeping with this objective, however, it is recognized that the defined model is not the most complete possible solution to predicting electromagnetic (EM) fields in a complex propagation environment. Other approaches such as the Integral Equation (IE) and Parabolic Equation (PE) methods could potentially provide more accurate full-wave solutions but with attendant limitations and a substantial increase in complexity. The model defined in this document relies on the geometric optic (ray-tracing) approach which basically deals with the transport of EM energy from location to another. It is an easy technique to visualize, and conceptually it is readily adapted to the 3D extension for predicting multipath and time-dispersion. Attempting to use IE or PE techniques in a full 3D mode for this purpose would be a daunting computational task, even on the largest computers.

5.2.1 Propagation Model Outline

For the purposes of this report, the Anderson 2D model has three basic elements which affect the predicted field strength at the receiver as follows:

1) Line-of-Sight (LOS) mode using basic two-ray theory with constraints

2) Non-line-of-sight (NLOS) mode using multiple wedge diffraction

3) Local clutter attenuation (see Section 5.4 of this report)

The LOS and NLOS modes are mutually exclusive - a given path between a transmitter and receiver is either LOS or not. The local clutter loss is an integral part of this model which is necessary to achieve correct signal level predictions in suburban, urban, and forested areas. It is describe separately in Section 5.4 of this Report.

The fundamental decision as to whether a path is LOS is based on the path geometry. It is described in the next section which defines the LOS mode for this model.

5.2.2 Line-of-Sight (LOS) Mode

The determination of whether a path between transmitter and receiver is LOS is done by comparing the depression angle of the path between the transmitter and receiver with the depression angle to each terrain elevation point along the path. The depression angle from transmitter to receiver is computed using an equation of the form of (6.15) in [15]:

$$\theta_{t-r} = \frac{h_r - h_t}{d_r} - \frac{d_r}{2a} \qquad \text{[Eq. 14]}$$

where:

θ_{t-r} is the depression angle relative to horizontal from the transmitter to the receiver in radians

h_t is the elevation of the transmit antenna center of radiation above mean sea level in meters

h_r is the elevation of the receive antenna center of radiation above mean sea level in meters

d_r is the great circle distance from the transmitter to the receiver in meters

a is the effective earth radius in meters taking into account the atmospheric refractivity

The atmospheric refractivity is usually called the K factor. A typically value of K is 1.333, and using an actually earth radius of 6340 kilometers, a would equal 8451 kilometers, or 8,451,000 meters.

Using an equation of the same form, the depression angle from the transmitter to any terrain elevation point can be found as:

$$\theta_{t-p} = \frac{h_p - h_t}{d_p} - \frac{d_p}{2a} \qquad \text{[Eq. 15]}$$

where:

θ_{t-p} is the depression angle relative to horizontal for the ray between the transmitter and the point on the terrain profile

h_p is the elevation of the terrain point above mean sea level in meters

d_p is the great circle path distance from the transmitter to the point on the terrain path in meters

h_t and a are defined above

The variable θ_{t-p} is calculated at every point along the path between the transmitter and the receiver and compared to θ_{t-r}. If the condition $\theta_{t-p} > \theta_{t-r}$ is true at any point, then the path is considered NLOS and the model formulations in Section 5.2.3 are used. If $\theta_{t-p} \leq \theta_{t-r}$ is true at every point, then the transmitter-receiver path is LOS and the formulations in this section apply.

For LOS paths the field strength at the receiver is calculated as the vector combination of a directly received ray and a single reflected ray. This calculation is presented in Section 5.2.2.1. If the geometry is such that a terrain elevation point along the path between the transmitter and receiver extends into the 0.6 Fresnel zone, then an additional loss ranging from 0 to 6 dB is included for partial Fresnel zone obstruction. This is discussed in Section 5.2.2.2.

5.2.2.1 Two-Ray Field Strength at the Receiver Using a Single Ground Reflection

For an LOS path, the field at the receiver consists of the directly received ray from the transmitter and number of other rays received from a variety of reflecting and scattering sources. For low antenna heights (on either the transmit or receive end of the path) the field at the receiver is dominated by the direct ray and a single reflected ray which intersects the ground near the transmitter. The *height-gain function* in which at field at the antenna increases as the height of the antenna above ground increases is a direct result of the direct and ground refection rays vectorially adding so that the magnitude of the resultant manifests this effect. The height-gain function is modeled here by considering the actual ground reflected ray and direct ray in vector addition. The magnitude of the direct ray is given by:

$$E_r = \frac{1}{d_r}\sqrt{\frac{P_T G_T \eta}{4\pi}} \qquad \text{[Eq. 16]}$$

where E_r is the field strength at the receive point, P_T is the transmitter power delivered to the terminals of the transmit antenna, G_T is the transmit antenna gain in the direction of the receiver point (or ray departure direction), η is the plane wave free space impedance (377 ohms), and d_r is the path distance from the transmitter to the receive point in kilometers.

Written in dB terms, this reduces to the familiar:

$$E_r = 76.92 - 20.0\log(d_r) + P_T \quad \mathrm{dB\mu V / m} \qquad \text{[Eq. 17]}$$

In [Eq. 17], P_T is effective radiated power (ERP_d) in dBW. The magnitude and phase of the ground-reflected ray is found by first calculating the complex reflection coefficient as follows:

$$R = R_s g \qquad \text{[Eq. 18]}$$

where R_s is the smooth surface reflection coefficient and g is the surface roughness attenuation factor (a scalar quantity).

For parallel and perpendicular polarizations, respectively, the smooth surface reflection coefficients are:

$$R_{s\|} = \frac{\sin\gamma_0 - \sqrt{\varepsilon - \cos^2\gamma_0}}{\sin\gamma_0 + \sqrt{\varepsilon - \cos^2\gamma_0}} \qquad \text{parallel polarization} \qquad \text{[Eq. 19]}$$

$$R_{s\perp} = \frac{\varepsilon\sin\gamma_0 - \sqrt{\varepsilon - \cos^2\gamma_0}}{\varepsilon\sin\gamma_0 + \sqrt{\varepsilon - \cos^2\gamma_0}} \qquad \text{perpendicular polarization} \qquad \text{[Eq. 20]}$$

where γ_0 is the angle of incidence and ε is the complex permittivity given by:

$$\varepsilon = \varepsilon_1 - j60\sigma_1\lambda \qquad \text{[Eq. 21]}$$

where ε_1 is the relative dielectric constant of the reflecting surface, σ_1 is the conductivity of the reflecting surface in Siemens/m, and λ is the (free space) wavelength of the incident radiation. For the case of a ground reflection, vertical polarization is parallel polarization and horizontal polarization is perpendicular polarization.

For the model defined here, it will be assumed that the local surface roughness is 0 (smooth surface) so that the term *g* in [Eq. 18] is one. Also, values of σ_1 = 0.008 Siemens/meter and that ε_i = 15 are commonly used for ground constants.

Since the length of the reflected path and the direct path is essentially the same (differing by only a few wavelengths or less), the amplitude of the two rays due to spatial attenuation (path length) is the assumed to be the same. The reflected ray, however, is multiplied by the reflection coefficient as given above and then shifted (retarded) in phase as a result of the longer path length compared to the direct ray. The vector addition of the two rays at the receiver is thus:

$$E_r = E_d \sin(\omega t) + E_d R \sin(\omega t + \Delta\varphi) \qquad \text{[Eq. 22]}$$

where:

E_d is the magnitude of the direct ray
ω is the carrier frequency in radians
R is the complex reflection coefficient given above
$\Delta\varphi$ is the phase delay of reflected ray in radians

The carrier term is usually suppressed so that [Eq. 22] becomes:

$$E_r = E_d\left(1 + |R|\angle(\varphi_r + \Delta\varphi)\right) \qquad \text{[Eq, 23]}$$

where φ_r is the phase angle of the reflection coefficient. The term $\Delta\varphi$ is found from the actual path length difference in meters. For a two-ray path geometry over a curved earth, the path length difference as given by (5.9) in [15] as:

$$\Delta r = \frac{2h_t' h_r'}{d_r} \qquad \text{[Eq. 24]}$$

where:

h_t' is the height of the transmit antenna *above the reflecting plane* in meters

h_r' is the height of the receive antenna *above the reflecting plane* in meters

so that:

$$\Delta\varphi = \frac{2\pi\Delta r}{\lambda} \quad \text{(modulo } 2\pi \text{ radians)} \qquad \text{[Eq. 25]}$$

The usual issue in using this approach is defining where the reflecting plane is for a complex terrain profile between the transmitter and receiver.

For the Anderson 2D model the reflection point is found by evaluating the angle of incidence and reflection at every terrain elevation point between the transmitter and receiver. The angle of incidence at any point along the profile (the evaluation point) is found from simple geometry as follows:

$$\gamma_t = \tan^{-1}[(h_t / d_t] \qquad \text{[Eq. 26]}$$

for the transmitter, and

$$\gamma_r = \tan^{-1}[(h_r / d_r] \qquad \text{[Eq. 27]}$$

for the receiver. The terms h_t, h_r, d_t, and d_r are the transmitter height above the evaluation point, the receive antenna height above the evaluation point, and the distances for the evaluation point to the transmitter and receiver, respectively. The evaluation point where $\gamma_t = \gamma_r$ is considered the reflection point. However, it is unlikely that these angles with ever be exactly equal. In such cases, at the two adjacent evaluation points where the angles inflect (i.e. γ_r becomes larger than γ_t), the reflection point is considered to exist along the profile segment defined by the adjacent points. The exact reflection point is then found along this profile segment using linear interpolation since the profile segment is by definition a linear slope. With the distance and elevation of the reflection point established, the reflection angle of incidence γ_0 is found using an equation of the form of [Eq. 25]. This value of γ_0 is then used in [Eq. 18] and [Eq. 19] to find the magnitude and phase of the reflection coefficients.

The effect of the nearby ground reflection will be to reduce the amplitude of the directly received ray because in general they will add out of phase and the amplitude of the reflected ray will be nearly equal to the direct ray because at low reflection angles of incidence, $|R| \cong 1.0$ for most practical combinations of frequency, conductivity, and permittivity. For an antenna placed very near the ground, the cancellation based on these formulas will be almost perfect so that the direct received (free space) ray will be reduced by 40 dB or more. However, it is unlikely that such a perfect cancellation will occur, therefore it is appropriate to put some reasonable limits on the change in the amplitude of the directly-received ray which can occur due to a reflection. Based upon measurement and theoretical data, the limits on the change in the free space amplitude due the reflection contributions will be -25 dB and + 6 dB.

Thus based on the preceding discussing, the path loss or attenuation term $A_{reflection}$ can be written as:

$$A_{reflection} = 20.0 \log \left[\left(1 + |R| \angle \left(\varphi_r + \Delta\varphi \right) \right) \right] \quad \text{dB} \qquad \text{[Eq. 28]}$$

with the limits that $-6.0\ \text{dB} \leq A_{reflection} \leq 25.0\ \text{dB}$.

5.2.2.2 Attenuation Due to Partial Obstruction of the Fresnel Zone

When a path is LOS but terrain obstacles are close to obstructing the path, additional attenuation will occur which cannot be accounted for using the ray approach from geometric optics. This is because the geometric optics only deals with energy transport, not phase. As such, the frequency is infinite and the wavelength is zero. With zero wavelength, the Fresnel zone radius is also zero. The failure of the ray approach to account for attenuation due to a "near miss" of obstacles on the path can be overcome to some extent by including a loss term in the LOS formulation which is based on the extent to which an obstacle penetrates the first Fresnel zone. From diffraction theory, when the ray just grazes an obstacle, the field on the other side is reduced by 6 dB (half the wavefront is obstructed). When the clearance between the obstacle and the ray path is 0.6 of the first Fresnel zone, the change in the field strength at the receiver is 0 dB, and with additional clearance a field strength increase of 6 dB can occur owing to the in-phase contribution from the ray diffracted from the obstacle. For additional clearance, an oscillatory pattern in the field strength occurs, as conveniently illustrated by Figure 7.1 in [15].

For the Anderson 2D model, if the ray path clears intervening obstacles by at least 0.6 of the first Fresnel zone, then no adjustment to the receiver field will occur. For the case when an obstacle extends into the 0.6 first Fresnel zone, a loss factor ranging from 0 to 6 dB will be applied based on a linear proportion of how much of the 0.6 First Fresnel zone is penetrated. This Fresnel zone path loss or attenuation term can be written as:

$$A_{Fresnel} = 6.0\left(\frac{C_{obs}(d_p)}{R_{FR}(d_p)}\right) \text{ dB} \qquad \text{[Eq. 29]}$$

where:

$C_{obs}(d_p)$ is the height difference in meters between the ray path and the terrain elevation at distance d_p along the path

$R_{FR}(d_p)$ is the 0.6 first Fresnel zone radius at distance d_p along the path

The values $C_{obs}(d_p)$ and $R_{FR}(d_p)$ are calculated taking into account the effective earth radius using the K factor. The 0.6 first Fresnel zone radius is given by:

$$R_{FR}(d_p) = 0.6\left[549.367\sqrt{\frac{d_p(d_r - d_p)}{f\,d_r}}\right] \text{ meters} \qquad \text{[Eq. 30]}$$

where f is the frequency in MHz and all distances are in kilometers.

The use of the partial Fresnel zone obstruction loss from 0 dB at 0.6 clearance to 6 dB at grazing also provides a smooth transition into the NLOS mode in which knife-edge diffraction loss just below grazing will start at 6 dB and increase for steeper ray bending angles to receive locations in the shadowed region. Note that this attenuation factor is found only for the terrain profile point which extends farthest into the 0.6 first Fresnel zone, not for every profile point which extends into the Fresnel zone.

5.2.2.3 Summary of the Calculation of the Field Strength at the Receiver Under LOS Conditions

All of the formulations for computing the field strength at the receiver under LOS conditions are now in place. They can be summarized with the following simple equation:

$$E_r = 76.92 - 20\log(d_r) + P_T - A_{reflection} - A_{Fresnel} - A_{clutter} \text{ dBuV / m} \qquad \text{[Eq. 31]}$$

where $A_{reflection}$ is the change due the reflection in dB from [Eq. 28], $A_{Fresnel}$ is the partial Fresnel zone obstruction loss from [Eq. 29]. $A_{clutter}$ is a local clutter loss number which will range from 0 dB to 17 dB as discussed in Section 5.4 and as shown in Table 12. The term P_T is the effective radiated power (ERP_d) in dBW in the direction of the receiver.

In terms of path loss between two antennas with gains of 0 dBi in the path direction, [Eq. 31] can be written as:

$$L_{LOS} = 32.45 + 20.0\log f + 20\log d_r + A_{reflection} + A_{Fresnel} + A_{clutter} \quad \text{dB} \qquad \text{[Eq. 32]}$$

5.2.3 Non-Line-of-Sight (NLOS) Mode

The decision on when to use the LOS mode and when to use the NLOS modes was set forth at the beginning of Section 5.2.2. If the model has elected to use the NLOS formulations, it means that one or more terrain or other features obstructs the ray path directly from the transmitter to the receiver. In this case, the free space field strength is further reduced for the attenuation cause by the obstacles. For the model defined here, the calculation of obstruction loss over an obstacle will be done by assuming the obstacle is a perfect electrical conductor (PEC) rounded obstacle with a height equal to the elevation of the obstruction and a radius equal to 1 meter. Diffraction loss in this model is calculated assuming individual obstacles on the path can be modeled as isolated rounded obstacles. The loss from each isolated obstacle is then combined using the Epstein-Peterson technique [10] as extended to more than two obstacles. The NLOS mode also include loss for partial Fresnel zone obstruction due to sub-path obstacles along the path from the transmitter to the first obstacle and from the last obstacle to the receiver. This partial Fresnel zone obstruction losses are found exactly as described in Section 5.2.2.2.

5.2.3.1 Diffraction Loss

The loss over an individual rounded obstacle is computed using the formulas taken from [1]. It is primarily a function of the parameter ν which is related to the path clearance over the obstacle. The total diffraction loss, $A(v,\rho)$, in dB is the sum of three parts $-A(v,0), A(0,\rho)$, and $U(v,\rho)$ The equations to calculate each part are given below:

$$A(v,\rho) = A(v,0) + A(0,\rho) + U(v,\rho) \quad \text{[Eq. 33]}$$

$$A(v,0) = 6.02 + 9.0v + 1.65v^2 \quad \text{for } -0.8 \le v \le 0 \quad \text{[Eq. 34]}$$

$$A(v,0) = 6.02 + 9.11v - 1.27v^2 \quad \text{for } 0 < v \le 2.4 \quad \text{[Eq. 35]}$$

$$A(v,0) = 12.953 + 20\log_{10}(v) \quad \text{for } v > 2.4 \quad \text{[Eq. 36]}$$

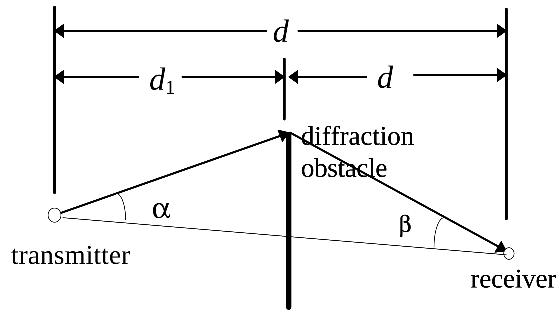
$$A(0,\rho) = 6.02 + 5.556\rho + 3.418\rho^2 + 0.256\rho^3 \quad \text{[Eq. 37]}$$

$$U(v,\rho) = 11.45v\rho + 2.19(v\rho)^2 - 0.206(v\rho)^3 - 6.02 \text{ for } v\rho \le 3 \quad \text{[Eq. 38]}$$

$$U(v,\rho) = 13.47v\rho + 1.058(v\rho)^2 - 0.048(v\rho)^3 - 6.02 \quad \text{for } 3 < v\rho \le 5 \quad \text{[Eq. 39]}$$

$$U(v,\rho) = 20v\rho - 18.2 \quad \text{for } v\rho > 5 \quad \text{[Eq. 40]}$$

where the curvature factor is:



$$\rho = 0.676 R^{0.333} f^{-0.1667} \sqrt{\frac{d}{d_1 d_2}}$$ [Eq. 41]

The obstacle radius *R* is in *kilometers*, and the frequency *f* is in MHz. The distance term *d* is the path length from the transmitter (or preceding obstacle) to the receiver (or next obstacle), d_1 is the distance from the transmitter (or preceding obstacle) to the obstacle, and d_2 is the distance from the obstacle to the receiver (or next obstacle). When the radius is zero, the obstacle is a knife edge and $A(v,\rho) = A(v,0)$.

The parameter ν in the above equations takes into account the geometry of the path and can be thought of as the bending angle of the radio path over the obstacle. It is computed as:

$$v = \sqrt{\frac{2d \tan(\alpha)\tan(\beta)}{\lambda}}$$ [Eq. 42]

where *d* is the path length from the transmitter (or preceding obstacle) to the receiver (or next obstacle), α is the angle relative to a line from the transmitter (or preceding obstacle) to the receiver (or next obstacle), and β is the angle relative to a line from the receiver (or next obstacle) to the transmitter (or preceding obstacle). The definition of α and β are shown if Figure 8. For the multiple obstacle case, obstacles are treated successively as transmitter-obstacle-receiver triads to construct the path geometry and bending angle ν over each obstacle. The value of *v* is then used to calculate the diffraction loss over each obstacle. The resulting obstacle losses are summed to arrive at the total obstacle diffraction loss for the path.

Figure 8. Geometry for computing ν

5.2.3.2 Handling Anomalous Terrain Profiles

Experience with terrain elevation databases covering the United States has shown that occasional anomalous profiles can be produced. A typical example is a “false” plateau in which the several adjacent data points all have the same or nearly the same value, and that value is usually exactly equal to a contour elevation line (like 400 or 600 feet) on the original 1:250,000 scale maps from which the original database was developed. Under LOS conditions these plateaus are usually not a problem but if they form an obstruction to the ray between the transmitter and the receiver, using the Epstein-Peterson type geometry, it may occur that every point on the top of the plateau appears to be an obstacle. The result is a string of diffracting points, many with grazing incidence, and a predicted diffraction attenuation in excess of what would actually occur. The following method is included in the Anderson 2D model for detecting and dealing with such anomalies.

When the model finds more that two consecutive points along the terrain profiles are obstacles using the geometry described above, it ignores all the intermediate obstacles. Instead, it preserves the obstacles at the beginning and at the end of the sequence as two rounded obstacles with a radius of 1 meter and calculates the diffraction loss over each as described above. In urban ray-tracing models, using this two-edge diffraction approach is common for computing ray attenuation over real plateau-like features such as buildings, provided slope diffraction coefficients are used at the second edge. For terrain profiles, this approach provides a simple way of resolving the anomalies which will also be approximately correct for real plateau obstacle features along the path.

5.2.3.3 Summary of the Calculation of the Field Strength at the Receiver Under NLOS Conditions

The field strength at the receiver in the NLOS mode can then be written as:

$$E_r = 104.77 - 20\log(d_r) + P_T - A_{diff} - A_{T,Fresnel} - A_{R,Fresnel} - A_{clutter} \text{ dBuV / m} \quad \text{[Eq. 43]}$$

where all the terms have the same definitions as given in Section 5.2.2.3 and the term A_{diff} is defined as:

$$A_{diff} = \sum_{n=1}^{\mathrm{NOBS}} A_n(\nu,\rho) \quad \text{dB} \qquad \text{[Eq. 44]}$$

where $A(\nu,\rho)$ is defined in [Eq. 33]. The terms $A_{T,Fresnel}$ and $A_{R,Fresnel}$ are the partial Fresnel zone obstruction attenuations on the path segments from the transmitter to the first obstacle, and from the last obstacle to the receiver, respectively, as described above.

The corresponding path loss between antennas with 0 dBi gain in the path direction can be written as

$$L_{NLOS} = 32.45 + 20.0\log f + 20\log d_r + A_{diff} + A_{T,Fresnel} + A_{R,Fresnel} + A_{clutter} \quad \text{dB} \quad \text{[Eq. 45]}$$

5.3 Terrain Elevation Database

The propagation prediction model defined in this specification inherently depends on the terrain database to compute the effective base antenna height for use in Section 5.1 and for the geometry computations for the shadow loss formulations in Section 5.2. In the United States, there are currently three terrain databases which are commonly used:

1. The 30 arc second National Geophysical Data Center (NGDC) database

2. The 3 arc second (U.S. Geological Survey (USGS) or Defense Mapping Agency (DMA)) database

3. The 30 meter (USGS) database

The 30 second database is primarily used by the FCC and those filing FCC applications to determine 2-10 miles (3-16 km) average terrain along radials emanating from a transmitter site for the purpose of determining the location of coverage of interference signal contours. Because of its wide point spacing (nearly 1 km), its use for more detailed propagation studies is not common.

The 3 arc second database is the one most commonly used for propagation studies. Its point spacing of about 90 meters north-south by an average point 70 meters east-west seems appropriate for many planning purposes, especially when wide-area systems with service radii of 50 km or more are being considered. Considering coverage and interference on a grid with spacings less than 100 meters is rarely necessary. The 3 arc second database is also a convenient size for use on personal computers since with reasonable compression techniques the entire database can fit and be used from an inexpensive CD-ROM drive.

The main drawback to the 3 arc second database is its vertical accuracy. For the most part it was derived from the 1:250,000 series of maps covering the US. Most of these maps have contour intervals of 200 feet. The result is that many ridges and hills with peak elevations that lie between 200 foot contour intervals are not properly represented. Even some peaks where USGS benchmarks are shown on the maps were not properly digitized. Occasionally, elevation errors occur, some as great as 200 meters.

The 30 meter data contains elevation data points spaced at 30 meter intervals rather than intervals based on latitude and longitude. Its development has been a on-going effort by the USGS over the last several years. It is fundamentally derived using contour and other information from the 7.5 minute quadrangle series maps which cover the US. Since the data source is a much larger scale the data source for the 3 second database, the vertical accuracy achieved is significantly better. Unfortunately, at this time the data files for only slightly more than 60% of the USA have been completed and released. New data files are released on a monthly basis.

Despite these uncertainties, the much improved vertical accuracy of the 30 meter data warrants consideration for a development effort of an up-to-date propagation model. With incomplete coverage, some techniques will need to be developed to handle transitions from the 30 meter data to the 3 second data which does cover the entire country. Smoothing individual terrain profiles at the transition is one possibility. Another, much more extensive, approach is to meld or re-grid the 30 meter data into the existing 3 second format. The net result of this effort would be a database with 3 second point spacings (sufficiently close as explained above) but with the improved vertical accuracy of the 30 meter data. Re-gridding data in this way is computationally extensive but straightforward. It is similar to the effort required when developing a terrain elevation database from information which has been digitized from a topographic map. The raw digitized data consists of a collections of flat lines, slope lines, and points. By considering all this information together, a smooth, internally consistent set of Digital Elevation Model (DEM) matrices can be produce with essentially arbitrary grid point spacings.

Because of its nationwide coverage, for the model defined here, the 3 arc second database will be the fundamental terrain database used. Where 30 meter data is available, its use is preferred. The calculation results based on the use of 30 meter data *along the entire transmitter-receiver path* shall take precedence over the those based on the 3 arc second data when there is a dispute about model prediction results.

5.3.1 Establishing Terrain Elevation Points Along a Profile Using the Terrain Database

In practice the model requires a terrain elevation profile to be defined between the transmitter and the receiver. This profile is fundamental to the path loss prediction techniques in Sections 5.1 and 5.2. The elevation points on this profile are to be extracted from the terrain database by first determining the great circle path from the transmitter to the receiver. Spacing between adjacent data points shall not exceed 0.2 km or 0.2% of the path length, whichever is less. Either method can be used regardless of the horizontal resolution of the database. Either of the following extraction techniques is acceptable.

5.3.1.1 Bilinear Interpolation

A profile elevation point spacing is selected. At a point some distance *d* from the transmitter along the great circle path where the profile elevation is to be found, the latitude-longitude or other coordinates of the point (the lookup point) are determined using double precision spherical trigonometry. These coordinates are then used to find the four surrounding elevation points; linear interpolation is used to establish the elevation at the lookup point. This process is used to find the elevation at each of the points along the profile from the transmitter to the receiver.

5.3.1.2 "Snap to Nearest Point" Method

The equation of the line segment between the transmitter and the receiver is established. Using conventional spherical trigonometry techniques, the distances from all points to the line are determined. The elevations of all points within 0.5x (the horizontal resolution of the database) are used. Their corresponding horizontal positions along the profile are the crossing points of perpendiculars from the points to the line. This method produces profiles with unequal horizontal spacings, but the results produce equally valid results as those using the method described above.

5.4 Local Clutter Loss Attenuation Standard Values

The path loss predictions in Sections 5.1 and 5.2 can be improved by applying a local clutter loss factor. Whether an urban, suburban, or foliage loss should be applied is determined by a land use or groundcover type associated with the receiver location. The land use/land cover (LULC) database which is currently available for the US comes from the USGS. It is actually available in two forms - as vector data describing the boundaries of land use region types, and as composite theme grids (CTG) files in which 1 of 37 land use types have been assigned to 200 meter square grid cells covering the entire country. The CTG files are the ones which are generally used with propagation models.

This database also has shortcomings. Much of the information was taken from 1:250,000 scale maps which limits its resolution. It also can be rapidly dated as new construction turns farms into subdivisions and factories. But at this time, the USGS LULC database is the only one that is readily available and for that reason, it is specified that it be used in this model for the purpose of determining the local clutter classification at each receive point. If a new database, or better database information becomes available through LandSat or other sources, then that better data can supersede the LULC data.

With the exception of categories 11-17, the remaining land use classifications in the LULC database are much too fine-grained for radio propagation use. Table 11 shows a recommended way of reducing the 37 classifications to 10. Table 12 shows the value of $A_{clutter}$ to be used for each of the reduced classifications as a function of frequency. For frequencies not shown in the table, linear interpolation in dB is to be used between the values shown in the table. To use the values in Table 12 with the programs in Section 5.1.2, the “open” mobile environment must be chosen. Choosing other mobile environments will result in predictions showing signals weaker than actual.

5.5 The Anderson 3D Propagation Methods for Time Dispersion and Multipath Predictions

It is well know that in hilly and mountainous areas reflections from terrain features can cause signal echoes to arrive at the receiver some time after the signal received directly from the transmitter. If the directly-received signal is weak due to shadowing, these reflections can actually be of comparable or higher amplitude than the directly-received signal. Such multipath effects are commonly recognized as “ghosting” in broadcast television. Depending on the data rate, for some types of digital transmission systems in mobile communications these echoes can have an important effect on bit error rate (BER) performance of the system [16].

The prediction of multipath time dispersion from terrain features is a relatively new field of research. As will be discussed, many specific parameters controlling the performance of the model cannot yet be established with the limited amount of field measurement data which is available. Therefore, the intent of this section is to outline an approach to prediction multipath time dispersion in a 3D terrain environment with sufficient detail that others can reproduce the method if desired. No equations are provided here.

The method described here can be viewed as an extension of the 2D transmitter-receiver path profile path loss prediction analysis set forth in Sections 5.1 and 5.2.

The first step in the method is to determine the terrain surrounding the base station which is illuminated by the base station. This can be done by finding the terrain elevations in a grid surrounding the base station out to distance beyond that where the signal dispersion is actually required. This is because terrain features beyond the coverage area of interest may reflect back into the area of interest. An initial test value considers terrain at distances 50% greater than the service area distance of interest.

Within this grid area, elevation points are extracted from the database at 0.5 or 1 km intervals. The four corner elevation points form a square through which a plane can approximately be constructed using standard analytical geometry. The vector normal to the plane defines its orientation. The normal to the plane is found as the cross product of any two vectors in the plane. The vectors can be those connecting two of the corner grid points, for example, if they lie on the plane.   The normal vector is resolved in to x,y,z components. The ratio of the z component to the overall magnitude of the normal vector is inversely proportional to the slope of the plane. The first sorting process after constructing the terrain grid is to evaluate the magnitude of the slope for each plane and determine whether it has sufficiently slope to be further considered as a reflecting source.   The threshold for such consideration has not yet been established.

For those planes or patches which have sufficient slope, the next step is to find the angle between the normal vector and the vector from the patch back to the base station. If the angle of this vector is less that 90 degrees, the patch is oriented such that it is illuminated to some degree. Angles greater than 90 degrees indicate the patch is oriented away from the base station and will not be illuminated. As a practical matter, the difference angle should be substantially less than 90 degrees so that the patch has sufficient cross-sectional area to intercept and reflect a relevant amount of energy.

At this point all the patches have been sorted so that only those with sufficient slope and which “face” the transmitter are still considered. For each of these patches a terrain profile is constructed from the transmitter to the center of the patch to determine whether it is line-of-sight (LOS) or shadowed. If it is shadowed, it is excluded from further consideration since it probably won’t have a strong enough incident field strength to produce important reflections. In mountainous areas with sheer rock faces, however, this assumption is probably not valid.

All the LOS patches determined to this point are illuminated by the base station. The next step is to considered each receive location in a grid or other study configuration, and at each such point, determine if the geometry of reflection is such that a reflected signal will illuminate the receiver. This is done using the vector from the patch to the base station, the vector from the patch to the receive location, and the vector normal to the patch. If the geometry is such that a reflection could be important, then the final step is to find the terrain profile from the patch to the receiver to see if it is LOS. If is shadowed, it will be assumed (as above) that the reflection will not be important at the receiver. If it is LOS, than the reflected signal amplitude can be found at the receiver. Some available empirical data by Dreissen [11] suggest that the return loss on the reflection is on the order of 20 to 30 dB. Therefore, a first estimate of the reflected signal amplitude at the receiver would be the amplitude the signal would have had given an LOS path of equivalent transmitter-reflector-receiver length, but reduced by 20 to 30 dB. The reflected echo transit time is easily found using the total transmitter-reflector-receiver path length. The time delay of the echo is this time minus the time for the directly received signal to arrive given its (shortest) path length.

Using the reflected echo amplitudes and time delays, a power delay profile can be constructed and the RMS delay spread determined (Section 6.7). The RMS delay spread can then be used to estimate BER values for various digital systems.

The mechanics of the described method have been implemented in a software model. An ensemble of multipath echoes at the received occur as expected. However, to breathe real life into the model, all those places in the model description where words like “substantial” and “significant” need to be replaced by quantifiable thresholds. Further development of the model will be directly toward establishing these thresholds.

A similar technique for 3D terrain scattering for radio propagation prediction purposes, including mathematical details, can be found in reference [12].

5.6 Propagation Modeling and Simulation Benchmarks

The following referenced path profiles and tabulated path losses are to serve as benchmark results of the propagation prediction model. Those interested in creating computer implementations of the model described in this section can use these tests to verify their implementation.

From the NBS measurement program reported by McQuate *et al* [17] and studies by Hufford [20], path numbers:

R1-20-T1	R2-10-T3	T1-10-R1
R1-20-T3	R2-10-T4	T1-10-R3
R1-20-T7	R2-10-T7	T1-10-R6
R1-50-T4	R2-20-T5	T1-20-R5
R1-50-T5	R2-20-T8	T1-80-R7
R1-50-T6	R2-20-T9o	T4-50-R7
R1-50-T7	R2-50-T3	T5-20-R7
R1-50-T8	R2-50-T4	T6-10-R2
R1-50-T9	R2-50-T5o	T7-80-R6o
R1-80-T1	R2-120-T2	
R1-120-T5	T1-5-R1	

The exact endpoint coordinates for these paths are contained in [17] [20]. Measured path loss as a function of receive antenna height above ground and at several frequencies are shown on graphs in [17] [20].

Because the paths in [17] [20] are for antenna heights relatively low to the ground, and no over-water nor desert paths are involved, a few additional paths are included for the model benchmark. Five paths start from Mt. San Bruno (North latitude 37° 44' 22", West 122° 26' 10") south of San Francisco where several transmit facilities on different frequencies are in operation. The model benchmark paths from Mt. San Bruno are azimuths 0, 45, 90, 135, and 180 degrees out to a distance of 50 km. Measured path loss data at different frequencies are not yet available for the Mt. San Bruno paths.

Two paths start from Onyx Peak (North latitude 34° 11' 31", West longitude 116° 42' 29" elevation 2778 m AMSL) east of Los Angeles, where several transmit facilities on different frequencies are in operation. The model benchmark paths are azimuths 41 and 95 degrees out to a distance of 90 km. Measured path loss data at different frequencies are not yet available for the Onyx Peak paths.

5.7 Recommendations Concerning Tiled vs. Radial Metaphors

A number of possibilities exist for defining the plane of the service area. The most widely used are the following:

- The Radial method
- The Stepped Radial method
- The Grid Mapped from Radial Data method
- The Tiled Method

5.7.1 Radial Method

In the radial method, many radials are drawn at equal angular intervals from the site to the far edge of the service area. Elevation points are extracted from the database at intervals along each radial. Each point represents an annular segment of service area. Since the radials get farther and farther apart as the distance from the site increases, care must be taken to ensure that the number of radials is sufficient to adequately characterize the area near the outer edge.

5.7.2 Stepped Radial Method

In the stepped radial method, the angular interval is stepped with distance. For example, in the CSPM [13] method, 8 radials are drawn from 0 to 2 km, 16 radials for 2 to 4 km, and so on up to 2,048 radials at distances of greater than 128 km. This results in a distance between radial ends not exceeding 1.57 km for all distances up to 256 km. Once again, each point along a radial represents an annular segment.

5.7.3 Grid Mapped from Radial Data Method

With this method, basic path loss information is calculated at points along radials as described in Section 5.7.1 or 5.7.2, and this information is then mapped into a uniform grid using linear or other interpolation methods. The derived signal levels at the grid locations can be then used for analyzing signals from multiple transmitters at common locations. This method combines the calculation speed advantages of radial methods over tiled methods, while still providing a common grid or tile structure for uniform multi-transmitter system analysis.

5.7.4 Tiled Method

In the tiled method, rectangular[2] tiles of a given size are predefined throughout the service area. Radials are drawn to each of these tiles. This results in unequal angular spacing and a greater number of required radials to predict signal levels in a given geographical area. The advantage is that a specific path loss calculation has been done to each tile centroid rather then being interpolated from nearby path loss calculation points.

5.7.5 Discussion of Methods

In predicting signal strength, only the radial method presents any kind of problem and, if the user is willing to increase the number of radials sufficiently, that problem can be averted. In predicting interference or simulcast performance, however, new problems arise. In the tiled method, all predictions from all sites are done to the same set of endpoints. Therefore, signal strength and delay spread prediction values can be calculated at those points. The grid mapped from radial data method provides a similar feature by using a set of interpolated endpoints.

Conversely, however, either radial method predicts to arbitrary endpoints. For a two-site system, the situation is not hopeless. The program must calculate the crossing points between the radials originating at the two sites and calculate its capture ratios, signal strengths, and delay spreads at those points. However, radial crossings become extremely far apart at angles approximating the azimuth between the two sites. Overall, the results of the radial approach to simulcast or interference prediction in a two-site system are mediocre at best.

In a system of three or more sites, the problem becomes more complicated. The tiled method still works well because the calculation points are predefined. The grid mapped from radial method also does the job. The radial method, however, becomes even more problematic. It is highly improbable that there will be ANY crossings that exist between radials from three or more sites. This means that any straight radial system cannot be used.

[2] In practice, the tiles may be squares or curvilinear trapezoids as well.

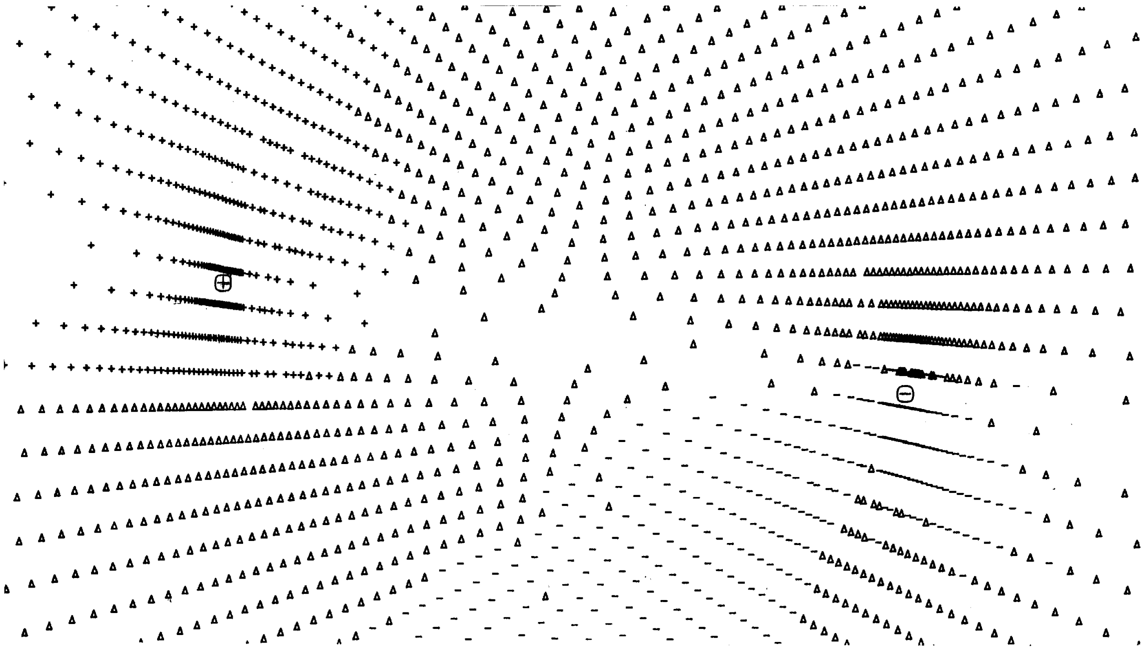


Figure 9. Radial Crossings in a 2-Site System

Notes to Figure 9:
Note 1: Figure is a randomly-selected capture ratio map
Note 2: Symbols:
Circled "+" = Site 1
"+" = Signals from Site 1 exceed those from Site 2 by predetermined ratio
Circled "-" = Site 2
"-" = Signals from Site 2 exceed those from Site 1 by predetermined ratio
"▲" = Capture ratio does not exceed predetermined value
Note 3: In the example, the "+" site is omni and the "-" is directional toward 240°

5.7.6 Summary and Recommendations

All four of the methods listed above can provide acceptable results for predicting signal strengths in the region around a single transmitter if proper consideration is given to the resolution of the study method and the objectives of the signal strength prediction. However, for simulcast, interference, best server, and other studies involving two or more transmitters, of the four methods listed, the grid mapped from radial method (Section 5.7.3) and the tiled method (Section 5.7.4) are best suited to providing acceptable results and are therefore recommended for such applications.

5.8 Reliability Prediction

The prediction of mean signal strength at a given location can vary from the measured signal for many reasons, including the following:

- Prediction algorithm not adequate
- Terrain database imperfections
- Land cover database imperfections
- Measurement made at slightly different location than prediction

Because of this, the signal at any one location can vary from that predicted by the model. It is recommended that a 1 dB margin be added for these "uncertainty" effects.

Additionally, signal variations due to land clutter tend to follow a lognormal distribution with a standard deviation of 5.6 dB. This value is applicable only when the terrain database recommendations of Section 5.3 are followed, including the local clutter database from Table 11 in Appendix-A and the shadow loss method of Section 5.2. A measurement error with a standard deviation of 1 dB is also included.

In determining the amount of extra margin to include, the user should set a required reliability level, and (because the only interest is in the signal equaling or exceeding a given value, rather than being in a given range) apply the "one-tailed" statistical test. Values of suggested margins for particular predicted reliabilities follow; these values are applicable only when the terrain database recommendations of Section 5.3 are followed:

Service Area Reliability	**Clutter Margin**	**Uncertainty Margin**	**Total Margin**
90 %	**7.2 dB**	**1.0 dB**	**8.2 dB**
95 %	**9.2 dB**	**1.0 dB**	**10.2 dB**
97 %	**10.5 dB**	**1.0 dB**	**11.5 dB**

No additional margin is required for time (temporal reliability). Time is considered to be 100%. This implies that measurements taken at different times over the same locations would produce similar results. Seasonal changes should be evaluated for worst case scenarios, such as trees losses with leaves rather than without.

5.9 Interference Calculations

Two methods of calculating interference from multiple lognormally-distributed sites are presented here: Monte Carlo simulation, and the "Equivalent Interferer" method. The Monte Carlo method can produce a more precise representation for the sum of lognormal interferers. However, for this application, the inherent accuracy of both methods is limited by the accuracy with which the constituent interference distributions are known.

5.9.1 Equivalent Interferer Method

If there is only one potential interferer, use its mean and standard deviation. If there are more than one, calculate the statistics of the "equivalent interferer" as follows:

1. $$\mu_j = 10^{\frac{m_{jdB}}{10}} \times \exp\left(\frac{\sigma_{jdB}\ln(10)}{20}\right) \qquad \text{[Eq. 46]}$$

$$D_j^2 = 10^{\frac{m_{jdB}}{5}} \times \left[\exp\left(\frac{\sigma_{jdB}\ln(10)}{5}\right) - \exp\left(\frac{\sigma_{jdB}\ln(10)}{10}\right)\right]$$

2. $\mu = \sum \mu_j$ [Eq. 47]

$D^2 = \sum D_j^2$

3. $\sigma_{nat}^2 = \ln\left(\dfrac{D^2}{\mu^2} + 1\right)$ [Eq. 48]

4. $m_{eq(nat)} = \ln(\mu) - \dfrac{\sigma_{nat}^2}{2}$ [Eq. 49]

$$m_{eq(dB)} = m_{eq(nat)} \times 10\log_{10}(e)$$

where:

$m_{jdB} \equiv$ The mean signal level of the j^{th} potential interferer in dB
$\sigma_{jdB} \equiv$ The standard deviation of the j^{th} potential interferer in dB
$m_{eq(dB)} \equiv$ The median strength of the equivalent interferer

Note: Use the same standard deviation for all interferers, except for the background noise level. Use a 0 standard deviation for the background noise.

If SIGN[$\tau/(2s_d)$] $\neq$ -1, substitute into the following equation:

$$R = 1 - 0.5erfc\left[\tau / \left(2s_d\right)\right]$$

where:

$\tau = m_d - m_{eq} - C/I_{req}$
i.e., the mean desired - equivalent interferer - required in C/I in dB
s_d = the standard deviation of the desired signal in dB, ***not*** the calculated value in natural units

If SIGN[$\tau/(2s_d)$] = -1, solve for R by substituting the absolute value of $\tau/(2s_d)$ for $\tau/(2s_d)$ in the equation for R, then by subtracting this result from 1.

Example of Equivalent Interferer Method:

Assume the following:

- A proposed analog FM system desiring DAQ-3 coverage.
- At a given location, the desired station has a signal strength of -75 dBm.
- Three potential interferers of -102, -108, and -111 dBm.

- Standard deviation of 5.7 dB.
- Noise for an ENBW of 16 kHz at 150MHz in a residential district.

m_{1dB} = -102	σ_{1dB} = 5.7	μ_1 = 121.6186E-12	D_1^2 = 401.6300E-22
m_{2dB} = -108	σ_{2dB} = 5.7	μ_2 = 30.5492E-12	D_2^2 = 25.3411E-22
m_{3dB} = -111	σ_{3dB} = 5.7	μ_3 = 15.3109E-12	D_3^2 = 6.3654E-22

Calculate m_{4dB}, the noise value, from Section 4.2.

m_{4dB} = -114	σ_{4dB} = 0	μ_4 = 3.9811E-12	D_4^2 = 0

$$\mu = \sum \mu_j = 171.4598E-12 \qquad D^2 = \sum D_j^2 = 433.3366E-22$$

$$\sigma_{nat}^2 = \ln\left[\frac{4.333366\times 10^{-20}}{\left(1.714598\times 10^{-10}\right)^2} + 1\right] = \ln\left(\frac{4.333366\times 10^{-20}}{2.939846\times 10^{-20}} + 1\right) = \ln(2.474011) = 0.9058408$$

$$m_{eq(nat)} = \ln\left(1.714598\times 10^{-10}\right) - \frac{.90958408}{2} = -22.93959 nats$$

$$m_{eq(dB)} = -22.93959\times 4.343 = -99.6 dB$$

Substituting into (5.9-2), m_{eq} = -99.6 dB

τ = -75 - (-99.6) - 17 = 7.6

Where -17 is the C/I value corresponding to CM3.

$$R = 1 - 0.5\times erfc\left(\frac{7.6}{2\times 5.7}\right)$$
$$= 1 - 0.5\times .346$$
$$= 0.827$$

5.9.2 Monte Carlo Simulation Method

Treating the remaining sites as potential interferers, run Monte Carlo simulations for points uniformly distributed over the proposed service area. For each point in the proposed service area, do the following in Sections 5.9.2.1 through 5.9.2.6.

5.9.2.1 Calculate Deterministic Signal Strengths

Calculate the (deterministic) signal strengths from the desired station and for all potential interferers at the location currently of interest using the methods of §§ 5.0 - 5.8. The results' should be expressed in dB values (e.g., dBm).

5.9.2.2 Draw from a Pseudorandom Number File

For the proposed station and for all potential interferers, draw a small number of times (e.g., 500) from a pseudorandom number file which has the following distribution: Type = Normal, standard deviation = 1, mean = 0. [For a proposed station and three potential interferers, this will result in 2000 draws, 500 corresponding to each station.]

5.9.2.3 Multiply by Known Standard Deviation

Multiply the values thus found by the known standard deviation for the area under consideration. See Section 5.8.

5.9.2.4 Offset the Calculated Signal Strengths

Offset the calculated signal strengths by the values just calculated; i.e., add 500 of the values calculated in Section 5.9.2.3 to the proposed station value calculated in Section 5.9.2.1, add the next 500 to the first interferer's value, etc. Note that, since the values calculated in Section 5.9.2.3 will have both positive and negative values, the results of Section 5.9.2.3 will sometimes be larger and sometimes smaller than Section 5.9.2.1.

5.9.2.5 Calculations for Each of the Samples

For each of the (500) samples, convert the values for the potential interferers to absolute (not dB) values, sum them, and convert the sum back to dB. Subtract this value from the value for the corresponding draw for the desired signal. If this number equals or exceeds the C/(I+N) goal, it is a "pass". Otherwise, it is a "fail".

5.9.2.6 Determine the Probability of a "Pass"

To determine the probability of a "pass" at a given location, divide the number of "passes" by the total number of samples (in the example, 500).

6.0 Performance Confirmation

This section addresses the issues associated with the empirical validation and quantification of wireless communications system performance. This process may be integral to a proof-of-performance or acceptance test or to quantify the actual interference environment versus simulated predictions in interference limited systems.

Conformance testing will validate that the user can expect to obtain the design reliability over their service area by measuring at a statistically significant number of random test locations, uniformly distributed throughout the service area. The entire concept of conformance testing rests on statistics.

The semantics of some of the terms used is critical to properly understanding this methodology. The service area is divided by a grid pattern to produce a large number of uniformly sized tiles, or test tiles. In one method, within each test tile a test location is randomly selected. At each of these test locations, a series of sequential measurements (**subsamples**) is made. This test location measurement, containing a number of subsamples, constitutes the **test sample** for this location.

Alternatively, the grid pattern is used to develop a test route that is uniformly distributed throughout the service area with an approximately equal distance traveled in each grid. This test route shall pass once through each test tile while collecting data. Thus, a large number of test samples is collected and evenly distributed throughout the service area.

6.1 Service Area Reliability

The service area reliability shall be determined by the requisite percentage of the test locations that meet or exceed the CPC.

$$\text{Service Area Reliability (\%)} = \frac{T_p}{T_t} \bullet 100\% \qquad \text{[Eq. 50]}$$

where:

T_p = Total of tests passed

T_t = Total number of tests

6.2 Determination of Number of Test Tiles

The "estimate of proportions" shall be used to determine with a high degree of confidence that sufficient test grids have been developed to accurately determine the Area Reliability.

6.2.1 Estimate of Proportions

$$T_l = \frac{Z^2 \bullet p \bullet q}{e^2}. \qquad \text{[Eq. 51]}$$

where:

T_l = Number of Test Locations

Z = Standard Deviate Unit (Corresponding to the confidence level)

p = True Service Area Reliability (decimal)

q = 1 - p

r = Service Area Reliability Criterion (decimal)

e = Sampling error allowance (decimal)

This is subject to a limit such that :

$T_l \geq 100$ [Eq. 52]

The requirement is that T_l be the larger of the two values..

Values for the standard deviate are available in most statistics books. Some standard values for one sided (tail) tests [Z_α] and two sided (tails) tests [$Z_{\alpha/2}$] are shown in Table 13 of Appendix-A.

6.3 Pass/Fail Test Criteria

The following pass/fail criteria are possible:

- The “Greater Than” Test
- The “Acceptance Window” Test.

6.3.1 The “Greater Than” Test

The “Greater Than” Test requires that the percentage of test locations which meet the CPC must equal or exceed the service area reliability requirement. This necessitates a slight “overdesign” of the system by *e*% to provide the statistical margins for passing the conformance test as defined. For this test configuration, Z has one-tail [Z_α] and *e* is the decimal percentage of overdesign.

6.3.2 The “Acceptance Window” Test

The “Acceptance Window” test allows the percentage of test locations which meet the CPC to fall within an error window, ±*e*, which is centered on the service area reliability requirement to consider the acceptance test a pass. This eliminates the requirement for “overdesign”, but necessitates a two tail Z [$Z_{\alpha/2}$] which increases the number of test samples to be evaluated.

6.4 Confidence

6.4.1 Confidence Level

The greater the number of test locations, the higher the confidence level. The confidence level should reflect a high confidence that the measured values will indicate what the true value is. A confidence level of 99% should be used unless this choice forces the size of the test grids for the desired service area to become too small; i.e., < 100λ.

6.4.2 Confidence Interval

This defines the limits within which the true value should fall. Using the preceding example of an acceptance window test with a 99% confidence level and 2% error and a service area reliability requirement of 95%, the statement would be, “I am 99% confident that the true value lies between 93 and 97%”.

6.4.3 Size Constraints

Test grid tiles (areas) should be $\geq 100\lambda$ by 100λ, but less than 2 km by 2 km. All test grids shall be of equivalent shape and area. A reasonable aspect ratio of 3:2 through 2:3 is considered to be square for the purpose of sizing test grid tiles of that shape. A tile created using other shapes, such as triangles and hexagons of equivalent areas is an acceptable alternative to a rectangularly shaped tile.

6.4.4 Accessibility

Locations with inaccessible test grids shall be specified, prior to testing, and treated per one of the following options:

- Eliminated from the calculation
- Considered a pass
- Estimated based on adjacent grids (single grids only)

6.5 Measurements

6.5.1 Carrier Power

The local mean power shall be measured with a receiver calibrated at its antenna port. See Section 6.8.3. The use of a mean power value requires a linear or logarithmic transfer function. Alternatively, if the transfer function of the detection system is known, but is non-linear, a suitable set of correction factors may be developed and applied to correct the non-linear ranges of the transfer function.

Other distributions may be captured and used for additional analysis of fading.

6.5.2 Distance

The distance (D) for measurements of the carrier local mean in a test grid shall be:
$29\lambda \leq D \leq 100\lambda$. The preferred distance is 40λ as it smoothes out Rayleigh fading. Shorter distances have a large impact from the Rayleigh fading. Larger distances tend to include changes in the local value due to the location variability starting to change. At lower frequencies, less than 40λ may be necessary.

Bit Error measurements may require longer distances and/or time intervals to capture the required number of test sub samples. It is recommended that separate local $40\ \lambda$ values be captured so that failures can be analyzed.

6.5.3 Bit Error Rate

BER shall be measured using a suitable pseudorandom test pattern, e.g., the ITU-T V.52 or O.153 patterns. See Section 6.6.1.3.

6.5.4 Number of Subsamples Per Test Sample

The number of subsamples taken for each test sample to measure the mean or median power at each location shall be greater than 50. This is to produce a 90% confidence interval that the measured value is within ±1 dB of the actual value. To calculate different confidence intervals, use the following formulas, where T_s is the number of sub sample data points taken:

$$\text{90\% Confidence Interval (dB)} = 20 Log\left(1+\left(\frac{1.65}{\sqrt{T_s}}\bullet\sqrt{\frac{4-\pi}{\pi}}\right)\right) \qquad \text{[Eq. 53]}$$

$$\text{95\% Confidence Interval (dB)} = 20 Log\left(1+\left(\frac{1.96}{\sqrt{T_s}}\bullet\sqrt{\frac{4-\pi}{\pi}}\right)\right) \qquad \text{[Eq. 54]}$$

$$\text{99\% Confidence Interval (dB)} = 20 Log\left(1+\frac{2.58}{\sqrt{T_s}}\bullet\sqrt{\frac{4-\pi}{\pi}}\right) \qquad \text{[Eq. 55]}$$

6.6 Adjacent Channel Transmitter Interference Assessment

A copy of the normalized power-density spectrum table is obtained for each adjacent channel transmitter within approximately 297 km (180 miles) of the station, and ± 25 kHz of the channel being coordinated. Each power-density value in the table is subsequently multiplied by the ERP of its respective transmitter, and each frequency value is shifted by an amount equal to the difference between the channel being coordinated and the assigned channel frequency of the transmitter being assessed. Further, the power density table is scaled by the amount of attenuation associated with the station separation in accordance with the standard propagation model described in Section 5. (Note: subsequent paragraphs 6.6.1.1 and 6.6.1.2 explain how a power-density spectrum table may be generated from transmitter measurements.)

In addition, a normalized receiver selectivity characteristic table is obtained for the receivers to be used on the channel being coordinated. The adjacent channel coupled power (ACCP) from each transmitter is then determined by multiplying the adjacent channel transmitter table described above with the receiver table over the bandwidth of the receiver. The ACCP from all adjacent channel transmitters is then summed to give the total receiver ACCP.

In Appendix C, Tables C-1 and C-2 contain a coarse table for 11K0F3 and 16K0F3 modulation. Tables C-3 through C-9 contain the calculated ACCP for some of the common IF configurations. The amount of power intercepted by the stated IF is shown for the VHF and UHF bands with the normal spacing and split channel spacings. Three analog configurations are shown that are varied for the transmitter low pass filter appropriate for the band. For VHF it is a 2 pole filter; for UHF is a 3 pole filter; and for narrowband FM it is a 5 pole filter. For analog FM (e.g., 20K0F3E) a standard TIA 603 receiver shall be assumed unless noted otherwise.

The total ACCP is added to the IF noise power as determined in Section 3.6.3 to result in the level of interference plus noise power to be overcome by the received power of the desired signal. The received power of the desired signal is determined by using the propagation model and the ERP of the desired transmitter.

The desired signal power is numerically divided by the interference power to determine the system signal to interference plus noise ratio, and converted to dB. If the resulting value is greater than the value necessary for the desired channel performance criterion (CPC) for the given technology according to Table 5, the result is said to be a "pass"; otherwise the result is said to be a failure.

6.6.1 Normalized Power-density Spectrum Table

A transmitter's emissions may be characterized by a measurement of its power-density spectrum over a specified frequency span using an adjacent channel power (ACP) analyzer or a spectrum analyzer. This type of analyzer typically presents the emission spectrum using an oscilloscopic display of a locus of 501 to 1001 discrete data points, each data point representing the amount of power measured in a "frequency bin". The frequency "sweep" is accomplished by incrementing the measurement frequency one step at a time at a value determined by the span setting.

The measured values are properly compensated by the analyzer for the characteristics of the filter used for the measurement. A table of the amplitude and frequency of each data point may then be obtained via the analyzer bus, or a floppy disk interface, and subsequently formatted into a computer file which may be used for assessment analysis. This file can be normalized by determining the power-density in each "frequency bin" of the measured span relative to the total power of the emission, and making the center frequency equal to zero.

To measure both on-channel and adjacent channel power it is necessary that the frequency span of the measurement be at least 3 times the channel spacing.

To facilitate assessment computations, it is desirable to have only one value of frequency step, and it must not exceed the resolution bandwidth. Since there is a 2:1 range in the frequency step size used between manufacturers and models of currently available analyzers, but most have an adjustable span, then a power-density spectrum table with a uniform step size may be obtained by setting the span to equal N-1 times the step size as shown in Table 8 of Appendix-A. Table 8 lists the frequency span and resolution bandwidth to use for the various channel spacings that will be encountered.

It is recognized that the trace data output sequence, data retrieval and analyzer bus control commands, and floppy disk formats (not universally available at this time) differ between the various spectrum analyzer vendors so the captured transmitter power-density spectrum data table may need to be converted into the table format needed for performing the interference analysis via a floppy disk or Internet data transfer means.

6.6.1.1 Power-density Spectrum Table for an Analog Modulated Transmitter

1. Connect the equipment as illustrated in the following diagram, with the transmitter set to produce rated RF at the assigned frequency, and the signal analyzer set to use average power detection and the span and resolution bandwidth given in Table 8. (Note that the audio mixer may be eliminated if the audio generators are series connected.)

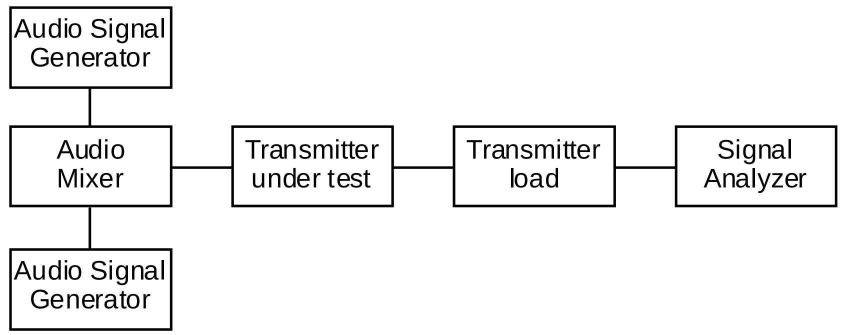


Figure 10. Two Tone Modulation Setup

2. Adjust the frequency of one audio generator to the lower frequency of the frequency pair given in Table 7 of Appendix-A for the modulation technology under test.

3. With the other audio generator off, modulate the transmitter with the low frequency audio tone only and adjust the generator output voltage to produce 50% of rated modulation. Record this level, then reduce the low frequency tone level by at least 40 dB.

4. Turn on the other audio signal generator and set its frequency to modulate the transmitter with the higher frequency tone of the frequency pair.. Adjust the generator output voltage to produce 50% of rated modulation and record this level.

5. Increase the output level of each signal generator respectively to a level 10 dB greater than the levels recorded in steps 3 and 4.

6. Capture the emission on the signal analyzer using a span no less than the appropriate span listed in Table 8. Generate a power-density spectrum table by recording the center frequency of, and the power in, each frequency bin of the spectrum produced by the emission.

7. Sum (linearly, not using logarithms) the power values in each bin of the spectrum produced by the signal analyzer, then record this total power value as the transmitter power.

8. Normalize the table by dividing the power value in each bin by the total power recorded in step 7, and setting the center frequency of the spectrum to 0 Hz. This is the normalized power-density spectrum table.

6.6.1.2 Power-density Spectrum Table for a Digitally Modulated Transmitter

1. Connect the equipment as illustrated below with the transmitter set to produce rated RF power at the assigned frequency, and the signal analyzer set to use average power detection with a span and resolution bandwidth per Table 8.

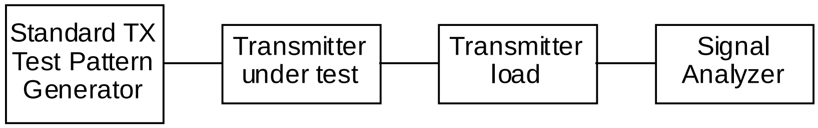


Figure 11. Digital Modulation Measurement Setup

2. Set the test pattern generator to produce the test pattern given in Table 7 of Appendix-A at the normal modulation level plus the maximum operating variance for the modulation technology under test.

3. Capture the emission on the signal analyzer using a display span no less than the appropriate value listed in Table 8. Generate a power-density spectrum table by recording the center frequency of, and the power in, each frequency bin of the spectrum produced by the emission.

4. Sum (linearly, not using logarithms) the power values in each bin of the spectrum produced by the signal analyzer, then record this total power value as the transmitter power.

5. Normalize the table by dividing the power value in each bin by the total power recorded in step 4), and setting the center frequency of the spectrum to 0 Hz. This is the normalized power-density spectrum table.

6.6.1.3 Digital Test Pattern Generation

The digital test patterns are based on the ITU-T V.52 pseudo-random sequence. The FORTRAN procedure given below generates this pattern for binary and four level signals.

```
         function v52()

C Function produces the V.52 bit pattern called for in the digital FM
C interference measurement methodology.  Each time this function is
C called, it produces one bit of the V.52 pattern.

         integer v52         ! The returned V.52 bit.
         integer register    ! The shift register that holds the current
                             ! state of the LSFR.

         data register/511/  ! The initial state of the shift register.
         save register       ! Saving the shift register between calls.

C Returning the value in the LSB of the shift register.
         v52=and(register,1)

C Performing the EXOR and feedback function.
         if(and(register,17) .eq. 1 .or. and(register,17) .eq. 16) then
           register=register+512
         end if

C Shifting the LSFR by one bit.
```

```
        register=rshft(register,1)
        end
```

The data from the procedure above is binary, and can be used to drive binary data systems directly. Since many modulations utilize four level symbols, the binary symbols from the V.52 sequence must be pared up into 4-level symbols. This can be done with this procedure:

```
        function v52_symbol()

C Function produces a di-bit symbol based on the V.52 sequence and
C the Layer 1 translation table.


        integer v52            ! External V.52 function.
        integer bit_1,bit_0    ! The two bits of the di-bit pair.
        integer v52_symbol     ! Four level V.52 symbol.
        integer table(0:1,0:1) ! Translation table to map bits into 4-
                               ! level symbols.

C Setting up the translation table.
        data table/+3,-3,+1,-1/

C Making the V.52 draws and translating them to a 4-level symbol level
C with the translation table.
        bit_1=v52()
        bit_0=v52()
        v52_symbol=table(bit_1,bit_0)
        end
```

6.7 Delay Spread Methodology and Susceptibility

A method of quantifying modulation performance in simulcast and multipath environments is desired. Hess describes such a technique [4], pp. 240-246. Hess calls the model the "multipath spread model." The model is based on the observation that for signal delays that are small with respect to the symbol time, the bit error rate (BER) observed is a function of RMS value of the time delays of the various signals weighted by their respective power levels. This reduces the entire range of multipath possibilities to a single number. The multipath spread for N signals is given by:

$$T_m = 2\sqrt{\frac{\sum_{i=1}^{N} P_i d_i^2}{\sum_{i=1}^{N} P_i} - \frac{\left[\sum_{i=1}^{N} P_i d_i\right]^2}{\left[\sum_{i=1}^{N} P_i\right]^2}}$$

[Eq. 56]

Since BER is proportional to T_m, any value of N can be represented as if it were due to two rays of equal signal strength, as shown here:

$$T_m\Big|_{i=2;\,P_1=P_2} = |d_2|$$

[Eq. 57]

Hess describes a method where multipath spread and the total signal power required for a given BER criteria are plotted and used in a computer program to determine coverage. Figure 12 in Section 6.7.1 shows this graph for QPSK-c class modulations at 5% BER given a 12 dB noise figure receiver. The points above and to the left of the line on the graph represent points that will have 5% BER or less, and thus meet the 5% BER criterion. The points below or to the right of the line have greater than 5% BER and thus do not meet the 5% BER criterion.

A figure of merit for delay spread is the asymptote on the multipath spread axis, which is the point at which it becomes impossible to meet the BER criterion at any signal strength. This is easily measured by using high signal strength and increasing the delay between two signals until the criterion BER is met. The two signal paths are independently Rayleigh faded. The other figure of merit for a modulation is the signal strength required for a given BER at T_m=0 µS. Given these attributes, the delay performance of the candidate modulation is bounded. It should be noted that these parameters are the figures of merit for the modulation itself; practical implementations, e.g., simulcast infrastructures, may change these curves. Figure 13 in Section 6.7.1 below shows the BER verses T_m at high signal strength for both QPSK-c class modulations.

6.7.1 QPSK-c Class Delay Spread Performance (12.5 and 6.25 kHz) Digital Voice

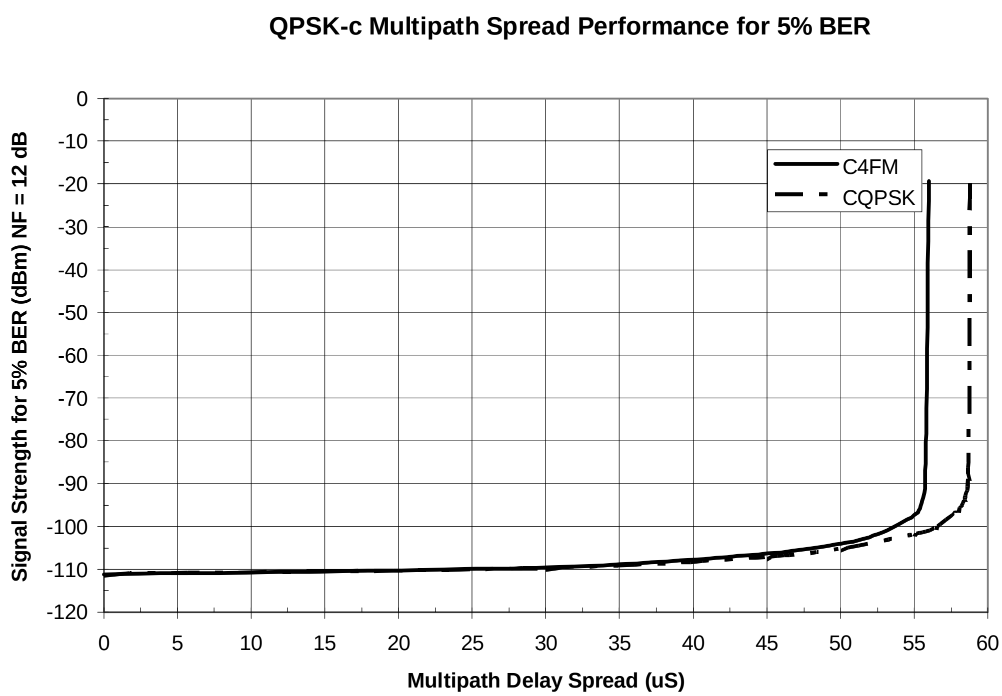


Figure 12. Multipath (Differential Phase) Spreads for APCO 25 Modulations

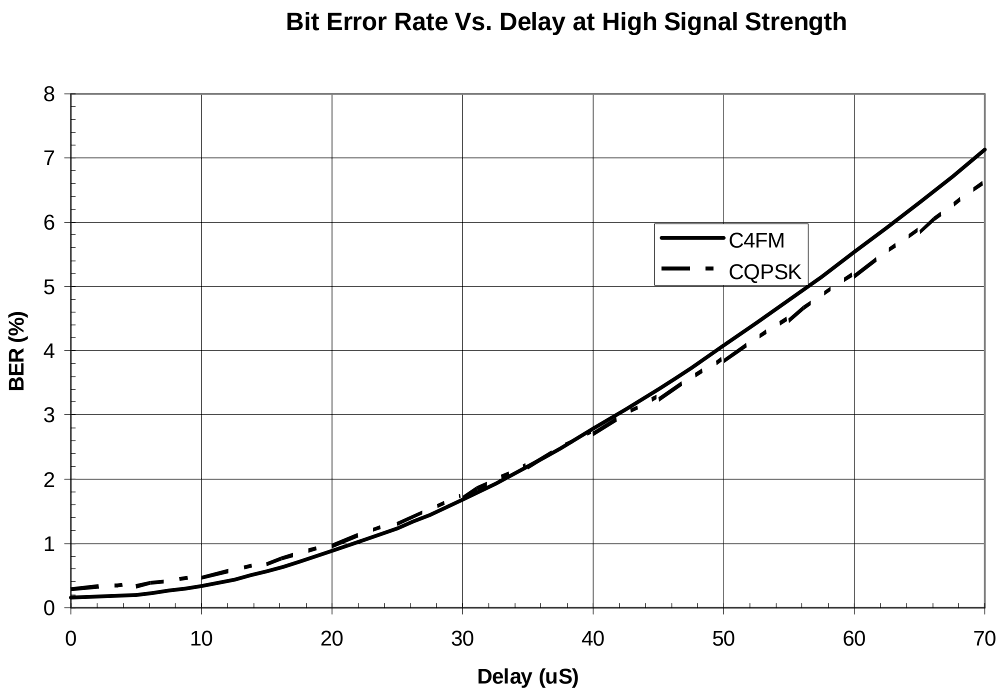


Figure 13. Simulcast Performance of APCO 25 Modulations

6.7.2 QPSK-c Type Delay Spread Performance (12.5 and 6.25 kHz) Digital Data

TBD

6.7.3 CVSD-XL Delay Spread Performance (25 kHz) Digital Voice

TBD

6.7.4 CVSD-XL NPSPAC Delay Spread Performance (12.5 kHz) Digital Voice

TBD

6.7.5 π/4 DQPSK (F-TDMA down link) Delay Spread Performance (12.5 kHz) Digital Voice

TBD

6.7.6 π/4 DQPSK (F-TDMA down link) Delay Spread Performance (12.5 kHz) Digital Data

TBD

6.7.7 EDACS® PRISM F-TDMA Delay Spread Performance (12.5 kHz) Digital Voice

TBD

6.7.8 EDACS® PRISM F-TDMA Delay Spread Performance (12.5 kHz) Digital Data

TBD

6.7.9 EDACS® Aegis Delay Spread Performance (25 kHz) Digital Voice

TBD

6.7.10 EDACS® Aegis Delay Spread Performance (25 kHz) Digital Data

TBD

6.7.11 EDACS® Aegis Delay Spread Performance (12.5 kHz) Digital Voice

TBD

6.7.12 EDACS® Aegis Delay Spread Performance (12.5 kHz) Digital Data

TBD

6.7.13 DIMRS Delay Spread Performance (25 kHz) Digital Voice

TBD

6.7.14 DIMRS Delay Spread Performance (25 kHz) Digital Data

TBD

6.7.15 TTIB/FFSR 16 QAM LM Delay Spread Performance (5 kHz) Digital Voice

TBD

6.7.16 TTIB/FFSR 16 QAM LM Delay Spread Performance (5 kHz) Digital Data

TBD

6.7.17 TTIB/FFSR 128 QAM LM Delay Spread Performance (5 kHz) Digital Data

TBD

6.7.18 RZ-SSB 16 QAM LM Delay Spread Performance (5 kHz) Digital Voice

TBD

6.7.19 RZ-SSB 16 QAM LM Delay Spread Performance (5 kHz) Digital Data

TBD

6.8 Conformance Measurements

6.8.1 Local Mean

In cases where dynamic range may be limited, the local mean should be captured and compared against the local median. The upper and lower deciles and standard deviation of the samples should also be measured. Characterization of these parameters is not required at each test sample location. Subsampling to perform these measurements should be done with a receiver calibrated at its antenna port. The use of a mean power value generally requires a detection system possessing either a linear or logarithmic transfer function. Alternatively, if the transfer function of the detection system is known, but is non-linear, a suitable set of correction factors can be developed and applied to correct the non-linear ranges of the transfer function. Local median values may be employed. If the difference between the local mean and local median exceeds approximately 2 dB, the distribution of the statistic shall be evaluated and an appropriate analysis performed.

6.8.2 Talk Out vs. Talk In Testing

Conformance testing need only be done in the Talk Out (outbound) direction. Reciprocity will apply and an offset correction value may be used to evaluate talk in (inbound) performance. If there is a large difference in height between the site transmit antenna and receive antenna, the assumption of reciprocity may not be valid. The additional expense and complexity of a talk in test may be justified in the following cases:

- Antenna distortions due to antenna support structure
- High ambient noise levels at site or in field
- Different Selectivity or Mode for Talk Out (down link) and Talk In (up link)
- Diversity
  - ♦ Macro (Voting)

- ♦ Micro (On Site Receiver Combiner)
- Different Horizontal Antenna Patterns

6.8.3 Calibration of a CPC Evaluation Receiver

A CPC evaluation receiver should be calibrated to its antenna input port using a signal source whose absolute level accuracy is specified within ± 1.0 dB. Coaxial cable losses shall be calibrated out. The calibration signal source shall have been calibrated within the time interval recommended by its manufacturer, but in no event more than one year prior to calibrating the test receiver. Prior to calibrating the CPC evaluation receiver, the calibration signal source shall have been warmed up according to its manufacturer's recommendation for guaranteed amplitude accuracy, but in no event for less than 30 minutes.

When BER is the criterion, the CPC evaluation receiver should have attenuators added so that its reference sensitivity is obtained at its specified power level. This is necessary to prevent a very sensitive receiver from biasing the test results. When received power is being measured, it is unnecessary to derate a receiver to its simulated test reference sensitivity.

6.8.4 RSSI Mobile

Using a substitution method, the loss of the calibration coaxial cable should be measured and the receiver calibration table adjusted to represent the median signal strength required to produce RSSI indications over the dynamic range of the RSSI circuit. The maximum step size should be 1 dB from the RSSI threshold for 20 dB, then 2 dB size steps for 20 dB, and 5 dB steps thereafter. Local Mean Power shall be measured with a receiver calibrated at its antenna port. The use of a mean power value generally requires a detection system possessing a linear or logarithmic transfer function. Alternately, if the transfer function of the detection system is known but is non-linear, a suitable set of correction factors can be developed and applied to correct the non-linear ranges of the transfer function.

6.8.5 RSSI Fixed End

Using a substitution method, the loss of the calibration coaxial cable should be measured and the receiver calibration table adjusted to represent the median signal strength required to produce RSSI indications over the dynamic range of the RSSI circuit. The maximum step size should be 1 dB from the RSSI threshold for 20 dB, then 2 dB size steps for 20 dB, and 5 dB steps thereafter. Local Mean Power shall be measured with a receiver calibrated at its antenna port. The use of a mean power value generally requires a detection system possessing a linear or logarithmic transfer function. Alternatively, if the transfer function of the detection system is known, but is non-linear, a suitable set of correction factors can be developed and applied to correct the non-linear ranges of the transfer function.

6.8.5.1 Multicoupler Correction

When a receiver is fed by a receiver multicoupler or has a tower mounted preamplifier installed, a calibration curve should be created to compensate for the additional gain and amplified noise that will exist. This is a practical measure as injecting signals at the amplifier input can interrupt service for other receivers. A Noise Gain offset to calibrate the RSSI will apply, but the weak signal region will required a separate calibration.

The RSSI Noise offset will consist of the Surplus Gain, the overall gain between the first amplifier input and the subsequent losses prior to the input of the test base receiver, less the Effective Multicoupler Gain (EMG), which is the effective improvement in reference sensitivity between the input of the first amplifier stage and the reference sensitivity of the base receiver alone.

$$\text{RSSI Noise Gain Offset} = \text{Surplus Gain} - \text{EMG} \qquad \text{[Eq. 39]}$$

EMG = Reference sensitivity at first amplifier input - base reference sensitivity w/o amplifiers, but with amplifiers providing their noise contribution. This requires a directional coupler methodology for measuring the effect of the base receiver.

Referring to Figure 14:

a) Measure and record the test receiver static reference sensitivity through a calibrated directional coupler, C1, with its input terminated in 50 Ω S-1 to A. Record the insertion loss of the calibrated directional coupler C1.

b) Repeat and record the measurement through directional coupler C1 with its input port connected to the amplifier chain, S-1 to B and S-2 to A, terminated in 50 Ω .

c) Measure and record the test receiver static reference sensitivity through the calibrated directional coupler C2 with its input terminated in 50 Ω, S-1 to B, S-2 to A. Record the insertion loss of the calibrated directional coupler C2.

d) Calculate the EMG, Step (a) power minus Step (c) power, both corrected for coupler insertion losses.

e) Calculate the Total Gain, Step (b) power minus Step (c) power, both corrected for coupler insertion losses.

f) Calculate the RSSI Noise Gain Offset. Step (b) power minus Step (a) power , both corrected for coupler insertion losses. This should also equal the difference calculated in steps (d) and (e).

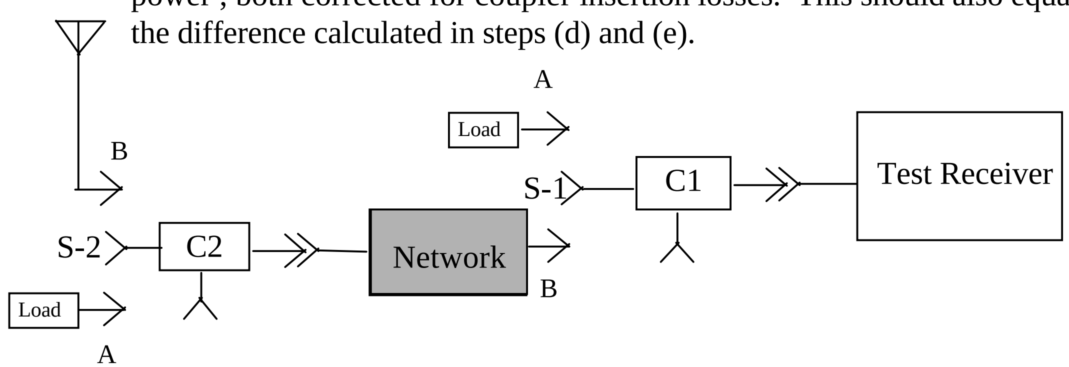


Figure 14. Multicoupler Calibration

g) Calibrate the RSSI by normalizing the input power level at C1 to that of a receiver that isn't connected to a multicoupler scheme. This would require that the "normalized" input power be Greater than the reference sensitivity by the RSSI Noise Gain Offset in dB.

h) For example, assume that the reference static sensitivity is -119 dBm, the Cs/N is 7 dB which infers that the noise floor of the receiver is -126 dBm. The corrected measurement a) would be -119 dBm. Corrected Measurement b) is -115.3 dBm and corrected measurement c) is -123.3 dBm. From this measurements, the EMG is (-119 -(-123.3)) = 4.3 dB. The Total Gain is (-115.3 - (-123.3)) = 8 dB. The RSSI Noise Gain Offset is (-115.3 -(-119)) = 3.7 dB. Thus the receiver requires a -115.3 dBm signal power to produce the same reference performance as a -123.3 dBm signal would at the input to the first amplifier. Thus by injecting the calibration signal at the input of the receiver at the RSSI Noise Gain Offset value, it is equivalent to injecting a signal at the input of the first amplifier which is EMG dB greater than the reference sensitivity of the receiver by itself, which isn't always practical when a system is in service.

6.9 Identifying Interference

Interfering carriers have the impact of affecting performance similar to an increase in noise. Since BER and RSSI can be measured, a reasonable calculation of interference can be made from evaluating these two related parameters.

BER can be mapped into $C_f/(I+N)$, e.g., a BER of 2% might, for example correspond to a ratio of 17 dB (50). RSSI is essentially C + I + N. When calibrated this might indicate that when a particular test yielded a measured 2% BER, the total power was for example -90 dBm (10^{-12} W). Thus the C and I + N components can be solved for. Measurements of idle channels can resolve the value of N to a reasonable value. Thus the C and I values can be solved for. High BER measurements at normal RSSI indications would represent increased I or N contributions. In the previous example, it would appear that if N is at -124 dBm, then there is an interferer at approximately -108 dBm.

7.0 Definitions and Abbreviations

There is a comprehensive Glossary of Terms, Acronyms, and Abbreviations listed in Appendix-A of TIA TSB102. In spite of its size, numerous unforeseen terms will have to be defined for the Compatibility aspects. Additional TIA/EIA references include; 603, Land Mobile FM or PM Communications Equipment Measurement and Performance Standards; TSB102.CAAA Digital C4FM/CQPSK Transceiver Measurement Methods; TSB102.CAAB, Digital C4FM/CQPSK Transceiver Performance Recommendations. ANSI/IEEE Std 100-1984. IEEE Standard Dictionary of Electrical and Electronic Terms will also be included as applicable. Items being specifically defined for the purpose of this document will be indicated as (New). All others will be referenced to their source as follows:

ANSI TIA/EIA-603	[T/E-603]
TIA TSB102, Appendix A	[102/A]
TIA TSB102.CAAA	[102.CAAA]
TIA TSB102.CAAB	[102.CAAB]
IEEE Standard Dictionary	[IEEE]
ITU-R [8A/XB]	[ITU-R]
New for this document	[New]

7.1 Definitions

ACCPR Adjacent Channel Coupled Power Ratio. The energy coupled into a victim receiver from an interfering carrier, relative to its average power on its assigned channel. The selectivity of the victims receiver and the Spectral Power Density of the interfering carrier interact to calculate this parameter.

ACIPR Adjacent Channel Interference Protection Ratio. See Adjacent Channel Rejection in TSB102.CAAA.

Adjacent Channel. The RF channel assigned adjacent to the licensed channel. The difference in frequency is determined by the channel bandwidth.

Adjacent Channel Coupled Power (ACCP). The energy from an adjacent channel transmitter that is intercepted by a victim receiver, relative to the power of the emitter.

Adjacent Channel Rejection [TSB102.CAAA]. The adjacent channel rejection is the ratio of the level of an unwanted input signal that causes the BER produced by a wanted signal 3 dB in excess of the reference sensitivity to be reduced to the standard BER, of the reference sensitivity. The analog adjacent channel rejection is a measure of the rejection of an unwanted signal that has an analog modulation. The digital adjacent channel rejection is a measure of rejection of an unwanted signal that has a digital modulation.

Cross analog to digital or digital to analog, require that the adjacent channel be modulated with its appropriate standard Interference Test Pattern modulation and that the test receiver use its reference sensitivity method.

Aegis [New]. Trademarked name for Ericsson trunked radio system.

"Area" Propagation Model. A model that does not predict power levels based upon the characteristics of path profiles.

Boltzmann's Constant (k). A value 1.3805x 10^{-23} J/K (Joules per Kelvin) At room temperature K = 290°.

C4FM [TSB102]. A 4-ary FM modulation technique that produces the same phase shift as a compatible CQPSK modulation technique. Consequently, either modulation may be received by the same receiver.

Co-Channel. Another licensee, potential interferer, on the same center frequency.

Confidence Interval. A statistical term where a confidence level is stated for the probability of the true value of something being within a given range which is the interval.

Confidence Level. The degree of confidence used in conjunction with a confidence interval to state the probability that the true value lies within that interval.

"Contour" Reliability. The probability of obtaining the CPC at the boundary of the Service Area. It is essentially the minimum allowable design probability for a specified performance.

CQPSK [TSB102]. The acronym for Quadrature Phase Shift Keyed (QPSK) AM transmitter which uses QPSK-c modulation to work with a compatible frequency discriminator detection receiver. See also C4FM.

Channel Performance Criterion [New]. The BER and vehicular Doppler fading rate required to deliver a specific MOS for the specific modulation. The CPC should be in the form of C_f/N, C_f/(I+N) @ X Hz Doppler.

DAQ [New]. The acronym for Delivered Audio Quality, a reference similar to Circuit Merit with additional definitions for digitized voice and a static SINAD equivalent intelligibility when subjected to multipath fading.

DIMRS [ITU-R]. The acronym for Digital Integrated Mobile Radio Service, representing a trunked digital radio system using multi-subcarrier digital QAM modulation.

Dipole. A half wave dipole is the standard reference for fixed station antennas. The gain is relative to a half wave dipole and is in dBd.

Effective Multicoupler Gain (EMG). The effective improvement in reference sensitivity between the input of the first amplifier stage and the reference sensitivity of the base receiver alone.

Equivalent Noise Bandwidth (ENBW). The noise bandwidth of a receiver. As it is very close to the $\pm$ 3 dB bandwidth, that value can be substituted if the exact effective bandwidth is unknown.

Faded Reference Sensitivity [TSB102.CAAA] The faded reference sensitivity is the level of receiver input signal at a specified frequency with specified modulation which, when applied through a faded channel simulator, will result in the standard BER at the receiver detector.

FFSR. Feed Forward Signal Regeneration: An adaptive equalization technique developed by McGeehan and Bateman to correct for amplitude and phase perturbations in a received signal by means of a reciprocal fading generator. FFSR is used in conjunction with TTIB.

FFSR/TTIB LM. A linear modulation implementation employing McGeehan and Bateman's channel linearization and equalization techniques.

Height Above Average Terrain (HAAT). The height of the radiating antenna above the average terrain which is determined by averaging equally spaced data points along radials from the site or the tile equivalents. Only that portion of the radial between 3 and 16 km should be averaged.

IMBE [TSB102]. The acronym for Improved Multi Band Excitation, the standard vocoder per TSB102.BAAA. "A voice coding technique based on Sinusoidal Transform Coding (analog to digital voice conversion)."

Inferred Noise Floor. The noise floor of a receiver calculated when the Reference Sensitivity is reduced by the static C_s/N required to obtain the Reference Sensitivity.

Interference Limited. The case where the CPC is dominated by the Interference component of C/(I+N).

Isotropic. An isotropic radiator is an idealized model where its energy is uniformly distributed over a sphere. Microwave point to point antennas are normally referenced to dBi.

Linear Modulation. Phase linear and amplitude linear frequency translation of baseband to passband and radio frequency

Lee's Method. The method of determining how many subsamples of signal power should be taken over a given number of wavelengths for a specified confidence that the overall sample is representative of the actual signal within a given number of decibels.

Local Mean. The mean power level measured when a specific number of samples are taken over a specified number of wavelengths. Except at frequencies less than 300 MHz, the recommended values are 50 samples and 40λ.

Local Median. The median value of measured values obtained while following Lee's method to measure the Local Mean.

Location Variability. The standard deviation of measured power levels that exist due to the variations in the local environment such as terrain and environmental clutter density variations.

Macro Diversity. Commonly used as "voting", where sites separated by large distances are compared and the best is "voted" to be the one selected for further use by the system.

Mean Opinion Score. The opinion of a grading body that has evaluated test scripts under varying channel conditions and given them a MOS.

Measurement Error. The variability of measurements due to the measuring equipment's accuracy and stability.

Micro Diversity. Receivers at the same site are selected among or combined to enhance the overall quality of signal used by the system after this process.

Noise Gain Offset (NGO). The difference between the overall gain preceding the base receiver (Surplus Gain) and the improvement in reference sensitivity (EMG).

Noise Limited. The case where the CPC is dominated by the Noise component of C/(I+N).

Normalized Power-Density Spectrum. The power-density in each frequency bin relative to the total power of the emission.

Number of Test Grids. The number of uniformly distributed but randomly selected test locations used to measure the CPC. It is calculated using the Estimate of Proportions formula and the specified Area Reliability, Confidence Interval and Sampling Error.

π/4 DQPSK [TSB102]. The acronym for "Differential Quadrature Phase Shift Keying", "quadrature" indicates that the phase shift of the modulation is a multiple of 90 degrees. Differential indicates that consecutive symbols are phase shifted 45 degrees (π/4) from each other.

Point to Point Model. A model that uses path profile data to predict path loss between points.

Power-Density Spectrum (PDS) [IEEE]. A plot of power density per unit frequency as function of frequency.

Power Spectral Density (PSD). The energy in dB relative to peak or rms power per Hertz.

Protected Service Area (PSA) [New]. That portion of a licensee's service area or zone that is to be afforded protection to a given reliability level from co-channel and adjacent channel interference and is based on predetermined service contours.

QPSK-c. The acronym for the Quadrature Phase Shift Keyed family of compatible modulations, which includes CQPSK and C4FM.

Reference Sensitivity [TSB102.CAAA]. The reference sensitivity is the level of receiver input signal at a specified frequency with specified modulation which will result in the standard BER at the receiver detector. [TIA/EIA-603] The reference sensitivity is the level of receiver input signal at a specified frequency with specified modulation which will result in the standard SINAD at the output of the receiver.

Sampling Error. The error from not being able to measure the true value by sampling the entire population.

Service Area. The boundary of the geographic area of concern for a user. Usually a political boundary such as a city limits, county limit or similar definition for the users business. Can be defined relative to site coordinates or an irregular polygon where points are defined by latitude and longitude.

Signal-to-Noise Ratio (SINAD). [E/T-603] The Signal-to-Noise Ratio (SINAD) is:

$$SINAD(dB) = 20\log_{10}[\frac{Signal + Noise + Distortion}{Noise + Distortion}]$$

where: Signal = Wanted audio frequency signal voltage due to standard test modulation. Noise = Noise voltage with standard test modulation. Distortion = Distortion voltage with standard test modulation.

Spectral Power Density (SPD) [IEEE]. The power density per unit bandwidth.

Standard BER [102.CAAA]. Bit Error Rate (BER) is the percentage of the received bit errors to the total number of bits transmitted. The value of the standard bit error rate (BER) is 5%.

Standard Deviate Unit (SDU). Also “Standard Normal Deviate.” That upper limit of a truncated normal (Gaussian) curve with zero mean and infinite lower limit which will produce a given area under the curve (e.g., Z = +1.645 for Area =0.95).

Standard Interference Test Pattern. [TSB102.CAAA] The standard digital transmitter test pattern is a continuously repeating 511 binary pseudo random noise sequence based on ITU-T V.52. Refer to ANSI TIA/EIA-603-196, 2.1.7 for the analog version. The standard analog digital transmitter test pattern is two tones, one at 650 Hz at a deviation of 50% of the maximum permissible frequency deviation, and another at 2200 Hz at a deviation of 50% of the maximum permissible frequency deviation.

Standard SINAD [ANSI EIA/TIA-603] The value of the standard signal-to-noise ratio is 12 dB. The standard signal-to-noise ratio (SINAD) allows comparison between different equipment when the standard test modulation is used.

Subsample. A single measured value. Part of a Test Sample.

Surplus Gain. The sum of all gains and losses from the input of the first amplified stage until the input to the base receiver.

Talk Out. From the fixed equipment outward to the "mobile" units. Also referred to as a forward link or down link.

Talk In. From the "mobile equipment" inbound to the fixed equipment. Also referred to as a reverse link or up link.

TTIB. Transparent Tone In Band: A technique developed by McGeehan and Bateman to provide an unambiguous frequency, phase and amplitude reference for use in linear modulation systems. TTIB is used in conjunction with FFSR.

Test Grid. The overall network of tiles where random samples of the CPC are taken.

Test Sample. A group of subsamples which are measured at a Test Tile.

Test Tile. The location where the random sample of CPC will be taken.

Tile Reliability [New]. The number of tiles which contain a margin equal to or greater than the Tile Reliability Margin, divided by the total number of tiles in the service area, expressed as a percentage. This is a direct way to calculate the CPC Area Reliability.

Tile Reliability Margin [New]. The margin, in dB, provided to create a minimum acceptable probability of achieving the required CPC in a tile. This is not to be confused with the CPC Area Reliability.

Voting. The process of comparing received signals and selecting the instantaneous best value and incorporating it into the system.

7.2 Abbreviations

4CPM	4-ary (Four Level) Continuous Phase Modulation
APCO	Association of Public Safety Communications Officials International, Inc.
ACCPR	Adjacent Channel Coupled Power Ratio
ACIPR	Adjacent Channel Interference Protection Ratio
ACPR	Adjacent Channel Protection Ratio
ANSI	American National Standards Institute
ATP	Acceptance Test Plan
BER	Bit Error Rate
C4FM	4-ary FM QPSK-C; Compatible Four Level Frequency Modulation
CAE	Counter Address Encoder
CCIPR	Co Channel Interference Protection Ratio (capture)

CCIR	International Radio Consultative Committee (Now ITU-R)
CFB	Cypher Feedback
CPC	Channel Performance Criterion
$C_f/(I+N)$	Faded Carrier to Interference plus Noise ratio
C_f/N	Faded Carrier to Noise ratio
C/I	Carrier to Interference signal ratio
CQPSK	AM QPSK-C; Compatible Quadrature Phase Shift Keying
C_s/N	Static Carrier to Noise ratio
CTG	Composite Theme Grids
CVSD	Continuously-Variable Slope Delta modulation
DAQ	Delivered Audio Quality
dBd	Decibels relative to a half wave dipole
dBqw	Decibels relative to a quarter wave antenna
dBi	Decibels relative to an isotropic radiator
dBμ	Decibels referenced to 1 microvolt per meter (1 μV/m)
DEM	Digital Elevation Model
DIMRS	Digital Integrated Mobile Radio System
DMA	Defense Mapping Agency
DQPSK	Differential Quadrature Phase-Shift Keying
DVP	Digital Voice Protection
$\frac{E_b}{N_0}$	Energy per bit per Hertz
EDACS®	Enhanced Digital Access Communication System
EMG	Effective Multicoupler Gain
ENBW	Equivalent Noise Bandwidth
erf	Error Function
erfc	Complementary Error Function
FDMA	Frequency Division Multiple Access
FFSR	Feed Forward Signal Regeneration
F-TDMA	Frequency, Time Division Multiple Access
HAAT	Height Above Average Terrain
iDEN™	Integrated Digital Enhanced Network
IMBE	Improved Multi Band Excitation
IMR	Intermodulation Rejection
ITU-R	International Telecommunication Union - Radiocommunication Sector
ITU-T	International Telecommunication Union - Telecommunication Sector
LM	Linear Modulation
LULC	Land Usage/Land Cover
MOS	Mean Opinion Score
NASTD	“National Association of State Telecommunications Directors”
NF	Noise Factor
NF_{db}	Noise Figure

NGDC	National Geophysical Data Center (under the Department of Commerce, located in Boulder, Colorado
NPSPAC	National Public Safety Planning Advisory Committee
OHD	Okumura/Hata/Davidson model
PEC	Perfect Electrical Conductor
PSA	Protected Service Area
QAM	Quadrature Amplitude Modulation
QPSK	Quadrature Phase-Shift Keying
QPSK-c	Quadrature Phase-Shift Keying - Compatible
QQAM	Quad Quadrature Amplitude Modulation (see TSB102)
RSSI	Receiver Signal Strength Indication
RZ SSB	Real Zero Single Sideband
SPD	Spectral Power Density
TBD	To Be Determined
TDMA	Time Division Multiple Access
TIREM	Terrain Integrated Rough Earth Model
TTIB	Transparent Tone In-Band
USGS	United States Department of the Interior, Geological Survey
Z	Standard Deviate Unit

8.0 References

[1] Reudink, Douglas O., "Chapter 2: Large-Scale Variations of the Average Signal", in: W. C. Jakes ed., *Microwave Mobile Communications*, New York: Wiley, 1974, Reprinted by IEEE Press, 1993, ISBN 0-471-43720-4

[2] Parsons, J. David, *The Mobile Radio Propagation Channel*, New York-Toronto, Halsted Press, 1992, ISBN 0-471-21824-X

[3] Lee, Wm. C.Y., *Mobile Communications Design Fundamentals*, New York, John Wiley & Sons, Inc., 1993, ISBN 0-471-57446-5.

[4] Hess, G., *Land Mobile Radio System Engineering*, Boston, Artech House, 1993, ISBN 0-89006-680-9.

[5] Freeman, Roger L., *Radio System Design for Telecommunications (1-100 GHz)*, New York: Wiley, 1987, ISBN 0-471-81236-6.

[6] Edwards, R. and J. Durkin, "Computer prediction of service areas for v.h.f. mobile radio networks", *Proc IEE*, Vol. 116, No. 9, September 1969, pp. 1496-97, §§3.2 - 3.2.4.

[7] *Digital Topographic Databases for Propagation Studies*, Recommendation RPN-1058, International Telecommunication Union, Geneva, 1994.

[8] *Draft Modification to Recommendation 1058*, Document [1994-98]: 3J/21 (United States of America), International Telecommunication Union, Geneva, 1995.

[9] Spaulding, A.D. & R.T. Disney (1974), *Man-Made Radio Noise*, OT Report 74-38, U.S. Department of Commerce, Office of Telecommunications, Boulder, 1974.

[10] J. Epstein and D.W. Peterson. “An experimental study of wave propagation at 850 Mc,” *Proc. IRE*, Vol. 41, No. 5, pp. 595-611, May 1953.

[11] P. F. Driessen. “Multipath delay characteristics in mountainous terrain at 900 MHz,” *42nd Vehicular Technology Society Conference Record*, Denver, CO, May, 1992, pp. 520-523, ISBN 0-7803-0673-2.

[12] M. Lebherz, W. Wiesbeck,. and W. Krank, A versatile wave propagation model for the VHF/UHF range considering three-dimensional terrain,” *IEEE Trans. on Antennas and Propagation*, Vol. 40, No. 10, October 1992, pp. 1121-1131.

[13] Jennings, Raymond D. and S. Jean Paulson. *Communication System Performance Model for VHF and Higher Frequencies.* OT Report 77-128, Office of Telecommunications, U.S. Department of Commerce, October 1977, NTIS Accession No. PB274458.

[14] Okumura, Y. *et al*, “Field Strength and Its Variability in VHF and UHF Land Mobile Radio Service”, *Rev Elect Comm Lab*, v16, 9-10, September-October 1968, pp. 825-873.

[15] *Transmission Loss Predictions for Tropospheric Communication Circuits*, Technical Note 101, NTIS access number AD 687-820, National Technical Information Service, US Department of Commerce, Springfield, Virginia.

[16] H.R. Anderson, “Site-Specific BER Analysis in Frequency-Selective Channels Using a Ray-Tracing Propagation Model,” *Proceedings of the Globecom ‘94 Conference*, San Francisco, CA, November, 1994, pp. 1441-1445.

[17] McQuate, P.L. et al., *Tabulations of Propagation Data over Irregular Terrain in the 230- to 9200 MHz Frequency Range, Parts I and II*, ESSA Technical Report ERL 65-ITS 58, Institute for Telecommunications Sciences, Boulder, CO, March 1968.

[18] Hata, Masaharu, “Empirical formula for propagation loss in Land Mobile radio services”, *IEEE Trans Veh Tech*, vol 29, 3, Aug 80, pp 317-325.

[19] Yao and Sheikh, “Investigation into Cochannel Interference in Microcellular Mobile Radio Systems”, *IEEE Trans. on Vehicular Tech.*, vol. 41, no. 2, May 1992, pp 114-123.

[20] Hufford, G., *Tabulations of Propagation Data over Irregular Terrain in the 75 to 8400 MHz Frequency Range, Part V: Virginia*, NTIA Report No. 91-282, December 1991.

Appendix-A. Tables

TABLE OF TABLES

Table 1. Delivered Audio Quality

DAQ Delivered Audio Quality	Subjective Performance Description	SINAD equivalent intelligibility[1,2]
1	Unusable, Speech present but unreadable	<8 dB
2	Understandable with considerable effort. Frequent repetition due to Noise/Distortion	12 ± 4 dB
3	Speech understandable with slight effort. Occasional repetition required due to Noise/Distortion	17 ± 5 dB
3.4	Speech understandable without repetition. Some Noise/Distortion	20 ± 5 dB
4	Speech easily understood. Occasional Noise/Distortion	25 ± 5 dB
4.5	Speech easily understood. Infrequent Noise/Distortion	30 ± 5 dB
5	Speech easily understood.	>33 dB

[1] CPC is set to the midpoint of the range.
[2] SINAD values are ***<u>NOT</u>*** to be used for system performance assessment.

Table 2. Antenna Reference Conversions

		Antenna Reference	
Specific Antenna	**Isotropic**	**λ/2 halfwave dipole**	**λ/4 quarterwave**
Isotropic	0 dBi	-2.15 dBd	-1.15 $dB_{l/4}$
λ/4 quarterwave	1.15 dBi	-1.0 dBd	0 $dB_{l/4}$
λ/2 halfwave dipole	2.15 dBi	0 dBd	1 $dB_{l/4}$

Table 3. IF Filter Specifications for Prototype Receivers

Modulation Type	ENBW (kHz)	IF Filter Simulation	Bandwidth (kHz)[1]
Analog FM (25 kHz) ±5 kHz	12.6	*	±7.5
Analog FM (25 kHz) ±4 kHz NPSPAC	10.1	*	±6.0
Analog FM (12.5 kHz) ±2.5 kHz	7.8	*	±4.6
C4FM	5.7	*	±3.4
CQPSK	5.7	12-pole Inverse Cheby†	±3.65
CVSD (25 kHz) ±4 kHz	12.6	*	±7.5
CVSD (25 kHz) ±3 kHz NPSPAC	10.1	*	±6.0
π/4 DQPSK (IMBE) TDMA (12.5 kHz)	9.5	*	±5.6
EDACS® (IMBE) (25 kHz)	9.8	*	±5.8
EDACS® (IMBE) (12.5 kHz)	9.1	*	±5.4
DIMRS	16.0	RRC, α=0.2	±8.75
TTIB FFSR	3.9	8-pole Inverse Cheby†	±2.5
RZ SSB	3.9	8-pole Inverse Cheby†	±2.5

[1] Filter Parameters for use with the formulas.

* Cascade of three 4-pole Butterworth stages, each 4-pole stage having a 3 dB bandwidth as given in the "Bandwidth" column.

† ε = 0.0003

RRC = Root Raised Cosine filter

Table 4. Filter Formulas

Butterworth Filter Equation
$$\text{Attenuation} = C \bullet 10 \bullet \log_{10}\left[1+\left(\frac{\Delta f}{\Delta f_0}\right)^{2n}\right]$$
C = The number of cascades Δf = The frequency offset from the IF center frequency Δf_0 = The frequency offset of the corner frequency n = Number of poles

Inverse-Chebyshev Equation
$$\text{Response in dB} = 10 \bullet \log_{10}\left[\frac{\varepsilon^2 \bullet \left[C_n\left(\frac{\Delta f_0}{\Delta f}\right)\right]^2}{1+\left\{\varepsilon^2 \bullet \left[C_n\left(\frac{\Delta f_0}{\Delta f}\right)\right]^2\right\}}\right]$$
$C_n(x) = \cosh[n * \cosh^{-1}(x)]$ for $x > 1$ $C_n(x) = \cos[n * \cos^{-1}(x)]$ for $x \le 1$ Δf = The frequency offset from the IF center frequency Δf_0 = The frequency offset of the corner frequency ε = Minimum stop-band attenuation n = Number of poles

Root Raised Cosine Equations
$$M(f) = 0dB\ ;\ \frac{f}{f_0} \le 1-\alpha$$ $$M(f) = 10\log_{10}\left\{\cos\left[\frac{\pi\left(\frac{f}{f_0}-1+\alpha\right)}{4\alpha}\right]^2\right\};\ 1-\alpha < \frac{f}{f_0} \le 1+\alpha$$ $$M(f) = -\infty\ ;\ 1+\alpha < \frac{f}{f_0}$$
f_0 = symbol frequency/2, i.e., for 9.6 kb/s with four levels the symbol frequency is 4.8 kS/s, therefore f_0 is 2,400.

Table 5. ***Projected*** **CPC Requirements for Different DAQs**

Modulation Type, (channel spacing)	**Static[1]** $ref / \frac{C_s}{N}$	**DAQ-3.0[2]** $BER\% / \frac{C_f}{(I+N)}$	**DAQ-3.4[3]** $BER\% / \frac{C_f}{(I+N)}$	**DAQ-4.0[4]** $BER\% / \frac{C_f}{(I+N)}$
Analog FM ± 5kHz (25 kHz)	12 dBS/4dB	na/17 dB	na/20 dB	na/27 dB
Analog FM ± 4kHz (25 kHz)[5]	12 dBS/5dB	na/19 dB	na/22 dB	na/29 dB
Analog FM ± 2.5kHz (12.5 kHz)	12 dBS/7dB	na/23 dB	na/26 dB	na/33 dB
C4FM (IMBE) (12.5 kHz)[6]	5%/5.4 dB	2.6%/15.2 dB	2.0%/16.2 dB	1.0%/20.0 dB
C4FM (IMBE) (12.5 kHz)[7]	5%/7.6 dB	2.6%/16.5 dB	2.0%/17.7 dB	1.0%/21.2 dB
CQPSK (IMBE) (12.5 kHz)[6]	5%/5.4 dB	2.6%/15.2 dB	2.0%/16.2 dB	1.0%/20.0 dB
CQPSK (IMBE) (12.5 kHz)[7]	5%/7.6 dB	2.6%/16.5 dB	2.0%/17.7 dB	1.0%/21.2 dB
CQPSK (IMBE) (6.25 kHz)	5%/7.6 dB	2.6%/16.5 dB	2.0%/17.7 dB	1.0%/21.2 dB
CVSD "XL" CAE (25 kHz)	8.5%/4 dB	5%/12.0 dB	3%/16.5 dB	1%/20.5 dB
CVSD "XL" CAE (NPSPAC)[8]	8.5%/4 dB	5%/14.0 dB	3%/18.5 dB	1%/22.5 dB
C4FM (VSELP)* (12.5 kHz)[6]	5%/5.4 dB	1.8%/17.4 dB	TBD	0.85%/21.6 dB
C4FM (VSELP)* (12.5 kHz)[7]	5%/7.6 dB	1.8%/17.4 dB	TBD	0.85%/21.6 dB
EDACS® Aegis (25 kHz)	TBD	TBD	TBD	TBD
EDACS® Aegis (12.5 kHz)	TBD	TBD	TBD	TBD
π/4 DQPSK (IMBE) TDMA (12.5 kHz)	5%/6.9 dB	2.6%/15.2 dB	2.0%/16.4 dB	1.0%/19.5 dB
EDACS® (IMBE) (25 kHz)	5%/5.3 dB	2.6%/14.7 dB	2.0%/15.7 dB	1.0%/19.2 dB
EDACS® (IMBE) (12.5 kHz)	5%/7.3 dB	2.6%/16.7 dB	2.0%/17.7 dB	1.0%/21.2 dB
DIMRS (25 kHz)	5%/12.5 dB	2.0 %/22 dB	TBD	1%/25 dB
TTIB/FFSR Analog LM (5 kHz)	TBD	TBD	TBD	TBD
TTIB/FFSR 16 QAM LM (5 kHz)	TBD	TBD	TBD	TBD
TTIB/FFSR 128 QAM LM (5 kHz)	TBD	TBD	TBD	TBD
RZ-SSB Analog LM (5 kHz)	TBD	TBD	TBD	TBD
RZ-SSB 16 QAM LM (5 kHz)	TBD	TBD	TBD	TBD

Note: These values were obtained from the manufacturers and should be verified with the manufacturer prior to usage.
* These VSELP values represent worst case, low speed.

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[1] Static is the reference sensitivity of a wireless detection sub-system (receiver) and is comparable to 12 dB SINAD in an analog system
[2] DAQ-2.0 (not shown) is comparable to 12 dB SINAD equivalent intelligibility,
DAQ-3.0 is comparable to 17 dB SINAD equivalent intelligibility
[3] DAQ-3.4 is comparable to 20 dB SINAD equivalent intelligibility, used for minimum CCP for some public safety entities.
[4] DAQ-4.0 is comparable to 25 dB SINAD equivalent intelligibility
[5] This is a NPSPAC configuration, 25 kHz channel bandwidths, but 12.5 kHz channel spacing. 20 dB ACIPR receiver assumed
[6] A wide IF bandwidth assumed as part of a migration process
[7] A narrow IF bandwidth is assumed after migration is completed.
[8] Reduced deviation for NPSPAC requirement.

Table 6. Protected Service Areas

Candidate Channel #1 Frequency XXX.YYYY		**PSAs Associated with Candidate Channel #1**						**Candidate Channel #1**	
Test AAR%	Order	Co-1 AAR%	Order	Co-2 AAR%	Order	Adj-1 AAR%	Order	Sum Orders	Rank
95	1	90	2	90	3	92	1	7	1
94	2	85	4	91	2	90	2	10	3
93	3	92	1	92	1	87	3	8	2
92	4	87	3	88	4	n/a		11	4

Candidate Channel #2	**PSAs Associated with Candidate Channel #2**	**Candidate Channel #2**
Channel #2 Results	Channel #2 Results	Channel #2 Results

Table 7. Test Signals

Modulation Type	**Modulation Test Signal**
Analog FM (± 5 kHz)	650 Hz tone & 2.2 kHz tone per Section 6.6.6.2
Analog FM, NPSPAC (± 4 kHz)	650 Hz tone & 2.2 kHz tone per Section 6.6.6.2
Analog FM (± 2.5 kHz)	650 Hz tone & 2.2 kHz tone per Section 6.6.6.2
C4FM (12.5 kHz)	ITU-T V.52 per TSB102.CAAA
QPSK-c (6.25 kHz)	ITU-T V.52 per TSB102.CAAA
CVSD - Normal (± 4 kHz)	12.0 kb/s binary ITU-T V.52 sequence
CVSD - NPSPAC (± 3 kHz)	12.0 kb/s binary ITU-T V.52 sequence
EDACS® Aegis (25 kHz)	9.6 kb/s binary ITU-T V.52 sequence
EDACS® Aegis (12.5 kHz)	9.6 kb/s binary ITU-T V.52 sequence
π/4 DQPSK (IMBE) TDMA (12.5 kHz)	18 kb/s 4-level ITU-T V.52 sequence
EDACS® (IMBE) (25 kHz)	9.6 kb/s binary ITU-T V.52 sequence
EDACS® (IMBE) (12.5 kHz)	9.6 kb/s binary ITU-T V.52 sequence
DIMRS (25 kHz)	TBD
TTIB/FFSR Analog LM (5 kHz)	TBD
TTIB/FFSR 16 QAM LM (5 kHz)	TBD
TTIB/FFSR 128 QAM LM (5 kHz)	TBD
RZ SSB Analog LM (5 kHz)	400 Hz tone & 1.0 kHz tone @ equal levels
RZ SSB 16 QAM LM (5 kHz)	TBD

Table 8. Power-Density Spectrum Measurement Bandwidths

Channel Separation (kHz)	Measurement Span (kHz) 501<N<1001	Resolution Bandwidth (Hz)
6.25	(N-1) x 50	100
7.50	(N-1) x 50	100
12.50	(N-1) x 100	100
15.00	(N-1) x 100	100
25.00	(N-1) x 150	300
30.00	(N-1) x 150	300

Note: N = number of points in analyzer span

Table 9. Noise Considerations

Frequency Range	Environment	Action
All	Fixed (site)	Consider Noise
< 400 MHz	Mobile	Consider Noise
near 821 MHz	Mobile	Consider Noise
≥ 400 MHz but not near 821	Mobile	Noise rarely an issue

Table 10. Difference between Powers (dB)

Difference	Add To		Difference	Add To		Difference	Add To
0.00	3.010		3.40	1.635		8.00	0.639
0.20	2.911		3.60	1.573		8.50	0.574
0.40	2.815		3.80	1.513		9.00	0.515
0.60	2.721		4.00	1.455		9.50	0.461
0.80	2.629		4.20	1.399		10.00	0.414
1.00	2.539		4.40	1.345		11.00	0.331
1.20	2.451		4.60	1.293		12.00	0.266
1.40	2.366		4.80	1.242		13.00	0.216
1.60	2.284		5.00	1.193		14.00	0.170
1.80	2.203		5.20	1.146		15.00	0.135
2.00	2.124		5.40	1.100		16.00	0.108
2.20	2.048		5.60	1.056		17.00	0.086
2.40	1.974		5.80	1.014		18.00	0.068
2.60	1.902		6.00	0.973		19.00	0.054
2.80	1.832		6.50	0.877		20.00	0.043
3.00	1.764		7.00	0.790		25.00	0.016
3.20	1.698		7.50	0.710		30.00	0.004

Table 11. Re-Classification of USGS Land Use/ Land Cover Codes

USGS Classification Number	USGS Classification Description	New Classification Number	New Classification Description
11	Residential	7	Residential
12	Commercial and services	9	Commercial/industrial
13	Industrial	9	Commercial/industrial
14	Transportation, communications, & utilities	1	Open land
15	Industrial and commercial complexes	9	Commercial/industrial
16	Mixed urban and built-up lands	8	Mixed urban/buildings
17	Other urban and built-up land	8	Mixed urban/buildings
21	Cropland and pasture	2	Agricultural
22	Orchards, groves, vineyards, nurseries, and horticultural	2	Agricultural
23	Confined feeding operations	2	Agricultural
24	Other agricultural land	2	Agricultural
31	Herbaceous rangeland	3	Rangeland
32	Shrub and brush rangeland	3	Rangeland
33	Mixed rangeland	3	Rangeland
41	Deciduous forest land	5	Forest land
42	Evergreen forest land	5	Forest land
43	Mixed forest land	5	Forest land
51	Streams and canals	4	Water
52	Lakes	4	Water
53	Reservoirs	4	Water
54	Bays and estuaries	4	Water
61	Forested wetland	5	Forest land
62	Non-forest wetland	6	Wetland
71	Dry salt flats	1	Open land
72	Beaches	1	Open land
73	Sandy areas other than beaches	1	Open land
74	Bare exposed rock	1	Open land
75	Strip mines, quarries, and gravel pits	1	Open land
76	Transitional areas	1	Open land
77	Mixed barren land	1	Open land
81	Shrub and brush tundra	1	Open land
82	Herbaceous tundra	1	Open land
83	Bare ground	1	Open land
84	Wet tundra	1	Open land
85	Mixed tundra	1	Open land
91	Perennial snowfields	10	Snow & ice
92	Glaciers	10	Snow & ice

Table 12. Local Clutter Attenuation in dB as a Function of Frequency and Land Use Classification

	Frequency	(MHz)					
Classification	**50.0**	**100.0**	**200.0**	**500.0**	**1000.0**	**2000.0**	**Reclassified Number**
open land	0.0	0.0	0.0	0.0	0.0	0.0	1
agricultural	0.0	0.0	0.0	0.0	0.0	0.0	2
rangeland	0.0	0.0	0.0	0.0	0.0	0.0	3
water	0.0	0.0	0.0	0.0	0.0	0.0	4
forest land	2.0	3.0	5.0	7.0	10.0	12.0	5
wetland	0.0	0.0	0.0	0.0	0.0	0.0	6
residential	3.0	5.0	7.0	10.0	12.0	15.0	7
mixed urban/ buildings	4.0	6.0	9.0	12.0	15.0	17.0	8
commercial/ industrial	4.0	6.0	9.0	12.0	15.0	17.0	9
Snow & ice	0.0	0.0	0.0	0.0	0.0	0.0	10

Warning: These attenuation values apply only when the Okumura/Hata/Davidson model is used in the 'Open' environment. Otherwise, attenuation may be included twice.

Table 13. Values for Standard Deviate Unit

Percentage (%)	$Z_{\propto}$	$Z_{\propto/2}$
50	*0*	*0*
70	*0.52*	*1.04*
80	*0.84*	*1.28*
85	*1.04*	*1.51*
90	*1.28*	*1.65*
95	*1.65*	*1.96*
97	*1.88*	*2.17*
99	**2.33**	**2.58**

Appendix-B. Recommended Data Elements for Automated Modeling, Simulation, and Spectrum Management of Wireless Communications Systems

Appendix-B. Recommended Data Elements for Automated Modeling, Simulation, and Spectrum Management of Wireless Communications Systems

The following information is required to facilitate Spectrum Management. Sufficient information is required to calculate the Effective Radiated Power (ERP_d) relative to a half wave dipole and the required signal levels for the minimum reliability for the Channel Performance Criterion (CPC) over the Protected Service Area. The existing systems must also be defined so that a bi-directional evaluation can be performed. The existing system(s) will be comprised of co-channel licensees, adjacent channel(s) and potentially alternate and second alternate channels for cases where a wide bandwidth channel is being utilized against narrow bandwidth channels.

Table B-1
Parameters of the Transmitter, [proposed]

1.1	**Site Latitude dd, mm, ss N/S**	
1.1.1	**Site Longitude ddd, mm, ss W/E**	
1.2	**Power supplied to the antenna**	**dBm**
1.3	**Antenna model and manufacturer**	
1.3.1	**Maximum Antenna Gain**	**dBd**
1.3.2	**Azimuth of directional gain if applicable**	**º from True North**
1.3.3	**Maximum Effective Radiated Power**	**dBm_d**
1.4	**Antenna Height Above Ground Level**	**(m) HAGL**
1.5	**Site Elevation, Height Above Mean Sea Level**	**(m) HAMSL**
1.6	**Tower Height**	**m**
1.7	**Modulation Type**	**Table 5**
1.7.1	**Vocoder type**	
1.7.2	**Adjacent Channel Power SPD histogram**	**dBc/bin**
1.8	**Bandwidth**	**kHz**
1.9	**Frequency**	**MHz**

Antenna Pattern **-** Provide manufacturer and model number so that an antenna pattern can be obtained. Leaving 1.3.2 blank implies omnidirectional and eliminates the requirement for an antenna pattern.

Table B-2
Parameters of the Receiver [proposed]

2.1	**Reference Static Sensitivity rel to 12 dBS or 5% BER**	**dBm**
2.2	**Receiver Equivalent Noise Bandwidth (ENBW)(Table 3)**	**kHz**
2.3	**Channel Performance Criterion, faded DAQ or % BER**	**Table 5**
2.3.1	**Usage Losses (in car or in building loss)**	**dB**
2.4	**Antenna Gain (include pattern and polarization losses)**	**dBd**
2.4.1	**Cable Loss**	**dB**
2.5	**Antenna Height Above Ground Level (HAGL)**	**m**
2.6	**Minimum Reliability for CPC at Service Area boundary**	**%**
2.7	**Frequency**	**MHz**
2.8	**Service Area definition**	
2.9	**Voting or Diversity? V(voting), DX (x branches)**	
2.10	**Simplex operation of mobile units? Y/N**	

Service area definition is required to determine where the mobile radios operate. It can be defined by:

- A radius around the site or a specific latitude/longitude.
- A rectangle with the opposite corners defined by latitude/longitude.
- Political boundary such as city, county, state.
- A political boundary plus an additional distance of "X" miles.
- A set of latitude/longitudes ordered in a counter clockwise direction so that when the points are connected, the resulting irregular polygon defines the required service area.

Simplex operation impacts adjacent channel reuse distance because of mobile to mobile potential interference.

Table B-3
Parameters for the Transmitter [existing]

3.1	**Site Latitude dd, mm, ss N/S**	
3.1.1	**Site Longitude ddd, mm, ss W/E**	
3.2	**Power supplied to the antenna**	**dBm**
3.3	**Antenna model and manufacturer**	
3.3.1	**Maximum Antenna Gain**	**dBd**
3.3.2	**Azimuth of directional gain if applicable**	**º from True North**
3.3.3	**Maximum Effective Radiated Power**	**dBm_d**
3.4	**Antenna Height Above Ground Level**	**(m) HAGL**
3.5	**Site Elevation, Height Above Mean Sea Level**	**(m) HAMSL**
3.6	**Tower Height**	**m**
3.7	**Modulation Type**	**Table 5**
3.7.1	**Vocoder type**	
3.7.2	**SPD histogram**	**dBc/bin**
3.8	**Bandwidth**	**kHz**
3.9	**Frequency**	**MHz**

Antenna Pattern - Provide manufacturer and model number so that an antenna pattern can be obtained.

Table B-4
Parameters of the Receiver [existing]

4.1	**Reference Static Sensitivity rel to 12 dBS or 5% BER**	**dBm**
4.2	**Receiver Equivalent Noise Bandwidth (ENBW)(Table 6)**	**kHz**
4.3	**Criterion Channel Performance, faded DAQ or % BER**	**Table 5**
4.3.1	**Usage Losses (in car or in building loss)**	**dB**
4.4	**Antenna Gain (include pattern and polarization losses)**	**dBd**
4.4.1	**Cable Loss**	**dB**
4.5	**Antenna Height Above Ground Level (HAGL)**	**m**

4.6	**Minimum Reliability for CPC at Service Area boundary**	**%**
4.7	**Frequency**	**MHz**
4.8	**Service Area Definition**	
4.9	**Voting or Diversity? V(voting), DX (x branches)**	
4.10	**Simplex operation of mobile units? Y/N**	

Service area definition is required to determine where the mobile radios operate. It can be defined by:

- A radius around the site or a specific latitude/longitude.
- A rectangle with the opposite corners defined by latitude/longitude.
- Political boundary such as city, county, state.
- A political boundary plus an additional distance of "X" miles.
- A set of latitude/longitudes ordered in a counter clockwise direction so that when the points are connected, the resulting irregular polygon defines the required service area.

If none of the above is available, use the method of Appendix-D. This applies to existing stations only.

Simplex operation impacts adjacent channel reuse distance because of mobile to mobile potential interference.

The evaluation will be made bi-directional, proposed to existing and existing to proposed, in the talk-out direction only, utilizing the worst case based on service area definitions.

Table B-5
Protected Service Area (PSA)

5.1	**Existing station protected availability (0 for unprotected)**	**%**
5.2	**Proposed station protected availability (0 for unprotected)**	**%**

The following field widths are recommended:

Table B-6
Field Widths

Sections		Input Data	Output Data
1.1	3.1	nn◊nn◊nn◊h (DMS)	±nn.nnnn (decimal degrees, not DMS)
1.1.1	3.1.1	nnn◊nn◊nn◊h (DMS)	±nnn.nnnn (decimal degrees, not DMS)
1.2	3.2	nn.n	nn.n
1.3	3.3	Mfr: 8 alpha char Model: 25 alpha char	Mfr: 8 alpha char Model: 25 alpha char
1.3.1	3.3.1	±nn.n	±nn.n
1.3.2	3.3.2	nnn	nnn
1.3.3	3.3.3	nn.n	nn.n
1.4	3.4	nnnn	nnnn
1.5	3.5	±nnnnn	±nnnnn
1.6	3.6	nnnn	nnnn
1.7	3.7	26 alpha char	26 alpha char
1.7.1	3.7.1	15 alpha char	15 alpha char
1.7.2	3.7.2	Up to 500 @ ±nn.n	Up to 500 @ ±nn.n
1.8	3.8	nn.nn	nn.nn
1.9	3.9	nnnn.nnnn	nnnn.nnnn
2.1	4.1	-nnn.n	-nnn.n
2.2	4.2	nn.nn	nn.nn
2.3	4.3	nn.n	nn.n
2.4	4.4	±nn.n	±nn.n
2.4.1	4.4.1	-nn.n	-nn.n
2.5	4.5	nnnn	nnnn
2.6	4.6	nn.n	nn.n
2.7	4.7	nnnn.nnnn	nnnn.nnnn
2.8	4.8	110 alpha characters	See Note 1
2.9	4.9	2 alpha characters	2 alpha characters
2.10	4.10	1 alpha character	1 alpha character
5.1	5.2	nn.n	nn.n

LEGEND:

h	=	hemisphere (N/S/E/W)
n	=	a numeric character
-	=	a minus sign (inserted for clarity)
±	=	a plus sign, a minus sign, or a blank (implying plus)
◊	=	a space (inserted for clarity)

. = a decimal point

Note 1: If the Service Area definition is in terms of a political boundary or a distance from a political boundary, the output data will consist of numerous pairs of latitude/longitude points. If the latitudes and longitudes are expressed in accordance with the RIGHT column for 1.1/3.1 and 1.2/3.2, each point will require 8 characters for each latitude and 9 for each longitude, excluding space characters between them. Political boundaries on coastlines or rivers will have numerous (possibly thousands of) points.

Note 2: For clarity, spaces must be included between fields (◊).

Note 3: Determine sign of output latitude/longitude from hemisphere. N & E are positive; S & W are negative. In the United States of America, latitudes are always positive and longitudes are generally negative. Some of the Aleutian Islands are in the Eastern Hemisphere.

Appendix-C. Simplified Explanation of Spectrum Management Process

Appendix-C. Simplified Explanation of Spectrum Management Process

C.1 Pull site elevation (AMSL) and antenna HAGL

C.2 Calculate ERP_d [Xmtr P_0 - cable losses - filtering losses + directional antenna gain (dB_d)]
e.g., 50 dBm - 2 dB - 4 dB + 8 dB = **52 dBm** (158.5 watts)

C.3 Use methods defined in this document to calculate the field strength at all points on the edge of the Service Area. If the field strength at any point on the edge of the Service Area exceeds 37 dBμ in the 150 MHz band or 39 dBμ in the 450 MHz band, the ERP must be reduced before proceeding.

C.4 Calculate Receiver requirements for CPC from reference sensitivity, in dBm or μV, Table 5.
C.4.1 Faded Performance Threshold = Ref Sensitivity - C_s/N + C_f/N (for CPC)
e.g., for C4FM (-116 dBm -7.6 dB + 16.5 dB (DAQ 3) = -107.1 dBm
C.4.2 Calculate ATP Target by adjusting for antenna gain, cable losses. building penetration margins, etc.
e.g., mobile with 2 dBd antenna and 3 dB cable loss, ATP Target = -106.1 dBm

C.5 Calculate coverage reliability for site independent of interference, noise only.
C.5.1 Pull Radial(s) from terrain data base. At each point, calculate propagation loss (L_1) for Open §5.1.2
C.5.2 Pull Environmental Loss from LULC cross reference (L_2) Table 12
C.5.3 Sum $L_l + L_2$ = Propagation Loss e.g., 136 + 10 (500 MHz residential) = 146 dB.
C.5.4 Calculate **Mean Signal Level** = ERPd - Propagation Loss e.g., 52 dBm -146 dBm = **-94 dBm.**
C.5.5 **Margin** = Mean Signal Level - ATP Target e.g., (-94 -(-106.1) =12.1 **dB**
C.5.6 **Z = Margin/σ** e.g., 12.1/5.6 = 2.1607
C.5.7 Calculate Noise-only Reliability. e.g., Z =2.16 ==> **98.46%.**
C.5.8 Store and continue iterating until PSA calculations are complete.

C.6 Calculate spectrally-equivalent ERP in adjacent channel
Using the power spectral density of the (proposed or existing) transmitter and the IF response of the (existing or proposed) receiver, calculate the amount of power intercepted by the receiver relative to one Watt for each instance.. Follow the procedure of Section 6.6.

C.6.1 For a proposed transmitter, collect the power spectral density on a bin by bin basis. If not in decibel units, put it into decibel units.
C.6.2 For an existing transmitter, if the **information** required in (C.6.1) is available use it, if not:
C.6.2.1 If the emission is appropriate for using Tables C-3 through C-9 with the appropriate IF responses, do so.
C.6.2.2 If the emission designator is 11KOF3E or 16KOF3E, and Tables C-3 - C-9 are not appropriate, use Table C-1 or C-2 and linear interpolation to form a bin-by-bin power spectrum.
C.6.2.3 Otherwise, form a bin-by-bin power spectrum based upon the relevant FCC Rule Section.

C.6.3 For applicants who have not yet selected specific equipment, or at the frequency **coordinator's discretion,** the method of C.6.2.3 may be used for any C.6.1 or C.6.2 situation.

e.g., (C.6.2.2) Create the SPD table in dBc/Hz for bins 2,500 Hz wide. For a lower RBW, linearly interpolate between data point when values are expresses as numeric, not in decibels.

e.g., (C.6.2.3) For a FCC mask which requires 60 dB adjacent channel rejection, the power in the adjacent channel is 60 dBc (10^{-6} watts) relative to the carrier. If it is to be divided into (for example) 100 bins, the power per bin would be -80 dBc (10^{-8} watts) per bin. The RBW is the frequency span divided by the number of bins. If this was 25 kHz, the RBW would be 250 Hz for the 100 bin example.

C.6.3 Determine the receiver IF response using the same RBW and bin centers as the transmitter SPD table.

C.6.3.1 If the receiver IF response is known, calculate a receiver response table using the formulas from Appendix-A, Table 4 and the parameters from Table 3.

C.6.3.2 If the receiver is existing, its IF response may be unknown. If so, use the values in TIA-603 § 3.1.6, 3.1.7, 4.1.6, 4.1.7, 5.1.6, or 5.1.7, as appropriate to determine adjacent channel rejection and apply this value uniformly across the entire adjacent channel. Assume zero on-channel rejection.

C.6.4 If it is desired to allow for frequency stability degradation, follow the rest of this section, otherwise assume no frequency error.

C.6.4.1 Use a standard deviation (σ) of 0. 4 times the sum of the FCC required stability (in Hz) for the combination of the fixed and mobile units At 450 MHz the 12.5 kHz channelization requires 4 ppm stability, thus σ is 0.4x1800 or 720 Hz.

C.6.4.2 Decide on a confidence factor (e.g., 95%) and find the corresponding Z value from Table 13. (e.g., Z=1.65)

C.6.4.3 Reduce the frequency separation between the adjacent channel separation by $\Delta f = Z\sigma\sqrt{2}$. At UHF this example would be (1.65)(720)(1.414) = 1680 Hz offset.

C.6.5 Sum the SPD file (dB) for the current bin with the corresponding receiver IF response (dB) bin and store in a file. These bins must have the same RBW and center frequencies. Convert the results of this addition to power. Sum the powers in all the bins and convert that sum back into a decibel value. Add 10 Log(RBW) (dB) to this sum. The result is the intercepted power of the victim receivers IF, relative to a one watt emitter which can be considered as a co-channel emitter. Reduce the ERPd of the interferer by this value for the simulation prediction.

C.7 Evaluate co- and adjacent-channel impact

C.7.1 Determine which sites to evaluate.

C.7.1.1 Find all existing sites on the frequency under consideration and both adjacent channels within 297 km.
[297 km is the sum of the 113 km protection distance plus line-of-site for k=1.33 for a 2,000 m HAAT mountain].

C.7.1.2 After the distance sorting process in Step C.7.1.1 above, the initial decision on whether to consider an interfering station further can be done using an analysis along the inter-station radial between the desired station and the interfering station. First, distance to the desired station coverage area boundary using the propagation method in Section 5. At the intersection of the inter-station radial and the designed station coverage area boundary, the magnitude of the interfering station signal is calculated, again using the Section 5 model. If the calculated interfering signal level at this intersection point is below the environmental noise level, this station need not be considered further as an interferer. For co-channel stations, if the desired median signal level at this point is 15 dB higher than the median interfering signal level plus the C/(I+N) allowance for CPC, then sufficient margin exists for adequate service and the interfering station need not be considered further as an interferer. For adjacent channel stations, if the desired median signal level at this point is 15 dB higher than the sum of the interfering median signal level plus the adjacent channel protection ratio minus the C/(I+N) allowance for CPC, then sufficient margin exists for adequate service and the interfering station need not be considered further as an interferer.

C.7.2 If the ratio of the desired station to interfering signal levels fall below the above criteria, or if the interferer is within the desired station coverage area, the interfering station will be subjected to further analysis. Voice systems may be subjected to either of the methods of Section 5.9.1 or of Section 5.9.2. If the results of the two methods conflict, the Monte Carlo Simulation is considered to be the more accurate, provided that the number of samples run is at least 5000. Because of “re-try” considerations, it is not practical to use the Simplified Estimate method for Data Systems. Thus, the Monte Carlo method must be used for data systems.

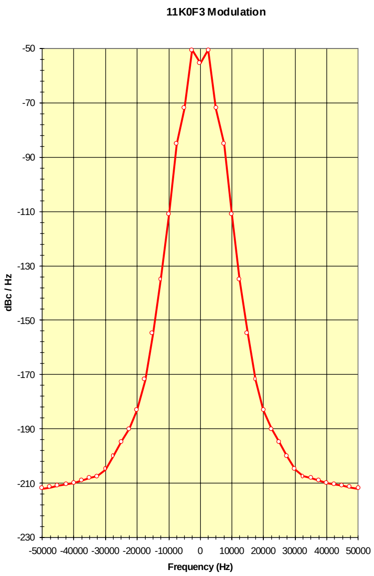
C.7.3 Interference Calculations
Calculate the interference potential using the methods of Sections 5.9.1 or 5.9.2.

C.8 If current evaluation was for proposed TX to existing RX and the existing TX to proposed RX evaluation hasn't yet been done, do that now by looping to C.2

C.9 Next configuration to evaluate. Loop to C.1

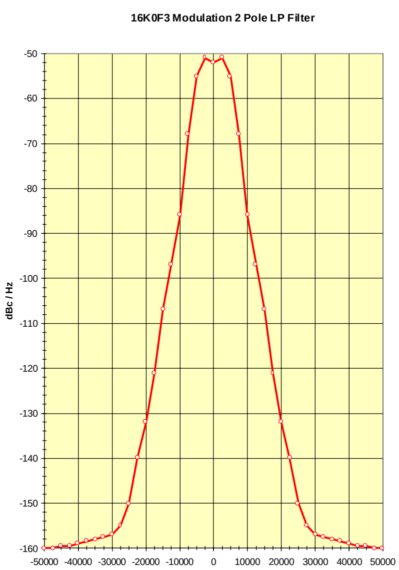
C.10 Continue to develop short list. Then evaluate short list in greater detail to determine the best recommendation.

Table C-1
Emission 11K0F3



Δf (kHz)	Response (dBc /Hz)
0	-55.5
2500	-50.5
5000	-72
7500	-85
10000	-111
12500	-135
15000	-155
17500	-172
20000	-183
30000	-205
50000	-212

Table C-2
Emission 16K0F3



Δf (kHz)	Response (dBc /Hz)
0	-52
2500	-51
5000	-55
7500	-68
10000	-86
12500	-97
15000	-107
17500	-121
20000	-132
30000	-157
50000	-160

The values indicated represent the intercepted power for a one Watt emitter for a 1 Hz bin at the frequency offset from the carrier as indicated. The associated chart which is part of the table is provided as an aid for interpolation. For Tables C-3 through C-5, the LP filter indicated is per the FCC rules are respectively for: VHF (2 pole filter, 5 kHz deviation; UHF/800 (3 pole filter, 5 kHz deviation; and narrowband FM (5 pole filter, 2.5 kHz deviation. Tables C-3 through C-9 were calculated for a RBW < 50 Hz, creating a better prediction than can be obtained from Tables C-1 and C-2.

Table C-3. Intercepted Power (ACCP) from FM 2 Pole Filter 5 kHz Deviation Emitter

	Intercepted Power from FM 2 Pole LP Filter 5 kHz Deviation			
	VHF		**UHF**	
IF Configuration	30 kHz	15 kHz	25 kHz	12.5 kHz
IF 12.6 kHz ENBW	-98.7	-29.8	na	na
IF 10.2 kHz ENBW	-100.0	-40.1	na	na
IF 7.8 kHz ENBW	-101.1	-47.5	na	na
IF 5.7 kHz ENBW	-102.4	-52.6	na	na

Table C-4. Intercepted Power (ACCP) from FM 3 Pole LP Filter 5 kHz Deviation Emitter

	Intercepted Power from FM 3 Pole LP Filter 5 kHz Deviation			
	VHF		**UHF**	
IF Configuration	30 kHz	15 kHz	25 kHz	12.5 kHz
IF 12.6 kHz ENBW	na	na	-92.5	-16.5
IF 10.2 kHz ENBW	na	na	-102.4	-24.5
IF 7.8 kHz ENBW	na	na	-110.7	-33.9
IF 5.7 kHz ENBW	na	na	-115.7	-42.0

Table C-5. Intercepted Power (ACCP) from FM 5 Pole 2.5 kHz Deviation Emitter

	Intercepted Power from FM 5 Pole LP Filter 2.5 kHz Deviation			
	VHF		**UHF**	
IF Configuration	30 kHz	15 kHz	25 kHz	12.5 kHz
IF 12.6 kHz ENBW	-138.3	-51.5	-119.5	-32.3
IF 10.2 kHz ENBW	-143.0	-66.6	-135.4	-45.5
IF 7.8 kHz ENBW	-144.2	-80.0	-139.1	-58.5
IF 5.7 kHz ENBW	-145.5	-91.0	-141.6	-68.9

Table C-6. Intercepted Power (ACCP) from C4FM Emitter

	Intercepted Power from C4FM			
	VHF		UHF	
IF Configuration	30 kHz	15 kHz	25 kHz	12.5 kHz
IF 12.6 kHz ENBW	-83.9	-50.4	-77.6	-32.1
IF 10.2 kHz ENBW	-85.2	-63.7	-79.3	-44.3
IF 7.8 kHz ENBW	-86.4	-70.2	-81.3	-56.4
IF 5.7 kHz ENBW	-87.7	-72.7	-83.3	-65.7

Table C-7. Intercepted Power from DVP Emitter

	Intercepted Power from DVP			
	VHF		UHF	
IF Configuration	30 kHz	15 kHz	25 kHz	12.5 kHz
IF 12.6 kHz ENBW	-83.0	-27.7	-61.4	-15.7
IF 10.2 kHz ENBW	-91.2	-30.7	-65.7	-24.3
IF 7.8 kHz ENBW	-93.2	-34.0	-71.0	-28.6
IF 5.7 kHz ENBW	-94.5	-37.8	-77.0	-30.1

Table C-8. Intercepted Power from EDACS® 12.5 kHz Emitter

	Intercepted Power from EDACS® 12.5 kHz			
	VHF		UHF	
IF Configuration	30 kHz	15 kHz	25 kHz	12.5 kHz
IF 12.6 kHz ENBW	-91.3	-38.4	-78.5	-26.8
IF 10.2 kHz ENBW	-94.6	-43.7	-81.4	-35.5
IF 7.8 kHz ENBW	-95.9	-51.9	-84.3	-38.7
IF 5.7 kHz ENBW	-97.2	-60.1	-87.6	-43.1

Table C-9. Intercepted Power from EDACS® 25 kHz Emitter

	Intercepted Power from EDACS® 25 kHz			
	VHF		UHF	
IF Configuration	30 kHz	15 kHz	25 kHz	12.5 kHz
IF 12.6 kHz ENBW	-80.3	-28.3	-53.1	-20.7
IF 10.2 kHz ENBW	-90.9	-32.3	-57.9	-26.1
IF 7.8 kHz ENBW	-93.3	-37.9	-65.5	-28.4
IF 5.7 kHz ENBW	-94.6	-42.7	-75.7	-31.7

Table C-10. Recommendations for ACCP (dB) from Adjacent Offset Channels

Emitter Modulation to Receiver Configuration (T to R)	7.5 kHz	6.25 kHz
25 kHz Wide band FM to 12.5 kHz Narrow IF Rcvr	10	na
25 kHz Wide band FM to 6.25 kHz Narrow IF Rcvr	TBD	? TBD
12.5 kHz Narrow FM to 25 kHz Wide FM IF Rcvr	6	na
12.5 kHz Narrow Digital to 25 kHz Wide FM IF Rcvr	6	na
12.5 kHz Narrow FM to 12.5 kHz Narrow IF Rcvr	27	17
12.5 kHz Narrow Digital to 12.5 kHz Narrow IF Rcvr	26.5	18
6.25 kHz Narrow Digital to 25 kHz Wide FM IF Rcvr	TBD	TBD
6.25 kHz Narrow Digital to 12.5 kHz Narrow IF Rcvr	TBD	TBD

Appendix-D. Methodology for Determining Service Area for Existing Land Mobile Licensees Between 30 and 512 MHz

Appendix-D. Methodology for Determining Service Area for Existing Land Mobile Licensees Between 30 and 512 MHz

We also have been asked by the Land Mobile Communications Council to address how the service area of an existing licensee should be determined. Such a task is not trivial as many of the data points present for new applications are not readily available for existing licensees.

The following contains an approach and methodology which, when used in conjunction with the overall WG 8.8 modeling and simulation methodology advanced in the body of this document, will permit the determination of a service area for most scenarios.

D.1 Information

It is possible to generalize a service area if certain basic elements are known or derived from the existing license which include:

- File or Reference Number
- Licensee Name
- Licensee Address (Mailing)
- Licensee Address (Physical)
- Latitude and Longitude Coordinates
- Ground Elevation AMSL
- Antenna Height AGL
- Fixed Station Class
- Mobile Station Class
- Fixed Station Transmitter Power Output
- Fixed Station Transmitter ERP (ref. half wave dipole). If ERP is not known, ERP will be inferred as follows:

Frequency Band	Inferred Fixed Station ERP
30-50 MHz	0.7x Transmitter Output Power
136-174 MHz	2.0x Transmitter Output Power
406-512 MHz	4.0x Transmitter Output Power

- Radio Service using current nomenclature, i.e., Police, Land Transportation, etc.

D.2 General Assumptions and Predicates

The WG 8.8 Modeling and Simulation Methodology will be employed as modified by the assumptions and predicates presented in this appendix.

D.2.1 Units and measures are consistently applied.

D.2.2 The modulation employed is analog frequency modulation with an emission designator of 16K0F3E .

D.2.3 The fixed station and mobile receiver performance meets TIA 603 concerning adjacent channel performance.

D.2.4 The fixed station and mobile transmitter sideband spectrum is represented by Tables C-1 and C-2.

D.2.5 Typical configurations of transmitter spectrum (ACCP) intercepted by various receiver configurations are tabulated in Tables C-3 through C-9.

D.2.6 Omni-directional fixed station antenna is used.

D.2.7 The mobile units operating with the associated base station/mobile relay operate within the coverage area of the base station/mobile relay.

D.2.8 Where handheld/portable units are licensed portable/handheld usage is assumed primary and the appropriate handheld/portable antenna correction factor shall be applied.

D.2.9 Handheld/portable antenna correction factor shall be applied as follows:

Frequency Band	Handheld/Portable Antenna Correction Factor
30-50 MHz	-15 dB
136-174 MHz	-10 dB
406-512 MHz	-6 dB

Note: Reference half wave dipole.

D.2.10 Coverage reliability is assumed as a function of radio service. The values are as follows:

Radio Service	Area Coverage Reliability
Public Safety	97%
LMR	90%

D.2.11 Average levels of ambient RF noise, referred to as kT_0b, are assumed. For 132-174 MHz this equates to a 6 dB derating value and a 3 dB derating value for 406-512 MHz. The RF noise level is defined in Section 4.2.

D.2.12 CPC is assumed as a function of radio service. The values are as follows:

Radio Service	CPC for Analog FM
Public Safety	DAQ-3.4 Equivalent
LMR	DAQ-2 Equivalent

Note: DAQ-3.4 is defined as 20 dB SINAD equivalent intelligibility.
DAQ-2 is defined as 12 dB SINAD equivalent intelligibility.

D.3 Discussion

This methodology assumes *a priori* that the information contained on the license is accurate and that the licensee is currently operating the station within the licensed parameters. However, when the parameters are evaluated in context of the overall modeling and simulation methodology proposed, a coverage area in the form of an irregular polygon may be determined for any existing licensed station.

In the event an existing licensee desired additional consideration above and beyond that provided by the above predicates, such a licensee could provide all of the information required of a new applicant. With more complete information a more finely tuned service area may be determined.

Appendix-E. User Choices

Appendix-E. User Choices

The main body of this document does not present a "hard and fast" methodology. It presents the user with a number of choices which must be made to perform the system design, spectrum management, and performance confirmation functions. The purpose of this Appendix is to present those choices in a simplified format so that users can clearly identify to others (e.g., prospective bidders) the specifics of the desired method.

Each choice is shown as a brief description along with a reference to the section of the main body where the choices are fully described. In those sections where optional choices can be made, no choice allows either selection to be used.

E.1 Identify Service Area - § 3.1 Use any of the methods of service area definition shown in Appendix B (between Tables B-2 and B-3).

E.2 Identify Channel Performance Criterion - §§ 3.2 & 3.5.1 For DAQ definitions, see Appendix-A, Table 1.

DAQ: ______

E.3 Identify Reliability Design Targets (both percentage and whether CPC contour or service area) - §§ 3.4 - 3.4.2 For advice, see **§** D.2.10.

______ %	☐ CPC Contour ☐ Service Area	(select one)

E.4 Identify the acceptable terrain profile extraction methods (check one or both) - §§ 5.3.1

- ☐ Bilinear Interpolation Method
- ☐ Snap to Grid Method

E.5 Identify acceptable interference calculation methods (check one or both) - §§ 5.9

- ☐ Equivalent Interferer Method
- ☐ Monte Carlo Simulation Method

E.6 Identify which metaphor(s) may be used to describe the plane of the service area - Select from those described in **§§ 5.7.1, 5.7.2, 5.7.3,** and **5.7.4.** See §§5.7 - 5.7.6 for discussion.

Select those that are acceptable (only the last two are acceptable for interference calculation or simulcast design):

- ☐ Radial Method
- ☐ Stepped Radial Method
- ☐ Grid Mapped from Radial Method
- ☐ Tiled Method

E.7 Determine required service area reliability to be predicted - §§ 3.6.2.2 and 5.8.

______ %

E.8 Determine Conformance Test confidence level - §§ 6.2.1, 6.4.1, and 6.5.4. This value is typically 99%.

______ %

E.9 Determine Sampling Error Allowance - §§ 6.2.1 & 6.4.2

± **______ %**

E.10 Determine which Pass/Fail Criterion to use - §§ 6.3 - 6.3.2

Select one:

- ☐ "Greater than" test
- ☐ Acceptance window test

E.11 Treatment of Inaccessible Grids - § 6.4.4

Select one:

- ☐ All are eliminated from the calculation
- ☐ All are considered a "pass"

- ☐ Single isolated inaccessible grids are estimated based upon “majority vote” of adjacent grids; multiple adjacent inaccessible grids are eliminated from the calculation
- ☐ Single isolated inaccessible grids are estimated based upon “majority vote” of adjacent grids; multiple adjacent inaccessible grids are considered a “pass”.

E.12 Adjacent channel drift confidence - § C.6.4.2

Confidence that combined drift due to desired and adjacent-channel stations will not cause degradation:

________ %