NTIA REPORT 86-198

BASEBAND PERFORMANCE DEGRADATION DUE TO INTERFERENCE IN THE FIXED-SATELLITE SERVICE

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AUGUST 1986

ABSTRACT

This report describes a set of receiver transfer characteristic (RTC) algorithms developed at NTIA for the analytical assessment of mutual interference effects in satellite communication services. The RTC algorithms convert the input carrier-to-interference power ratio (C/I) and modulation specifications of the desired (C) and interferer (I) signals into an output baseband performance degradation. The RTC algorithms also compute a C/I threshold margin for a given output baseband performance requirement. The modulation types and communication services include: (1) companded single-sideband, amplitude-modulated, multichannel telephony, (2) regular or companded, frequency-modulated, multichannel telephony, (3) frequency-modulated analog television, (4) digital coherent multiple-phase-shift keying, (5) single-channel-per-carrier with digital coherent multiple-phase-shift keying, and (6) single-channel-per-carrier with frequency-modulated analog voice.

KEY WORDS

Baseband Performance Degradation
Fixed-Satellite Service Communications
Interference Analysis Algorithms
Receiver Transfer Characteristics

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SECTION 1

INTRODUCTION

BACKGROUND

The National Telecommunications and Information Administration (NTIA) is responsible for managing the Federal Government's use of the radio frequency spectrum. Part of NTIA's responsibility is to "...establish policies concerning spectrum assignment, allocation and use, and provide the various departments and agencies with guidance to assure that their conduct of telecommunications activities is consistent with these policies" [NTIA, 1985]. In support of these requirements, NTIA performs spectrum resource assessments to identify existing or potential spectrum utilization and compatibility problems among the telecommunications systems of various departments and agencies. NTIA also provides recommendations to resolve any spectrum usage or allocation conflicts and to improve the spectrum management functions and procedures.

NTIA is engaged in the development and application of automated computer capabilities to be used by the Federal Government for the comprehensive assessment of national and international satellite communications systems. In particular, NTIA participated in the preparations for the 1985/88 Space World Administrative Radio Conference, and recently developed a Geostationary Satellite Orbit Analysis Program (GSOAP) dedicated to analyze the mutual interference effects between geostationary communications satellite networks [Hurt et al, 1986].

The coexistence of multiple satellite networks represents a critical concern from the orbital congestion, communications interference, and service impairment standpoints. The link transmissions and spectral characteristics produce unwanted interfering signals into the antenna patterns and receiver passbands of the satellite transponders and earth stations involved. This interference causes an output performance degradation on the communications services provided, which must be assessed to guide the orbit and spectrum resource management decisions.

The link geometries, transmission frequencies, antenna coverages, and power budgets all combine to specify the desired and interferer signal levels received. The receiver processing varies according to the desired modulation type and produces distinct output degradations according to the interferer modulation type. The modulation specifications thus govern the output performance degradation induced in the baseband information extracted (analog messages, digital symbols).

The development of receiver transfer characteristics (RTCs) is required to evaluate the output performance degradation for each possible combination of desired and interferer modulation types. The RTCs are dedicated algorithms that accept the desired and interferer modulation specifications and signal levels to compute an output performance degradation according to the baseband information content.

The RTC algorithms form an integral part of the GSOAP Program. The output performance degradations obtained are compared to preset threshold values to compute the available margins between the desired (C) and interferer (I) signal levels. These C/I threshold margins reflect the severity of the performance degradation effects and serve to assess the mutual interference between the communications satellite networks in question.

OBJECTIVE

The objective of this report is to present the RTC algorithms developed for the performance evaluation of mutual interference between geostationary communications satellite networks.

APPROACH

The existing and planned geostationary satellite networks were analyzed to identify the modulations associated with their communication services. The modulation types and parameter specifications encountered represent a wide variety of carrier frequencies, signal bandwidths, and spectral characteristics.

The RTC algorithms were developed by considering each and all pairs representing a distinct combination of desired and interferer modulation types. An output performance measure appropriate for each desired modulation type was used to formulate the output performance degradation caused by each interferer modulation type, as a function of the desired and interferer modulation parameters. The pertinent CCIR Recommendations and reports were employed, when applicable, and supplemented when they could not accommodate certain modulation types or parameter specifications of interest.

Each RTC algorithm representing a distinct pair of desired and interferer modulation types was designed in both a compact and a detailed version. The compact mode provides faster execution times, while the detailed mode can involve a more elaborate processing to improve accuracy at the expense of speed. The compact and detailed versions are identical in those cases where a simple formula or subroutine can be provided without accuracy compromises.

SECTION 2

SUMMARY

The RTC algorithms provide for the conversion of the input C/I levels and modulation specifications into an output baseband performance degradation and an input C/I threshold margin. This conversion varies with the modulation types and parameter values of the desired and interferer signals. A set of RTC algorithms has been developed for the following modulation types:

1. CSSB/AM: companded single-sideband, amplitude-modulated, multichannel telephony with suppressed carrier

2. FDM/FM: frequency-division multiplexed, frequency-modulated, multichannel telephony with or without companding

3. TV/FM: frequency-modulated television

4. DIG/PSK: digital coherent multiple-phase-shift keying

5. SCPC/PSK: single-channel-per-carrier, coherent multiple-phase-shift keying

6. SCPC/FM: single-channel-per-carrier, frequency-modulated voice

The inputs to the RTC algorithms consist of the following data for each pair of desired and interferer signals under consideration.

1. the two data sets of modulation parameters (MPARs) representing the desired and interferer modulation specifications

2. the frequency offset (F_O) between the desired and interferer signals

3. the ratio of the desired-to-interferer signal power (C/I), referred to the desired receiver input and excluding any interferer spectral truncation by the receiver passband

4. the ratio of the desired signal power to the thermal noise power density (C/N_0), referred to the desired receiver input

5. a binary flag designating the user selection of the compact or detailed computation mode.

The MPARs of each signal are structured into data sets fed to the RTC algorithms. One of the MPARs is a code word designating the modulation type from the options available. The RTC algorithms employ the desired and interferer designators to automatically select the one algorithm dedicated to a given modulation type pair. The RTC algorithm selected extracts automatically of those MPARs in the input data sets that are needed for the algorithm computations.

The RTC algorithms often consist of a main routine supported by dedicated subroutines that are automatically selected according to the input data values. This permits the RTC algorithms to self-adjust and accommodate many carrier offset, signal bandwidth and spectral shape variations in the desired and interferer signals within a given modulation type pair.

The RTC algorithms can automatically handle the interferer spectral truncation and power reduction as a function of the carrier offset and signal bandwidth values. All the spectral shaping and truncation occurs within the algorithm computations, since the spectral models employed are automatically triggered by the modulation specifications fed to the RTC algorithms.

The RTC algorithms automatically perform all the spectral generation and processing (e.g., sampling, integration, convolution) involved in the algorithm computations. A unified formulation is provided for each desired modulation type, insofar as possible, with the interferer type then inducing certain specific computations in the general formulation. This approach provides a unified foundation for interference analysis and permits modular processing and logical execution of the algorithm computations.

The RTC algorithms include automated simulation programs recently developed at NTIA. The FMSPC Program provides for the accurate generation of FDM/FM spectra, with automatic switching between a simple gaussian formula and an elaborate simulation process according to the modulation specifications. The SPCVL Program provides for the accurate execution of spectral convolution, which is required to generate the output interference spectra in FM demodulation cases (FDM/FM, TV/FM, SCPC/FM).

The RTC algorithms for PSK signals have distinct formulations automatically selected according to the interference spectral characteristics. The thermal noise effects are accounted for when computing the bit error rate performance and C/I threshold margin, with the latter automatically adapted according to the noise magnitude to reflect the decreased error margin available.

The following output data is provided for each RTC algorithm computation corresponding to a desired and interferer modulation pair:

1. the output performance measure representing the output baseband degradation

2. the C/I threshold margin referred to the desired receiver input.

The RTC algorithms represent a bridge between the input C/I level and the output baseband performance degradation. The fact that distinct modulation specifications produce a different output performance degradation with the same input C/I level emphasizes the need to use the RTC algorithms for the assessment of mutual interference between communications satellite networks.

The performance degradation dependence on the modulation parameter values also emphasizes the need to have complete and accurate modulation specifications. Moreover, the link power budgets that establish the input C/I levels also vary with the modulation specifications in practice, which must be accounted for in the input data preparation.

The GSOAP model is presented in Section 3 to illustrate the role of the RTC algorithms in the GSOAP computations. The modulation parameters corresponding to each modulation type are also identified in that section. Some application examples of the RTC algorithms are then presented in Section 4 to illustrate their results and potential.

The spectral models employed in the RTC algorithms are presented in Section 5 for each modulation type in question. The spectral characterization ranges from simple formula evaluations to elaborate process simulations, according to the modulation specifications and/or the compact/detailed computation mode.

The RTC formulations and algorithm computations are described in APPENDICES A through F. Each appendix considers a distinct desired modulation type and presents the different algorithms employed according to the interferer modulation type. The compact and detailed versions are clearly identified, along with the spectral models employed in each version.

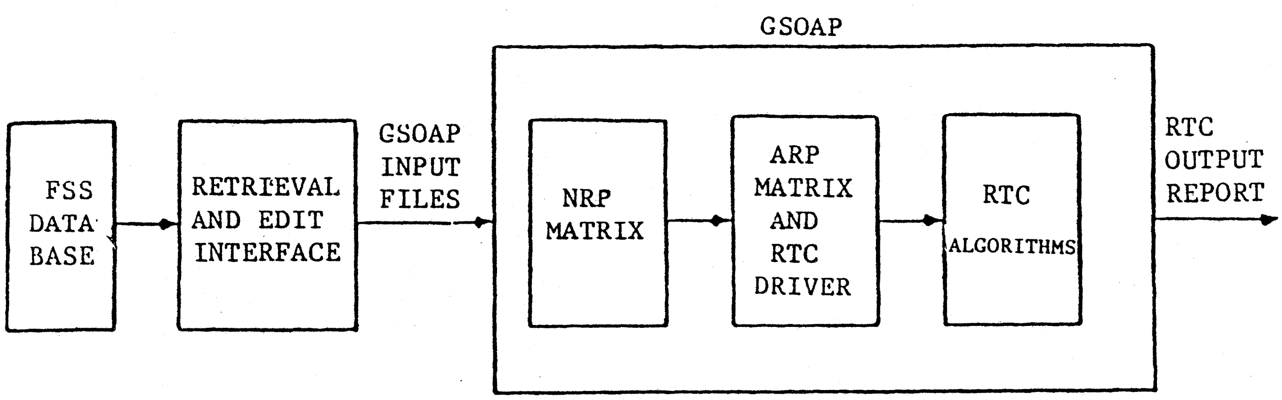
A set of additional formulas and measurements available for interference into TV/FM desired signals is presented in APPENDIX G. They include both input protection ratio and output signal-to-noise ratio as performance evaluation criteria.

SECTION 3

GSOAP MODEL

GSOAP MODEL DESCRIPTION

GSOAP is an analytical model developed by NTIA for the assessment of mutual interference between geostationary communications satellite networks [Hurt et al, 1986]. The GSOAP model is supported by a fixed-satellite service (FSS) data base dedicated to create the input scenarios via a user-interactive, menu-driven, retrieval and edit interface as shown in Figure 3-1.



GSOAP : Geostationary Satellite Orbit Analysis Program

FSS : Fixed Satellite Service

NRP : Normalized Received Power

ARP : Actual Received Power

RTC : Receiver Transfer Characteristic

Figure 3-1. Block diagram of GSOAP processing.

The scenario data is processed to obtain received signal levels for all uplink or downlink desired and interfering paths involved, as well as to select the worst C/I path conditions per desired and interfering network pair and modulation pair combination. The output performance degradation due to interference is then computed by the RTC algorithms, along with the C/I threshold margin available for each network-pair and modulation-pair combination in the scenario created.

FSS Data Base

The FSS data base consists of five files that contain the records of the geostationary communications satellite networks. A satellite network is defined as a space station (satellite) and its dedicated earth stations for uplink and/or downlink communications. A satellite network may exhibit various operational states based on the up/down link, transponder plan, satellite antenna, service area, carrier modulation, and link budget options available.

The five files in the FSS data base are: (1) Satellite Networks, (2) Service Areas, (3) Transponder Plans, (4) Satellite Antennas, and (5) Modulation and Link Budgets [Haines, 1985]. The file records consist of data fields whose entries specify the identification codes, parametric values, and generic characteristics of the satellite networks.

The first file specifies the space station and operational states (e.g., transponder plans, service areas) for each satellite network. The second file specifies each service area as a set of earth points with their geographical locations and topographical characteristics. The third file specifies each transponder plan as a set of uplink and downlink transmission frequencies. The fourth file specifies each satellite antenna pattern in the form of digitized shaped-beam contours and/or elliptical-beam parameters. The fifth file specifies the carrier modulation types and parameters, plus the link power budgets and earth station antennas associated with each modulation record.

An important feature of the GSOAP model is the provision of modulation-dependent link budgets. The equivalent isotropically radiated power (EIRP) and receive antenna gain specifications depend on the modulation type and on the modulation parameter values within a given type. This reflects the fact that a given input C/N (or C/I) amount will have a distinct effect on the output baseband performance depending on the modulation type and parameter values. Hence, the power budget is adapted, in practice, according to the modulation specifications.

Retrieval and Edit Interface

The retrieval and edit interface is used to interactively select the GSOAP input scenarios and automatically format the data for the GSOAP input files. These files support the automatic operation of all GSOAP program routines.

The retrieval and edit process is summarized in Figure 3-2, where the user-interactive stages are menu-driven with default options. The user can initially select among distinct versions of the data base (e.g., existing operational satellite networks versus proposed orbital spacing plans). The user has access to the first file to select specific satellite networks or all networks with a common characteristic such as orbital arc, frequency band, service region, or notifying administration.

The user can also select specific operational states from those available for each network, as well as antenna pattern and propagation models from the menu options. The first file contains codeword pointers that serve as automatic links into the subsequent files to provide for the automatic retrieval of linked record collections. The user can select specific modulation records from those retrieved for each network to control the scenario complexity (e.g., number of satellites, earth stations, modulations).

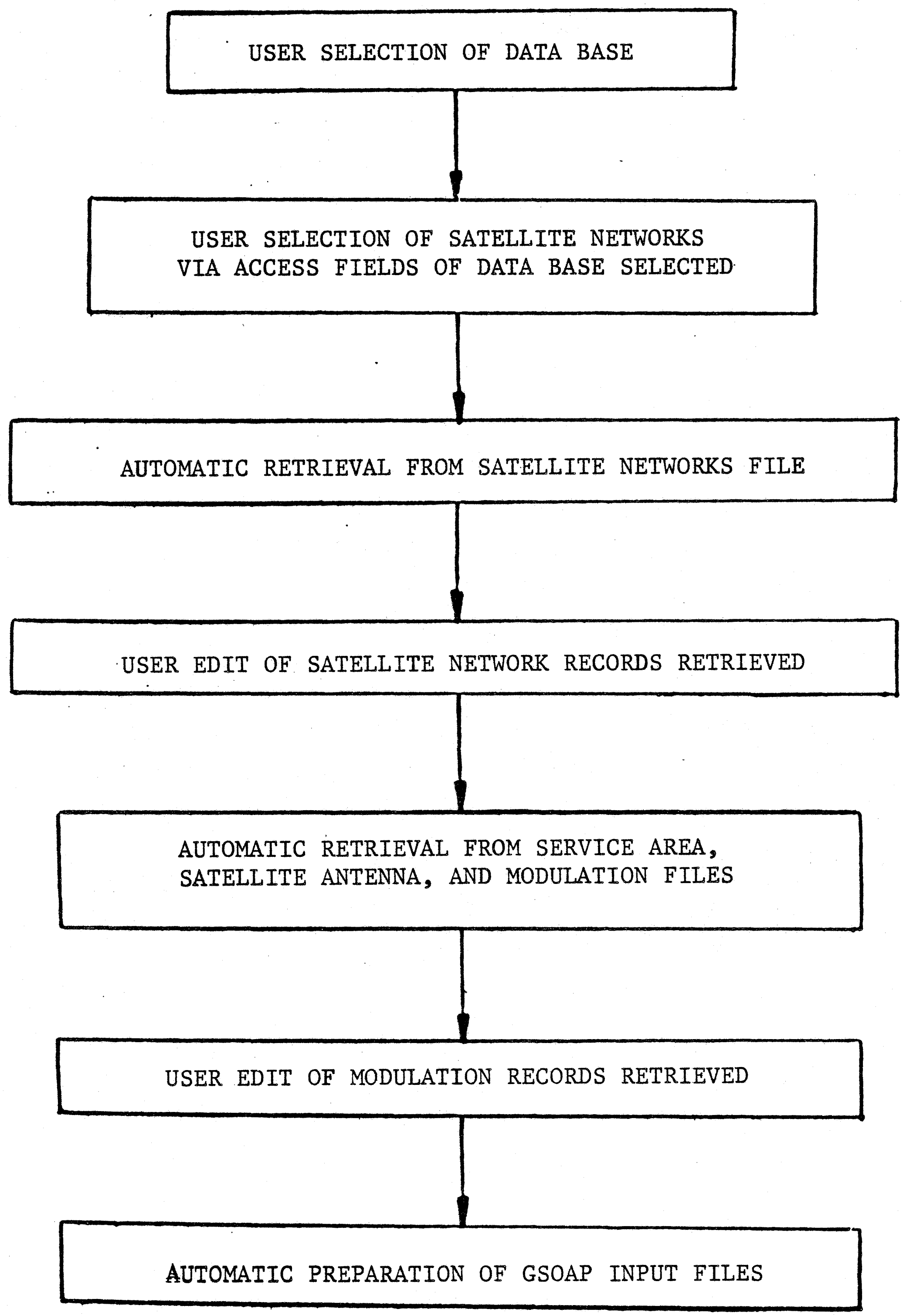


Figure 3-2. Block diagram of retrieval and edit process.

Normalized Received Power Matrix

GSOAP supports either uplink only, downlink only, or up-plus-down link scenarios. A normalized received power (NRP) matrix is computed for all desired and interfering signal paths in the uplink or downlink scenario, as illustrated in Figures 3-3 and 3-4. All the link EIRPs are set to 0 dB in the transmitting earth (uplink) or space (downlink) stations for the NRP computations.

The NRP computations are supported by a dedicated set of routines that account for the earth and satellite antenna patterns, propagation models, and polarization combining options selected by the user [Akima, 1985]. The NRP computations are fully automatic, with all necessary identification codes, parameter values, and option flags already provided in the GSOAP input files.

Actual Received Power Matrix

The EIRP specifications provided by the data base are used to compute the actual received power (ARP) and input power ratio (C/I) for all desired and interfering paths involved. The NRP to ARP conversion is performed by adding a modulation-dependent EIRP value per path, so that the ARP and C/I values vary with the modulation type and parameter values.

The worst C/I for a fixed desired and interfering network pair (i,j) and a fixed desired and interfering modulation pair (k,l) is determined by examining all possible desired and interfering paths for the two networks involved. The process is repeated by recycling over all possible network pair and modulation pair combinations, which results in a set of worst C/I values tagged by four index designators (i,j,k,l) that identify the desired network (i), interfering network (j), desired modulation (k), and interfering modulation (l) associated with each value.

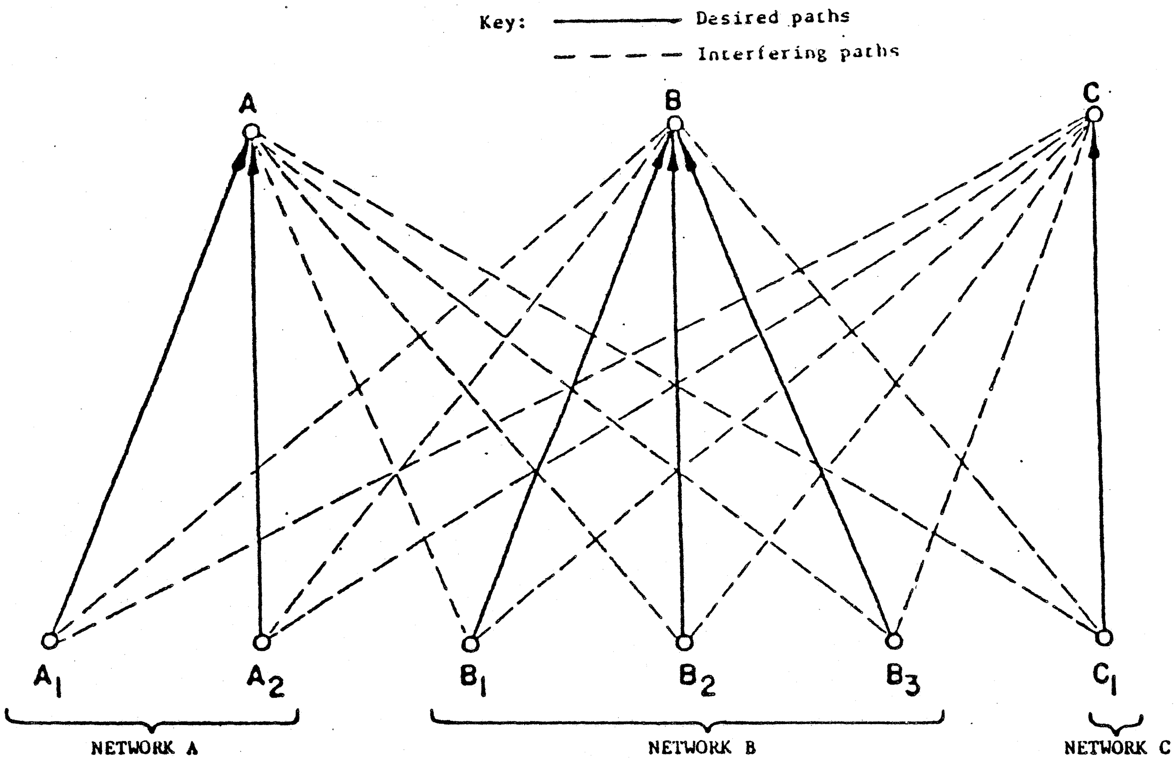


Figure 3-3. Illustration of uplink scenario paths.

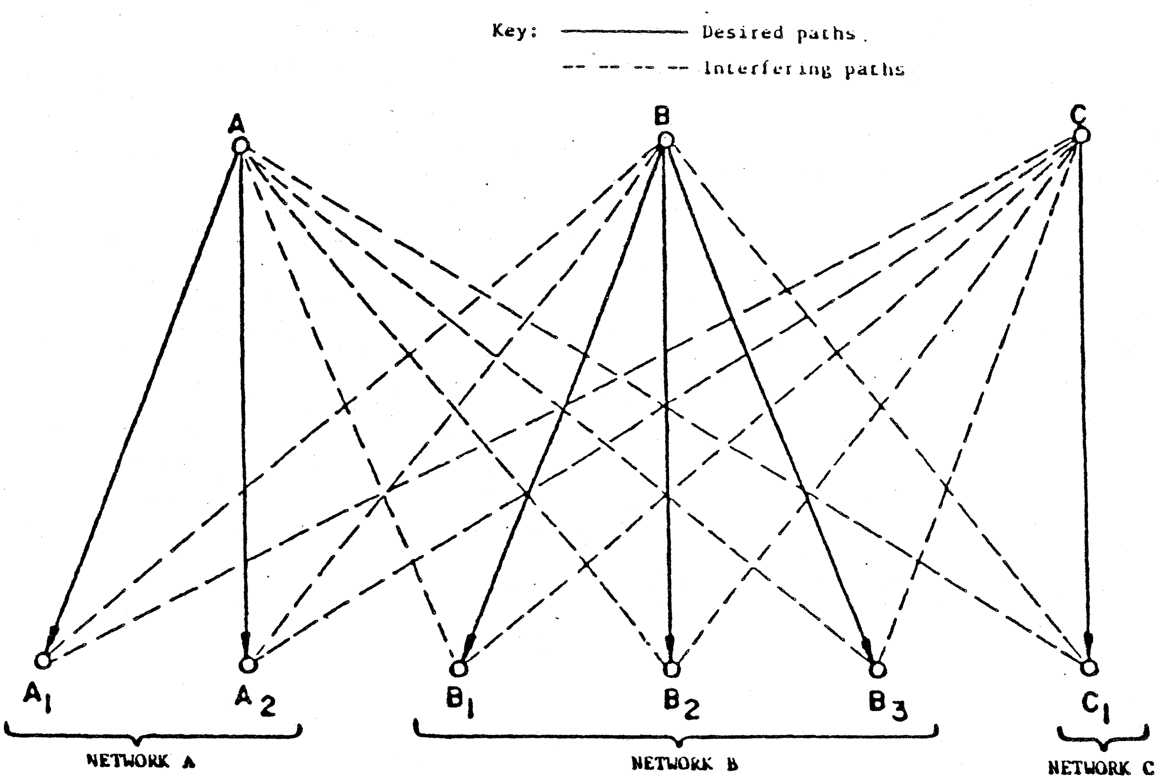


Figure 3-4. Illustration of downlink scenario paths.

Modulation Options

The modulation options supported by the RTC algorithms are now itemized, along with their data sets of modulation parameters used in the algorithm computations. A modulation specification consists of a set of modulation parameters pertinent to a given modulation type. The following six modulation types are presently supported by the RTC algorithms.

1. CSSB/AM: companded SSB/AM multichannel telephony
2. FDM/FM: FDM/FM multichannel telephony with or without companding
3. TV/FM: FM television
4. DIG/PSK: digital coherent multi-phase PSK
5. SCPC/PSK: single-channel-per-carrier with digital coherent multi-phase PSK
6. SCPC/FM: single-channel-per-carrier with analog FM voice

Modulation Parameters

The modulation parameters (MPARs) consist of data sets of generic or numerical parameters that provide the modulation specifications. The first and second MPAR are always a code designator and a text descriptor of the modulation in question. These two parameters serve to distinguish between different modulation types, as well as between different parameter values within the same type. In particular, the code designator provides for the automatic distinction of different MPAR data sets.

The other MPARs provide modulation data pertinent to the modulation type in question. The information content of each parameter varies with the modulation type as shown in TABLES 3-1 through 3-6. The data format (real, integer, alphanumeric) is maintained invariant for the same MPAR number over distinct modulation types to permit a simple processing of the MPAR data sets.

TABLE 3-1

COMPANDED SSB/AM MODULATION PARAMETERS

MPAR01	A#	= A serves to designate CSSB/AM, # serves to index CSSB/AM cases
MPAR02	CSSB/AM	= identifies what A above means
MPAR03	N_c	= number of voice channels
MPAR04	B_{all}	= allocated bandwidth
MPAR05	B_{occ}	= occupied bandwidth
MPAR06	f_L	= low baseband frequency
MPAR07	f_H	= high baseband frequency
MPAR08		(OPEN)
MPAR09		(OPEN)
MPAR10		(OPEN)
MPAR11		(OPEN)
MPAR12	T	= average talker level
MPAR13		(OPEN)
MPAR14	G	= companding gain
MPAR15		(OPEN)
MPAR16		(OPEN)

TABLE 3-2

FDM/FM MODULATION PARAMETERS

MPAR01	B#	= B serves to designate FDM/FM, # serves to index FDM/FM cases
MPAR02	FDM/FM	= identifies what B above means
MPAR03	N_c	= number of voice channels
MPAR04	B_{all}	= allocated bandwidth
MPAR05	B_{occ}	= occupied bandwidth
MPAR06	f_L	= low baseband frequency
MPAR07	f_H	= high baseband frequency
MPAR08	m	= rms multichannel modulation index
MPAR09	D_{RMS}	= rms multichannel frequency deviation
MPAR10	Λ	= peak/average multichannel baseband power ratio
MPAR11	β	= rms multichannel phase deviation
MPAR12	T	= average talker level
MPAR13	B_{DP}	= carrier energy dispersal bandwidth
MPAR14	G	= companding gain
MPAR15		(OPEN)
MPAR16	C,N	= flag designating companded (C) or not (N)

TABLE 3-3

TV/FM MODULATION PARAMETERS

MPAR01	C#	=	C serves to designate TV/FM, # serves to index TV/FM cases
MPAR02	TV/FM	=	identifies what C above means
MPAR03	1,2	=	number of TV channels per transponder
MPAR04	B_{all}	=	allocated bandwidth
MPAR05	B_{occ}	=	occupied bandwidth
MPAR06	f_{LV}	=	low baseband frequency of video
MPAR07	f_{HV}	=	high baseband frequency of video
MPAR08	M_V	=	peak modulation index of video
MPAR09	D_V	=	peak frequency deviation of video
MPAR10	D_{RSS}	=	rss of peak frequency deviation of video, audio subcarrier and dispersal waveform
MPAR11	f_{SC}	=	subcarrier frequency of highest audio subcarrier
MPAR12	b_{SC}	=	modulated bandwidth of highest audio subcarrier
MPAR13	D_{DP}	=	peak frequency deviation of dispersal waveform
MPAR14	G	=	gain (noise weighting and preemphasis)
MPAR15	V	=	TV-standard constant
MPAR16	U,L,F	=	flag designating dispersal type (U = unknown, L = line rate, F = frame rate)

TABLE 3-4

DIGITAL PSK MODULATION PARAMETERS

MPAR01	D#	=	D serves to designate DIG/PSK, # serves to index DIG/PSK cases
MPAR02	DIG/MPSK	=	identifies what D above means, including the number (M) of phase states
MPAR03	M	=	number of phase states
MPAR04	B_{all}	=	allocated bandwidth
MPAR05	B_{occ}	=	occupied bandwidth
MPAR06	α	=	unit power normalization factor, accounts for band-limited transmission
MPAR07	B_r	=	bit rate
MPAR08			(OPEN)
MPAR09			(OPEN)
MPAR10			(OPEN)
MPAR11			(OPEN)
MPAR12			(OPEN)
MPAR13			(OPEN)
MPAR14	G	=	coding gain
MPAR15	L	=	implementation loss
MPAR16			(OPEN)

TABLE 3-5

SCPC/PSK MODULATION PARAMETERS

MPAR01	E	=	E serves to designate SCPC/PSK # serves to index SCPC/PSK cases
MPAR02	SCPC/MPSK	=	identifies what E above means, including the number (M) of phase states
MPAR03	M	=	number of phase states
MPAR04	B_{all}	=	allocated bandwidth
MPAR05	B_{occ}	=	occupied bandwidth
MPAR06	α	=	unit power normalization factor, accounts for band-limited transmission
MPAR07	B_r	=	bit rate
MPAR08			(OPEN)
MPAR09			(OPEN)
MPAR10			(OPEN)
MPAR11			(OPEN)
MPAR12			(OPEN)
MPAR13			(OPEN)
MPAR14	G	=	coding gain
MPAR15	L	=	implementation loss
MPAR16			(OPEN)

TABLE 3-6

SCPC/FM MODULATION PARAMETERS

MPAR01	F#	=	F serves to designate SCPC/FM, # serves to index SCPC/FM cases
MPAR02	SCPC/FM	=	identifies what F above means
MPAR03	N	=	number of voice channels
MPAR04	B_{all}	=	allocated bandwidth
MPAR05	B_{occ}	=	occupied bandwidth
MPAR06	f_L	=	low baseband frequency
MPAR07	f_H	=	high baseband frequency
MPAR08	M_{PK}	=	peak modulation index
MPAR09	D_{PK}	=	peak frequency deviation
MPAR10	λ	=	peak/rms frequency deviation ratio
MPAR11			(OPEN)
MPAR12	P	=	loading factor
MPAR13			(OPEN)
MPAR14	G	=	gain (companding, noise weighting, preemphasis)
MPAR15			(OPEN)
MPAR16			(OPEN)

RTC Driver

The RTC driver prepares and delivers the pertinent data into the RTC algorithms. The MPAR data sets retrieved from the data base are associated with their corresponding worst C/I values, via the network and modulation designators previously discussed.

The carrier frequency offset (F_o) between desired and interfering signal pairs and the received carrier-to-noise density (C/N_o) in the desired paths involved are also included in the data sets fed to the RTC algorithms. The latter have a compact or detailed computation option selected by the user, and the RTC driver passes the flag designator corresponding to the mode selected.

The inputs provided by the RTC driver to the RTC algorithms consist of the following data, which is provided for each desired and interferer modulation pair under consideration:

1. the two data sets of MPAR's representing the desired and interferer modulation specifications

2. the frequency offset (F_o) between the desired and interferer signals

3. the ratio of the desired-to-interferer signal power (C/I), referred to the desired receiver input and excluding any interferer spectral truncation by the receiver passband

4. the ratio of the desired signal power to the thermal noise power density (C/N_o), referred to the desired receiver input

5. a binary flag designating the user selection of the compact or detailed computation mode.

RTC Algorithms

The RTC algorithms compute the output baseband degradation experienced by a desired signal in the presence of an interfering signal. The output baseband degradation varies with the desired and interfering modulation specifications, since the receiver processing is governed by the desired modulation and produces distinct effects in the output baseband according to the interferer modulation.

A distinct RTC algorithm is automatically selected to match each of the 36 possible combinations of desired and interferer modulation types. Each RTC algorithm also computes a specific baseband performance degradation measure according to the desired modulation type, (e.g., pWOp = picowatts of output interference for analog telephony, BER = bit error rate for digital data).

The output performance obtained in each case is compared to a preset threshold value. The performance threshold margin obtained is then used to compute the corresponding C/I margin (dB differential) referred to the receiver input. Hence, each case may have its distinct output performance measure, but they are all translated into a common index (the C/I threshold margin in dB) to permit sensitivity comparisons and worst-case identification.

The input/output operation of the RTC algorithms is fully automated once a compact or detailed computation mode is selected. The RTC driver delivers all the MPAR data sets and associated data (C/I, C/N_o, F_o), along with the four designators (i,j,k,l) previously discussed. The RTC algorithms logic recognizes each modulation type pair (k,l) to select and execute the appropriate routine dedicated to that pair, and then recycles over all possible network pair (i,j) and modulation pair (k,l) combinations. The output performance measures and C/I threshold margins computed are properly tagged by the four designators (i,j,k,l).

RTC Output

The output of the RTC algorithms consists of the following information for each desired and interfering network pair (i,j) and modulation pair (k,l) in the scenario:

1. the output performance measure (value and units) representing the output baseband degradation, as computed by the RTC algorithms

2. the actual C/I input (dB value) representing the worst-path condition at the desired receiver input, as computed from the ARP matrix

3. the C/I threshold and C/I threshold margin (dB values) referred to the desired receiver input, as computed by the RTC algorithms.

The C/I threshold margin is positive or negative depending on whether the actual C/I input is above or below the C/I threshold. The relation C/I (threshold margin) = C/I (input) - C/I (threshold) is always satisfied.

SECTION 4

APPLICATION OF THE RTC ALGORITHMS

PROCESSING FEATURES OF THE RTC ALGORITHMS

The purpose of this section is to illustrate the application of the RTC algorithms using representative examples. There are certain common designators and processing features shared by all the algorithms that are first described to develop a proper perspective of their application.

A unified notation is used for those modulation parameters and spectral characteristics pertinent to all algorithms, as shown in TABLE 4-1. The desired (C) and interferer (I) input powers are always referred to the desired receiver input. Any interferer spectral truncation is accounted for within the algorithm computations.

TABLE 4-1

RTC STANDARD NOTATION

C = desired signal power at desired receiver input

I = interferer signal power at desired receiver input

B_d = occupied bandwidth of desired emission spectrum

B_i = occupied bandwidth of interferer emission spectrum

B'_d = allocated bandwidth of desired emission spectrum

B'_i = allocated bandwidth of interferer emission spectrum

F_o = frequency offset between desired and interferer emission spectra

$S_d(f)$ = desired emission spectrum in normalized lowpass form

$S_i(f)$ = interferer emission spectrum in normalized lowpass form

$K_i(f)$ = effective interference spectrum (includes frequency offset and spectral truncation effects)

The desired $S_d(f)$ and interferer $S_i(f)$ emission spectra are specified in their normalized (unit power) form and generated according to the modulation type, as described in Section 5. Their occupied bandwidths are respectively denoted by B_d and B_i, and they are either directly specified or indirectly computed from the modulation parameters, as described in Section 5. The allocated bandwidths are respectively denoted by B_d' and B_i' to distinguish them from their occupied bandwidth counterparts.

A fundamental distinction is made between the interferer emission spectrum $S_i(f)$ and its effective interference spectrum $K_i(f)$. The latter is obtained by first shifting $S_i(f)$ relative to $S_d(f)$ by the frequency offset F_o, and then truncating this shifted replica according to the occupied bandwidth of the desired signal. The $K_i(f)$ spectrum represents the effective interference into the desired signal at the demodulator input.

The $K_i(f)$ spectrum may no longer be normalized, with the power truncation factor dependent on the B_d, B_i, and F_o values besides the spectral characteristic. If $B_i > B_d$, the spectrum $S_i(f)$ will be truncated regardless of the offset magnitude. If $B_i \leq B_d$, the spectrum $S_i(f)$ will be truncated if $F_o > 0.5\ |B_d - B_i|$, but not otherwise. If $F_o > 0.5\ (B_d + B_i)$, there is no spectral overlap and no interference is computed.

The truncated interferer power is given by $I' = \rho I$, where ρ is the spectral truncation factor obtained by integrating the spectrum $K_i(f)$ over the desired receiver passband. This spectrum varies with the different interferer modulation types via the emission spectrum $S_i(f)$, so that this factor is best computed within each algorithm as needed.

Another parameter of interest is the spectral overlap bandwidth (B_o) between the desired and interference signals, including the shift and truncation effects. This parameter is also dependent on the B_d, B_i, and F_o values, as shown in Equation 4-1. The value $B_o = 0$ corresponds to the no interference condition, in which case the algorithm computations are bypassed.

$$
\begin{aligned}
B_o \;=\;& \min\,(B_d, B_i), \quad \text{if} \quad F_o \leq |B_d - B_i| \\
& 0.5\,(B_d + B_i) - F_o, \quad \text{if} \quad 0.5\,|B_d - B_i| < F_o < 0.5\,(B_d + B_i) \\
& 0, \quad \text{if} \quad F_o \geq 0.5\,(B_d + B_i)
\end{aligned}
\tag{4-1}
$$

In the case of SCPC interference, the I value represents one of many possible interferers that can be contained within the desired signal bandwidth. When the SCPC interferers are narrowband relative to the desired signal, they will be assumed to fill the receiver passband with an activity factor to obtain realistic interference estimates in the algorithm computations.

The number of SCPC interferers that fits in the desired signal bandwidth is $N' = B_d/B_i'$ and this number is reduced to kN' by the activity factor (k). They will be characterized by a total power $(kN')I$ uniformly distributed over the bandwidth B_d, which corresponds to an effective interference spectrum $K_i = kN'/B_d = k/B_i'$ under normalized power conditions $(I = 1)$.

The $I' = \rho I$ relation can be used for multiple SCPC interference with a simple modification. The parameter $\rho = kN' \geq 1$ now represents a power increase factor due to the SCPC multiplicity, while the I' term represents the total interference power (accumulation instead of truncation) present in the desired signal bandwidth.

Example 1: All Modulation Types Under Cochannel Interference

This example illustrates the application of all the RTC compact algorithms under cochannel interference conditions. The six modulations shown in TABLE 4-2 are used to span all 36 combinations of desired (DES) and interferer (IFR) modulation pairs. The six modulations used represent typical communication services provided by the RCA SATCOM satellite network [Sharp, 1984].

A two-satellite, downlink-only, scenario was simulated to obtain the input C/I levels shown in TABLE 4-3 from the modulation-dependent power budgets. The modulation parameters of TABLE 4-2 and the input C/I levels of TABLE 4-3 were then fed to the RTC algorithms under cochannel conditions $(F_o = 0)$.

The output performance degradation computed by the compact versions of the RTC algorithms is shown in TABLE 4-4. The performance measure is the

TABLE 4-2

SATCOM MODULATION SPECIFICATIONS

RECORD	MPAR01	MPAR02	MPAR03	MPAR04	MPAR05	MPAR06	MPAR07	MPAR08
1	A1	CSSB/AM	5820	32.0 MHz	32.0 MHz	2 MHz	34.0 MHz	----
2	B1	CFDM/FM	2892	36.0 MHz	36.0 MHz	12 kHz	12.4 MHz	0.323
3	C1	TV/FM	1	36.0 MHz	33.0 MHz	25 kHz	4.2 MHz	2.560
4	D1	DIG/4PSK	4	36.0 MHz	28.6 MHz	1.33	60.0 MHz	----
5	E1	SCPC/2PSK	2	70.0 kHz	64.0 kHz	1.33	64.0 kHz	----
6	F1	SCPC/FM	-	47.8 kHz	36.8 kHz	300 Hz	3.4 kHz	4.412

RECORD	MPAR09	MPAR10	MPAR11	MPAR12	MPAR13	MPAR14	MPAR15	MPAR16
1	---	---	---	-21.0 dBm0	---	9.0 dB	---	----
2	4.00 MHz	10.00	6.6 rad	-29.1 dBm0	0.0 MHz	0.0 dB	---	C
3	10.75 MHz	10.98 MHz	6.8 MHz	0.7 MHz	1.0 MHz	12.8 dB	13.5 dB	U
4	---	---	---	---	---	0.0 dB	-4.4 dB	----
5	---	---	---	---	---	0.0 dB	-3.7 dB	----
6	15.00 kHz	---	---	-1.67 dB	---	16.0 dB	---	----

NOTE: The companding gain (MPAR14) is included in the talker level (MPAR12) for the FDM/FM case (but not for the CSSB/AM case) to match the original information source.

TABLE 4-3

C/I INPUT LEVELS FOR EXAMPLE 1

IFR \ DES	CSSB/AM	FDM/FM	TV/FM	DIG/PSK	SCPC/PSK	SCPC/FM
CSSB/AM	33.4 dB	37.4 dB	37.4 dB	37.4 dB	12.4 dB	3.4 dB
FDM/FM	29.4 dB	33.4 dB	33.4 dB	33.4 dB	8.4 dB	0.6 dB
TV/FM	29.4 dB	33.4 dB	33.4 dB	33.4 dB	8.4 dB	-0.6 dB
DIG/PSK	29.4 dB	33.4 dB	33.4 dB	33.4 dB	8.4 dB	-0.6 dB
SCPC/PSK	54.4 dB	58.4 dB	58.4 dB	58.4 dB	33.4 dB	24.4 dB
SCPC/FM	63.4 dB	67.4 dB	67.4 dB	67.4 dB	42.4 dB	33.4 dB

TABLE 4-4

OUTPUT PERFORMANCE VALUES AND C/I THRESHOLD MARGINS FOR EXAMPLE 1

IFR \ DES	CSSB/AM	FDM/FM	TV/FM	DIG/PSK	SCPC/PSK	SCPC/FM
CSSB/AM	145 pWOp (6.2 dB)	128 pWOp (6.7 dB)	37.4 dB C/I' (10.5 dB)	9.6×10^{-8} BER (8.6 dB)	1.1×10^{-7} BER (13.1 dB)	57 pWOp (10.2 dB)
FDM/FM	1162 pWOp (-2.9 dB)	346 pWOp (2.4 dB)	33.4 dB C/I' (6.5 dB)	1.9×10^{-7} BER (3.3 dB)	1.5×10^{-7} BER (4.1 dB)	456 pWOp (1.2 dB)
TV/FM	611 pWOp (-0.1 dB)	320 pWOp (2.7 dB)	26.2 dB PR (7.2 dB)	1.8×10^{-7} BER (3.6 dB)	8.8 dB PR (-9.4 dB)	3.7 dB PR (-4.3 dB)
DIG/PSK	516 pWOp (0.7 dB)	429 pWOp (1.5 dB)	33.4 dB C/I' (6.5 dB)	1.9×10^{-7} BER (3.3 dB)	1.2×10^{-7} BER (7.6 dB)	203 pWOp (4.7 dB)
SCPC/PSK	765 pWOp (-1.1 dBi)	193 pWOp (4.9 dB)	35.7 dB C/I' (8.7 dB)	1.0×10^{-7} BER (7.0 dB)	1.9×10^{-7} BER (4.4 dB)	288 pWOp (3.2 dB)
SCPC/FM	341 pWOp (2.5 dB)	35 pWOp (12.3 dB)	43.0 dB C/I' (16.1 dB)	8.4×10^{-8} BER (14.4 dB)	1.3×10^{-7} BER (13.4 dB)	156 pWOp (5.8 dB)

Notes: (1) The top entry is the baseband performance and the bottom entry is the C/I threshold margin.

(2) The DIG/PSK and SCPC/PSK columns assume C/N = 15 dB for M = 2 and C/N = 19 dB for M = 4.

picowatts of output interference power (pWOp) with a 0 dBm0 test-tone desired modulation for the CSSB/AM, FDM/FM, and SCPC/FM cases, the input signal-to-interferer power ratio with spectral truncation (C/I') or the protection ratio (P.R.) for the TV/FM cases, and the bit error rate (BER) for the DIG/PSK and SCPC/PSK cases.

The significance of the RTC algorithms is evident when comparing the C/I input levels in TABLE 4-3 to their corresponding C/I threshold margins shown in parentheses in TABLE 4-4. For example, there is a 4 dB differential in the C/I input level between DES = CSSB/AM and DES = FDM/FM with IFR = CSSB/AM. Yet this becomes a smaller 0.5 differential in the C/I threshold margin with the same performance threshold.

There is also a 4 dB differential in the C/I input level (TABLE 4-3) between DES = CSSB/AM and DES = FDM/FM with IFR = FDM/FM. However, this now becomes a larger 5.3 dB differential in the C/I threshold margin (TABLE 4-4) with the same performance threshold.

Example 2: FDM/FM Performance under Cochannel and Offset Interference

This example illustrates the application of the RTC detailed algorithms under cochannel and offset interference conditions using various FDM/FM emission spectra. The modulations shown in TABLE 4-5 were selected to provide three gaussian and three nongaussian spectra with different bandwidths.

These modulations represent typical communication services provided by the RCA SATCOM and AT&T COMSTAR satellite networks [Sharp, 1983]. The SATCOM signals have high modulation indices that validate a gaussian spectrum, while the COMSTAR signals have low indices that produce a nongaussian spectrum.

The signal spectra are illustrated in Figures 4-1 through 4-6, where the origin represents the carrier center frequency location, and the frequency axis units (f) represent the displacement away from the center frequency. These spectra were automatically generated by the FMSPC spectrum generation program described in Section 5.

The performance measure computed is the picowatts of output interferer power (pWOp) in a desired channel. The C/I threshold margin is also computed

TABLE 4-5

FDM/FM TELEPHONY SIGNALS AND MODULATION PARAMETERS

FDM/FM TELEPHONY SIGNALS

G1	=	SATCOM	FDM/FM	252/15 MHz	(gaussian spectrum)
G2	=	SATCOM	FDM/FM	432/25 MHZ	(gaussian spectrum)
G3	=	SATCOM	FDM/FM	972/36 MHz	(gaussian spectrum)
N1	=	COMSTAR	FDM/FM	360/15 MHz	(non-gaussian spectrum)
N2	=	COMSTAR	FDM/FM	600/25 MHz	(non-gaussian spectrum)
N3	=	COMSTAR	FDM/FM	1200/36 MHz	(non-gaussian spectrum)

MODULATION PARAMETERS (SEE TABLE 3-2)

FDM/FM Signal	N_c (# chan)	B_{all} (MHz)	B_{occ} (MHz)	f_L (kHz)	f_H (kHz)	m (rms)	D_{RMS} (kHz)	A (peak/ave)	B (rad)	T (dB)	B_{DP} (kHz)	G (dB)	C,N (flag)
G1	252	15.0	12.4	12	1052	1.548	1628	10.00	9.239	-16.0	0	0.0	N
G2	432	25.0	20.7	12	1796	1.501	2695	10.07	11.715	-16.0	0	0.0	N
G3	972	36.0	36.0	12	4028	1.104	4445	9.88	12.811	-16.0	0	0.0	N
N1	360	15.0	13.5	564	2044	0.500	1022	21.20	0.878	-15.0	2990	0.0	N
N2	600	25.0	22.5	564	3084	0.607	1871	19.05	1.179	-15.0	597	0.0	N
N3	1200	36.0	34.0	564	5772	0.486	2807	16.02	1.155	-15.0	605	0.0	N

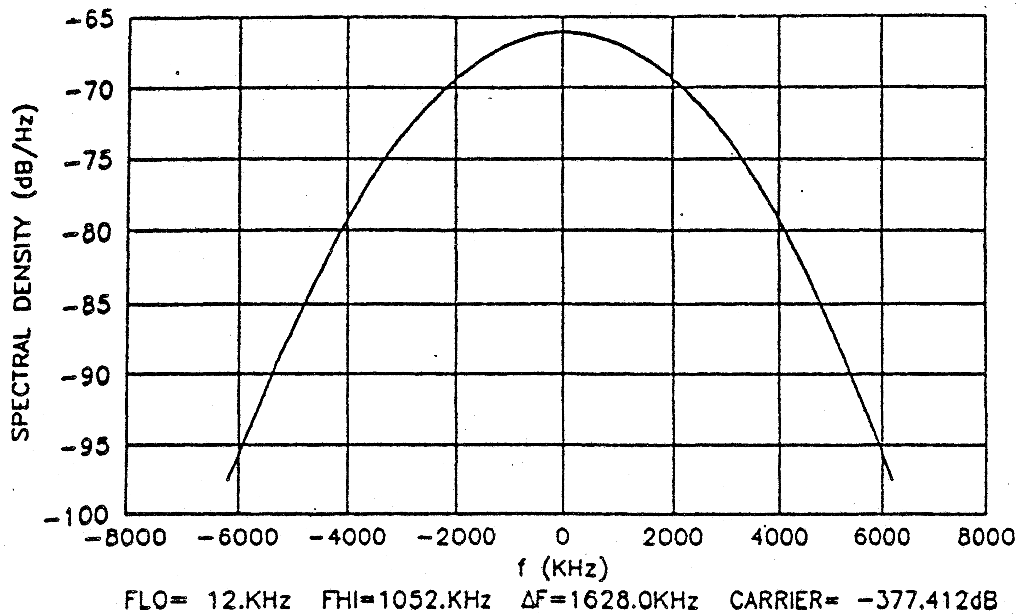


Figure 4-1. SATCOM FDM/FM 252/15 MHz spectrum.

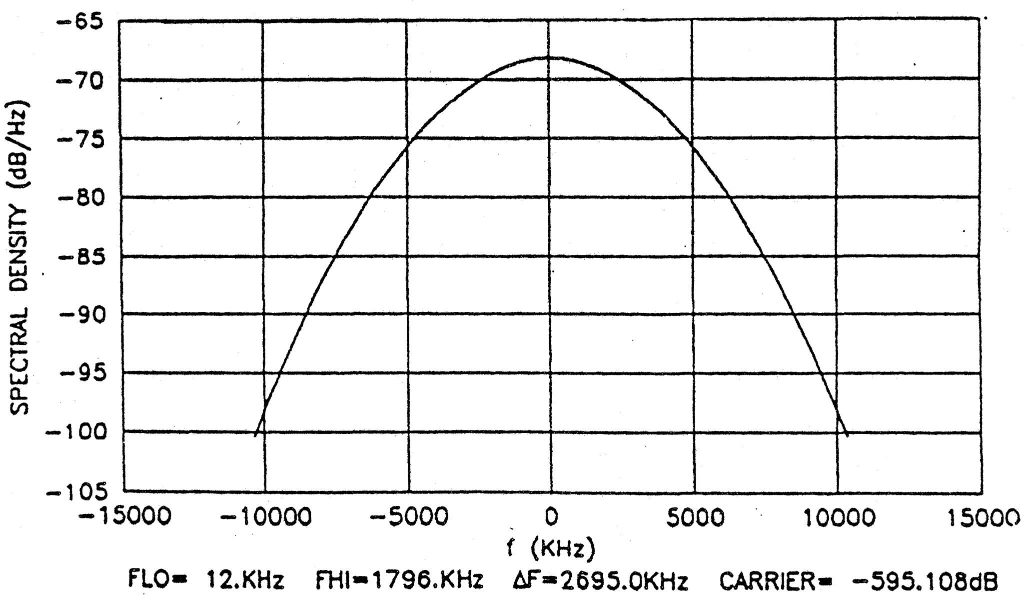


Figure 4-2. SATCOM FDM/FM 432/25 MHz spectrum.

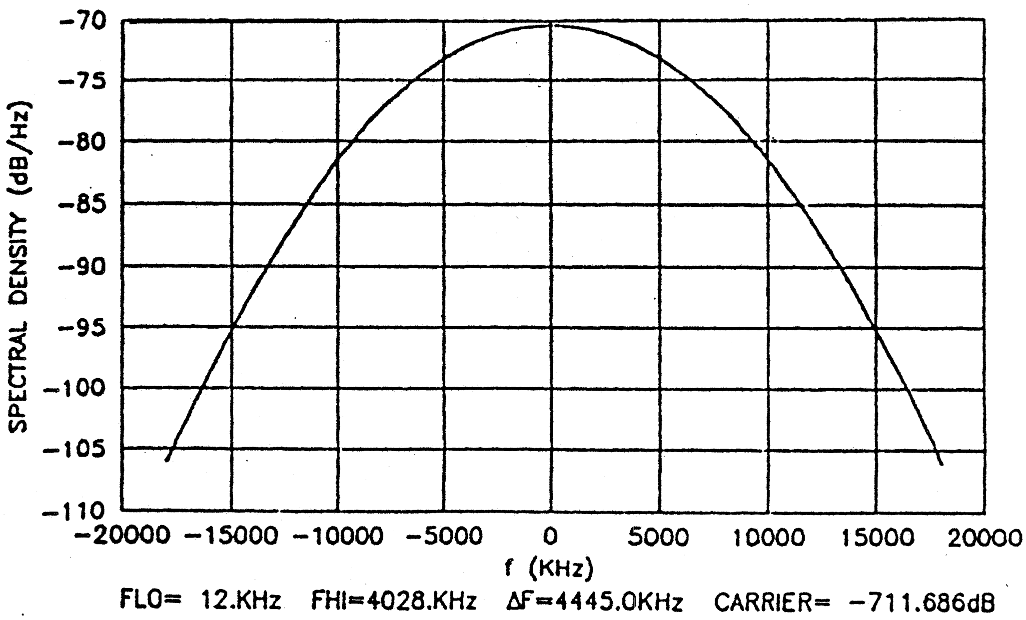


Figure 4-3. SATCOM FDM/FM 972/36 MHz spectrum.

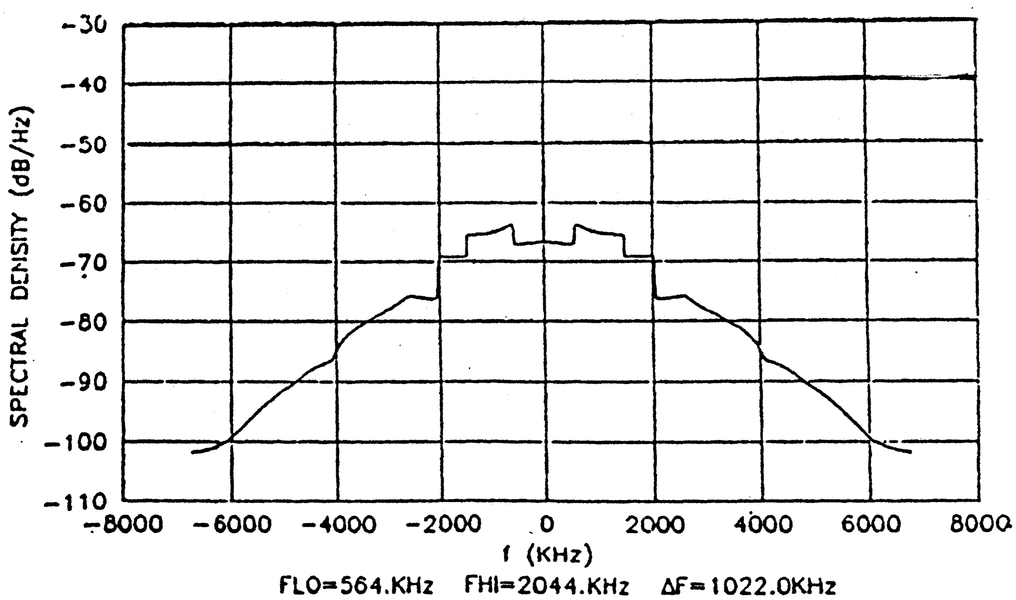


Figure 4-4. COMSTAR FDM/FM 360/15 MHz spectrum.

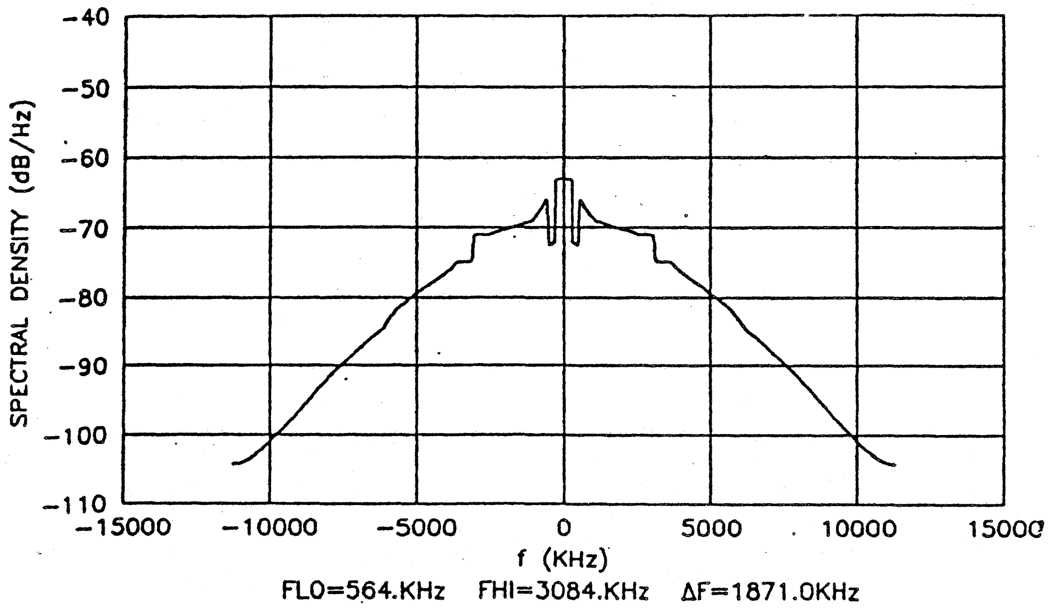


Figure 4-5. COMSTAR FDM/FM 600/25 MHz spectrum.

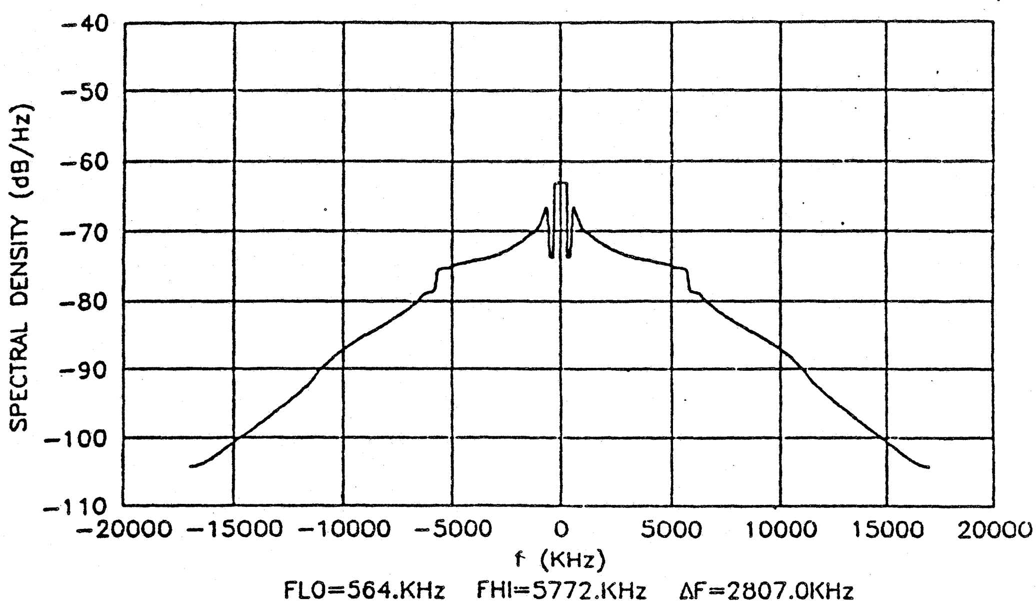


Figure 4-6. COMSTAR FDM/FM 1200/36 MHz spectrum.

assuming a 600 pWOp performance threshold. The performance evaluation requires the spectral convolution between the input spectra generated, but including any interferer spectral shift by the carrier offset or spectral truncation by the desired bandwidth. This convolution is automatically performed by the SPCVL spectral convolution program described in APPENDIX B.

The results obtained under various conditions are presented in Figures 4-7 through 4-18. Each figure consists of three pairs of graphs, with each pair showing the pWOp and C/I margin computed for the worst desired channel affected, as a function of the carrier offset and input C/I values. The three pairs per figure represent the three cases of narrow, medium, and wide input bandwidths as discussed below.

Figures 4-7 through 4-9 form a group having gaussian desired and interferer spectra. Figures 4-10 through 4-12 form another group having nongaussian desired and interferer spectra. Figures 4-13 through 4-15 form another group having nongaussian desired and gaussian interferer spectra. Figures 4-16 through 4-18 form another group having gaussian desired and nongaussian interferer spectra.

The first figure within each group (4-7, 4-10, 4-13, 4-16) maintains the desired signal invariant and varies the interferer signal bandwidth. The second figure within each group (4-8, 4-11, 4-14, 4-17) maintains the interferer signal invariant and varies the desired signal bandwidth. The third figure within each group (4-9, 4-12, 4-15, 4-18) maintains the desired and interferer bandwidths equal, but varies their value to span the narrow, medium, and wide cases.

This collection of results emphasizes the capability of the RTC algorithms to perform an exhaustive interference analysis or concentrate on specific parametric variations, while accounting for the distinct spectral characteristics of the signals involved. Another application would be to consider two (or more) signals to be designed, and vary certain modulation parameters to establish the performance sensitivity from a mutual interference standpoint.

The need for an accurate spectral generation and convolution is evident by comparing the all-gaussian group (Figures 4-7 through 4-9) with the all-nongaussian group (Figures 4-10 through 4-12). The former always has the

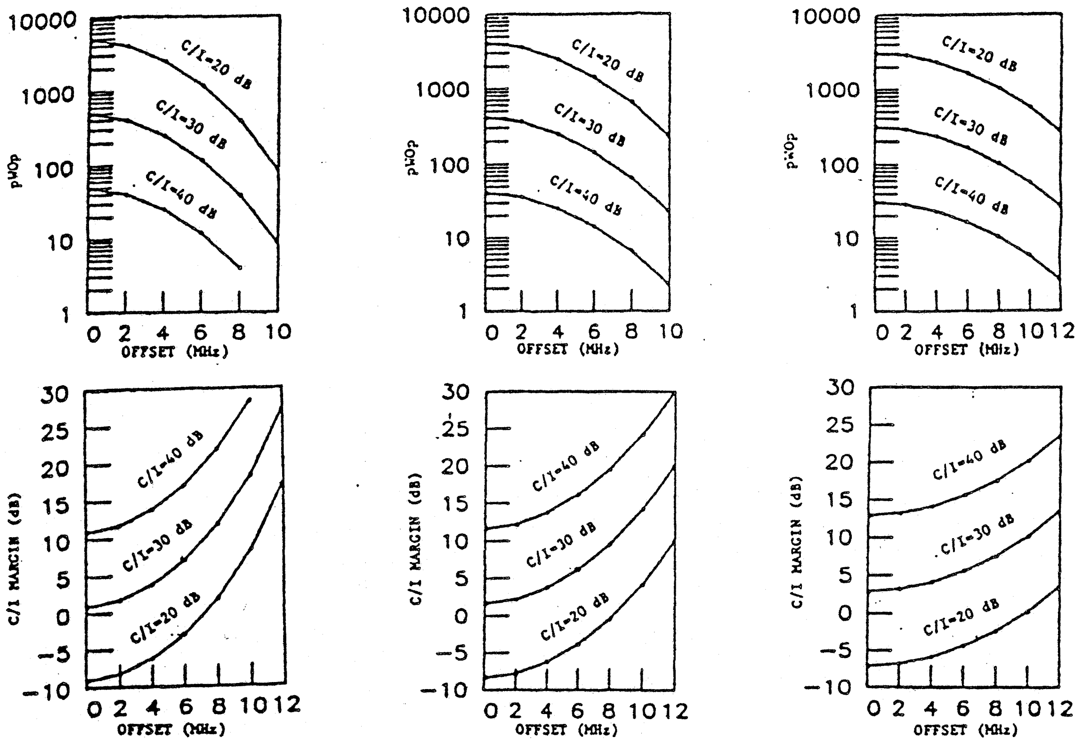


Figure 4-7. FDM/FM gaussian interferer into gaussian desired spectrum: desired invariant (G1 into G2, G2 into G2, G3 into G2).

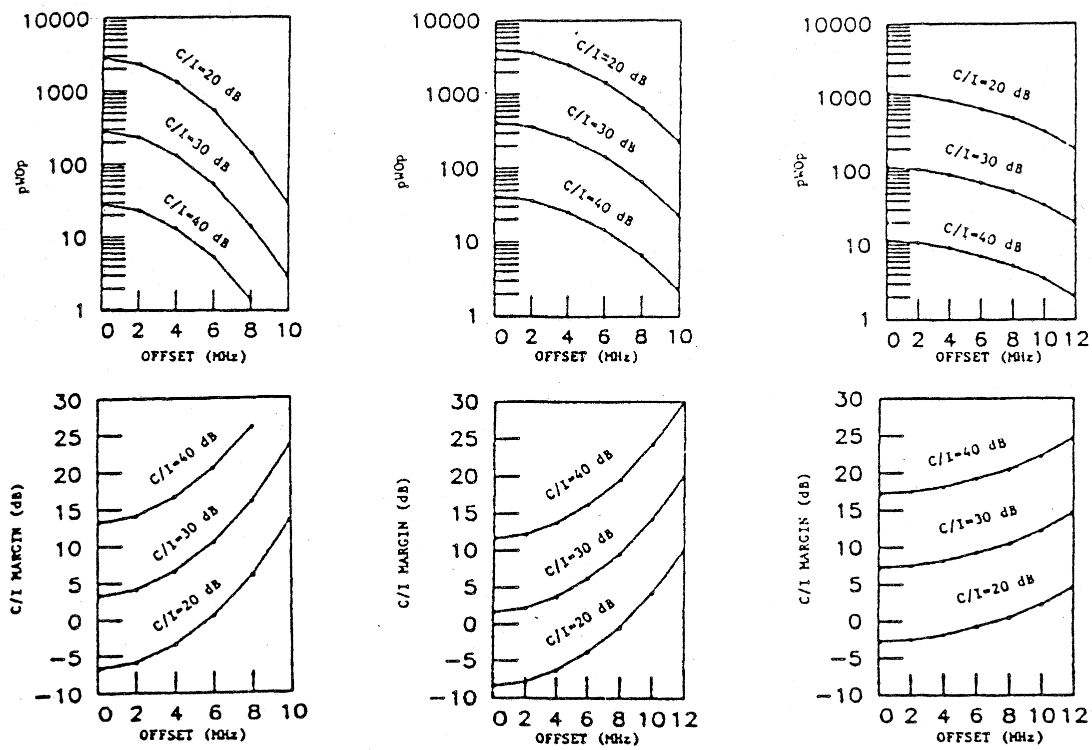


Figure 4-8. FDM/FM gaussian interferer into gaussian desired spectrum: interferer invariant (G2 into G1, G2 into G2, G2 into G3).

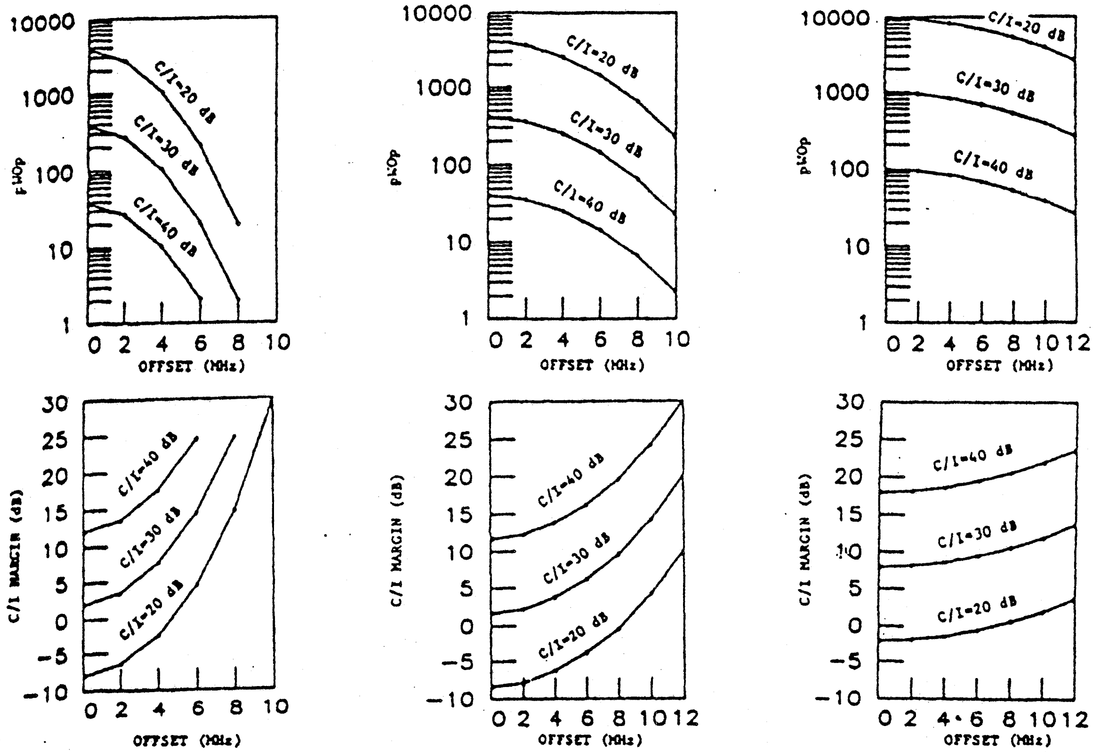


Figure 4-9. FDM/FM gaussian interferer into gaussian desired spectrum: desired = interferer (G1 into G1, G2 into G2, G3 into G3).

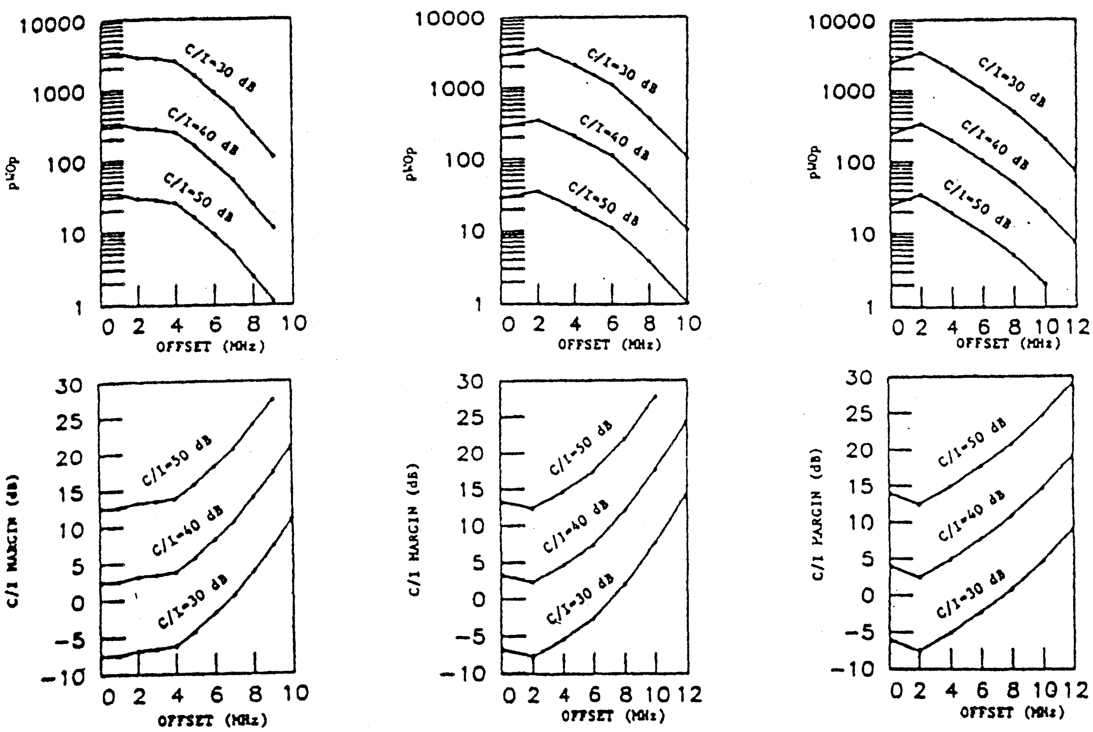


Figure 4-10. FDM/FM nongaussian interferer into nongaussian desired spectrum: desired invariant (N1 into N2, N2 into N2, N3 into N2).

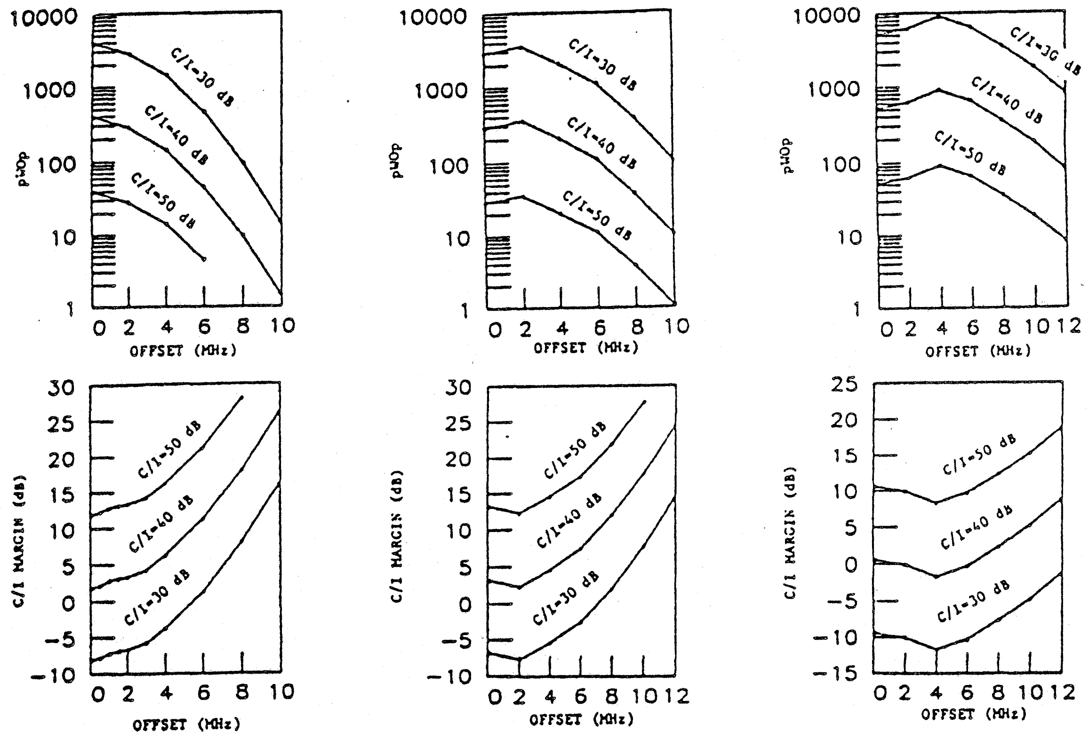


Figure 4-11. FDM/FM nongaussian interferer into nongaussian desired spectrum: interferer variant (N2 into N1, N2 into N2, N2 into N3).

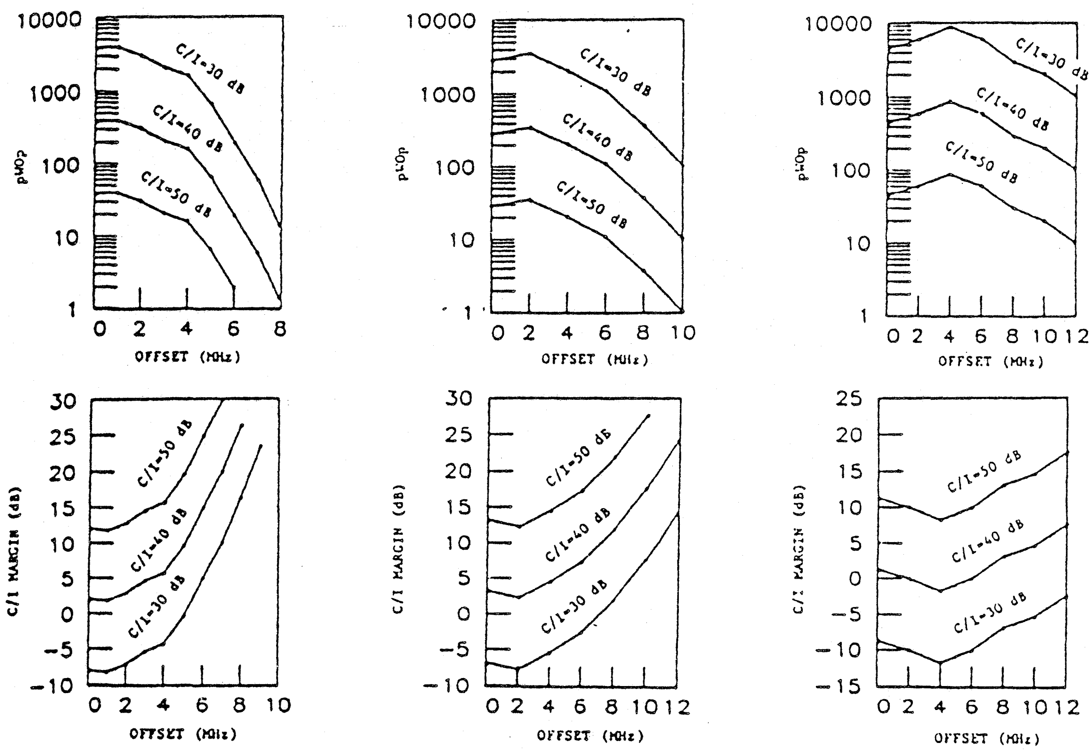


Figure 4-12. FDM/FM nongaussian interferer into nongaussian desired spectrum: desired = interferer (N1 into N1, N2 into N2, N3 into N3).

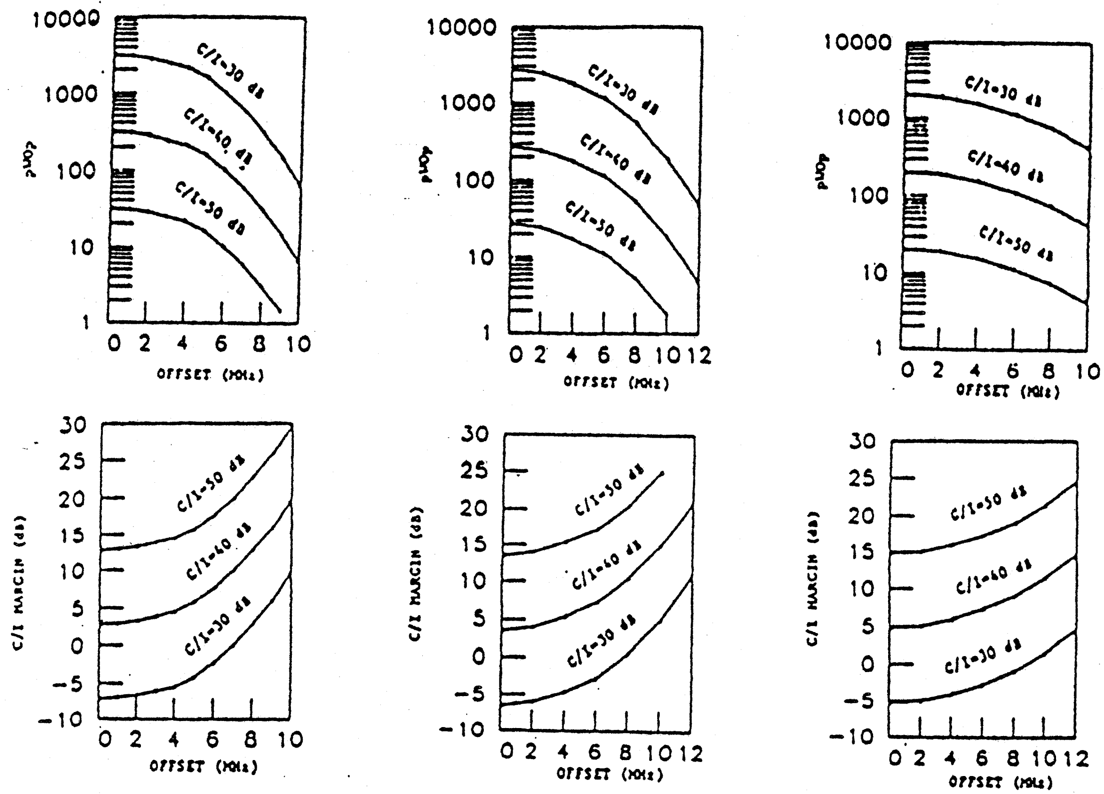


Figure 4-13. FDM/FM gaussian interferer into nongaussian desired spectrum: desired invariant (G1 into N2, G2 into N2, G3 into N3).

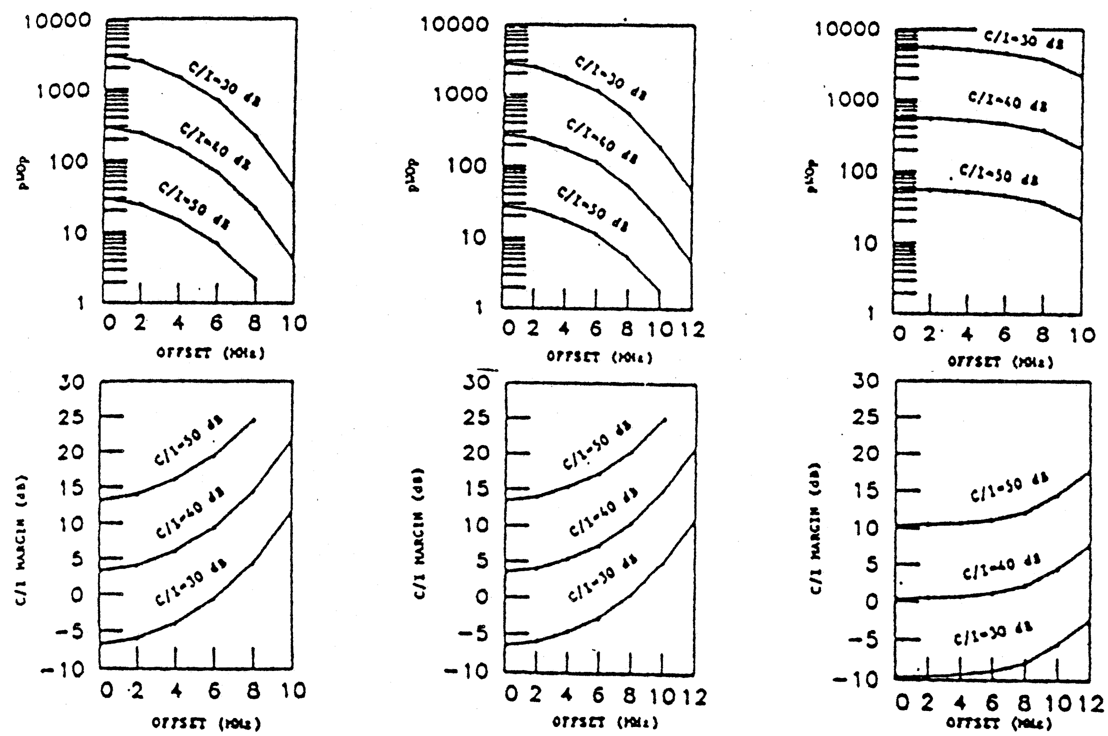


Figure 4-14. FDM/FM gaussian interferer into nongaussian desired spectrum: interferer invariant (G2 into N1, G2 into N2, G2 into N3).

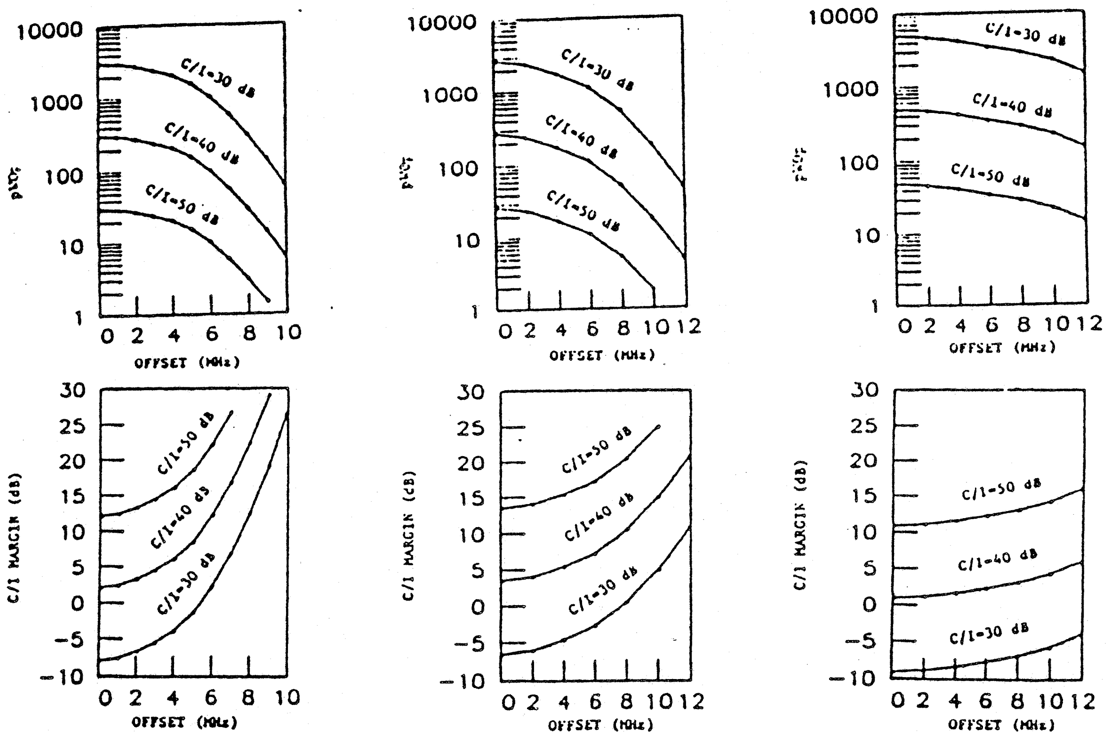


Figure 4-15. FDM/FM gaussian interferer into nongaussian desired spectrum: desired bandwidth = interferer bandwidth (G1 into N1, G2 into N2, G3 into N3).

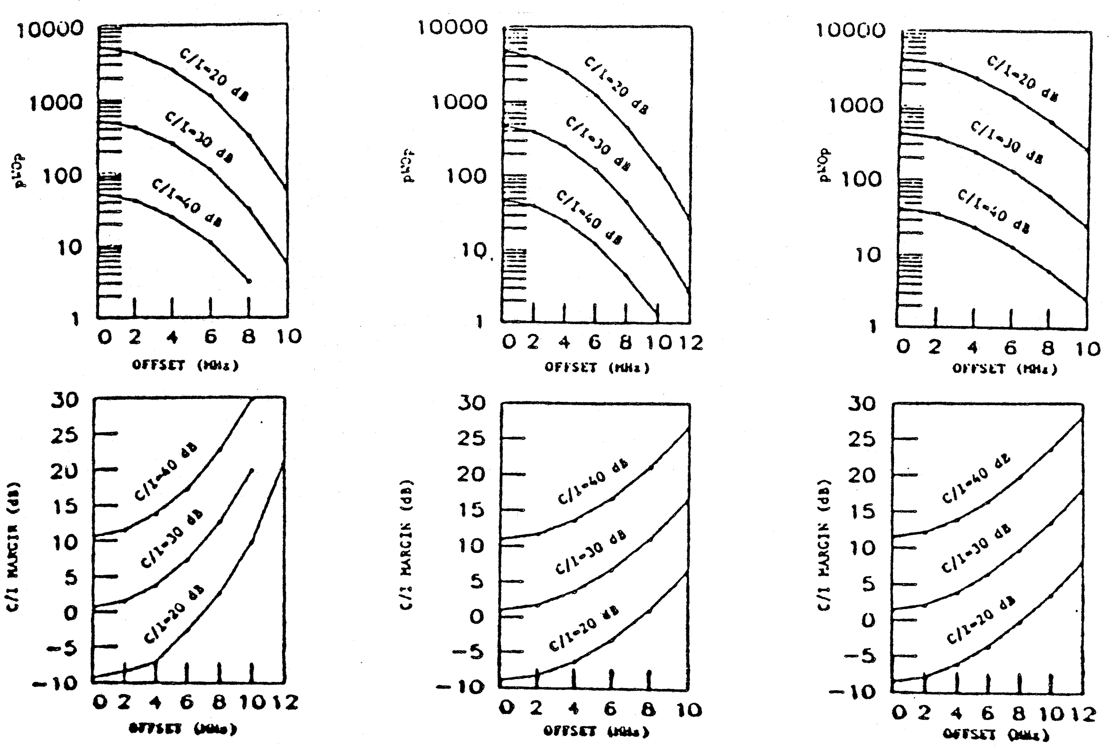


Figure 4-16. FDM/FM nongaussian interferer into gaussian desired spectrum: desired invariant (N1 into G2, N2 into G2, N3 into G2).

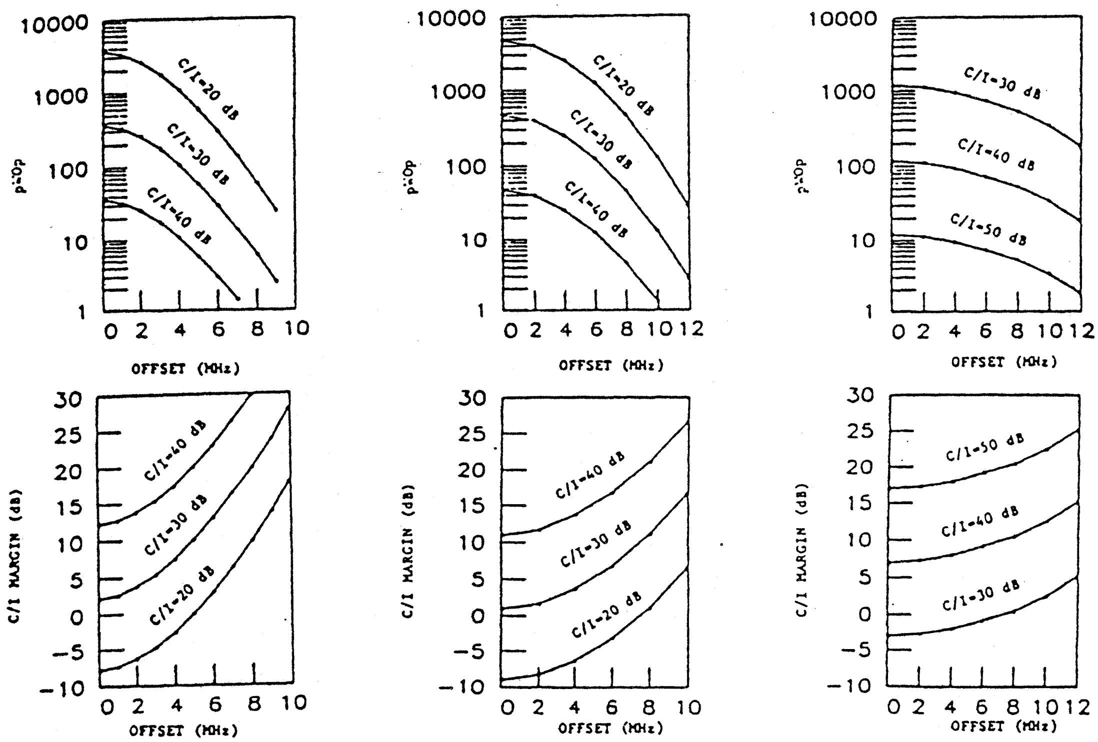


Figure 4-17. FDM/FM nongaussian interferer into gaussian desired spectrum: interferer invariant (N2 into G1, N2 into G2, N2 into G3).

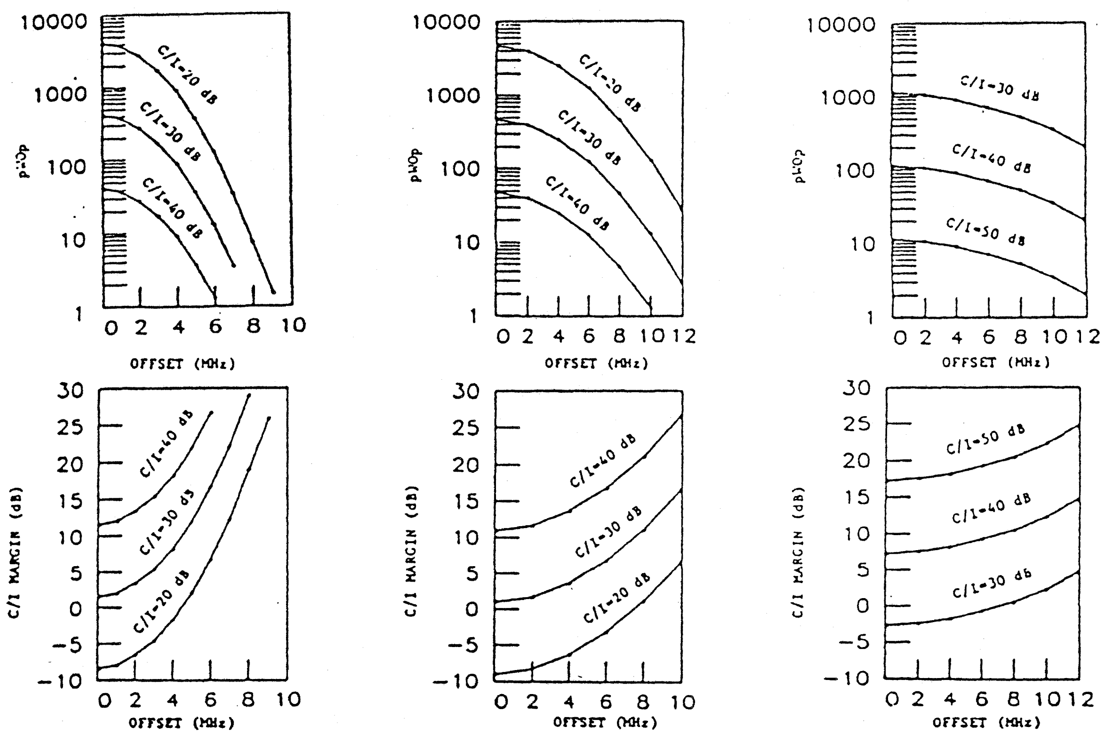


Figure 4-18. FDM/FM nongaussian interferer into gaussian desired spectrum: desired bandwidth = interferer bandwdith (N1 into G1, N2 into G2, N3 into G3).

cochannel condition as a worst occurrence, while the latter often has an offset condition as the worst occurrence. This happens because the side peaks in the nongaussian spectra do not coincide with two cochannel signals, unlike the central peaks of two gaussian spectra. Also, the spectral convolution tends to smooth the interferer output power, so that the worst offset value is not evident a priori.

Example 3: PSK Performance Under Varied Interference and Threshold Conditions

This example illustrates the application of the RTC algorithms under various interferer characteristics and threshold requirements using coherent PSK (CPSK) as the desired signal. The algorithms distinguish between white versus nonwhite interference spectra based on the modulation parameters to select the appropriate BER performance and C/I margin formulation, as described in APPENDIX D.

The BER performance is governed by an equivalent output SNR parameter (γ) that includes both thermal noise (C/N) and interference (C/I) effects. The C/N value affects the symbol energy to noise density ratio (E/N_o). The C/N to E/N_o conversion accounts for the number of phase states (M), the time-bandwidth product of the symbol duration (T) and the signal bandwidth (B_d), and the coding gain (G) and implementation loss (L) effects (see APPENDIX D).

The C/I value affects the BER performance in a different way depending on the white versus nonwhite interference characteristic, including any spectral truncation by the receiver passband. The nonwhite interference formulation is based on a simple upper bound to the actual performance that maintains the same BER order of magnitude in the 10^{-3} to 10^{-6} region, as discussed in APPENDIX D.

A comparison of the BER performance for white versus nonwhite interference is shown in Figure 4-19 for binary CPSK, with $B_dT = G/L = 1$ and no interference spectral truncation assumed for simplicity. The higher error performance exhibited by the white interference case as C/I decreases is a consequence of the distinct white and nonwhite interference effects, as discussed in APPENDIX D.

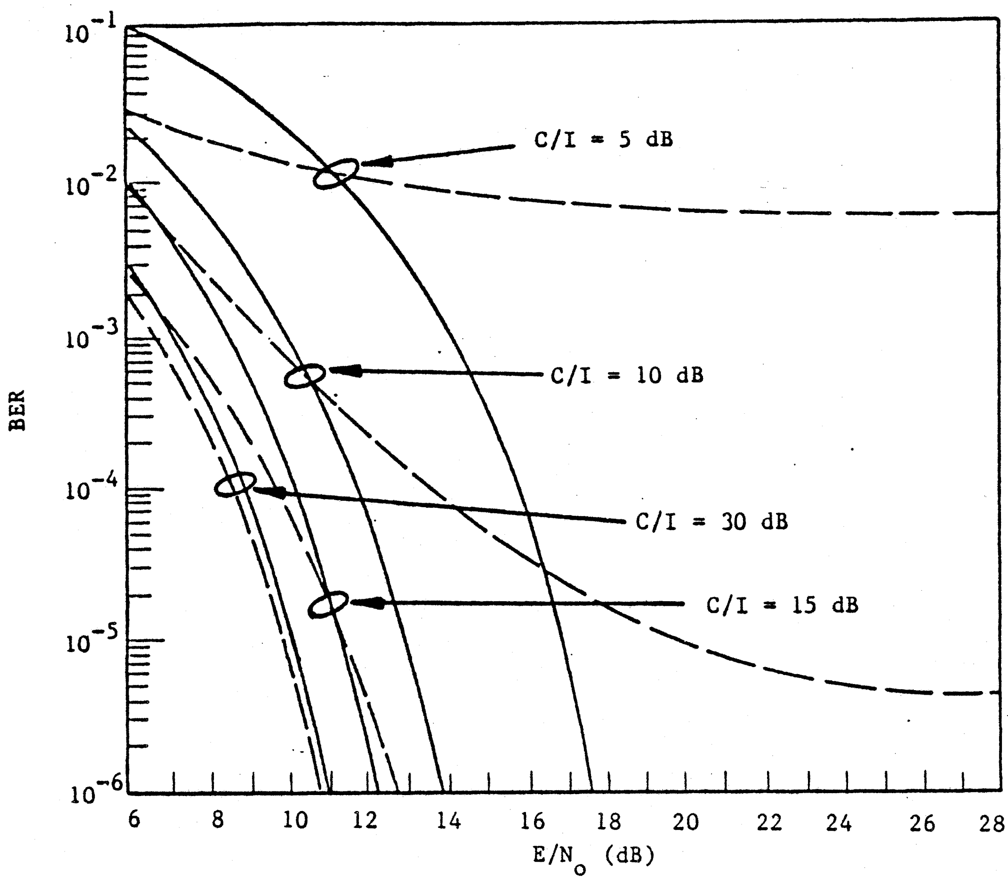


Figure 4-19. BER performance for white (dotted) and nonwhite (solid) interference into binary CPSK. (Note: upper bound used for nonwhite case.)

The input C/I threshold required to provide a given BER performance is shown in Figure 4-20. The C/I threshold variation with the BER performance requirement is shown in Figure 4-21. The vertical lines correspond to the specific E/N_O values that match the BER requirement, so that thermal noise alone causes the threshold performance. The vertical lines imply that there is no room for interference due to the thermal noise amount (i.e., the C/I threshold is infinite). The vertical line location shifts to the left as the BER requirement is relaxed, since more noise can then be tolerated before threshold occurs.

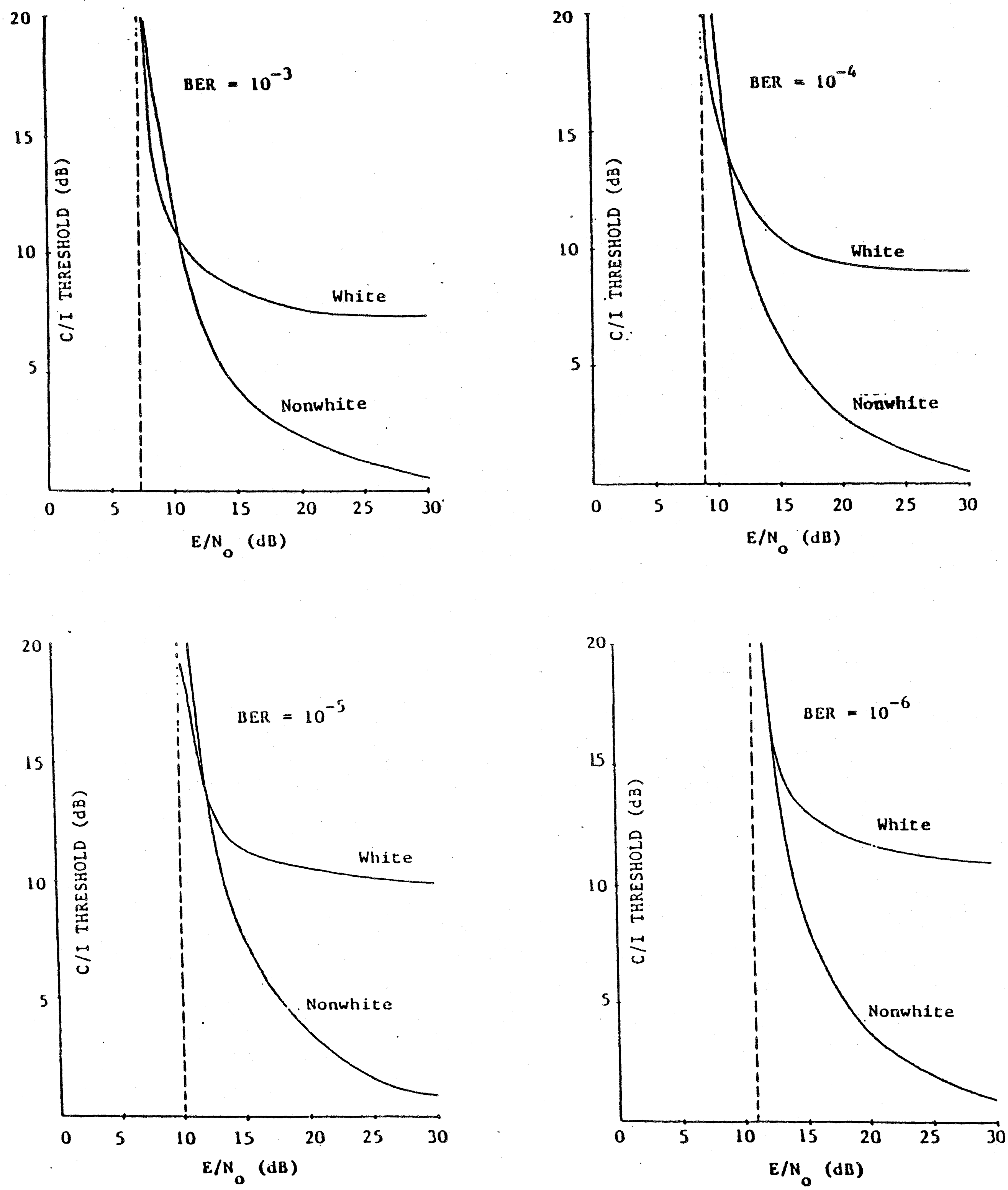


Figure 4-20. C/I threshold versus E/N_o for white and nonwhite interference into binary CPSK at various BER performance specifications.

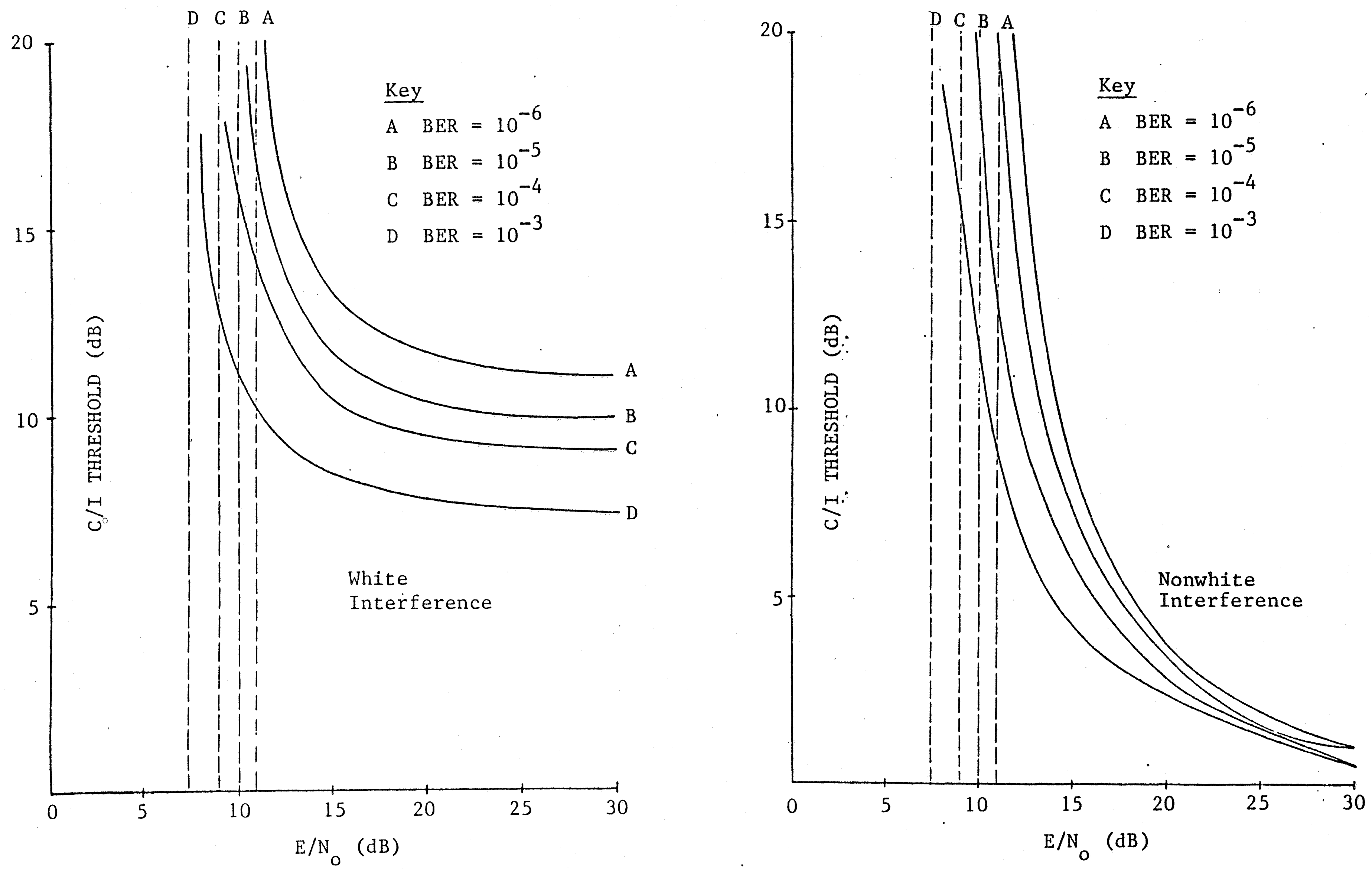


Figure 4-21. C/I threshold versus E/N_o and BER for white and nonwhite interference into binary CPSK.

SECTION 5

MODULATION SPECTRUM MODELS

GENERAL

A normalized power density spectrum is characterized in this section for each modulation type, to be used by the RTC algorithms, as needed. The spectrum is to be generated from the modulation parameter values specified for each modulation type, with the generation process ranging from a simple formula evaluation to an elaborate simulation process.

The normalized nature implies that the spectrum provided has unit power, which corresponds to unit area under the power density characteristic. A scaling factor is needed to maintain the normalization under band-limited transmission, though a unit factor can be assumed if the bandwidth limits represent a negligible power percentage in the spectrum tails truncated.

The normalized spectrum is provided in its lowpass equivalent form. This means that the spectrum shall be centered at zero frequency, with the actual transmission spectrum being a translated replica of this lowpass representation [Stein and Jones, 1967; Pontano et al, 1973]. The RTC algorithms need only generate such lowpass replicas and then accommodate the frequency offset and spectral truncation effects, as needed. This approach is compatible with the CCIR spectral representation for interference analysis [CCIR Report 388-4, 1982].

The provision of a simple formula for a given modulation spectrum does not necessarily imply a compact RTC algorithm implementation when such modulation is involved. The algorithm may include other computation processes besides spectral generation that require elaborate processing, even if the spectral generation itself is a simple formula evaluation. For example, the algorithm may involve spectral convolution once the spectra are generated and spectral integration after the spectra are convolved.

CSSB/AM TELEPHONY SPECTRUM

The signal baseband consists of a number (N_c) of telephony channels that are frequency multiplexed to form a multichannel baseband spectrum with specified low (f_L) and high (f_H) baseband frequencies. The individual channel center frequencies are determined from these parameters as given by Equation 5-1.

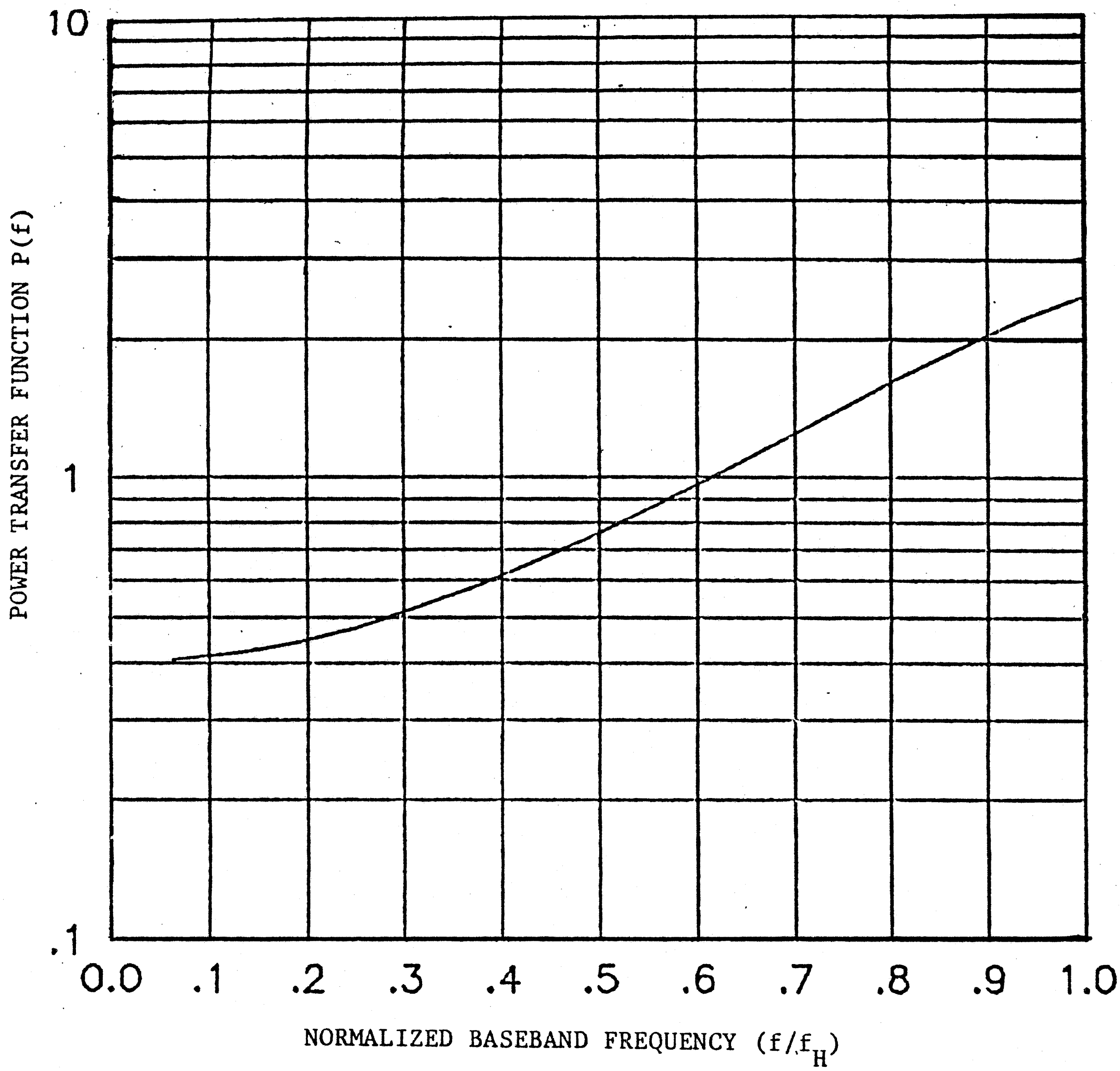
$$f_j = f_L + \frac{2j - 1}{2N_c}(f_H - f_L), \quad j = 1 \text{ to } N_c \tag{5-1}$$

The baseband spectrum is characterized as a uniform power distribution over its frequency range (f_L, f_H). The modulation process preserves this characteristic when the spectrum is linearly translated for transmission as single sideband on a suppressed carrier. The normalized emission spectrum is represented by a rectangle with a two-sided magnitude of $S(f) = 1/2B$ for $f_L < |f| < f_H$, where $B = f_H - f_L$ is the occupied bandwidth.

The center frequency of the emission spectrum is $0.5\,(f_H - f_L)$ away from the unmodulated carrier frequency. This center frequency is used when specifying frequency offsets between desired and interfering signals that involve CSSB/AM modulation.

FDM/FM TELEPHONY SPECTRUM

The original baseband is identical to that previously stated for CSSB/AM telephony, and Equation 5-1 again specifies the channel frequencies. However, the uniform multichannel baseband spectrum is now preemphasized and applied as frequency modulation on a transmission carrier. The preemphasis characteristic is shown in Figure 5-1 based on CCIR recommendations, and ideally preserves the rms frequency deviation of the multichannel baseband modulation so that it is identical before and after preemphasis [CCIR Recommendation 464-1, 1982; Panter, 1972].



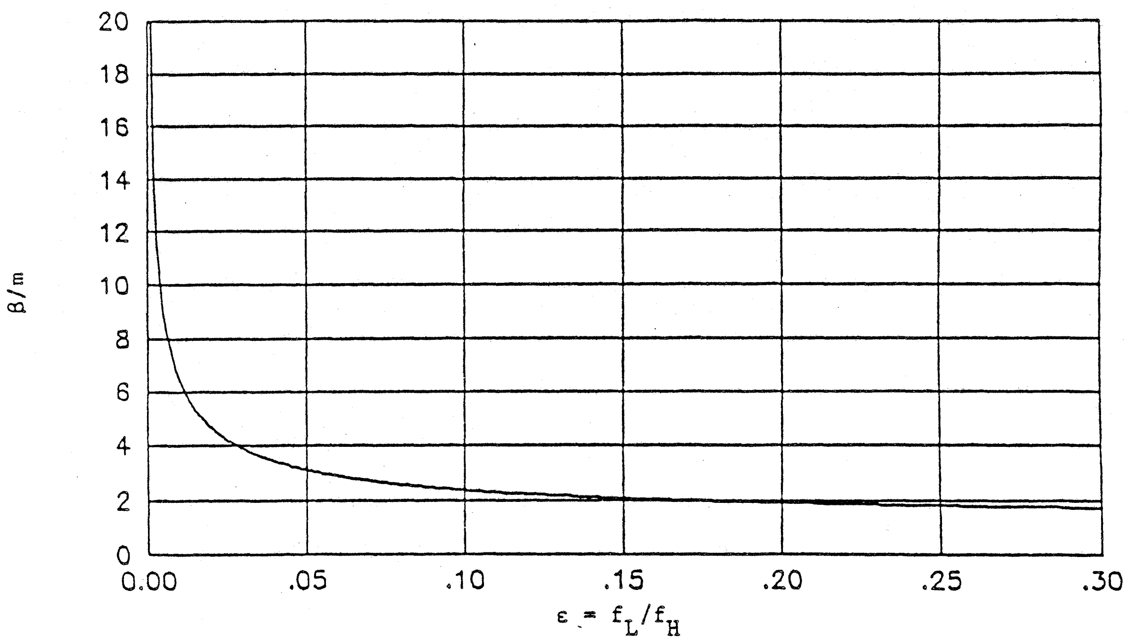
$$P(f) = (0.4) \cdot \frac{1 + 2.0796(f/f_H)^2 + 0.4096(f/f_H)^4}{1 - 0.8545(f/f_H)^2 + 0.4096(f/f_H)^4}$$

Figure 5-1. FDM/FM preemphasis characteristic.

The FDM/FM signal spectrum has a simple formulation under wideband FM conditions (i.e., high modulation index). The FDM/FM spectrum is then represented by a gaussian characteristic centered at the carrier frequency and fully specified by its standard deviation parameter (σ). The latter equals the rms multichannel deviation (D_{RMS}) of the preemphasized baseband, so that the normalized FDM/FM spectrum is given by Equation 5-2 [Schwartz, Bennett, and Stein, 1966].

$$S(f) = \frac{1}{\sqrt{2\pi}(\sigma)} e^{-0.5(f/\sigma)^2} , \text{ where } \sigma = D_{RMS} \tag{5-2}$$

The validity of the gaussian spectral representation depends on the modulation parameter values. An effective validation criterion is based on the magnitude of the rms phase deviation (β) of the preemphasized baseband. This parameter is a function of both the rms modulation index ($m = D_{RMS}/f_H$) and the low-to-high baseband frequency ratio ($\varepsilon = f_L/f_H$), as shown in Figure 5-2. The gaussian spectral representation has been validated via extensive



$$\beta \approx \frac{0.63m}{\sqrt{\varepsilon(1-\varepsilon)}} [1+2.89\varepsilon-3.17\varepsilon^2-0.72\varepsilon^4]^{1/2} \quad \text{(radian)}$$

Figure 5-2. RMS phase deviation of FDM/FM preemphasized baseband.

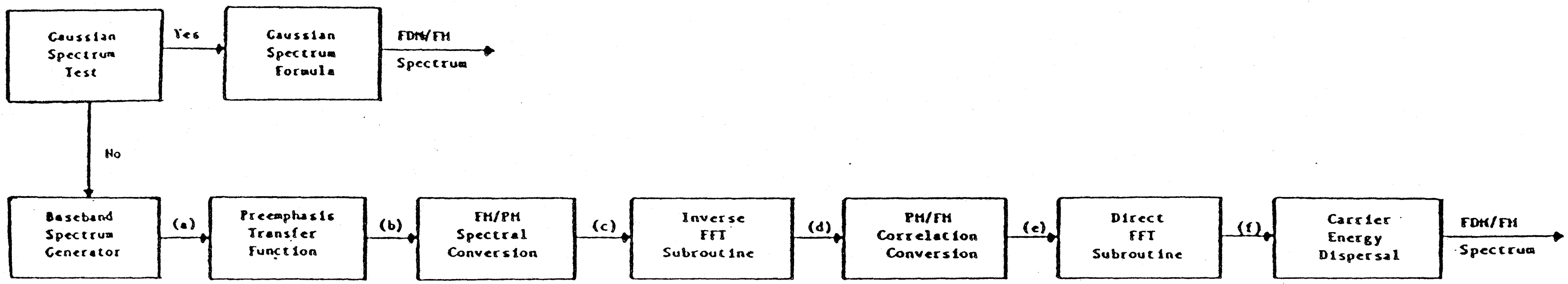
spectral analysis and computer simulation to be accurate over the predominant spectral region when $\beta \geqq 1.5$ radian. The criterion serves to account for low-baseband frequency effects (not reflected by the modulation index) that can significantly affect the spectral characteristic in certain cases of interest [Filippi, 1983].

The FDM/FM signal spectrum also has a simple formulation under narrowband FM conditions (i.e., low modulation index). The FDM/FM spectrum is then a translated replica of the preemphasized baseband spectrum, plus a residual carrier component. However, these narrowband conditions are not typical of the FDM/FM signals of interest. There remain the cases where neither wideband nor narrowband conditions exist. The FDM/FM spectrum does not have simple formulation in such cases.

NTIA has designed and implemented an FDM/FM Spectrum Generation Program (FMSPC) that simulates the modulation process and generates an accurate spectral representation in all cases [Filippi, 1983]. The program automatically recognizes the wideband FM cases by examining the modulation specifications and replaces the modulation process simulation by the gaussian formula evaluation when applicable, so as to simplify and expedite the spectral generation process.

The FMSPC spectrum generation program is illustrated in Figure 5-3. The FDM/FM modulation parameters are first analyzed to decide if a gaussian spectrum is valid, in which case the simple gaussian formula is used. Otherwise, the FDM/FM simulation process is engaged to generate the spectral samples. The process corresponds to a spectral convolution series, but it is easily implemented using Fast Fourier Transforms (FFTs) and correlation function properties, so as to avoid the multiple spectral convolutions otherwise required [Filippi, 1983].

The occupied bandwidth of the FDM/FM signal is to be based on Carson's Rule. The rms multichannel deviation can be converted into a peak multichannel deviation via $D_{PK} = \lambda\ (D_{RMS})$, where λ is the peak-to-rms frequency deviation, and $\Lambda = \lambda^2$ is the peak-to-average power ratio of the pre-emphasized baseband. The FDM/FM signal bandwidth is then given by $B = 2\ (\lambda\ D_{RMS} + f_H)$. A value of $\Lambda = 10$ is often used in practice when this parameter is not specified.



(a) Normalized Power Spectrum of Multichannel Baseband Signal: $S_B(f)$

(b) Normalized Power Spectrum of Frequency-Modulating Signal: $P(f) \cdot S_B(f)$

(c) Power Spectrum of Equivalent Phase-Modulating Signal: $S_P(f) = \left(\frac{D_{RMS}}{f}\right)^2 \cdot P(f) \cdot S_B(f)$

(d) Correlation Function of Equivalent Phase-Modulating Signal: $R_P(t)$ = Inverse Transform of $S_P(f)$

(e) Normalized Correlation Function of Frequency Modulated Signal: $R_F(t) = \mathrm{Exp}[-R_P(0) + R_P(t)]$

(f) Normalized Power Spectrum of Frequency Modulated Signal: $S_F(f)$ = Direct Transform of $R_F(t)$

Figure 5-3. FMSPC spectrum generation program for FDM/FM multichannel telephony.

The FDM/FM spectrum is to be restricted by the occupied bandwidth of the FDM/FM signal. The use of Carson's Rule as a bandwidth measure has been noted to be conservative under wideband conditions, so that the spectral tails removed by the occupied bandwidth limits have negligible power content. Hence, the bandwidth restriction can be bypassed to use the entire gaussian function for the desired signal, if deemed convenient for RTC algorithm simplification purposes.

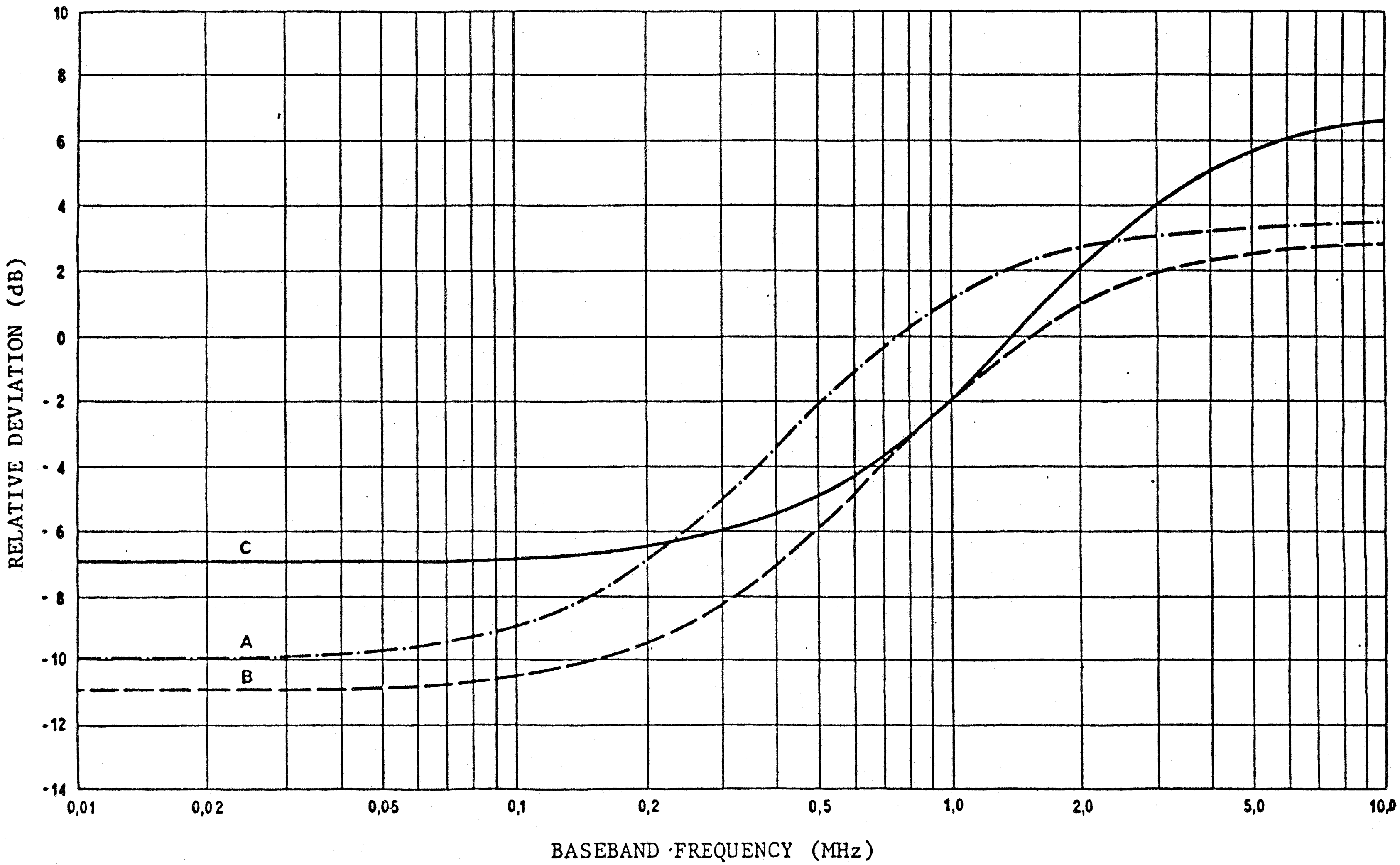
Conversely, the receiver passband may significantly truncate a gaussian interferer spectrum due to frequency offset or distinct bandwidth conditions. The use of the entire interferer spectrum can yield conservative interference estimates, but will preserve the simple algorithm formulation (e.g., gaussian-gaussian spectral convolution formula). A truncated gaussian spectrum will improve the computational accuracy, but may require an elaborate algorithm (e.g., truncated gaussian-gaussian spectral convolution routine).

TV/FM SPECTRUM

The signal baseband consists of the video component, the modulated audio subcarriers, and the energy dispersal waveform. This composite baseband is usually preemphasized prior to frequency modulation for transmission purposes. The preemphasis characteristics recommended by the CCIR are illustrated in Figure 5-4 for various television standards [CCIR Recommendation 405-1, 1982].

The video component itself contains two additive signals representing the luminance (luma) and chrominance (chroma) information. The luma signal is a lowpass baseband, while the chroma signal consists of two lowpass basebands simultaneously applied as quadrature amplitude modulations on a chroma subcarrier frequency (3.58 MHz for 525-line M/NTSC standard).

The TV/FM spectral characterization remains an open issue in the telecommunications community. The TV/FM spectral measurements show a notable variation with the modulation specifications, and investigations to characterize the TV/FM spectrum are underway [Miller, 1984]. There does not exist an accepted formulation or simulation to generate the TV/FM spectrum, while accommodating all modulation components, parametric dependences, and



Key:
Curve A: 525-line system
B: 625-line system
C: 819-line system

Figure 5-4. TV/FM preemphasis characteristics.

statistical properties in the composite baseband. The theoretical models and experimental results available in the open literature usually simplify the baseband modulation and/or restrict the parametric variations, which hinders a generalized spectral characterization.

One recent approach decomposes the video baseband into its luma and chroma spectral contributions. The TV/FM spectrum with video-only modulation is derived under certain assumptions (low-index luma, high-index chroma, gaussian video statistics) as a spectral convolution series and predominant terms are kept to obtain a simple parametric formula [Ali, 1983]. The normalized spectrum has a gaussian central lobe at the carrier frequency and a gaussian sidelobe shifted by the chroma subcarrier frequency. It has been noted to agree with color-bar test results except at the spectral tails where an algebraic correction is needed. However, the formulation requires the power levels of the luma and chroma signals as spectral parameters and further investigation is needed to establish reliable assignment guidelines.

Another approach is to always use a normalized gaussian spectrum, but assign its standard deviation such that it provides a spectral bound for the actual TV/FM spectrum based on experimental results. An example of this approach is illustrated in Figure 5-5, where the solid curves correspond to the spectral envelope tagged by the percentage of time not exceeded, and the dotted curve represents the gaussian bound with standard deviation (σ) equal to the video peak frequency deviation (D_V) divided by $\sqrt{2}$ [CCIR Study Groups, Documents 4/116-F and 9/105-F, 1984]. This approach has the advantage of providing a simple formulation for spectral generation purposes, without requiring specification of the audio subcarrier and energy dispersal parameters (which is not always available).

Another approach consists of having different spectral bound formulas whose relative predominance varies according to the modulation specifications. An example of this approach is illustrated in Figure 5-6, where one gaussian and three distinct rectangular characteristics represent the options available. The interest is to: (1) compare the gaussian to each rectangular function separately, (2) identify the largest function in each comparison, (3) use the one to three functions obtained to represent the TV/FM spectrum in interference algorithms, and (4) select the worst interference result after the corresponding one to three algorithm runs [CCIR Report 388-4, 1982].

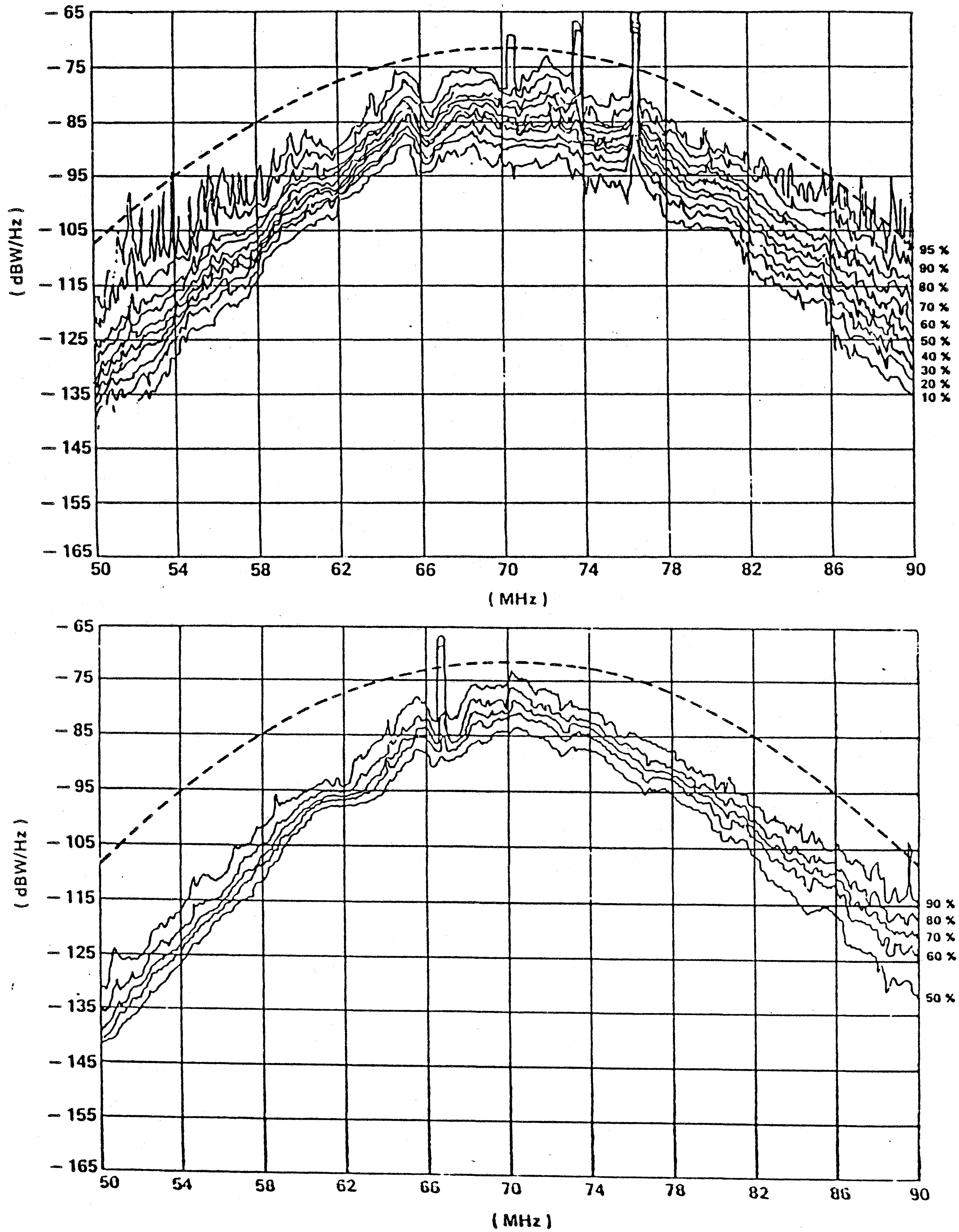
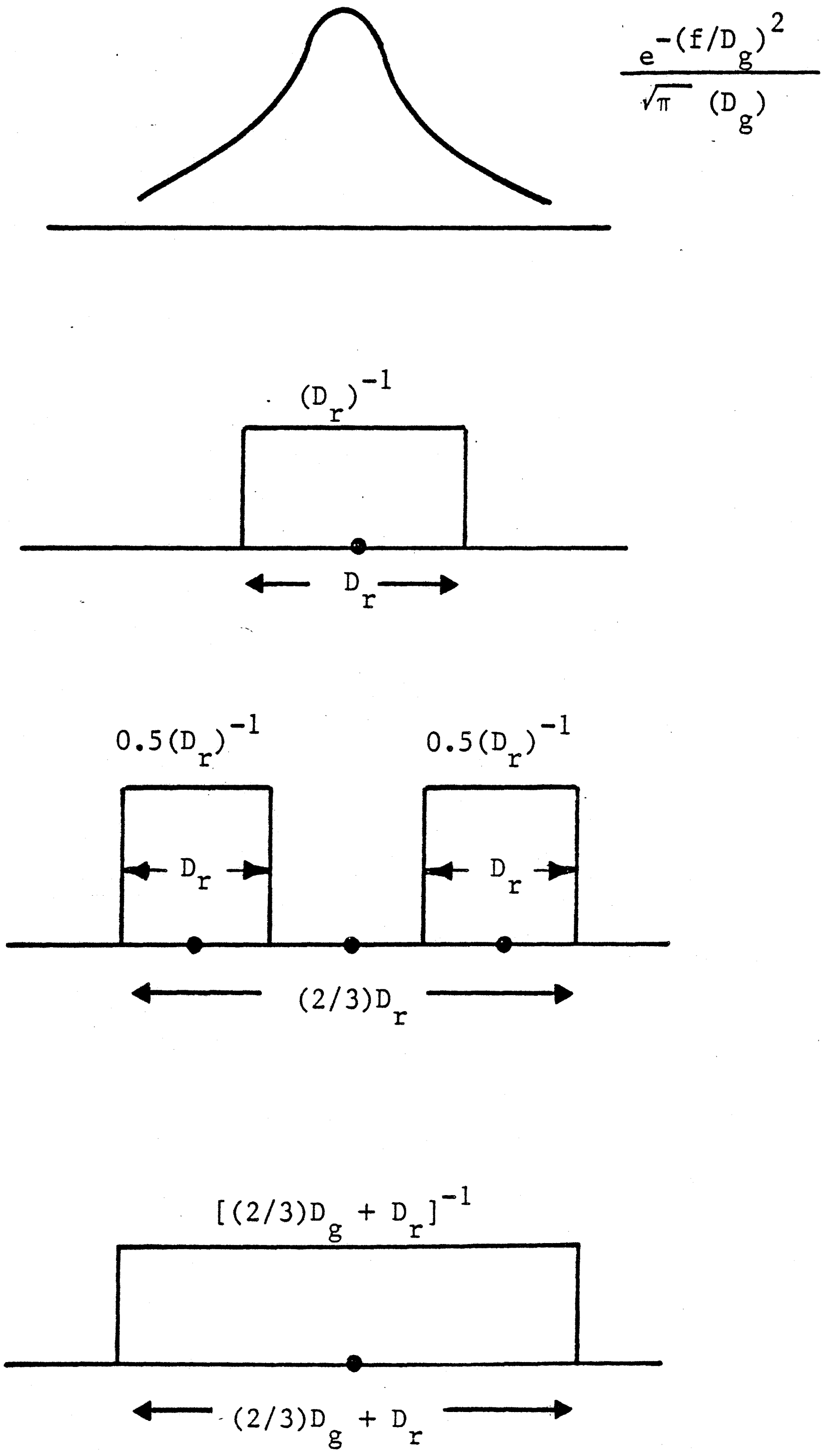


Figure 5-5. Gaussian spectral bound for TV/FM spectral measurements (top without energy dispersal, bottom with energy dispersal).



$D_g = D_V/(\sqrt{10})^{A/10}$, $D_r = D_{DP}$

A = 10 for 525-line, 11 for 625 line

Figure 5-6. CCIR TV/FM spectral models.

There are some inherent limitations in this last approach. One is the lengthy computation involving up to three algorithm runs per case. Another is the complex bookkeeping needed in automated programs, since the number of run repetitions per case varies with different modulation specifications. Another is that the largest function selection is ambiguous, since the gaussian may predominate at some frequencies and the rectangle at other frequencies in any of the comparisons. The use of a hybrid spectrum with gaussian and rectangular segments, based on their relative predominance, would bypass this ambiguity, but requires an area measurement to provide for the unit power normalization.

In summary, the gaussian spectral bound represents the best approach when a simple formulation is desired. The normalized gaussian spectrum given by Equation 5-2 can be used with $\sigma = D_V/\sqrt{2}$ where D_V is the peak frequency deviation of the video component. The dual gaussian spectral model provided by the luma-plus-chroma video decomposition is a promising alternative, but requires further investigation to provide reliable guidelines for the assignment of the parameters used in the formula. On this basis, the gaussian spectral bound is used for the compact algorithm formulations.

The hybrid spectral bound derived from the spectral characteristics shown in Figure 5-6 is used for the detailed algorithms. The TV/FM spectrum generation process (TVSPC) is illustrated in Figure 5-7 and consists of the selection of the gaussian or rectangular segments based on their spectral predominance. An area measurement is included to provide a scaling factor for the normalization of the hybrid spectrum.

The use of Carson's Rule for the TV/FM occupied bandwidth also requires attention. A conservative estimate results if the sum of the peak deviations of all baseband components (video, audios, dispersal) is used as the peak deviation in the rule. A more realistic estimate results if the root sum of squares (rss) of the peak deviations is used instead of their sum [CCIR Documents 9/1-E, 10-11S/3E, CMTT/5-E, 1983). In either case, the modulation frequency corresponds to the highest baseband frequency, including the modulated subcarrier bandwidths.

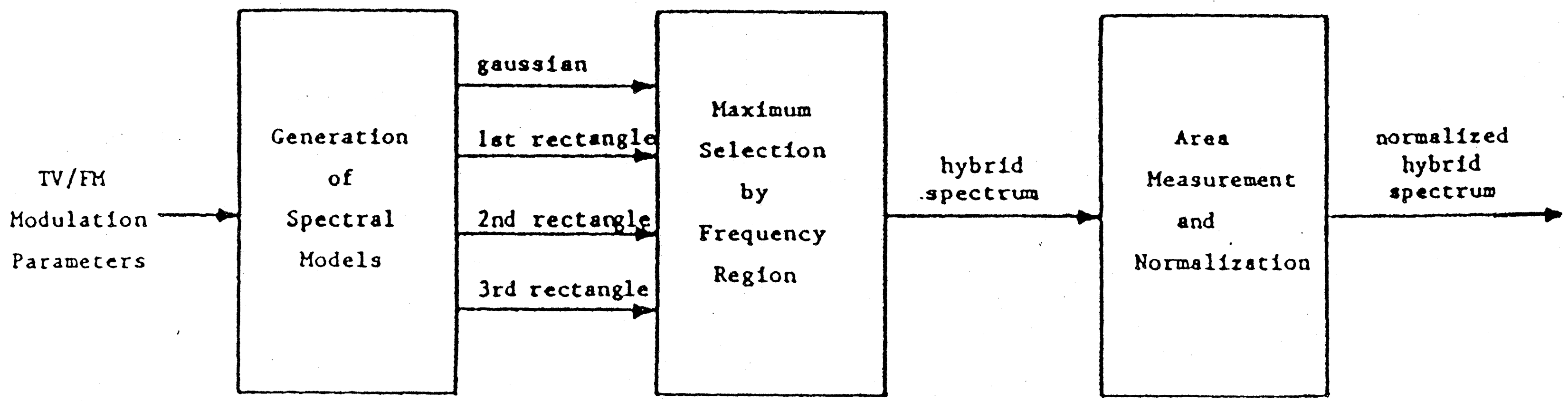


Figure 5-7. TVSPC generation program for TV/FM spectral bound.

Another approach to the TV/FM bandwidth estimate consists of using only the video component parameters (video deviation and video frequency) in Carson's Rule, but introducing a multiplier factor that expands the bandwidth thus obtained. A 1.1 multiplier has been suggested by the CCIR and has the advantage of standardizing the bandwidth evaluation for cases where the audio or dispersal modulation specifications are not available [CCIR Report 215-5, 1982].

These TV/FM bandwidth estimation approaches yield values compatible with existing modulation specifications under full-transponder operation, as shown in TABLE 5-1. The video deviation (D_V) approximates the rss deviation (D_{RSS}), and the 1.1 $(BW)_V$ bandwidth approximates the $(BW)_{RSS}$ bandwidth values. In particular, the 1.1 $(BW)_V$ rule always yields a value within the transponder bandwidth allocation. On this basis, this rule is used for full-transponder operation, since it bypasses the need for the audio and dispersal modulation specifications that are not always available.

However, the TV/FM bandwidth formula requires further revision under half-transponder operation to avoid excessive values, as shown in TABLE 5-1. The use of a 1.3 video overdeviation factor and time companding techniques has been noted to yield bandwidth compression without sacrificing broadcast quality for partial-transponder operation [Eng and Haskell, 1981]. This overdeviation yields values that match the half-transponder bandwidth allocation as shown in TABLE 5-1, and is used in the TV/FM bandwidth formula for half-transponder operation.

PSK SPECTRUM

The PSK spectrum model supports both wideband digital PSK and narrowband SCPC/PSK signals. The baseband is a digital data stream that keys the phase states of the transmission carrier. The phase states are equally spaced and the data symbols are assumed equiprobable, so there is no discrete carrier component in the signal spectrum.

The normalized spectrum, when all sidelobes are included, is given by $S(f) = T \operatorname{sinc}^2 (\pi Tf)$, where $\operatorname{sinc} X = (\sin X)/X$ and T is the digital symbol

TABLE 5-1

TV/FM BANDWIDTH ESTIMATES

Transponder Operation (#Signals)	Transponder Bandwidth (MHz)	Video Deviation (MHz)	RSS Deviation (MHz)	$(BW)_{RSS}$ Bandwidth (MHz)	$1.1(BW)_V$ Bandwidth (MHz)	$(BW)_{OVD}$ Bandwidth (MHz)
Full(1)	36.00	10.8	10.98	36.30	33.00	25.02
Full(1)	36.00	10.8	11.01	35.80	33.00	25.02
Full(1)	36.00	11.0	11.22	34.90	33.44	25.32
Full(1)	36.00	11.0	11.22	36.40	33.44	25.32
Full(1)	36.00	12.0	12.28	36.14	35.64	26.86
Half(2)	36.00	6.0	6.24	24.20	22.44	17.60
Half(2)	36.00	6.3	6.46	25.30	23.10	18.09

$$(BW)_{RSS} = 2\,(D_{RSS} + f_m)$$

$$(BW)_V = 2\,(D_V + f_v)$$

$$(BW)_{OVD} = 2\,\left(\frac{D_V}{1.3} + f_v\right)$$

where D_{RSS} = root sum of squares of video, audio-subcarriers and dispersal deviations

D_V = video deviation

f_m = highest modulation frequency, including modulated-subcarrier bandwidths

f_v = highest video frequency, including luma and chroma baseband components duration. If the number of phase states is M and the bit rate is B_r, then the symbol rate (1/T) is given by ($B_r/\log_2 M$). However, the emission spectrum is a truncated version of this characteristic, with the transmission passband usually contained within the mainlobe. A normalization factor (α) must then be included in the spectral formula to account for the band-limiting effect as shown in Equation 5-3, where B is the transmission bandwidth.

$$S(f) = \alpha T \operatorname{sinc}^2(\pi T f) \text{ , where } \alpha \geqq 1 \text{ and } |f| \leqq 0.5B \qquad (5\text{-}3)$$

The normalization factor (α) is a function of the transmission bandwidth specified. A plot of the power percentage ($1/\alpha$) preserved as a function of the bandwidth is shown in Figure 5-8, so that the normalization factor can be determined for a given bandwidth (B) and symbol rate (1/T) [Cohen, 1984)]. For example, if the passband exactly preserves the mainlobe of the spectrum, then $BT \approx 2$ and $\alpha = 1.1$ corresponds to 90% power preserved. The use of $BT \approx 1$ is often employed, with the bandwidth limits within the mainlobe, and $\alpha \approx 1.3$ corresponds to about 77% power preserved.

SCPC/FM SPECTRUM

The signal baseband consists of a single analog channel that is preemphasized and applied as frequency modulation on a transmission carrier. The statistical nature of the baseband modulation can vary according to the information content. For example, gaussian statistics can be assumed for FM music, whereas nongaussian statistics must be accepted for FM voice.

Some compact formulas have been proposed to model the nongaussian statistics of the single-channel voice baseband. The interest in such formulation is that the statistical probability density function (pdf) effectively represents the SCPC/FM spectrum under wideband FM conditions, assuming that the baseband statistics are not significantly altered by the preemphasis effect [Blackman and McAlpine, 1969].

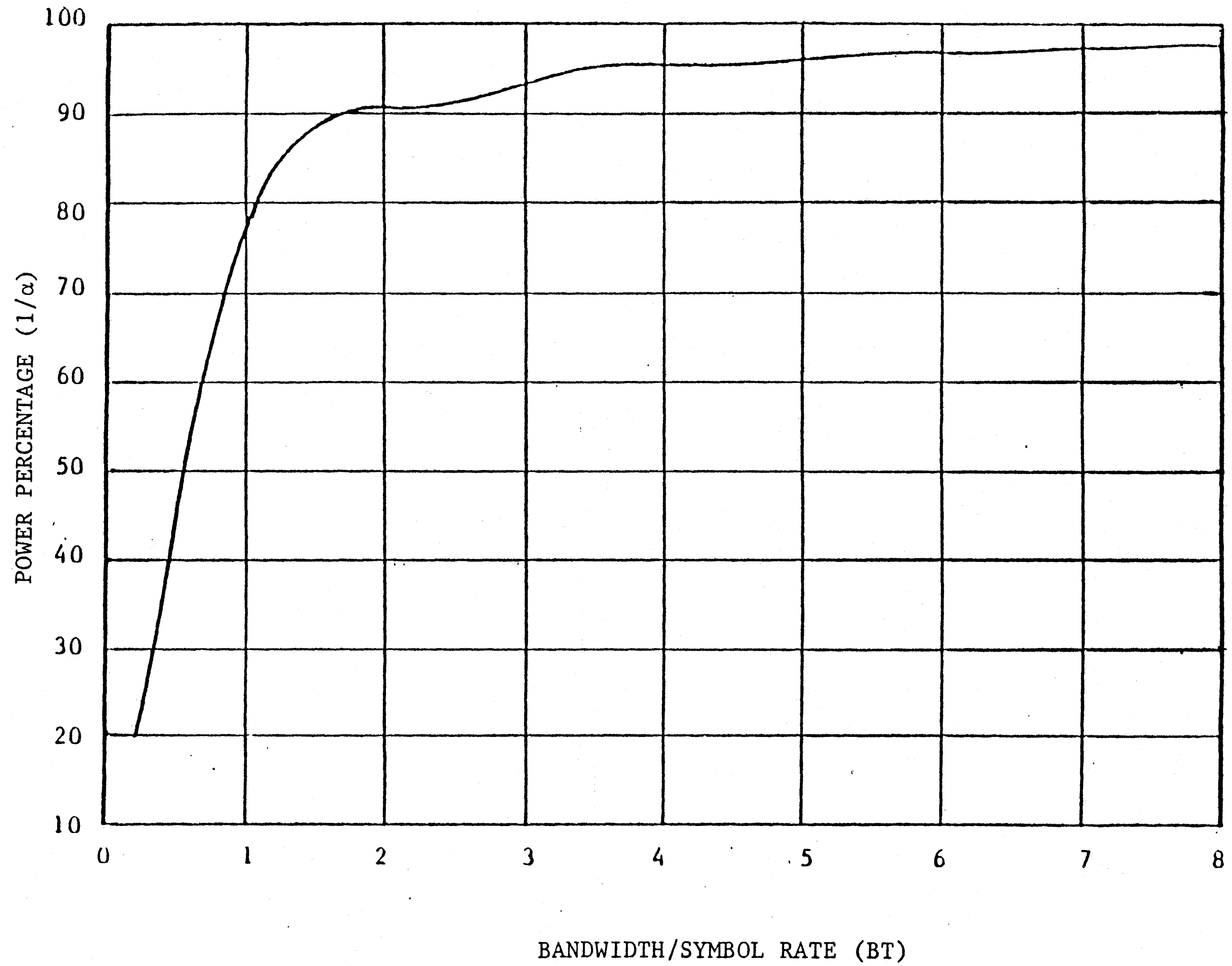


Figure 5-8. Power content in band-limited digital PSK modulation.

The three formulations shown in Equations 5-4 to 5-6 have been proposed to represent the pdf of single-channel voice. The exponential function is a simple one-parameter distribution, where (σ) is the rms frequency deviation. The modified gamma function is a two-parameter distribution (a,b) for improved fit potential, but it cannot be extended to the origin where it becomes infinite. The exponential plus gaussian function provides a four-parameter distribution based on the standard deviations (σ_1,σ_2) and weighting factors (k_1,k_2) of the two distributions involved.

Exponential:

$$p(\chi) = \frac{1}{\sqrt{2}\sigma} e^{-\sqrt{2}|\chi/\sigma|} \qquad (5\text{-}4)$$

Modified Gamma:

$$p(\chi) = \frac{b(b\chi)^{a-1}}{2\Gamma(a)} e^{-b/\chi}, \quad b = \sqrt{a(a+1)} \qquad (5\text{-}5)$$

Exponential plus Gaussian:

$$p(\chi) = \frac{k_1}{\sqrt{2}\sigma_1} e^{-\sqrt{2}|\chi/\sigma_1|} + \frac{k_2}{\sqrt{2\pi}\sigma_2} e^{-0.5(\chi/\sigma_2)^2} \qquad (5\text{-}6)$$

The exponential function can be used for spectral modeling by assigning the peak/rms frequency deviation factor ($\lambda = D_{PK}/D_{RMS}$) to match spectral measurements. The peak deviation (D_{PK}) is usually specified (either directly or via the peak modulation index) and must be converted into an rms frequency deviation (D_{RMS}) to permit the formula usage with $\sigma = D_{RMS}$. The approach is illustrated in Figure 5-9, where various λ values are assigned to the

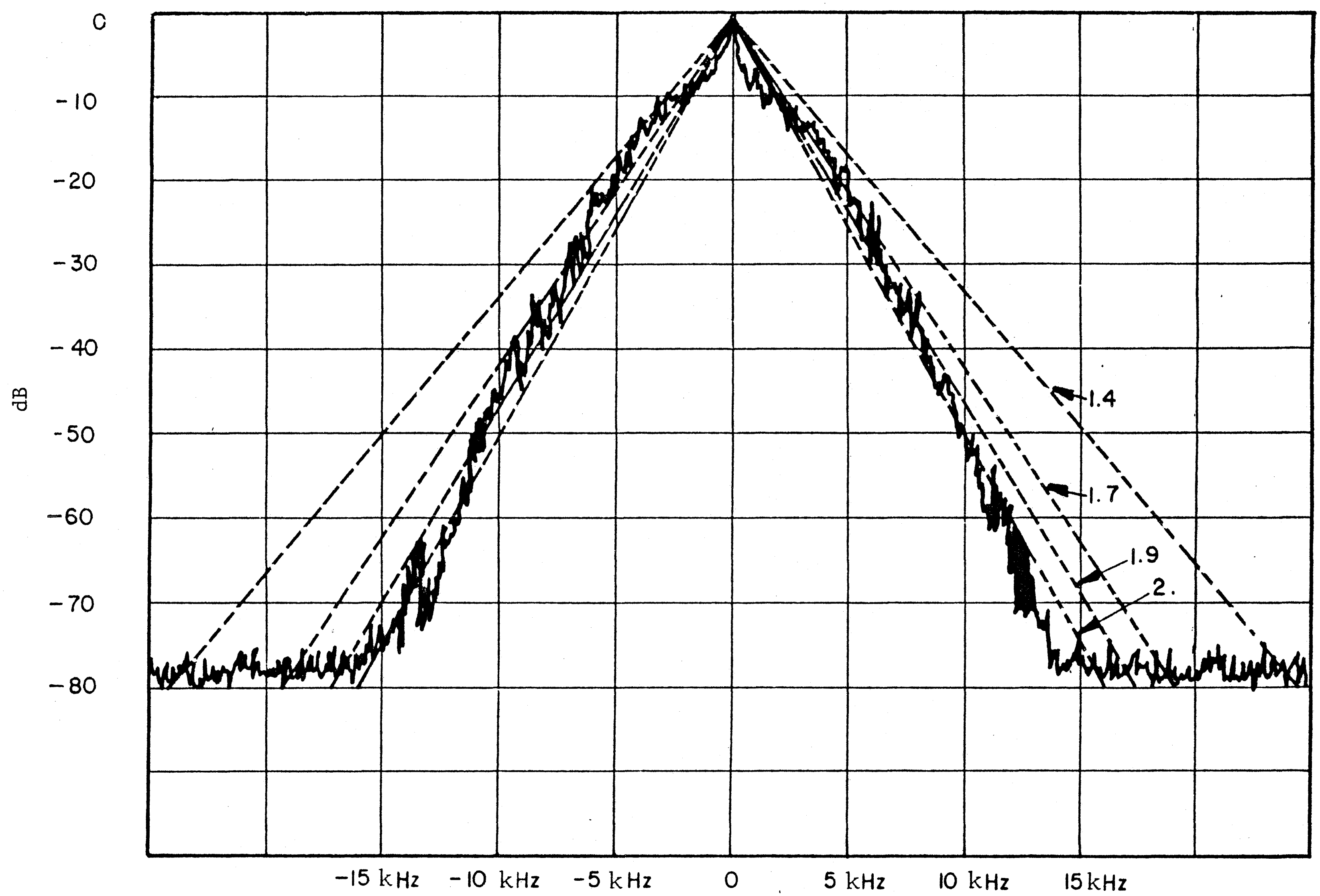


Figure 5-9. SCPC/FM voice spectrum and exponential modeling.

exponential characteristic (dotted), being superposed on a SCPC/FM spectrum measured with voice modulation [Haller and Van Deursen, 1983]. A value of $\lambda = 2$ is noted to provide a useful spectral bound when using the exponential model with $\sigma = D_{RMS} = (1/\lambda)\ D_{PK}$.

The modified gamma function approximates spectral measurements for $a < 1$, but is hindered by the discontinuity at the origin, which makes it impossible to perform a spectral normalization or convolution as is. The origin discontinuity can be avoided by assigning a lower frequency limit to the function, but such value must be well specified, since it determines the scaling factor needed for normalization purposes. There have been no guidelines hitherto available towards such specification, and the modified gamma function will not be used, since it cannot be normalized otherwise.

The exponential-plus-gaussian function is also hindered by the lack of guidelines to select its four parameters $(k_1, k_2, \sigma_1, \sigma_2)$. The k parameters control the relative weighting $(k_1 + k_2 = 1)$ of the two distributions involved, while the σ parameters control their significant spectral regions. The k_1/k_2 and σ_1/σ_2 relations are needed to determine the four parameters from the rms frequency deviation $(D_{RMS} = k_1\sigma_1 + k_2\sigma_2)$, since the latter is the only parameter that can be logically derived from the peak frequency deviation.

In summary, the SCPC/FM spectrum is modeled by a simple formula under wideband FM conditions provided the baseband modulation statistics are formulated. The case of music modulation can be assumed to have gaussian statistics and the spectrum takes the gaussian form of Equation 5-2. The case of voice modulation can be assumed to have exponential statistics and the spectrum takes the form of Equation 5-7, where $\lambda = 2$ is used when not specified.

$$S(f) = \frac{1}{\sqrt{2}\sigma} e^{-\sqrt{2}\,|f/\sigma|} \text{ , where } \sigma = (1/\lambda)\ D_{PK} \qquad (5\text{-}7)$$

APPENDIX A

RTC ALGORITHMS FOR CSSB/AM TELEPHONY

CSSB/AM TELEPHONY PERFORMANCE FORMULATION

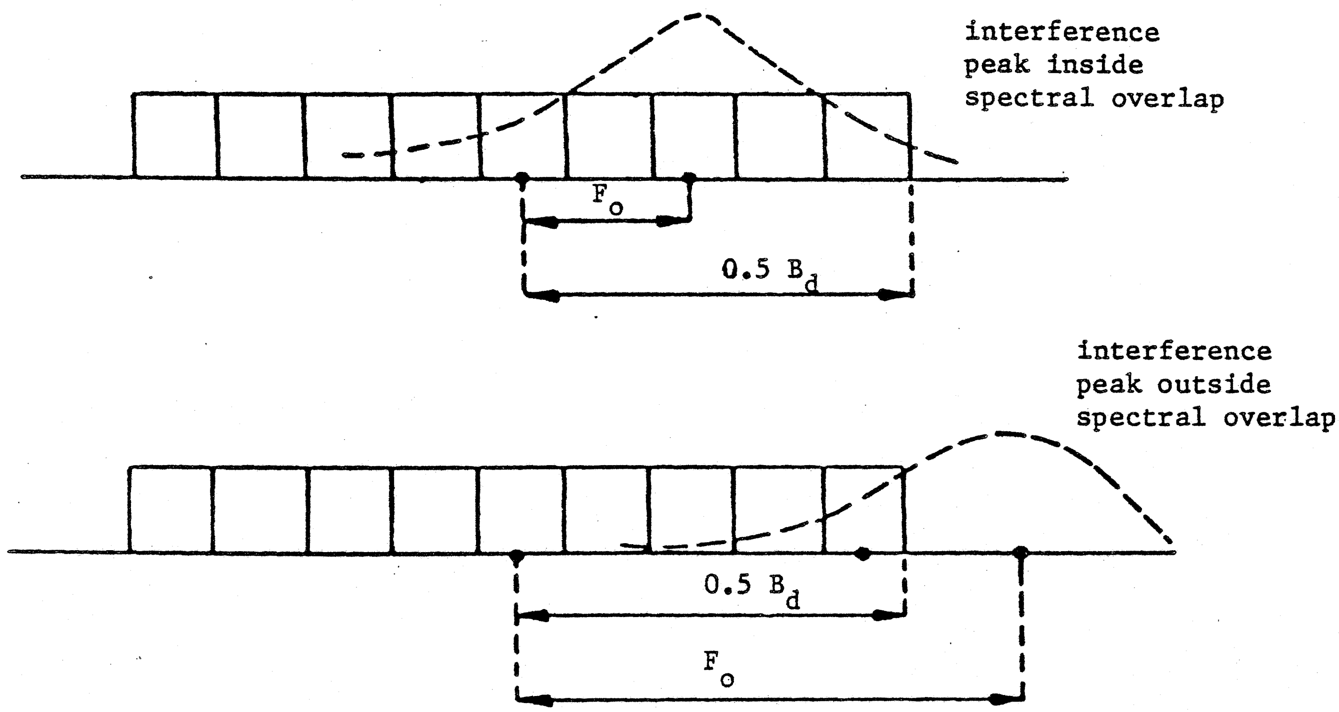
The CSSB/AM receiver extracts the uniform multichannel baseband spectrum from the desired input signal via linear frequency translation. An interfering signal spectrum contained within the receiver input passband is also linearly translated along with the desired input spectrum and produces an output interferer component added to the desired output baseband. The relative levels and spectral characteristics of the desired and interferer spectra are preserved through the demodulation process.

The desired output power (P_d) in a slot measurement bandwidth is the product of the uniform baseband spectral density (C/B_b) times the measurement bandwidth (B_m), where $B_b = f_H - f_L$ is the baseband bandwidth and $B_m = 3.1$ kHz is usually employed. The interferer output power (P_i) is obtained by integrating the interferer output spectrum $I \cdot K_i(f)$ over the measurement bandwidth.

The desired-to-interferer output power ratio (P_d/P_i) is called the noise power ratio (NPR), due to the uniform power distribution assumed in the multichannel baseband spectrum. The NPR performance varies over the different baseband channels when the interferer spectrum $K_i(f)$ is not uniform. The worst channel being affected corresponds to the location of the peak of the interference spectrum $K_i(f)$ within the desired baseband.

The effective interference spectrum $K_i(f)$ can differ from the interference emission spectrum $S_i(f)$, due to the frequency offset and spectral truncation conditions. The two possible situations are shown in Figure A-1: the peak of $S_i(f)$ can occur either inside or outside the spectral overlap region. The peak value of $K_i(f)$ equals that of $S_i(f)$ for the inside condition, but must be evaluated from the spectral characteristic as a function of the frequency offset for the outside condition.

A 0-dBm0 test-tone modulation is conventionally used as a measurement standard, instead of the multichannel baseband modulation. The output power



$$K_i = \frac{1}{\sqrt{2\pi}(\sigma)} e^{-0.5(f_p/\sigma)^2}$$ for FDM/FM or TV/FM interferer (gaussian spectrum model)

$$K_i = (\alpha T)\ \mathrm{sinc}^2\ (\pi T f_p)$$ for DIG/PSK interferer

where

$f_p = 0$ if $F_o \leq 0.5\ (B_d - B_c)$

$f_p = F_o - 0.5\ (B_d - B_c)$ if $F_o > 0.5\ (B_d - B_c)$

Figure A-1. Wideband interference into CSSB/AM telephony.

ratio becomes a desired-tone versus interference-noise ratio (TNR), and the performance measure employed is the output interference power in picowatts (pWOp).

The generalized NPR, TNR, and pWOp formulations are shown in TABLE A-1, along with the C/I threshold and margin corresponding to a 600 pWOp output performance requirement [CCIR Recommendation 466-3, 1982] The bandwidth ratio (BWR) accounts for the distinction between the baseband bandwidth (B_b) and the measurement bandwidth (B_m). The noise loading ratio (NLR) accounts for the distinction between the multichannel speech level (P-dBm0) and the test-tone reference level (0-dBm0). The noise weighting factor (NWF) accounts for the psophometric noise weighting effect (2.5 dB).

The NLR term includes the speech level (T_s) and companding gain (G) effects. The use of companding in CSSB/AM signals has been noted to produce a change in speech level from T_s to $T_s + X_s$ in dB, as well as to introduce a subjective noise improvement of A dB, which results in a net companding gain of $G = A - X_s$ in dB. The values of A = 16 dB, along with X_s = 4 dB for T_s = -15 dB and X_s = 7 dB for T_s = -21 dB, have been cited for design purposes [Jonnalagadda, 1982]. There are no CCIR guidelines available on this matter, so that A = 16 dB is used if not specified, along with the CCIR guidelines for speech level assignment in FDM/FM telephony [CCIR Report 708-1, 1982].

COMPACT AND DETAILED ALGORITHM COMPUTATIONS

The general formulation of TABLE A-1 is to be used for all interferer types, with each type distinguished by its distinct interference spectrum $K_i(f)$ being used in the P_i and NPR evaluation. The basic computation steps performed by a dedicated RTC algorithm are:

1. generation of the interferer input spectrum $K_i(f)$ for given modulation specifications

2. evaluation of the NPR performance for the worst channel affected, which requires evaluating the $K_i(f)$ spectrum if $B_i >> B_c$, or

TABLE A-1

CCSB/AM TELEPHONY PERFORMANCE FORMULATION

$$B_b = (f_H - f_L) \qquad \text{(baseband bandwidth)}$$

$$P_d = (C/B_b) \cdot (B_m) \qquad \text{(desired output power)}$$

$$P_i = (I) \cdot \int_{B_m} K_i(f)\,df \qquad \text{(interferer output power)}$$

$$= (I) \cdot (K_i B_m) \quad \text{if } B_i >> B_m$$

$$(NPR)_{dB} = 10 \log (P_d/P_i) = (C/I)_{dB} - 10 \log \left[(B_b/B_m) \cdot \int_{B_m} K_i(f)df\right]$$

$$= (C/I)_{dB} - 10 \log (K_i B_b) \quad \text{if } B_i >> B_m$$

$$(TNR)_{dB} = (NPR)_{dB} + (BWR)_{dB} - (NLR)_{dB}$$

$$= (NPR)_{dB} + 10 \log (B_b/B_m) - (P)_{dB}$$

$$pWOp = 10^{[90-(TNR)_{dB} - (NWF)_{dB}]/10} = 10^{[87.5 - (TNR)_{dB}]} \quad \text{(picowatts)}$$

$$(C/I)_{dB(margin)} = 10 \log \frac{600}{pWOp} = 10 \log 600 - [87.5 - (TNR)_{dB}]$$

$$(C/I)_{dB\ (threshold)} = (C/I)_{dB} - (C/I)_{dB(margin)}$$

<u>Notes</u>: B_m = 3.1 kHz is assumed

$P = T_s + U \log N_c - A$ is assumed as follows unless specified otherwise

$T_s = -15.0$ and $U = 10$ if $N_c \geqq 240$ without companding ($A = 0$)

$T_s = -11.2$ and $U = 10$ if $N_c \geqq 240$ with companding ($A = 16$)

$T_s = -1.0$ and $U = 4$ if $12 \leqq N_c < 240$ without companding ($A = 0$)

$T_s = +2.8$ and $U = 4$ if $12 \leqq N_c < 240$ with companding ($A = 16$)

600 pWOp performance threshold assumed

integrating the $K_i(f)$ spectrum otherwise, as well as identification of the worst channel affected

3. evaluation of the pWOp performance and C/I threshold margin from the worst NPR value using simple conversion formulas.

The distinction between the compact versus detailed algorithm versions is based on how simple it is to generate and integrate the $K_i(f)$ spectrum, as well as to identify the worst channel affected. The compact versions are characterized by simple formulas or subroutines for the spectrum generation and/or integration, plus simple procedures for the worst channel identification. The detailed versions can involve elaborate spectral generation and/or integration routines, plus recycled sampling of all channels for the worst channel identification.

Case of CSSB/AM Telephony Interferer: Compact and Detailed Algorithm

A CSSB/AM multichannel telephony interferer satisfies the wideband condition $B_i >> B_m$. Also, the interference spectrum is uniform with $K_i(f) = 1/B_i$, so that all desired channels within the spectral overlap region are equally affected. The NPR formula reduces to the simple form NPR = $(C/I) \cdot (B_i/B_b)$ for any and all channels being interfered with.

Case of FDM/FM Telephony Interferer: Compact Algorithm

An FDM/FM multichannel telephony interferer satisfies the wideband condition $B_i >> B_m$, but the interference spectrum is not uniform and the NPR values vary over the channels being interfered with. The FDM/FM emission spectrum $S_i(f)$ uses the gaussian formula given in Equation 5-2 for the compact algorithm.

The effective interference spectrum $K_i(f)$ is obtained according to the expressions in Figure A-1. If the gaussian peak is inside the spectral overlap region, then $K_i = 1/(\sqrt{2\pi}\ \sigma)$ is used in the NPR formula of TABLE A-1 for the worst channel affected. If the gaussian peak is outside the spectral overlap region, this value is reduced by the gaussian characteristic as indicated in Figure A-1.

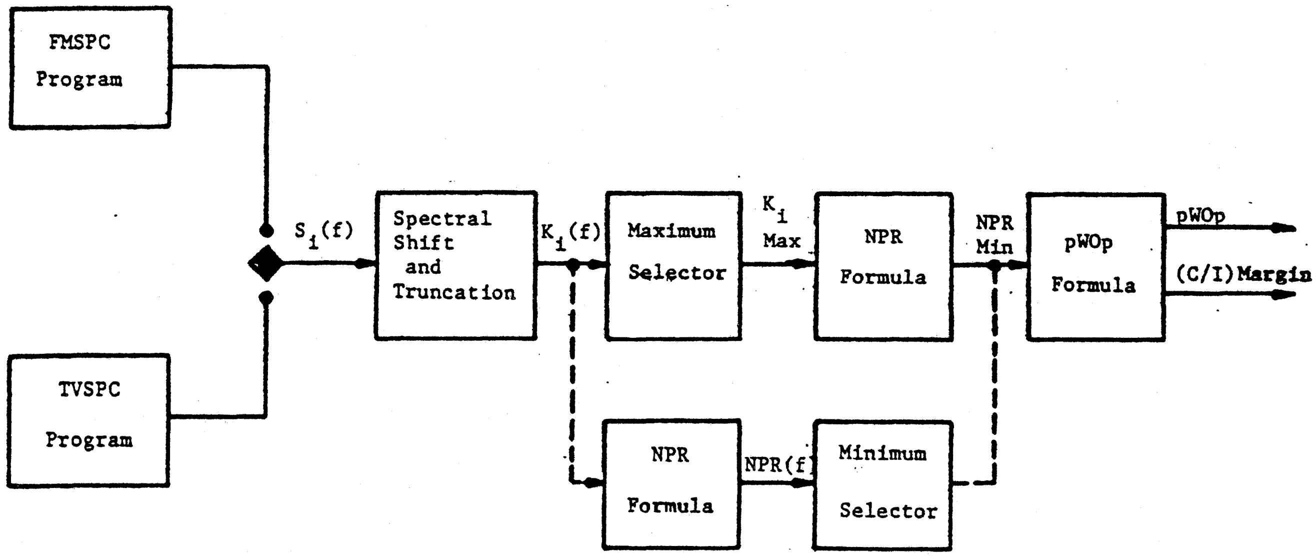
Case of FDM/FM Telephony Interferer: Detailed Algorithm

The wideband interferer condition $B_i \gg B_m$ is still applicable, but now a simple formula is not available for the interference spectrum. The FMSPC Program shown in Figure 5-3 (Section 5) is employed to generate the FDM/FM emission spectrum $S_i(f)$. The effective interference spectrum $K_i(f)$ is obtained by appropriately shifting and truncating $S_i(f)$, and the result must be examined to select its peak value within the spectral overlap region. This peak value is used for K_i in TABLE A-1 to represent the worst channel affected.

The processing involved in this detailed algorithm is summarized in Figure A-2. The two alternatives illustrated are essentially similar in processing complexity. One approach examines the $K_i(f)$ spectrum samples to select the maximum value and performs one NPR computation representing the worst channel. The other approach examines the NPR(f) samples at all channel frequencies and selects the minimum value representing the worst channel. This last approach is selected, since it has the advantage of providing all the channel NPRs to the user for comparison purposes.

Case of TV/FM Interferer: Compact Algorithm

A TV/FM interferer satisfies the wideband condition $B_i \gg B_m$ with a non-uniform spectrum. The TV/FM emission spectrum $S_i(f)$ uses the gaussian envelope (see Section 5) for the compact algorithm. The effective interference spectrum $K_i(f)$ for the worst channel affected is obtained according to the expressions in Figure A-1, which distinguish whether the gaussian peak lies inside or outside the spectral overlap region.



Key:
Dotted line shows processing alternative.

Figure A-2. Detailed algorithm for FDM/FM or TV/FM interference into CSSB/AM telephony.

Case of TV/FM Interferer: Detailed Algorithm

The wideband interferer condition $B_i >> B_m$ is still applicable, and the interferer emission spectrum $S_i(f)$ is generated by the TVSPC Program shown in Figure 5-7 (Section 5). The effective interference spectrum $K_i(f)$ is obtained by the proper shifting and truncating of $S_i(f)$, and its peak value K_i within the spectral overlap region is employed in TABLE A-2 for the worst channel affected. The two processing alternatives shown in Figure A-2 are possible, and the top one is selected for the reasons already discussed.

Case of Wideband Digital PSK Interferer: Compact and Detailed Algorithms

A wideband digital PSK interferer satisfies the $B_i >> B_m$ condition, with a sinc-squared emission spectrum that is band-limited to a ±0.5 B_i emission passband. The effective interference spectrum $K_i(f)$ is generated according to the formula shown in Figure A-1. If the mainlobe peak is inside the spectral overlap region, then $K_i = (\alpha T)$ is used in TABLE A-1. Otherwise, this value is reduced by the sinc-squared characteristic as indicated in Figure A-1.

Case of SCPC/PSK or SCPC/FM Interferers: Compact and Detailed Algorithms

A SCPC/PSK (digital data) or SCPC/FM (analog voice) interferer is narrowband compared to the previous interference types discussed. However, these SCPC interferers are usually wideband relative to the slot measurement bandwidth (B_m). The formulation that follows handles the SCPC interferers, regardless of whether they are narrow or wide relative to the measurement bandwidth.

The parameters $N_i = B_m/B_i$ and $N_i^{'} = B_m/B_i^{'}$ are first computed to determine how many SCPC interferers fit within the measurement bandwidth. If $N_i^{'} < 1$, then one truncated interferer fits if $N_i < 1$ or one complete interferer fits if $N_i \geqq 1$. If $N_i^{'} = 1$, then one complete interferer fits regardless of the N_i value. If $N_i^{'} > 1$, then more than one interferer fits, and an activity factor (k) is used to determine their effective number as either one if $N_i^{'} \leqq 1/k$ or many ($kN_i^{'}$) if $N_i^{'} > 1/k$. The value k = 0.4 is assumed for the activity factor.

If there is one truncated SCPC interferer within the desired channel, its effective interference spectrum $K_i(f)$ must be integrated over the measurement bandwidth. The interferer spectrum is assumed to be centered relative to the measurement bandwidth, so that the spectral integration limits are $\pm 0.5\ B_m$. A simple spectrum formula is used for both SCPC interferer types, as discussed in Section 5. A band-limited sinc-squared formula is used for SCPC/PSK, and an exponential formula is used for SCPC/FM.

However, only the SCPC/FM case has a simple integration, as shown in TABLE A-2. The SCPC/PSK case requires an integral subroutine for an exact evaluation, but an upper bound can be obtained by using a rectangular envelope instead of the sinc-squared function. The rectangular bound is used for the compact algorithm, and the integral subroutine for the detailed algorithm, in the SCPC/PSK case. The compact and detailed algorithms are identical in the SCPC/FM case.

If there is one complete SCPC interferer within the desired channel, then its spectral integration is unity and $P_i = I$ as shown in TABLE A-2. If there are many SCPC interferers (kN_i') within a channel bandwidth, then the interference power per channel is $P_i = (kN_i')I$ as shown in TABLE A-2. The compact and detailed algorithms are identical under these conditions for both SCPC/PSK and SCPC/FM cases.

TABLE A-2

CASE OF SCPC/PSK OR SCPC/FM INTERFERERS

Case of $N' \leqq 1$ and $N < 1$: one truncated interferer

$$\int_{B_m} K_i(f)\, df = \begin{cases} \int_{B_m} \alpha T\, (\mathrm{sinc}\, \pi f T)^2 df \leqq (\alpha T) B_m & \text{for SCPC/PSK interferer} \\ 1-\exp[(-\sqrt{2}/\sigma)(0.5\, B_m)] & \text{for SCPC/FM interferer} \end{cases}$$

Case of $N' < 1$ and $N \geqq 1$: one complete interferer

$$\int_{B_m} K_i(f)\, df = 1$$

Case of $N' = 1$: one complete interferer

$$\int_{B_m} K_i(f)\, df = 1$$

Case of $N' > 1$: possibly many interferers

$$\int_{B_m} K_i(f)\, df = 1 \quad \text{if} \quad N' \leqq 1/k \quad \text{(many interferers)}$$

$$\int_{B_m} K_i(f)\, df = kN' \quad \text{if} \quad N' > 1/k \quad \text{(many interferers)}$$

Note: $k = 0.4$ will be assumed

APPENDIX B

RTC ALGORITHMS FOR FDM/FM TELEPHONY

FDM/FM TELEPHONY PERFORMANCE FORMULATION

The FDM/FM receiver extracts the uniform multichannel baseband spectrum from the desired input signal via frequency demodulation and deemphasis. The total output power in the desired multichannel baseband is given by $(2\pi D_{RMS})^2$, where D_{RMS} is the rms multichannel frequency deviation. The desired output power (P_d) in a slot measurement bandwidth (B_m) is this total power reduced by the ratio (B_m/B_b), where $B_b = f_H - f_L$ is the baseband bandwidth and B_m = 3.1 kHz is usually employed.

An interfering signal within the receiver input passband produces an interferer output component added to the desired output baseband. The interferer output spectrum $K_o(f)$ differs from its effective input spectrum $K_i(f)$ due to the FM demodulation process. The interferer output power (P_i) in the measurement bandwidth is obtained by integrating the spectrum $K_o(f)$ over such bandwidth.

The desired-to-interferer power ratio (P_d/P_i) is called the noise power ratio (NPR), due to the uniform power distribution in the multichannel baseband spectrum. The NPR performance varies over the different baseband channels via the interferer output spectrum $K_o(f)$, which must be characterized to evaluate the NPR performance and identify the worst channel affected. The spectral characterization and performance evaluation next described are consistent with the CCIR formulation on this matter [CCIR Report 388-4, 1982].

The interferer output spectrum $K_o(f)$ has the form $(2\pi f)^2$ D(f) R(f), where D(f) is the deemphasis power transfer function and R(f) is the resultant of a spectral convolution series. The nth term in the series is proportional to $(C/I)^{-n}$, and requires the nth order convolution of the desired input spectrum $S_d(f)$ with the effective interference input spectrum $K_i(f)$. Under high C/I conditions, the series can be approximated by its predominant term, and only one spectral convolution is required to characterize the interferer output spectrum $K_o(f)$.

• The series reduces to $R(f) \approx (C/I)^{-1} \cdot Q(f)$ under high C/I conditions, where $Q(f) = 0.5[K(f) + K(-f)]$ and $K(f) = S_d(f) * K_i(f)$ is the spectral convolution. Under cochannel conditions, $K(f) = K(-f)$ is symmetric and $Q(f) = K(f)$ is obtained. Under offset conditions, $K(f) \neq K(-f)$ is asymmetric but $Q(f)$ always remains symmetric, as illustrated in Figure B-1 using two rectangular input spectra as an example. Hence, the interferer output spectrum $K_o(f)$ will always be symmetric, with a single formulation handling both cochannel and offset conditions.

The worst channel affected in the desired output baseband corresponds to the peak of the interferer output spectrum $K_o(f)$ within the desired baseband, since such peak maximizes the interferer output power (P_i). The worst channel often occurs near the higher frequency limit (f_H) due to the $(2\pi f)^2$ factor contribution. However, the deemphasis D(f) and convolution K(f) factors can alter this behavior in certain cases.

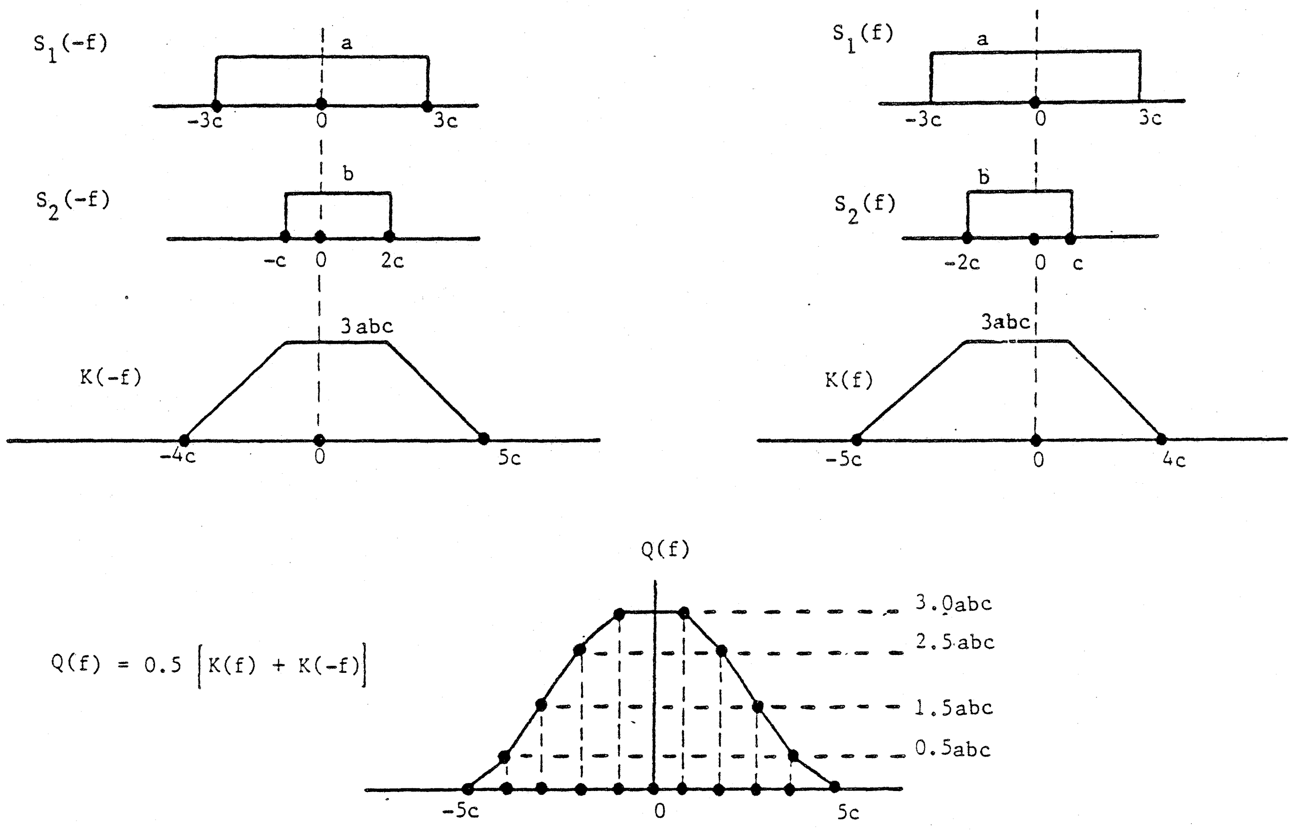


Figure B-1. Example of spectral convolution K(f) and superposition Q(f) under offset conditions.

The generalized NPR, TNR, and pWOp formulations for FDM/FM telephony are shown in TABLE B-1, along with the C/I threshold and margin corresponding to a 600 pWOp output performance requirement [CCIR Recommendation 466-3, 1982]. The notation and development is similar to that already presented for CSSB/AM telephony (see TABLE A-1). The main distinction lies in the different expressions for the desired and interferer output powers (P_d, P_i), which introduce distinct modulation parameters and functional relations into the NPR, TNR, and pWOp computations.

There are CCIR guidelines for the speech level and companding gain assignment in FDM/FM telephony [CCIR Report 708-1, 1982]. The RTC algorithms use the CCIR formulas when the speech level is not specified, along with a companding improvement A = 16 dB. The CCIR formulas assign the loading factor (P) as a function of the number of channels (N_c) and the companding flag (yes/no), as shown in TABLE B-1.

COMPACT AND DETAILED ALGORITHM COMPUTATIONS

The general formulas of TABLE B-1 are applicable to all interferer types, with each type distinguished by its effective input spectrum $K_i(f)$ producing a distinct convolution spectrum K(f) and output spectrum $K_o(f)$. The basic computation steps performed by a dedicated RTC algorithm are:

1. generation of the desired $S_d(f)$ and interferer $K_i(f)$ input spectra (if needed) from the modulation specifications

2. generation of the convolution K(f) and superposition Q(f) spectra from the desired and interferer input spectra

3. evaluation of the NPR performance for the worst channel affected, which requires characterization of the interferer output spectrum $K_o(f)$, as well as identification of the worst channel affected

4. evaluation of the pWOp performance and the C/I threshold margin from the worst NPR value using simple conversion formulas.

TABLE B-1

FDM/FM TELEPHONY PERFORMANCE FORMULATION

$B_b = (f_H - f_L)$ (baseband bandwidth)

$P_d = (2\pi D_{RMS})^2 \cdot (B_m/B_b)$ (desired output power)

$P_i = \int_{B_m} K_o(f)df \doteq K_o B_m$ (interferer output power)

where $K_o(f) = (2\pi f)^2 D(f)R(f)$ (interferer output spectrum)

$D(f) = 1/P(f)$ (see Figure 5-1 in Section 5)

$R(f) = (C/I)^{-1} \cdot Q(f)$ if $C/I >> 1$

$Q(f) = 0.5\,[K(f) + K(-f)]$

$K(f) = S_d(f) * K_i(f)$ (* denotes convolution)

$$(NPR)_{dB} = 10 \log (P_d/P_i)$$

$$= (C/I)_{dB} + 20 \log D_{RMS} - 10 \log [f^2 D(f)Q(f) \cdot B_b] \quad \text{if } C/I >> 1$$

$$(TNR)_{dB} = (NPR)_{dB} + (BWR)_{dB} - (NLR)_{dB}$$

$$= (NPR)_{dB} + 10 \log (B_b/B_m) - (P)_{dB}$$

$$pWOp = 10^{[90-(TNR)_{dB} - (NWF)_{dB}]/10} = 10^{[87.5 - (TNR)_{dB}]} \quad \text{(picowatts)}$$

$$(C/I)_{dB(margin)} = 10 \log \frac{600}{pWOp} = 10 \log 600 - [87.5 - (TNR)_{dB}]$$

$$(C/I)_{dB\ (threshold)} = (C/I)_{dB} - (C/I)_{dB(margin)}$$

Notes: B_m = 3.1 kHz is assumed

$P = T_s + U \log N_c$ is assumed as follows unless specified otherwise

T_s = -15.0 and U = 10 if $N_c \geq 240$ without companding (A = 0)

T_s = -11.2 and U = 10 if $N_c \geq 240$ with companding (A = 16)

T_s = -1.0 and U = 4 if $12 \leq N_c < 240$ without companding (A = 0)

T_s = +2.8 and U = 4 if $12 \leq N_c < 240$ with companding (A = 16)

600 pWOp performance threshold assumed

The distinction between the compact versus detailed algorithm versions is based on how simple it is to generate the spectral convolution K(f), (including any needed generation of the desired and interferer input spectra), and to identify the worst channel affected. The compact versions are characterized by simple formulas or subroutines for the spectral convolution outcome, as well as simple procedures for the worst channel identification. The detailed versions can involve elaborate spectral convolution and/or generation routines, plus recycled sampling for the worst channel identification.

NTIA has designed and implemented a detailed Spectrum Convolution Program (SPCVL) that is employed whenever the spectral convolution K(f) does not have a simple formula or subroutine. The SPCVL Program accepts the desired $S_d(f)$ and effective interferer $K_i(f)$ spectral samples as inputs to automatically perform the spectral convolution K(f) and generate the superposition function Q(f) needed for the NPR computation.

The SPCVL Program is summarized in Figure B-2. The realization exploits the fact that the spectral convolution in the frequency domain corresponds to the correlation function multiplication in the time domain. This permits the use of simple operations (multiplication, interpolation) along with Fast Fourier Transforms (FFTs) to avoid the many shift-multiply-integrate cycles otherwise needed in a frequency domain realization. The bandwidth expansion of the convolution process is accounted by properly controlling the sample size before and after the correlation function multiplication in the time domain.

The NPR computation for all the baseband channels to choose the worst one can result in a long run, even if the spectral convolution and NPR computation are simple formulas. The compact algorithms avoid long runs by restricting the number of channels being sampled to a certain number of equally spaced channels that can be selected by the user and always includes the corner channels. The top channel is automatically chosen when only one baseband channel is to be sampled. The detailed algorithms always sample all channels to identify the worst NPR condition.

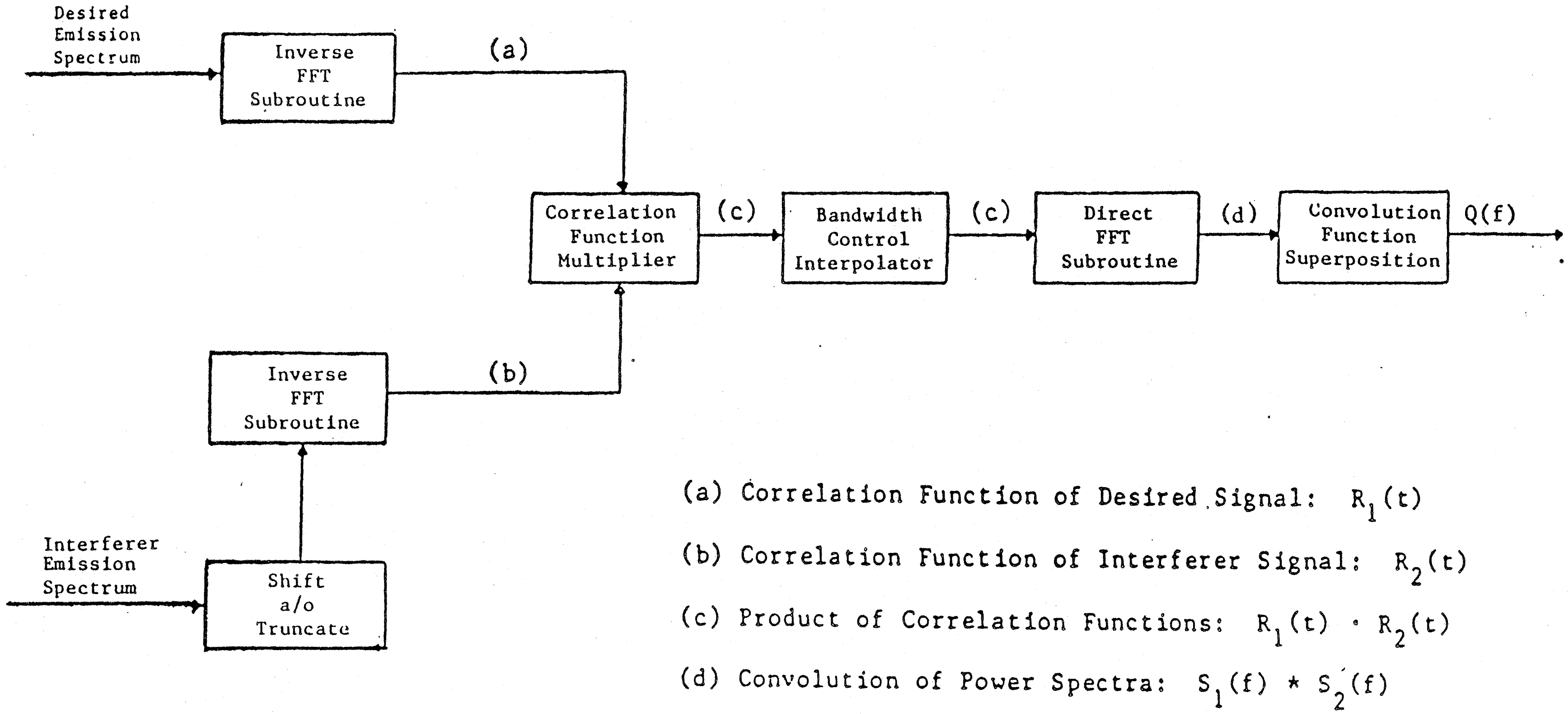


Figure B-2. Block diagram of the SPCVL spectral convolution program.

Case of CSSB/AM Telephony Interferer: Compact Algorithm

The FDM/FM desired spectrum $S_d(f)$ has the gaussian formula given by Equation 5-2 in Section 5, and the CSSB/AM interferer spectrum is rectangular with $K_i(f) = 1/B_i$ over the spectral overlap region. The convolution of gaussian and rectangular spectra requires the gaussian integral evaluation over limits that vary as the convolution function develops. A simple subroutine is available for the standardized gaussian integral and is used to generate the samples Q of the spectrum Q(f), as summarized in TABLE B-2. A baseband channel frequency (f) is used to compute a Q value to be used in TABLE B-1 to compute the NPR for that channel. It should be noted that the entire Q(f) spectrum need not be generated, since only its values at those channels being sampled are needed for the compact algorithm.

Case of CSSB/AM Telephony Interferer: Detailed Algorithm

The FMSPC Program shown in Figure 5-3 (Section 5) is used to generate the FDM/FM desired spectrum. The CSSB/AM interferer spectrum is rectangular with $K_i(f) = 1/B_i$. The SPCVL program shown in Figure B-2 is used to generate the spectrum Q(f) needed for the NPR computation in TABLE B-1. All desired baseband channels are sampled to determine the worst NPR performance.

Case of FDM/FM Telephony Interferer: Compact Algorithm

The FDM/FM desired $S_d(f)$ and interferer $S_i(f)$ emission spectra are both gaussian, and the effective interference spectrum $K_i(f)$ is not truncated within the desired receiver passband. The spectral convolution K(f) has a simple gaussian formula and the NPR evaluation in TABLE B-1 reduces to the simple formula shown in TABLE B-3. This expression is consistent with the CCIR formulation corresponding to the gaussian-gaussian case [CCIR Report 388-4, 1982].

TABLE B-2

RECTANGULAR-GAUSSIAN SPECTRAL CONVOLUTION

DEFINITIONS

$a = F_o - 0.5\ B_i\ ,\ b = F_o + 0.5\ B_i\ ,\ K_i = 1/B_i\ ,\quad \sigma = D_{RMS}$

$$J(x) = \int_x^\infty (\sqrt{2\pi})^{-1} \exp(-0.5\ y^2)\ dy \quad \text{(gaussian integral subroutine)}$$

LOGIC

IF $F_o > 0.5\ (B_d + B_i)$ THEN no interference

ELSE IF $F_o > 0.5\ |B_d - B_i|$ THEN use Formula 1

ELSE IF $B_d \le B_i$ THEN use Formula 2

ELSE use Formula 3

FORMULAS

(1) $Q = 0.5\ K_i\ (Q_1 + Q_2)$

IF $f \ge B_d$ THEN $Q_1 = 0$

ELSE $L = f - 0.5\ B_d$, $U = \min(f-a,\ 0.5\ B_d)$

$Q_1 = J(L/\sigma) - J(U/\sigma)$

IF $f + a \ge 0.5\ B_d$ THEN $Q_2 = 0$

ELSE $L = f + a$, $U = 0.5\ B_d$

$Q_2 = J(L/\sigma) - J(U/\sigma)$

(2) IF $f \ge B_d$, THEN $Q = 0$

ELSE $L = F - 0.5\ B_d$, $U = 0.5\ B_d$

$Q = K_i\ [J(L/\sigma) - J(U/\sigma)]$

(continued)

TABLE B-2 (continued)

(3) IF $F_o = 0$ THEN $Q = Q_1$

ELSE $Q = 0.5\ K_i\ (Q_1 + Q_2)$

IF $f - b \geq 0.5\ B_d$ THEN $Q_1 = 0$

ELSE $L = f - b$, $U = \min\ (f-a,\ 0.5\ B_d)$

$Q_1 = J(L/\sigma) - J(U/\sigma)$

IF $f + a \geq 0.5\ B_d$ THEN $Q_2 = 0$

ELSE $L = f + a$, $U = \min\ (f + b,\ 0.5\ B_d)$

$Q_2 = J(L/\sigma) - J(U/\sigma)$

TABLE B-3

NPR FORMULA FOR GAUSSIAN-GAUSSIAN CASE

$$(NPR)_{dB} = (C/I)_{dB} + 10 \log \frac{2\sqrt{2\pi}\,(m)\,(\sigma_d^2)}{(1-\varepsilon) f^2 D(f) E(f)}$$

where f = baseband channel frequency

$m = \sigma/f_H$

$\varepsilon = f_L/f_H$

$\sigma = (\sigma_d^2 + \sigma_i^2)^{1/2}$

σ_d = rms frequency deviation of desired signal

σ_i = rms frequency deviation of interferer signal

F_o = carrier frequency offset

$D(f) = 1/P(f)$ (see Figure 5-1 in Section 5)

$E(f) = \exp\,[-(f - F_o)^2/(2\sigma^2)] + \exp\,[-(f + F_o)^2/(2\sigma^2)]$

Case of FDM/FM Telephony Interferer: Detailed Algorithm

Either the desired $S_d(f)$ or interferer $S_i(f)$ emission spectrum is not gaussian, or they are both gaussian but the effective interference spectrum $K_i(f)$ is truncated by the desired receiver passband. The FMSPC Program shown in Figure 5-7 (Section 5) is used to generate the emission spectra, and the interferer spectrum is then shifted and truncated appropriately. The SPCVL Program shown in Figure B-2 is employed to perform the spectral convolution and generate the spectrum Q(f) for the NPR computation in TABLE B-1. All desired baseband channels are sampled to determine the worst NPR performance.

Case of TV/FM Interference: Compact and Detailed Algorithms

The compact algorithm is similar to that used for the FDM/FM interferer case, except that the TV/FM frequency deviation is used to obtain the gaussian standard deviation, as discussed in Section 5. The detailed algorithm is similar to that used for the FDM/FM interferer case, except that the TVSPC Program shown in Figure 5-7 (Section 5) is used to generate the interferer emission spectrum used in Figure B-2.

Case of Wideband Digital PSK Interference: Compact Algorithm

The FDM/FM desired spectrum $S_d(f)$ has the gaussian formula given by Equation 5-2 in Section 5. The PSK interferer spectrum $S_i(f)$ has the sinc-squared formula given by Equation 5-3 in Section 5, but is replaced by its rectangular bound $S_i(f) = \alpha T$ to permit a simple convolution subroutine. The process already presented in TABLE B-2 is applicable, with $K_i = \alpha T$ being the only distinction. The Q values obtained are used for the NPR computation in TABLE B-1.

Case of Wideband Digital PSK Interferer: Detailed Algorithm

The FMSPC Program shown in Figure 5-3 (Section 5) is used to generate the FDM/FM desired spectrum. The PSK interferer spectrum $S_i(f)$ is the sinc-squared formula given by Equation 5-3 in Section 5. The SPCVL Program shown in Figure B-2 is employed to generate the spectrum Q(f) for the NPR computation in TABLE B-1. All baseband frequency channels are sampled to determine the worst NPR performance.

Case of SCPC/PSK and SCPC/FM Interferers: Compact and Detailed Algorithms

The effective number of SCPC/PSK (digital data) or SCPC/FM (analog voice) interferers that fits in the desired signal bandwidth (B_d) is $kN' = k\,(B_d/B_i')$, where B_i' is the allocated bandwidth of one SCPC interferer and k is an activity factor. The total interferer power is given by $(kN')I$ and the equivalent power density is given by $(kN')I/B_d = (k/B_i')I$, based on a uniform distribution over the desired passband. The value k = 0.4 is assumed for the activity factor.

The compact and detailed algorithms previously discussed for the CSSB/AM interferer case can now be used, with $K_i = k/B_i'$ as the uniform interference spectrum distributed over the desired signal bandwidth B_d. The C/I term appearing in TABLE B-1 is computed using only one SCPC interferer for the I value, since the SCPC multiplicity is already accounted in the K_i term.

APPENDIX C

RTC ALGORITHMS FOR TV/FM

TV/FM PERFORMANCE EVALUATION

The desired signal receiver extracts the composite audio-video baseband via frequency demodulation and deemphasis. An interfering signal contained within the desired receiver passband produces an interferer output component whose spectral characteristics differ from those of the interference input spectrum. The interferer output component is added to the desired output baseband and ultimately degrades the video (picture) and audio (sound) information extracted.

There are two main approaches to the evaluation of the video performance degradation. One approach relies on subjective measurements of the output picture quality based on statistical observations. Another approach relies on objective measurements of an output performance index, with magnitude requirements based on empirical results. These approaches can be combined into a unified performance evaluation procedure as illustrated below.

The subjective measurements for the video performance evaluation are based on an output picture quality rating, with a five-point grade scale usually employed [CCIR Recommendation 500-2, 1982]. The conversion from the quality grade Q to an input C/I requirement is called the protection ratio (PR) and is formulated in terms of the modulation specifications for the case of a cochannel TV/FM interferer into TV/FM, as shown in TABLE C-1. The formula has been found to be reliable for $4 < Q < 5$ on the conservative side, as discussed in APPENDIX G.

Various protection ratio formulas have been proposed to handle an offset TV/FM interferer, with a considerable variation in their parametric relations and characteristic features. The present trend consists of piece-wise linear functions of the offset magnitude, as summarized in TABLE C-2. Their relative features are discussed in APPENDIX G, where they are also compared to experimental measurements reported in the literature.

TABLE C-1

PROTECTION RATIO FORMULA FOR COCHANNEL TV/FM INTERFERER
[CCIR Recommendation 500-2, 1982]

$$(PR)_{dB} = V - 20 \log (D/12) - Q + 1.1Q^2$$

where V = 13.5 for 525-line M/NTSC

18.5 for 625-line K/SECAM

12.5 for 625-line I/PAL, G/PAL, L/SECAM

D = peak-to-peak frequency deviation in MHz

Q = picture quality grade

The development of protection ratios for the other interferer modulation types of interest is rather limited, both in the formulations and measurements available. The few experiments conducted lack sufficient variation in the modulation specifications to permit reliable generalizations or parametric formulations. The only pattern is that the formula in TABLE C-1 is too conservative for the other interferer types in question.

The objective measurements for the video performance evaluation are based on an output S/I measure. It is defined as the power ratio of the peak-to-peak luma signal to the weighted rms interference in the output video bandwidth [CCIR, Recommendation 567-1, 1982]. The S/I evaluation can be formulated for all interferer types if the interferer output spectrum is characterized to permit its integration over the video bandwidth.

The conversion from the output S/I performance into an input C/I threshold requires specification of an output threshold value. There are some CCIR guidelines for the case of thermal noise input [CCIR Report 215-5, 1982], which are applicable to any interferer having a uniform input spectrum over the desired receiver passband. Moreover, the output power ratio can be related to the picture quality grade, and this establishes a conversion between the objective and subjective performance measures.

TABLE C-2

PROTECTION RATIO FORMULAS FOR OFFSET TV/FM INTERFERER

(1) Bouchard and Chouinard, 1983

$$
\begin{aligned}
(PR)_{dB} &= 36.90 - 28.40\,|F/B| && \text{for } |F/B| \leq 0.379 \\
&= 53.25 - 71.63\,|F/B| && \text{for } 0.379 < |F/B| \leq 0.52 \\
&= 30.44 - 27.73\,|F/B| && \text{for } 0.52 < |F/B| \leq 0.7 \\
&= 44.64 - 48.05\,|F/B| && \text{for } |F/B| > 0.7
\end{aligned}
$$

where (F/B) = frequency offset/desired bandwidth

(2) CCIR/BSS Region 2, 1983

$$
\begin{aligned}
(PR)_{dB} &= 28 && \text{for } |F| \leq 8.36 \\
&= 51.09 - 2.762\,|F| && \text{for } 8.36 < |F| \leq 12.87 \\
&= 30.40 - 1.154\,|F| && \text{for } 12.87 < |F| \leq 21.25 \\
&= 48.38 - 2.000\,|F| && \text{for } |F| > 21.25
\end{aligned}
$$

where F = frequency offset in MHz

(3) CCIR/CPM Region 2, 1982

$$
\begin{aligned}
(PR)_{dB} &= (P.R.)_o && \text{for } |F/B| \leq 0.274 \\
&= (P.R.)_o - 35.6\,(|F/B| - 0.274) && \text{for } 0.274 < |F/B| \leq 0.92 \\
&= (P.R.)_o - 23 - 71\,(|F/B| - 0.92) && \text{for } |F/B| > 0.92
\end{aligned}
$$

where $(P.R.)_o$ = cochannel protection ratio in dB

(F/B) = frequency offset/desired bandwidth

(4) CCIR Report 634-2, 1982

$$
\begin{aligned}
(PR)_{dB} &= 30 && \text{for } -3 \leq F \leq 10 \\
&= 30 - (5/3)(F - 10) && \text{for } F > 10 \\
&= 30 - (10/9)(F + 3) && \text{for } F < -3
\end{aligned}
$$

where F = frequency offset in MHz

(5) CCIR Report 388-4, 1982

$$
\begin{aligned}
(PR)_{dB} &= (35 - 20 \log D') + (6/5)\,|F| && \text{for } |F| \leq 5 \\
&= (41 - 20 \log D') - (5/D')\,(|F| - 5) && \text{for } |F| > 5
\end{aligned}
$$

where $D' = 0.5D/(\sqrt{10})^{A/10}$, D = peak-to-peak deviation in MHz

A = 10 for 525-lines, 11 for 625-lines television

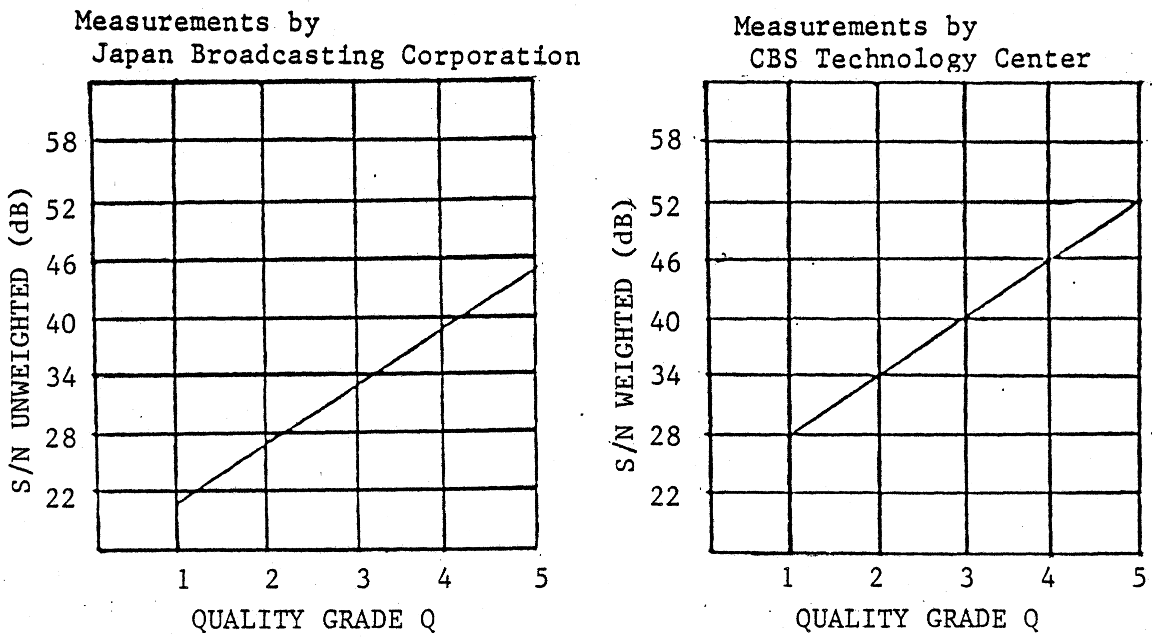
A linear relation between the output power ratio and the picture quality grade Q has been obtained by two recent independent measurements with M/NTSC television and thermal noise input, as shown in Figure C-1 [Izumi and Matsumae, 1982; Pritchard and Radin, 1984]. The same slope was obtained in both cases, and the two lines coincide when an 8-dB noise weighting is assumed, which is a reasonable value [CCIR Report 215-5, 1982].

An analogous relation can be developed for a TV/FM cochannel interferer. An empirical formula is available for the output S/I versus input C/I relation as discussed in APPENDIX G [CCIR Report 449-1, 1982]. It can be combined with the protection ratio formula in TABLE C-1 to obtain the output S/I versus picture quality relation formulated in Figure C-2.

A comparison between the thermal noise and the TV/FM cochannel interferer cases indicates that the thermal noise produces less subjective degradation for the same output power ratio. A weighted S/N = 49 dB is needed for Q = 4.5 with thermal noise, while a weighted S/I = 58.9 dB is needed with the TV/FM interferer. However, it should be cautioned that the protection ratio formula of TABLE C-1 has been found to be conservative for average program material, as discussed in APPENDIX G.

The output S/I consists of the output power ratio of the peak-to-peak luma signal to the weighted rms interference. The output power in the video component is given by $(2\pi D_{RMS})^2$, where D_{RMS} is the rms frequency deviation. This power is increased by a factor of $4\lambda^2$ for the peak-to-peak conversion, where λ is the peak/rms deviation ratio. The power is also reduced by a factor of μ^2 representing the fractional content in the luma signal of the video component. The values $\lambda^2 = 2$ and $\mu^2 \approx 0.5$ are used in practice [Ball and Rubin, 1978], which results in a desired output power (P_d) of $4(2\pi D_{RMS})^2$ for the peak-to-peak luma signal, as shown in TABLE C-3.

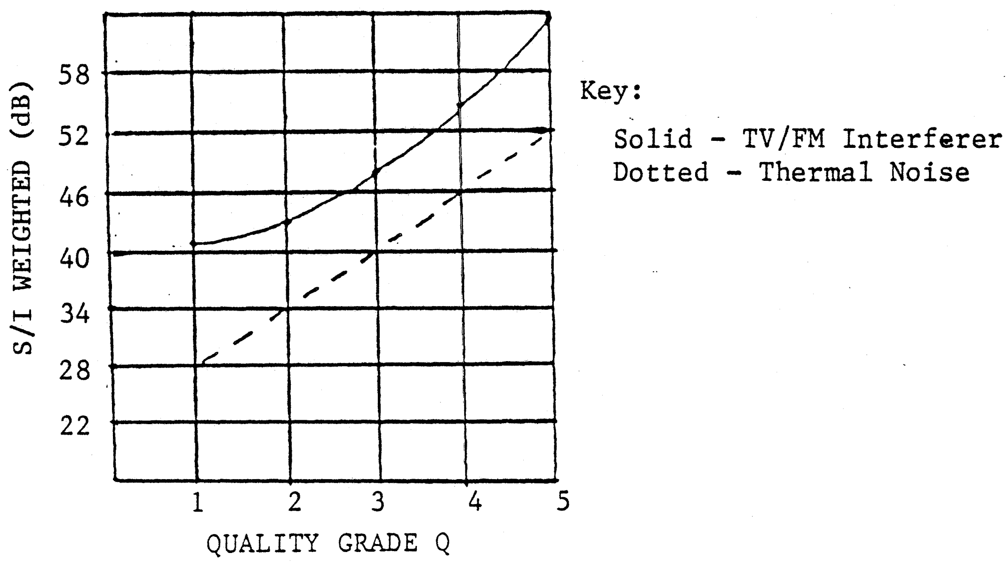
The unweighted interferer output power (P_i) is obtained by integrating the interferer output spectrum $K_o(f)$ over the video bandwidth (B_v). The generation of the $K_o(f)$ spectrum has the same development as in TABLE B-1 (APPENDIX B), since an FM demodulation process is involved in both cases. However, a conventional approach for TV/FM is to formulate the $K_o(f)$ spectrum without the deemphasis effect and scale the results by a gain factor (G_{PW}) that jointly accounts for the preemphasis and noise weighting effects. The



$$(S/N)_{dB} = 6Q + 14$$
unweighted

$$(S/N)_{dB} = 6Q + 22$$
weighted

Figure C-1. Output power ratio versus picture quality for M/NTSC television and thermal noise input.



$(S/I)_{dB}$ weighted $= (C/I)_{dB} + 20 \log D + 6$ (see APPENDIX G)

$= (V + 6 + 20 \log 12) - Q + 1.1Q^2$ (see TABLE C-1)

$= 41.1 - Q + 1.1Q^2$

Figure C-2. Output power ratio versus picture quality for M/NTSC television and TV/FM cochannel interferer.

TABLE C-3

TV/FM OUTPUT S/I PERFORMANCE FORMULATION

$$P_d = (2\lambda\mu)^2 (2\pi D_{RMS})^2 = 2(2\pi D_{pk})^2 \qquad \text{(desired output power)}$$

$$P_i = \int_{B_v} K_o(f)\, df \qquad \text{(interferer output power)}$$

where

$$K_o(f) = (2\pi f)^2 D(f) R(f) \qquad \text{(interferer output spectrum)}$$

$$D(f) = 1/P(f) \qquad \text{(see Figure 5-4 in Section 5)}$$

$$R(f) \approx (C/I)^{-1} \cdot Q(f) \quad \text{if } C/I \gg 1$$

$$Q(f) = 0.5\,[K(f) + K(-f)]$$

$$K(f) = S_d(f) * K_i(f) \qquad \text{(* denotes convolution)}$$

$$(S/I)_{dB(unweighted)} = 10 \log(P_d/P_i)$$

$$= (C/I)_{dB} + 20 \log D_{PK} - 10 \log \left[\int_{B_v} f^2 D(f)\, Q(f)\, df\right]$$

$$(S/I)_{dB(weighted)} = (S/I)_{dB(unweighted)} + (G_W)_{dB}$$

$$= (C/I)_{dB} + 20 \log D_{PK} - 10 \log \left[\int_{B_v} f^2 D(f)\, Q(f)\, df\right] + (G_{PW})_{dB}$$

$$(C/I)_{dB(margin)} = (S/I)_{dB(weighted)} - (S/I)_{dB(threshold)}$$

$$(C/I)_{dB(threshold)} = (C/I)_{dB} - (C/I)_{dB(margin)}$$

<u>Note:</u> $\lambda = \sqrt{2}$ and $\mu = 1/\sqrt{2}$ assumed

value G_{PW} = 12.8 dB is often assumed for M/NTSC television can be verified to be representative for both test-tone and random-noise modulation [Loo, 1977].

The generalized S/I formulation is shown in TABLE C-3. The analogy between the unweighted S/I expression for TV/FM in TABLE C-3 and the NPR expression for FDM/FM in TABLE B-1 is evident. They both represent desired-to-interferer, unweighted, output power ratios following an FM demodulation process. One distinction is the D_{PK} parameter appearing in the TV/FM case versus the D_{RMS} parameter in the FDM/FM case, as a consequence of the peak-to-peak luma power content being used in the TV/FM performance measure.

Another distinction has to do with the interferer output power (P_i) computation. The interferer output spectrum $K_o(f)$ is essentially constant over a narrow telephony channel bandwidth (B_c) in the FDM/FM case, but not over the wide video bandwidth (B_v) in the TV/FM case. Hence, the P_i computation only requires the $K_o(f)$ spectrum evaluation at the channel frequency (plus the channel bandwidth multiplication) in the FDM/FM case, while the $K_o(f)$ spectrum integration must be performed in the TV/FM case.

The general formulation in TABLE C-3 can be bypassed if the effective interference spectrum is uniform, with $K_i = 1/B_i$ over the desired signal bandwidth $B_d \leq B_i$. The conventional FM formula for the output power ratio with thermal noise input can be used, but the interferer power truncation by the (B_d/B_i) ratio must be included. The results are summarized in TABLE C-4 and can be verified to reproduce the standard thermal-noise formula when I = N and $B_i = B_d$ [CCIR Report 215-5, 1982].

TABLE C-4

OUTPUT S/I FORMULA FOR WHITE INTERFERENCE CASE

$P_d = 2\,(2\pi D_{PK})^2$ (desired signal power)

$K_i(f) = 1/B_i$ (normalized interferer input spectrum)

$K_o(f) = (IK_i/C)\,(2\pi f)^2$ (interferer output spectrum w/o deemphasis)

$P_i = \int_{B_v} K_o(f)df = (2\pi)^2(IK_i/C)\,(B_v^3/3)$ (interferer ouput power)

$$
\begin{aligned}
(S/I)_{dB(weighted)} &= 10\log(P_d/P_i) + (G_{PW})_{dB} \\
&= (C/I)_{dB} + 10\log[6(D_{PK}/B_v)^2\,(K_i\,B_v)^{-1}] + (G_{PW})_{dB} \\
&= (C/I)_{dB} = 10\log[3(D_{PP}/B_v)^2\,(B_i/2B_v)] + (G_{PW})_{dB}
\end{aligned}
$$

The use of TABLE C-3 requires a S/I threshold value to convert the output S/I performance into the input C/I threshold margin. The results of Figure C-1 indicate that S/I = 49 dB weighted is needed for Q = 4.5 picture quality with M/NTSC television and thermal noise. The CCIR cites thresholds of 45 dB (0.1% criterion) and 53 dB (1% criterion) for band-limited noise, with the 1% criterion further recommended for the fixed-satellite service [CCIR Recommendations 354-2, 483-1 and 567-1, 1982].

COMPACT AND DETAILED ALGORITHM COMPUTATIONS

The formulation of the S/I output performance based on TABLE C-3 as a function of the interferer modulation is not easy. Each interferer type is distinguished by its effective input spectrum $K_i(f)$ producing a distinct spectral convolution K(f) and output spectrum $K_o(f)$. The basic computation steps to be performed by a dedicated RTC algorithm are:

1. generation of the desired $S_d(f)$ and interferer $K_i(f)$ input spectra (if needed) from the modulation specifications

2. generation of the convolution K(f) and superposition Q(f) spectra from the desired and interferer input spectra

3. integration of the interferer output spectrum $K_o(f)$ (without deemphasis) over the video bandwidth B_v to determine the interferer output power P_i (the integration may not be simple due to the $(2\pi f)^2$ factor present in the $K_o(f)$ spectrum)

4. evaluation of the weighted output S/I performance and the C/I threshold margin using simple conversion formulas.

One approach to bypass the S/I formulation complexity is to require that the effective interference input power (spectral truncation included) be limited to 4% of the input C/N needed for an output S/N = 53 dB weighted [CCIR Recommendation 483-1, 1982]. The C/N value is obtained via TABLE C-4 (replacing I by N and B_i by B_d), and the effective C/I threshold is 14 dB lower. This approach is used for the compact algorithm versions.

The input C/I level is reduced by 10 log ρ in the compact algorithms, where ρ is an interference spectral truncation factor. The latter corresponds to the effective input spectrum $K_i(f)$ integrated over the desired receiver passband, which can be formulated in terms of the overlap bandwidth B_o given by Equation (4-1) in Section 4. The difference between the effective C/I input and threshold values yields the C/I threshold margin for the compact algorithms.

The formulation in TABLE C-3 is not used for the TV/FM interferer case, since the nature of the picture quality degradation caused by this interference has been noted to differ from thermal noise effects [Barnes, 1979]. An empirical output S/I expression is available for a TV/FM cochannel interferer, as discussed in APPENDIX G, but its extension to offset conditions is hitherto unknown. Also, the relation between the output S/I expression and the picture quality Q is only available under cochannel conditions, as formulated in Figure C-2.

However, the C/I threshold is available under cochannel and offset conditions, since it corresponds to the formulas in TABLES C-1 and C-2. In particular, only the third formula in TABLE C-2 is compatible with TABLE C-1, and it is selected for the C/I threshold formulation on this basis. The C/I margin is obtained as the difference between the C/I input and threshold values, and the compact and detailed algorithm versions are identical for the TV/FM interferer case. A value of Q = 4.5 is used in TABLE C-1.

Case of CSSB/AM Telephony Interferer: Compact and Detailed Algorithms

A CSSB/AM telephony interferer has a rectangular spectrum with $K_i(f) = 1/B_i$ over the spectral overlap region. The compact algorithm uses this expression to derive the spectral truncation factor $\rho = B_o/B_i$. The detailed algorithm uses the white interference formulation in TABLE C-4 when the spectral overlap spans the desired signal bandwidth.

The white interference model becomes more conservative as the spectral overlap becomes more restricted. The general formulation in TABLE C-3 is then needed in the detailed algorithm, and it requires the convolution of the desired and interferer input spectra, followed by the generation and integration of the interferer output spectrum.

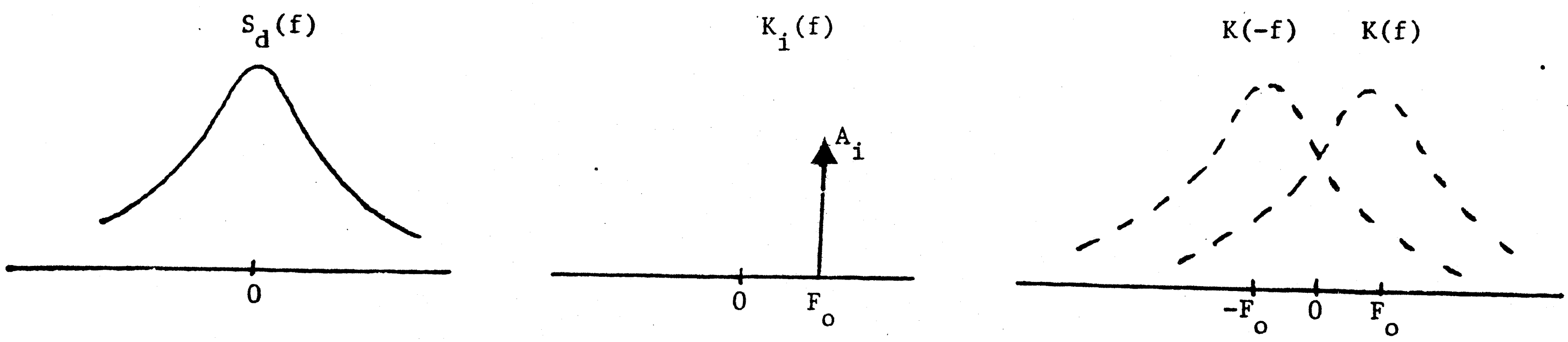
A gaussian envelope representation for the TV/FM spectrum (see Section 5) permits the use of the procedure in TABLE B-2 (see APPENDIX B) to perform the spectral convolution with the rectangular interferer spectrum. However, such a procedure has to be recycled to generate all spectral convolution samples. Each sample is then weighted by $(2\pi f)^2$ to generate the interferer output spectrum, and all samples are fed to an integral subroutine to obtain the interferer output power.

The longer processing time involved is the recycling of the gaussian area evaluations as the output samples are generated. These repeated evaluations can result in a long execution time, so other options sacrificing accuracy for expediency can be included.

The computation time can be reduced if the effective interference is narrowband enough to have an impulse spectrum representation. The gaussian-impulse convolution has a simple formula as shown in Figure C-3. The spectral convolution samples are derived by evaluating the Q(f) formula so that TABLE B-2 is bypassed. However, the multiple sample generation and processing is still needed to perform the video baseband integration, since the $(2\pi f)^2Q(f)$ expression does not have a simple integration.

The gaussian-impulse model in Figure C-3 does not accommodate the intermediate case where the effective interference is neither wideband nor narrowband. This can be resolved by using the white interference model (TABLE C-4) down to a spectral overlap breakpoint as a conservative estimate, and using the gaussian-impulse model thereafter. The breakpoint can be preset to a fixed value (say 50% spectral overlap), and the ratio B_o/B_d is used to automatically trigger the appropriate model and formulas by computing the spectral overlap bandwidth B_o given by Equation 4-1 in Section 4.

Another approach that would bypass the multiple sample processing still required consists of using a rectangular envelope with the same gaussian peak



$$S_d(f) = (\sqrt{2\pi}\ \sigma)^{-1}\ e^{-0.5(f/\sigma)^2}$$

$$K_i(f) = A_i\ \delta(f-F_o) \quad \text{where} \quad A_i = \begin{cases} 1 \text{ if } B_i < B_d \\ B_o/B_d \text{ if } B_o < B_d < B_i \end{cases}$$

and B_o = see Equation 3-1 in Section 3

$$K(f) = A_i\ (\sqrt{2\pi}\ \sigma)^{-1}\ e^{-0.5(f-F_o)^2/\sigma^2}$$

$$Q(f) = 0.5\ A_i\ (\sqrt{2\pi}\ \sigma)^{-1}\ [e^{-0.5(f-F_o)^2/\sigma^2} + e^{-0.5(f + F_o)^2/\sigma^2}]$$

Figure C-3. Gaussian-impulse spectral convolution.

and unit power for the desired spectrum, along with the rectangular interference spectrum under cochannel conditions, as shown in Figure C-4. The cochannel condition is assumed to provide a conservative estimate, rather than claim no interference when the spectral overlap occurs in the gaussian tails truncated by the rectangular envelope.

The rectangle-rectangle model in Figure C-4 yields a $(2\pi f)^2 Q(f)$ expression with a simple integration as shown in TABLE C-5. A fast execution time is provided at the expense of spectral representation accuracy (rectangular envelope for desired signal) and conservative performance evaluation (cochannel interference conditions).

TABLE C-5

CSSB/AM INTERFERER OUTPUT POWER USING COCHANNEL-RECTANGLES CONVOLUTION FOR NARROWBAND CASE

Case of $B_o \geq B_d$: Use TABLE C-4

Case of $B_o < B_d$:

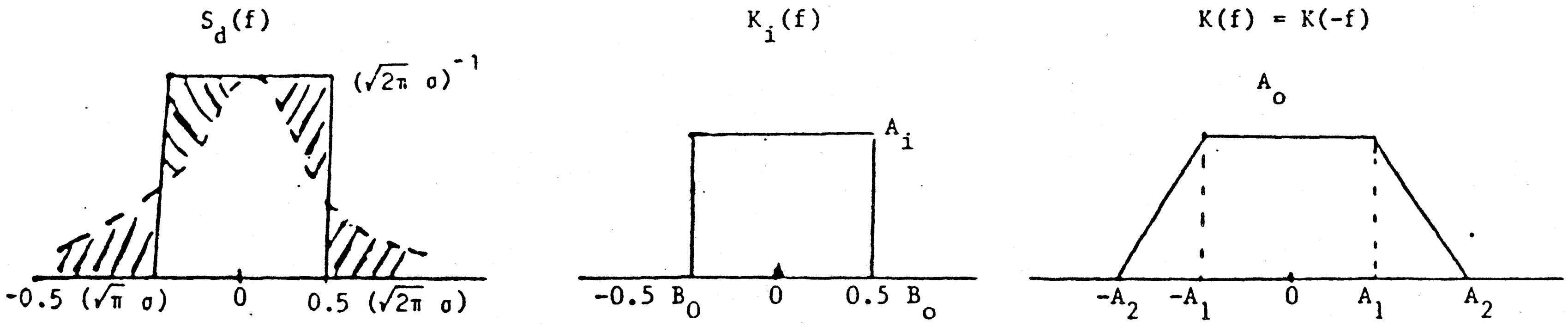
IF $B_o \geq \sqrt{2\pi}\ \sigma$ then use TABLE C-4

IF $B_o < \sqrt{2\pi}\ \sigma$ then use as follows

$$P_i = (2\pi)^2 (A_o I/3C)\, B_v^3 \quad \text{if } B_v \leq A_1$$

$$= (2\pi)^2 (A_o I/3C)\, [A_1^3 + [1 + (A_1/B_o)]\, (B_v^3 - A_1^3) - (3/4\, B_o)\, (B_v^4 - A_1^4)] \quad \text{if } A_1 \leq B_v \leq A_2$$

$$= (2\pi)^2 (A_o I/3C)\, [A_1^3 + [1 + (A_1/B_o)]\, (A_2^3 - A_1^3) - (3/4\, B_o)\, (A_2^4 - A_1^4)] \quad \text{if } B_v > A_2$$



$S_d(f) \;=\; (\sqrt{2\pi}\,\sigma)^{-1}$ for $|f| \leq 0.5\,(\sqrt{2\pi}\,\sigma)$

$K_i(f) \;=\; A_i$ for $|f| \leq 0.5\,B_o$, where $A_i = 1$(if $B_i < B_d$) or B_o/B_d (if $B_o < B_d < B_i$)

B_o = see Equation 4-1 in Section 4

$$Q(f) \;=\; K(f) \;=\; A_o \quad \text{if} \quad |f| \leq A_1$$

$$= \; A_o - (A_o/B_o)\,(|f| - A_1) \quad \text{if} \quad A_1 \leq |f| \leq A_2$$

where $A_o \;=\; (A_iB_o)/(\sqrt{2\pi}\,\sigma)$

$$A_1 \;=\; 0.5\,(\sqrt{2\pi}\,\sigma - B_o)$$

$$A_2 \;=\; 0.5\,(\sqrt{2\pi}\,\sigma + B_o)$$

Figure C-4. Rectangle-rectangle spectral convolution ($B_i < \sqrt{2\pi}\,\sigma$).

The detailed algorithm processing compromise is summarized in Figure C-5. The complex option for the narrow interference case uses the TVSPC Program of Figure 5-7 (Section 5) to generate the TV/FM spectrum, and the SPCVL Program in Figure B-2 (APPENDIX B) to perform the spectral convolution with the rectangular interferer spectrum. The complex option also includes the weighting and integration subroutines following the spectral convolution process.

Case of FDM/FM Telephony Interference: Compact and Detailed Algorithms

The use of gaussian spectral models for the TV/FM desired and FDM/FM interferer signals permits a simple gaussian integral subroutine to derive the spectral truncation factor ρ in the compact algorithm, as well as a simple formula for the spectral convolution in the detailed algorithm (see Figure C-6). However, the latter still requires the generation and processing of multiple samples to derive the interferer output power P_i, since the $(2\pi f)^2 Q(f)$ integration does not have a simple formula.

This last complexity can be bypassed by replacing the gaussian terms in Q(f) by their rectangular envelopes with the same peak value and power content. However, the gaussian tails will be truncated around the origin for $F_o > 0.5\sqrt{2\pi}\,\sigma$, as shown in Figure C-7, so that their contributions to the interferer output power in the $|F_o - B_v|$ portion of the video baseband integration will not be accounted for. The error magnitude is a function of the (B_v, F_o, σ) parameter values.

The cochannel condition in Figure C-7 can be assumed to provide a conservative estimate rather than neglect some interference power. The video baseband integration of the $(2\pi f)^2 Q(f)$ spectrum has the simple formula shown in TABLE C-6. A fast execution time is provided at the expense of spectral convolution accuracy (rectangular approximation to gaussian outcome) and conservative performance evaluation (cochannel interference conditions).

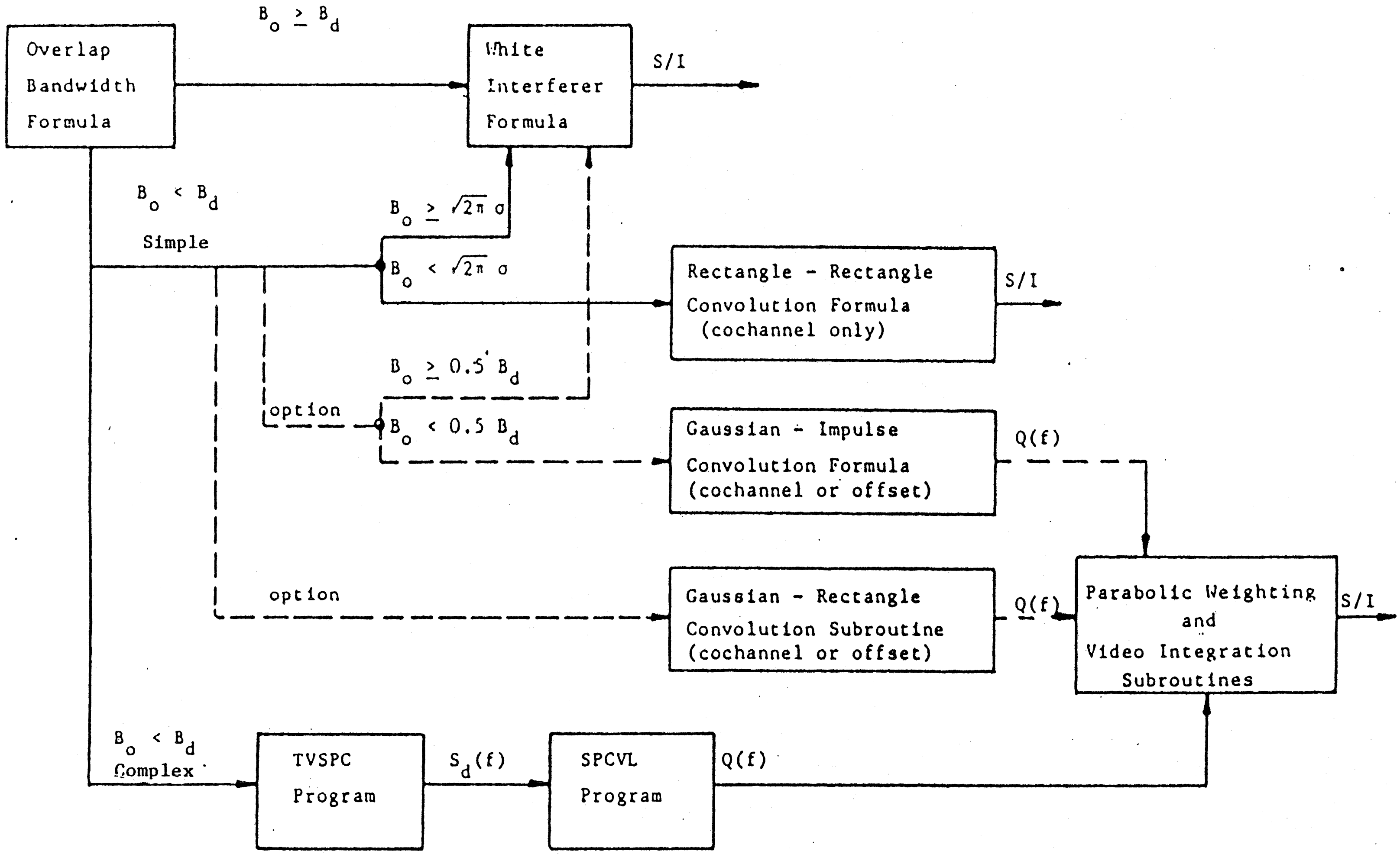
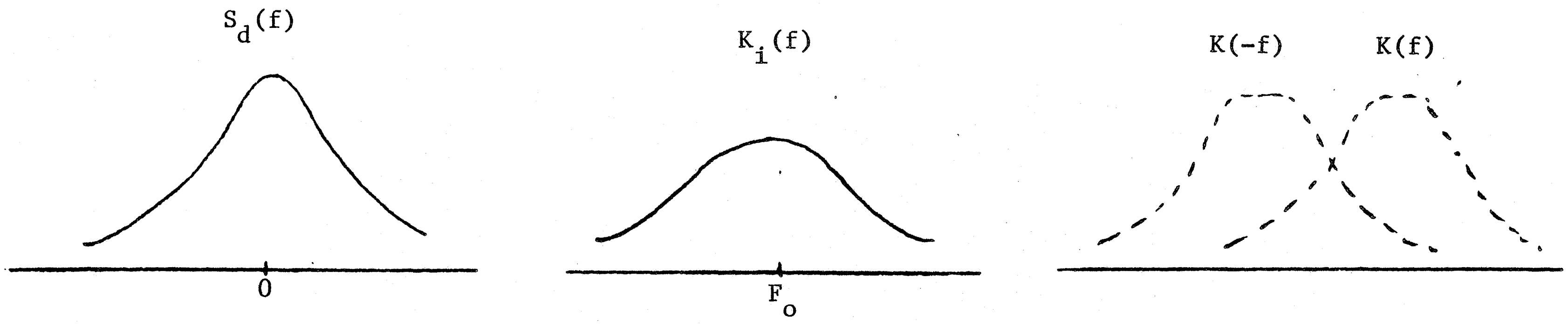


Figure C-5. Detailed algorithm processing compromise for CSSB/AM interference into TV/FM.



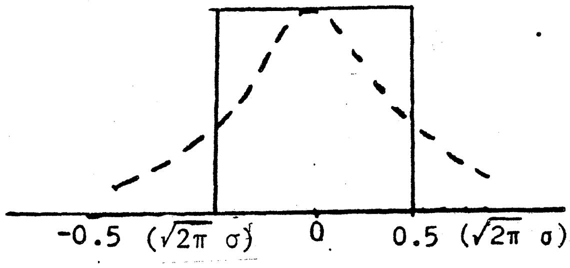
$$S_d(f) = (\sqrt{2\pi}\,\sigma_1)^{-1}\, e^{-0.5(f/\sigma_1)^2}$$

$$K_i(f) = (\sqrt{2\pi}\,\sigma_2)^{-1}\, e^{-0.5(f-F_o)^2/\sigma_1^2}$$

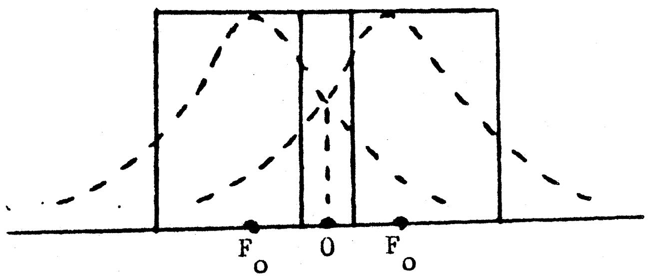
$$K(f) = (\sqrt{2\pi}\,\sigma)^{-1}\, e^{-0.5(f-F_o)^2/\sigma^2}, \quad \text{where } \sigma^2 = \sigma_1^2 + \sigma_2^2$$

$$Q(f) = 0.5(\sqrt{2\pi}\,\sigma)^{-1}\, [e^{-0.5(f-F_o)^2/\sigma^2} + e^{-0.5(f+F_o)^2/\sigma^2}]$$

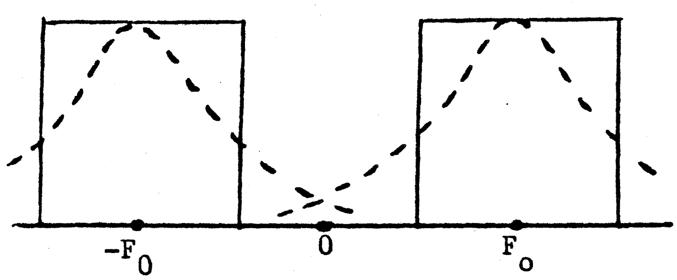
Figure C-6. Gaussian-gaussian spectral convolution.



Case of $F_o = 0$



Case of $F_o < 0.5(\sqrt{2\pi}\ \sigma)$



Case of $F_o > 0.5(\sqrt{2\pi}\ \sigma)$

Figure C-7. Spectral tail truncation by rectangular approximation to gaussian terms after spectral convolution.

TABLE C-6

FDM/FM INTERFERER OUTPUT POWER USING EQUIVALENT COCHANNEL RECTANGLE INTEGRATION

$$P_i = (2\pi)^2 \, (I/3C) \, (\sqrt{2\pi}\,\sigma)^{-1} \, B_v^{\,3} \quad \text{if } B_v \leq 0.5\,(\sqrt{2\pi}\,\sigma)$$

$$P_i = (2\pi)^2 \, (I/3C) \, (\sqrt{2\pi}\,\sigma)^{-1} \, (0.5\,\sqrt{2\pi}\,\sigma)^3$$

$$= (2\pi)^2 \, (I/24C) \, (\sqrt{2\pi}\,\sigma)^2 \quad \text{if } B_v > 0.5\,(\sqrt{2\pi}\,\sigma)$$

Note: The rectangular approximation to the gaussian characteristic is introduced after the spectral convolution operation, unlike the CSSB/AM interferer case in TABLE C-5 where the rectangle approximation is on the spectrum itself before convolution.

The detailed algorithm processing compromise is summarized in Figure C-8. The complex version uses the TVSPC and FMSPC Programs shown in Figures 5-7 and 5-3 (Section 5) to generate the desired and interferer spectra, along with the SPCVL Program in Figure B-2 (APPENDIX B) to perform the spectral convolution. The complex version also includes the weighting and integration subroutines following the spectral convolution process.

Case of Wideband Digital PSK Interferer: Compact and Detailed Algorithms

A wideband digital PSK interferer has the sinc-squared emission spectrum given by Equation 5-3 in Section 5. The rectangular bound $K_i(f) = \alpha T$ is used to derive the spectral truncation factor $\rho = B_o(\alpha T)$ for the compact algorithm.

The detailed algorithm processing compromise is similar to the one previously described for the CSSB/AM interferer case, except that the SPCVL Program now performs the spectral convolution with the interferer spectrum generated from the sinc-squared formula instead of a rectangular formula.

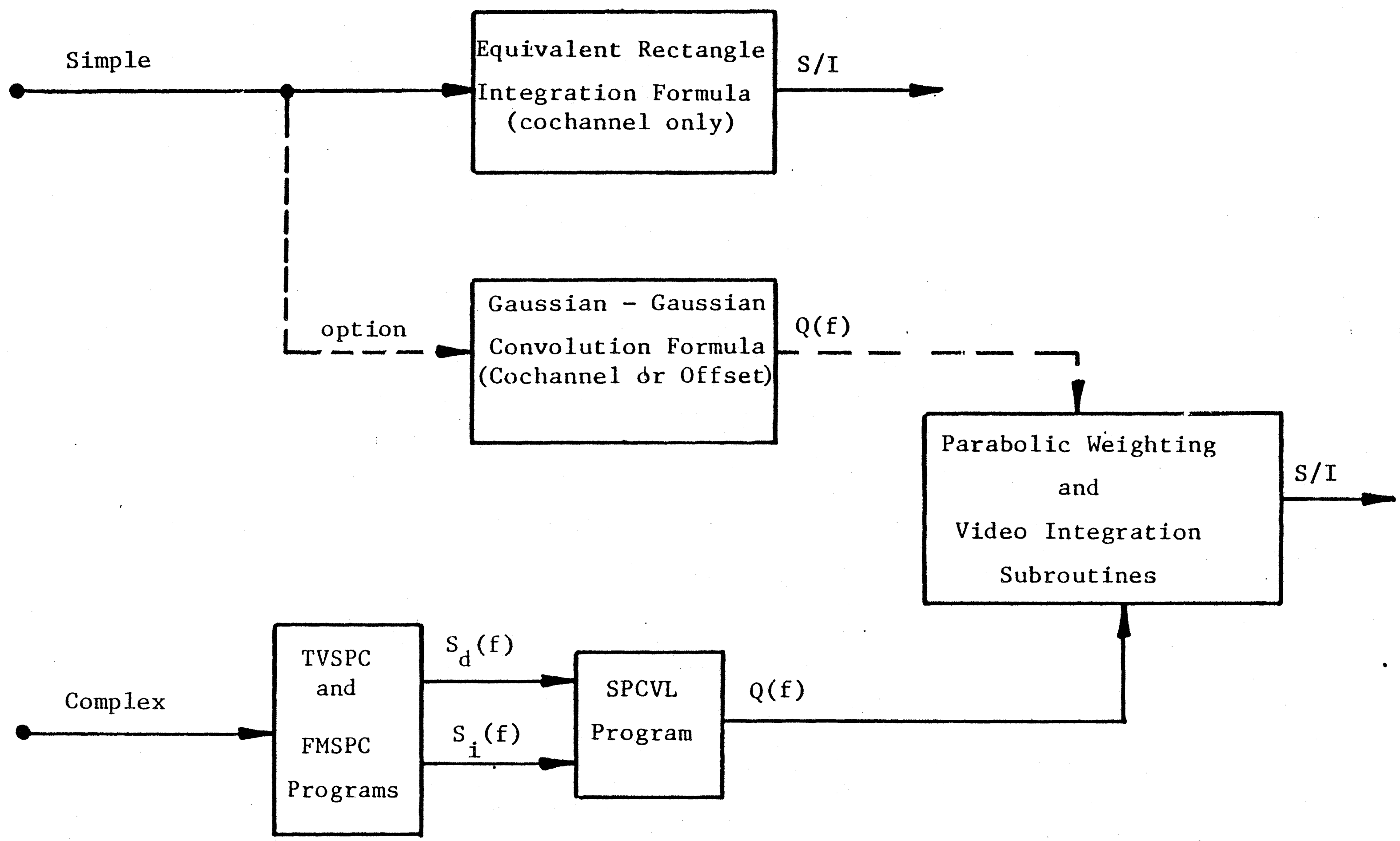


Figure C-8. Detailed algorithm processing compromise for FDM/FM interference into TV/FM.

Case of SCPC/PSK and SCPC/FM Interferers: Compact and Detailed Algorithms

The effective interference spectrum is uniform with $K_i = k/B_i'$ representing the collective effect, as described in APPENDIX B for these interferer types. The compact algorithm has $\rho = K_i B_c$ (spectral accumulation instead of spectral truncation), and the detailed algorithm uses the white interference formula of TABLE C-4. A value $k = 0.4$ is assumed for the activity factor.

APPENDIX D

RTC ALGORITHMS FOR DIGITAL CPSK

DIGITAL CPSK PERFORMANCE EVALUATION

The PSK receiver demodulates the desired input signal to extract the analog waveform representing the digital data symbols. The latter are then examined to estimate which one of the M possible phase states is being represented. A CPSK receiver coherently extracts the analog waveforms via product demodulation, which is followed by a filter-and-sample process over a symbol duration. This produces an output state decision variable whose value is recognized as representing one of the possible phase states according to a decision logic.

The M phase states can be viewed as equidistant points on a circle corresponding to signal phase angles $\theta = 360°/n$ where $n = 1$ to M. These phase states are separated by 360°/M, which represents a ± 180°/M phase error margin that can be tolerated without making incorrect decisions. Any undesired phase perturbations in the demodulated signal creates departures from the nominal state values, which reduces the error margin tolerance and causes decision errors when the margin is exceeded.

The presence of thermal noise causes random phase perturbations and output state variations characterized by a statistical probability of exceeding the error margin and making incorrect decisions. This decision error probability is used as a performance measure for digital systems and has a conventional formulation in terms of the input signal-to-noise ratio (C/N) when equiprobable digital symbols are assumed. The bit and symbol error rates (BER, SER) for thermal noise and CPSK matched-filter detection are formulated in TABLE D-1 [Stein and Jones, 1967], including coding gain and implementation loss effects.

An interfering signal within the desired input passband also contributes phase perturbations at the demodulator output. These interference effects have a reduced error margin in the presence of thermal noise, since the

TABLE D-1

CPSK ERROR PERFORMANCE FOR THERMAL NOISE

$$\text{BER} = (\log_2 M)^{-1} \cdot (\text{SER})$$

$$\text{SER} = K_1 \text{ erfc } \sqrt{\gamma} = K_1 \left[(2/\sqrt{\pi}) \int_{\sqrt{\gamma}}^{\infty} \exp(-x^2)dx\right]$$

where $K_1 = 0.5$ (if $M = 2$) or 1.0 (if $M > 2$)

$$\gamma = K_2(C/N) = (\sin \pi/M)^2 (G/L) (E/N_o)$$

$$K_2 = (\sin \pi/M)^2 (G/L) (B_d T)$$

$$E/N_o = (B_d T)(C/N) = (\log_2 M)(B_d/B_r)(C/N)$$

and

B_d = desired signal bandwidth

B_r = bit rate

E = symbol energy

G = coding gain

L = implementation loss

M = number of phase states

N_o = noise power density

T = symbol duration

±180°/M tolerance must now be shared by the interference and thermal noise perturbations. Moreover, the error probability contributions of the interference and thermal noise effects cannot be independently added, since the error performance is a nonlinear function of their combined perturbations.

If the interference has a uniform spectrum over the desired signal bandwidth, then the formulation of TABLE D-1 can be extended to include the interference effects, as shown in TABLE D-2. The N_o term is replaced by $N_o' = N_o + (I/B_i)$ to represent the total power density due to both thermal noise and interference. The interference contribution includes the spectral truncation by the receiver passband and the truncated interference power is $I' = (B_d/B_i)I$. The net effect is the C/N term in TABLE D-1 being replaced by the $C/N_{TOTAL} = C/(N + I')$ term in TABLE D-2.

The effects of a nonwhite interference spectrum on the error performance in the presence of thermal noise is discussed next. The cases of a sinusoidal CW interferer with random phase and a wideband FM interferer spanning the receiver bandwidth are considered under cochannel conditions. These two cases correspond to an impulse spectrum and a gaussian spectrum as the interference characteristic.

The results obtained from theoretical formulations are presented in TABLES D-3 and D-4 for uncoded binary (M = 2) and quaternary (M = 4) CPSK as the desired signal [Rosenbaum, 1969; Morinaga and Namekawa, 1982]. The tables show the (C/I, C/N) paired values that support a given error rate performance. The results clearly illustrate the need to jointly account for interference and thermal noise effects when evaluating the PSK error performance.

A comparison between the results of TABLES D-3 and D-4 for a fixed M value shows that there is no significant difference in the error performance for the two interferer types considered. There is little variation in the (C/I, C/N) pairings needed for a given error performance between the two tables, regardless of the desired signal case (M = 2 or 4) or the error performance requirement (10^{-3} to 10^{-6}).

A third interference type considered is a binary PSK cochannel interferer on a binary CPSK desired signal. This case corresponds to a

TABLE D-2

CPSK ERROR PERFORMANCE FOR WHITE INTERFERENCE PLUS THERMAL NOISE

$$BER = (\log_2 M)^{-1} \cdot (SER)$$

$$SER = K_1 \operatorname{erfc} \sqrt{\gamma} = K_1 \left[(2/\sqrt{\pi}) \int_{\sqrt{\gamma}}^{\infty} \exp(-x^2)dx\right]$$

where $K_1 = 0.5$ (if $M = 2$) or 1.0 (if $M > 2$)

$$\gamma = K_2(C/N_{TOTAL}) = (\sin \pi/M)^2 (G/L)(E/N_o')$$

$$K_2 = (\sin \pi/M)^2 (G/L)(B_d T)$$

$$E/N_o' = (B_d T)(C/N_{TOTAL}) = (\log_2 M)(B_d/B_r)(C/N_{TOTAL})$$

$$C/N_{TOTAL} = C/(N + I') = [(C/N)^{-1} + (C/I')^{-1}]^{-1}$$

$$= (C/N)[1 + \rho \ (C/N)(C/I)^{-1}]^{-1}$$

and

B_d = desired signal bandwidth

B_i = interference signal bandwidth

B_r = bit rate

E = symbol energy

G = coding gain

L = implementation loss

M = number of phase states

N_o = noise power density

N_o' = noise plus interference power density

T = symbol duration

I' = ρI = truncated interferer input power

ρ = B_d/B_i = interferer spectral truncation factor

TABLE D-3

CW SINUSOIDAL (IMPULSE SPECTRUM) COCHANNEL INTERFERENCE
INTO BINARY AND QUATERNARY CPSK

Binary CPSK (M = 2)			Quaternary CPSK (M = 4)		
BER	C/I(dB)	C/N(dB)	SER	C/I(dB)	C/N(dB)
10^{-3}	30	6.6	10^{-3}	30	10.2
"	20	7.0	"	20	10.9
"	15	7.4	"	15	11.7
"	10	8.7	"	10	14.0
"	5	12.0	"	5	22.0
10^{-4}	30	8.4	10^{-4}	30	11.8
"	20	8.7	"	20	12.2
"	15	9.2	"	15	13.4
"	10	10.6	"	10	15.7
"	5	14.0	"	5	24.2
10^{-5}	30	9.5	10^{-5}	30	13.0
"	20	10.0	"	20	13.7
"	15	10.3	"	15	14.7
"	10	12.0	"	10	17.0
"	5	15.8	"	5	N/A
10^{-6}	30	10.7	10^{-6}	30	14.0
"	20	11.0	"	20	14.7
"	15	11.7	"	15	15.7
"	10	13.0	"	10	18.0
"	5	16.8	"	5	N/A

Note: The analytical model is based on ideal coherent detection, instantaneous phase sampling, and no intersymbol interference.

TABLE D-4

WIDEBAND FM (GAUSSIAN SPECTRUM) COCHANNEL INTERFERENCE
INTO BINARY AND QUATERNARY PSK

Binary CPSK (M = 2)			Quaternary CPSK (M = 4)		
BER	C/I(dB)	C/N(dB)	SER	C/I(dB)	C/N(dB)
10^{-3}	∞	6.8	10^{-3}	∞	10.4
"	20	7.0	"	20	10.8
"	10	8.8	"	10	14.0
"	5	12.0	"	5	N/A
10^{-4}	∞	8.4	10^{-4}	∞	11.8
"	20	8.8	"	20	12.5
"	10	10.6	"	10	15.5
"	5	14.0	"	5	N/A
10^{-5}	∞	9.7	10^{-5}	∞	13.0
"	20	10.0	"	20	13.7
"	10	12.0	"	10	17.0
"	5	N/A	"	5	N/A
10^{-6}	∞	10.5	10^{-6}	∞	14.0
"	20	11.0	"	20	14.7
"	10	13.0	"	10	N/A
"	5	N/A	"	5	N/A

Note: The analytical model assumes ideal product demodulation, instantaneous phase sampling, and no intersymbol interference.

sinc-squared spectrum as the interference characteristic. The interferer data rate was set equal to or less than the desired signal, so that the interference spectrum is contained within the receiver bandwidth.

The theoretical results obtained are presented in TABLE D-5 for two distinct receiver input filters [Celebiler and Coupe, 1978]. There was no significant variation as a function of the interferer data rate in both narrowband and wideband interference cases. The filter distinction can also be noted to have a secondary effect in the error performance. A comparison with the results of TABLES D-3 and D-4 again shows little variation in the (C/I, C/N) pairings needed to support a given error performance, so that there is no significant difference between the three interference types considered.

Hence, the formulation corresponding to the CW sinusoidal case can be used to represent a variety of interference spectra (impulse, gaussian, sinc-squared) under cochannel conditions. This variety suggests that the formulation can also be used as an effective bound under offset conditions, since the ρ term accounts for the interference spectral truncation. This approach is consistent with the CCIR formulation on this matter, which does not distinguish between cochannel and offset conditions but includes spectral truncation effects [CCIR Report 388-4, 1982].

The nonwhite interference formulation is shown in TABLE D-6 [Rosenbaum, 1969]. The phase-averaging integration required can be bypassed by setting $\cos \phi = -1$ as a worst condition. The upper bound thus obtained only requires the erfc function evaluation. A comparison of the results obtained with the averaging and upper-bound expressions is shown in Figure D-1, assuming $M = 2$, $G/L = 1$, and $\rho = 1$ for simplicity. The upper-bound yields the same order of magnitude in the 10^{-3} to 10^{-6} BER performance region for $C/I \geqq 10$ dB, so that its simpler formulation can be used for practical purposes.

However, the formulas in TABLE D-6 do not support the white interference case (TABLE D-2) under moderately low C/I conditions. A comparison of the two formulations is shown in Figure D-2, assuming the averaging formula with $G/L = 1$ and $\rho = 1$ in TABLE D-6 and the condition $B_d = B_i$ in TABLE D-2. The white interference case matches the nonwhite interference error performance for $C/I \geqq 30$ dB and maintain the same order of magnitude in the error performance for $15 \leq C/I < 30$ dB.

TABLE D-5

BINARY PSK (SINC-SQUARED SPECTRUM) COCHANNEL INTERFERENCE INTO BINARY CPSK

2-Pole Butterworth Filter			3-Pole Butterworth Filter		
BER	C/I(dB)	C/N(dB)	BER	C/I(dB)	C/N(dB)
10^{-3}	∞	7.00	10^{-3}	∞	7.05
"	26	7.05	"	26	7.10
"	20	7.15	"	20	7.25
"	18	7.25	"	18	7.35
"	16	7.40	"	16	7.50
"	14	7.65	"	14	7.75
"	12	8.05	"	12	8.15
"	10	8.65	"	10	8.85
"	8	9.60	"	8	9.90
10^{-4}	∞	8.60	10^{-4}	∞	8.65
"	26	8.65	"	26	8.75
"	20	8.85	"	20	8.95
"	18	9.00	"	18	9.10
"	16	9.20	"	16	9.35
"	14	9.45	"	14	9.65
"	12	10.00	"	12	10.15
"	10	10.65	"	10	12.00
"	8	11.85	"	8	12.15
10^{-5}	∞	9.80	10^{-5}	∞	9.85
"	26	9.90	"	26	10.00
"	20	10.15	"	20	10.30
"	18	10.30	"	18	10.45
"	16	10.55	"	16	10.70
"	14	10.95	"	14	11.10
"	12	11.45	"	12	11.65
"	10	12.15	"	10	12.55
"	∞	N/A	"	8	N/A
10^{-6}	∞	10.75	10^{-6}	∞	10.85
"	26	10.85	"	26	10.95
"	20	11.15	"	20	11.30
"	18	11.35	"	18	11.50
"	16	11.65	"	16	11.75
"	14	12.00	"	14	12.25
"	12	12.65	"	12	12.85
"	10	N/A	"	10	N/A
"	8	N/A	"	8	N/A

Note: The analytical model assumes ideal product demodulation, lowpass-filtered phase sampling, and includes intersymbol interference effects due to bandpass filtering.

TABLE D-6

CPSK ERROR PERFORMANCE FOR NONWHITE INTERFERENCE PLUS THERMAL NOISE

$$BER = (\log_2 M)^{-1} \cdot (SER)$$

$$SER = \frac{1}{\pi} \int_o^{\pi} K_1 \text{ erfc } \sqrt{\gamma(\phi)}\, d\phi$$

$$= \frac{1}{\pi} \int_o^{\pi} K_1 \left[(2/\sqrt{\pi}) \int_{\sqrt{\gamma(\phi)}}^{\infty} \exp(-x^2) dx\right] d\phi$$

where $K_1 = 0.5$ (if $M = 2$) or 1.0 (if $M > 2$)

$$\gamma(\phi) = K_2 (C/N) \left[1 + \sqrt{\rho (\sin \pi/M)^{-2} (C/I)^{-1}} \; (\cos\phi)\right]^2$$

$$K_2 = (\sin \pi/M)^2 (G/L) (B_d T)$$

$$E/N_o = (B_d T) (C/N) = (\log_2 M) (B_d/B_r) (C/N)$$

and

B_d = desired signal bandwidth

B_r = bit rate

E = symbol energy

G = coding gain

L = implementation loss

M = number of phase states

N_o = noise power density

T = symbol duration

I' = ρI = truncated interferer power

$$\rho = \int_{B_d} K_i(f) df = \text{interferer spectral truncation factor}$$

<u>Upper Bound Formulation: $\cos \phi = -1$</u>

$$SER \leq K_1 \text{ erfc } \sqrt{\gamma} = K_1 \left[(2/\sqrt{\pi}) \int_{\gamma}^{\infty} \exp(-x^2)\, dx\right]$$

where $$\gamma = K_2 (C/N) \left[1 - \sqrt{\rho (\sin \pi/M)^{-2} (C/I)^{-1}}\,\right]^2$$

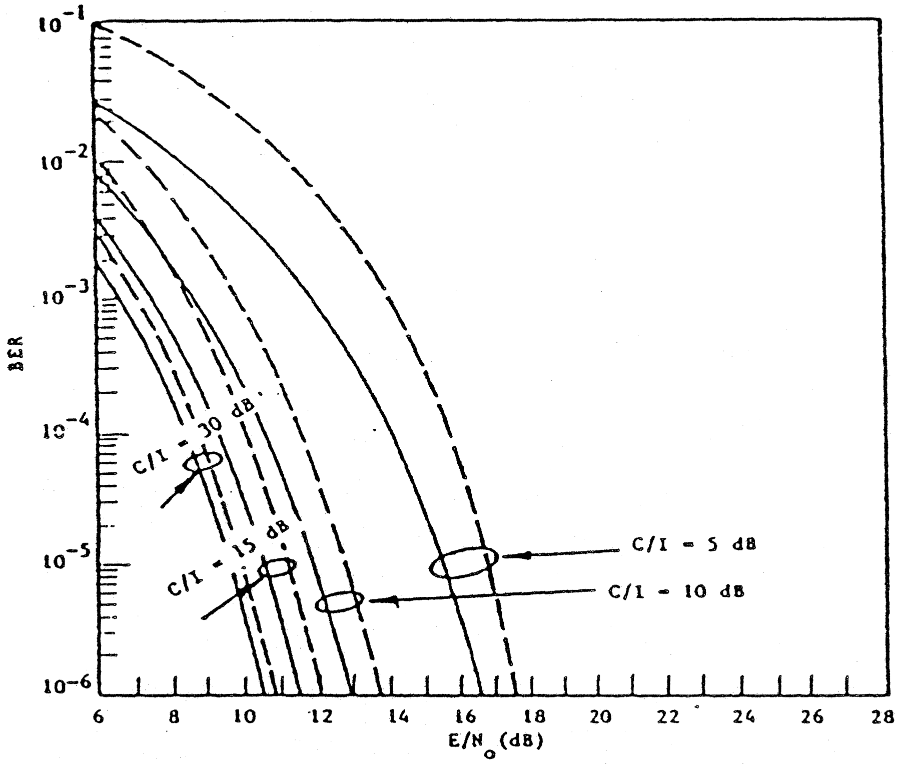


Figure D-1. Comparison of averaged (solid lines) versus upper bound (broken lines) performance for nonwhite interference into binary CPSK.

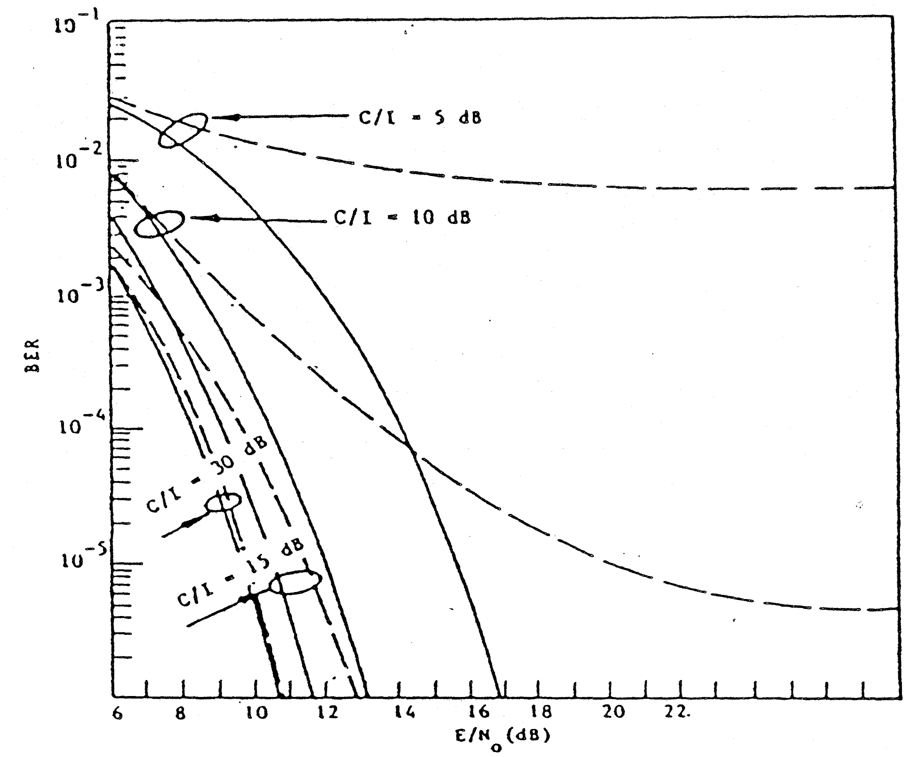


Figure D-2. Comparison of nonwhite (solid lines) versus white (broken lines) interference performance into binary CPSK.

The different behavior of the white and nonwhite interference performance as C/I decreases is a consequence of the distinct perturbation effects introduced in the state decision variable. The white interference case causes an independent random variation similar to that caused by thermal noise. The total variance corresponds to the algebraic sum of the noise and interference powers (N + I).

The nonwhite interference case is based on the cochannel sinusoidal model, where a sine wave of amplitude $\sqrt{I}$ and phase ϕ is added to the desired signal. The state decision variable has an amplitude $\sqrt{C} + \sqrt{I}\cos\phi$ with a random variation caused by the thermal noise. Hence, the interference affects the mean value while the noise affects the variance (unlike the white interference case).

If the interference has angle modulation (e.g., FM or PSK), the mean value varies with time and the performance expressions hold if $\cos\phi(t)$ is slowly varying relative to a symbol duration (T). If the interference has both amplitude (a) and angle (ϕ) modulation (e.g., SSB/AM), the mean value perturbation is $\sqrt{I} \cdot a \cdot \cos\phi$ and the performance expressions hold if $a \cdot \cos\phi$ is slowly varying relative to a symbol duration (T).

An offset sinusoidal interferer corresponds to $\phi = 2\pi F_o t$ and $2\pi F_o T \ll 1$ is needed for slowly varying conditions. The desired signal bandwidth (B_d) usually satisfies $B_d T \leq 2$ (i.e., at most one spectral lobe is preserved), so that $4\pi F_o \ll B_d$ suffices to validate the performance expressions and make the cochannel results representative.

The CCIR formulation on this matter is based on using the nonwhite interference case (based on the cochannel sinusoidal model) for angle-modulated interference and the white interference case for amplitude-modulated interference [CCIR Report 388-4, 1982]. On this basis, the nonwhite interference formulas (TABLE D-6) are used for FDM/FM, TV/FM, or PSK interferers, while the white interference formulas are used for CSSB/AM or multiple SCPC interferers.

The γ parameter in the formulas represents an equivalent output signal-to-noise ratio (SNR) that accounts for both interference and thermal noise effects and controls the BER performance via the erfc function. The value of γ is obtained from the input (C/N, C/I) values and the modulation parameters

(B_d, B_i, G, L, M, T), where the latter are combined into a single multiplier term (K_2). The white versus nonwhite interference nature selects the formula to be used for the γ evaluation (TABLE D-2 or D-6).

There remains the C/I threshold and margin formulations for a given BER performance specification. A BER = 10^{-6} requirement is representative of 8-bit PCM data and uncoded CPSK modulation, but it may be relaxed if analog or digital encoding techniques are employed (e.g., delta modulation, convolutional coding, soft-decision decoding).

A BER performance specification becomes an SER specification via the SER = $(\log_2 M)$ BER relation. This corresponds to a minimum γ value (output SNR threshold) to be provided via the erfc function inversion for both white and nonwhite interference cases, when the upper bound formula is used for the latter. This $\gamma(\min)$ value varies with M (via the BER to SER conversion) and is used to determine the corresponding C/I threshold according to the interference spectral nature, as shown below.

However, the critical step in this process is the erfc function inversion. A simple procedure that effectively approximates the inversion is summarized in TABLE D-7. It is based on the existence of relatively simple bounds for the erfc function as shown in the table [Wozencraft and Jacobs, 1965]. The upper bound is used in the inversion procedure to provide a threshold estimate on the conservative side. The inversion result is a simple formula for the $\gamma(\min)$ required to support a given BER performance.

The inversion procedure assumes $\log \gamma \ll (\log e)\,\gamma$ to replace a transcendental equation by a linear one that can be inverted. The assumption was validated via a round-trip computation, where the erfc values were computed exactly for γ = -3 to 30 dB and then inverted via the procedure shown in TABLE D-7 to recover the γ values. The recovered values were always conservative, as intended, and the inversion errors are shown in Figure D-3 to be small for practical purposes.

The C/I threshold and margin for any given BER performance specification are formulated in TABLE D-8. The inversion procedure yields the $\gamma(\min)$ value after the BER to SER conversion. The C/I threshold is then obtained using the $\gamma(\min)$, C/N, and modulation parameters, with a distinct formula employed

TABLE D-7

SRC ERFC-FUNCTION INVERSION PROCEDURE

Lower and Upper Bounds

$$\left(1 - \frac{1}{2\gamma}\right) \cdot \frac{1}{\sqrt{\pi\gamma}} e^{-\gamma} < \operatorname{erfc}\sqrt{\gamma} < \frac{1}{\sqrt{\pi\gamma}} e^{-\gamma}$$

SER Inversion Procedure: $SER = K_1 \operatorname{erfc}\sqrt{\gamma}$

$$SER = K_1 (\sqrt{\pi\gamma})^{-1} \exp(-\gamma) \qquad \text{(upper bound)}$$

$$\log(1/SER) = \log(\sqrt{\pi}/K_1) + \log\gamma + (\log e)\gamma$$

$$\approx \log(\sqrt{\pi}/K_1) + (\log e)\gamma$$

$$\gamma \cong (\log e)^{-1} [\log(1/SER) - \log(\sqrt{\pi}/K_1)]$$

Note: If $SER \geq 0.25$, then $\log(1/SER) = \log(1/0.25)$ is used in the last expression to avoid $\gamma < 0$ in non-dB units. The case $SER \geq 0.25$ represents an extreme undesirable error performance that will always render unacceptable C/I threshold margins, regardless of the exact value in the $0.25 \leq SER \leq 0.5$ range.

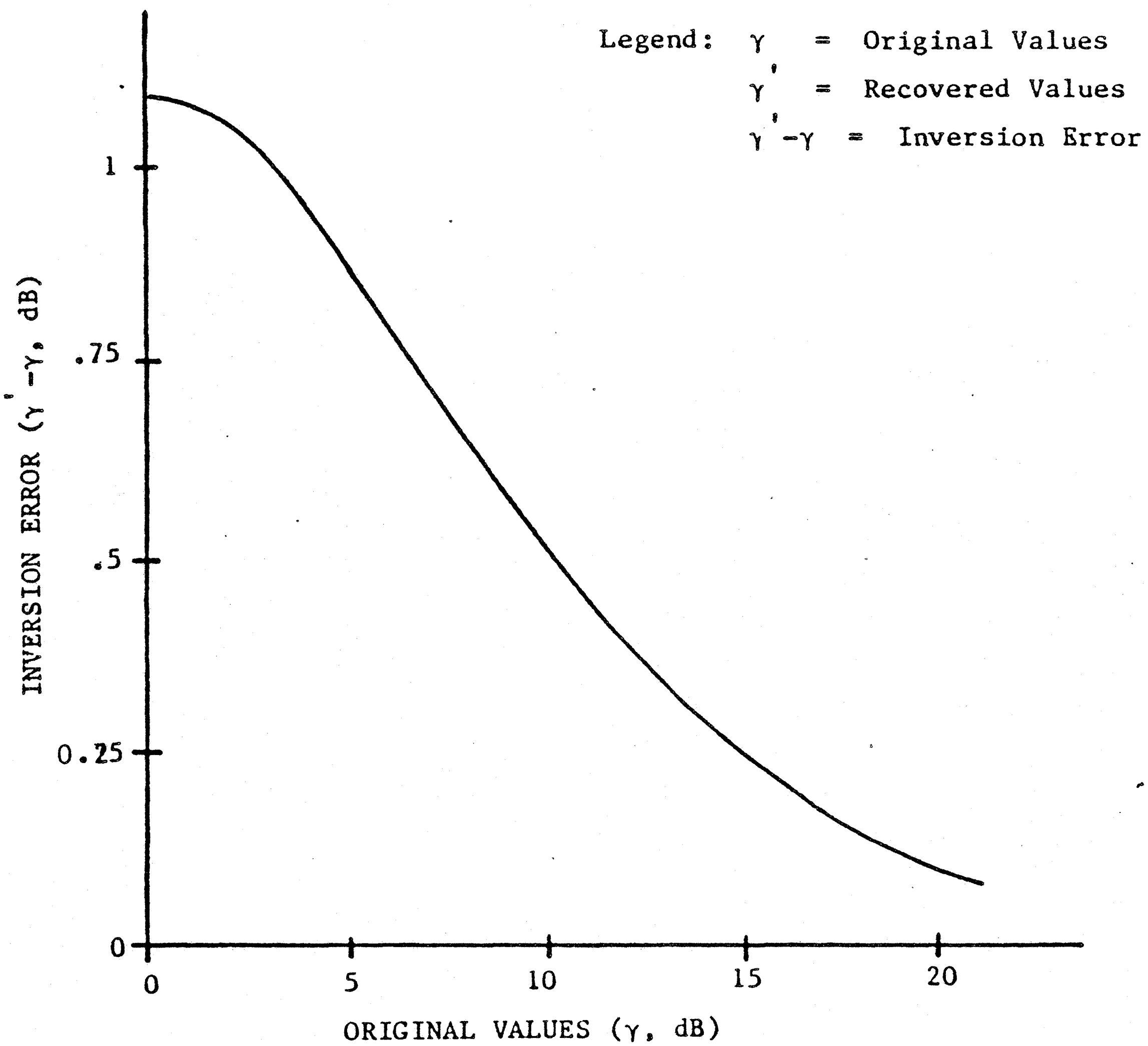


Figure D-3. Validation of erfc-function inversion procedure.

TABLE D-8

CPSK THRESHOLD AND MARGIN FORMULATION

BER = performance specification

SER = $(\log_2 M) \cdot \text{BER}$

$$\gamma = (\log e)^{-1} \quad [\log (1/\text{SER}) - \log (\sqrt{\pi}/K_1)] \text{ for SER} \leq 0.25$$

$$= (\log e)^{-1} \quad [\log (1/0.25) - \log (\sqrt{\pi}/K_1)] \text{ for SER} > 0.25$$

K_1 = 0.5 (if M = 2) or 1.0 (if M > 2)

K_2 = $(\sin \pi/M)^2 \, (B_d T) \, (G/L)$

Case of White Interference: $\rho = B_d/B_i$

$$(C/I)_{dB(threshold)} = 10 \log \frac{\rho \, (\gamma/K_2)}{1 - (\gamma/K_2) \, (C/N)^{-1}}$$

$$(C/I)_{dB(margin)} = (C/I)_{dB} - (C/I)_{dB(threshold)}$$

Case of Nonwhite Interference: $\rho = \int_{B_d} K_i(f)$

$$(C/I)_{dB(threshold)} \approx 10 \log \frac{\rho(\sin \pi/M)^{-2}}{\left(1 - \sqrt{(\gamma/K_2) \, (C/N)^{-1}}\right)^2}$$

$$(C/I)_{dB(margin)} = (C/I)_{dB} - (C/I)_{dB(threshold)}$$

Note: The upper bound BER formula is used for the nonwhite interference case (see TABLE D-6).

according to the interference spectral nature. The C/I margin is the difference between the C/I actual and threshold values.

The C/I threshold formulas in TABLE D-8 can be verified to have a logical functional dependence. If the BER requirement becomes tighter (i.e., the BER and SER values decrease), then the $\gamma(\min)$ and C/I threshold can be noted to increase. Also, if the BER remains fixed (i.e., the $\gamma(\min)$ value is fixed), but C/N decreases, the C/I threshold can be noted to increase. This behavior can be verified to hold for both the white and nonwhite interference cases.

In the white interference case, the C/I threshold formula in TABLE D-8 can be presented in the equivalent form shown in TABLE D-9, so as to permit a comparison with the CCIR formulation [CCIR Document 4/186-E, 1984]. The latter specifies the threshold (C/I') of the truncated interferer power ($I' = \rho I$), assuming a BER = 10^{-6} performance specification and a white interference model ($\rho = B_d/B_i$). The CCIR formula can be noted to represent a special case of the general formula provided in this report.

However, it should be emphasized that the approach presented in TABLE D-9 is based on a constant value assigned a priori to the fixed I'/N_{TOTAL} power ratio. This corresponds to a fixed value being assumed for the (C/I)/(C/N) ratio regardless of its actual value, as made evident by Equation D-1.

$$\frac{I'}{N_{TOTAL}} = \frac{I'}{I' + N} = \left[1 + \frac{N}{I'}\right]^{-1} = \left[1 + (B_i/B_d)\frac{(C/I)}{(C/N)}\right]^{-1} \qquad \text{(D-1)}$$

The use of a fixed value for I'/N_{TOTAL} is representative of the operational status only when the actual values of (C/N, C/I, B_d/B_i) reproduce the I'/N_{TOTAL} value assumed (-7, -8.2, or -14 dB). Otherwise, the C/I threshold computed from the I'/N_{TOTAL} value assumed will reflect an operational status that differs from the actual one, which can lead to rather conservative or liberal estimates of the C/I threshold and margin that actually prevail.

TABLE D-9

EQUIVALENT C/I THRESHOLD FORMULATION FOR WHITE INTERFERENCE

Equivalent BER Formulation

$$\text{BER} = (\log_2 M)^{-1} \cdot (\text{SER}) = (\log_2 M)^{-1} \cdot (K_1 \text{ erfc } \sqrt{\gamma})$$

where $\gamma = K_2(C/N_{TOTAL}) = K_2(C/I') \cdot (I'/N_{TOTAL})$

$$K_2 = (\sin \pi/M)^2 \, (B_d T) \, (G/L)$$

$$I' = \rho I = (B_d/B_i) I$$

Equivalent C/I Threshold Formulation

$$(C/I')_{dB} = 10 \log (\gamma/K_2) - (I'/N_{TOTAL})_{dB}$$

$$= 10 \log \gamma - 20 \log(\sin \pi/M) - 10 \log(B_d T) - (G/L)_{dB} - (I'/N_{TOTAL})_{dB}$$

$$(C/I)_{dB} = (C/I')_{dB} - 10 \log \rho = (C/I')_{dB} - 10 \log (B_i/B_d)$$

CCIR Formulation

$$(C/I')_{dB} = 10.8 - 20 \log (\sin \pi/M) + (L)_{dB} - (I'/N_{TOTAL})_{dB}$$

Note: $10 \log \gamma = 10.8$ for SER $= 10^{-6}$

$B_d T = 1$ and $G = 1$ assumed

$$L = \begin{cases} 3.0 + 0.7 \, (\log_2 M) \text{ for TDMA} \\ 2.5 + 0.5 \, (\log_2 M) \text{ for FDMA} \end{cases}$$

I'/N_{TOTAL} = -7 dB (20%), - 8.2 dB (15%) or -14 dB (4%) as per CCIR Recommendation 523-1, 1982.

For example, consider the case of $I'/N_{TOTAL} = -14$ dB. Equation D-1 then yields Equation D-2 below as an input status condition. Only those desired and interferer signals that satisfy Equation D-2 will reflect the actual operational status in the threshold calculation. The other signals will produce C/I threshold values in excess or defect, depending on whether the actual input C/I values are higher or lower than those provided by Equation D-2.

$$C/I\ (\mathrm{dB}) = C/N\ (\mathrm{dB}) + 13.8 - 10 \log (B_i/B_d) \qquad \text{(D-2)}$$

The basic issue is that the I'/N_{TOTAL} term emerged directly from $C/N_{TOTAL} = (C/I')(I'/N_{TOTAL})$ and cannot be treated as a constant of the scenario. All receivers would have to exhibit the same (C/I)/(C/N) ratio for all possible interferers to maintain the I'/N_{TOTAL} invariant as per Equation D-1. The actual value of I'/N_{TOTAL} varies with the (C/N, C/I) magnitudes encountered and can be easily computed from these input parameters, rather than assumed to be a scenario constant.

Moreover, the I'/N_{TOTAL} variations actually reflect how more or less interference (threshold) can be tolerated depending on how much thermal noise is present. The decision error margin (±180°/M) is to be shared by both disturbances compromising each other, and the interference threshold should be adapted according to the thermal noise magnitude. The I'/N_{TOTAL} variations provide such self-adjustment, since the actual values reflect the thermal noise changes.

The approach presented in TABLE D-8 can handle any C/N, C/I, ρ input values, regardless of whether or not they match a preset I'/N_{TOTAL} value, so that the actual operational status is being analyzed. This approach also accommodates any BER performance requirement and automatically computes the output SNR threshold (γ) from which the C/I threshold and margin are easily obtained. The approach presented in TABLE D-8 also distinguishes between white and nonwhite interference cases to compute the C/I threshold and margin accordingly.

The coding (G) and implementation loss (L) appearing in all the BER formulas are sometimes absent in the modulation specifications. The CCIR formulation in TABLE D-9 has assignment guidelines for the implementation loss and when not specified otherwise. The worst case is employed if no distinction between TDMA and FDMA is available in the modulation specifications.

The coding gain, if any, is a function of the input C/N value, since the coding purpose is to improve the BER performance established from the E/N_o value as per TABLE D-1. The coding gain also varies with the encoding and decoding procedures, as illustrated in Figure D-4 for various binary convolutional codes and decoding options [Sklar, 1983]. The dotted lines correspond to the BER performance including the coding effect and can be used to specify the coding gain versus E/N_o required to provide a given BER performance when coding is needed.

COMPACT AND DETAILED ALGORITHM COMPUTATIONS

The formulations in TABLES D-2 and D-6 are used to evaluate the BER performance, while TABLE D-8 is used to evaluate the C/I threshold margin. The upper bound in TABLE D-6 is used for the nonwhite interference case to maintain compatibility with its C/I threshold margin evaluation in TABLE D-8. A BER specification of 10^{-6} is assumed with an override capability provided for user selection of another BER requirement. The basic computation steps performed by a dedicated RTC algorithm are:

1. identification of the effective interference spectrum $K_i(f)$ as representing a white or nonwhite interference case

2. evaluation of the BER performance from the appropriate formulation (the generation and integration of the interference spectrum may be required to derive the spectral truncation factor ρ for the nonwhite interference case)

3. evaluation of the C/I threshold for the given BER performance specification.

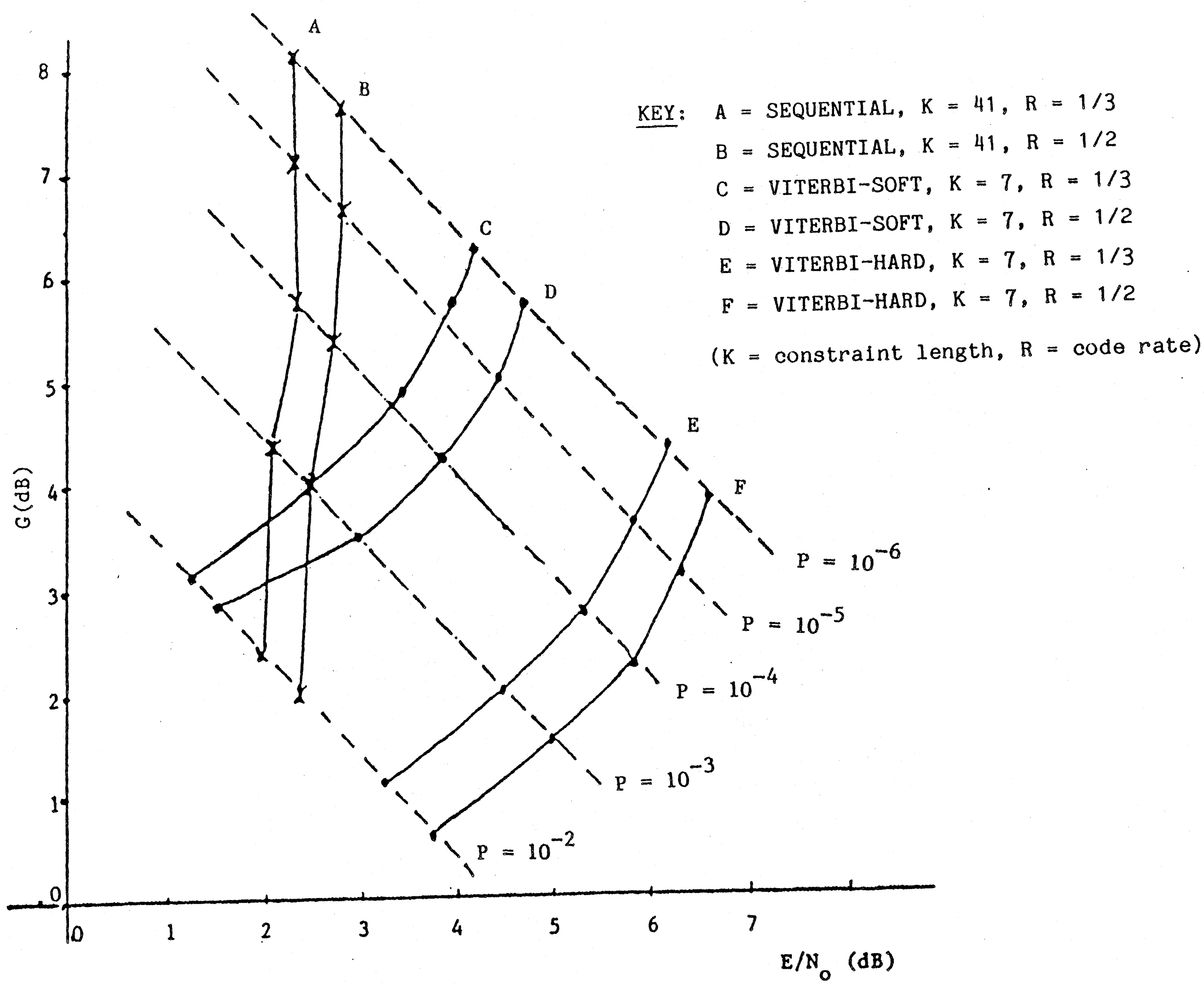


Figure D-4. Coding gain (G) versus output SNR (E/N_o) and BER performance (P) for convolutional codes and ideal binary CPSK.

The distinction between the compact and detailed algorithm versions is based on how simple it is to evaluate the interferer spectral truncation factor ρ. The compact versions are characterized by simple formulas or subroutines based on simple spectral models. The detailed versions are characterized by an elaborate spectral generation process followed by an integral subroutine that involves multiple sample processing.

Case of CCSB/AM Telephony Interferer: Compact and Detailed Algorithms

A CSSB/AM telephony interferer has a uniform spectrum with $K_i(f) = 1/B_i$ over the spectral overlap bandwidth. If $B_i \geq B_d$ and $F_o \leq 0.5\ (B_i - B_d)$, the overlap corresponds to the desired signal bandwidth and the white interference formula is applicable. If $B_i < B_d$, or if $B_i \geq B_d$, but $F_o > 0.5\ (B_i - B_d)$, the overlap is narrower and the white interference model becomes progressively conservative. The main limitation of using the white interference formula with a narrow interferer is that it yields a higher BER performance under low C/I conditions, as illustrated in Figure D-2.

A practical compromise would be to switch between the white and nonwhite interference models based on the amount of spectral overlap. The switch could be triggered automatically from the ratio B_o/B_d, where B_o is the overlap bandwidth given by Equation 4-1 in Section 4. However, there are no guidelines to select a switch value, so that the white interference model shall be used as a conservative estimate.

Case of FDM/FM Telephony Interferer: Compact and Detailed Algorithms

The nonwhite interference formulation is used. The gaussian formula of Equation 5-2 in Section 5 is used for the FDM/FM emission spectrum $S_i(f)$ in the compact algorithm. The spectral truncation factor ρ is obtained by integrating the shifted FDM/FM spectrum over the desired signal bandwidth, as shown in TABLE D-10. A simple gaussian integral subroutine that does not require multiple sample processing is employed.

TABLE D-10

SPECTRAL TRUNCATION FACTOR FOR GAUSSIAN INTERFERER SPECTRUM

$S_i(f) = (\sqrt{2\pi}\,\sigma)^{-1} \exp[-0.5(f/\sigma)^2]$ (interferer emission spectrum)

$\rho = \int_{-0.5B_D}^{0.5B_d} S_i(f - F_o)\, df = \int_L^U S_i(f) df$ (spectral truncation factor)

Case of $B_i \leq B_d$ and $F_o \leq 0.5\,(B_d - B_i)$: $\rho = 1$ (full interference)

Case of $B_i \leq B_d$ and $0.5\,(B_d - B_i) < F_o < 0.5\,(B_d + B_i)$

$L = -0.5\,B_i$, $U = 0.5\,B_d - F_o$

Case of $B_i > B_d$ and $F_o \leq 0.5\,(B_i - B_d)$

$L = -0.5\,B_d - F_o$, $U = 0.5\,B_d - F_o$

Case of $B_i > B_d$ and $0.5\,(B_i - B_d) < F_o < 0.5(B_d + B_i)$

$L = -0.5\,B_i$, $U = 0.5\,B_d - F_o$

Case of $F_o \geq 0.5\,(B_d + B_i)$: $\rho = 0$ (no interference)

Note: $\int_L^U S_i(f)df = J(L/\sigma) - J(U/\sigma)$, where $J(x) = \int_x^\infty (\sqrt{2\pi})^{-1} \exp\,(-y)^2 dy$

is a simple gaussian integral subroutine.

The detailed algorithm uses the FMSPC Program shown in Figure 5-3 (Section 5) to generate the FDM/FM emission spectrum. The latter is shifted and truncated to develop the effective interference spectrum $K_i(f)$. The spectral truncation factor ρ is obtained by integrating the $K_i(f)$ spectrum with an integral subroutine that requires multiple sample processing, as shown in Figure D-5.

Case of TV/FM Interferer: Compact and Detailed Algorithms

The nonwhite interference formulation is used. The compact algorithm uses the gaussian envelope for the TV/FM emission spectrum as discussed in Section 5. The spectral truncation factor is obtained using TABLE D-10. The detailed algorithm uses the TVSPC Program shown in Figure 5-7 (Section 5) to generate the TV/FM emission spectrum, and the spectral truncation factor is obtained as shown in Figure D-5.

Case of Wideband Digital PSK Interferer: Compact and Detailed Algorithms

The nonwhite interference formulation is applicable. The interferer emission spectrum $S_i(f)$ has the sinc-squared formula given by Equation 5-3 in Section 5, but it is approximated by its rectangular bound $S_i(f) = \alpha T$ in the compact algorithm to permit a simple integration. The spectral truncation factor is then given by $\rho = 1$ if $B_o = B_i$ and $\rho = (\alpha T)B_o$ if $B_o < B_i$, where B_o is the overlap bandwidth given by Equation 4-1 in Section 4. The detailed algorithm uses the exact sinc-squared spectrum as shown in Figure D-5.

Case of SCPC/PSK and SCPC/FM Interferers: Compact and Detailed Algorithms

The collective effect of multiple SCPC interferers is represented by a uniform power distribution over the desired signal bandwidth. The total interference power is given by $(kN')I$, where k is an activity factor and $N' = (B_d/B_i')$ is the number of interferers with allocated bandwidth B_i' that

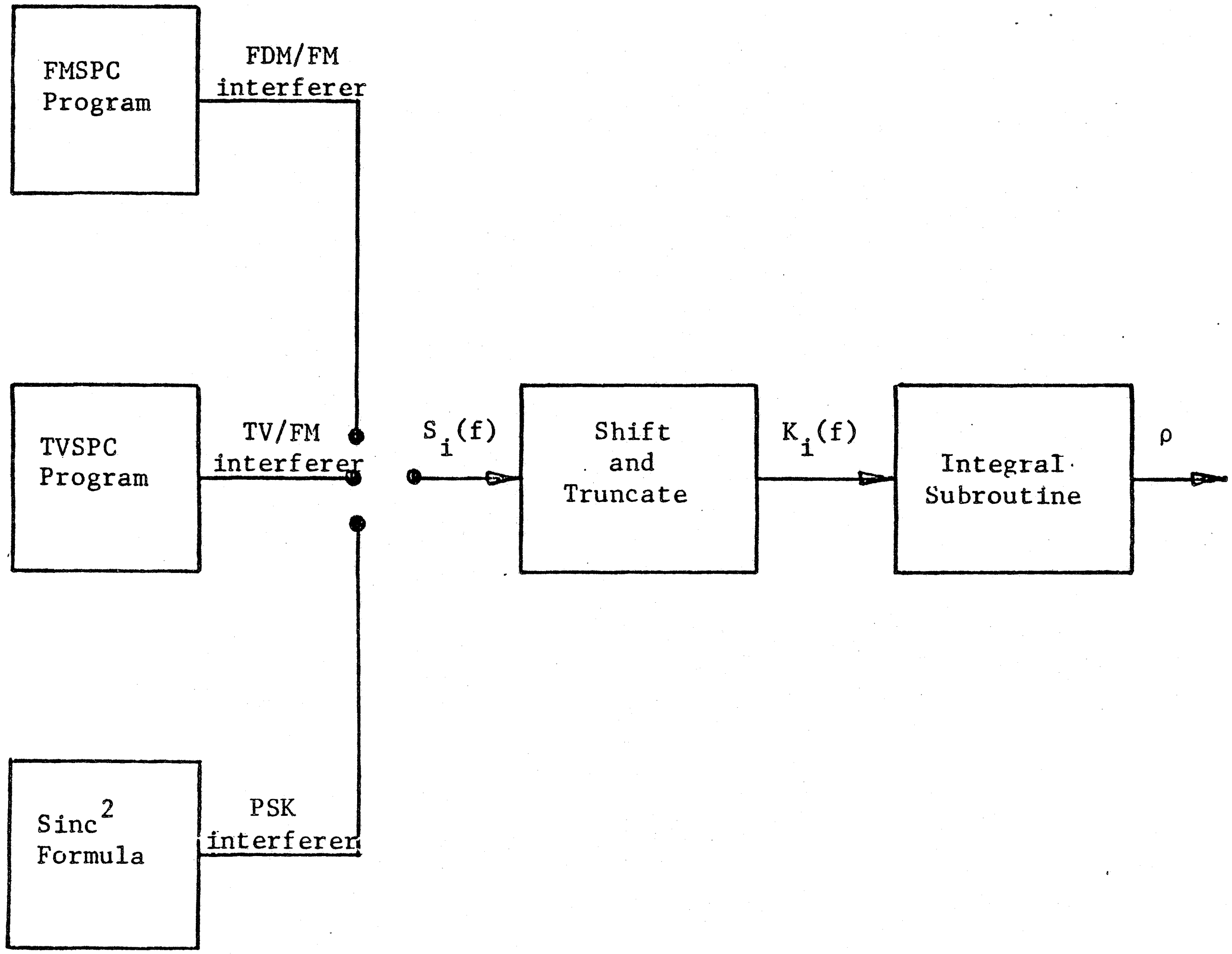


Figure D-5. Detailed algorithm evaluation of spectral truncation factor for nonwhite interference.

fits in the desired signal bandwidth B_d. The value $k = 0.4$ is assumed for the activity factor.

The compact and detailed algorithms are identical, and the white interference formulation is applicable with one important modification. The spectral truncation factor $\rho = B_i/B_d$ in TABLES D-2 and D-8 is now replaced by the value $\rho = kN'$ to reflect the presence of kN' interferers inside the desired bandwidth (i.e., the spectral power truncation is replaced by a spectral power enhancement to account for the SCPC multiplicity).

APPENDIX E

RTC ALGORITHMS FOR SCPC/PSK

SCPC/PSK PERFORMANCE FORMULATION

The PSK receiver demodulates the desired input signal to extract the digital data and perform the phase state decisions, as discussed in APPENDIX D. The general formulation presented there is applicable, with the exception of the TV/FM interferer case, as discussed below. The general formulation can also be simplified under wideband interference conditions, since a nonwhite interferer spectrum will appear to be white over the narrow desired bandwidth.

The effective interference spectrum $K_i(f)$ is essentially constant when $B_d << B_i$ and has a magnitude $K_i = S_i(F_o)$ obtained from the interferer emission spectrum $S_i(f)$ evaluated at the offset F_o. The white interference formulas in TABLES D-2 and D-8 are then applicable, except that the spectral truncation factor is given by $\rho = K_i B_d$. The nonwhite interference formulas of TABLES D-6 and D-8 are needed when the effective interference spectrum is not wideband relative to the desired signal. The basic computation steps performed by a dedicated RTC algorithm using these formulations are discussed in APPENDIX D. The distinction between the compact and detailed versions is based on how simple it is to evaluate the spectral truncation factor, including any spectral generation or integration required.

The TV/FM interferer case requires special attention due to the energy dispersal effects on the narrowband desired signal. A slowly dispersed carrier becomes a strong component that periodically sweeps the narrow SCPC spectrum with a duty cycle specified by the desired signal bandwidth and the peak dispersal deviation. The phenomenon has been compared to a pulsed interference occurrence, in which case the APPENDIX D formulations are not applicable.

A different formulation is required for the TV/FM interferer case. The white interference formulas in APPENDIX D can be used with $K_i = S_i(F_o)$ and $\rho = K_i B_d$ for a TV/FM interferer only if there is no dispersal, or if the latter has a peak-to-peak deviation smaller than the desired signal bandwidth, but these conditions are not representative with SCPC desired signals. The APPENDIX D formulations can be used for the other interferer types as discussed below.

Case of CSSB/AM Telephony Interferer: Compact and Detailed Algorithms

A CSSB/AM telephony interferer has a uniform spectrum with $K_i = 1/B_i$, so that the spectral truncation factor $\rho = B_d/B_i$ is used with the white interference formulation in TABLES D-2 and D-8.

Case of FDM/FM Telephony Interferer: Compact and Detailed Algorithms

The gaussian formula given by Equation 5-2 in Section 5 is used for the FDM/FM emission spectrum $S_i(f)$ in the compact algorithm. The effective interference spectrum is given by $K_i = (\sqrt{2\pi}\ \sigma)^{-1} \exp[-0.5(F_o/\sigma)^2]$, and the spectral truncation factor $\rho = K_iB_d$ is used with the white interference formulation in TABLES D-2 and D-8.

The detailed algorithm uses the FMSPC Program shown in Figure 5-3 (Section 5) to generate the FDM/FM emission spectrum $S_i(f)$. This spectrum is sampled at the offset frequency F_o to obtain its effective magnitude K_i and spectral truncation factor $\rho = K_iB_d$ used with the white interference formulation in TABLES D-2 and D-8.

Case of Wideband Digital PSK Interferer: Compact and Detailed Algorithms

The PSK interferer emission spectrum $S_i(f)$ has the sinc-squared formula of Equation 5-3 in Section 5. The effective interference spectrum is given by $K_i = \alpha T(\mathrm{sinc}\ \pi F_oT)^2B_d$ for $F_o \leq 0.5\ B_i$ ($K_i = 0$ otherwise), and the spectral truncation factor $\rho = K_iB_d$ is used with the white interference formulation in TABLES D-2 and D-8.

Case of SCPC/PSK and SCPC/FM Interferers: Compact and Detailed Algorithms

The number of SCPC/PSK (digital data) and SCPC/FM (analog voice) interferers that fits within the desired signal bandwidth B_d is $N' = B_d/B_i'$, where B_i' is the allocated bandwidth of one SCPC interferer. An activity factor (k) reduces this number to kN', and a value k = 0.4 is assumed.

If $N' > 1/k$, a multiple SCPC interferer condition occurs, and the white interference formulas in TABLES D-2 and D-8 are used with $\rho = kN'$ representing the SCPC multiplicity effect. If $N' \leq 1/k$, a single SCPC interferer condition occurs, and the nonwhite interference formulas in TABLES D-6 and D-8 are used. This one SCPC interferer case has $\rho = 1$ (no spectral truncation) if $N \geq 1$ where $N = B_d/B_i$. Otherwise, the spectral truncation factor is obtained by integrating the interference spectrum over the desired signal bandwidth, as discussed next.

The cases with one SCPC/PSK or SCPC/FM interferer have the simple sinc-squared and exponential formulas given by Equations 5-3 and 5-7 in Section 5 for their emission spectra. However, only the SCPC/FM exponential spectrum has a simple integration to provide identical compact and detailed algorithm versions. The SCPC/PSK sinc-squared spectrum is approximated by its rectangular envelope $S_i(f) = \alpha T$ in the compact version to permit a simple integration, while the exact characteristic is used along with an integral subroutine in the detailed version, as formulated in TABLE E-1.

Case of TV/FM Interferer: Compact and Detailed Algorithms

The white interference formulation in TABLES D-2 and D-8 is applicable only if $F_o \geq D_{DP} + 0.5\ B_d$ where F_o is the offset, B_d is the desired signal bandwidth and D_{DP} is the peak dispersal deviation. The interferer emission spectrum $S_i(f)$ is given by the gaussian envelope formula for the compact algorithm, and generated by the TVSPC Program shown in Figure 5-7 for the detailed algorithm (see Section 5). The spectral truncation factor $\rho = S_i(F_o)B_d$ is obtained from the interferer spectrum and used in the white interference formulas.

TABLE E-1

CASE OF SCPC/PSK OR SCPC/FM INTERFERERS

Case of $N' \leqq 1$ and $N < 1$: one truncated interferer

$$\int_{B_d} K_i(f)\, df = \begin{cases} \int_{B_d} \alpha T\,(\mathrm{sinc}\, \pi f T)^2 df \leqq (\alpha T) B_d & \text{for SCPC/PSK interferer} \\ 1-\exp[(-\sqrt{2}/\sigma)(0.5\, B_d)] & \text{for SCPC/FM interferer} \end{cases}$$

Case of $N' < 1$ and $N \geqq 1$: one complete interferer

$$\int_{B_d} K_i(f)\, df = 1$$

Case of $N' = 1$: one complete interferer

$$\int_{B_d} K_i(f)\, df = 1$$

Case of $N' > 1$: possibly many interferers

$$\int_{B_d} K_i(f)\, df = 1 \quad \text{if } N' \leqq 1/k \quad \text{(many interferers)}$$

$$\int_{B_d} K_i(f)\, df = kN' \quad \text{if } N' > 1/k \quad \text{(many interferers)}$$

Note: $k = 0.4$ will be assumed

The case of smaller offsets requires a different approach to handle the duty cycle effect. Equations E-1 and E-2 below represent a conservative estimate of the C/I threshold for a BER = 10^{-6} performance at C/N = 14 dB, when the SCPC/PSK signal is quaternary uncoded (M = 4, G = 1) and the TV/FM interferer has a dispersal duty cycle given by $\delta = B_d/2D_{DP}$ [CCIR Report 867, 1982].

$$(C/I)_{dB} = 27.5 + 6 \log \delta \text{ for frame-rate dispersal} \qquad \text{(E-1)}$$

$$(C/I)_{dB} = 27.5 + 10.5 \log \delta \text{ for line-rate dispersal} \qquad \text{(E-2)}$$

There are measurements available using a SCPC/QPSK signal with 64 kbps rate, 38 kHz bandwidth, and with or without 3/4-rate error-correction coding. The TV/FM interferer consisted of flat-field video plus dispersal modulating a carrier with peak-to-peak deviations of 4.75 MHz for video and 1 MHz for dispersal. The TV/FM spectrum was generated via computer simulation to identify its peak locations, and the symmetric SCPC/QPSK spectrum was centered at the largest peak as a worst condition [Yam, 1980].

A comparison between the C/I threshold measurements and the CCIR formulas is shown in Figures E-1 and E-2. The two thermal noise conditions represent the extreme cases of no noise and near-threshold noise, since C/N = 14.5 dB corresponds to a BER = 10^{-6} in uncoded SCPC/QPSK without interference. It is evident that the formulas become more conservative as the thermal noise decreases from its threshold value.

The C/I threshold differential between the two extreme noise cases in the measurements emphasizes the need to account for the C/N magnitude in the PSK error performance. The ±180°/M phase error margin is shared by the thermal noise and interference effects, with each restricting the amount allowed for the other. The C/I threshold decreases (increases) as the input C/N increases (decreases), since there is more (less) room for phase perturbations before the error margin is saturated.

The CCIR Equations E-1 and E-2 are used in the RTC algorithms corresponding to a TV/FM interferer with a duty cycle $\delta < 1$. However, further measurements are needed to establish the C/I threshold variations with intermediate C/N values and properly account for the C/I versus C/N compromise.

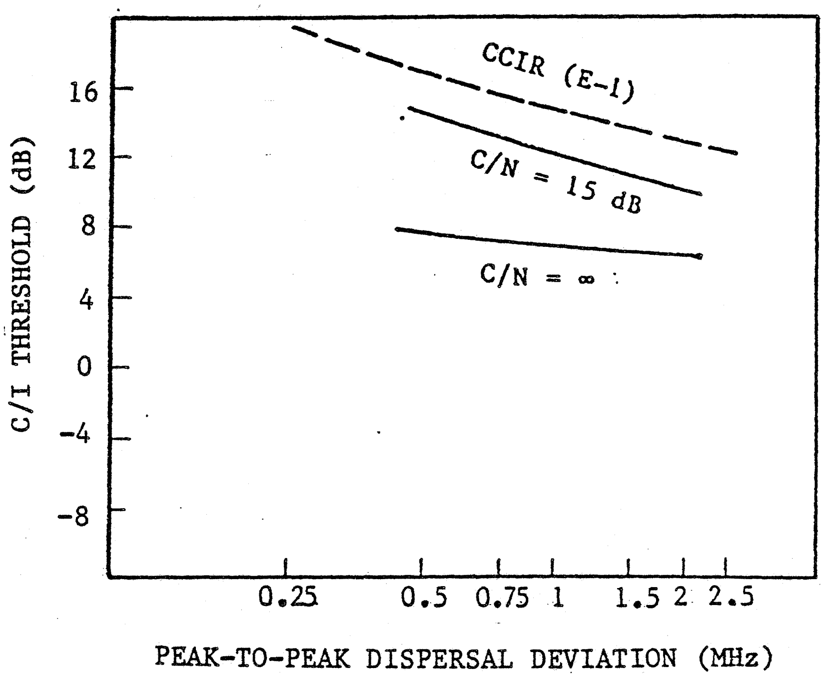


Figure E-1. Frame-rate dispersal: comparison of CCIR Equation E-1 with measurements for BER = 10^{-6}.

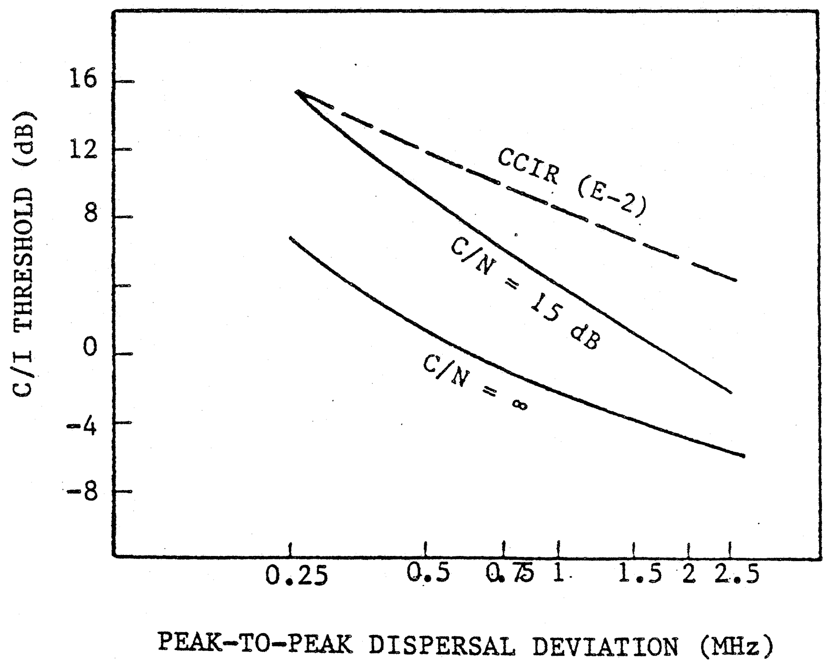


Figure E-2. Line-rate dispersal: comparison of CCIR Equation E-2 with measurements for BER = 10^{-6}.

The implementation loss can be assumed to be accounted for in the CCIR formulas, since they support the uncoded SCPC/QPSK measurements on the conservative side. However, the coding improvement (if any) must be subtracted from the formula values for general application. The coding gain is provided as a modulation parameter, but further investigation is needed if default values are to be assigned when not specified. The measurements produced coding gains of about 9 dB at C/N = 15 dB and 2 dB at C/N = ∞ for $\delta = 0.038$ and BER = 10^{-6}, so that the need to account for the C/N dependence is again emphasized.

The effect of the number of phase states (M) must also be included for generalization purposes. There are no guidelines available on this matter, though a 10 log $(\sin \pi/M)^2$ dependence can be proposed based on the nonpulsed interference results for BER = 10^{-6} (see TABLE D-9 in APPENDIX D). A 3 + 20 log(sin π/M) term is subtracted from the CCIR formula on this basis.

These extended CCIR formulas correspond to a BER = 10^{-6} performance specification. There are also measurements available for the C/I threshold at other BER requirements, both for frame-rate and line-rate dispersal [Yam, 1980]. However, there is no formulation available, and a simple extension of the BER = 10^{-6} formula is not possible due to the various parameters involved (BER, C/N, δ).

This experimental data needs further analysis to develop reliable empirical relations from the measurements available. The interest is to develop formulas for the BER performance as a function of the (C/N, C/I, δ) parameters for both dispersal modes. Once this is accomplished, the C/I threshold can follow from the formula inversion. However, an accurate formulation is hindered by the fact that only certain C/N values are empirically available. Moreover, the C/N = 15 dB no longer represents a near-threshold condition when BER $\neq 10^{-6}$ is considered.

In summary, the C/I threshold formulation used for TV/FM interference with $F_o > D_{DP} + 0.5\,B_d$ consists of the CCIR Equations E-1 and E-2, plus an extra term $-G - (3 + 20 \log \sin \pi/M)$ to account for the coding gain (G) and number of states (M). A BER = 10^{-6} performance specification is assumed in this formulation and the results will be conservative, since near-threshold C/N conditions are used in the CCIR equations.

APPENDIX F

RTC ALGORITHMS FOR SCPC/FM VOICE

SCPC/FM PERFORMANCE EVALUATION

The SCPC/FM performance evaluation requires a distinction between TV/FM interference and other cases, as happened with SCPC/PSK in APPENDIX E. The TV/FM carrier dispersal produces a slow interference component that periodically sweeps across the narrow SCPC/FM signal bandwidth with a certain duty cycle. Its effects are distinguished from those obtained with a continuous interference present in the desired signal bandwidth. The latter case is representative of the other interference types in question.

The SCPC/FM receiver extracts the baseband spectrum from the desired input signal via frequency demodulation and deemphasis. The total output power (P_d) in the desired baseband is given by $(2\pi D_{RMS})^2$, where D_{RMS} is the rms frequency deviation. However, the modulation specifications usually provide the peak frequency deviation D_{PK}, so that the peak factor $\lambda = D_{PK}/D_{RMS}$ is needed to formulate the desired output power. The value $\lambda = 2$ is used based on experimental measurements, as discussed in Section 5.

An interfering signal within the receiver input passband produces an interferer output component added to the desired output baseband. The interferer output spectrum $K_o(f)$ differs from its effective input spectrum $K_i(f)$ due to the FM demodulation process. The interferer output power (P_i) is obtained by integrating the spectrum $K_o(f)$ over the baseband bandwidth (B_b).

If the interferer output spectrum $K_o(f)$ is wideband relative to the baseband bandwidth, a uniform $K_o(f)$ spectrum can be assumed and the integration process is replaced by the K_oB_b product. This condition is satisfied whenever the interference input spectrum $K_i(f)$ is wideband relative to the baseband bandwidth ($B_i >> B_b$), since the output spectrum $K_o(f)$ is wider than the input spectrum $K_i(f)$.

The interferer output spectrum $K_o(f)$ has the form $(2\pi f)^2 D(f)R(f)$, where $D(f)$ is the deemphasis power transfer function and $R(f)$ is the resultant of a spectral convolution series. Under high C/I conditions, the series can be approximated by its predominant term and only one convolution between the desired $S_d(f)$ and interference $K_i(f)$ spectra is required. The notation and development is similar to that already presented for the FDM/FM and TV/FM cases, since an FM demodulation process is always involved (see APPENDICES B and C).

A conventional approach for SCPC/FM is to formulate the $K_o(f)$ spectrum without any deemphasis effect and scale the output power ratio by a composite gain factor (G_{PWC}) that jointly accounts for the preemphasis, noise-weighting and companding effects. There are measurements available to guide the assignment of gain values when they are not specified [Rogers et al, 1978]. The companding improvement measured under varied conditions is summarized below. There is little variation with talker level and a slight variation with the thermal noise density. Companding gains of 16.0 dB and 12.7 dB were found to correspond to the 50th and 90th percentile at C/N_o = 64 dB-Hz when all subjects were included.

Average Talker Level (25-28 subjects per case)	Companding Gain (Mean Values) C/N_o = 64 dB-Hz	C/N_o = 57 dB-Hz
Female, Soft (-33 dBm0)	16.6 dB	18.3 dB
Male, Soft (-27 dBm0)	16.3 dB	15.5 dB
Male, Average (-16 dBm0)	15.9 dB	15.0 dB
Male, Loud (-8 dBm0)	16.0 dB	15.8 dB

These companding gain results were obtained with VOX and preemphasis circuitry disconnected. The VOX-only gain (without companding) was only 4.1 dB for the two stronger talkers (the other two cases could not maintain the VOX activated). The VOX plus companding gain (without preemphasis) was 14.4 to 16.4 dB and the addition of preemphasis produced composite gains of 15.0 to 15.9 dB. These results indicate that the companding gain is the predominant contribution and a composite gain of 16.0 dB is assumed in the RTC algorithms when not specified.

The output power ratio with voice modulation in the desired signal is denoted by SNR. The output power ratio with test tone modulation instead of baseband voice is denoted by TNR. The SNR-to-TNR conversion is done via a loading factor P and the output performance measure is the interference output power in picowatts (pWOp) corresponding to a 0-dBm0 test tone.

The generalized SNR, TNR, and pWOp formulations are shown in TABLE F-1. If the interference spectrum $K_i(f)$ is wideband, the conventional FM formula for the output SNR in thermal noise can be used. The interference input power density is $I \cdot K_i$, where K_i is constant over the desired signal bandwidth. The output SNR expression in TABLE F-1 then simplifies, as shown in TABLE F-2.

COMPACT AND DETAILED ALGORITHM COMPUTATIONS

The general formulas in TABLE F-1 are applicable to all interferer types except TV/FM, with each type distinguished by its effective interference spectrum $K_i(f)$ producing a distinct convolution spectrum $K(f)$ and output spectrum $K_o(f)$. The white interference formulas in TABLE F-2 can be used when $B_i \gg B_d$ is satisfied. Otherwise, the basic computation steps performed by a dedicated RTC algorithm are:

1. generation of the desired $S_d(f)$ and interferer $K_i(f)$ input spectra (if needed) from the modulation specifications

2. generation of the convolution $K(f)$ and superposition $Q(f)$ spectra from the desired and interferer input spectra

3. integration of the output spectrum $K_o(f)$ (without deemphasis) over the baseband bandwidth B_b to determine the interferer output power P_i

4. evaluation of the pWOp performance and C/I threshold margin from the output SNR expression using simple conversion formulas.

TABLE F-1

SCPC/FM VOICE PERFORMANCE EVALUATION

$B_b = (f_H - f_L)$ (baseband bandwidth)

$P_d = (2\pi D_{RMS})^2 = (2\pi D_{PK}/\lambda)^2$ (desired output power)

$P_i = \int_{B_b} K_o(f)df$ (interferer output power)

where $K_o(f) = (2\pi f)^2 D(f)R(f)$ (interferer output spectrum)

$D(f)$ = deemphasis power transfer function

$R(f) \approx (C/I)^{-1} \cdot Q(f)$ if $C/I >> 1$

$Q(f) = 0.5\ [K(f) + K(-f)]$

$K(f) = S_d(f) * K_i(f)$ (* denotes convolution)

$$(SNR)_{dB} = 10 \log (P_d/P_i) + (G_{WC})_{dB}$$

$$= (C/I)_{dB} + 20 \log (D_{PK}/\lambda) - 10 \log [\int_{B_b} f^2 D(f)Q(f)df] + (G_{WC})_{dB}$$

$$= (C/I)_{dB} + 20 \log (D_{PK}/\lambda) - 10 \log [\int_{B_b} f^2 Q(f)df] + (G_{PWC})_{dB}$$

assuming $C/I >> 1$

$(TNR)_{dB} = (SNR)_{dB} - (P)_{dB}$ (test-tone modulation)

$pWOp = 10^{[90-(TNR)_{dB}]/10}$ (picowatts)

$$(C/I)_{dB(margin)} = 10 \log \frac{600}{pWOp} = 10 \log 600 - [90 - (TNR)_{dB}]$$

$$(C/I)_{dB\ (threshold)} = (C/I)_{dB} - (C/I)_{dB(margin)}$$

Notes: $\lambda = 2$ will be assumed

600 pWOp performance threshold assumed

TABLE F-2

OUTPUT SNR FORMULA FOR WHITE INTERFERENCE

$$P_d = (2\pi D_{PK}/\lambda)^2 \quad \text{where } \lambda = D_{PK}/D_{RMS}$$

$$P_i = \int_{B_b} (IK_i/C)\,(2\pi f)^2 df = (2\pi)^2 (IK_i/3C)\,(f_H)^3 (1-\varepsilon^3)$$

where $\varepsilon = f_L/f_H$

$$(SNR)_{dB} = 10 \log (P_d/P_i) + (G_{PWC})_{dB}$$

$$= (C/I)_{dB} + 10 \log [\,(3/\lambda^2)\,(D_{PK}/f_H)^2\,(K_i f_H)^{-1}\,(1-\varepsilon^3)^{-1}] + (G_{PWC})_{dB}$$

The distinction between the compact versus detailed algorithm versions is based on how simple it is to generate the spectral convolution K(f) and to integrate the output spectrum $K_o(f)$. The compact versions are characterized by a simple formula or fast subroutine for the integration process, which always requires a simple formula for the spectral generation and convolution. The detailed versions can involve elaborate spectral generation, convolution, or integration routines.

Case of CSSB/AM Telephony Interferer: Compact and Detailed Algorithm

A CSSB/AM telephony interferer has a wideband uniform spectrum with $K_i = 1/B_i$ over the desired signal bandwidth. The white interferer formulas in TABLE F-2 are applicable.

Case of FDM/FM Telephony Interferer: Compact and Detailed Algorithms

An FDM/FM telephony interferer also has a wideband spectrum, so that the formulas in TABLE F-2 are applicable. However, the value of K_i depends on the frequency offset between the desired and interferer signals, since the FDM/FM spectrum is not uniform.

The compact algorithm uses the gaussian formula in Equation 5-2 (Section 5) for the interferer emission spectrum $S_i(f)$. The value of K_i used in TABLE F-2 is $K_i = (\sqrt{2\pi}\ \sigma)^{-1} \exp[-0.5 (F_o/\sigma)^2]$, where σ is the rms frequency deviation of the FDM/FM interferer and F_o is the frequency offset.

The detailed algorithm uses the FMSPC Program shown in Figure 5-3 (Section 5) to generate the interferer emission spectrum $S_i(f)$. This spectrum is then sampled at the offset frequency F_o to obtain the K_i value used in TABLE F-2.

Case of Wideband Digital PSK Interferer: Compact and Detailed Algorithm

The formulas in TABLE F-2 are applicable. The PSK spectrum given by Equation 5-3 in Section 5 yields the value $K_i = (\alpha T)\ \mathrm{sinc}^2\ (\pi F_o T)$, where $F_o \leq 0.5\ B_i$, to be used in TABLE F-2.

Case of SCPC/PSK and SCPC/FM Interferers: Compact and Detailed Algorithms

The effective number of SCPC/PSK (digital data) or SCPC/FM (analog voice) interferers that fits in the desired signal bandwidth (B_d) is at most one if $N' \leqq 1/k$ or many (kN') if $N' > 1/k$, where $N' = B_d/B_i'$, B_i' is the allocated bandwidth of one SCPC interferer and k is an activity factor. A value of $k = 0.4$ is assumed.

If $N' > 1/k$ (multiple SCPC interferers), the white interference formulas in TABLE F-2 are used with $K_i = k/B_i'$. The total interference power is $(kN')I$ and it is uniformly distributed over the desired bandwidth B_d, which represents an equivalent power density $K_i I = (kN')I/B_d = (k/B_i')I$. The (C/I) term in Equation F-1 is computed using only one interferer for the I value, since the SCPC multiplicity is accounted for in the K_i term.

If $N' \leq 1/k$ (one SCPC interferer), the white interference formulas become questionable and the general formulas in TABLE F-1 are used. The spectral convolution $K(f)$ must be performed to obtain the output spectrum $K_o(f)$ to be integrated over the desired baseband B_b.

The desired spectrum $S_d(f)$ has the exponential characteristic of Equation 5-7 in Section 5, with standard deviation $\sigma = \sigma_1$ specified from the modulation parameters. The interferer emission spectrum $S_i(f)$ has an exponential characteristic with $\sigma = \sigma_2$ for the SCPC/FM interferer case, or the sinc-squared characteristic of Equation 5-3 in Section 5 for the SCPC/PSK interferer case.

The convolution of two exponential spectra has a simple expression to yield a simple formula for the interferer output power (P_i), so that the compact and detailed algorithms are identical in the SCPC/FM interferer case. However, the convolution of the exponential and sinc-squared spectra does not have a simple expression, so that the rectangular envelope $S_i(f) = \alpha T$ is used for the SCPC/PSK interferer spectrum in the compact algorithm. The sinc-squared spectrum is used in the detailed algorithm for the SCPC/PSK interferer, and the SPCVL Program shown in Figure B-7 (APPENDIX B) is employed to perform the spectral convolution, as summarized in Figure F-1.

The exponential-exponential (SCPC/FM interferer) and exponential-rectangular (SCPC/PSK interferer) spectral convolutions were formulated under cochannel conditions to derive the interferer output spectrum $K_o(f)$. The letter was then integrated over the desired baseband limits (f_L, f_H) to obtain the normalized interferer output power ($\hat{P}_i$) corresponding to C/I = 0 dB. The results are summarized in TABLES F-3 and F-4, and the output SNR expression in TABLE F-1 then has a simple formula for both cases.

The cochannel condition assumed represents a worst occurrence, since the desired and interferer in question come from a distribution of many SCPC signals in their respective transponders. The desired and interferer spectra have single central peaks, so that the cochannel condition yields a higher power content at low frequencies than the offset condition when the spectral convolution $K(f)$ is performed. The deemphasis characteristic $D(f)$ compensates

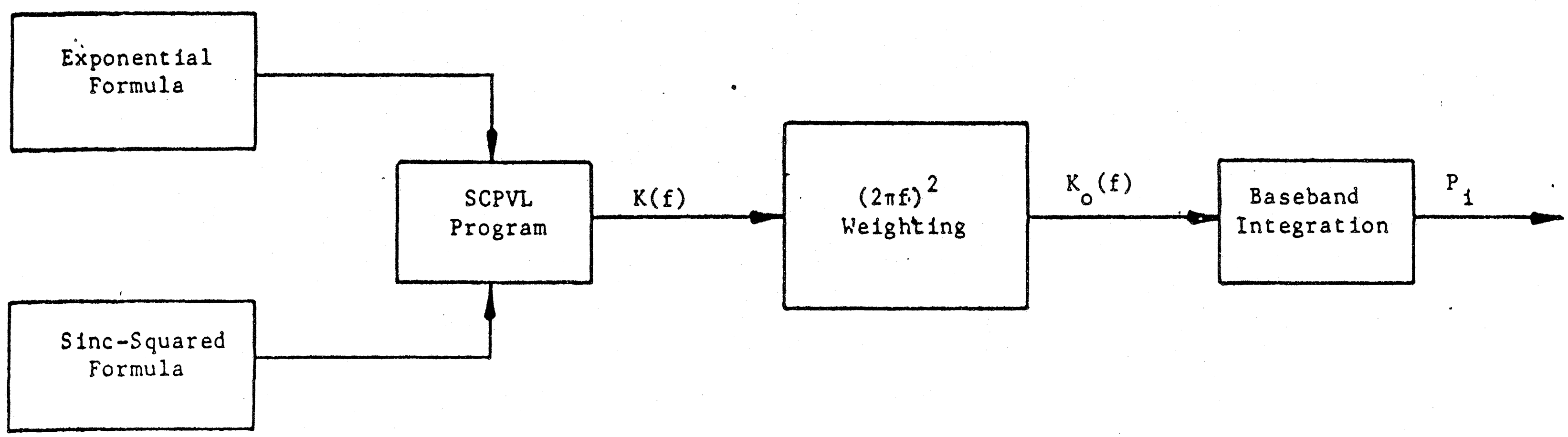


Figure F-1. Interferer output power for one cochannel SCPC interferer (detailed algorithm only).

TABLE F-3

INTERFERER OUTPUT POWER FOR ONE COCHANNEL SCPC/FM INTERFERER
(COMPACT AND DETAILED ALGORITHMS)

Case of $\sigma_1 \neq \sigma_2$

$$\hat{P}_i = \frac{(2\pi)^2}{2(\sigma_1^2 - \sigma_2^2)} \Big\{ \sigma_1^4 [(Y_{L1} - Y_{H1}) + 0.5 (Z_{L1} - Z_{H1}) \exp(-0.5 X_{D3})]$$
$$- \sigma_2^4 [(Y_{L2} - Y_{H2}) + 0.5 (Z_{L2} - Z_{H2}) \exp(-0.5 X_{D3})] \Big\}$$

where $\sigma_3 = \sigma_1 \sigma_2 / (\sigma_1 + \sigma_2)$

Case of $\sigma_1 = \sigma_2$

$$\hat{P}_i = \frac{(2\pi)^2 \sigma_1^2}{4} [(Y_{L1} - Y_{H1}) + (W_{L1} - W_{H1}) + (Z_{L1} - Z_{H1}) \exp(-X_{D1})]$$

Notation $X_{D1} = \sqrt{2}\, B_d/\sigma_1$ $X_{D3} = \sqrt{2}\, B_d/\sigma_3$

$X_{L1} = \sqrt{2}\, f_L/\sigma_1$ $X_{H1} = \sqrt{2}\, f_H/\sigma_1$

$X_{L2} = \sqrt{2}\, f_L/\sigma_2$ $X_{H2} = \sqrt{2}\, f_H/\sigma_2$

$W_{\#\#} = (x^3 + 3x^2 + 6x + 6) \exp(-x)$ evaluated at $x = X_{\#\#}$

$Y_{\#\#} = (x^2 + 2x + 2) \exp(-x)$ evaluated at $x = X_{\#\#}$

$Z_{\#\#} = (x^2 - 2x + 2) \exp(+x)$ evaluated at $x = X_{\#\#}$

where ## stands for D1, D3, H1, H2, L1, or L2, as needed.

TABLE F-4

INTERFERER OUTPUT POWER FOR ONE COCHANNEL SCPC/PSK INTERFERER
(COMPACT ALGORITHM ONLY)

Case of $B_d \leqq B_i$

$$\hat{P}_i = k_o \left\{ 1 - 0.5 \left[1 - 3X^{-1}(Z_L - Z_H) \right] \exp(-0.5\ X_D) \right\}$$

Case of $B_i < B_d \leqq 2\ B_i$ and $f_H \leqq 0.5\ (B_d - B_i)$

$$\hat{P}_i = k_o \left\{ 1 - 1.5\ X^{-1} \left[(Y_L - Y_H) - (Z_L - Z_H) \right] \exp(-\ 0.5\ X_D) \right\}$$

Case of $B_i < B_d \leqq 2\ B_i$ and $0.5\ (B_d - B_i) < f_H \leqq 0.5\ B_i$

$$\hat{P}_i = k_o \left\{ 1 - 0.5\ r_1 \exp(-\ 0.5\ X_D) - 1.5\ X^{-1} \left[(Y_L - Y_K) - (Z_L - Z_K) \right] \exp(-\ 0.5\ X_D) \right\}$$

Case of $B_i < B_d \leqq 2\ B_i$ and $f_H > 0.5\ B_i$

$$\hat{P}_i = k_o \left\{ 1 - r_2 - 0.5\ r_1 \exp(-\ 0.5\ X_D) + 1.5\ X^{-1}\ (Y_J - Y_K) \exp(+\ 0.5\ X_I) - 1.5\ X^{-1} \left[(Y_L - Y_K) - (Z_L - Z_K) \right] \exp(-\ 0.5\ X_I) \right\}$$

Case of $B_d > 2\ B_i$ and $f_H \leqq 0.5\ B_i$

$$\hat{P}_i = k_o \left\{ 1 - 1.5\ X^{-1} \left[(Y_L - Y_K) - (Z_L - Z_K) \right] \exp(-\ 0.5\ X_I) \right\}$$

(continued)

TABLE F-4 (continued)

Case of $B_d > 2\,B_i$ and $0.5\,B_i < f_H \leq 0.5\,(B_d - B_i)$

$$\hat{P}_i = k_o \left\{ 1 - r_2 + 1.5\,X^{-1}\,(Y_J - Y_H)\,\exp\,(+\,0.5\,X_I) - 1.5\,X^{-1}\,[(Y_L - Y_H) + (Z_J - Z_H)]\,\exp\,(-0.5\,X_I) \right\}$$

Case of $B_d > 2\,B_i$ and $f_H > 0.5\,(B_d - B_i)$

$$\hat{P}_i = k_o \left\{ 1 - r_2 - 0.5\,r_1\,\exp\,(-\,0.5\,X_D) + 1.5\,X^{-1}\,(Y_J - Y_H)\,\exp\,(+\,0.5\,X_I) - 1.5\,X^{-1}\,[(Y_L - Y_K) - (Z_L - Z_J)]\,\exp\,(-\,0.5\,X_I) \right\}$$

Notation: $k_o = (2\,\pi)^2\,(\alpha\,T/3)\,(f_H)^3\,(1 - \varepsilon^3)$

$r_1 = (1 - \varepsilon_1)/(1 - \varepsilon)$, $r_2 = (1 - \varepsilon_2)/(1 - \varepsilon)$

$\varepsilon = f_L/f_H$, $\varepsilon_1 = 0.5\,(B_d - B_i)/f_H$, $\varepsilon_2 = 0.5\,B_i/f_H$

$X = X_H{}^3\,(1 - \varepsilon^3)$, $X_L = \sqrt{2}\,f_L/\sigma_1$, $X_H = \sqrt{2}\,f_H/\sigma_1$

$X_D = \sqrt{2}\,B_d/\sigma_1$, $X_I = \sqrt{2}\,B_i/\sigma_1$, $X_J = \sqrt{2}\,B_i/2\sigma_1$,

$X_K = \sqrt{2}\,(B_d - B_i)/2\sigma_1$

$Y_\# = (x^2 + 2x + 2)\,\exp\,(-\,x)$ evaluated at $x = X_\#$

$Z_\# = (x^2 - 2x + 2)\,\exp\,(+\,x)$ evaluated at $x = X_\#$

where # stands for H, I, J, K, or L, as needed.

for the $(2\pi f)^2$ enhancement of high frequencies in the output interference spectrum $K_o(f)$, in which case the cochannel condition represents a worst occurrence.

Case of TV/FM Interferer: Compact and Detailed Algorithm

The formulation in TABLE F-2 is applicable with a TV/FM interferer when $F_o \geq D_{DP} + 0.5\ B_d$, where F_o is the offset, B_d is the desired signal bandwidth and D_{DP} is the peak dispersal deviation of the TV/FM carrier. The compact and detailed algorithms are then similar to the FDM/FM interference case, except that the TV/FM frequency deviation is used for the gaussian spectrum in the compact algorithm and the TVSPC Program shown in Figure 5-7 is used to generate the TV/FM spectrum in the detailed algorithm (see Section 5).

The case of a TV/FM interferer with smaller offsets requires a different approach to handle the duty cycle effect. The empirical Equation F-1 below has been obtained for the C/I threshold, based on subjective measurements of the output speech quality. The SCPC/FM modem was operated at C/N = 12 dB input and TNR = 50 dB output, including an 18 dB companding gain. A frame-rate dispersal waveform with duty cycle $\delta = B_d/2\ D_{DP}$ was the only modulation applied to the interfering carrier [CCIR Report 867, 1982].

$$(C/I)_{dB} = 26 + 8 \log \delta \quad \text{for frame-rate dispersal} \qquad \text{(F-1)}$$

The empirical Equations F-2 and F-3 below have also been obtained as conservative estimates for the C/I threshold, based on impulse-count measurements according to CCITT Recommendation M1020. The SCPC/FM modem was operated as described above, and the dispersal waveforms were the only modulation applied to the interfering carrier in each case [CCIR Document 4/194-E, 1984].

$$(C/I)_{dB} = 24 + 9 \log \delta \quad \text{for frame-rate dispersal} \qquad (F-2)$$

$$(C/I)_{dB} = 19 + 10 \log \delta \quad \text{for line-rate dispersal} \qquad (F-3)$$

The relation between the subjective speech quality and the output impulse count has not been established. The speech quality Equation F-1 always yields a higher C/I threshold than the impulse count Equation F-2. The dB differential is 2-log δ and is always positive, with a minimum of 2 dB at $\delta = 1$ and increasing by 1 dB as δ decreases by a factor of ten. The values $B_d = 20$ to 40 kHz for SCPC/FM and $2D_{DP} = 1$ to 4 MHz for TV/FM are representative and induce a range of $\delta = 5 \times 10^{-3}$ to 4×10^{-2}, which results in a differential of 3.4 to 4.3 dB between the two formulas.

In summary, the Equations F-2 and F-3 are used for the C/I threshold in the RTC algorithms, when $F_o < D_{DP} + 0.5\ B_d$, since they account for both frame-rate and line-rate dispersal. The Equation F-1 for frame-rate dispersal is more conservative with higher C/I thresholds, but it does not have a line-rate counterpart at this stage.

APPENDIX G

REVIEW OF INTERFERENCE FORMULAS FOR TV/FM SIGNALS

PROTECTION RATIO CRITERION

A performance criterion used to evaluate interference effects on TV/FM signals is the protection ratio. It consists of the ratio of the desired-to-interfering signal power at the receiver input that is required to provide a certain picture quality based on a subjective assessment. The protection ratio also varies with other signal characteristics, such as the television standard, frequency deviation, energy dispersal, preemphasis, carrier offset, etc.

Cochannel TV/FM Interferer

The protection ratio (PR) for the case of two TV/FM signals with identical modulation specifications under cochannel conditions has been formulated as follows [CCIR Report 634-2, 1982]:

$$PR(dB) = V - 20 \log (D/12) - Q + 1.1\, Q^2 \qquad \text{(G-1)}$$

where V is a constant of the television standard (V = 12.5 for 625-line I/PAL or G/PAL or L/SECAM, V = 18.5 for 625-line K/SECAM, V = 13.5 for 525-line M/NTSC), D is the peak-to-peak frequency deviation in MHz, and Q is the quality grade parameter on a 5-point scale. This formula has been noted to match experimental results for high Q values representing good picture quality, but to become inaccurate at low Q values representing picture degradation [Groumpos and Vernon, 1981; Whyte, 1983].

Some measurements with cochannel TV/FM interference were recently performed to establish the quality grade dependence on the input C/I ratio [Groumpos and Vernon, 1981]. Both signals were M/NTSC standards with D = 12 MHz, no energy dispersal, and CCIR preemphasis. The receiver thermal noise produced a video output S/N = 42 dB in the absence of interference. The measurements included different static pictures in the desired signal, while the interferer contained dynamic program material in all cases.

A comparison with the PR formula indicates that the latter is somewhat conservative for grades $Q \leq 3$. Also, the formula yields $Q = 4.5$ for $C/I = 31.3$ dB, while the measurements match or exceed such a grade for $C/I = 30$ dB, except for the worst case of color bars in the desired picture pattern. The formula reaches the ideal $Q = 5$ grade at $C/I = 36$ dB, while the measurements remain above $Q = 4.5$ but below $Q = 5$ even at higher C/I ratios (with the color bars exception). However, the experimental standard deviation ranged from 0.67 for $C/I = 5$ dB to 0.45 for $C/I = \infty$, which accommodates a 0.5 margin in the quality grades.

Another set of recent protection ratio measurements is available for M/NTSC standards and two distinct viewer population groups [Bouchard and Chouinard, 1983]. The signals employed had peak-to-peak deviations of 9.52 MHz for video and 2 MHz for two audio subcarriers, no energy dispersal, and CCIR preemphasis. The receiver thermal noise produced a video output $S/N = 42$ dB (weighted) for an input $C/N = 12$ dB in the absence of interference.

In general, the PR formula again represents higher C/I requirements than those measured. For example, C/I values of 19.8 to 27.4 dB for one group and 23.4 to 27.3 dB for the other were measured for $Q = 4.5$, while the formula yields 31.3 dB. The measurements also produced $Q > 4.7$ in all cases for $C/I = 30$ dB, so that the formula again appears to be a conservative bound for high-quality video performance.

The discussion has so far assumed identical modulation specifications for the desired and interfering signals. Some limited measurements are also available for the cochannel interference having a different frequency deviation than the desired signal [CCIR Report 634-2, 1982]. The unequal deviation results are within 0.5 dB of those obtained for the equal deviation case, assuming the desired signal deviation for both signals. These results indicate that the PR formula can also be used for the unequal deviation case, with the D parameter corresponding to the frequency deviation of the desired signal.

Some modulation specifications have been noted to have a secondary effect in the cochannel protection ratio performance [CCIR Report 634-2, 1982]. The use of none, one, or two audio subcarriers, or the presence or absence of preemphasis, did not affect the results obtained significantly. However, the use of dynamic program material instead of static picture patterns has been noted to relax the cochannel protection ratio by as much as 5 to 10 dB. Also, the use of preemphasis has been noted to affect the adjacent-channel protection ratio (unlike the cochannel case).

The effects of the input C/N and the output S/N are next considered. The table below summarizes recent measurements Q(meas) obtained at C/I = 30 dB under various conditions [Goldberg and Jones, 1983]. The measured values show a differential improvement of about 0.5 to 0.7 in the $4 < Q < 5$ range corresponding to a C/N or S/N reduction of 9 dB. The theoretical values Q(theor) based on the PR formula with C/I = 30 dB are noted to lie within the measured range for both deviation cases considered.

D(MHz)	C/I(dB)	C/N(dB)	S/N(dB)	Q(meas)	Q(theor)
12	30	14.8	46	4.05	4.35
12	30	23.8	55	4.72	4.35
16	30	12.3	46	4.13	4.65
16	30	21.3	55	4.64	4.65

It has been suggested that the cochannel protection ratio be reduced by 49 - (S/N) for S/N < 49 dB to account for the masking of interference by thermal noise effects [CCIR Report 634-2, 982]. However, the same report also cites experimental measurements where S/N variations in the 42 to 50 dB range did not produce any interference masking at a Q = 4.5 grade quality. It is evident that further investigation is needed to reach a consensus regarding the S/N bound corresponding to interference masking effects.

Adjacent-Channel TV/FM Interferer

A protection ratio formulation for adjacent-channel interference is more difficult to establish. There exist empirical curves relating the protection ratio to the carrier frequency offset, but only for specific deviation cases and with a considerable variation in the subjective criteria employed to evaluate picture quality [CCIR Report 634-2, 1982]. These limitations hinder a generalization of the data to account for the various modulation parameters under adjacent-channel conditions. An unequal deviation in the desired and interfering signals has been noted to substantially alter the protection ratio relative to the equal deviation case, which did not happen under cochannel interference conditions [CCIR Report 634-2, 1982].

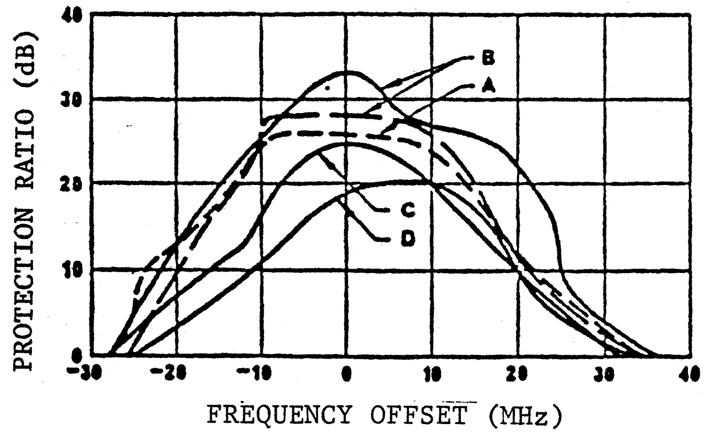
A set of subjective measurements with M/NTSC standard and D = 18 MHz in both signals produced the upper and lower bounds in the protection ratio shown in Figure G-1 [CCIR Report 634-2, 1982] The actual curves were contained within these bounds, with their relative location varying with the desired picture content involved. The picture quality criterion was the visual emergence of perceptible interference, and the cochannel PR formula can be noted to reproduce the origin point in the upper bound when $Q \approx 5$ is assumed for the impairment grade.

Other sets of subjective measurements are shown in Figure G-2 for two distinct television standards [CCIR Report 634-2, 1982]. The first set relied on a modified quality grade criterion, while the second set employed a simple binary (good-bad) picture quality criterion. The origin points match the cochannel PR formula in the first figure for $4.5 < Q < 5$ and in the second figure for $4 < Q < 4.5$.

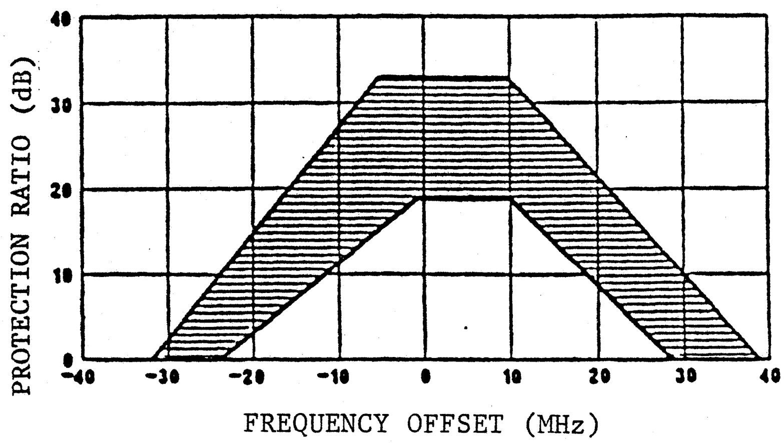
An empirical formula has been suggested for the protection ratio as a function of the carrier frequency offset and unequal desired and interference frequency deviations [Jeruchim, 1977]. The formula consists of four terms selected to fit experimental data as follows:

$$PR(DB) = K - 20 \log M_1 - F\ (M_1^{-0.85})$$

$$- 0.475\ (M^{-2.5})\ (F^{0.645M})\ \log M \qquad \text{(G-2)}$$

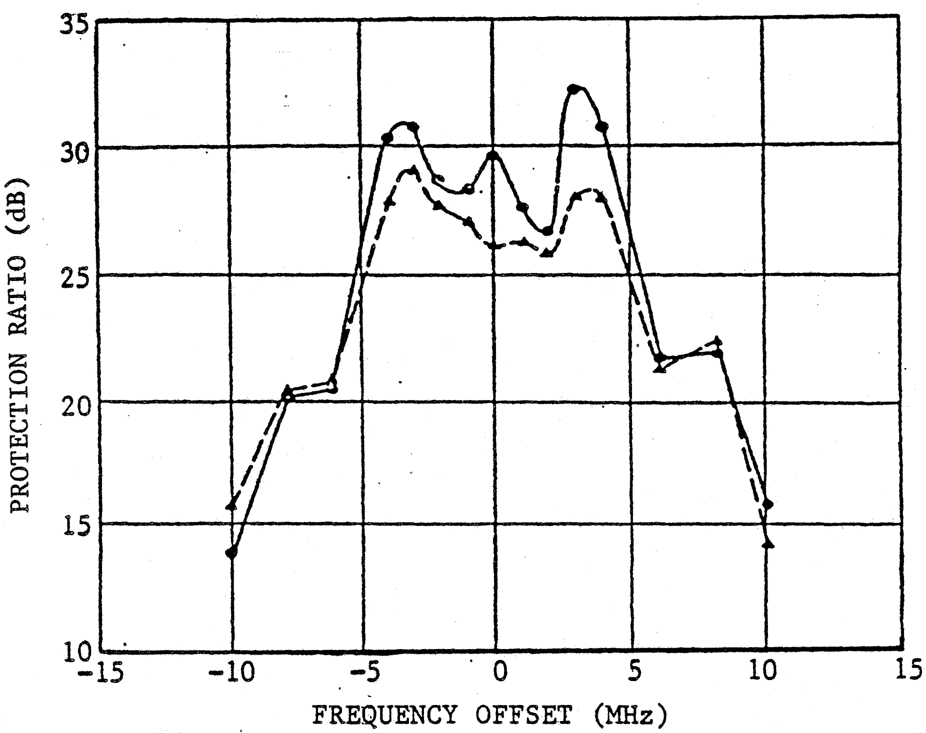


(a) Actual Measurements

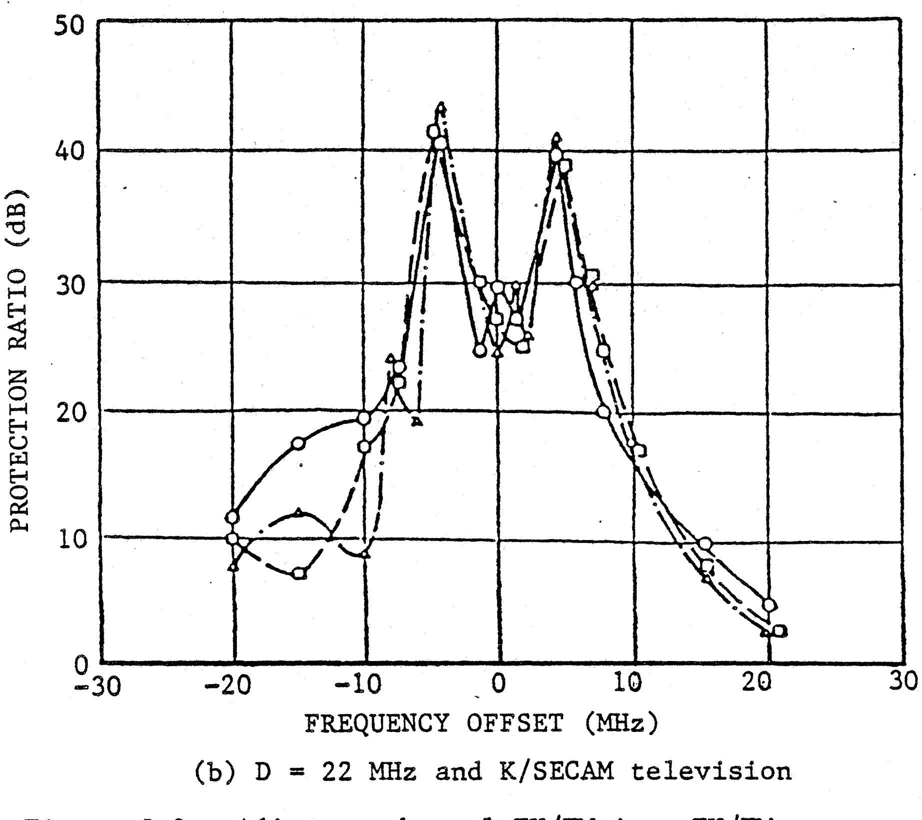


(b) Upper and Lower Bounds

Figure G-1. Adjacent channel TV/FM into TV/FM measurements for D = 18 MHz and M/NTSC television [CCIR Report 634-2, 1982].



(a) D = 8.4 MHz and M/NTSC television



(b) D = 22 MHz and K/SECAM television

Figure G-2. Adjacent channel TV/FM into TV/FM measurements [CCIR Report 634-2, 1982].

where K is a constant, F is the carrier frequency offset, $M = M_1/M_2$, $M_1 = D_1/8$ and $M_2 = D_2/8$ are the modulation indices corresponding to the peak-to-peak frequency deviations of the desired (D_1) and interfering (D_2) signals assuming a 4 MHz baseband bandwidth.

The use of $F = 0$ and $D_1 = D_2$ in Equation G-2 yields the relation $PR(dB) = K - 20 \log (D/8)$ for the cochannel protection ratio under equal deviations. A comparison with Equation G-1 shows that $K = V + 20 \log (12/8) - Q + (1.1)Q^2$ is needed to match both formulas. The value K = 29.5 was originally proposed, and it corresponds to Q = 3.85 for V = 13.5 (M/NTSC) and Q = 3.10 for V = 18.5 (K/SECAM). The use of $31 < K < 40$ for M/NTSC and $36 < K < 45$ for K-SECAM may be preferable, since it corresponds to $4 < Q < 5$.

The use of $F \neq 0$ and $D_1 = D_2$ in Equation G-2 yields the adjacent-channel protection ratio under equal deviations. It corresponds to a linear reduction from the cochannel value, since the third term in the formula is proportional to the offset F and the fourth term is zero for M = 1. This linear reduction starts immediately at the origin, so that the cochannel protection ratio will always exceed the adjacent-channel protection ratio.

The cochannel value and the decay slope provided by Equation G-2 vary with the frequency deviation, as shown in Figure G-3. A higher (lower) cochannel protection ratio is accompanied by a faster (slower) reduction in the adjacent-channel protection ratio. The K = 29.5 value proposed for M/NTSC can be noted to be deficient to support the measurements shown in Figure G-1, with a K = 45 value actually needed to cover the upper bound, as shown in Figure G-4.

Equation G-2 is also limited in its capability to accommodate the peak overshoots shown in Figure G-2, as a consequence of the linear reduction starting immediately at the cochannel value. If the cochannel protection ratio is matched, as shown in Figure G-5, then Equation G-2 does not support the overshoots. A higher value of K can be used to shift the formula upwards, as shown by the broken line in Figure G-5. However, the cochannel protection ratio would then become conservative (e.g., an extra 6 dB for M/NTSC and 16 dB for K/SECAM).

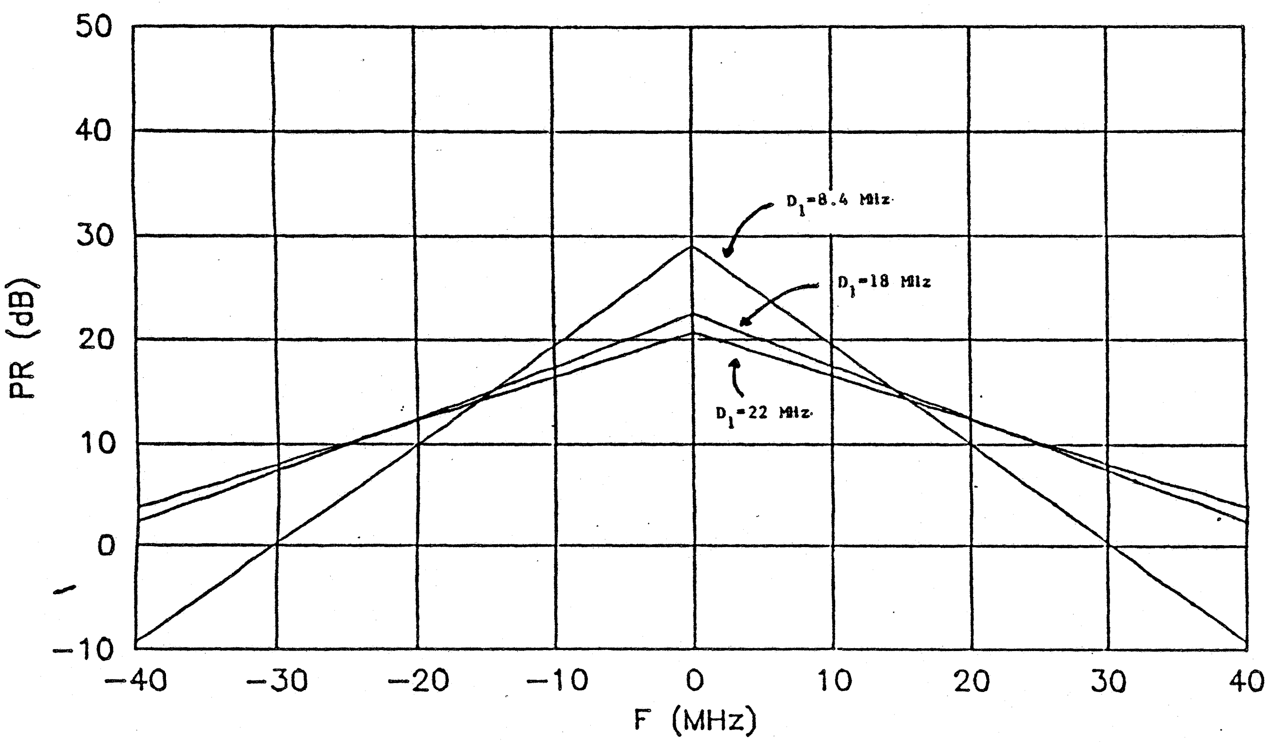


Figure G-3. Equation G-2 versus frequency offset (F) for K = 29.5 and $D_1 = D_2$.

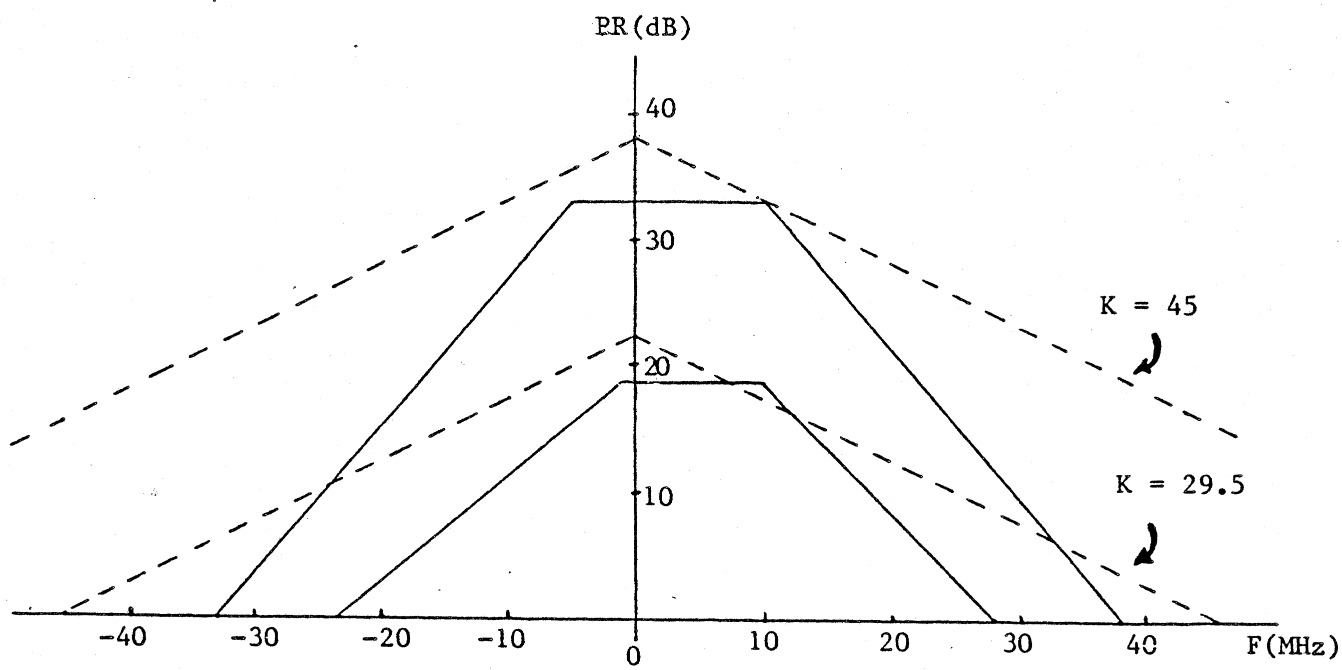
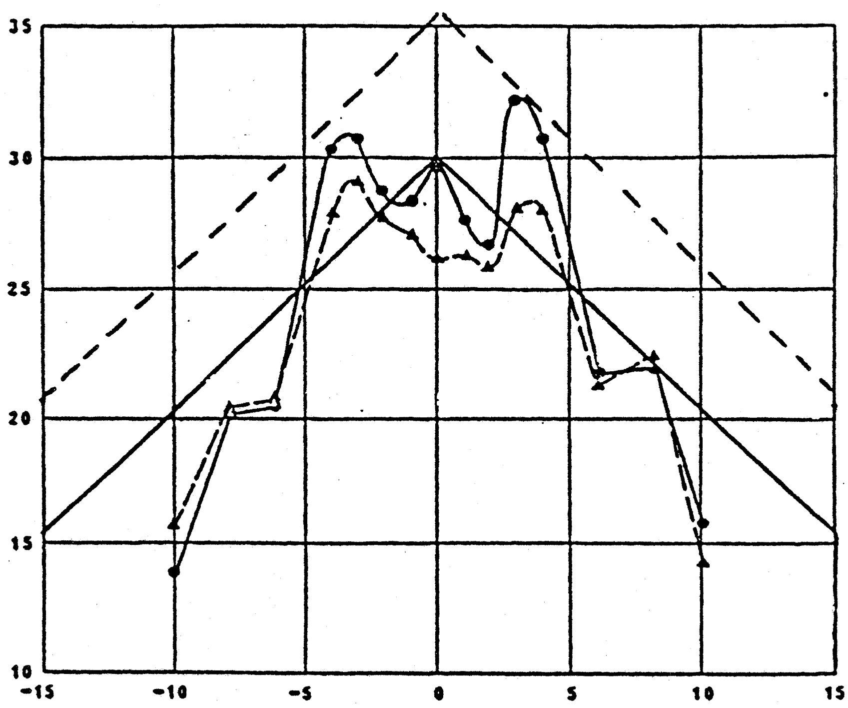
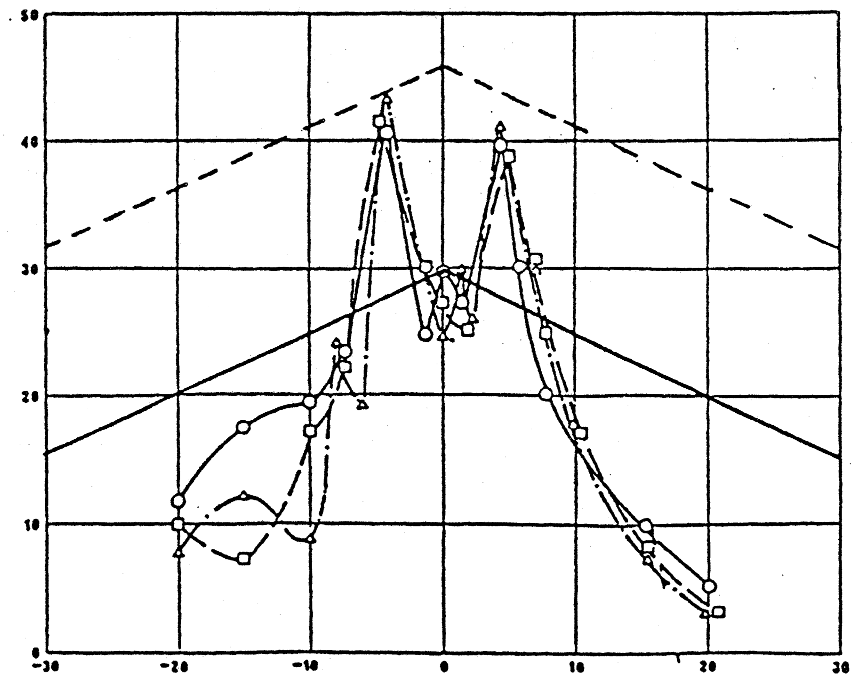


Figure G-4. Comparison of Equation G-2 with measurements in Figure G-1.



FREQUENCY OFFSET (MHz)

(a) K = 30.4 for 30 dB cochannel protection with $D_1 = D_2 = 8.4$ MHz and M/NTSC television



FREQUENCY OFFSET (MHz)

(b) K = 38.8 for 30 dB cochannel protectior with $D_1 = D_2 = 22$ MHz and K/SECAM television

Figure G-5. Comparison of Equation G-2 with experimental measurements in Figure G-2.

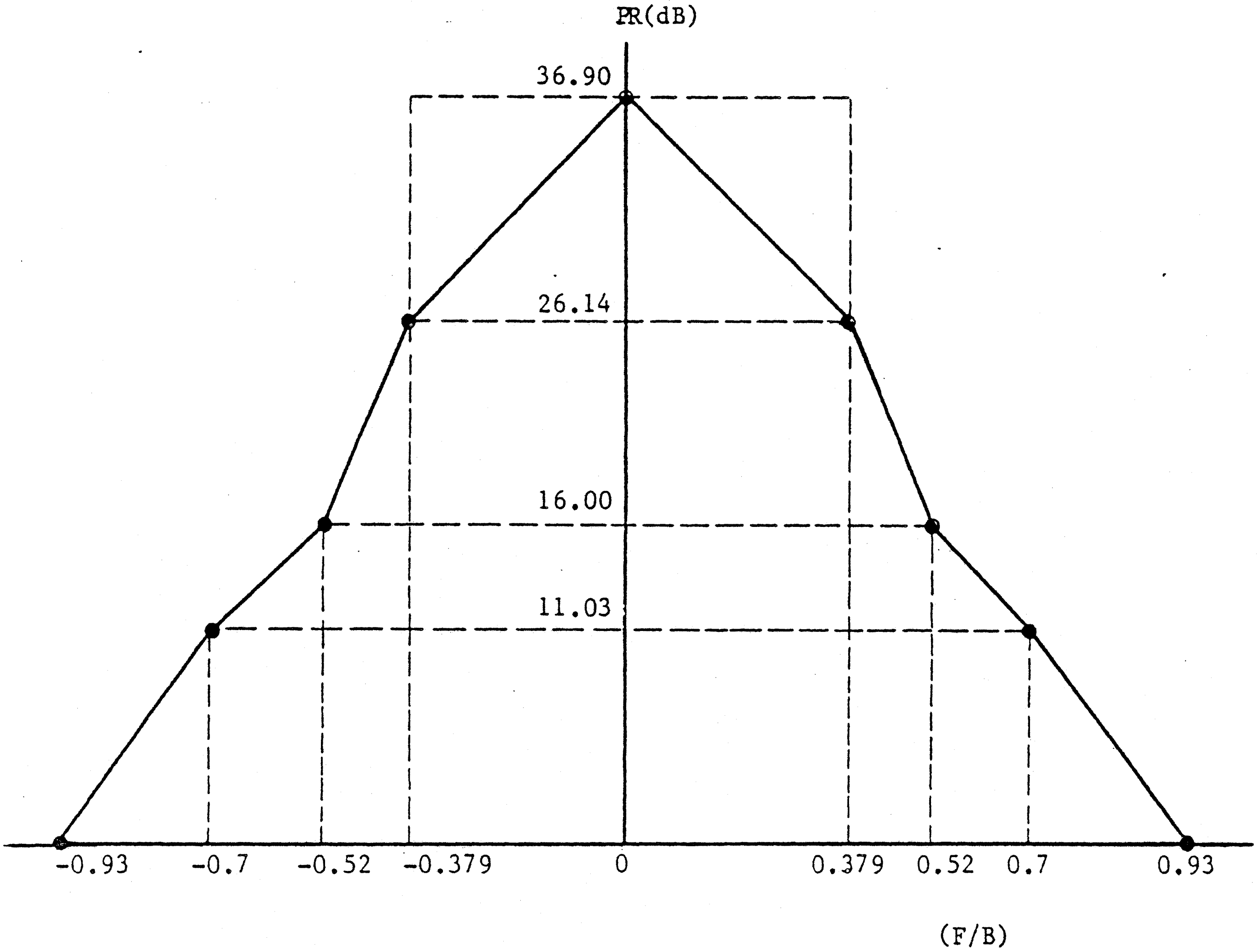
The shift also yields conservative protection ratios at offset values exceeding the overshoot region (e.g., an extra 10 dB for M/NTSC and 16 dB for K/SECAM at F = 10 MHz). This effect can be controlled by providing a decay pattern with piece-wise linear segments of varying slopes to fit experimental measurements. This approach is illustrated in Figure G-6, where the protection ratio template was obtained as a practical bound to support M/NTSC measurements with a varied static picture content [Bouchard and Chouinard, 1983].

Various protection ratio templates are available from CCIR documents, as shown in Figures G-7 through G-10. The patterns differ in various characteristics such as: (1) symmetric or asymmetric, (2) fixed or adjustable cochannel value, (3) flat or rising slope at the origin, (4) single or multiple breakpoints, (5) fixed or adjustable decay slope, (6) absolute or normalized frequency offset.

It is evident that there are significant differences between the various patterns and they have varied effects. A fixed cochannel value cannot accommodate the deviation dependence. A flat top can accommodate the overshoots only at the expense of a conservative cochannel value. A single breakpoint is more prone to produce conservative protection at offset values beyond the overshoot region. An absolute offset parameter cannot accommodate signal bandwidth restrictions.

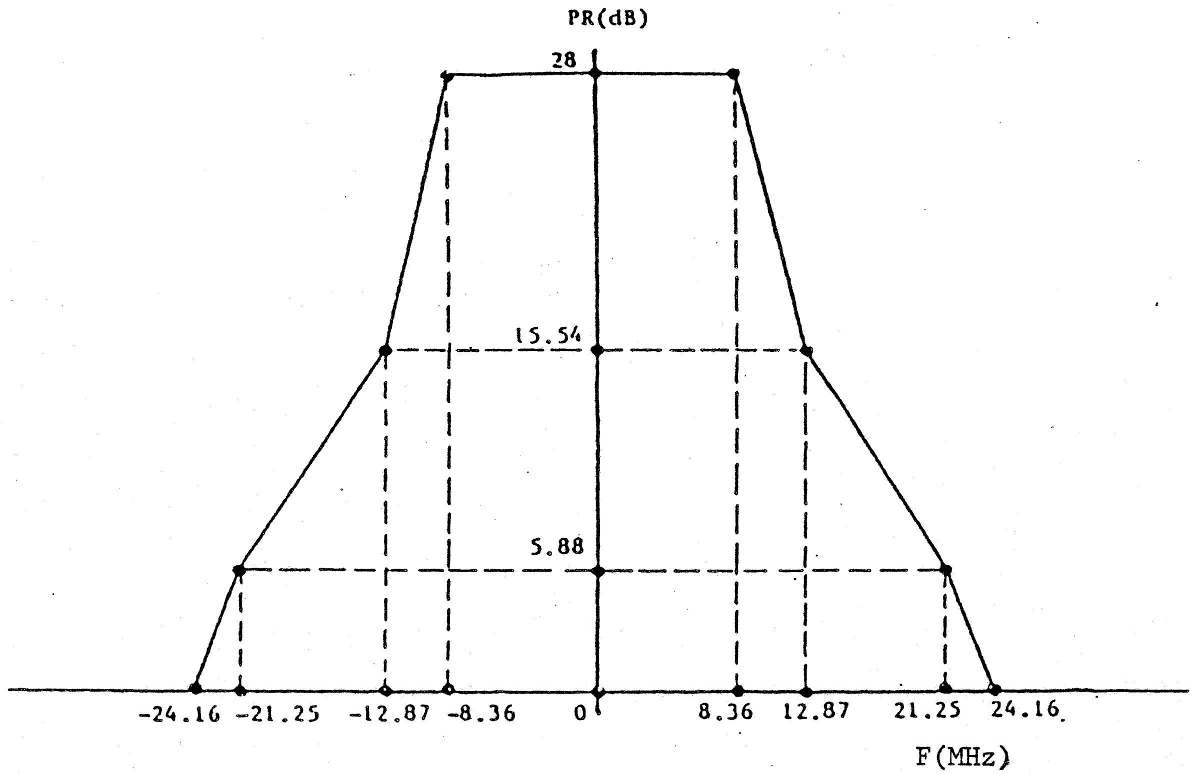
The templates in Figures G-8 and G-10 are the only two that provide for a cochannel protection that is dependent on the deviation parameter. The template in Figure G-8 permits the use of Equation G-1 for the cochannel protection ratio. The template of Figure G-10 represents a cochannel protection ratio of 29 + A - 20 log D (in dB), so that it matches Equation G-1 at Q = 4.3 for M/NTSC television and at Q = 3.8 for K/SECAM television.

The template in Figure G-10 is the only one that has the parametric capability to accommodate overshoots without a conservative cochannel value. It is also the only one with a decay slope that varies with the deviation parameter instead of being constant. This template was obtained with an unmodulated sinusoidal interference, and the extension to a modulated case is not straightforward [CCIR Report 388-4, 1982].



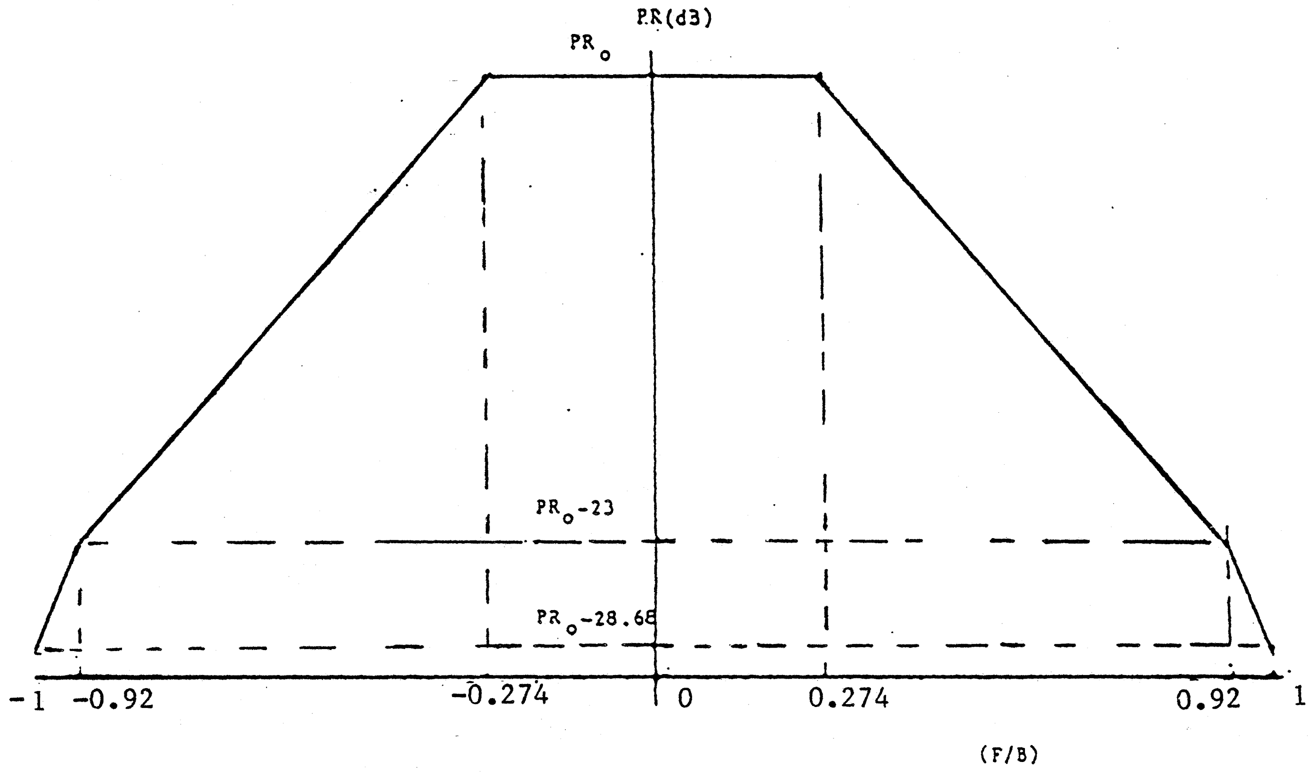
$$
\begin{aligned}
PR &= 36.90 - 28.40\,|F/B| && \text{for } |F/B| \le 0.379 \\
&= 53.25 - 71.63\,|F/B| && \text{for } 0.379 < |F/B| \le 0.52 \\
&= 30.44 - 27.73\,|F/B| && \text{for } 0.52 < |F/B| \le 0.7 \\
&= 44.64 - 48.05\,|F/B| && \text{for } |F/B| > 0.7
\end{aligned}
$$

Figure G-6. TV/FM protection ratio template [Bouchard and Chouinard, 1983].



$$
\begin{aligned}
PR &= 28 && \text{for } |F| \le 8.36 \\
&= 51.09 - 2.762\,|F| && \text{for } 8.36 < |F| \le 12.87 \\
&= 30.40 - 1.154\,|F| && \text{for } 12.87 < |F| \le 21.25 \\
&= 48.38 - 2.000\,|F| && \text{for } |F| > 21.25
\end{aligned}
$$

Figure G-7. TV/FM protection ratio template [CCIR/BSS, Region 2, Geneva, 1983].



$$PR = (PR)_o \quad \text{for } |F/B| \le 0.274$$

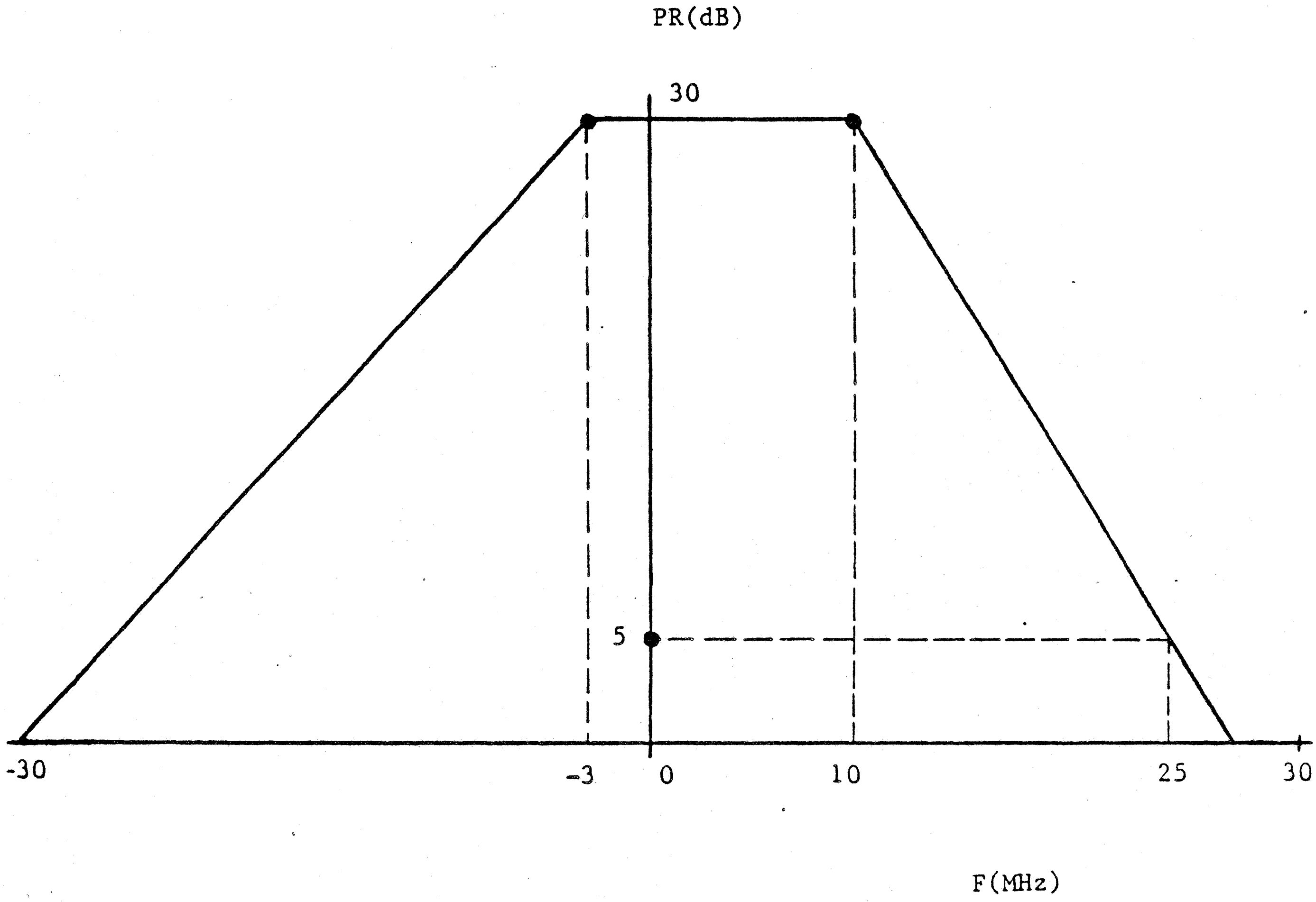
$$= (PR)_o - 35.6\,(|F/B| - 0.274) \quad \text{for } 0.274 < |F/B| \le 0.92$$

$$= (PR)_o - 23 - 71\,(|F/B| - 0.92) \quad \text{for } |F/B| > 0.92$$

where $(PR)_o$ = cochannel protection ratio

(F/B) = frequency offset/occupied bandwidth

Figure G-8. TV/FM protection ratio template [CCIR/CPM, Region 2, Geneva, 1982].

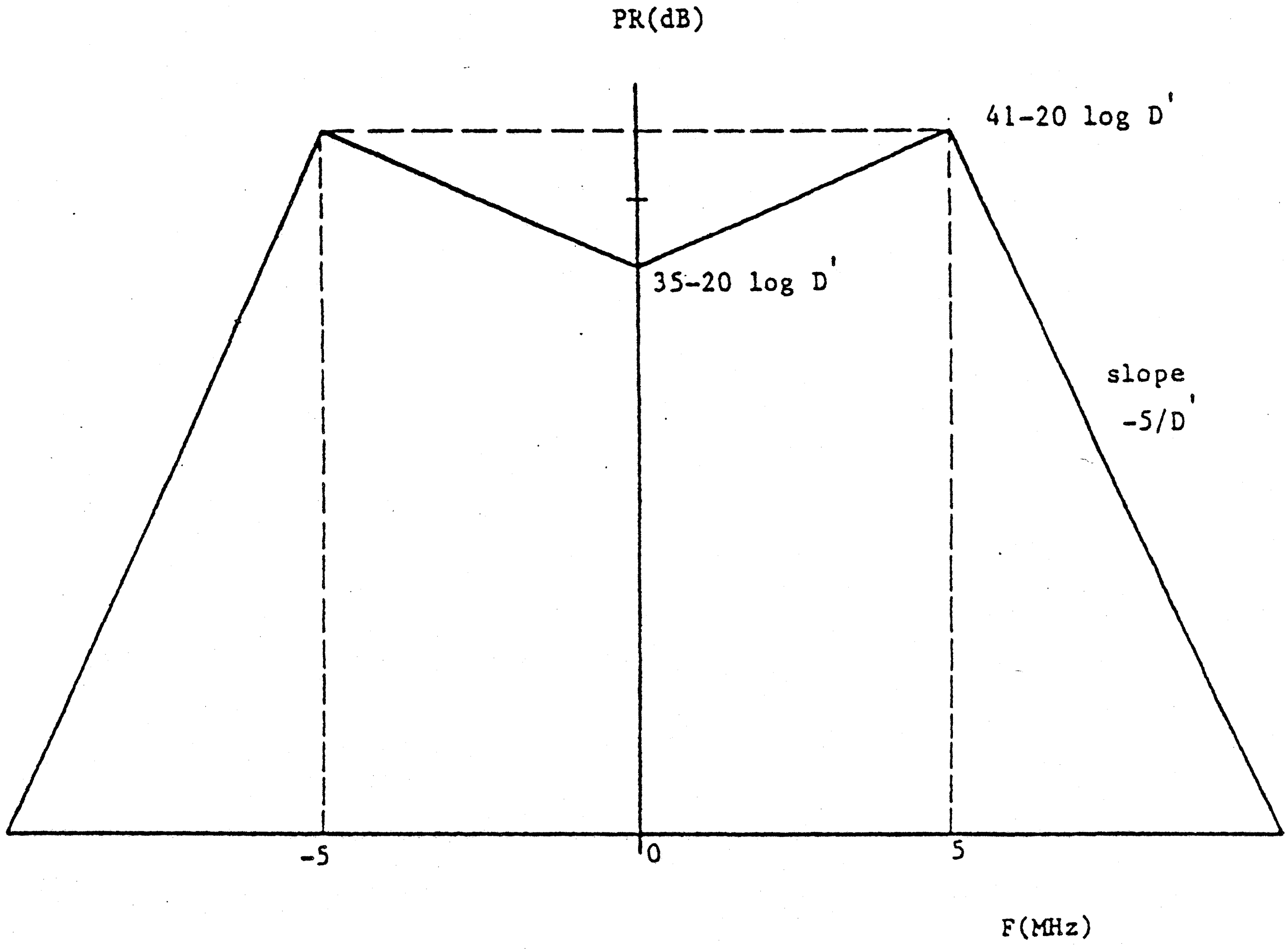


$$PR = 30 \quad \text{for} \quad -3 \leq F \leq 10$$

$$= 30 - (5/3)\ (F - 10) \quad \text{for} \quad F > 10$$

$$= 30 + (10/9)\ (F + 3) \quad \text{for} \quad F < -3$$

Figure G-9. TV/FM protection ratio template [CCIR Report 634-2, 1982].



$$PR = (35 - 20 \log D') + (6/5)\,|F| \quad \text{for} \quad |F| \le 5$$

$$= (41 - 20 \log D') - (5/D')\,(|F| - 5) \quad \text{for} \quad |F| > 5$$

where $D' = 0.5D/(\sqrt{10})^{A/10}$ where D = peak-to-peak deviation

$A = 10$ for 525-lines, 11 for 625 lines television

Figure G-10. TV/FM protection ratio template [CCIR Report 388-4, 1982].

A comparison of the templates in Figures G-6 through G-10 with the M/NTSC measurements in Figures G-1 and G-2 is presented in Figures G-11 through G-15. A comparison of the templates in Figures G-8 and G-10 with the K/SECAM measurements in Figure G-2 is also presented in Figure G-16, since these two templates have the parametric dependence to accommodate distinct television standards.

FDM/FM Telephony Interference

There is limited information concerning the effects of FDM/FM telephony interference on TV/FM. One report cites measured cochannel protection ratios of 32 and 35 dB for just perceptible interference from 60-channel and 970-channel FDM/FM telephony with 270 kHz and 800 kHz rms test-tone deviations [CCIR Report 634-2, 1982]. However, the TV/FM signal had a specific modulation consisting of preemphasized video with 14 MHz peak-to-peak deviation and preemphasized audio on a 5.5 MHz subcarrier with 75 kHz peak deviation.

Another report presents the experimental measurements shown in Figure G-17 for just perceptible interference from 132-channel FDM/FM telephony, using LOS-relay modulation standards for the TV/FM signal [CCIR Report 449-1, 1982). The cochannel values obtained are noted to be lower than those cited above, so that further investigation is needed to establish the dependence on the modulation parameters of the TV/FM and FDM/FM signals.

One approach has been to treat the FDM/FM signal as an equivalent TV signal of the same bandwidth and then use the TV/FM into TV/FM interference results. The numerical constants in Equation G-2 were modified to fit some FDM/FM telephony into TV/FM measurements as follows [Jeruchim and Kane, 1970]:

$$PR(dB) = 24.1 - 20 \log M_1 - F\,(M_1^{-1.5}) - 0.85\,(M^{-3})\,(F^{0.5M}) \log M \qquad (G\text{-}3)$$

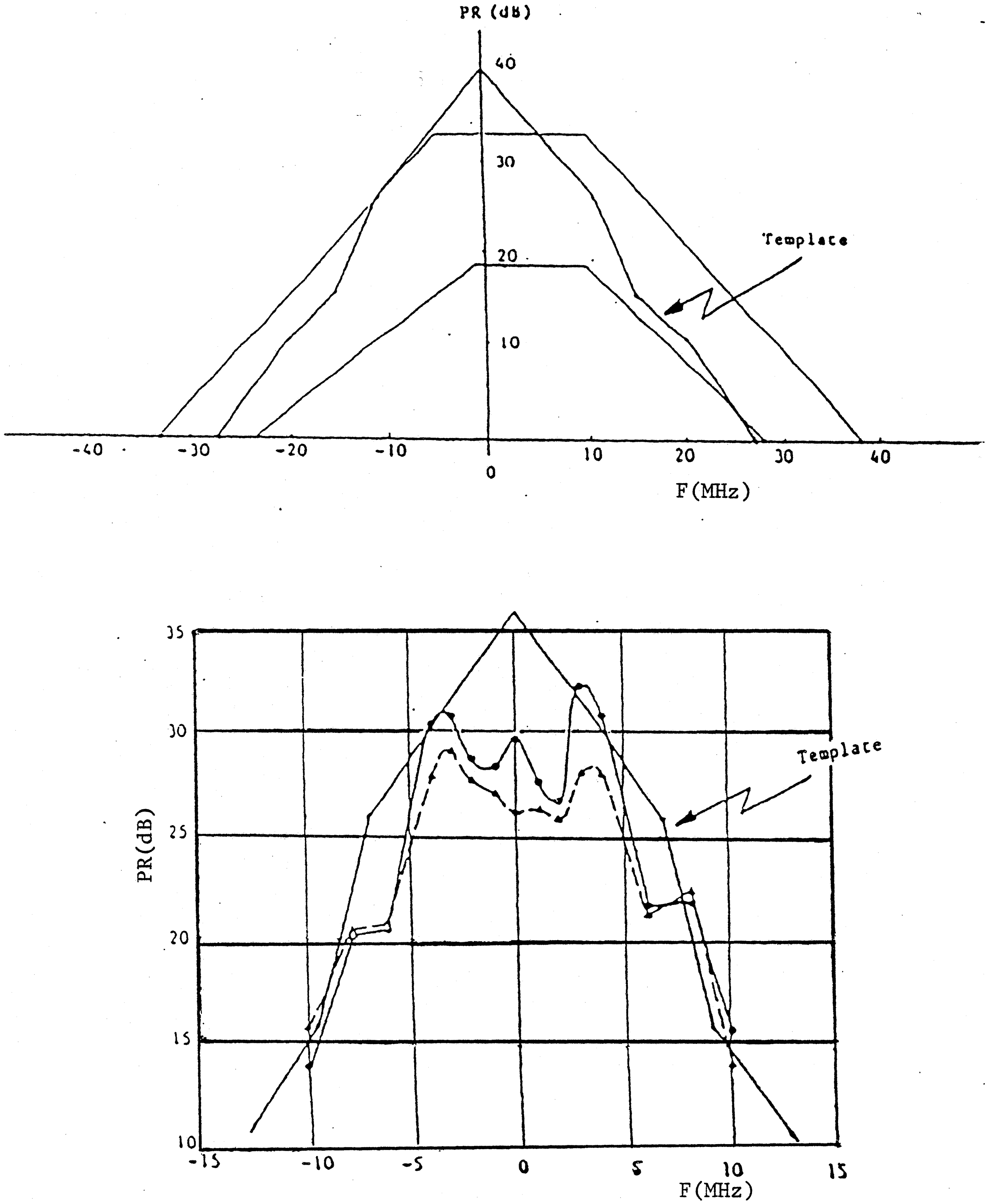
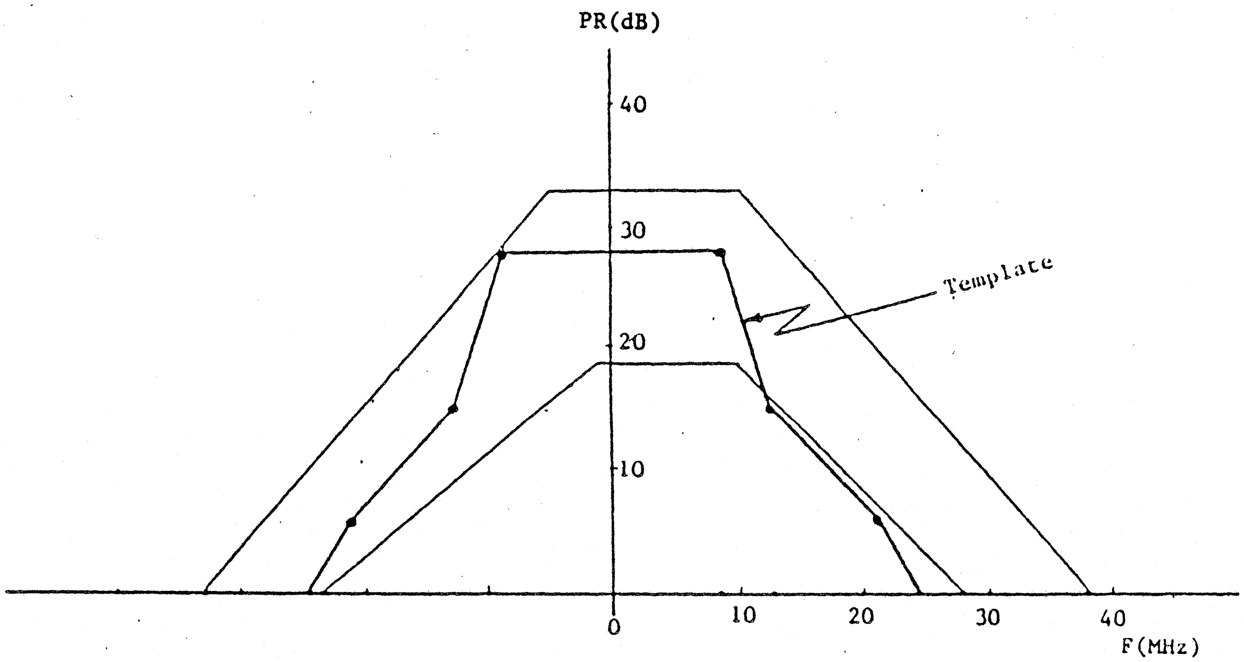


Figure G-11. Comparison of template in Figure G-6 with M/NTSC measurements in Figures G-1 (top) and G-2 (bottom).



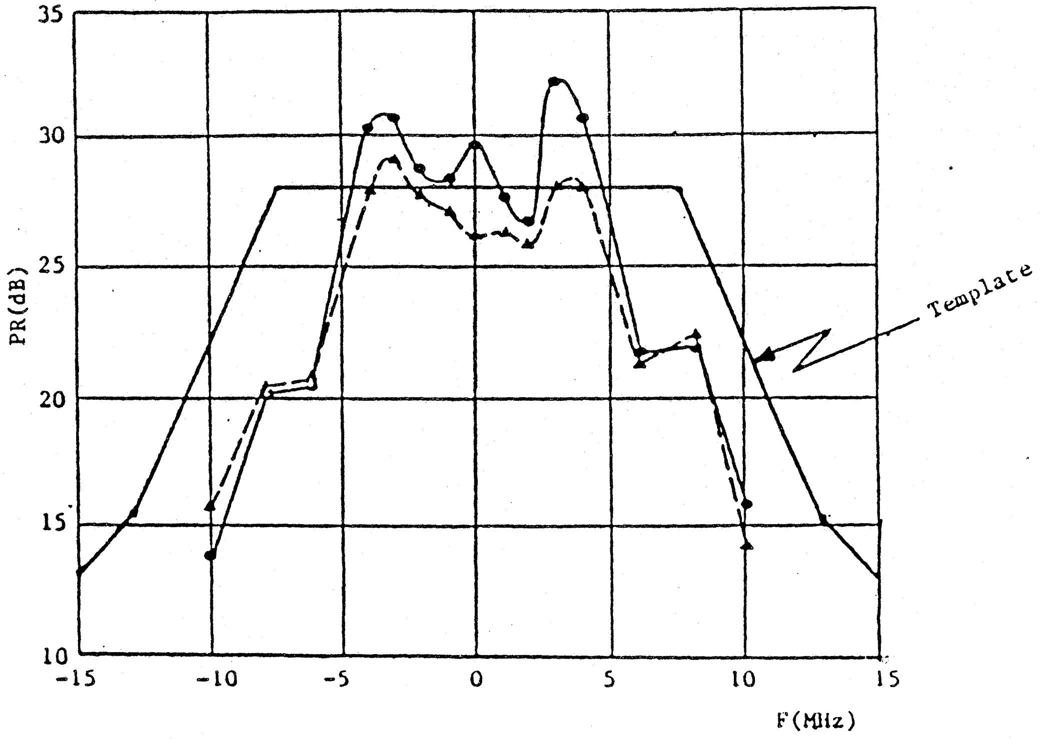
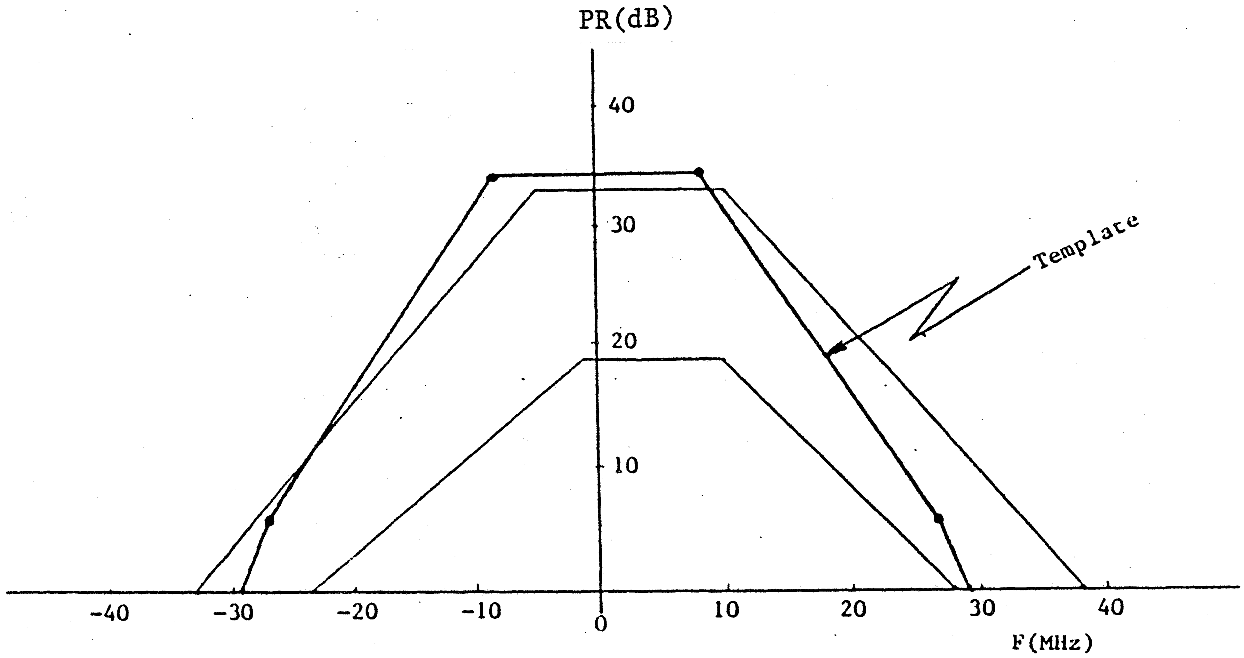
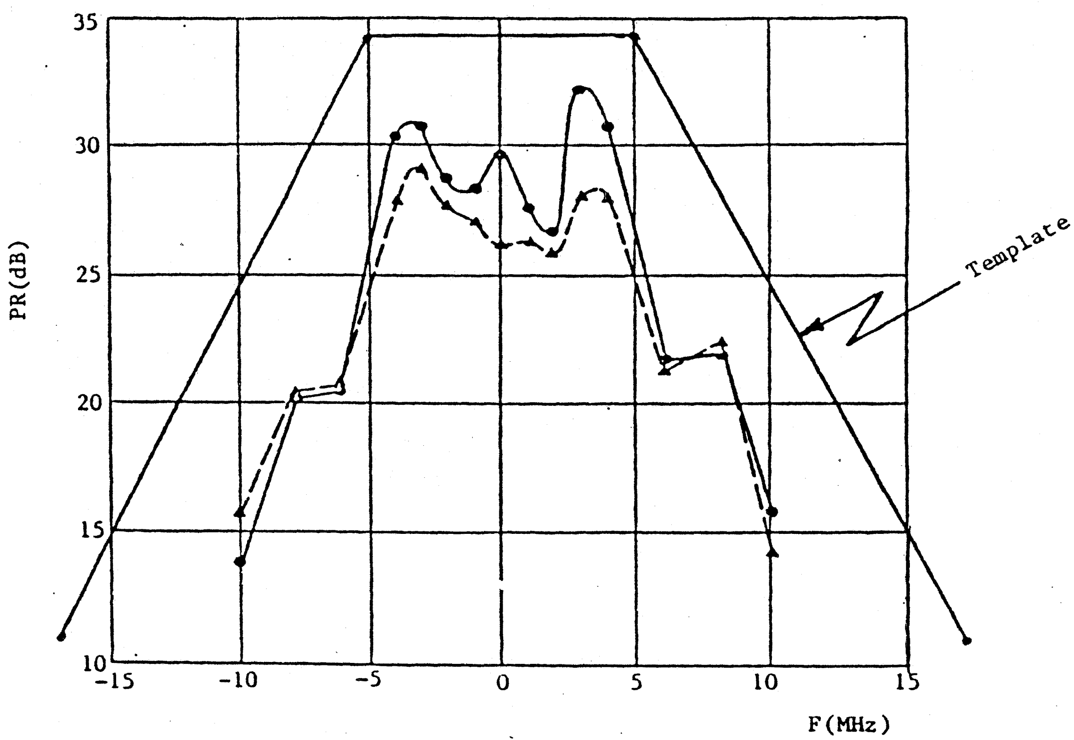
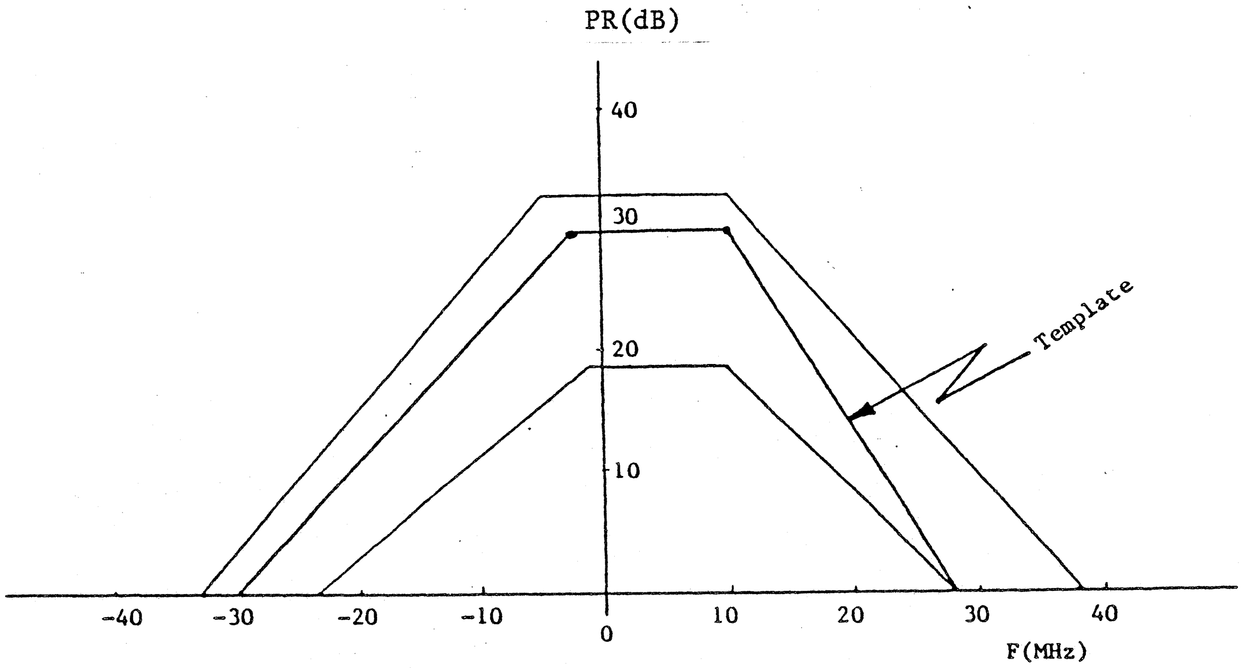


Figure G-12. Comparison of template in Figure G-7 with M/NTSC measurements in Figures G-1 (top) and G-2 (bottom).





G-13. Comparison of template in Figure G-8 with M/NTSC measurements in Figures G-1 (top) and G-2 (bottom).



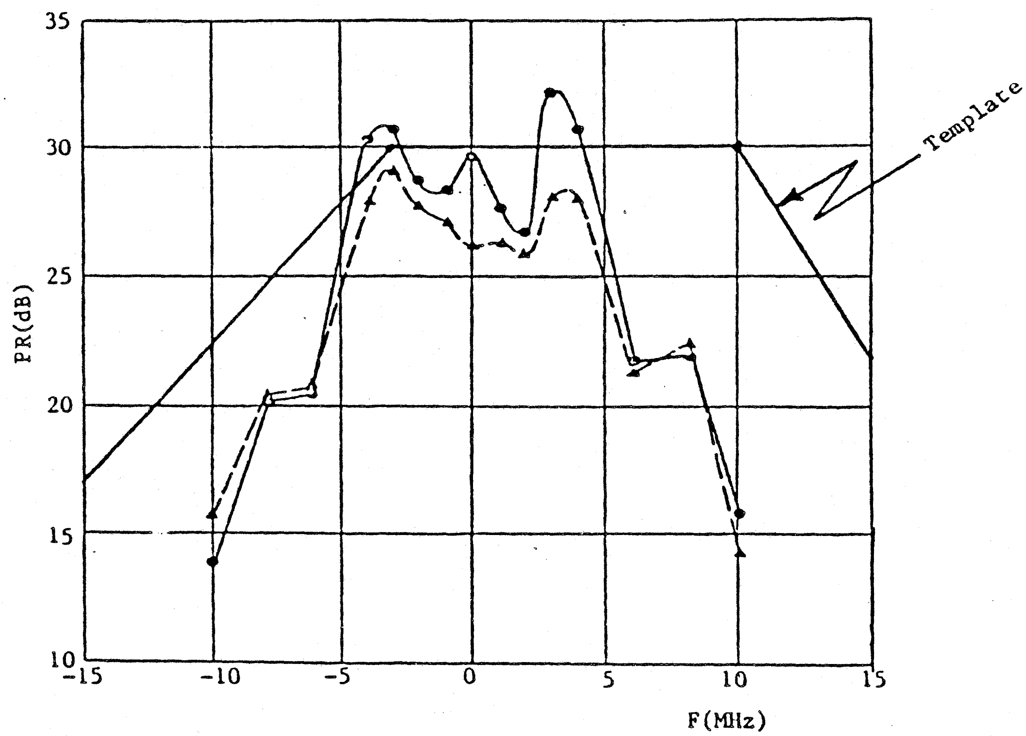
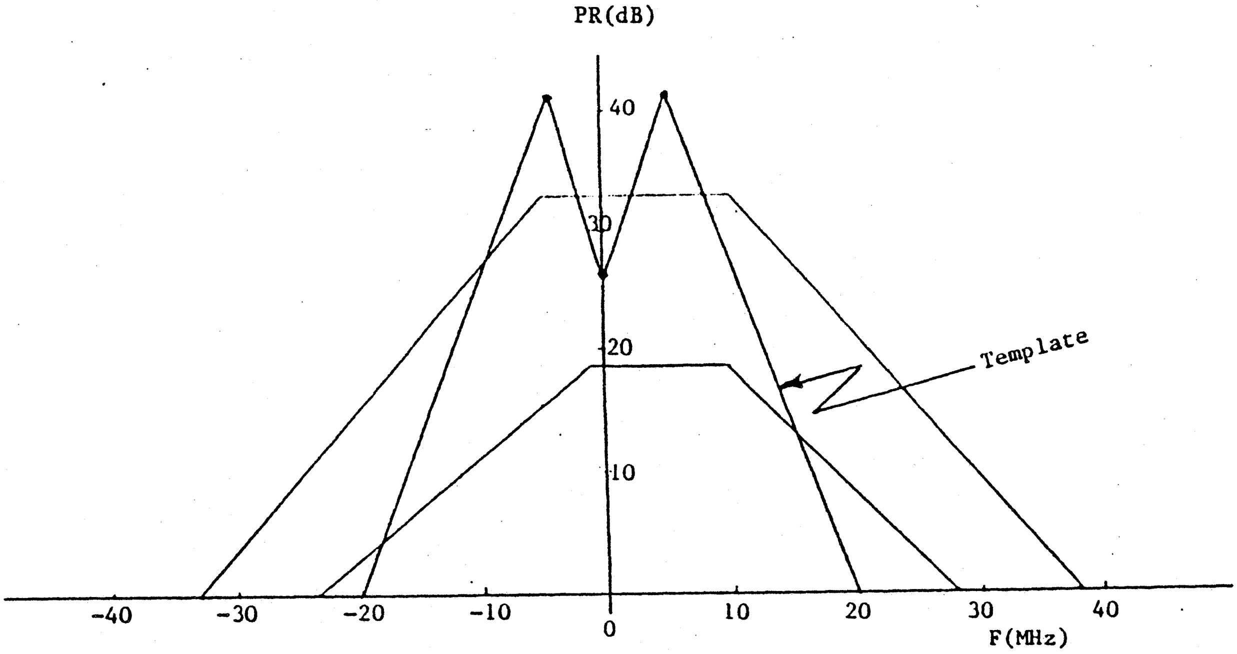


Figure G-14. Comparison of template in Figure G-9 with M/NTSC measurements in Figures G-1 (top) and G-2 (bottom).



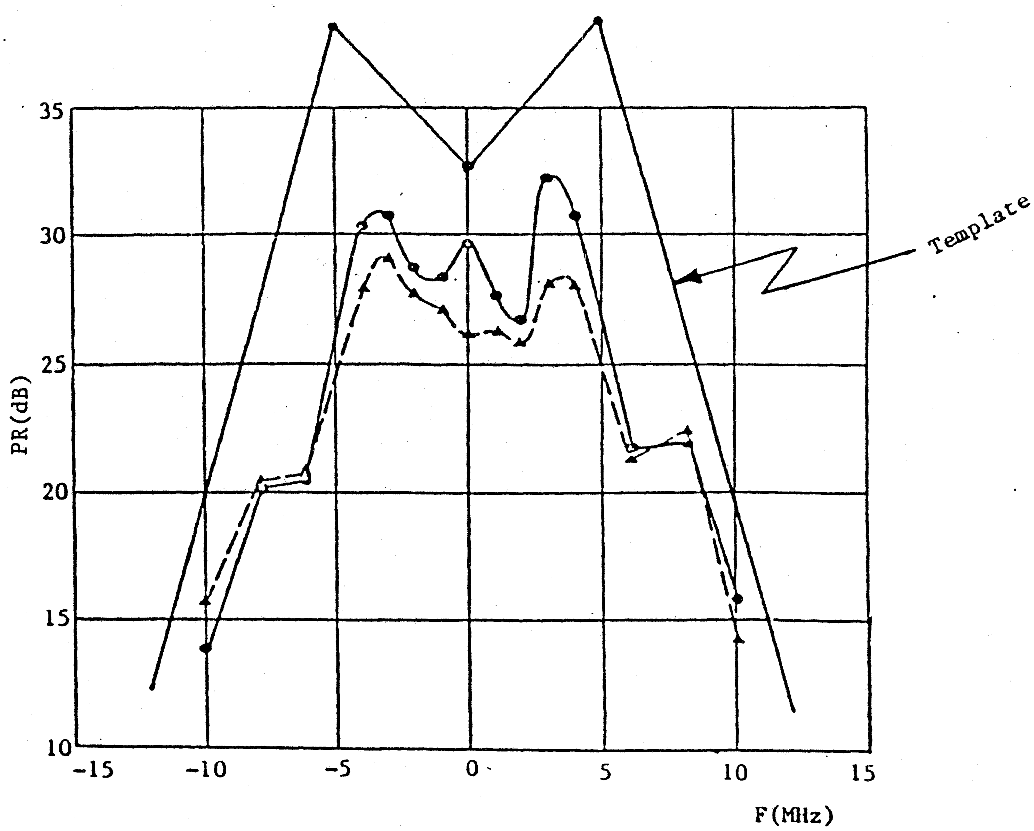


Figure G-15. Comparison of template in Figure G-10 with M/NTSC measurements in Figures G-1 (top) and G-2 (bottom).

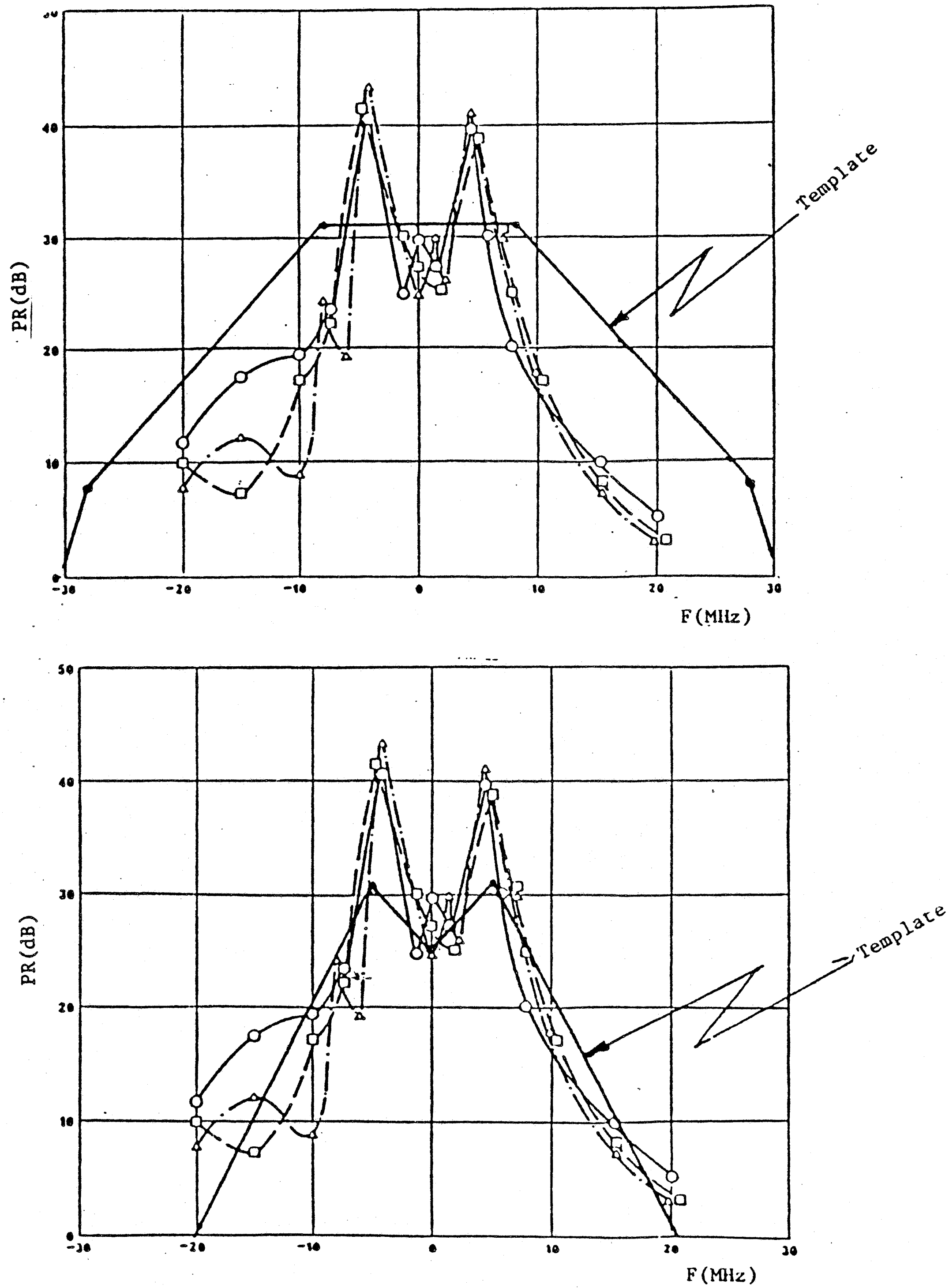


Figure G-16. Comparison of template in Figure G-8 (top) and G-10 (bottom) with K/SECAM measurements in Figure G-2.

The symbols are as before, except for $M_2 = (B_2/8) - 1$, where B_2 is the FDM/FM signal bandwidth in MHz. The M_2 parameter represents the peak modulation index of an equivalent TV signal occupying the same RF bandwidth as the FDM/FM signal in question (i.e., $B_2 = 2\ (M_2 + 1)\ (4\ \text{MHz})$ is Carson's Rule for the equivalent TV signal, and B_2 from the FDM/FM signal is used to compute the equivalent TV index M_2). The limitations of this formula are the limited amount of the data points actually employed in the original fit, plus the fact that the original formula was itself found deficient by not supporting the peak overshoots of TV/FM interference under offset conditions, again present in Figure G-17.

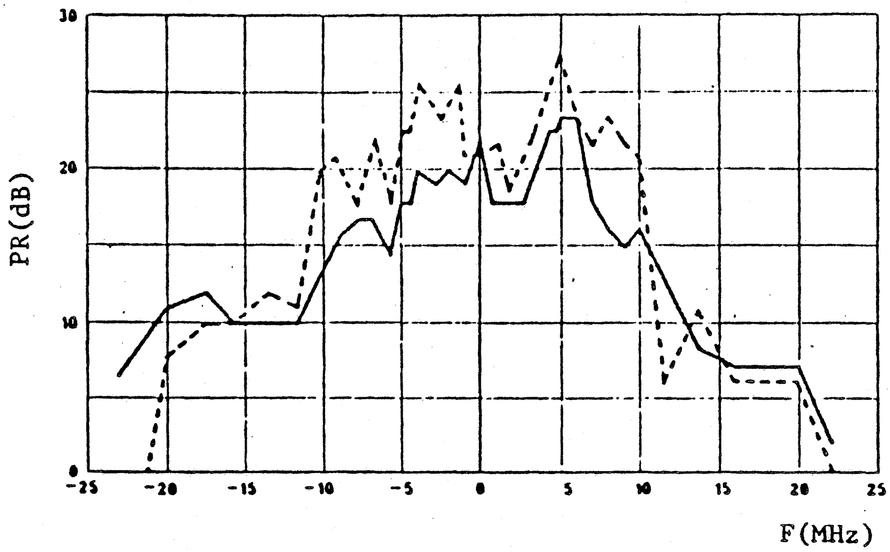
OUTPUT S/I CRITERION

The use of the protection ratio as a performance index has the disadvantage of relying on subjective measurements. Conversely, the use of an output power ratio (S/I) would represent an objective approach free from subjective effects. However, the output S/I must still be linked to output picture quality to properly evaluate performance degradation for practical purposes.

Cochannel TV/FM Interferer

A set of measurements involving both thermal noise and cochannel interference into TV/FM has been performed using D = 24 MHz M/NTSC and real program material in both desired and interfering signals. The composite output SNR was measured as the power ratio of the peak-to-peak signal to the weighted rms composite noise plus interference at the demodulator output. The results are shown in Figure G-18 where the (A) and (B) lines respectively correspond to the interference-predominating (low C/I) and noise-predominating (high C/I) performance regions [CCIR Report 449-1, 1982].

In particular, the (A) lines were used to derive the empirical relation S/I(dB) = C/I(dB) + 33.5 for D = 24 MHz M/NTSC television without thermal



Key: solid - full load
dotted - no load

Figure G-17. Adjacent-channel FDM/FM into TV/FM measurements [CCIR Report 449-1, 1982].
1982].

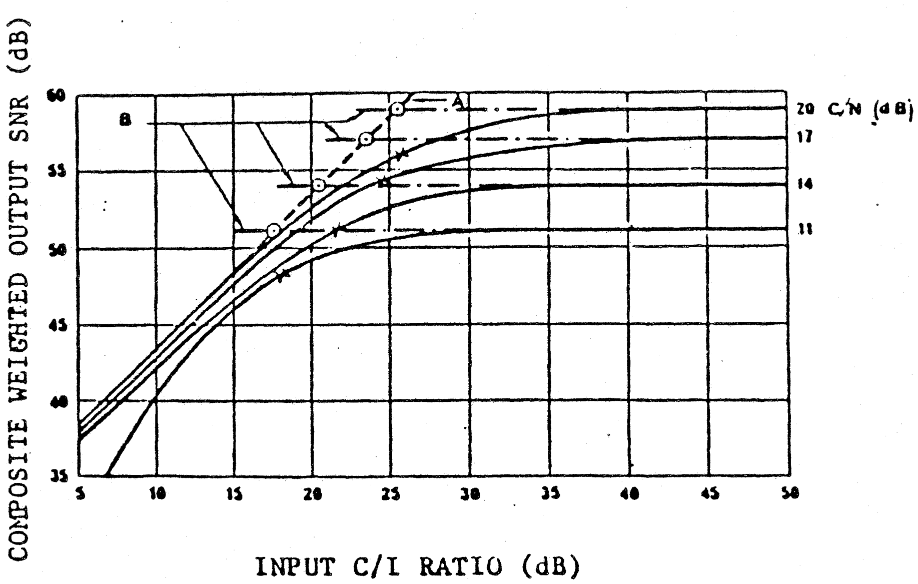


Figure G-18. Cochannel TV/FM into TV/FM measurements for D = 24 MHz NTSC television [CCIR Report 449-1, 1982].

noise. The relation was then extended to other deviations, as given in Equation 6-4 below, by assuming the 20 log D parametric dependence already found in the cochannel protection ratio formula.

$$S/I(dB) = C/I(dB) + 20 \log D + 6 \qquad (G\text{-}4)$$

The B lines in the figure can be verified to match the conventional formula for thermal noise only. This formula is given by Equation G-5 below, where B_V = 4.2 MHz for M/NTSC television, $B = D + 2B_V$ is the input bandwidth estimate, and G is the preemphasis plus noise-weighting gain. The use of D = 24 MHz yields S/N = C/N + 25.8 + G, and the value G = 13.8 dB [CCIR Report 215-5, 1982] yields S/N = C/N + 39.6. This relation matches the B lines of Figure G-18 within 1 dB.

Another formula has been proposed for the output S/I under cochannel or offset conditions [Jeruchim and Kane, 1970]. The cochannel relations were first established under fixed modulation parameters, and then extended via the previous Equation G-2 to accommodate other parameter values. The formula shown below was obtained for M/NTSC television, and it can be noted to represent an extension of the Equation G-2 already discussed.

$$S/I(dB) = C/I(dB) + 23 + 20 \log M_1 + F(M_1^{-0.85})$$
$$+ 0.475\ (M^{-2.5})\ (F^{0.645M}) \log M \qquad (G\text{-}6)$$

A comparison between the Equations G-4 and G-6 can be made for the cochannel case by letting F = 0, M = 1, M_1 = D/8 in Equation G-6. The two expressions then differ by only 1 dB for M/NTSC television. An offset condition is not handled by Equation G-4, unlike Equation G-6, but the latter is based on a formulation that has already been found deficient to accommodate the overshoots present under offset conditions, as previously discussed.

Cochannel and Adjacent-Channel PSK Interferer

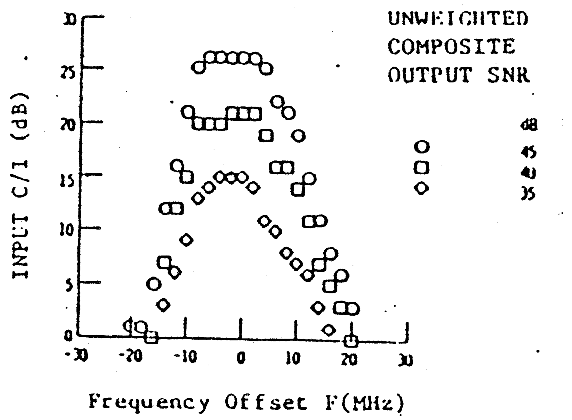
The nature of the degradation effects caused by thermal noise versus TV/FM interference in the output picture has been noted to be distinct. The effects caused by PSK interference have been noted to be similar to the thermal noise case, so that the output power ratio is a useful measurement for this interference type.

The results of recent measurements with differential QPSK interference are shown in Figure G-19 [Barnes, 1979]. The desired TV/FM signal consisted of M/NTSC television with D = 12 MHz and CCIR preemphasis. The unweighted output S/N with thermal noise only was set to 50 dB. The input C/I required to produce a given composite output SNR (35, 40, 45 dB unweighted) including both thermal noise and interference contributions, was then measured as a function of the carrier offset for two interferer data rates (22, 43 Mbps).

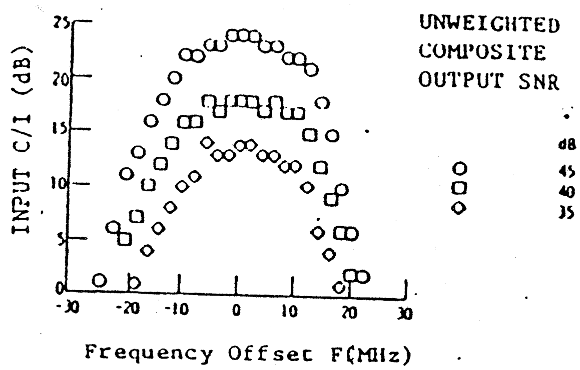
The data points were used along with a polynomial fit routine at NTIA to develop the empirical relations included in the figure. The 22 Mbps case shows a 6 dB C/I increment per 5 dB SNR increment, and the 43 Mbps case shows a definite pattern for the frequency offset dependence. However, more data is needed to generalize any apparent trends, as well as to include the effects of other parameters maintained invariant in these measurements (e.g., TV/FM frequency deviation, number of PSK phase states).

These same C/I measurements are also presented in a CCIR report as protection ratio values [CCIR Report 634-2, 1982]. However, it should be emphasized that these do not represent protection ratios in terms of a subjective picture quality. They actually represent the input C/I values needed to deliver a certain composite output SNR (35, 40, 45 dB unweighted) for given offset and data rate conditions. Any picture quality interpretation further requires the conversion of these output SNRs into a picture quality grade, (see APPENDIX C of this report).

EXPERIMENTAL MEASUREMENTS



(a) Interferer Data Rate = 22 Mbps



(b) Interferer Data Rate = 43 Mbps

POLYNOMIAL FIT RESULTS

(a) Case of Interferer Data Rate = 22 Mbps

$C/I = 15.0 - 6.67(F/10)^2 - 5.67(F/10)^4 + 1.33(F/10)^6$ for SNR = 35 dB

$C/I = 21.0 - 3.11(F/10)^2 - 3.77(F/10)^4 + 0.88(F/10)^6$ for SNR = 40 dB

$C/I = 27.0 - 0.11(F/10)^2 - 8.11(F/10)^4 + 2.22(F/10)^6$ for SNR = 45 dB

(b) Case of Interferer Data Rate = 43 Mbps

$C/I = 14.1 - 50.7(F/10)^2 + 1.21(F/10)^4 - 0.21(F/10)^6$ for SNR = 35 dB

$C/I = 18.2 - 3.86(F/10)^2 + 0.92(F/10)^4 - 0.20(F/10)^6$ for SNR = 40 dB

$C/I = 24.2 - 3.86(F/10)^2 + 0.92(F/10)^4 - 0.20(F/10)^6$ for SNR = 45 dB

Figure G-19. Differential QPSK interference into TV/FM
(D = 12 MHz and M/NTSC television).

BIBLIOGRAPHY

Abramson, N. (1963), "Bandwidth and Spectra of Phase and Frequency Modulated Waves," IEEE Trans. on Communication Systems, December.

Aein, J.M. and Turner, R.D. (1973), "Effect of Cochannel Interference on CPSK Carriers," IEEE Trans. on Communications, Vol. COM-21, No. 7, July.

Boyhan, J.W. (1984), "Normalized Power Spectra for FDM/FM Signals and Noise Power Ratio Calculations," IEEE Trans. on Communications, Vol. COM-32, No. 1, January.

Brown, R.J., Guha, M.L., Hedinger, R.A., and Hoover, M.L. (1982), "Companded Single Sideband Satellite Transmission," Globecom Conference Record, Miami, Florida.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. IV-1, Recommendation 464-1, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. IV/IX-2, Report 792-1, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. IX-1, Recommendation 275-3, Geneva.

CCIR/USA (1983), "Increasing Geostationary Orbit Capacity By Using Companded Single Sideband Transmission," CCIR Study Groups, Document USSG IWP 4/1, February.

Chakraborty, D. and Robertson, A. (1974), "An Experimental Investigation of 8-Phase CPSK Signal Transmission in Band-Limited Satellite Channels", Comsat Technical Review, Vol. 4, No. 2, Fall.

Das, A. and Sharp, G. (1978), "Convolution Method of Interference Calculation," Federal Communications Commission, Office of Chief Engineer, Research and Standards Division, Washington, D.C., FCC/OCE RS75-04, April.

Fang, R. and Shimbo, O. (1973), "Unified Analysis of a Class of Digital Systems in Additive Noise and Interference", IEEE Trans. on Communications, Vol. COM-21, No. 10, October.

Ferguson, M.E. (1982), "Sideband: Analog's Quantum Leap," Satellite Communications, August.

Freeling, M.R. and Braun, W.H. (1980), "Maximum Transponder Capacity for Transmission of FDM/FM Channels," RCA Review, Vol. 41, September.

Hamer, R. (1961), "Radio Frequency Interference in Multichannel Telephony FM Radio Systems," I.E.E., June.

Jeruchim, M.C. and Kane, D.A. (1970), "Orbit/Spectrum Utilization Study," VOl. IV, General Electric, Space Systems Organization, Valley Forge Space Center, Philadelphia, Pennsylvania, Document No. 70SD4293, December.

Jeruchim, M.C. (1971), "Interference in Angle Modulated Systems with Predetection Filtering," IEEE Trans. on Communication Technology, October.

Johannsen, K., Tustison, G. and Egli, S. (1980), "Notes on SCPC/FM System Design and Performance," IEEE Trans. on Aerospace and Electronic Systems, Vol. AES-16, No. 5, September.

Johns, P.B. and Rowbotham, T.R. (1972), "Communication System Analysis," Van Nostrand Reinhold Company.

Juroshek, J.R. (1976), "Performance of Digital Modems with Cochannel Interference and Gaussian Noise", U.S. Department of Commerce, Office of Telecommunications, OT-76-82, February.

Koh, E.K. and Shimbo, O. (1976), "Computation of Interference into Angle Modulated Systems Carrying Multichannel Telephone Signals," IEEE Trans. on Communications, February.

Kurjan, D. and Wachs, M. (1976), "TV Cochannel Interference on a PCM-PSK SCPC System," Comsat Technical Review, Vol. 6, No. 2, Fall.

Medhurst, R.G., Hicks, E.M., and Grosset, W. (1958), "Distortion in FDM/FM Systems Due to An Interfering Carrier," I.E.E., May.

Medhurst, R.G., (1961), "RF Spectra and Interfering Carrier Distortion in FM Trunk Radio Systems with Low Modulation Ratios," IRE Trans. on Communications Systems, June.

Prasanna, S. Sharp, G., and Das, S. (1971), "Interference Between FM Carriers and a Digitally Modulated SCPC Circuit," IEEE Trans. on Aerospace and Electronic Systems, July.

Rosenbaum, A.S. and Glave, F.E. (1974), "An Error Probability Upper Bound for CPSK with Peak-Limited Interference", IEEE Trans. on Communications, Vol. COM-22, No. 1, January.

Shimbo, O. and Fang, R. (1973), "Effects of Cochannel Interference and Gaussian Noise in M-ary PSK Systems", Comsat Technical Review, Vol. 3, No. 1, Spring.

Skevington, R.C. (1979), "Compandors for FDM/FM Voice Traffic," National Telemetry Conference, Washington, D.C., November.

Stojanovic, Z.D., Dukic, M.L. and Stojanovic, I.S. (1981), "A New Demodulation Method Improving FM System Interference Immunity," IEEE Trans. on Communications, Vol. COM-32, No. 7, January.

Szarvas, G.G. and Suyderhoud, H.G. (1981), "Enhancement of FDM/FM Satellite Capacity by Use of Compandors," COMSAT Technical Review, Vol. 11, No. 1, Spring.

REFERENCES

Akima, H. (1985), "Technical Basis for the Geostationary Satellite Orbit Analysis Program, Version 2," U.S. Department of Commerce, National Telecommunications and Information Administration, NTIA Report 85-183.

Ali, A.A. (1983), "Spectral Analysis of Analog Video and PSK Data Transmitted by FM," IEEE Trans. on Broadcasting, Vol. BC-9, No. 1, March.

Ball, J.E. and Rubin, P.A. (1978), "Communication Satellites for Public Television," IEEE Trans. on Broadcasting, Vol. BC-24, No. 2, June.

Barnes, S.P. (1979), "Carrier-Interference Ratios for Frequency Sharing Between Satellite Systems Transmitting Frequency Modulated and Digital Television Signals", National Telemetry Conference, Washington, D.C.

Blackman, N.M. and McAlpine, G.A. (1969), "The Spectrum of a High-Index FM Waveform," IEEE Trans. on Communication Technology, Vol. COM-17, No. 2, April.

Bouchard, M. and Chouinard, G. (1983), "Subjective Evaluation of the Interference Protection Ratio for Frequency Modulated NTSC Television Signals," Communications Research Centre, Dept. of Communications, Ottawa, Canada, April.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. IV-1, Recommendation 354-2, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. IV-1, Recommendation 466-3, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. IV-1, Recommendation 483-1, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. IV-1, Recommendation 523-1, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. IV-1, Report 455-3, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. IV-1, Report 708-1, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. IV-1, Report 867, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. IV/IX-2, Report 388-4, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. IV/IX-2, Report 449-1, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. IX-1, Recommendation 405-1, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. X/XI-2, Report 215-5, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. X/X1-2, Report 634-2, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. XII, Recommendation 500-2, Geneva.

CCIR (1982), Recommendations and Reports of the CCIR, XVth Plenary Assembly, Vol. XII, Recommendation 567-1, Geneva.

CCIR (1983), CCIR Study Groups, Period 1982-1986, Documents 9/1-E, 10-11S/3E, CMTT/5-E, April 25, 1983.

CCIR (1984), CCIR Study Groups, Period 1982-1986, Document 4/186-E, May 15, 1984.

CCIR (1984), Documents, CCIR Study Groups, Period 1982-1986, Document 4/194-E, May 17, 1984.

CCIR (1984), CCIR Study Groups, Period 1982-1986, Documents 4/116-F and 9/105-F, April 30, 1984.

CCIR/CPM (1982), "Technical Bases for the Regional Administrative Radio Conference 1983 for the Planning of the Broadcasting Satellite Service in Region 2," Geneva.

CCIR/BSS (1983), "Final Acts of the Regional Administrative Conference for the Planning of the Broadcasting Satellite Service in Region 2", Geneva.

Celebiler, M.I. and Coupé, G.M. (1978), "Effects of Thermal Noise, Filtering and Cochannel Interference on the Probability of Error in Binary Coherent PSK Systems," IEEE Trans. on Communications, Vol. COM-26, No. 2, February.

Cohen, D.J. (1984), "Necessary Bandwidth and Spectral Properties of Digital Modulation," U.S. Dept. of Commerce, National Telecommunications and Information Administration, NTIA Report 84-168, February.

Eng, K.Y. and Haskell, B.G. (1981), "TV Bandwidth Compression Techniques using Time Companded Differentials and their Applications to Satellite Transmissions," National Telemetering Conference, New Orleans, Louisiana.

Filippi, C.A. (1983), "FM Spectral Modeling and FDM/FM Simulation Programs," U.S. Dept. of Commerce, National Telecommunications and Information Administration, NTIA Report 83-134, October.

Goldberg, A.A. and Jones, B.L. (1983), "Subjective Assessment of Cochannel Protection Ratio for DBS," CBS Technology Center, Stamford, Connecticut, April.

Groumpos, P.P. and Vernon, G.D. (1981), "Experimental Results for the Interference Between FM Television Signals," National Telemetry Conference, New Orleans, Louisiana.

Haines, R.H. (1985), "Fixed-Satellite Service Data Base for ITU Region 2," U.S. Department of Commerce, National Telecommunications and Information Administration, NTIA-TN-86-1, October.

Haller, R. and VanDeursen, H. (1983), "Amplitude Compandored Sideband Compared to Conventional Frequency Modulation for VHF Mobile Radio," Federal Communications Commission, Office of Science and Technology, FCC/OST TM-83-7, October.

Hurt, G., Akima, H., Ng, H. and Filippi, C. (1986), "Geostationary Satellite Orbit Analysis Program (GSOAP)," U.S. Department of Commerce, National Telecommunications and Information Administration, AIAA Communications Satellite Systems Conference, San Diego, California, March 1986.

Izumi, T. and Matsumae, A. (1982), "Transmission System for the Television Broadcasting Satellite," IEEE Trans. on Broadcasting, Vol. BC-28, No. 4, December.

Jeruchim, M.C. and Kane, D.A. (1970), "Orbit/Spectrum Utilization Study", Vol. IV, General Electric, Space Systems Organization, Valley Forge Space Center, Philadelphia, Pennsylvania, Doc. No. 70SD4293, December.

Jeruchim, M.C. (1977), "A Survey of Interference Problems and Applications to Geostationary Satellite Networks", Proceedings of the IEEE, Vol. 65, No. 3, March.

Jonnalagadda, K. (1982), "SSB/AM Satellite Voice Communication System Having 6000 Channels for Transponder," RCA Review, Vol. 43, September.

Loo, C. (1977), "S/N Improvement of an FM Signal at or Near Threshold due to Deemphasis and Noise Weighting," IEEE Trans. on Broadcasting, Vol. BC-23, No. 3, September.

Miller, E. (1984), "FM Spectra of Selected Video Signals of Interest," NASA Lewis Research Center, Ohio, October 29, 1984.

Morinaga, N. and Namekawa T. (1982), "Intersystem Interference Between Analog and Digital Systems," International Communications Conference, Philadelphia, Pennsylvania.

NTIA (1985), "Manual of Regulations and Procedures for Federal Radio Frequency Management," U.S. Department of Commerce, National Telecommunications and Information Administration (NTIA), Revised May 1985.

Panter, P.F. (1972), "Communication Systems Design," McGraw Hill Book Company.

Pontano, B.A., Fuenzalida, J.C., and Chitre, N.K. (1973), "Interference into Angle Modulated Systems Carrying Multichannel Telephony Signals," IEEE Trans. on Communications, Vol. COM-21, No. 6, June.

Pritchard, W.L. and Radin, H.W. (1984), "Direct Broadcast Satellite Service by Direct Broadcast Satellite Corporation," IEEE Communications, March.

Rogers, J.C., Madigan, R.J. and Elliott, J.C. (1978), "Subjective Tests of a Companded SCPC/FM Communication System," IEEE Trans. on Communication, Vol. COM-26, No. 6, June.

Rosenbaum, A.S. (1969), "PSK Error Performance with Gaussian Noise and Interference," Bell System Technical Journal, February.

Schwartz, M., Bennett, W.R. and Stein, S. (1966), "Communication Systems and Techniques," McGraw Hill Book Company.

Sharp, G.L. (1983), "Reduced Domestic Satellite Orbital Spacings at 4/6 GHz," Federal Communications Commission, Office of Science and Technology, Technical Analysis Division, Washington, D.C., FCC/OST R83-2, May.

Sharp, G. (1984), "Reduced Domestic Satellite Orbit Spacing," AIAA Communications Satellite Systems Conference, Orlando, Florida, March.

Sklar, B. (1983), "A Structured Overview of Digital Communication: A Tutorial Review, Part II," IEEE Communications Magazine, October.

Stein, S. and Jones, J.J. (1967), "Modern Communications Principles," McGraw-Hill Book Company.

Whyte, W. (1983), "Analysis of Adjacent-Channel Interference for FM Television Systems," NASA Lewis Research Center, Ohio, June.

Wozencraft, J.M. and Jacobs, I.M. (1965), "Principles of Communication Engineering," John Wiley and Sons, Inc.

Yam, S.E. (1980), "New TV Energy Dispersal Techniques for Interference Reduction," Comsat Technical Review, Vol. 10, No. 1, Spring.

FORM **NTIA-29**
(4-80)

U.S. DEPARTMENT OF COMMERCE
NAT'L. TELECOMMUNICATIONS AND INFORMATION ADMINISTRATION

BIBLIOGRAPHIC DATA SHEET

	1. PUBLICATION NO.	2. Gov't Accession No.	3. Recipient's Accession No.
	NTIA TR-86-198		

4. TITLE AND SUBTITLE

BASEBAND PERFORMANCE DEGRADATION DUE TO INTERFERENCE IN THE FIXED-SATELLITE SERVICE

5. Publication Date

JULY 1986

6. Performing Organization Code

7. AUTHOR(S)

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9. Project/Task/Work Unit No.

9014101

8. PERFORMING ORGANIZATION NAME AND ADDRESS

National Telecommunications & Information Administration
Annapolis Office
179 Admiral Cochrane Drive
Annapolis, MD 21401

10. Contract/Grant No.

11. Sponsoring Organization Name and Address

U.S. Department of Commerce/NTIA
179 Admiral Cochrane Drive
Annapolis, MD 21401

12. Type of Report and Period Covered

TECHNICAL

13.

14. SUPPLEMENTARY NOTES

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

This report describes a set of receiver transfer characteristic (RTC) algorithms developed at NTIA for the analytical assessment of mutual interference effects in satellite communication services. The RTC algorithms convert the input carrier-to-interference power ratio (C/I) and modulation specifications of the desired (C) and interferer (I) signals into an output baseband performance degradation. The RTC algorithms also compute a C/I threshold margin for a given output baseband performance requirement. The modulation types and communication services include: (1) companded single-sideband, amplitude-modulation, multichannel telephony, (2) regular or companded, frequency-modulated, multichannel telephony, (3) frequency-modulated analog television, (4) digital coherent multiple-phase-shift keying, (5) single-channel-per-carrier with digital coherent multiple-phase-shift keying, and (6) single-channel-per-carrier with frequency-modulated analog voice.

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

BASEBAND PERFORMANCE DEGRADATION; FIXED-SATELLITE SERVICE COMMUNICATIONS; INTERFERENCE ANALYSIS ALGORITHMS; RECEIVER TRANSFER CHARACTERISTICS

17. AVAILABILITY STATEMENT	18. Security Class. *(This report)*	20. Number of pages
☒ UNLIMITED.	UNCLASSIFIED	190
☐ FOR OFFICIAL DISTRIBUTION.	19. Security Class. *(This page)* UNCLASSIFIED	21. Price