RECOMMENDATION UIT-R SF.766*

METHODS FOR DETERMINING THE EFFECTS OF INTERFERENCE ON THE PERFORMANCE AND THE AVAILABILITY OF TERRESTRIAL RADIO-RELAY SYSTEMS AND SYSTEMS IN THE FIXED-SATELLITE SERVICE

(1992)

The ITU Radiocommunication Assembly,

considering

a) that it is necessary to evaluate the effects of interference on the performance and the availability of terrestrial radio-relay systems and systems in the fixed-satellite service;

b) that, in general, the determination of interference criteria requires suitable calculation methods;

c) that calculation methods for determining interference to FDM-FM systems are fairly well established;

d) that calculation methods for interference to single-channel-per-carrier (SCPC) FM telephony are to be established;

e) that calculation methods for interference to FM television are to be established;

f) that calculation methods for interference to amplitude modulated (AM) telephony are to be established;

g) that calculation methods for interference to digital transmissions are to be established;

h) that, in future, calculation methods for interference to systems employing new modulation techniques may need to be established;

j) that it is desirable to provide spectra of signals for determination of interference from the general formulation,

recommends

1. that the methods described in Annex 1 be used for calculation of interference to FDM-FM systems;

2. that in the absence of more accurate information the methods described in Annex 2 be used provisionally for wanted signals other than FDM-FM.

ANNEX 1**

Calculation methods for the interference of FDM-FM systems

Given below is the method of calculation to determine the effects of interference to the FDM-FM systems in terrestrial radio-relay systems and systems in the fixed-satellite service.

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* Radiocommunication Study Groups 4 and 9 made editorial amendments to this Recommendation in 2000 in accordance with Resolution ITU-R 44.

** *Note from the Director CCIR* – For information, derivation of the formulae and historical development of this Annex are given by References contained in CCIR Report 388 (Düsseldorf, 1990).

1. Calculation methods

1.1 General formulation

The relationship (this linear relationship is only valid for the lower levels of interference into FDM-FM telephony signals) between baseband interference power in a telephone channel and the carrier-to-interference ratio involves the interference reduction factor B (in dB), defined as follows:

$$B = 10 \log \frac{S/N_i}{C/I} \qquad (1)$$

where:

- S : test signal power in a telephone channel = 1 mW
- N_i : unweighted interference power in a telephone channel (bandwidth: 3.1 kHz)
- C : power of the wanted signal carrier (W)
- I : power of the interfering signal carrier (W).

The weighted interference power N_p (pW) is obtained as unweighted power in 1.75 kHz, which gives:

$$10 \log N_p = 87.5 - B - 10 \log (C/I) \qquad (2)$$

The interference reduction factor B is expressed as:

$$B = 10 \log \frac{2(\delta f)^2 \, p(f/f_m)}{b \, f^2 \, D(f, f_0)} \qquad (3)$$

with:

$$D(f, f_0) = \int_{-\infty}^{+\infty} S(F) \, P_1(f + f_0 - F) \, \mathrm{d}F + \int_{-\infty}^{+\infty} S(F) \, P_1(f - f_0 - F) \, \mathrm{d}F + S(f + f_0) \, P_{10} +$$

$$+ S(f - f_0) \, P_{10} + S_0 \, P_1(f + f_0) + S_0 \, P_1(f - f_0) + \frac{S_0 \, P_{10}}{b} \, \delta(f - f_0) \qquad (4)$$

$$P_1(f) = P(f) \, A^2(f) \qquad (5)$$

$$P_{10} = P_0 \, A^2(0) \qquad (6)$$

$$\delta(f - f_0) = 1 \quad \text{when} \quad f = f_0$$
$$\delta(f - f_0) = 0 \quad \text{when} \quad f \neq f_0 \qquad (6a)$$

where:

- δf : r.m.s. test tone deviation (without pre-emphasis) of the wanted signal (kHz)
- f : centre-frequency of channel concerned, within the wanted signal baseband (kHz)
- f_m : upper frequency of the wanted signal baseband (kHz)
- $p(f/f_m)$: pre-emphasis factor for centre-frequency of channel concerned, within the wanted carrier baseband
- b : bandwidth of telephone channel (3.1 kHz)
- f_0 : separation between carriers of the wanted and interfering signals (kHz)
- $S(f)$: continuous part of the normalized power spectral density of the wanted signal with pre-emphasis (Hz^{-1})
- S_0 : normalized vestigial carrier power of the wanted signal
- $P(f)$: continuous part of the normalized power spectral density of the interfering signal (Hz^{-1})
- P_0 : normalized vestigial carrier power of the interfering signal
- $A(f)$: amplitude-frequency response of the wanted signal receiving filter, the origin of the frequencies being the centre frequency of the interfering signal carrier.

The power spectral densities are normalized to unity and are assumed to be one-sided (only positive frequencies).

The expression of N_p in terms of the ratio C/I is derived from expressions (2) and (3). In order to determine N_p, it is necessary to determine:

– the wanted signal spectrum (analogue telephony),

– the interfering signal spectrum.

The expressions of these spectra are given in § 2 below and in § 3 of Annex 2.

1.2 Interference from a low-modulation-index FDM/FM signal to a high-modulation-index FDM-FM signal

This case represents a terrestrial radio-relay system interfering into a system of the fixed-satellite service. The baseband channel which receives the most interference is not easily identified. However, the worst interference condition results when the wanted-to-unwanted carrier frequency separation is equal to, or less than, the top baseband frequency of the wanted signal.

The factor B can be determined from the following formula:

$$B = 10 \log \frac{1}{b\, f^2} \left\{ \frac{2(\delta f)^2\, p(f/f_m)}{P(f_0 - f) + P(f_0 + f)} \right\} \qquad (7)$$

If the modulation index of the wanted signal is greater than 3, the signal spectrum shape is near Gaussian, and formula (7) takes the following form:

$$B = 10 \log \frac{1}{b\, f^2} \left\{ \frac{2\sqrt{2\pi}\, (\delta f)^2\, p(f/f_m)\, f_s}{\exp\left[\dfrac{-(f_0 - f)^2}{2f_s^2}\right] + \exp\left[\dfrac{-(f_0 + f)^2}{2f_s^2}\right]} \right\} \qquad (7a)$$

The definitions of the parameters in formulae (7) and (7a) have been given in § 1.1 with the exception of the following:

f_s : r.m.s. multi-channel deviation of the wanted signal (kHz)

$$= \delta f \cdot 10^y \cdot (LF)^{½} \qquad (8)$$

LF : load factor, which is less than unity when not in the busy hour;

$$\begin{aligned} y &= (-15 + 10 \log N_c)/20 && \text{for } N_c \geq 240 \\ &= (-1 + 4 \log N_c)/20 && \text{for } 60 \leq N_c < 240 \\ &= (2.6 + 2 \log N_c)/20 && \text{for } 12 \leq N_c < 60 \end{aligned} \qquad (9)$$

N_c : number of voice channels in the baseband.

1.3 Interference from a high-modulation-index FDM-FM signal to a high-modulation index FDM-FM signal

The same comments as in § 1.2 apply concerning the baseband channel with the most interference and the worst frequency separation. Moreover, the factor B is identical to that given in formula (7) of § 1.2 with substitution of F_s for f_s.

F_s is defined as follows:

$$F_s = \sqrt{f_{s_1}^2 + f_{s_2}^2} \qquad (10)$$

where f_{s_1} and f_{s_2} are the r.m.s. multi-channel frequency deviations of the wanted and interfering signals (kHz).

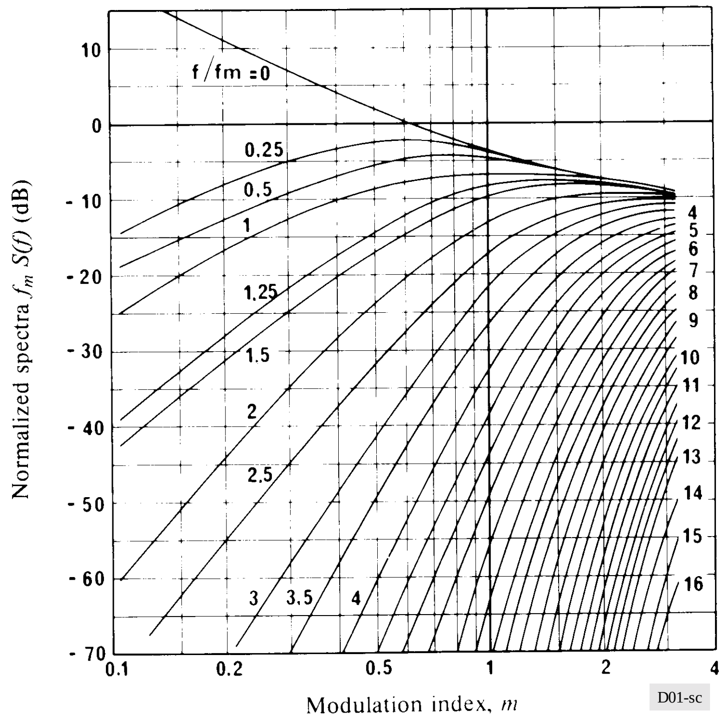
1.4 Interference between FDM-FM signals with intermediate modulation indices

Figure 1 contains a number of curves of normalized spectra as a function of the modulation index for given normalized frequency values. These curves may readily be used to plot the spectrum graph for any modulation index from 0.1 to 3. When $m > 3$, the signal spectrum shape is near Gaussian. If the modulation indices of the wanted and interfering signals are greater than 3, formula (7) should be applied to calculate interference, taking into account § 1.3.

In certain special cases, where the interfering signal may be characterized by its r.m.s. modulation index, and the upper baseband frequency is equal to the wanted signal (i.e. $f_{m_1} = f_{m_2} = f_m$) there is the possibility of calculating the interference function, $D(f, f_0)$, very simply from the normalized curves of Fig. 1.

FIGURE 1

Normalized spectral density of FDM-FM signals



The equivalent modulation index is determined by:

$$m = \left[m_1^2 + m_2^2 \right]^{1/2} \tag{11}$$

and for this value of m on the curves in Fig. 1 we find the values $f_m S(f_1)$ and $f_m S(f_2)$,

where:

$$f_1 = \frac{(f_0 + f)}{f_{m_1}} \quad \text{and} \quad f_2 = \frac{(f_0 - f)}{f_{m_1}} \tag{12}$$

and further:

$$D(f, f_0) = \frac{1}{f_{m_1}} \left[f_m S(f_1) + f_m S(f_2) \right] \tag{13}$$

The same method may be used for the approximate determination of D (f, f_0) according to the value of the "equivalent" modulation index:

$$m = \left[m_1^2 + m_2^2 \left(\frac{f_{m_2}}{f_{m_1}} \right)^2 \right]^{½} \tag{14}$$

when:

$$f_{m_2} < f_{m_1} \quad \text{and} \quad m_2^2 \left(\frac{f_{m_2}}{f_{m_1}} \right)^2 << m_1^2 \tag{15}$$

The symbols used are defined as follows:

f_0 : carrier frequency separation

f_{m_1}, f_{m_2} : mid frequency of the top baseband channel of the desired and interfering signals respectively

m_1, m_2 : r.m.s. modulation indices of desired and interfering signal respectively.

1.5 Interference from a high-modulation-index FDM-FM signal to a low-modulation-index FDM-FM signal

This case is typically that of a system in the fixed-satellite service causing interference in a terrestrial radio-relay system. Low-index angle modulation can be regarded as quasi-linear with respect to some types of interfering signal; the calculation of interference in these cases is performed by a simple procedure analogous to that employed for linear DSB-AM.

The following approximate formula can be used:

$$\left\{ \frac{\text{Interference power in telephone channel}}{\text{Thermal noise power in telephone channel}} \right\} = \left\{ \frac{\text{Interfering signal power in two appropriate 4 kHz bands at the receiver input}}{\text{Thermal noise power in same two 4 kHz bands at the receiver input}} \right\} \tag{16}$$

1.6 Interference from angle-modulated digital signals into FDM-FM signals

Digital systems using PSK or FSK modulation are classes of angle-modulated systems. Consequently, the interference from these systems into analogue, angle-modulated systems is computed by the convolution integral. However, the spectral densities of digital, angle-modulated signals cannot be easily generalized; a specific spectrum is, however, provided in § 3.2 of Annex 2. More generalized computation would involve the calculation of the digital spectral density (see § 3.2 of Annex 2), the calculation of the analogue spectral density, the convolution of the two densities, and the computation of the factor B.

When a high-modulation-index FDM-FM carrier receives interference from angle-modulated digital signals that occupy a bandwidth small compared with that of the wanted signal, factor B is given roughly by formula (7).

If a wanted FDM-FM signal suffers interference from an unwanted PCM-PSK or DPSK-PM signal that occupies a bandwidth which is large compared with that of the wanted signal, factor B is given by the following simplified formula:

$$B = 10 \log \frac{1}{b\, f^2} \left\{ \frac{2(\delta f)^2\, p(f/f_m)}{P(f_0 - f) + P(f_0 + f)} \right\} \tag{17}$$

The normalized spectral power density of the interfering signal $P(f)$ used in this formula is determined by the formulae (36a - 36d) given in § 3.2 of Annex 2.

1.7 Interference from AM signals into FDM-FM signals

The quasi-linear properties of low-modulation-index angle-modulated signals with respect to interfering signals whose spectral densities do not exhibit excessive variations within the receiver passband, permit the use for such cases of the following approximate formula:

$$\left\{ \frac{\text{Interference power in telephone channel}}{\text{Thermal noise power in telephone channel}} \right\} = \left\{ \frac{\text{Interfering signal power in two appropriate 4 kHz bands at the receiver input}}{\text{Thermal noise power in same two 4 kHz bands at the receiver input}} \right\}$$

Two 4 kHz bands are used in the formula since there may be asymmetry of the interfering spectrum with respect to the wanted carrier. When a high-modulation-index angle-modulated system receives interference from amplitude-modulated digital signals that occupy a bandwidth small compared with that of the wanted signal, factor B is given roughly by the formula of § 1.2.

1.8 Interference from a narrow-band system into an FDM-FM system

The theoretical expression of § 1.1 can be applied to the case of an interfering signal of arbitrary modulation, but with a bandwidth small compared with that of the interfered-with signal. Interference from SCPC to FDM-FM signals is an example of such a situation.

In particular, for evenly spaced SCPC carriers, the aggregate interference power in the baseband from all SCPC interference entries from one interfering network is close to thermal noise with equal power starting from five to six carriers.

1.9 Interference from FM-TV signals into FDM-FM signals

When the FM-TV signal modulated only by the dispersal waveform is the interfering signal, the FDM-FM wanted signal with a low number of telephone channels has a spectrum with a width commensurate with that of the interfering signal spectrum, and the carrier frequencies coincide, then formula (4) takes the form:

$$D(f, 0) = P \left[\int_{f-\Delta f/2}^{f+\Delta f/2} S(F)\,\mathrm{d}F - \int_{f+\Delta f/2}^{f-\Delta f/2} S(F)\,\mathrm{d}F \right] = 2P \int_{f-\Delta f/2}^{f+\Delta f/2} S(F)\,\mathrm{d}F \qquad (18)$$

where:

Δf : frequency deviation of dispersal waveform (peak-to-peak)

P : spectral power density of interfering signal (see Fig. 4, $i = 1$)

$= 1/\Delta f$.

In the conditions described above, and with reference to formula (3), we may consider that:

$$\int_{f-\Delta f/2}^{f+\Delta f/2} S(F)\,\mathrm{d}F \approx 1 \qquad \text{when} \qquad f < f_{m_1} \qquad (19)$$

so that:

$$B = 10 \log \frac{(\delta f)^2\, \Delta f\; p\,(f/f_m)}{f^2\, b} \qquad (20)$$

1.10 Residual tone interference of FDM-FM system with low modulation index

Special attention should be given to the severe effects of tone interference due to the residual carrier in FDM-FM systems with a low modulation index. The interference noise power $P_{\varepsilon\delta}$ at the point of zero relative level is given by:

$$P_{\varepsilon\delta} = \frac{1}{2m_1^2} 10^{\frac{1}{10}y} \sum_{n=1}^{\infty} \frac{k^{2n}}{n^2} D_n \frac{\sigma^2}{p(\sigma)} \delta(\sigma - n\sigma_d)_I \tag{21a}$$

where:

$k = (I/C)^{½}$

$D_n = e^{-nm_e^2}$ for unmodulated carrier interference (21b)

$= e^{-2nm_e^2}$ for same type-modulated FDM-FM interference (21c)

$$m_e^2 = \left[C_0 + (C_2/\beta) + \frac{C_\mu}{3\beta^3} (\beta^2 + \beta + 1) \right] m_{e0}^2 \tag{21d}$$

$$m_{e0}^2 = m_1^2 \beta \tag{21e}$$

β : upper to lower frequency ratio in the wanted signal baseband

$$\sigma = f / f_{m_1} \tag{21f}$$

$$p(\sigma) = C_0 + C_2(\sigma)^2 + C_\mu(\sigma)^4 \tag{21g}$$

$$\delta(\sigma - \sigma_d)_I = 1 \quad \text{for } \sigma = \sigma_d \tag{21h}$$

$$= 0 \quad \text{for } \sigma \neq \sigma_d$$

The influence of tone interference tends to become substantial in cases where the frequency separation f between carriers of the wanted and interfering signals falls into the baseband or particular regions, such as the pilot, the frequency slot of detected noise, a possible super-baseband service channel and DAV subscriber. In this case, the design requirements with respect to interference should be determined in the light of the tone interference.

2. Signal spectra

2.1 Analogue telephony (FDM-FM) signal

The signal's normalized power spectral density centred on the carrier frequency is expressed as:

$$P(f) = e^{-a} \left[\delta(f) + \sum_{n=1}^{\infty} \frac{m^{2n}}{n!} S(f)^n * S(f) \right] \tag{22}^{*}$$

where:

$\delta(f)$: Dirac delta function

$S(f)_n * S(f)$: convolution of the function $S(f)$ n times itself

$S(f)$: normalized spectral density of the signal phase:

$$S(f) = \frac{f_m \, p \, (f/f_m)}{2f^2 \, (1 - \varepsilon)} \tag{23}$$

where ε is the lower to upper frequency ratio in the wanted signal baseband.

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* Although the series of formula (22) converges for all values of system parameters, it does not always provide the most appropriate algorithm for numerical computation, particularly in cases where the normalized r.m.s. multi-channel phase and/or frequency deviation (a and m respectively) are large.

The ITU-R pre-emphasis characteristics are well approximated by the expression:

$$p\,(f/f_m) = 0.4\,00 + 1.35\left(\frac{f}{f_m}\right)^2 + 0.75\left(\frac{f}{f_m}\right)^4, \quad \text{when} \quad \varepsilon \le \frac{f}{f_m} \le 1 \tag{24}$$

Here:

$$a = R_s(0) - R_s(\infty) \simeq \frac{m^2}{\varepsilon}\,(0.4 + 1.6\,\varepsilon + 0.25\,\varepsilon^2 + 0.25\,\varepsilon^3) \approx \frac{m^2}{\varepsilon}\,(0.4 + 1.6\,\varepsilon) \tag{25}$$

where:

$R_S(\tau)$: the autocorrelation function of $S(f)$.

The normalized power of the vestigial signal carrier is expressed as e^{-a}.

When $m > 1$:

$$P(f) = \frac{1}{f_s\sqrt{2\pi}}\, e^{-\frac{f^2}{2f_s^2}} \left\{ 1 + \sum_{n=2}^{\infty} (-1)^n \frac{C_{2n}}{m^{2n}\,2^n}\, H_{2n}\left(\frac{f}{f_s\sqrt{2}}\right) \right\} \approx$$

$$\approx \frac{1}{f_s\sqrt{2\pi}}\, e^{-\frac{f^2}{2f_s^2}} \left\{ 1 + \frac{6.375 \cdot 10^{-2}}{m^2} H_4^*\left(\frac{f}{f_s\sqrt{2}}\right) - \frac{7.416 \cdot 10^{-3}}{m^4} H_6^*\left(\frac{f}{f_s\sqrt{2}}\right) + \right.$$

$$\left. + \left(\frac{2.37 \cdot 10^{-2}}{m^4} + \frac{7.16 \cdot 10^{-4}}{m^6}\right) H_8^*\left(\frac{f}{f_s\sqrt{2}}\right) - \left(\frac{9.929 \cdot 10^{-3}}{m^6} + \frac{5.854 \cdot 10^{-5}}{\mathrm{m}^8}\right) H_{10}^*\left(\frac{f}{f_s\sqrt{2}}\right) \right\} \tag{26}$$

where:

f_S : r.m.s. multi-channel signal frequency deviation

$H_{2n}^*(x) = (-1)^n \dfrac{n!}{(2n)!} H_{2n}(x)$: normalized Hermite polynomial.

Figures 2a to 2e contain spectral graphs plotted according to formulae (22) and (26) for modulation indices m adopted in typical radio-relay and communication-satellite systems.

The curves are approximate in the region f/f_m near 0 and 1. The exact values depend upon the particular value of ε. The exact curves for several values of ε are given in Figs. 2f to 2j for f/f_m near zero. (The inset curves in Figs. 2d and 2e are also accurate enough for f/f_m near zero if ε is equal to or greater than 0.02.)

For modulation indices greater than 1.1, the following empirical formula has been found to fit adequately the curves of $P(f)$ and is a good approximation of equation (26):

$$f_m \cdot P(f) = \frac{1}{m\sqrt{2\pi}}\, e^{-\frac{x^2}{2m^2(1 + 0.01337\,x^2 \cdot m^{-3;.67})}} \tag{26a}$$

where:

$x = f/f_m$.

This empirical formula is an adaptation of the Gaussian formula for large modulation indices.

FIGURE 2a

Normalized power spectral densities for various modulation indices

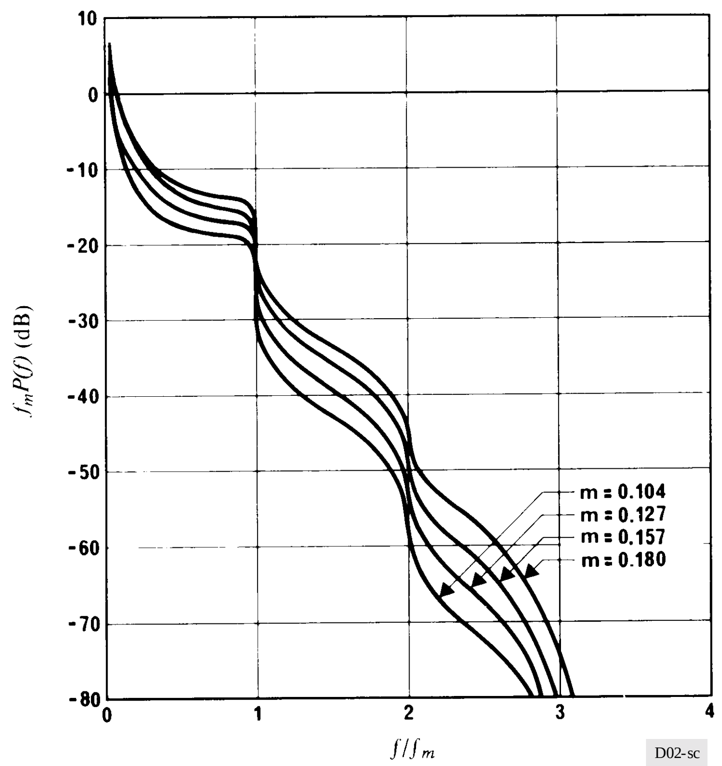


FIGURE 2b

Normalized power spectral densities for various modulation indices

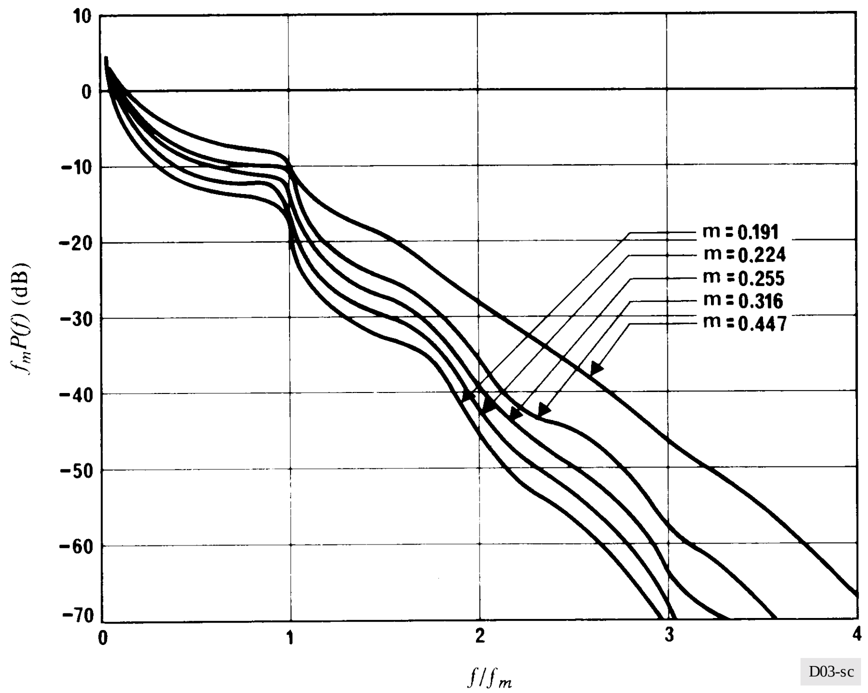


FIGURE 2c

Normalized power spectral densities for various modulation indices

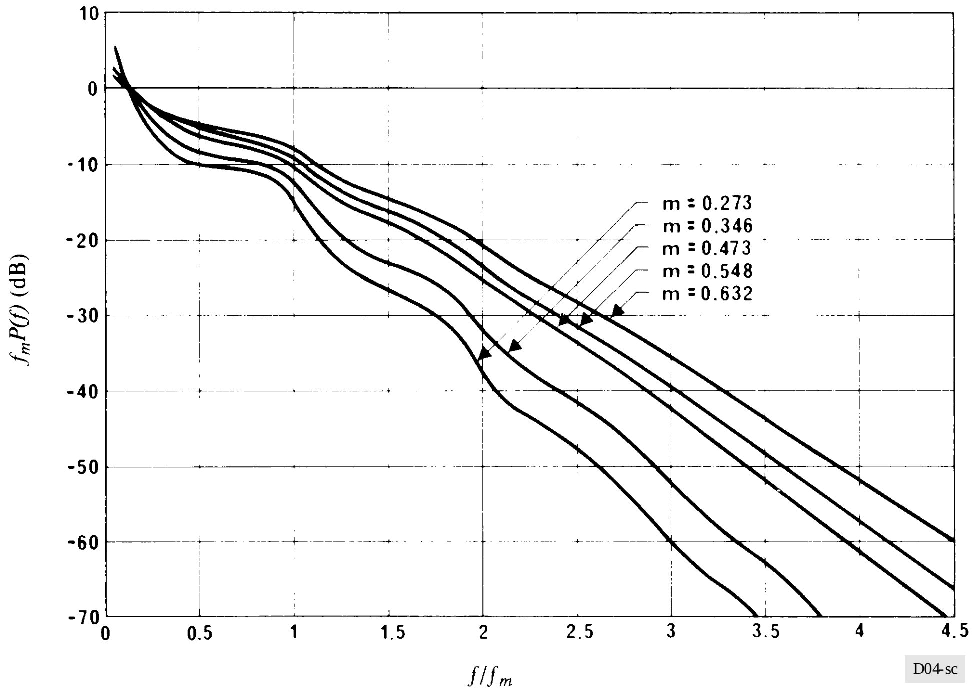


FIGURE 2d

Normalized power spectral densities for various modulation indices

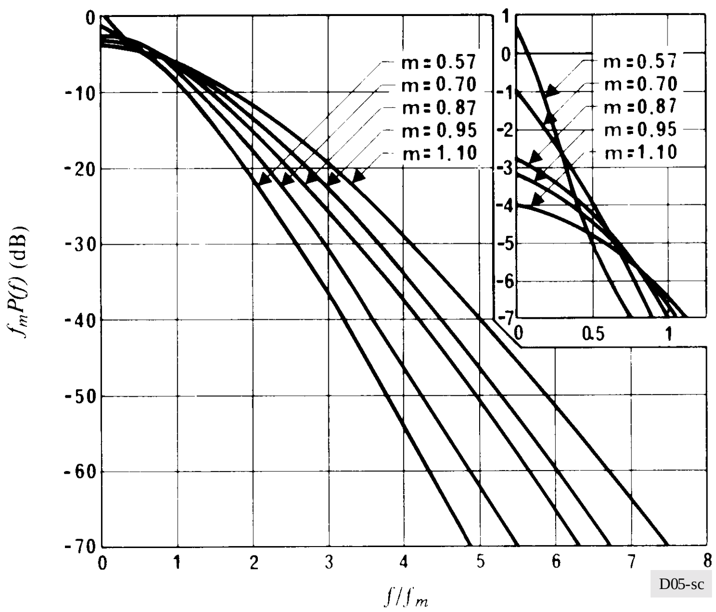


FIGURE 2c

Normalized power spectral densities for various modulation indices

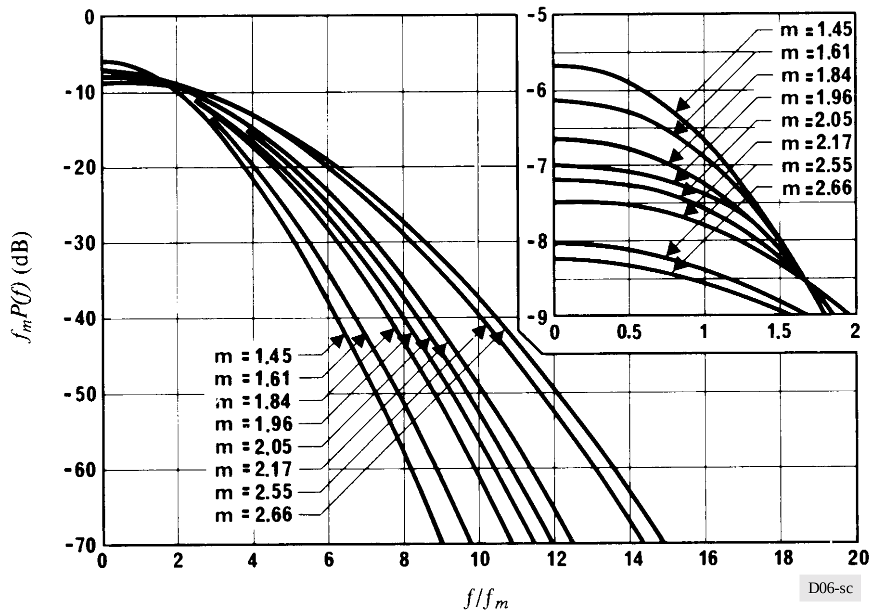
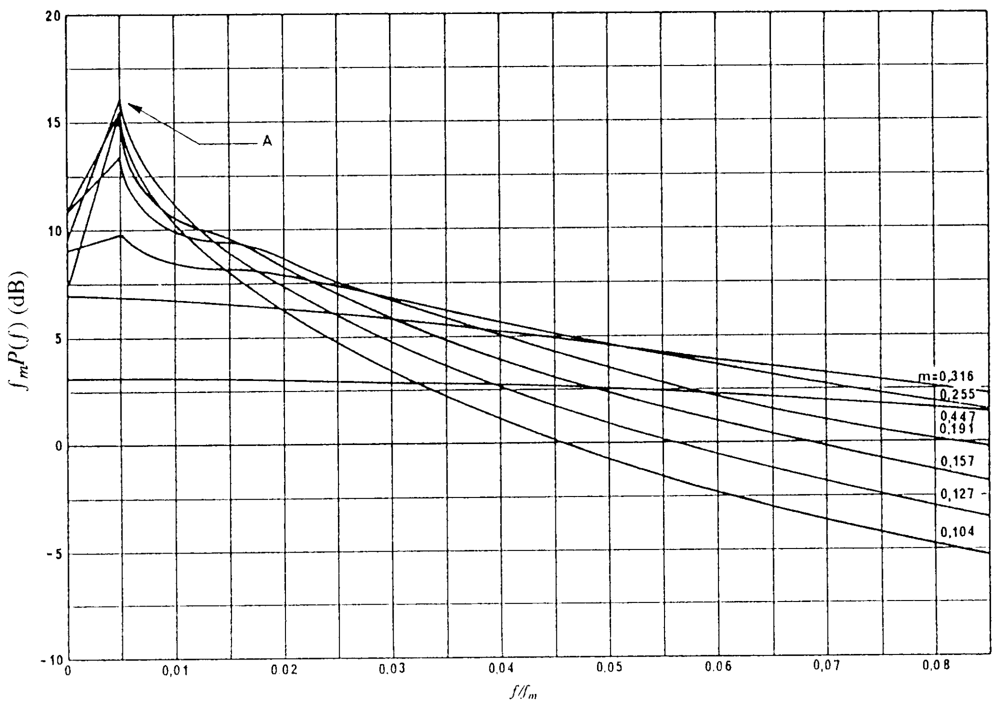


FIGURE 2f

Normalized power spectral densities for various modulation indices and for ε = 0.005

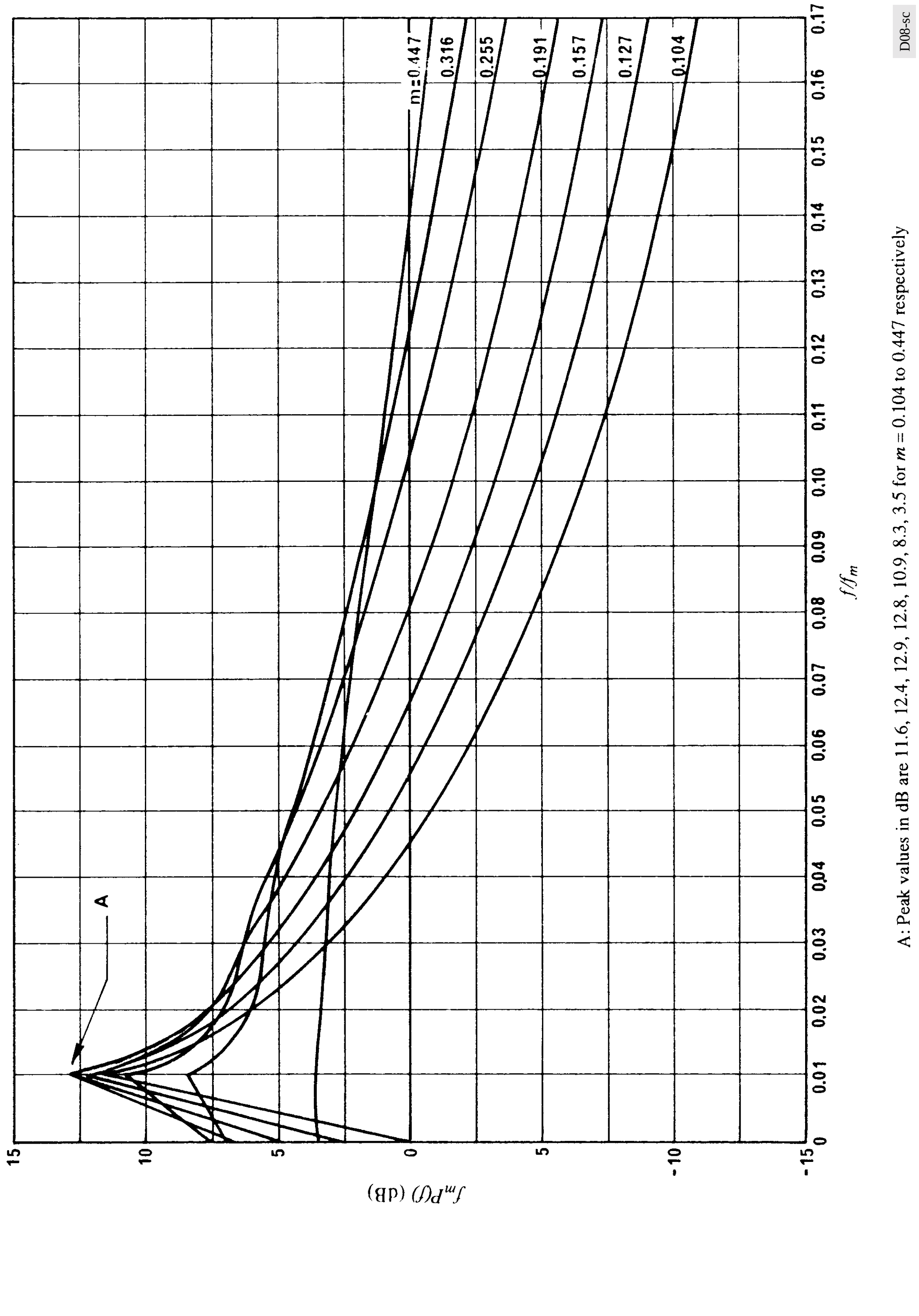


A: Peak values in dB are15.9, 15.9, 15.2, 13.6, 9.8, 7.1, 3.2
for m = 0.104 to 0.447 respectively

D07-sc

FIGURE 2g

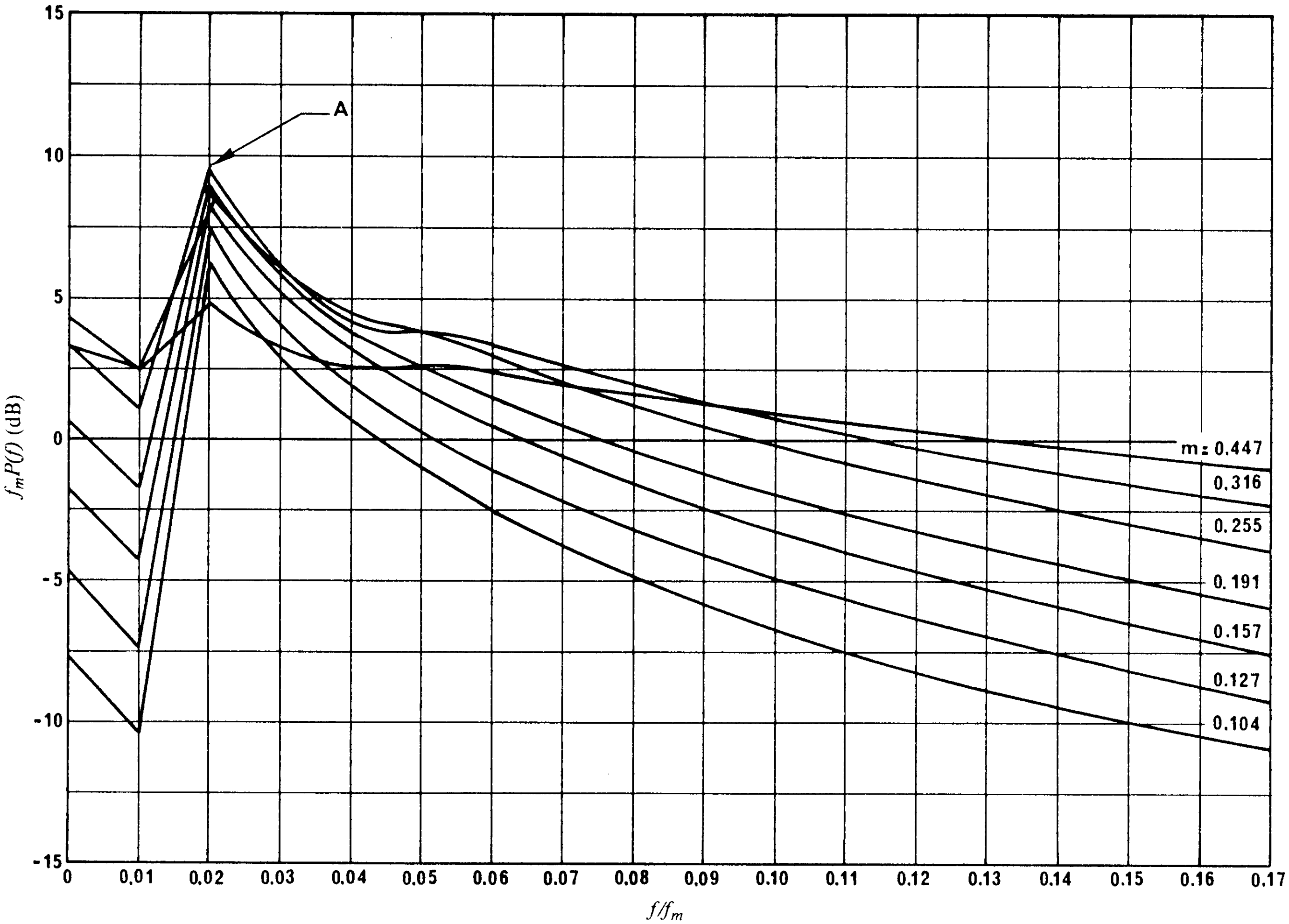
Normalized power spectral densities for various modulation indices and for ε = 0.01



A: Peak values in dB are 11.6, 12.4, 12.9, 12.8, 10.9, 8.3, 3.5 for m = 0.104 to 0.447 respectively

FIGURE 2h

Normalized power spectral densities for various modulation indices and for ε = 0.02

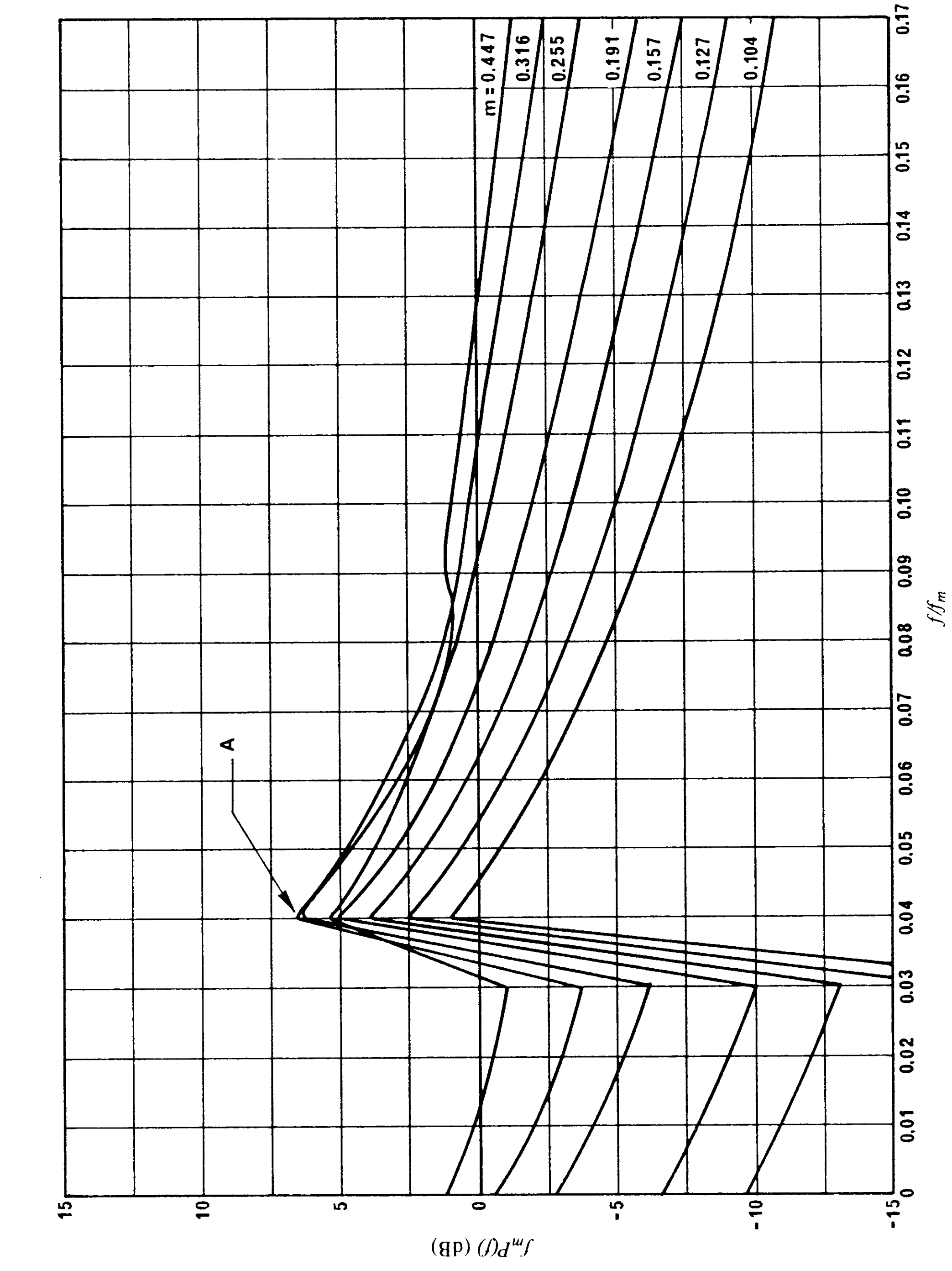


A: Peak values in dB are 6.5, 7.7, 8.8, 9.5, 9.6, 8.7, 4.8 for m = 0.104 to 0.447 respectively

D09-sc

FIGURE 2i

Normalized power spectral densities for various modulation indices and for ε = 0.04

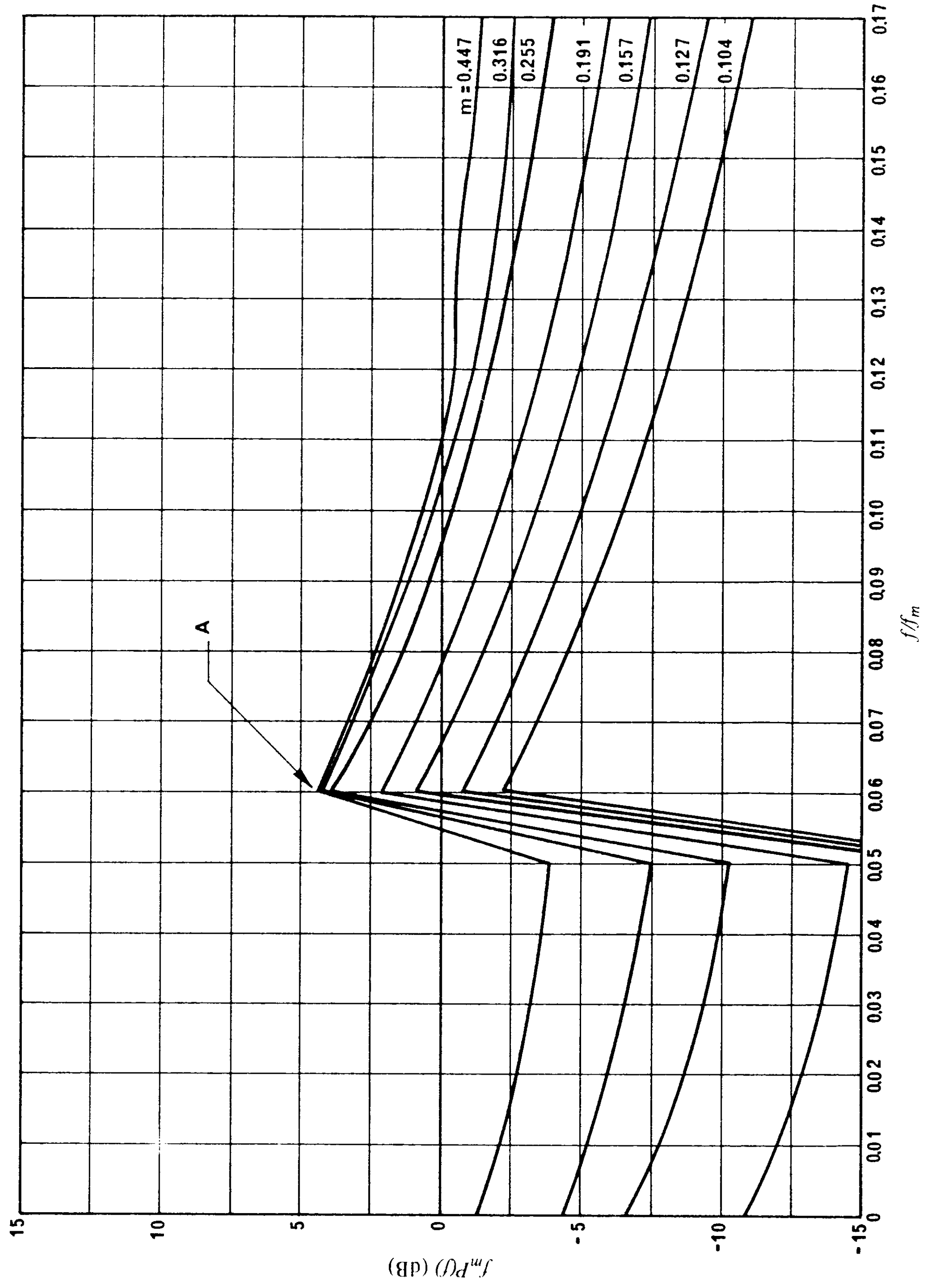


D10-sc

A: Peak values in dB are 1.0, 2.5, 3.9, 5.1, 6.3, 6.5, 5.2 for m = 0.104 to 0.447 respectively

FIGURE 2j

Normalized power spectral densities for various modulation indices and for ε = 0.06



A: Peak values in dB are −2.3, −0.7, 0.9, 2.2, 3.7, 4.4, 4.3 for m = 0.104 to 0.447 respectively

D11-sc

ANNEX 2*

Calculation methods for interference to systems other than FDM-FM systems

Given below is the method of calculation for wanted signals other than FDM-FM

1. General

Formulae and/or graphs in which the degradation due to interference can be readily seen are presented for most cases. Spectra of signals are also provided to allow determination of interference from the general formulation, and to assist in calculations of power densities used in Recommendation ITU-R SF.675.

Further studies are necessary on analogue single-sideband (SSB) telephony, companded SSB, companded FDM-FM, hybrid data-and-voice (DAV) and data-and-video (DAVID), multiplexed analogue component television MAC-TV, high definition TV (HDTV), time division multiplex access (TDMA), spread spectrum code division multiple access (CDMA) signals, etc.

The performance degradation of analogue telephony transmission can be expressed in terms of noise (pW) and unavailability. In the case of digital transmission, it can be expressed in terms of bit error ratio (BER), severely errored seconds, degraded minutes and unavailability. In the case of FM television, the expressions given in this Annex make it possible to estimate the permissible value of carrier-to-interference ratio.

Precautionary notes are included with respect to interference effects which are not predictable by determination based on spectra and with respect to non-linear channel effects.

2. Interference formulations

2.1 Single-channel-per-carrier FM telephony wanted signal

Further studies are required on this item.

2.2 Frequency-modulated television wanted signal

A protection ratio R which can be introduced, represents the carrier-to-interference ratio corresponding to a given impairment. As a result of tests carried out in France, with the interfering signal being an unmodulated carrier, the values of R given in Fig. 3 are expressed as a function of the frequency separation f_0 between the wanted and interfering signal carriers. The curve in Fig. 3, composed of two straight line segments and two half-lines, is an empirical curve plotted from test data (ΔF = frequency deviation in the low frequencies of the wanted signal (MHz)).

The subjective interference level chosen was that corresponding to the perceptibility threshold without thermal noise, for an observer placed in a dimly lit room at a distance from the screen equal to six times the height of the picture.

The permissible value $(C/I)_a$ of this ratio is obtained from the expression:

$$\left(\frac{C}{I}\right)_a = \int_{-\infty}^{+\infty} R(f - f_0)\, A(f)\, [P(f) + P_0\, \delta]\, \mathrm{d}f \qquad (27)$$

where $P(f)$, P_0 and $A(f)$ have the same meaning as in § 1.1 of Annex 1.

Calculation of $(C/I)_a$ can be performed once the interfering spectrum is specified (see § 3).

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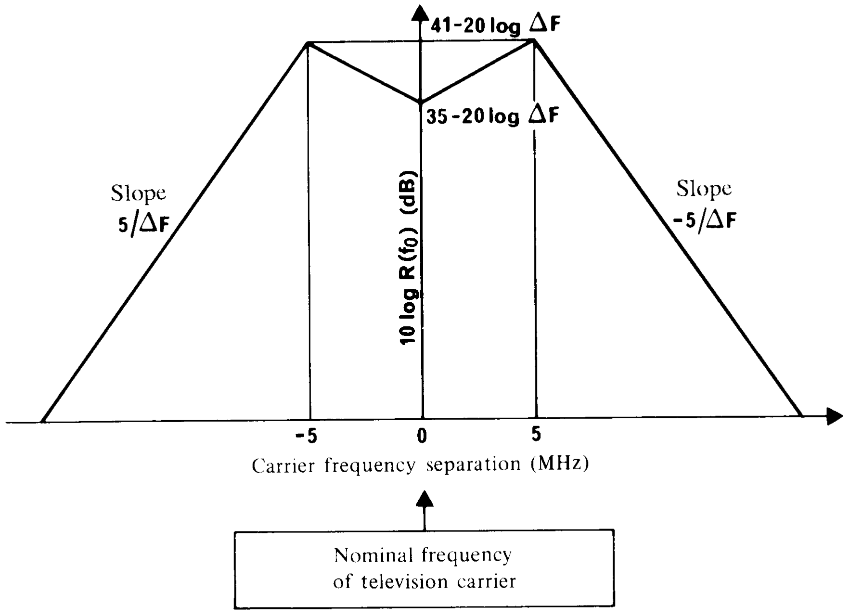
* *Note from the Director CCIR* – For information, derivation of the formulae and historical development of this Annex are given by References contained in CCIR Report 388 (Düsseldorf, 1990).

FIGURE 3

Protection ratio *R* (dB)



Wanted signal: frequency-modulated TV carrier
Interfering signal: pure carrier

ΔF : frequency deviation in the low frequencies of the wanted signal (frequency-modulated TV) (MHz)

D12-sc

2.3 Amplitude modulated telephony wanted signal

2.3.1 General information

Further studies are required on this item.

2.3.2 Interference between amplitude-modulated signals

The K_4 factor is defined as the amount (dB) by which the signal-to-interference power ratio exceeds the ratio of the signal spectral density in the appropriate 4 kHz band at the receiver input to the interference density at the same bandwidth.

In consequence of the property of linear modulation of translating interfering signals directly to baseband, the value of the factor K_4 is simply 0 dB for SSBSC, and 3 dB for DSBSC.

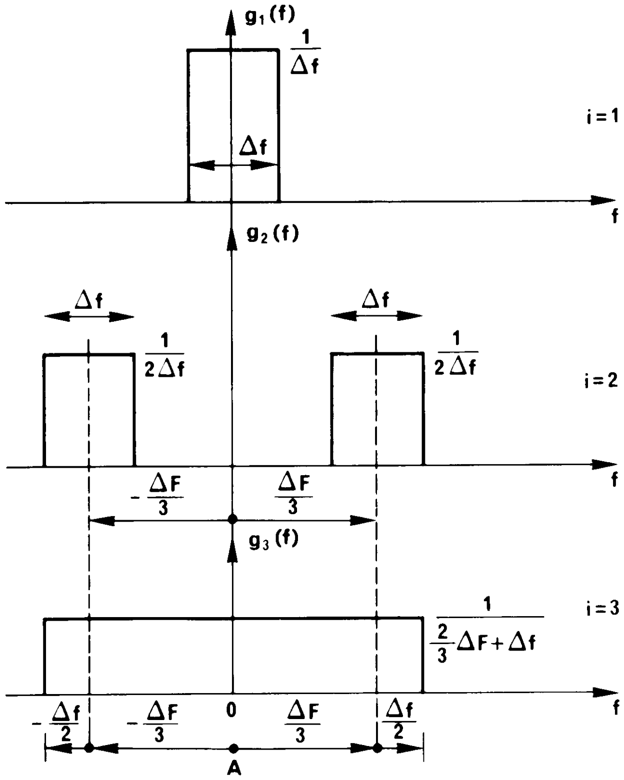
2.3.3 Interference to amplitude-modulated signals from angle-modulated signals

The values of factor K_4 are again 0 dB for SSBSC, and 3 dB for DSBSC.

The baseband spectrum of the interference will be identical with that of the RF interfering spectrum in the SSBSC case, and with the sum of the RF interfering spectra falling on the upper and lower sidebands in the DSBSC case. As a result, angle-modulated interference with strong carriers will generate tone-interference at baseband. The channel arrangement of AM systems will generally need to take account of this mode of interference.

FIGURE 4

Frequency-modulated television signal



Models used to represent the central part of the spectrum

A: Nominal frequency

D13-sc

2.4 Digital wanted signal

The expressions for the performance of non-coded coherent digital modulation systems in a Gaussian channel are well-known. However, in practice a perfect Gaussian channel environment rarely occurs. The received signal is a random process consisting of two components, the first contributed by thermal white Gaussian noise and the second by all other sources such as co-channel interference (CCI), adjacent channel interference (ACI) and intersymbol interference (ISI). The effects on the error probability performance due to this interference can be obtained in principle. Possible methods include direct calculation/simulation, numerical method with computer simulation Gram-Charlier series, Gaussian quadrature rule, complex integration, and bound methods.

2.4.1 Gaussian interference environment

The probability of error performance of binary phase shift keying (BPSK = 2 PSK), M-ary PSK (MPSK) ($M > 2$), M-ary quadrature amplitude modulation (M-QAM) and M-ary quadrature partial response (M-QPR) and differential BPSK (DBPSK) modulation schemes is given by the following expressions:

$$P_B \doteq \frac{1}{2}\, erfc\left(\sqrt{\gamma_b}\right) \qquad \text{(BPSK)} \qquad (28)$$

$$P_{DBPSK} = \frac{1}{2}\, e^{-\gamma_b} \qquad (29)$$

$$P_M \doteq erfc\left(\sqrt{k\gamma_b}\,\sin\frac{\pi}{M}\right) \qquad \text{(MPSK, } M > 2\text{)} \qquad (30)$$

$$P_L \doteq \left(1 - \frac{1}{L}\right) erfc\left(\sqrt{\frac{3}{M-1}\,\frac{1}{2}\,\gamma_{av}}\right) \qquad \text{(M-QAM)} \qquad (31a)$$

$$P_L \doteq \left(1 - \frac{1}{M}\right) erfc\left(\sqrt{\frac{3}{M-1}\,\frac{1}{2}\left(\frac{\pi}{4}\right)^2\gamma_{av}}\right) \qquad \text{(M-QPR)} \qquad (32a)$$

$$P_M = 2P_L\left(1 - \frac{1}{2}\,P_L\right) \qquad \text{(M-QAM and M-QPR)} \qquad (31b)\ (32b)$$

where:

$$\gamma_b = \frac{E_b}{N_0} = \frac{C}{N}\,\frac{B}{f_b} \qquad (33)$$

P_B : probability of error performance of BPSK system

P_M : symbol error rate for the MPSK ($M > 2$), M-QAM and M-QPR systems

P_L : probability of error of the baseband signal in each of the two quadrature components of QAM or QPR system

P_{DBPSK} : probability of error performance of differential BPSK system

γ_{av} : average signal-to-noise ratio per *k*-bit symbol

$k =$ log *M* where *M* is the number of states

L : number of baseband levels, i.e. $M = L^2$

γ_b : energy per bit-to-noise ratio

C/N : carrier-to-thermal noise ratio

f_b : bit rate (bit/s)

B : double-sided noise bandwidth (Hz). We assume *B* is equal to the double-sided Nyquist bandwidth.

Expressions (28), (30) and (31) give the probability of error performance curves illustrated in Fig. 5.

In the case of M-QAM systems, which employ very tight filtering (such as Nyquist raised-cosine) the interference may be treated as Gaussian-like noise. A victim M-QAM receiver may be interfered with from one or several sources. The amplitude distribution of a tightly filtered interferer exhibits a high peak-to-average ratio which could be approximated by an equivalent Gaussian-like noise source. In the case of several interferers the sources of interference are considered to be independent random variables. The central-limit theorem says that, under certain general conditions, the resultant equivalent interference probability density function approaches a normal Gaussian curve as the number of sources increases. In both the single and multiple interference cases, the equivalent interference may be treated as Gaussian-like noise. This practical approach yields useful performance curves in which the degradation due to interference can be readily seen.

The Gaussian-like interference is combined with the assumed white Gaussian noise channel to produce a total carrier-to-noise ratio $(C/N)_T$ given by:

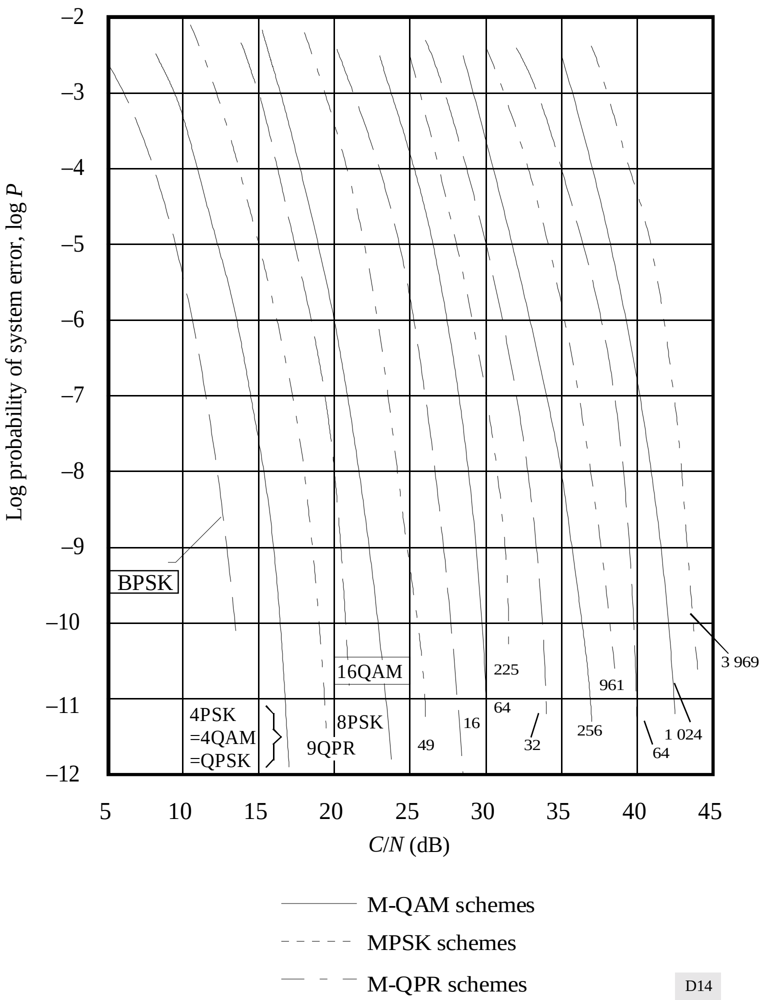
$$(C/N)_T = (N/C + I/C)^{-1} \qquad (34a)$$

$$I/C = I_1/C + I_2/C + \ldots + I_n/C \qquad (34b)$$

where N/C is the thermal noise-to-carrier ratio, I/C is the equivalent interference-to-carrier ratio, and I_i/C ($i = 1, \ldots, n$) is the interference-to-carrier ratio of the *i*-th random source. Expressions such as equations (28) to (31) are used to calculate performance of coherent digital modulation systems in the presence of interference by replacing C/N by $(C/N)_T$ and having C/I as a variable parameter. Inclusion of C/I as a variable parameter produces a series of curves shown in Figs. 6 and 7. The degradations (dB) of $(C/N)_T - C/N$ for $P_e = 10^{-6}$ versus C/I for the M-QAM systems are summarized in Fig. 8. If a carrier-to-interference ratio is at least 10 dB higher than a carrier-to-thermal noise ratio required for the $P_e = 10^{-6}$, the degradation due to interference will be less than 1 dB. Although not shown in Fig. 8, it can be calculated that if the C/I is at least 6 dB higher than a C/N for $P_e = 10^{-3}$, the degradation due to interference will be less than 1 dB.

FIGURE 5

The average probability of error performance of the M-ary schemes versus the carrier-to-thermal noise ratio (measured in double-sided Nyquist bandwidth) for a white Gaussian channel only)



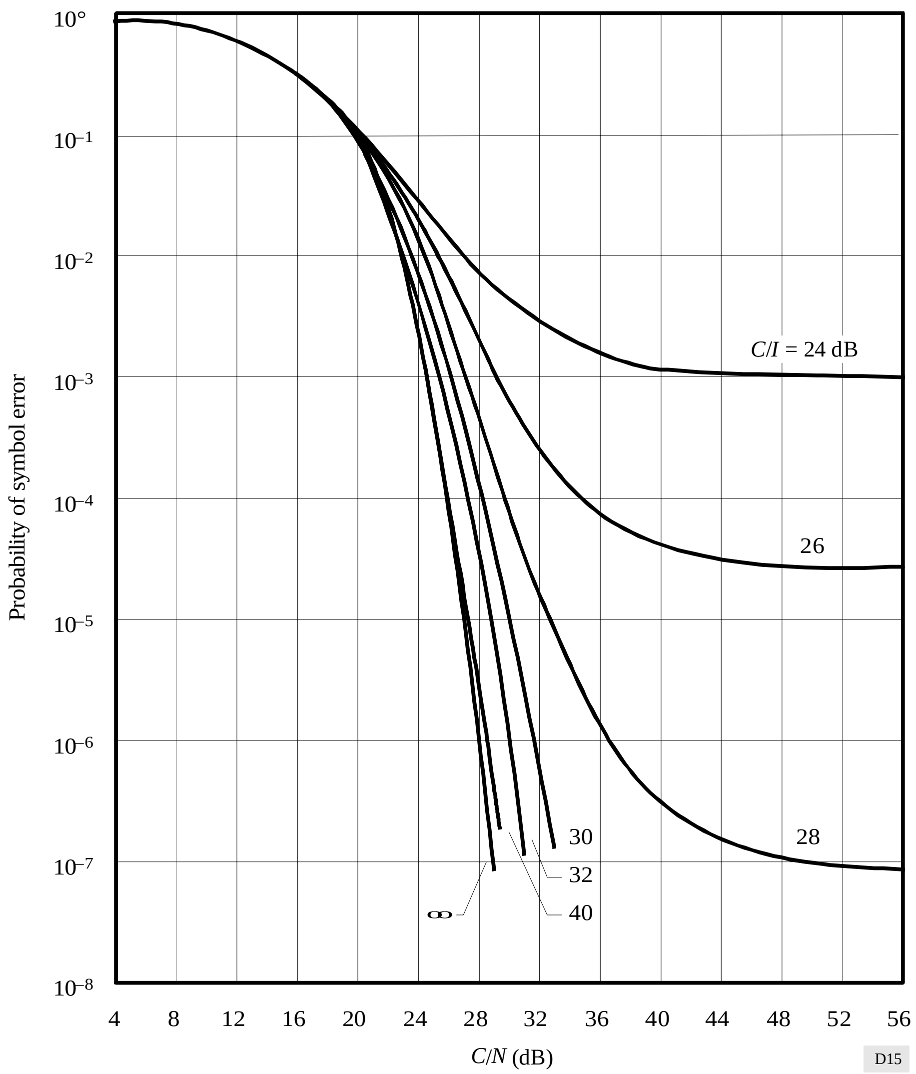
However, Gaussian-like interference is not necessarily the worst case. Administrations are urged to study new methods for determining the mutual effects of interference between M-QAM and other digital and analogue schemes.

2.4.2 Gauss quadrature rule method

This section presents performance curves for bit error ratios of 10^{-3} and 10^{-6} and C/I and C/N as variables. These curves are obtained by using the Gauss quadrature rule method. These curves refer to co-adjacent and channel interference in 4-PSK, 16-QAM and 64-QAM carriers due to several types of interfering carriers. Transmit and receive filters of wanted and interfering systems are assumed to have square root raised cosine transfer functions with roll-off factors of 0.4 and 0.5 typical for satellite systems. The transmit filter also includes an aperture equalizer to achieve the intersymbol interference free condition. A curve obtained assuming Gaussian interference (see § 2.4.1) is also indicated in each figure, for comparison purposes.

FIGURE 6

The probability of error performance curves of 64-QAM modulation system versus carrier-to-thermal noise ratio and a carrier-to-interference ratio as a parameter (double-sided Nyquist bandwidth)

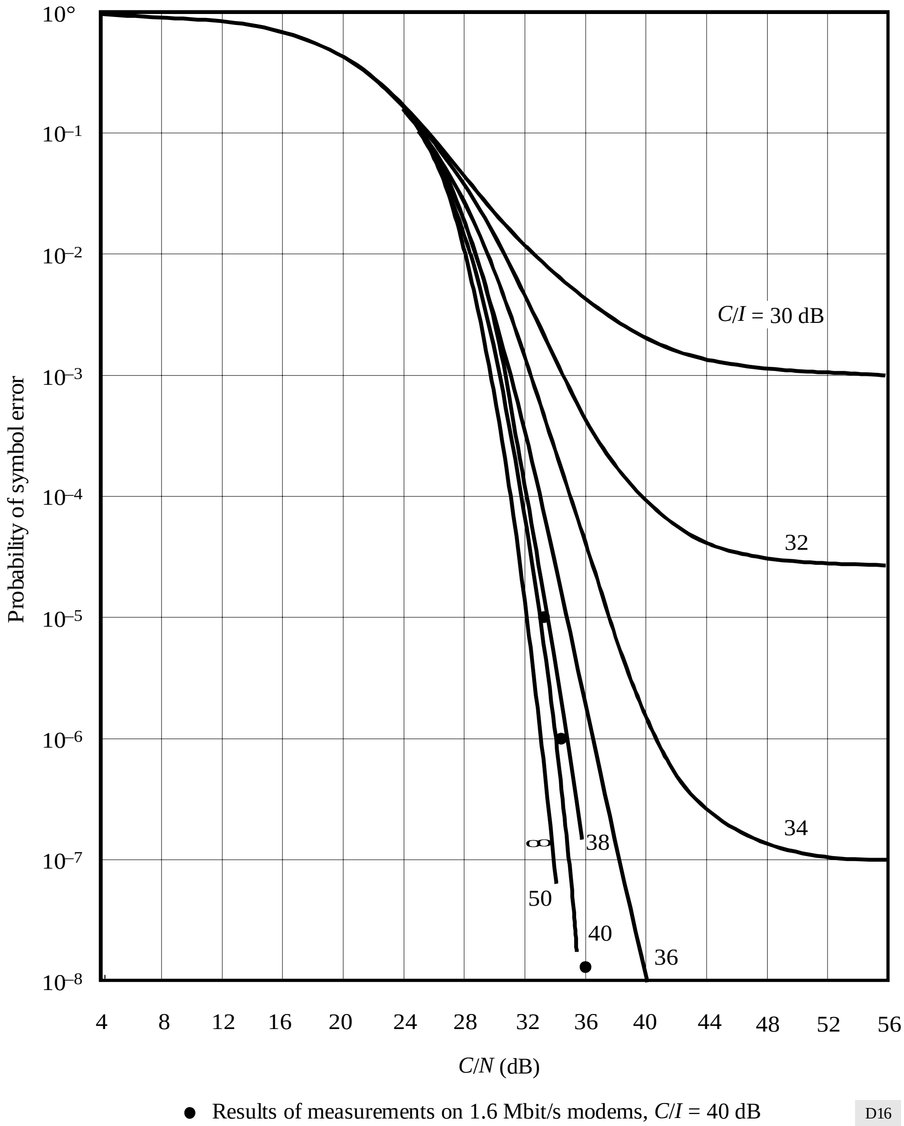


In Figs. 9 to 13, *C*/*I* ′ is defined as the ratio between carrier power at the receive filter input and interference power at the receive filter output. Carrier-to-interference power ratio at the receive filter input can be determined by subtracting the corresponding interference reduction factor which is given in the figures. Also, in these figures, *C*/*N* represents the ratio between the carrier power at the receive filter input and the noise power at the receive filter output. The carrier-to-noise ratio at the receive filter output is around 0.5 dB lower because of the attenuation of the spectrum of the desired carrier by the receive filter.

Figures 9 to 11 refer to co-channel and adjacent channel interference in systems with 4-PSK, 16-QAM and 64-QAM modulations, considering several values of frequency separation between two equally modulated carriers. Figure 12 addresses co-channel interference between two 4-PSK carriers with different relative bandwidths. Figure 13 shows the interference effect of different modulations on the performance of a 4-PSK system.

FIGURE 7

The probability of error performance curves of 256-QAM modulation system versus carrier-to-thermal noise ratio and a carrier-to-interference ratio as a parameter (double-sided Nyquist bandwidth)



The following general conclusions can be drawn by inspection of the figures:

– when the interfering signal power is equal to, or larger than, the thermal noise power, the effect of angle-modulation interference is considerably less than that of an equal amount of white Gaussian noise power;

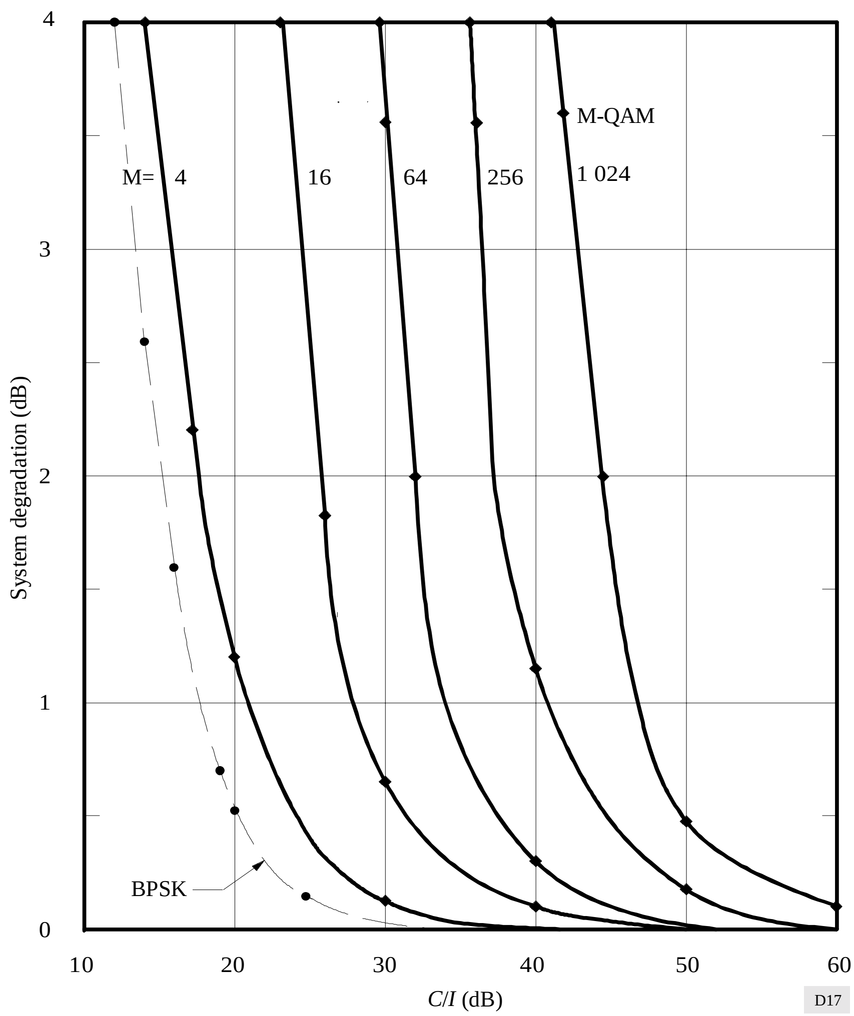
– when the interfering signal power is small compared to the thermal noise power, the effect on error rate can be estimated safely by assuming that the interfering signal is equivalent to Gaussian noise of equal power;

– at a given carrier-to-interference ratio, the vulnerability to interference increases substantially as the number of transmitted symbols, *M*, increases;

– for the same interfering power after filtering, interference effects tend to become larger as the frequency separation between carriers increases. These effects also tend to increase with the interfering carrier bandwidth and with the number of interfering carriers. They are approximately the same for an interfering 4-PSK or 8-PSK carrier but they increase with the number of symbols for a QAM interfering signal. All the above situations can be interpreted in terms of an increase in the interference peak factor: for large values of frequency separation (adjacent channel interference), for large values of interfering carrier bandwidth and for a large number of interfering carriers, the interference effect approaches that of an equal amount of white Gaussian noise.

FIGURE 8

The degradation of the M-QAM systems (dB) with respect to the theoretical value necessary to achieve the $P_e = 10^{-6}$ performance, versus the carrier-to-interference ratio (dB) in the double-sided Nyquist bandwidth



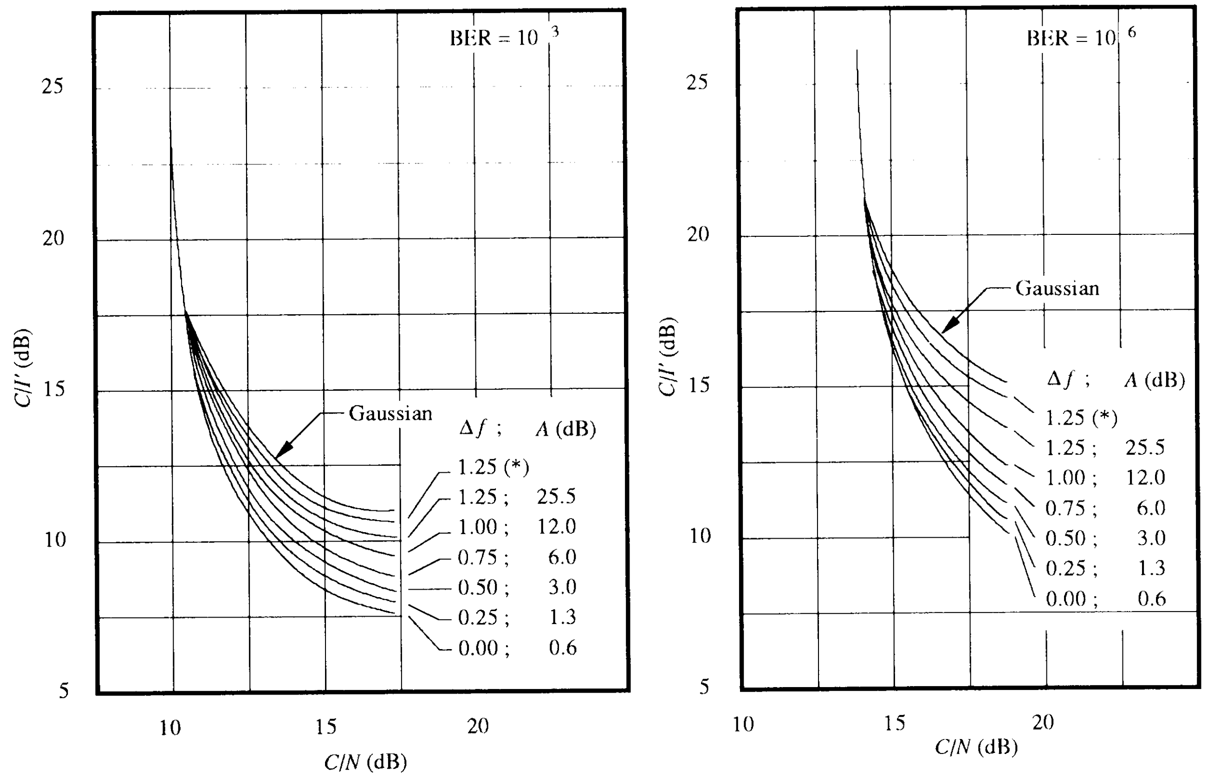
2.4.3 Numerical method with computer simulation

A computation method (numerical method with computer simulation) can be used to evaluate the performance of general PSK and QAM multi-state modulation schemes staggered, and non-staggered, in an environment of additive noise, interference and distortion, including modified constellation QAM schemes. Figure 14 shows the computation results of single interference for a 16-QAM system. The curves represent the results obtained from the computation method mentioned above. For purposes of comparison, the symbol + on the curves in Fig. 14 has been used to indicate the results obtained by using pure numerical computation of a series expansion technique. They are in good agreement with each other.

This method can also be used for analysis of the combined effects of interference and implementation distortion.

FIGURE 9

***C/I'* versus *C/N* for 4-PSK transmission with 4-PSK interfering carriers for 10^{-3} and 10^{-6} BER**



Δf: frequency separation between carriers (normalized with respect to the symbol rate)

A (dB) : interference reduction factor defined as the ratio between the interfering powers at the input and at the output of the receive filter

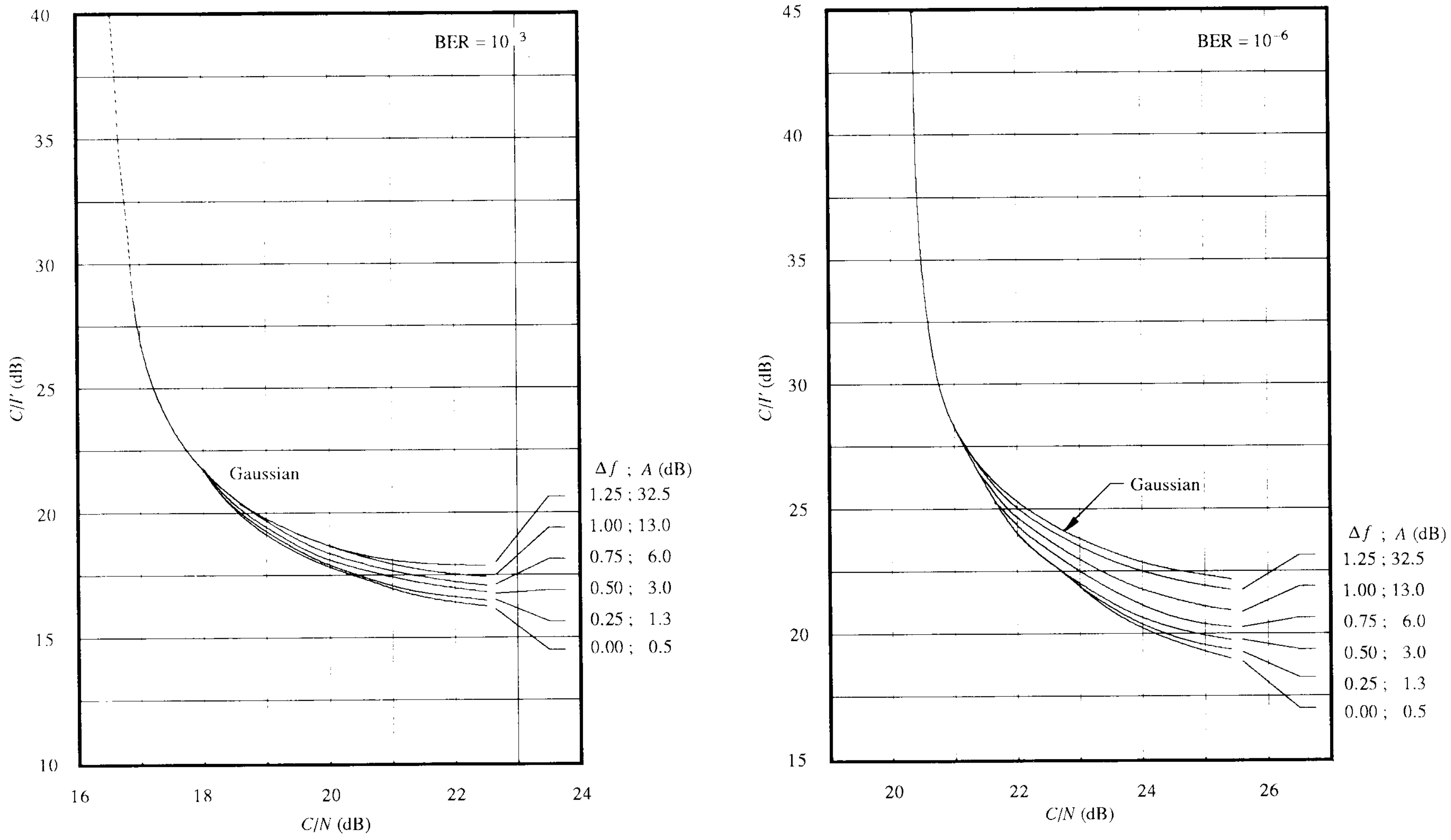
Roll-off = 0.5

* 2 adjacent carriers.

D18-sc

FIGURE 10

***C/I'* versus *C/N* for 16-QAM transmission with 16-QAM interfering carriers for 10^{-3} and 10^{-6} BER**



Δf : frequency separation between carriers (normalized with respect to the symbol rate)

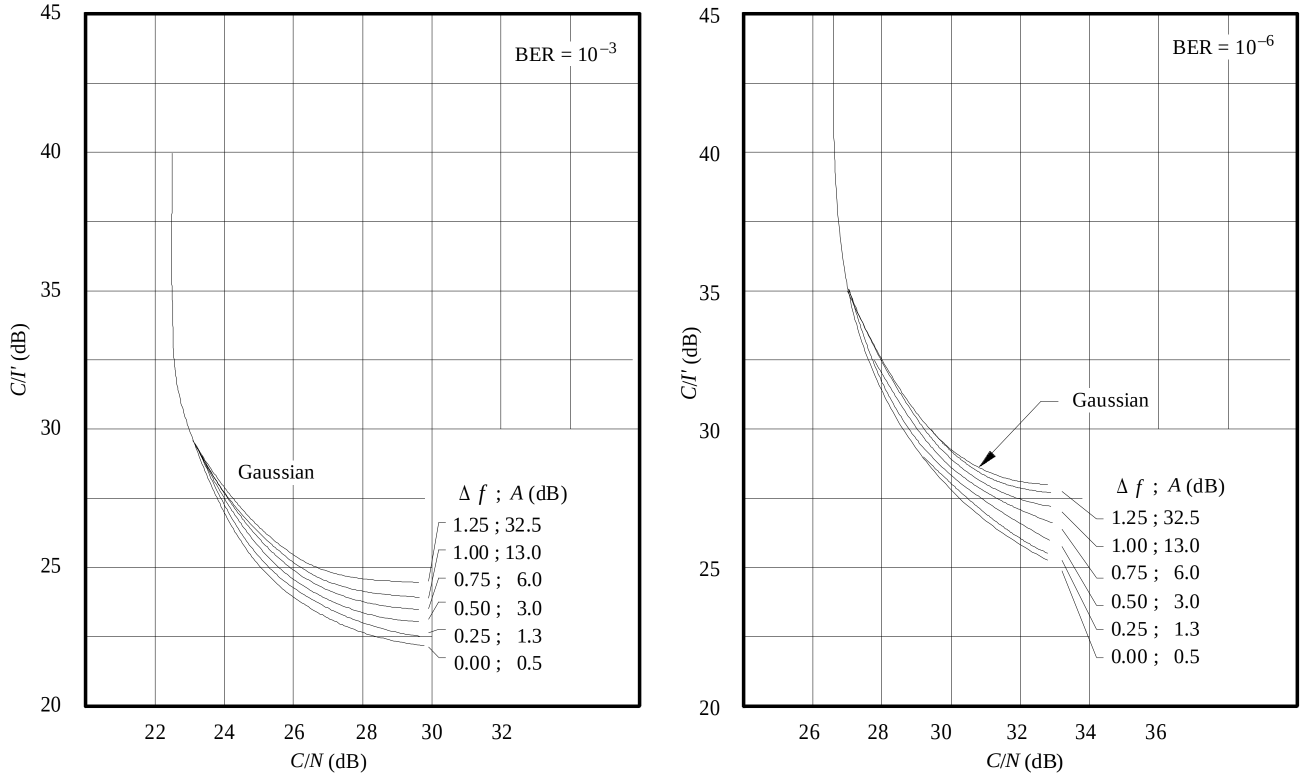
A (dB) : interference reduction factor defined as the ratio between the interfering powers at the input and at the output of the receive filter

Roll-off = 0.4

D19-sc

FIGURE 11

C/I' versus C/N for 64-QAM transmission with 64-QAM interfering carriers for 10^{-3} and 10^{-6} BER



Δf : frequency separation between carriers (normalized with respect to the symbol rate)

A (dB) : interference reduction factor defined as the ratio between the interfering powers at the input and at the output of the receive filter

Roll-off = 0.4

D20

2.4.4 Bounds methods

In many practical situations where an exact statistical distribution of the various interferences is not available, a useful technique is to compute an upper bound on the probability of error. This method requires knowledge only of the carrier-to-noise ratio at the demodulator input, C/N, the peak-to-r.m.s. ratio of the interference and the ratio of the powers of the wanted signal and interference, C/I. It should be noted that the results apply to a theoretical system and take no account of practical system restraints; they may be substantially modified by the presence of jitter and other degradations encountered in practical systems.

Other studies provide results for various cases of practical interest, including the effect of frequency separation between the wanted and unwanted carriers.

Curves of combinations of C/N and C/I ratios that give rise to upper-bound bit-error probabilities of 10^{-3} and 10^{-7} are presented in Figs. 15 and 16 respectively. These curves apply to cases of single or multiple interferences. The parametric curves are presented as a function of the interference peak factor, PF:

$$PF = 20 \log \frac{R}{\tau_r} \qquad (35)$$

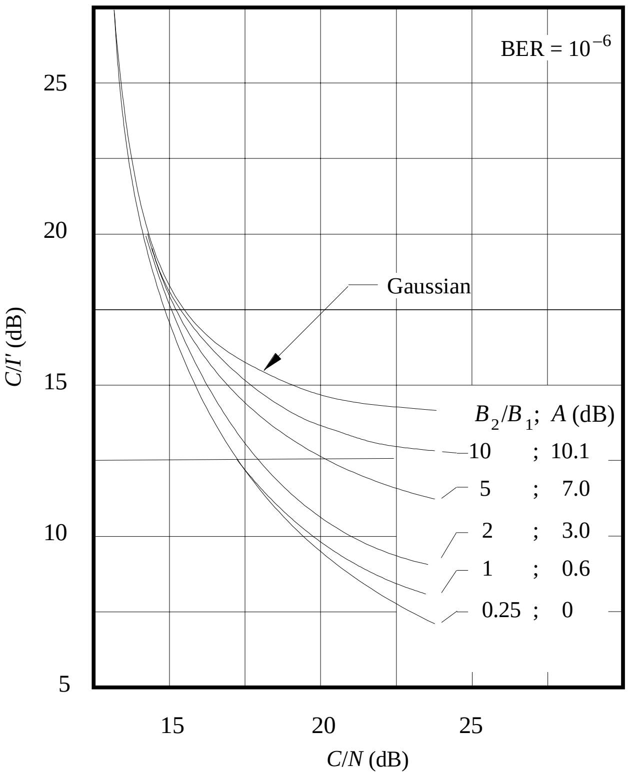
where:

R : peak value of the interference envelope

τ_r : root mean square value of the interference envelope.

FIGURE 12

C/I' versus C/N for 4-PSK transmission with different bandwidth 4-PSK interfering carriers for 10^{-6} BER



B_1: wanted carrier bandwidth

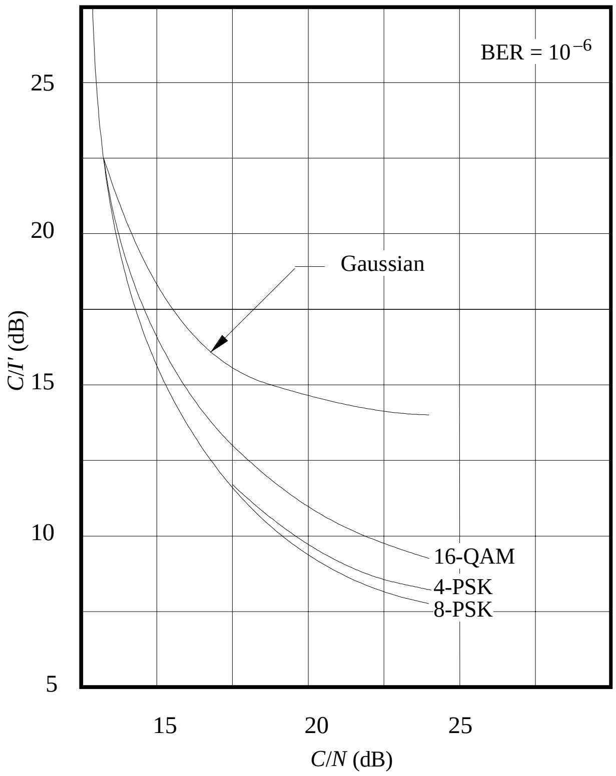
B_2: interfering carrier bandwidth

A (dB): interference reduction factor, defined as the ratio between the interfering powers at the input and at the output of the receive filter

Roll-off = 0.5

FIGURE 13

C/I' versus C/N 4-PSK transmission with equal bandwidth PSK and QAM interfering carriers for 10^{-6} BER



Roll-off = 0.5

D21

An unfiltered angle-modulated signal has a value of:

$$PF = 0$$

Results identical with those for a single angle-modulated signal (FM, PM, CPSK, or DPSK) interfering with binary CPSK can be obtained directly from the $PF = 0$ curves of Figs. 15 and 16. The corresponding results for interference to ternary and quaternary CPSK can be obtained indirectly from the same curves through the use of the foregoing formulae.

The following general conclusions can be drawn by inspection of the figures:

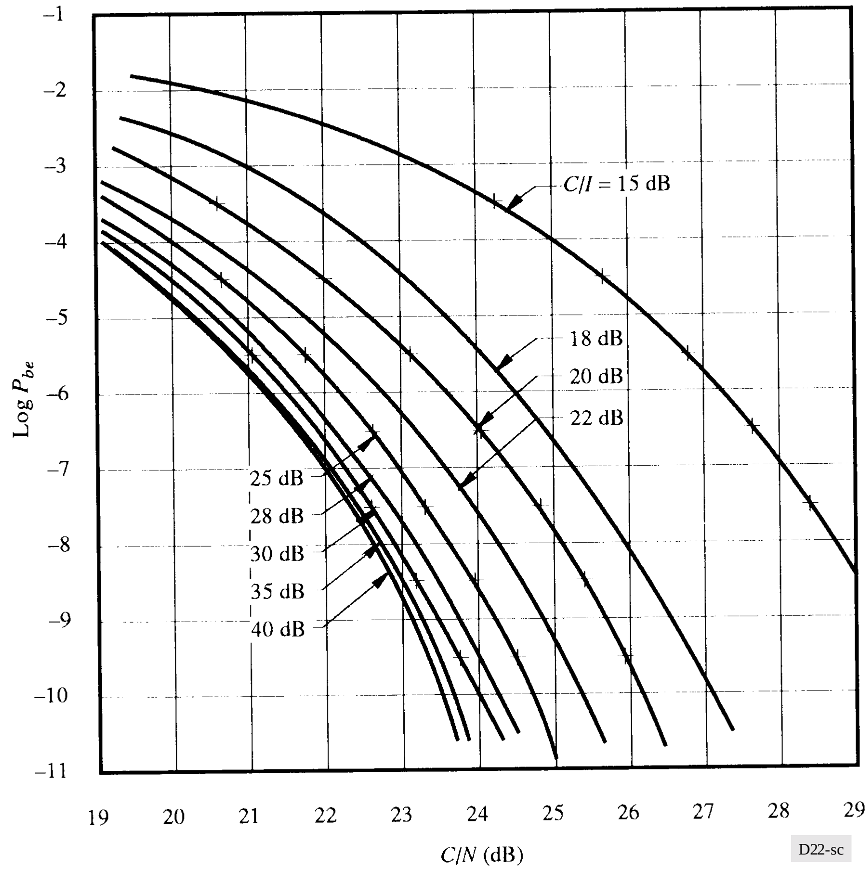
– when the interfering signal power is equal to, or larger than, the thermal noise power, the effect of angle-modulation interference is considerably less than that of an equal amount of white Gaussian noise power;

– when the interfering signal power is small compared to the thermal noise power, the effect on error ratio can be estimated safely by assuming that the interfering signal is equivalent to Gaussian noise of equal power;

– at a given carrier-to-interference ratio, the vulnerability to interference increases substantially as the number of transmitted symbols, M, increases.

FIGURE 14

Computational results of single interference to a 16-QAM system



2.4.5 Interference to DSPK signals from angle-modulated signals

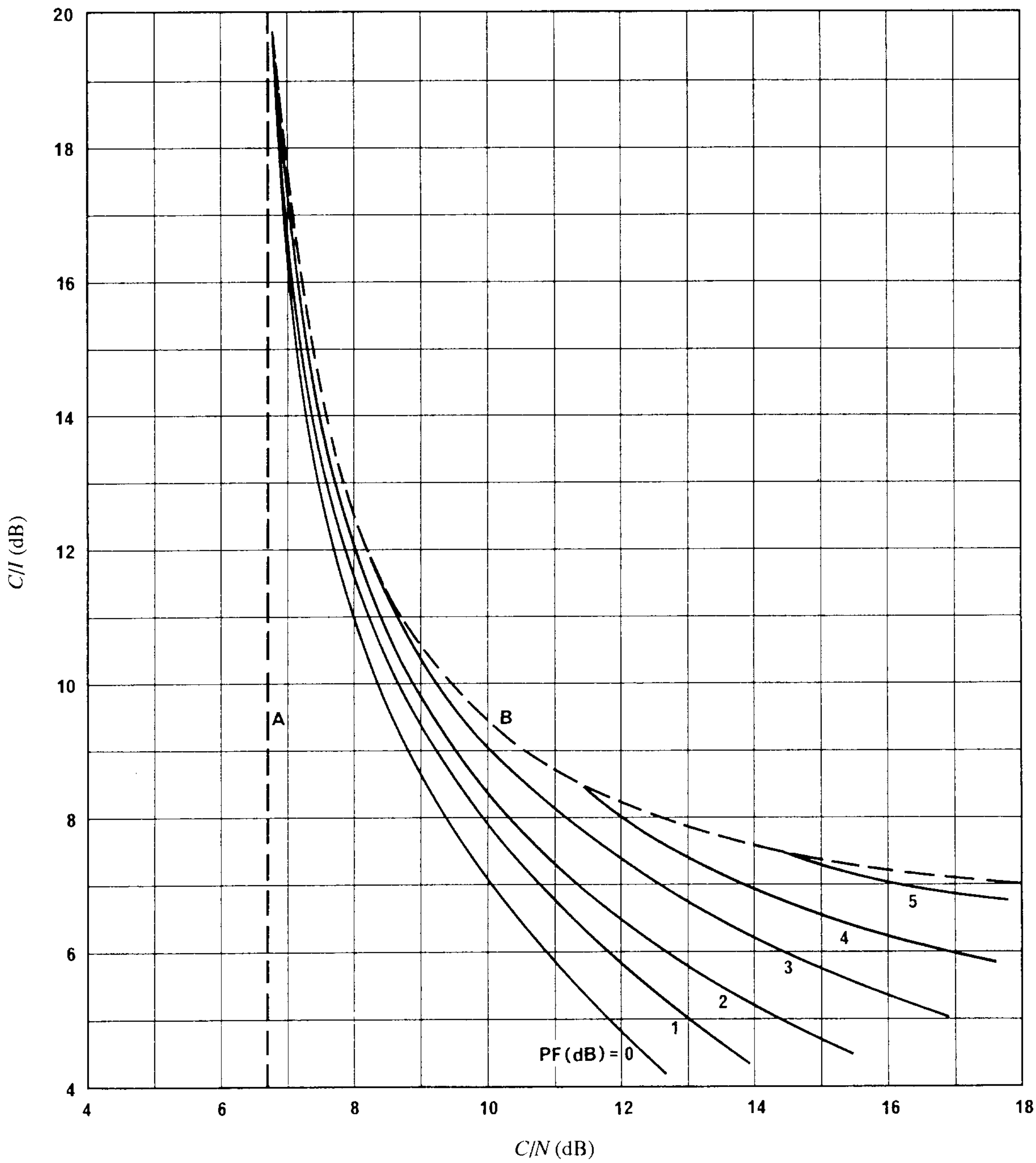
Curves of symbol error ratio against *C/N* ratio, with *C/I* ratio as a parameter, for differentially-coherent signals with 2, 4, 8 and 16 transmitted phases, are shown in Fig. 17. The error probability for differential detection is seen to be dependent on an additional parameter, θ, which is the relative phase slip of the interference from one sample to the next. However, the θ dependency diminishes as the number of transmitted phases increases. As a result, θ is assumed as a uniformly distributed random variable for systems with higher than four transmitted phases. Hence, average error probabilities have been derived for M = 8, 16; and probability bounds have been derived for the binary and quaternary cases.

The curves for DPSK imply the same conclusions as to the CPSK curves regarding the relative interference effects of white noise and angle-modulated signals, and the dependence of these effects on M. In addition, it can be seen that, in general, differential detection suffers more degradation than coherent detection, except that binary DPSK performs about as well as binary CPSK. Interference degradation is used as a basis for comparison because any disparities in the noise-only performance are reconciled.

A method, simulating an FM signal passing through an ideal band filter, has been used to calculate the error probability for FM interference on a binary PSK system. Figure 18 shows the error probability P_e as a function of the ratio of the r.m.s frequency deviation of FM interference f_g to the PSK signal receiver filter band Δf_c. Calculations were carried out at five different interference levels relative to intrinsic noise: 3 dB (curve A), 0 dB (curve B), –3 dB (curve C), –6 dB (curve D), –10 dB (curve E). The signal-to-noise ratio was taken as 12.4 dB, since the error probability (also in the presence of interference) does not exceed 10^{-6}, which corresponds to a signal-to-noise ratio of 10.5 dB, while the total interference margin from all terrestrial and satellite systems amounts to at least 35%. The upper modulating frequency of the FM interference f_B was taken to be Δf_c. Figure 18 thus also shows the error probability P_e as a function of the effective FM interference modulation index, as well as the error probability values in the presence of additional thermal noise, instead of FM interference, at the same demodulator input levels as for FM interference (horizontal lines A′, B′, C′, D′ and E′).

FIGURE 15

***C/I* versus *C/N* for 10^{-3} BER**



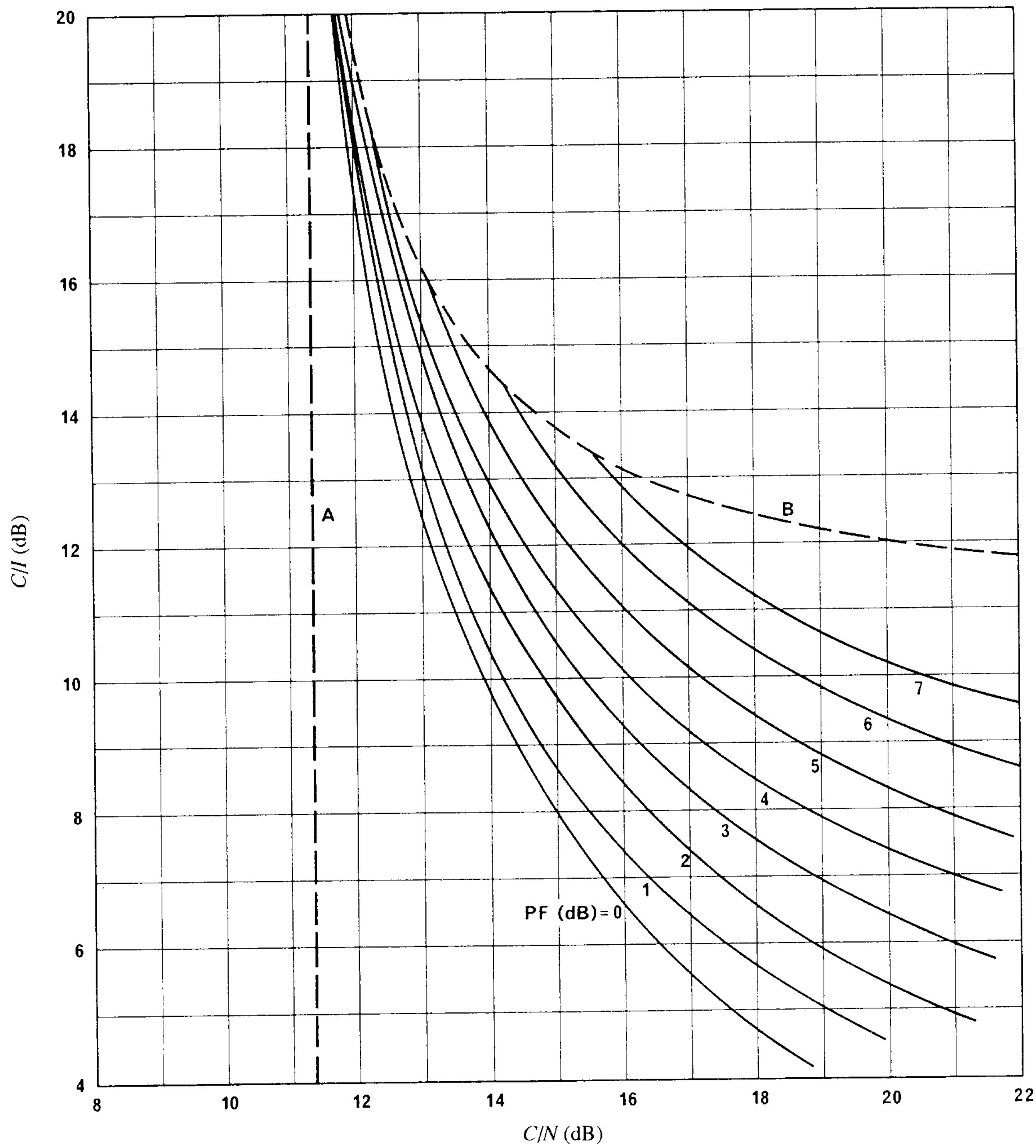
A : *C/N* in absence of interference
B : interference with characteristics of thermal noise
PF : interference peak factor

Note 1 – The curves are theoretical and take no account of practical system restraints.

D23-sc

FIGURE 16

C/I versus *C/N* for 10^{-7} BER



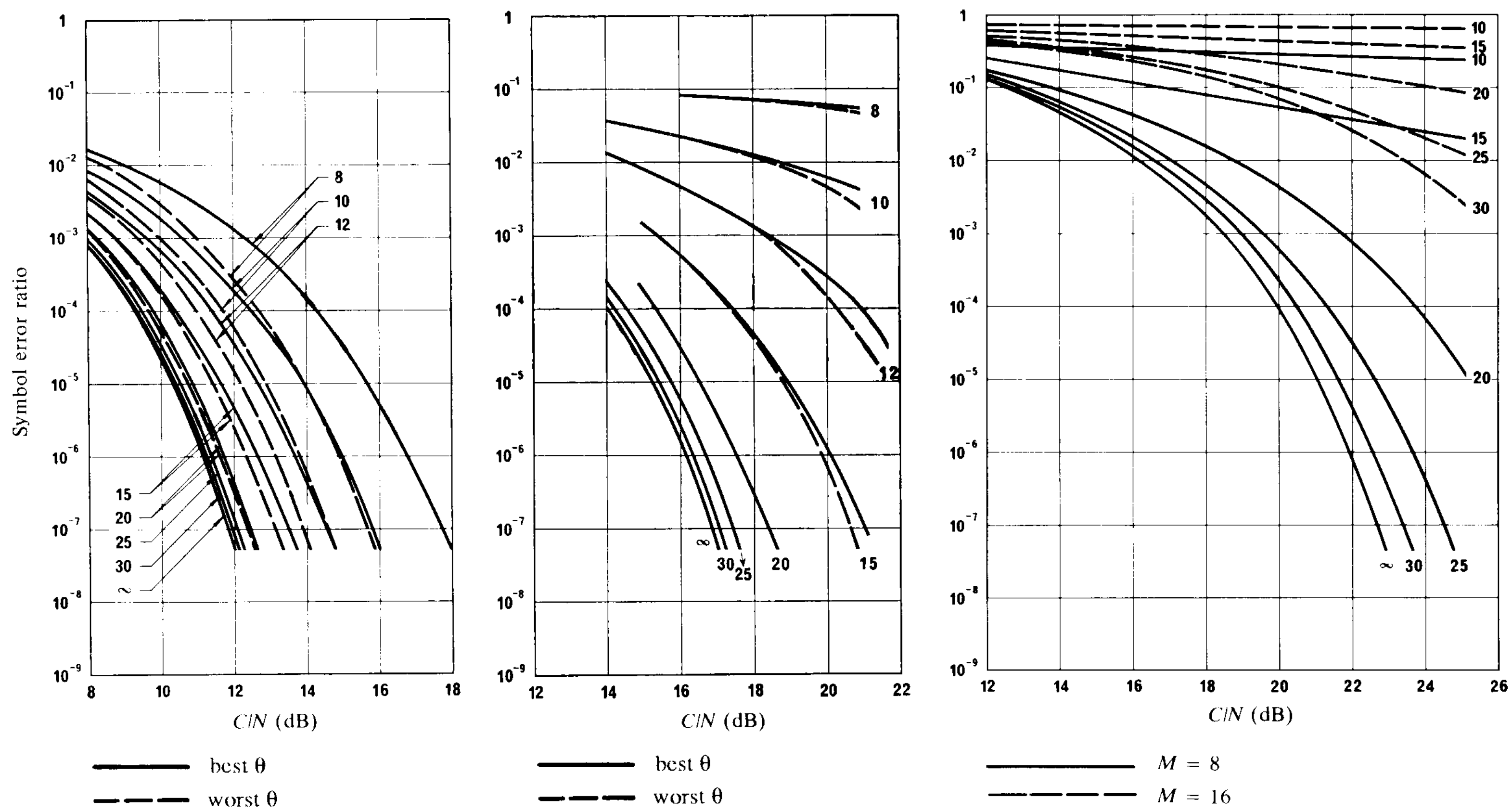
A : *C/N* in absence of interference
B : interference with characteristics of thermal noise
PF : interference peak factor

Note 1 – The curves are theoretical and take no account of practical system restraints.

D24-sc

FIGURE 17

Interference to a DPSK signal from an angle-modulated signal



a) *Binary (M = 2)*

b) *Quaternary (M = 4)*

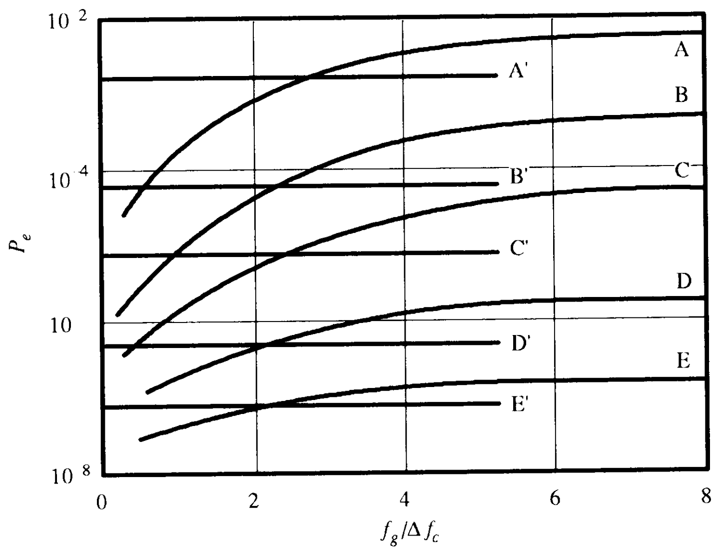
c) *Eight-phase (M = 8) and sixteen-phase (M = 16)*

(The carrier-to-interference ratio *C/I* (dB) is shown on each curve)

In practice, satellite and radio-relay systems operate with effective modulation indices of not more than 3. In analyzing the effect of existing FM systems on PSK systems, a Gaussian approximation provides the upper estimate. The increase in error probability proves substantially lower for FM interference than for thermal noise of the same power level, so that the permissible level of FM interference may be increased over the 6% provisionally established in Recommendation ITU-R S.523, for this particular case up to 1.4 dB.

FIGURE 18

Bit error probability (P_e) versus the $f_g/\Delta f_c$



Curves A: I/N = 3 dB
B: 0 dB
C: –3 dB
D: –6 dB
E: –10 dB
Lines A', B', C', D', E': Gaussian approximation

D26-sc

3. Signal spectra

3.1 Single channel per carrier FM telephony

Further studies are required.

3.2 Digital modulation signal of PSK, QAM and CPM type

The normalized power spectral density of the signal centred on the carrier frequency is expressed as follows:

$$P(f) = T_s \cdot \frac{\sin^2(\pi f T_s)}{(\pi f T_s)^2} \qquad \text{for M-ary PSK and QAM} \qquad (36a)$$

$$P(f) = 4T_s \cdot \frac{1 + \cos(2\pi f T_s)}{\pi^2 [1 - (4f^2 T_s^2)]^2} \qquad \text{for MSK} \qquad (36b)$$

$$P(f) = 2T_s \cdot \left[\frac{\sin(\pi f T_s)}{\pi f T_s} \right]^2 \cdot \left[\frac{\cos(\pi f T_s)}{1 - (4f^2 T_s^2)} \right]^2 \quad \text{for quadrature-overlapped raised cosine modulation} \tag{36c}$$

$$P(f) = \frac{\sin^2(2\pi f T_s)}{2\pi^2 f^2 T_s (1 - 4f^2 T_s^2)^2} \quad \text{for IJF-OQPSK} \tag{36d}$$

3.3 Frequency-modulated television signal (FM-TV)

After examining the spectrum, the following expression is taken as the upper bound of the normalized signal spectral power density centred on the carrier frequency:

$$P(f) = \text{Sup}\left\{ \frac{1}{\sqrt{2\pi}\,\Delta F} \exp\left[-\frac{1}{2}\left(\frac{f}{\Delta F}\right)^2 \right], g_i(f) \right\} \tag{37}$$

where i may assume three different values (Sup (x, y) designates the greater of the two functions x and y). The interference obtained for each one of these values is examined in turn, and the highest level of interference is adopted.

Measurements have shown that the $P(f)$ of an FM-TV signal with dispersion is more accurately defined by the following formula:

$$P(f) = \text{Sup}\left\{ \frac{1}{\sqrt{\pi}\,\Delta F} \exp\left[-\left(\frac{f}{\Delta F}\right)^2 \right], g_i(f) \right\} \tag{37a}$$

When determining the allowable level of interference for 20% of the time from an FM-TV signal with dispersion, this value can be assumed to be 10 dB lower than that calculated by formula (37a).

The first part of the expression between square brackets represents the "continuous background" of the spectrum, which is Gaussian; ΔF having the meaning given in § 2.2 and f being the frequency (MHz). The second part, $g_i(f)$, represents the "central" part of the spectrum essentially linked with the lines corresponding to "black" and "white". If Δf is the frequency deviation of the energy dispersal, $g_i(f)$ has the values given in Fig. 4 for $i = 1$, 2 and 3. These values correspond respectively to the case of a uniform picture (black or white), strongly contrasted (typically, "half-line bar" test pattern), slightly contrasted (typically, "staircase" test pattern). The effect of the synchronization line and the colour sub-carrier was disregarded in these models, since the lines concerned are less important in terms of power than the lines taken into account in these models.

However, the model corresponding to $i = 1$ can only be used as it stands for AC coupled modulators in which case the spectrum remains centred on the nominal frequency for a black (or white) picture. However, if a DC coupled modulator is used, the nominal frequency corresponds in all cases to medium grey; the function $g_i(f)$ must then be centred on a frequency separated by $\pm\,\Delta F/3$ from the nominal frequency.

3.4 Amplitude modulated telephony signal

If f_{min} and f_{max} are the lower and upper frequencies of the baseband signal, the normalized spectral power density is equal to:

$$P(f) = \begin{cases} \dfrac{1}{f_{max} - f_{min}} & \text{(SSB – SC case)} \\[2ex] \dfrac{1}{2(f_{max} - f_{min})} & \text{(DSB – SC case)} \end{cases} \tag{38}$$

inside the bandwidth of the signal; equal to zero outside.

4. Non-spectral interference effects – linear channel

Besides the spectral interference effects, consideration must be given to effects which are not predictable from power spectral densities. Various interference degradations require examination of time related characteristics. Examples of such degradations are:

– impulse noise in FDM-FM communications systems can result from adjacent channel FM interference. In this case an FDM-FM carrier located in an adjacent frequency band will occasionally deviate into the desired carrier's passband. If the interference to desired carrier power ratio and the time versus deviation statistics are improper, impulsive or click noise will result;

– interference to television can result from a burst transmission carrier such as TDMA. In this case the envelope of the interfering carrier may have frequency components to which the video signal is sensitive. Frequencies near the television line or frame rate may be expected to provide subjectively disturbing degradations;

– interference effects may result from a large carrier, modulated only by the energy dispersal waveform, sweeping past a small narrow passband carrier such as single-channel-per-carrier (SCPC). This situation produces transient effects related to the interference duty factor and sweep rate.

This list of examples is not complete but is meant to illustrate several time dependent interference mechanisms.

Another non-spectral effect in relation to interference performance is dependent on the demodulation technique. Depending on the nature of the interference, one demodulation technique may be preferable. As an example, adjacent channel induced impulsive noise in a wideband FM system may be reduced by the use of a properly designed phase locked loop or frequency modulated feedback demodulator. In the case of digital reception, different carrier and clock timing recovery techniques will have differing sensitivities to specific types of interferences.

5. Non-linear channel effects

5.1 General

Most satellite transmission channels in use today have non-linear transmission properties resulting from the transponder and earth station equipment employed. A non-linear relation exists in the transponder between the input and output amplitude (AM-AM) in addition to which the phase transfer function is related to input amplitude (AM-PM conversion). These characteristics have implications for the interference susceptibility of a given communications system. With both the desired signal and interference present at the input of the non-linear device, a multiplicative (non-additive) degradation is generated. Depending on the modulation technique employed, this degradation will manifest itself on baseband performance.

5.2 Analogue FDM-FM telephony wanted signal

In dealing with interference to FM analogue signals, two areas should be considered. The presence of the desired carrier and an interfering carrier(s) at the non-linear device input will result in the generation of intermodulation spectral components. These components may then appear as additional interfering carriers. The second area of consideration is when the input combination of desired and interfering signals results in amplitude modulation; this modulation is converted to phase modulation by the AM-PM conversion process. The phase modulation is imparted on the desired carrier and when finally demodulated at the receiver, results in baseband degradation.

Incomplete suppression of the amplitude modulation of the wanted signal by the limiter of the receiver can generate baseband interference, or adjacent channel interference on the slope of the wanted channel's filter can be amplitude modulated; this AM converted to PM thus appears at baseband. The non-linearity of power amplifiers and demodulators are the usual sources of this type of interference.

The non-linear interference can have a severe subjective effect because it can appear as direct cross-talk. Moreover, it can degrade the threshold of the receiver, and this effect is particularly applicable to satellite signals where the level of the wanted signal is usually near the threshold level, and adjacent channel interference may generate a burst of threshold noise.

The non-linear interference mechanisms require investigation when the more conventional linear mechanisms appear to produce negligible interference. The calculation of this interference requires information on the specific receivers, filters, and AM-PM conversion constants.

In the investigation and analysis of FDMA-FM systems for the transmission of multi-channel telephony, calculations of interference noise in individual channels should take account of the following sources of interference:

– non-linearity of a realizable limiter;

– non-linearity of a realizable frequency detector;

– threshold effect of an FM receiver (taking account of the modulation index of the interference);

– AM-PM conversion in the RF channel.

5.3 Digital PSK wanted signal

The treatment of interference to digital PSK carriers is more complex than the analogue case. Bandpass filtering of the PSK carrier to minimize bandwidth requirements results in significant envelope amplitude modulation at frequencies related to the symbol rate. This, when converted to phase modulation by the AM-PM conversion mechanism, reduces the interference immunity of the system. Separate consideration must be given to the performance of the carrier and clock reference recovery functions of the system. Specification of the modulator and demodulator characteristics with respect to filtering, carrier and clock reference recovery techniques and sampling methods, may be expected to have significant impact on the interference immunity of the system. At the present time there are no analytic expressions available for the computation of the interference effect to PSK carriers transmitted over a non-linear channel. Laboratory measurements on various specific systems have been presented and can be used for guidance.

6. Measurements of interference into digital systems

One study shows that a considerable reduction in interference is possible from angle-modulation systems into pulse-code modulation systems using phase-shift keying as compared to the mutual interference between two angle-modulation systems.

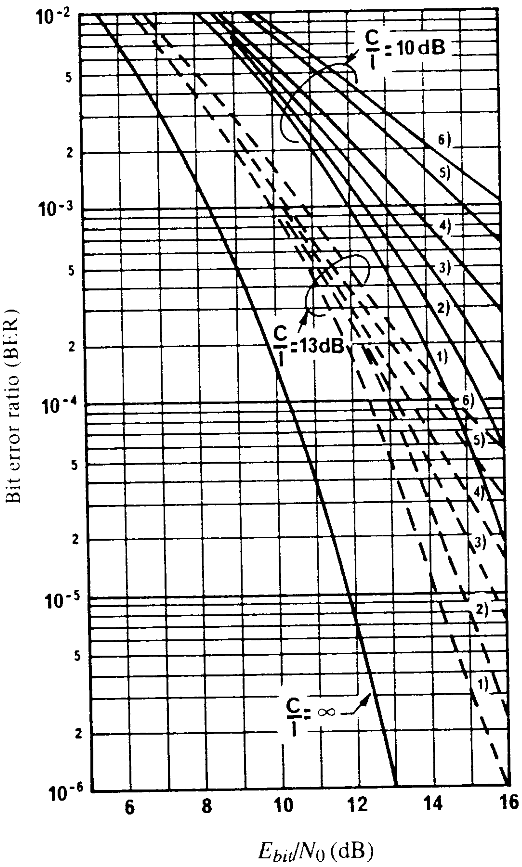
Other tests showed agreement between theory and measured data.

Experiments conducted on the effect of PSK interference and noise on PSK signal demodulators make it possible to determine the validity of a Gaussian approximation in estimating the effect of PSK interference. Figure 19 shows the error ratio of a coherent 4-PSK demodulator as a function of the energy/bit-to-noise density ratio for two fixed C/I ratios, 10 and 13 dB, and different ratios between interference R_i and signal R_s channel transmission rates (R_i/R_s = 0; 0.5; 1; 2; 5). The carrier-to-interference ratio was established at the demodulator receiving filter output with a band 1.1 times that of the Nyquist band. Figure 20 shows the error ratio as a function of the R_i/R_s ratio for a fixed C/N ratio = 13 dB and three values of C/I ($C/I = C/N$, $C/I = C/N + 2$ dB, $C/I = C/N - 2$ dB). Figure 21 shows the relationship for the use in the wanted signal channel of a convolution code codec at $\gamma = 1/2$ for the code speed with Viterbi decoding.

Examination of the results obtained shows that the representation of co-channel PSK interference as Gaussian noise is correct for R_i > (4 – 5) R_S, and this applies both to the ordinary channel and to systems using coding, although in the latter case the character of the variation in error ratio is not monotonic. In the region of values of levels of interference commensurate with the thermal noise level, wideband PSK interference produces an increase in error ratio roughly of an order of magnitude in comparison with unmodulated interference of the same level, which is equivalent to a difference in their levels of up to 3 to 4 dB at a constant error ratio. It should also be noted that 2-PSK interference produces a more perceptible effect on error ratio than 4-PSK interference.

FIGURE 19

Error ratio as a function of energy/bit-to-noise density, E_{bit}/N_0 and carrier-to-interference *C/I ratios*



1) $R_i/R_s = 0$

2) $R_i/R_s = 0.5$

3) $R_i/R_s = 1$

4) $R_i/R_s = 2$

5) $R_i/R_s = 5$

6) Interference in the form of noise

D27-sc

FIGURE 20

Coherent 4-PSK demodulator error ratio as a function of R_i/R_s transmission speed ratio

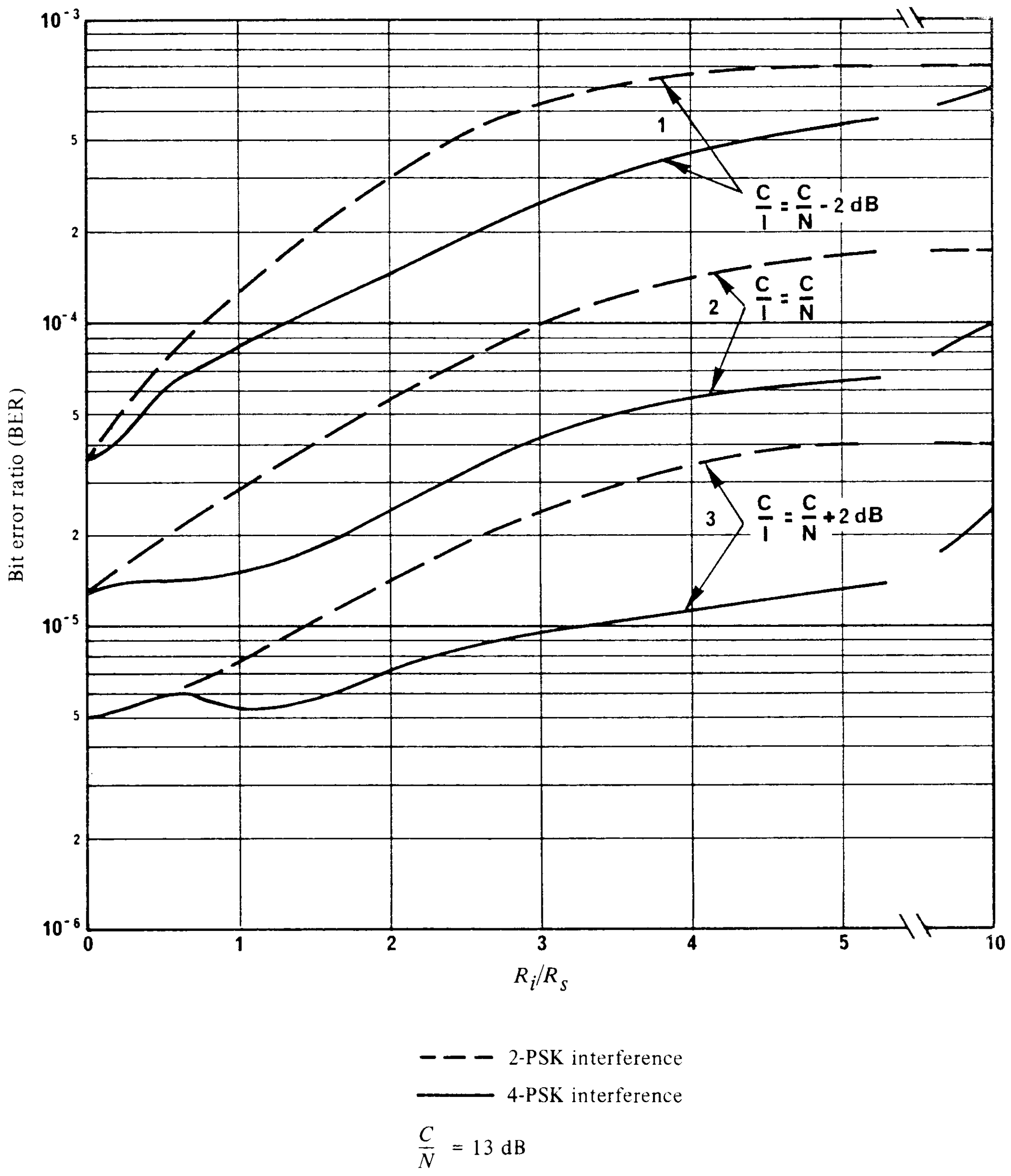


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FIGURE 21

Error ratio at convolution code decoder output as a function of R_i/R_s transmission speed ratio