FINAL REPORT

FREQUENCY ASSIGNMENT FOR LAND MOBILE RADIO SYSTEM: DIGITAL TRANSMISSION OVER LAND MOBILE CHANNELS-INTERFERENCE CONSIDERATIONS

by

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Department of Electrical Engineering

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The University of British Columbia

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Department of Communications

Ottawa, Ontario, CANADA

under

Department of Supply and Services Contract Serial No: 0SU79-00138

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LIST OF ILLUSTRATIONS

A B S T R A C T

Increasing demand for digital transmissions over Land Mobile radio channels requires technical data in support of guidelines relating to transmitted spectra, data rates, bit-error rates, and co-channel and adjacent-channel interference. This report focuses on interference from both FM voice and data signals.

Land Mobile channels, which show signal level fluctuations in excess of 90 dB, are characterized for purposes of the present study. Spectral emission constraints, including those currently proposed by Communications Canada, are described and discussed. The effects on bit-error probability of signal level variations during data reception are examined.

Spectra of both transmitted data signals and FM voice signals are determined. Maximum data rates permitted by spectral emission constraints are determined for various binary signalling schemes, including PAM, PSK and FSK. Transmitted FM spectra for voice modulation are calculated, based on two different spectral (Gaussian) models for speech. For each model, the maximum rms bandwidth permitted by the spectral emission constraints is determined.

A general expresion for the output signal-to-noise-plus-interference ratio (SNIR) is derived for digital data demodulators. The expression includes the effects of co-channel and adjacent-channel interference from both FM voice and other data signals. Output SNIR depends on the ratio of the received signal-to-interfering-power levels, the receiver baseband filter characteristic, the channel separation Δ relative to the data rate and, for FM voice interference, the voice signal's rms bandwidth B as well as the desired signal's data rate. Graphs and tables are presented to show how the interference-to-signal ratio varies with Δ for the signal formats listed above.

Effects of data and voice interference on FM voice signals are determined, and IF bandpass filter to reduce these interference effects are described.

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I INTRODUCTION

I-1 Scope and Objectives of the Study

This report presents the results of a Communications Canada Contract study whose motivation and purpose are summarized below:

MOTIVATION:

Recent advances in land mobile system development has provided increasing evidence that digital transmission of dispatch data as well as other digitally encoded information will constitute an important part of land mobile communication traffic. The need for the development of guidelines identifying transmission characteristics, requirements, and limitations of digital transmission is becoming increasingly urgent. These guidelines would lead to the development of new regulations and standards governing the technical characteristics and operation of equipment and systems employing digital transmission of information over land mobile communication channels.

The following items are of specific interest:

1). Co-channel interference criteria for various combinations of digital and analog transmission over Land Mobile channels

2). Adjacent-channel interference criteria for various combinations of analog and digital transmission over Land Mobile channels

3). Receiver selectivity characterisitics for minimum degradation due to adjacent-channel interference.

PURPOSE:

To produce technical data in support of the development of new standards governing the use and operation of Land Mobile radio equipment utilizing digital modulation techniques over Land Mobile channels.

Results obtained would be expressed in terms of carrier-to-interference (C/I) ratio, receiver selectivity characteristics, and spectral emission constraints for Land Mobile channels.

It was recognized at the outset that an important part of the study is to select from among various alternative approaches one which could provide a useful set of technical data consistent with the study's motivation and purpose.

I-2 Executive Summary

The following detailed summary is presented here for convenience, and replaces summaries which might otherwise appear at the conclusion of each chapter.

Chapter 2 briefly characterizes land mobile communication channels, which show fluctuations in received signal level in excess of 90 dB. Various alternatives are available to combat these fluctuations including space diversity, time diversity via forward error correction (FEC) coding, and variable-rate transmission via error detection and retransmission (ARQ). The need for spectral emission constraints is established for spectrum sharing based on conventional channelization, and a spectral emission constraint issued by Communications Canada is described. It is shown that for frequencies in the 900 MHz region, received signal levels are relatively constant during the reception of a single bit at rates exceeding approximately 1 kilobits/sec. (Kbs).

In Chapter 3 power spectral densities are determined for digital signals including PAM, PSK and FSK, as well as for analog FM signals modulated by pre-emphasized speech. Speech is modelled as a stationary Gaussian process whose power spectral density approximates the long-term average spectral density of continuous speech. Transmitted FM spectra are calculated and displayed for various values of the root-mean-square bandwidth B of the transmitted signal. The output signal-to-noise ratio increases with B; for this reason the maximum value of B for which the FM spectra fit within a defined spectral region was determined as 2.9 kHz.

In considering multi-level pulse amplitude modulation (PAM) for digital signalling it was recognized at the outset that the large received signal level variations inherent in Rayleigh channels would require adaptive receiver decision level adjustments; to obviate this difficulty, PAM signalling was restricted to binary antipodal. Modulator pulse shapes were restricted to those which generate zero intersymbol interference (isi) following receiver matched filtering, again to obviate the need for signal-level-dependent receiver adjustments. Within these limits consideration was given to conventional rectangular baseband pulses and to pulses with raised-cosine spectra which provide a tradeoff between relative excess bandwidth α and bit rate. With maximum excess bandwidth and double sideband modulation the spectral emission constraints permit data rates of up to 20 Kbs. Use of zero excess bandwidth signalling permits rates of 31 Kbs. For conventional rectangular pulse shapes, the spectral constraints limit the bit rate to 4 Kbs.

Conventional PSK signals have a constant amplitude from which it immediately follows that binary PSK is identical to conventional binary PAM. However, PSK need not be restricted to two levels since detection is based on the ratio of outputs from quadrature filters, and this ratio is independent of signal level. Noise rather than bandwidth limits the PSK data rate since the transmitted signal spectra is independent of the number of PSK levels. For bandwidth-efficient PSK signalling, non-constant amplitude pulse shapes may be used; the resulting spectra are identical to those of PAM with the same pulse shapes. Use of pulses with raised-cosine spectra permit bit rates of $(\log_2 L)\ R_{PAM}$ where L is the number of PSK levels and R_{PAM} is the corresponding PAM data rate; R_{PAM} = 31 Kbs, 20 Kbs and 4 Kbs for minimum excess bandwidth pulses, maximum excess bandwidth pulses and conventional rectangular pulses, respectively.

To obviate the need for carrier phase estimation differential PSK (DPSK) is available, with a maximum power penalty of 3 dB relative to PSK. Normally the spectra of DPSK is identical to that of PSK. For environments where bandwidth limits data rates, DPSK with raised-cosine signalling having a maximum excess bandwidth appears attractive, since phase estimation is unnecessary and timing error tolerance is relatively high.

FSK signal spectra are not simply related to the baseband modulating signal spectrum, and may be wider than or narrower than that of the baseband signal. The spectrum depends on the baseband signal pulse shape and the peak-to-peak frequency deviation relative to the bit rate. However, for a deviation ratio of 0.5 minimum shift keying (MSK) results, and exact determination of the transmitted spectra is possible. We show that for binary MSK the data rate permitted by the spectral emission constraints is 36 Kbs.

The relatively high data rates permitted for digital signalling occur because the transmitted spectra of appropriately designed digital signals neatly fit the defined channel whose bandwidth from zero to -26 dB below the unmodulated carrier power is ±10 KHz; the bandwidth falls linearly (in dB) to ±20 KHz at -60 dB. However, a high bit rate R implies a low energy per bit and thus a potentially high bit error rate, in which case noise rather than bandwidth may limit R. A reduction of R and, for FSK a corresponding reduction in peak deviation does increase the signal-to-noise ratio (SNR) in proportion to 1/R. However, some bandwidth is thereby unused, and more effective alternatives exist which use all of the bandwidth. These alternatives include channel encoding, or a reduced bit rate together with bandwidth expanding modulation of an FM carrier which increases SNR in proportion to R^{-3}.

All of the bit rates quoted are for double sideband modulation. Use of single-sideband or vestigital sideband modulation increases these allowable rates up to a factor of two, but at the expense of increased sensitivity to demodulator timing errors.

Existing spectral constraints for land mobile radio channels require that 99% of the spectral energy be with ±8 kHz of the carrier frequency. These constraints result in much lower allowable data rates than those quoted above. For example, the maximum rates for MSK, PAM with maximum excess bandwidth and PAM with zero excess bandwidth are 13.3 Kbs, 8.1 Kbs and 16.2 Kbs, respectively. The maximum allowable rms bandwidth for speech FM is also reduced.

Chapter 4 considers the effects of co-channel and adjacent-channel interference as well as noise on digital transmission. Expressions are derived for output interference-to-signal ratios for interference arising from both FM voice and data transmission. These expressions are of the form $(P_i/P_s)C_i$ where P_i/P_s is the ratio of the received interference to signal powers, and interference coefficient C_i depends on the format of the desired data signal, the demodulator filter transfer characteristic, the data rate, and the separation Δ, relative to the data rate T^{-1}, between the interfering and desired carrier signal frequencies; $\Delta=0$ corresponds to co-channel interference and $\Delta\neq 0$ corresponds to adjacent-channel interference. When the interference is from an FM voice signal, the rms bandwidth B of that signal also affects C_i. When the interference is from a data signal with a format different from that of the desired data signal, C_i also depends on the interfering signal's format and data rate.

Interference coefficients were actually determined for voice interference as well as for data interference when the interfering and desired data signals were of identical format and data rate. Data formats considered include conventional PSK (rectangular-pulse binary PAM), MSK, PAM with raised-cosine spectra and maximum excess bandwidth, and sinusoidal FSK. For FM voice interference, both B and T^{-1} assumed the maximum values permitted by the spectral emission constraints, in order to maximize output signal-to-noise ratio for voice transmission and data rate for data transmission. For all data signal formats considered, C_i for voice interference decreased as Δ^{-2}, because of the relatively slow spectral roll-off of FM voice signals. Because of this slow spectral roll-off, use of different types of receiver filters for data demodulation would not significantly reduce interference from FM voice transmissions on the same or adjacent channels. The only way to reduce this interference, other than by reducing T^{-1} or B is to preprocess the voice signal prior to FM modulation or to employ digital transmission of voice. The availability of adaptive differential source encoders which transmit good quality speech at approximately 20 Kbs makes this latter alternative a viable one, provided subjective quality tests confirm that digital transmission on mobile channels compares favourably with conventional FM transmission. Both alternatives are of real interest and merit further study.

Adjacent-channel interference from another data signal results in a decrease with Δ in the coefficient C_i; the rate of decrease is governed by the data signal with the slowest spectral roll-off. When both the desired and interfering signals use the same formats, C_i decreases as Δ^{-2} for conventional PSK, Δ^{-4} for MSK and Δ^{-8} for SFSK. From the point of view of minimum adjacent channel interference, SFSK is clearly the most attractive, notwithstanding the facts that its maximum data rate is 32 Kbs (against 36 Kbs for MSK), and its spectrum is in slight violation of the emission constraints at this data rate.

When bandwidth efficient PAM is used on both the desired and interfering channel, zero adjacent-channel interference results when Δ exceeds some value, which value is $2T^{-1}$ when both signals have data rate T^{-1}. For $T^{-1} = 20$ Kbs, for example, the required separation is $\Delta = 40$ kHz.

Results quoted above apply to digital data demodulator filters matched to data signal pulse shapes, although some other filters were considered for purposes of implementation simplicity.

In comparing co-channel interference coefficients at maximum data rates, variations of less than approximately ±1 dB from $C_i \simeq -3$ dB are observed among various signalling formats subject to either voice or data interference, unless conventional PSK is used in which case $C_i \simeq -7.5$ dB for voice; this rather low value results from the relatively low (4 Kbs) data rate. Reducing the data rate in the presence any given interference does reduce C_i for both co-channel and adjacent-channel interference, since C_i varies as T^{-1}. However, such rate reductions result in unused bandwidth which can be employed to further reduce bit-error rates, as explained earlier.

To estimate bit-error probability p which result from interference, one can replace the resulting interference-plus-noise from the demodulator output by Gaussian noise of equivalent power, and calculate p for a given level of signal and interfering powers. This value of p must then be averaged over both the signal and interference levels to give p for the region where the average signal and interference levels are relatively constant.

In Chapter 5 the latest version of Communication Canada's proposed Radio Standards Specifications is considered. These standards restrict the peak frequency deviation of the transmitted signal to ±5 kHz, and require the measured energy in a 16 kHz bandwidth centred ±25 kHz from a carrier modulated by a standard test signal to be 60 dB below the unmodulated carrier signal power. Since the standards involve limitations on peak frequency deviations, they are not really applicable to PAM or PSK signalling. They do apply to FSK and imply a maximum data rate of 20 Kbs for binary MSK. The ±5 kHz maximum deviation permits estimation of the transmitted spectrum of an FM voice signal with rms bandwidth $B \gtrsim 2.5$ kHz. Standards based on transmitted spectra of actual signals (rather than test signals) would be suitable for PAM and PSK formats and would more tightly control emission levels. However, such standards would be more difficult to administer than those proposed.

Chapter 5 also includes consideration of an alternative speech spectrum model, based on non-linear compression (companding) and lowpass filtering at 3 kHz prior to PM modulation of a carrier. For this spectral model the transmitted spectrum decreases with frequency at a very much faster rate than does that of the earlier speech spectral model. For binary MSK, the voice interference coefficient C_v is comparable to the data interference coefficient C_d. These results clearly indicate the need for speech models consistent with any speech preprocessing operations inherent in microphones, compandors and filters. The results also support earlier suggestions that appropriate preprocessing can reduce considerably adjacent-channel interference from transmitted FM (or PM) voice signals.

Chapter 5 considers the effects of both voice and data-signal interference to FM voice signals. Bandpass IF filters to minimize the effects of this interference are considered, and the IF signal-to-noise ratio is determined. The interference by data signals on FM voice signals is shown to equal that of FM voice on data signals. The interference coefficient J_v from other FM voice transmissions is calculated for the two speech spectral models considered. For the spectral model based on actual speech J_v exceeds C_v, whereas the converse is true for the model based on companded and lowpass filtered speech.

Finally, Chapter 5 considers receiver filters for reducing interference. The decrease in adjacent-channel interference with increasing channel separation Δ depends on the spectral roll-off rate of the interfering signal and the receiver matched filter, whichever is less. It follows that non-matched filters would not reduce this roll-off rate when the desired and interfering signals have the same format. Reduction of adjacent-channel interference is therefore based on design of signals with desirable spectral properties, as illustrated by recent work on MSK-type signals. To reduce adjacent-channel interference from conventional FM demodulators the IF filter should have a spectral roll-off rate not less than that of the interfering FM voice or data signal. Co-channel interference lies in the frequency band of the signal, and is not normally removable by receiver filter design modifications.

II DIGITAL COMMUNICATION OVER LAND MOBILE CHANNELS

II-1 Fluctuations in Received Signal Strength

Bandwidth and noise limit the rate at which information can be transmitted over any communication channel. In land mobile radio channels, legislation limits the bandwidth occupied by any user. Noise originates from various sources [J1, A1], including (Gaussian) receiver front-end amplifier noise, (impulsive) automobile ignition noise, co-channel interference from other users of the frequency channel, adjacent-channel interference from users on other frequency channels, and randon phase changes associated with received signal level fluctuations described below. Because the relative geographic positions of both co-channel and adjacent-channel users change over time the associated interference levels also vary.

The level of a signal received by a mobile varies as the mobile moves. This level variation includes a fading component, illustrated in Fig. 2-1, resulting from multipath interference changes. Adjacent peaks in the received signal envelope are separated by one-half a carrier wavelength. The amplitude probability density $f_r(\alpha)$ of the received signal level r is reasonably well approximated by the Rayleigh distribution as follows [W1, F1]:

$$f_r(\alpha) = \begin{cases} (2\alpha/r_0)\exp(-\alpha^2/r_0) & \alpha > 0 \\ 0 & \alpha \leq 0 \end{cases} \tag{2-1}$$

where $s = \sqrt{\pi r_0}/2$ is the local mean level of the received signal and r_0 is the average energy in the received signal.

The distribution of r^2 the received signal energy is [W1, F1]:

$$f_{r^2}(\alpha) = \begin{cases} (1/r_0)\exp(-\alpha/r_0) & \alpha \geq 0 \\ 0 & \alpha < 0 \end{cases} \tag{2-2}$$

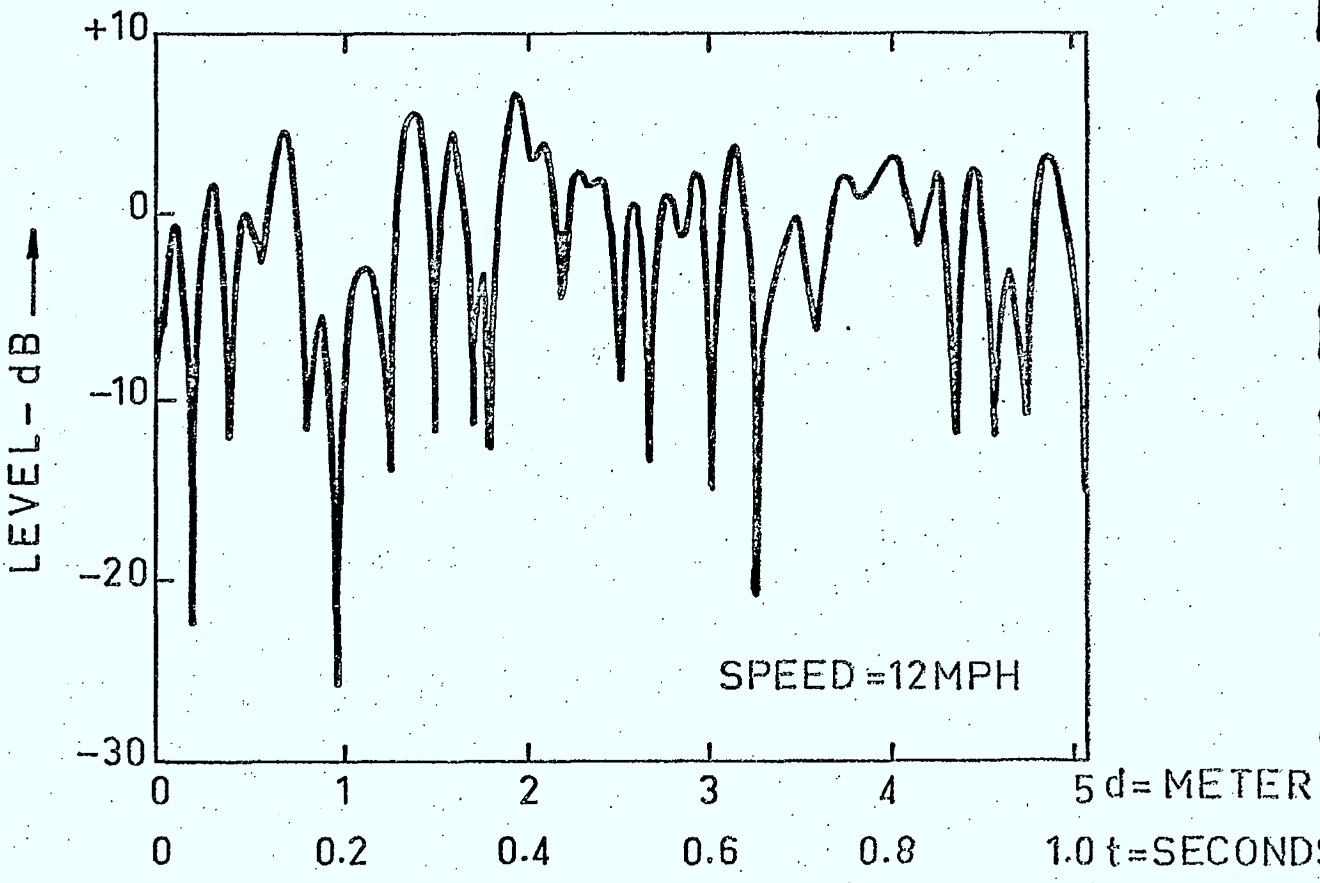

Fig. 2-1 Received signal level vs. time in an urban environment. Vehicle speed: 12 mph., carrier frequency: 850 MHz (after Arrendondo and Smith [A1].)

The local mean signal level s varies slowly over several wavelengths as a result of changes in path transmission characteristics. These changes result from variations in shadowing caused by topographical features such as street width, building height and hills. The amplitude probability density of s in dB is reasonably well approximated by the log-normal distribution, as follows [J1, F1]:

$$f_s(\alpha) = \frac{1}{\sqrt{2\pi}\sigma} \exp\left(-(\alpha-m)^2/2\sigma^2\right) \tag{2-3}$$

In (2-3) m is the mean signal level in dB averaged over several wavelengths (typically 50 m), and σ is the standard deviation in dB of the local mean; σ has been quoted as 6 dB in London [F1] and as 8-12 dB in Japan and the USA [F1, J1].

Finally, the local mean m varies over the mobile coverage area as the distance D between transmitter and receiver changes. This variation is proportional to D^{-n} where $2<n<4$ [J1, O1] for $D \overset{\sim}{<} 40$ km (25 miles). The rate of attenuation of n tends to increase with D. A recent study shows that $n \simeq 3.7$ in urban Philadelphia [O1].

II-2 Communication Over Land Mobile Radio Channels

Land mobile radio channels present problems additional to those associated with conventional point-to-point links. The signal level variations resulting from fading, local mean variations, and variations in transmitter-receiver separation D can exceed 20 dB, 10 dB and 60 dB, respectively, which together can cause a total received signal level variation in excess of 90 dB.

One way to combat the large signal-to-noise ratio fluctuations referred to above is to use one or more types of diversity. For example, spatial diversity in which several base station antennas are located at the edge of the radio coverage area can reduce fluctuations in base-station received signal level resulting from all these of the above-noted causes [J1]. By switching these antennas during transmission from the base station, variations in mobile receiver signal level due to shadowing and D^{-n} loss can also be reduced considerably [J1]. Space diversity systems utilize the fact that the probability of two or more transmission paths being poor at the same time is considerably less than the corresponding probability for any individual path. Forward error correcting [L1, G1] codes involve use of time diversity.

An alternative to diversity involves transmitting information at a rate which depends either directly or indirectly on the signal-to-noise ratio (SNR) at the receiver. Transmission rates are high when SNR is high, and conversely. Variable-rate transmission normally implies a two-way channel between transmitter and receiver. The receiver may estimate its signal-to-noise ratio and transmit this estimate to the transmitter, whose information transmission rate is adjusted accordingly [C1, C2]. Alternatively, the decoder in the receiver may detect, rather than correct errors in transmitted bit sequences, and request retransmission of those sequences received in error. Retransmission probability is highest and information throughput is lowest when SNR is low; thus, the actual information rate adjusts automatically to the state of the channel.

In determining the performance of various digital transmission schemes, the rate of signal level variation relative to vehicle speed is important. The salient facts are summarized below.

The carrier signal wavelength $\lambda=c/f$ where c is the velocity of light (3×10^8 m/sec) and f the carrier frequency. At 900 MHz, $\lambda \simeq$ 1ft. Thus, fading minima in Fig. 2-1 correspond to vehicle movements of 1/2ft. (1/6 m). At 15 mph (22'/sec) a mobile moves 44 half-wavelengths in one sec. For data rates of 10 Kb/s, 227 bits are transmitted as a vehicle moving at 15 mph moves one-half wavelength. For data rates of 1 Kb/s, 11 bits are transmitted as a vehicle moving at 30 mph moves one-half wavelength. Thus, at 10 Kb/s and 15 mph a 100 bit data block is transmitted as the vehicle moves one-quarter wavelength. At 1 Kb/s and 30 mph a 100 bit block is transmitted as the vehicle moves 5 wavelengths. It follows that the received signal level is relatively constant over one bit period, but not necessarily during the transmission of one data block. The mean signal level, however, being relatively constant over 50 m, is relatively constant over one transmitted data block.

Vehicle motion also broadens the power spectral density of the radiated signal on the order of 50 Hz at 20 mph. Because radiated spectra normally have bandwidths in excess of 10 KHz, this spectrum broading can be neglected below 1 GHz, for spectral occupancy calculations.

II-3 Spectral Emission Constraints

Conventional spectrum management policies involve division of that portion of the radio frequency spectrum assigned to the land mobile radio service into frequency channels. To prevent a user of any channel from interfering unduly with users in other channels, limitations on the spectrum of the transmitted signal are imposed. Also imposed are limitations on the total transmitted power, in order to avoid excessive interference with others using the same frequency channel in other geographic locations.

Fig. 2-2 shows a spectral emission chart issued by Communications Canada [G2]. In Chapter 3 we demonstrate the determination of maximum digital transmission rate and the selection of modulator pulse shapes to enable the transmitted signal spectrum to tightly fit inside the proposed emission characteristic. We also determine the maximum allowable root-mean square bandwidth B for analog FM transmission of voice signals; this maximum value of B determines the maximum output signal-to-noise ratio. In Chapter 5 spectral emission constraints are re-examined in the context of Communication Canada's current Radio Standards (see Appendix) and their implications on the transmission of data and FM voice signals.

The division of the land mobile spectrum into frequency channels is not the only viable spectrum management option. Considerable interest exists in alternative schemes whereby each user shares the entire frequency band using spread spectrum multiple access (SSMA) [D1, C3, S1]. Definitive comparisons of the capabilities of SSMA with the more conventional channel assignment schemes have become available [C3].

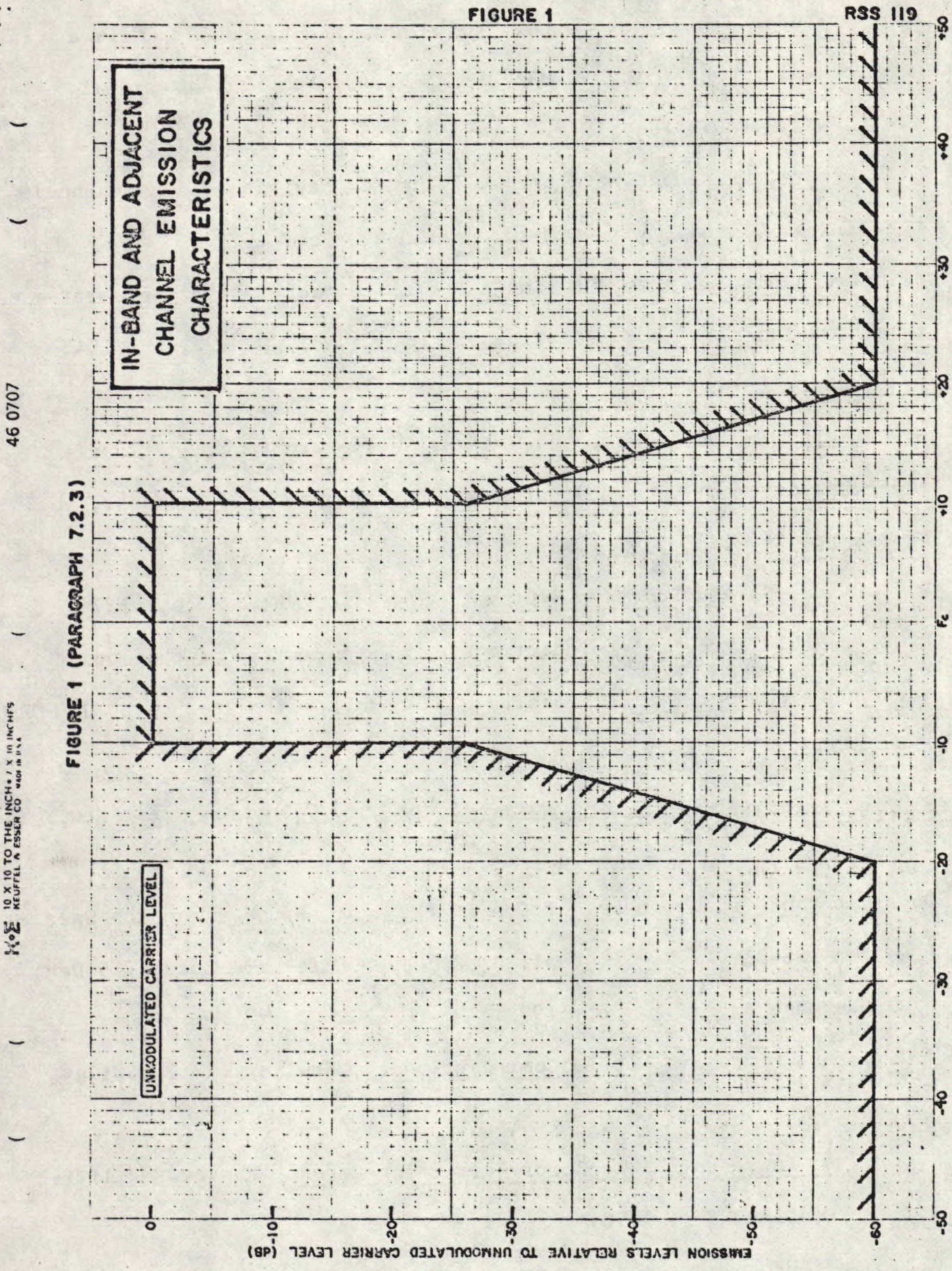

Fig. 2-2 In-band and adjacent-channel emission chart issued by Communications Canada.

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III SPECTRA OF ANALOG FM SIGNALS AND DIGITAL SIGNALS

III-1 Spectra of Voice Signals Transmitted Using Frequency Modulation

The transmitted signal s(t) which results when a message signal m(t) FM modulates a sinusoidal signal of carrier frequency f_c is

$$s(t) = \sqrt{2P} \cos [2\Pi f_c t + \theta + \int_{-\infty}^{t} m(\tau)\, d\tau] \qquad (3\text{-}1)$$

In (3-1) P is the average transmitted power and θ is a random phase angle uniformly distributed between zero and 2Π.

It is well known that an analytical expression for the power spectral density of a FM modulated signal is available if the message m(t) is Gaussian. In this case the "baseband" power spectral density V(f) of the transmitted signal is [R1]

$$V(f) = \int_{-\infty}^{\infty} v(\tau) \exp (-j2\Pi f\tau) d\tau \qquad (3\text{-}2a)$$

$$= 2\int_{0}^{\infty} v(\tau) \cos 2\Pi f\tau d\tau \qquad (3\text{-}2b)$$

where

$$v(\tau) = \exp [-2\int_{\infty}^{\infty} M(f) (\frac{\sin\Pi f\tau}{2\Pi f})^2 df] \qquad (3\text{-}3)$$

and M(f) is the power spectral density of the modulating signal m(t).

The spectral density of the signal actually transmitted is [R1]

$$S(f) = \frac{P}{2} [V(f+f_c) + V(f-f_c)] \qquad (3\text{-}4)$$

where f_c is the FM carrier frequency. In order that the power in the transmitted signal equal the unmodulated carrier power P, it is necessary that $\int_{-\infty}^{\infty} V(f) df = 1$. This latter condition is implicit in (3-2) and (3-3); inverse Fourier transformation of (3-2a) yields

$$v(\tau) = \int_{-\infty}^{\infty} V(f)\, e^{j2\Pi f\tau} df \qquad (3\text{-}2c)$$

from which $v(0) = \int_{-\infty}^{\infty} V(f) df$ follows. From (3-3) one sees that v(0) = 1 for any M(f).

For non-Gaussian signals $V(f)$ can be obtained analytically in two limiting cases [T1, R1]:

(1) when the transmitted signal has vanishingly small bandwidth in which case

$$V(f) = P_1(M(f)/f^2) + P_2\ \delta(f) \tag{3-5}$$

where constants P_1 and P_2 depend on the detailed form of $M(f)$ and on the fact that $\int_{-\infty}^{\infty} V(f)df = 1$.

(2) when the FM signal has a bandwidth which is wide relative to that of the message signal but narrow relative to the carrier frequency f_c. In this case

$$V(f) = P\ p(f) \tag{3-6}$$

where $p(f)$ is the amplitude probability density of $m(t)$. In the narrowband case, a fairly large fraction of the power may be in the carrier.

Both the Laplacian [C1, S1] and Gamma [C1, R2] functions defined below provide reasonable fits to the amplitude probability density of speech:

$$p(x) = \frac{1}{2b} \exp\ (-|x-a|/b) \qquad \text{(Laplacian)} \tag{3-7}$$

$$p(x) = \frac{c}{2\Gamma(d)}\ [C|x|]^{d-1} \exp\ (-c|x|) \qquad \text{(Gamma)} \tag{3-8}$$

Non-negative constants a, b, c and d are selected to fit the actual data; $\Gamma(x)$ is the Gamma function [P1] defined by $\Gamma(x) = x\Gamma(x-1)$ for any $x>0$. If x is an integer $\Gamma(x) = x!$

Clearly, speech is <u>not</u> a Gaussian random process, and an analytical expression for $V(f)$ is unavailable. To determine $V(f)$ for speech modulation, we are left with three alternatives:

1. Assume that speech is Gaussian and use (3-2) to obtain $V(f)$.
2. Use approximate numerical techniques to calculate $V(f)$, possibly

based on "typical" speech segments.

3. Estimate $V(f)$ using actual measured data.

The most attractive alternative from the point of view of the effort required is the first one, particularly since we will require $V(f)$ for various values of modulation index, and because convolution of $V(f)$ with various filter responses will be required for interference calculations in Chapter 4. Use of numerical techniques requires much care in the selection of "typical" speech samples and much computation. Spectral estimation presents its own difficulties [C2], particularly in low spectral energy regions, and involves considerable computation. For these reasons, and because $p(x)$ bears some resemblance to the Gaussian bell shape we select alternative 1 realizing that results based on the Gaussian assumption would ultimately require experimental verification.

The following equation has been proposed as an approximation to the long term average spectral density of continous speech [J1, F1]:

$$C(f) = f^4/(f^2+(70)^2)(f^2+(180)^2)(f^2+(400)^2)(f^2+(700)^2) \qquad (3\text{-}9)$$

where f is the frequency in Hz.

Communications Canada's proposed radio standards specification described in Chapter 2 require that the demodulator output signal prior to audio filtering should not fall within the cross-hatched region of Fig. 3-1. Thus, if the transmitter is FM modulated and FM demodulation is used, then speech must be filtered,prior to modulating the FM carrier,by a filter with transfer function

$$H(f) = \frac{jf+300}{jf+3000} \qquad (3\text{-}10)$$

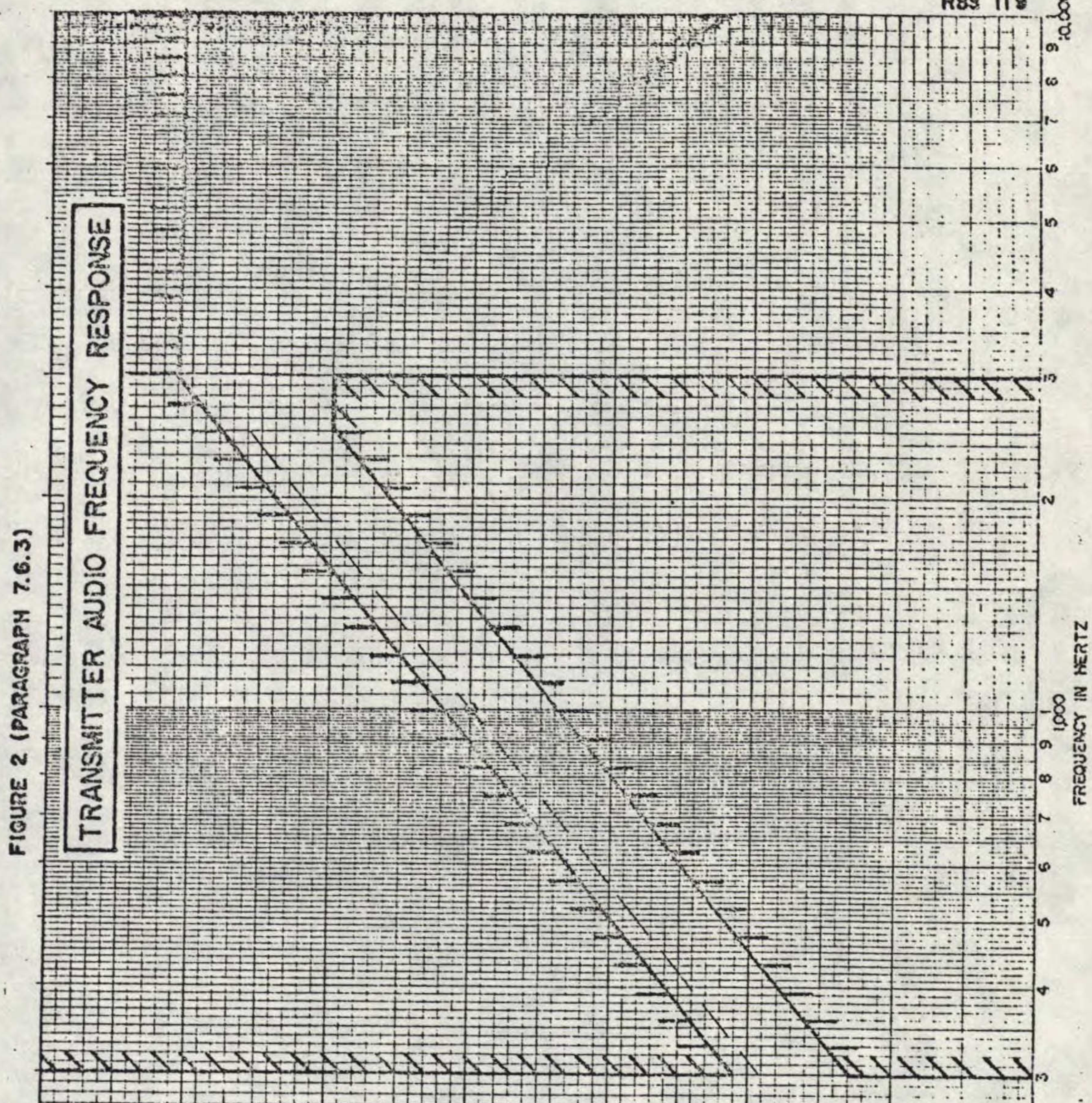

Fig. 3-1 Transmitter audio frequency response for land mobile channels.

where f is again frequency in Hz. To determine V(f) in (3-2) we use

$$M(f) = K|H(f)|^2 C(f) \tag{3-11}$$

where positive constant K determines the power level in M(f), and thus the FM spectral bandwidth. Fig. 3-2 shows M(f) and $M(f)/f^2$ where the spectral amplitudes are scaled such that the area under these spectra equals unity.

To actually calculate $v(\tau)$ in (3-3) we first write $M(f)/f^2$ as follows, were $a_1 = 70$, $a_2 = 180$, $a_3 = 400$, $a_4 = 700$ and $a_5 = 3000$:

$$M(f)/f^2 = \sum_{i=1}^{5} A_i/(f^2+a_i^{\,2}) \tag{3-12}$$

where

$$A_i = [(f^2+a_i^{\,2})M(f)/f^2]_{f^2 \to a_i^{\,2}} \tag{3-13}$$

The values of A_i obtained from (3-13) when K = 1 in (3-11) appear in Table 3-1.

Determination of $v(\tau)$ now requires evaluation of

$$I(a, \tau) = 2\int_{-\infty}^{\infty} \frac{(\sin \pi f|\tau|)^2}{(f^2+a^2)}\, df$$

$$= \int_{-\infty}^{\infty} \frac{df}{f^2+a^2} - \int_{-\infty}^{\infty} \frac{\cos 2\pi f|\tau|}{f^2+a^2}\, df$$

$$= (\Pi/a)[1 - \exp(-2\Pi a|\tau|)] \tag{3-14}$$

From (3-3), (3-12), (3-13) and (3-14) one obtains

$$v(\tau) = \exp[-\sum_{i=1}^{5}(A_i/4\Pi a_i)\{1 - \exp(-2\Pi a_i|\tau|)\}] \tag{3-15}$$

$$= \prod_{i=1}^{5} \exp(-b_i) \prod_{j=1}^{5} \exp[b_j \exp(-2\pi a_j|\tau|)] \tag{3-16a}$$

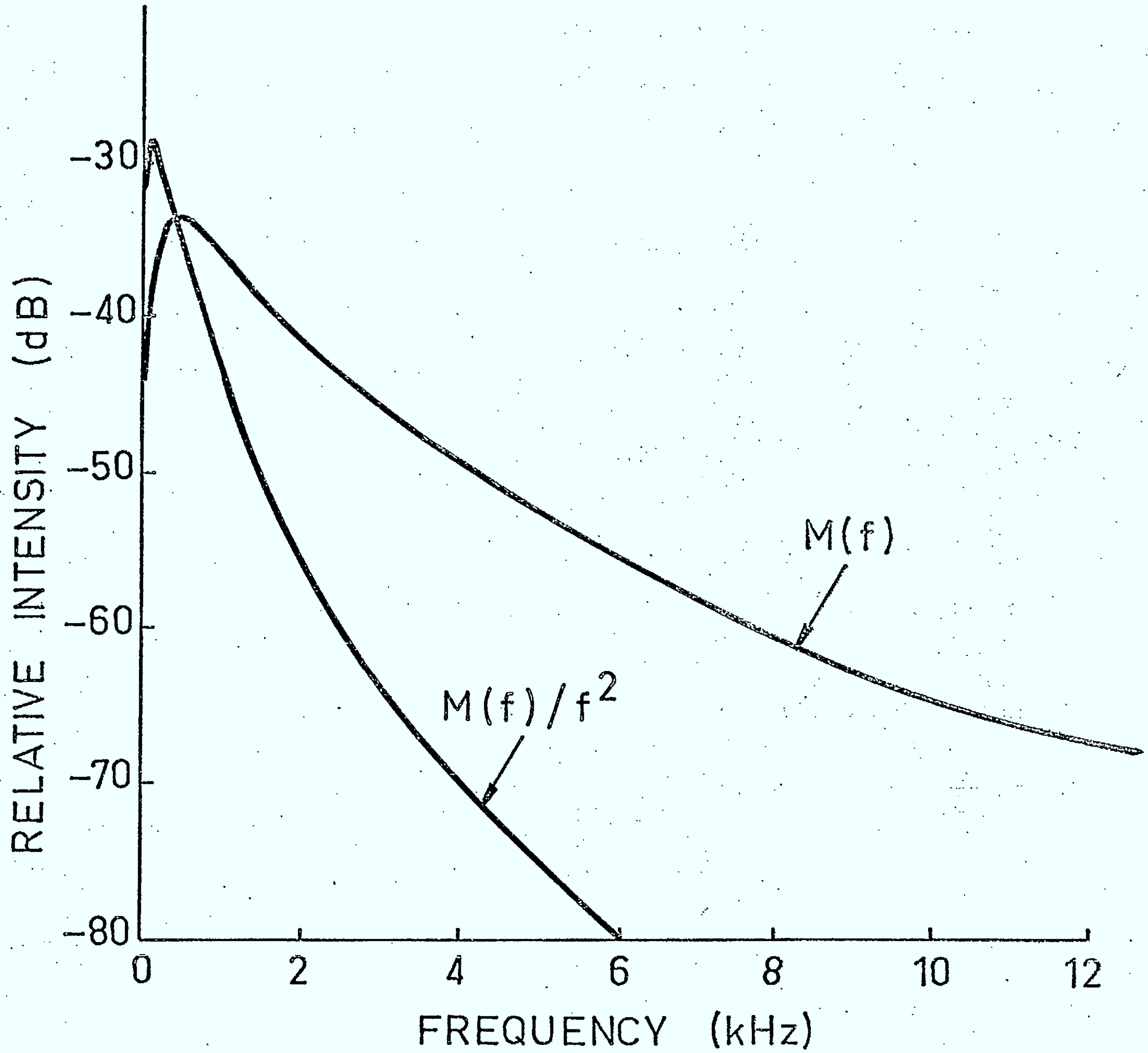

Fig. 3-2 Spectra of M(f) and $M(f)/f^2$.

i	a_i	$A_i (\times 10^{-13})$	$B_i/a_i (\times 10^{-13})$	$A_i/a_i (\times 10^{-16})$
1	70	-0.224	15.7	-3.20
2	180	1.40	-252	7.78
3	400	1.94	-776	4.85
4	700	-3.14	2198	-4.49
5	3000	0.132	-396	0.044

$$\sum_{i=1}^{5} A_i/a_i = 4.99 \times 10^{-16} \qquad \sum_{i=1}^{5} B_i/a_i = 7.90 \times 10^{-11}$$

$$\frac{1}{4\pi} \sum_{i=1}^{5} A_i/a_i = 3.97 \times 10^{-17} \qquad \frac{1}{4\pi} \sum_{i=1}^{5} B_i/a_i = 6.28 \times 10^{-12}$$

$$\pi \sum_{i=1}^{5} A_i/a_i = 1.57 \times 10^{-15} \qquad \pi \sum_{i=1}^{5} B_i/a_i = 2.48 \times 10^{-10}$$

Table 3-1 Values of A_i from (3-13), $B_i/a_i = -a_i A_i$, and A_i/a_i for $K = 1$

where

$$b_i = A_i/4\Pi a_i \tag{3-16b}$$

Substitution of (3-15) into (3-3) yields V(f) which is probably most easily evaluated for various values of f by using a numerical integration algorithm. An alternative is to express V(f) as follows, were $\otimes$ denotes convolution:

$$V(f) = [\prod_{i=1}^{5} \exp (-b_i)] \; H_1(f) \otimes H_2(f) \otimes H_3(f) \otimes H_4(f) \otimes H_5(f) \tag{3-17}$$

and $H_i(f)$ is the Fourier transform of $\exp[b_i \exp(-2\Pi a_i \, |\tau|)]$. This latter term can be represented by an infinite series as follows:

$$\exp\,[b_i \; \exp\;(-2\pi a_i|\tau|)] = \sum_{k=0}^{\infty} \frac{[b_i \; \exp(-2\pi a_i|\tau|)]^k}{k!} \tag{3-18}$$

From (18) one obtains

$$H_i(f) = \sum_{k=0}^{\infty} \frac{(b_i)^k}{k!} \; [\frac{ka_i/\pi}{f^2+(ka_i)^2}] \tag{3-19}$$

Since each term in the series decreases rapidly as k increases, the first few terms may be used to approximate $H_i(f)$ and the error can be calculated. This approach was used by Ferris [F2] to calculate the spectral characteristics of FDM-FM multiplexed signals.

Techniques are available to replace V(f) in (3-2) by bounds [P2, P3, R1, A1] which, unfortunately are rather complex.

Eqn. (3-15) can be written as follows,

$$v(\tau) = \exp\;[\{R_f(\tau) - R_f(0)\}] \tag{3-20}$$

where

$$R_f(\tau) = \sum_{i=1}^{5} \left(\frac{A_i}{4\Pi a_i}\right) \exp(-2\Pi a_i|\tau|) \tag{3-21}$$

denotes the auto-correlation function of $f(t) = \int m(t)\,dt$. Fig. 3-3 shows $\rho(\tau) = R_f(\tau)/R_f(0)$ vs τ. Since the pre-emphasis function $H(f)$ in (3-10) acts as a differentiator over much of the audio frequency range, $\rho(\tau)$ is similar to the actual correlation function of speech as obtained from inverse Fourier transformation of $C(f)$ in (3-9).

From (3-21) one obtains

$$R_f(0) = \sum_{i=1}^{5} A_i/4\Pi a_i \tag{3-22}$$

For $K = 1$ in (3-11), $R_f(0) = 3.97 \times 10^{-17}$ as indicated in Table 3-1. If $K = \alpha K_0$ where $K_0 = (3.97 \times 10^{-17})^{-1} = 2.52 \times 10^{16}$ then all $\{A_i\}$ are scaled by αK_0, $R_f(0) = \alpha$ and

$$\begin{aligned} v(\tau) &= \exp(\alpha[\rho(\tau)-1]) \\ &= [\exp[\rho(\tau)-1])^{\alpha} \\ &= e^{-\alpha} e^{\alpha\rho(\tau)} \end{aligned} \tag{3-23}$$

The more narrow is pulse $v(\tau)$ the larger must be α, and the wider is the spectrum $V(f)$.

It is seen that $V(f)$ has finite energy (a spectral impulse) at $f=0$ equal to $v(\infty) = e^{-\alpha}$. As α increases this spectral impulse in the form of carrier frequency power decreases, and the energy goes into the sidebands of $V(f)$. The power in these sidebands $\int_{-\infty}^{\infty} V_{ac}(f)\,df$ where

$$\begin{aligned} V_{ac}(f) &= 2\int_0^{\infty}(v(\tau)-e^{-\alpha})\cos 2\Pi f\tau d f \\ &= 2e^{-\alpha}\int_0^{\infty} [e^{\alpha\rho(\tau)}-1] \cos 2\Pi f\tau d\tau \end{aligned} \tag{3-24}$$

A bandwidth measure of interest is the root mean square bandwidth B(Hz) of the "baseband" spectrum $V(f)$ in (3-2) where we define [H1]:

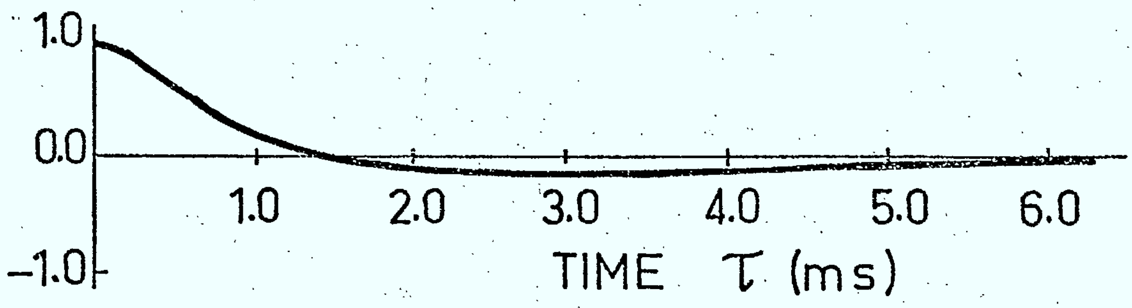

Fig. 3-3 - Autocorrelation function $\rho(\tau)$ for speech prefiltered by the filter with transfer characteristic in Fig. 3-1.

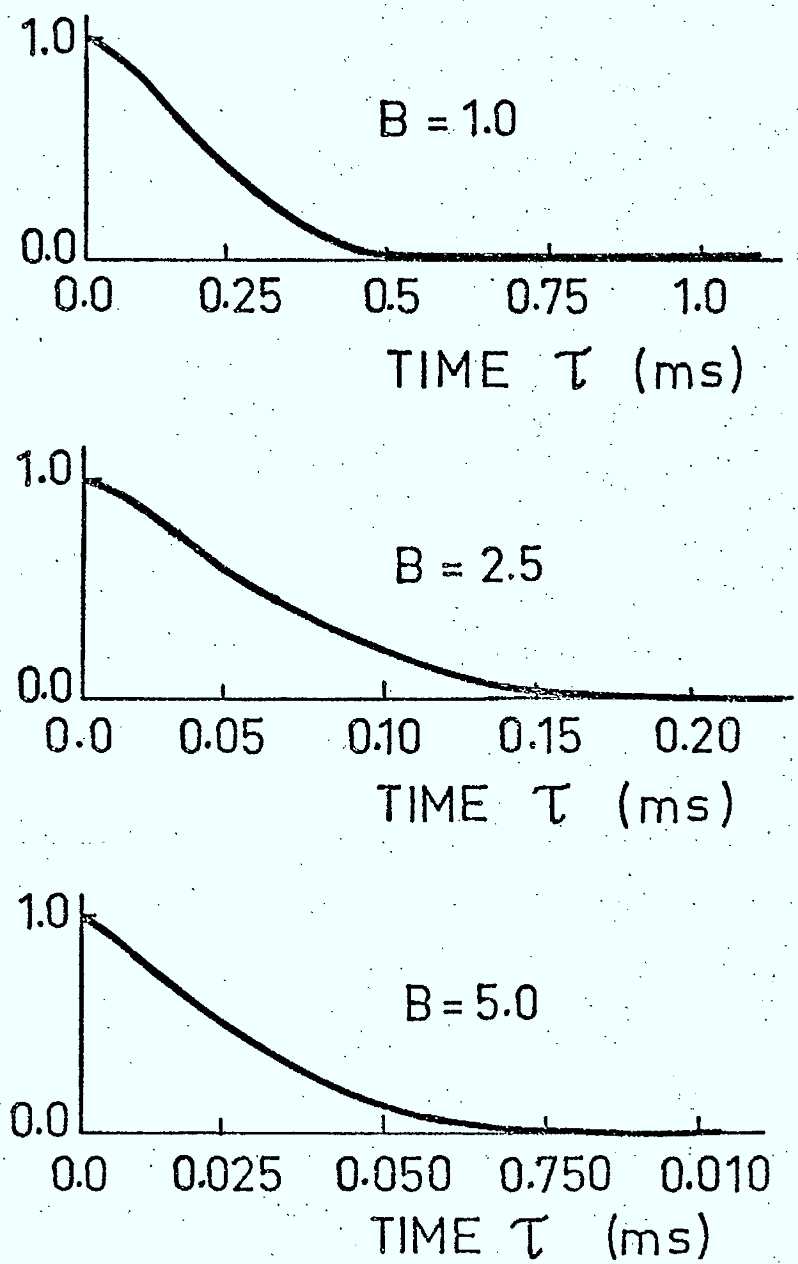

Fig. 3-4 - Autocorrelation function $v(\tau) - v(\infty)$ for FM speech modulated signal.

$$B = [\int_{-\infty}^{\infty} f^2 V(f)df \;/\; \int_{-\infty}^{\infty} V(f)df]^{1/2} \tag{3-25}$$

(Various other definitions for B exist and differ from ours by a constant factor.) For Gaussian signals [Hl]

$$B = (1/2\Pi)[\int_{-\infty}^{\infty} M(f)df]^{1/2} \tag{3-26}$$

From (3-26), (3-22) and the definition of $\alpha=R_f(0)$ it follows that B varies in direct proportion to $\sqrt{\alpha}$. Evaluation of

$$\int_{-\infty}^{\infty} M(f)df = \alpha K_o \sum_{i=1}^{5} \int_{-\infty}^{\infty} (\frac{B_i}{f^2+a_i^2})df$$

$$= \alpha K_o [\Pi \sum_{i=1}^{5} B_i/a_i) \tag{3-27}$$

where

$$B_i = (f^2+a_i^2)M(f)]_{f^2=a_i^2}$$

$$= -a_i^2 A_i \tag{3-28}$$

yields

$$B^2 = \alpha K_o [\frac{1}{4\Pi} \sum_{i=1}^{5} (B_i/a_i)] \tag{3-29}$$

From Table 3-1 and $K_o = 2.52 \times 10^{16}$ it follows that

$$B = \sqrt{(2.52 \times 10^{16})(6.28 \times 10^{-12})}\sqrt{\alpha}$$

$$= 0.398\sqrt{\alpha} \quad \text{(kHz)} \tag{3-30}$$

Fig. 3-4 shows $v(\tau)-e^{-\alpha}$ vs τ and Fig. 3-5 shows $V_{ac}(f)$ vs f for various values of B. It is seen the spectral emission constraint at $f = \pm 20$ kHz governs the allowable value of B.

Fig. 3-6 shows V(f) vs B for f = 20 kHz. One sees that $B \simeq 2.9$ is the largest value for which the spectral emission constraint is satisfied.

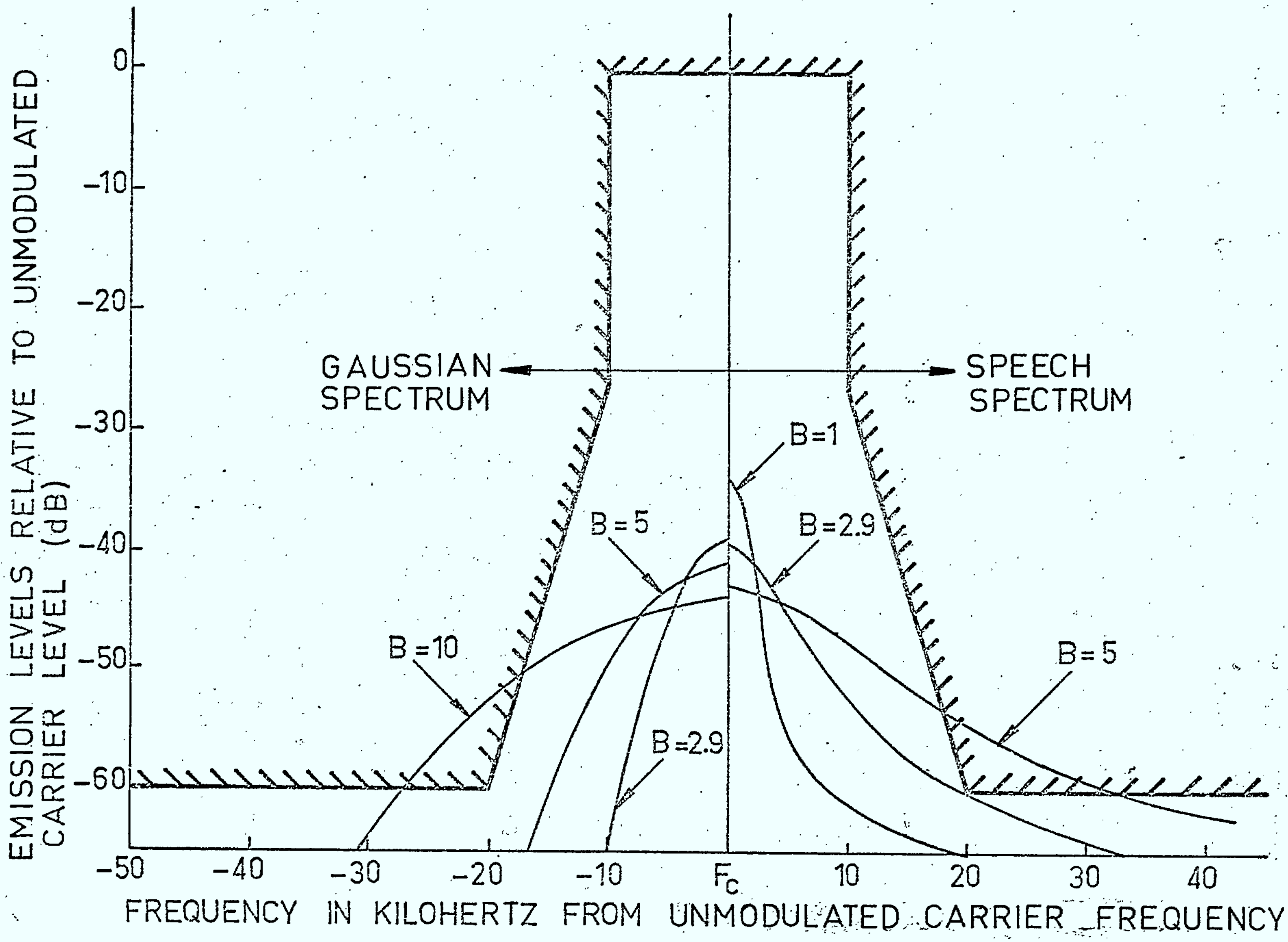

Fig. 3-5 Power spectral density $V_{ac}(f)$ for an FM signal modulated by prefiltered speech for rms bandwidths B = 1.0, 2.9, and 5.0 kHz. Also shown is the Gaussian spectrum for these values of B.

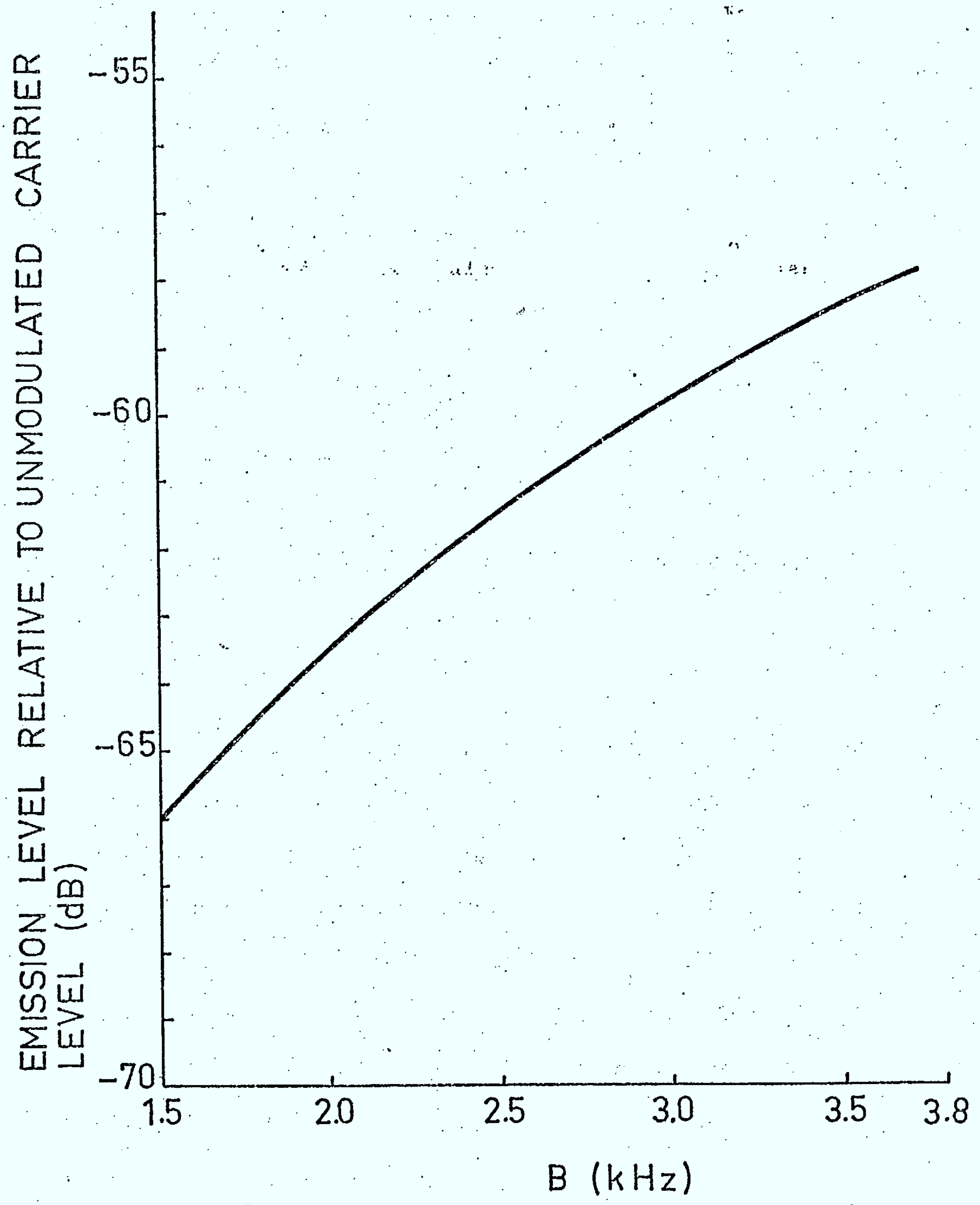

Fig. 3-6 V(f) vs. rms bandwidth B for f = 20 kHz.

Also plotted on Fig. 3-5 is

$$G(f) = (\sqrt{2\Pi}\ B)^{-1} \exp(-f^2/2B^2) \tag{3-31}$$

for various values of B. One sees that the wideband Gaussian approximation referred to earlier is a poor approximation to V(f) for B $\stackrel{\sim}{<}$ 5. Comparison of Fig. 3-5 and $M(f)/f^2$ in Fig. 3-2 indicates that the narrowband approximation for V(f) is also a poor one. For the transmission bandwidths considered, actual calculation of the spectral density V(f) is required.

III-2 Digital Signal Spectra-Linear Modulation (PAM)

In pulse amplitude modulation (PAM) the digital data modulator in Fig. 3-7 generates a signal s(t) whose constitutive pulse amplitudes depend on the symbol sequences $\{a_k\}$ impressed upon the modulator [L1]. Thus

$$s(t) = [\sum_k a_k\ g(t-kT)]\ \sqrt{2P}\ \cos\ (\omega_c t + \theta) \tag{3-32}$$

where a_k is the amplitude corresponding to the k^{th} input symbol, g(t) is the basic modulator pulse shape, ω_c is the (radian) carrier signal frequency, θ is the carrier phase, $T=R_B^{-1}$ where R_B is the pulse transmission rate (the baud rate) and P is the power of the unmodulated carrier signal. In L-level PAM, a_k can assume one of L values and the bit rate $R=R_B \log_2 L$. In binary PAM L=2 and R_B=R. The PAM levels are normally spaced in accordance with the relation

$$a_k = 2kd-(L-1)d \qquad (k = 0,1, \ldots, L-1) \tag{3-33}$$

A special case of binary PAM is conventional ON-OFF amplitude shift keying (ASK) where $a_0=0$, $a_1=A$ and

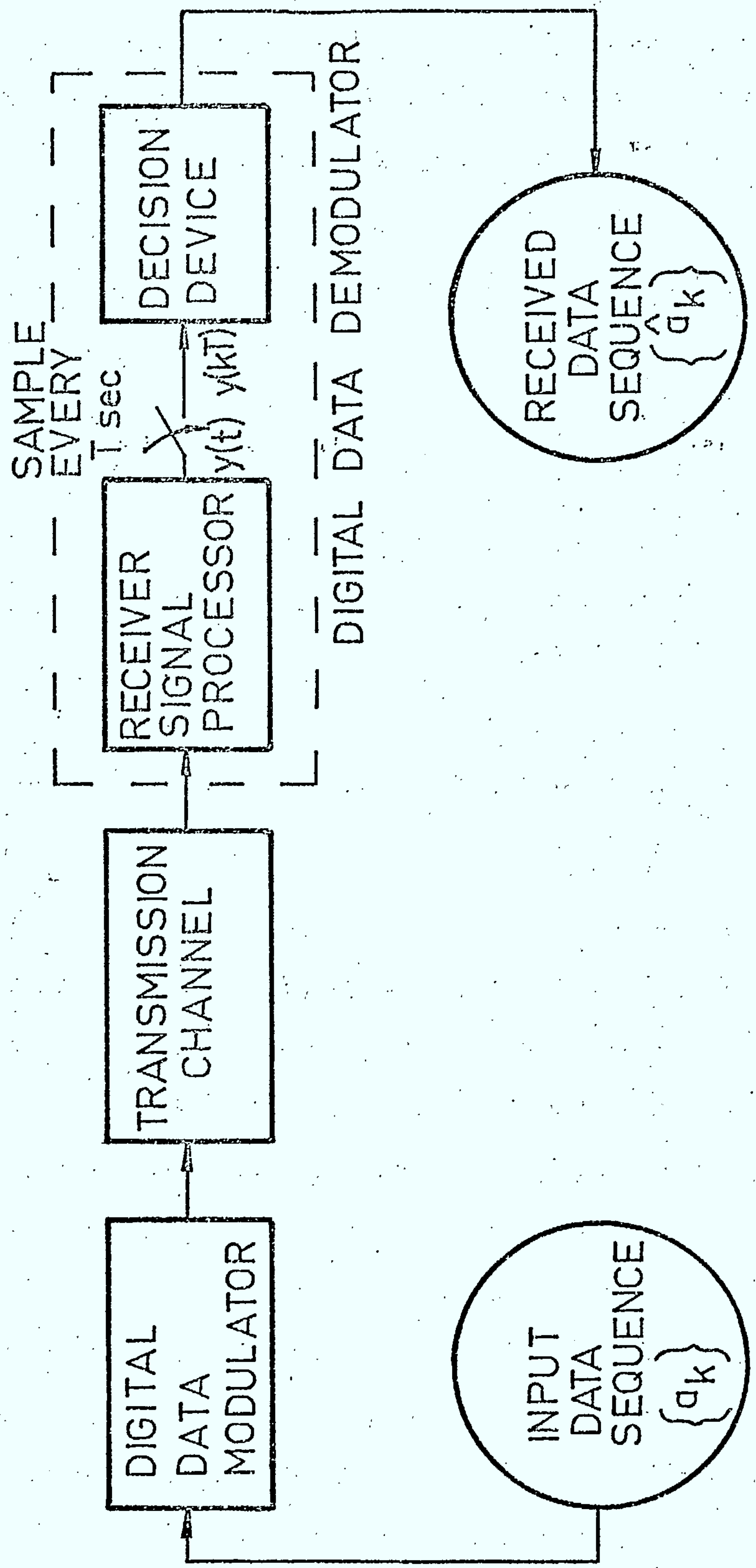

Fig. 3-7 Block diagram of a digital data transmission system

$$g(t) = \begin{bmatrix} 1 & 0 \leq t \leq T \\ 0 & t < 0;\ t > T \end{bmatrix} \qquad (3\text{-}34)$$

Each pulse $g(t)$ is limited in duration to $[0,T]$ and the absence or presence of a signal conveys the information.

Conventional ASK has three disadvantages:

1. The bandwidth of the transmitted signal $s(t)$ is excessive.
2. The decision (slicing) threshold at the receiver depends on the received signal level which can range over 90 dB.
3. The peak-power capabilities of the transmitter are excessive by 3 dB, since (for equiprobable bits) no energy is transmitted half of the time.

The last two disadvantages are obviated by using binary antipodal ASK where $a_0=-A$ and $a_1=A$ in which case the decision level remains at zero independent of A. The disadvantage of this "improvement" is that coherent detection is required; with ON-OFF ASK incoherent detection is feasible. The excessive bandwidth which results from rectangular-pulse baseband signalling can be reduced by shaping the pulses $g(t)$ and removing the constraint that they be non-zero only for $0 \leq t < T$. The result is some increase in the complexity of the digital data modulator and demodulator. However, implementation of modulators for PAM pulse shaping is now greatly facilitated by digital electronic and microprocessor technology for digital waveform generation [G1, S2]. The ever decreasing costs of digital components relieves the designer of the need to rely solely on analog components to generate the required modulator pulse shapes.

Use of more than two signal levels in the presence of large and rapid signal level variations requires adaptive adjustment of decision thresholds [W1]. Signal level estimation [L2, F3] needed for such adjustment would complicate the receiver and introduce estimation errors and ensuing error rate degredations.

Large and rapid signal level variations also require use of pulse shapes which are zero at all sampling times except the zeroth to avoid intersymbol interference (isi). Existing isi-combatting receiver techniques [B1, S3, T2] are not directly applicable since these assume fixed and known signal levels.

The restriction to two-level PAM also prevents use of partial response signalling [K1, K2] which requires at least two decision thresholds both of which cannot be zero.

The power spectral density of a transmitted L-level PAM signal s(t) is S(f) where

$$S(f) = \frac{1}{2}\left[P(f+f_c) + P(f-f_c)\right] \tag{3-35}$$

In (3-35) P(f) is the power spectral density of the baseband signal $p(t) = \sum_k a_k\, g(t-kT)$ in(3-32). If the symbols a_k are statistically independent then [L2]

$$P(f) = \frac{1}{T^2}\sum_{j=-\infty}^{\infty}\left|\sum_{k=1}^{L} p_k a_k G(j/T)\right|^2 \delta\!\left(f-\frac{k}{T}\right)+\frac{1}{T}\sum_{k=1}^{L} p_k {a_k}^2 |G(f)|^2 \tag{3-36}$$

where p_k is the probability of level a_k, $\delta(f)$ is the unit impulse, and G(f) is the Fourier transform of g(t); thus

$$G(f) = \int_{-\infty}^{\infty} g(t)\, e^{-j2\pi ft}\, dt \tag{3-37}$$

The first term in (3-36) is a non-information bearing line spectrum which vanishes when any positive level a_k is balanced by a corresponding negative level $a_j=-a_k$ with probability $p_j= p_k$. Thus, for binary antipodal PAM with $a_1=A$ and $a_2=-A$, the first term in (3-36) is zero. For binary ON-OFF PAM the first term is not zero. In any case this first term depends on G(f) only at discrete values of frequency f=j/T, where j is an integer.

The challenge for the designer of the digital data modulator is to specify g(t) (or equivalently G(f)) to meet existing bandwidth constraints such as those in Fig. 2-2, to maximize the bit rate R, to enable digital data demodulation which is easily implemented and which yields an acceptable bit-error rate in the presence of noise and interference, and to maintain tolerable levels of interference to other users.

Fig. 3-8 shows three different pulse shapes, all of which show zero intersymbol interference at the sampling times t=nT, $n\neq 0$. The two non-rectangular pulses are derived from the family of pulse shapes

$$f(t) = K\left(\frac{\sin \Pi t/T}{\Pi t/T}\right)\left(\frac{\cos\, \alpha\, \Pi t/T}{1-(2\alpha t/T)^2}\right) \tag{3-38}$$

where α ($0\leq\alpha\leq 1$) controls the pulse "excess" bandwidth [L1] and K the pulse power. For $\alpha=0$, $f_a(t)$ in Fig. 3-8 results, whereas $f_c(t)$ occurs for $\alpha=1$. The larger is α, the less is the performance degredation caused by sample-timing errors [L1].

The actual modulator pulse g(t) is derived from the corresponding modulator pulse shape f(t) in Fig. 3-8 as follows: for the rectangular pulse, $g_b(t) = f_b(t)$ while for the family specified by (3-38) g(t) is the inverse Fourier transform of $\sqrt{F(f)}$ where F(f) is the Fourier transform of

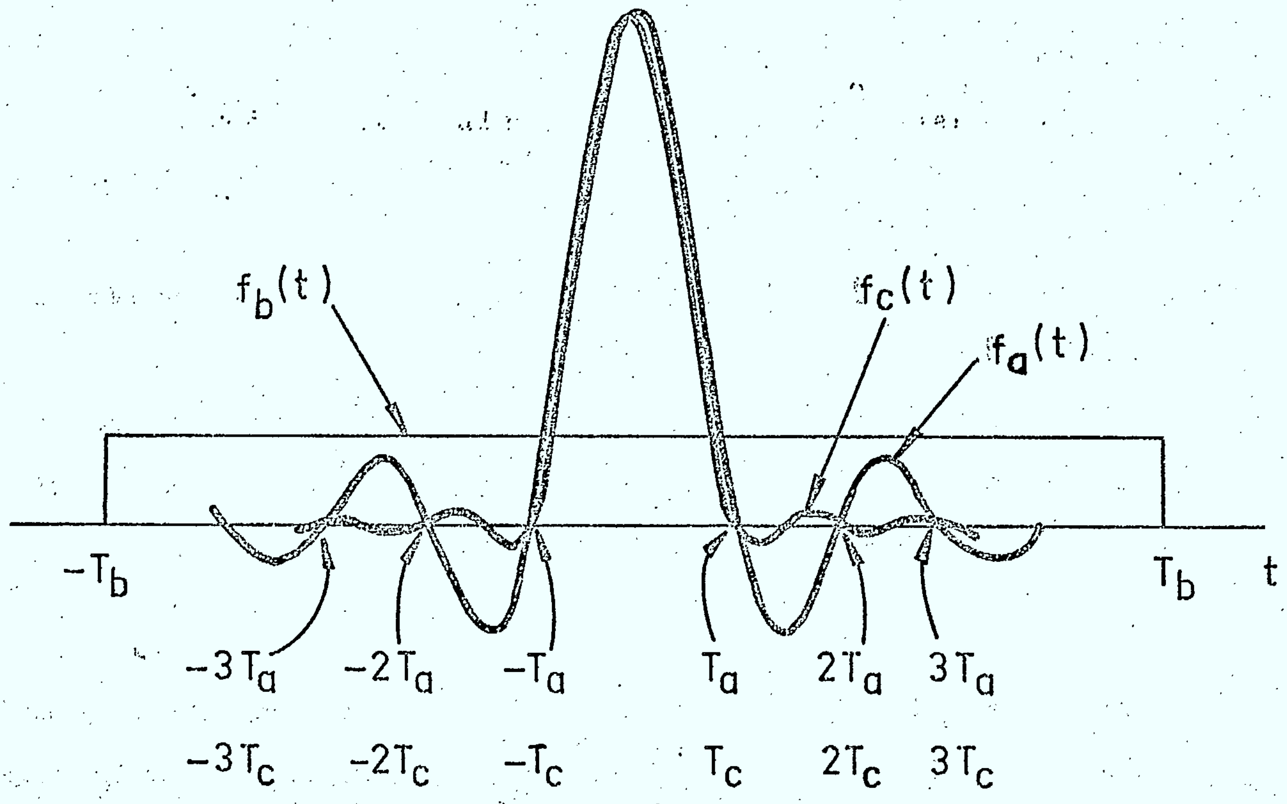

Fig. 3-8 PAM pulse shapes where f(t) is the pulse shape following matched filtering at the demodulator.

f(t). These comments follow from the fact that minimization of bit error probability on a white Gaussian noise all-pass channel requires that the modulator pulse shape and receiver baseband filter impulse response be identical [L1].

Fig. 3-9 shows the corresponding plots of S(f); the sampling periods have been selected to enable the spectra to fit tightly inside proposed spectral emission constraint of Fig. 2-2. Pulse shape (b) is time-limited but has spectral components at all frequencies, whereas pulse shapes (a) and (c) are not time-limited but are strictly limited in bandwidth.

Determination of the maximum permissible data rate for binary signalling is easiest for the sin t/t - type pulse with spectral density

$$S_a(f) = \begin{cases} K_a = 1/2\, W_a & |f| < W_a \\ 0 & |f| > W_a \end{cases} \tag{3-39}$$

Amplitude K_a is selected to ensure that the transmitted power is equal to that of the unmodulated carrier signal. The transmitted power P_T in the transmitted signal

$$s(t) = p(t)\,\sqrt{2P}\,\cos\,(\omega_c t + \phi) \tag{3-40}$$

is

$$P_T = P \int_{-\infty}^{\infty} P(f)\,df \tag{3-41}$$

from which it follows immediately that $\int_{-\infty}^{\infty} P(f)\,df = 1$, where P(f) is the power spectral density of baseband signal p(t).

Use of (3-39) together with

$$10\,\log_{10} E(f) = 8 - 3.4\,|f| \qquad 10 \leq |f| \leq 20 \tag{3-42}$$

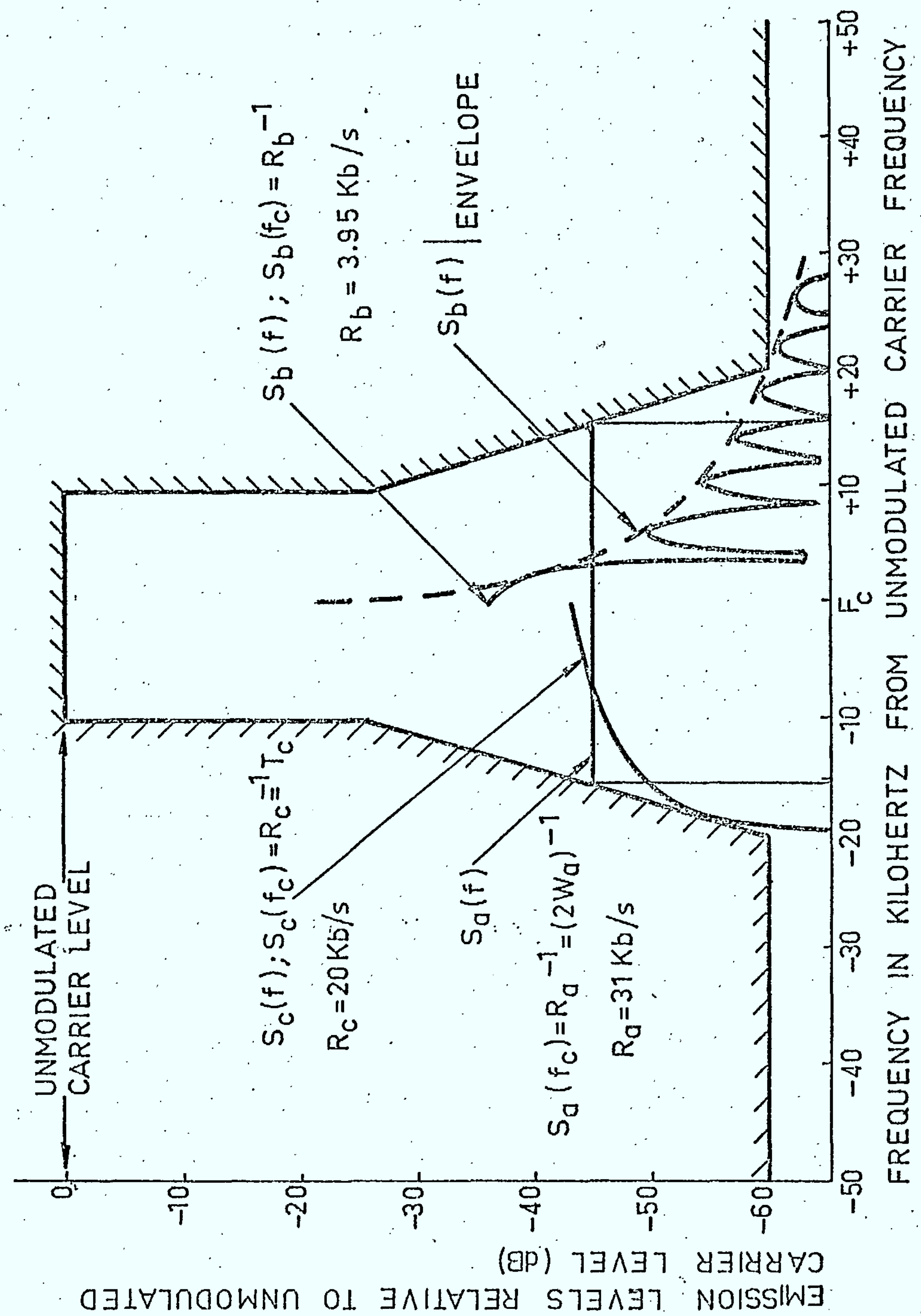

Fig. 3-9 Power spectral density for the pulse shapes in Fig. 3-8 under maximum bit-rate conditions.

where E(f) is the spectral emission constraint boundary in Fig. 2-2 for frequencies $10 \leq |f| \leq 20$ (f in kHz), yields the maximum value of $W_a \simeq 15.5$ kHz. This value of W_a solves the following equation:

$$10 \log_{10} (1/2\ W_a) = 8 - 3.4\ W_a \tag{3-43}$$

It follows that the maximum bit rate is $R_a = 2\ W_a \simeq 31$ Kb/s. It should be noted that the total absence of spectral energy outside the frequency band $f_c \pm 15.5$ kHz in Fig. 3-9 implies zero interference with adjacent channels whose separation from f_c exceeds 31 kHz.

A similar determination of the maximum bit rate can be made for conventional binary antipodal ASK for which

$$S_b(f) = K_b\ T_b\ (\sin \pi f T_b/\pi f T_b)^2 \tag{3-44a}$$

To ensure $\int_{-\infty}^{\infty} S_b(f)df = 1$, $K_b = 1$

To estimate the maximum bit rate $R_b = T_b^{-1}$ for which $S_b(f)$ tightly fits the spectral emission constraint one can determine the minimum T_b for which the envelope $1/(\pi f)^2 T_b$ equals -60 dB for f = 20 kHz. The resulting value of R_b is obtained from

$$R_b/(\pi\ 20{,}000)^2 = 10^{-6} \tag{3-44b}$$

which yields $R_b \simeq 3.95$ Kb/s. Thus, the maximum bit rate permissible for conventional rectangular pulses under the proposed emission constraint is 12.7% of that permitted for sin t/t - type pulses. In addition some interference is generated for adjacent channels at arbitrarily large frequencies from f_c.

The power spectral density $S_c(f)$ of the transmitted signal with $\alpha = 1$ is

$$S_c(f) = \begin{cases} K_c = T_c & |f| < (1-\alpha)/2T_c \\ \frac{T_c}{2}\left[1-\sin[(\pi T_c/\alpha)(f-\frac{1}{2T_c})]\right] & (1-\alpha)/2T_c \leq |f| \leq (1+\alpha)/2T_c \\ 0 & |f| > (1+\alpha)/2T_c \end{cases} \quad (3\text{-}45)$$

To fit $S_c(f)$ tightly inside the spectral emission characteristic one plots $10 \log_{10} S_c(f)$ for various values of T_c and thereby determines the maximum bit rate $R_c = T_c^{-1}$. The result will always exceed $(1 - \frac{\alpha}{2})R_a$ where R_a is the maximum bit rate for the case $\alpha=0$. Fig. 3-9 yields $R_c \simeq 20$ Kb/s which, because of the emission constraint roll-off from -26dB to -60dB, greatly exceeds the lower bound $R_a/2 \simeq 15.5$ Kb/s. To avoid co-channel interference, adjacent frequency channels must be separated by at least 40 kHz.

Reduction of α would increase the maximum allowable bit rate as well as the timing error sensitivity. Thus, selection of excess bandwidth α permits a trade-off between these quantities.

Selection of other pulse shapes limited to [0, T] such as triangles, half-cosine pulses and other zero-isi pulses not limited to [0, T] results in bit rates which lie between $R_a \simeq 31$ Kb/s and $R_b \simeq 4$ Kb/s.

III-3 Digital Signal Spectra-Non-Linear Modulation (FSK and PSK)

Define

$$x(t) = \sum_{k=-\infty}^{\infty} a_k \, g(t-kT) \tag{3-46}$$

where $\{a_k\}$ denotes the symbol sequence, $g(t)$ a basic pulse shape and T the symbol period. A digital FM modulated carrier signal $s(t)$ with radian frequence ω_c and average power P results when $x(t)$ modulates the carrier, as follows:

$$s(t) = \sqrt{2P} \cos \left(\omega_c t + \omega_d \int_{-\infty}^{t} x(\tau)\, d\tau + \theta\right) \tag{3-47}$$

where θ is a (random) phase angle and the instantaneous frequency is $\omega_c + \omega_d\, x(t)$. When $x(t)$ is normalized to have unity peak amplitude, ω_d is the peak frequency deviation of $s(t)$. In general, increasing ω_d increases the bandwidth of $s(t)$ and reduces the obtainable bit error rate.

When $g(t)$ in (3-46) is a rectangular pulse

$$g(t) = \begin{cases} 1 & 0<t\leq T \\ 0 & t\leq 0,\ t>T \end{cases} \tag{3-48}$$

and conventional frequency shift keying (FSK) results. It is not yet clear how best to select $g(t)$ to give $s(t)$ desired spectral properties and maintain low error rates.

A conventional digital phase modulated signal is of the form

$$s(t) = \sqrt{2P} \sum_{k=-\infty}^{\infty} g(t-kT)(\cos \omega_c t + \phi_k) \tag{3-49a}$$

where rectangular pulse $g(t)$ is defined in (3-48). Phase sequency $\{\phi_k\}$ varies in accordance with symbol sequence $\{a_k\}$; for example $\phi_k = 2k\pi/L$ $(k=1,2,\ldots,L)$ where L is the number of symbol levels.

Analog FM and PM spectra are similar to one another and differ only because of an integration or differentiation of the baseband waveform. Digital FSK and PSK differ by virtue of the fact that phase continuity is normally preserved across FSK symbol boundaries, but not across PSK symbol boundaries. Thus, in contrast to digital FSK, digital PSK does not permit bandwidth expansion and is similar in spectral characteristics to PAM.

Digital FM signals have a constant amplitude or envelope, as do PSK signals when g(t) in (3-49a) consists of rectangular pulses. Constant envelope signals are desirable under certain conditions, including satellite communication applications where constant output power amplifiers are more efficient than quasi-linear power amplifiers [R4].

Eqn. (3-48) may be rewritten as follows:

$$s(t) = \sqrt{2P}\{[\sum_{k=-\infty}^{\infty} b_k\, g(t-kT)]\cos\omega_c t + [\sum_{k=-\infty}^{\infty} c_k\, g(t-kT)]\sin\omega_c t\} \qquad (3\text{-}49b)$$

where

$$b_k = \cos\phi_k \qquad (3\text{-}49c)$$

$$-c_k = \sin\phi_k \qquad (3\text{-}49d)$$

$$\phi_k = 2\pi k/L \qquad (3\text{-}49e)$$

Thus, conventional PSK consists of quadrature PAM signals. For binary PSK $\phi_k = 0$ or π, in which case $b_k = \pm 1$ and $c_k = 0$. The resulting signal s(t) is identical to PAM with rectangular baseband pulses. From the discussion in Section III-2 it follows that S(f), the power spectral density of s(t) in (3-49) is given by the response $S_b(f)$ in Fig. 3-9.

Optimal detection of conventional PSK on Gaussian white noise channels involves synchronous demodulation of the quadrature carriers sin

$\omega_c t$ and cos $\omega_c t$, processing of each of these demodulated signals by filters matched to g(t), sampling of the filtered outputs and optimal decision making [L1, W1]. Because the optimum decision device compares ratios of numbers from the demodulated quadrature carriers, PSK on Rayleigh fading channels need not be restricted to binary.

Use of differential PSK (DPSK) obviates the need to track the carrier signal phase. In DPSK the received symbol phase is compared with that of the previously detected symbol. The added receiver complexity resulting from differential detection is usually more than offset by the absence of carrier phase recovery equipment. Some error rate increase occurs since errors tend to occur in pairs [L1, W1]; if the detected phase of a symbol is incorrect, the (differentially) detected phase of the following symbol is also likely to be in error. A transmitter power increase of not more than 3dB reduces the DPSK error rate to that obtainable using PSK. DPSK signal spectra are like PSK spectra.

It is well known [G2, L1, R1, R3, S4, T3] that determination of the power spectral density S(f) of the FSK signal in (3-47) is very difficult except in the two limiting cases discussed in Section III-1. In the case of low deviation ratio ω_d it follows from (3-5) that the bandwidth of the transmitted FM signal can actually be less than that of the modulating data signal. For arbitrary modulating signals s(t) in (3-46) or even for conventional rectangular FSK pulses g(t) in (3-48) closed form expressions for S(f) are unavailable, and approximate methods must be used to find S(f). Garrison's [G2] approach involves approximation of the baseband modulating signal pulses g(t) in (3-46) by a duration-limited step-wise approximation. The method is fairly general but as Garrison states [G2]:

".... the casual reader should not be mislead into believing that arbitrary configurations for premodulation filtering are amenable to analysis. Machine computation time can readily become prohibitive unless the pulse approximation can be conveniently bounded in both duration and quantization number."

When the peak frequency deviation $f_d=\omega_d/2\pi=1/4T=R/4$ where R is the baud rate, minimum shift keying (MSK) results [K3, A2, G3, R4]. For binary MSK the spectrum S(f) is known exactly [G3, S5]:

$$S_{MSK}(f) = \frac{8PT(1+\cos 4\pi fT)}{\pi^2(1-16T^2f^2)^2} \tag{3-50}$$

where f is the frequency offset from the carrier.

Fig. 3-10 shows $S_{MSK}(f)$ plotted to fit tightly within the spectral emission constraints of Fig. 2-2. These constraints permit a bit rate of R=36 Kbs.

It is possible to obtain closed form expressions for M-ary (M>2) MSK spectra [S5]. The rather complex expressions are not presented here, since noise rather than bandwidth would likely limit the transmission rate of MSK on actual mobile channels, in which case binary rather than M-ary signalling would be used.

MSK signals can be regarded as resulting from modulation of quadrature carriers by alternate data bits [S5, A2, K3]. It follows that coherent detection of these binary data streams is possible and results in an error

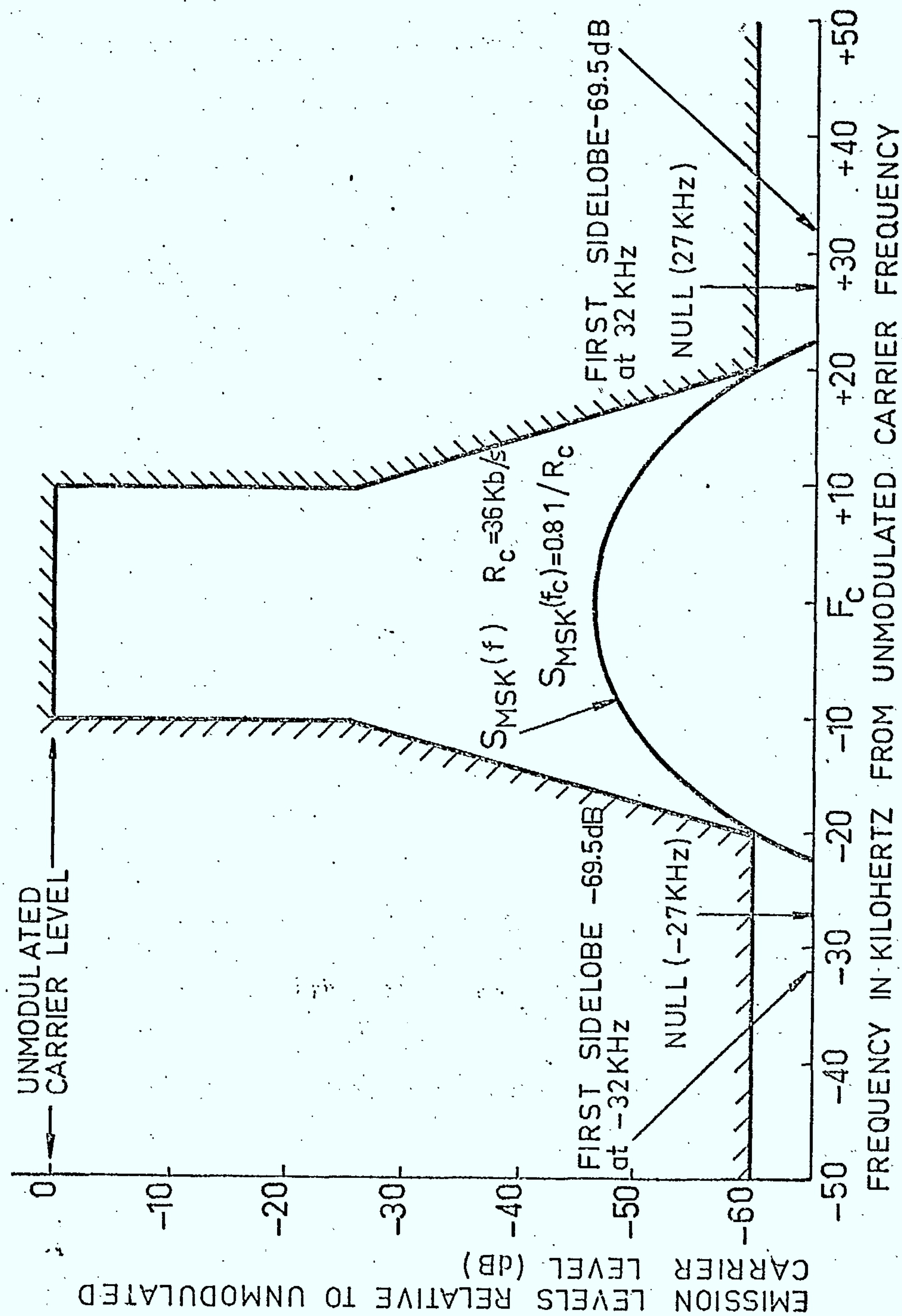

Fig. 3-10 Power spectral density for binary MSK signals.

probability obtainable using antipodal signalling, which represents a 3dB detectability improvement over that obtainable using detection of the MSK signals viewed as binary orthogonal [S5, K3].

Other types of signals, including offset quadrature shift keying (OQPSK) and sinusoidal frequency shift keying (SFSK) have spectra with most of the energy at low frequencies. However none of these permit data rates as high as 36 Kbs. Fig. 3-11 compares S(f) for OQPSK, SFSK and MSK.

The spectral roll-off rate is often shown for digital FM signals, as is available for MSK and MSK-type signals [G3, S5]. However, as noted by Reiffen and White [R4]:

> "The signal spectrum approach taken in many references in the literature, e.g., [9-11], focuses on how fast the signal spectrum "rolls off" with frequency. Although relevant and yielding useful results, this approach is incomplete in that it does not take the receiver window into account."

The interaction of receiver filters (windows) and adjacent-channel signals is considered in detail in Chapter 4; however we note here that for channel separations of less than approximately twice the data rate, adjacent-channel interference (crosstalk) is lower for MSK signals than for more general MSK-type signals.

Determination of the maximum data rate permitted under the spectral emission constraints in Fig. 2-2 for FSK signals with arbitrary deviation ratios $2f_d/R$ is a computationally tedious task. In effect, numerically calculated spectra would have to be plotted on Fig. 2-2 for various data

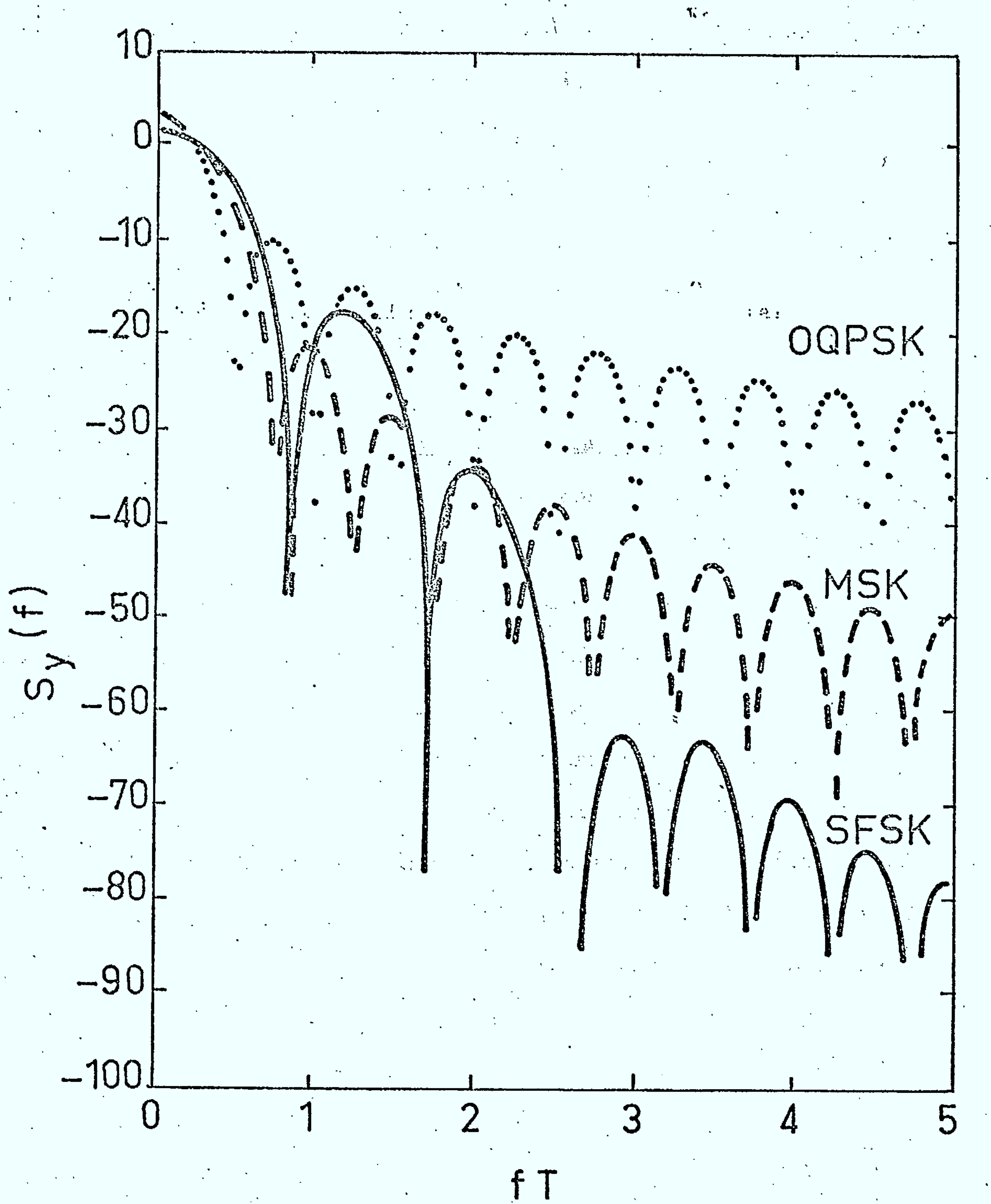

Fig. 3-11 FM spectra for MSK, OQPSK and SFSK signals (after Simon [S5]).

rates R. That value of R yielding a snug fit to the constraint would define the maximum R. In calculating spectra it would be necessary to determine the spectral amplitude relative to the unmodulated carrier power.

In considering ways to improve the spectral efficiency and hence the maximum permissible data rate for PSK and FSK signalling, we generalize (3-49) in two ways. First, the amplitudes of the rectangular pulses are replaced by arbitrary pulse shapes h(t). Second, the constant phase angles ϕ_n are replaced by phase functions $f_{\phi_n}(t)$. For simplicity of implementation, we restrict $f_{\phi_n}(t) = \phi_n f(t)$. With these generalizations, the phase modulated signal becomes

$$s(t) = \sqrt{2P} \sum_{k=-\infty}^{\infty} h(t-kT) \cos(\omega_c t + \sum_{n=-\infty}^{\infty} \phi_n f(t-nT)) \quad 0<t\leq T \qquad (3\text{-}51)$$

Neither h(t) nor f(t) need be restricted to the interval [0, T].

The first generalization yields quadrature PAM signals and therefore PAM spectra. Unlike PAM, however, multilevel signalling is feasible as explained earlier. Disadvantages of using non-rectangular pulses include more complex pulse-shaping instrumentation at the transmitter and receiver and loss of the constant envelope signalling property.

Concerning the second approach to improving PSK spectra, various authors [P4, G4] have determined power spectra when f(t) in (3-51) assumes a variety of pulse shapes; however only Prabhu [P4] appears to have calculated the resulting symbol error probability. Prabhu's detailed study yields two conclusions. First, considerable bandwidth savings result from using phase functions f(t) which are non-zero for one or two pulse periods. Specifically binary PSK with

$$f(t) = \begin{cases} 0.5\ [1+\cos(\pi t/T)] & |t| < T \\ 0 & |t| > T \end{cases} \tag{3-52}$$

results in a bandwidth which is 15% of that resulting from using constant phase functions ϕ_k; bandwidth being defined in terms of 99% spectral energy containment. A 1.4 dB transmitter power increase on a white Gaussian channel restores the bit error probability to that of conventional binary PSK.

A second result of interest occurs when conventional (rectangular pulse) PSK is filtered prior to transmission by four-pole Butterworth filters to produce a spectral bandwidth equal to that of similarly filtered PSK with a phase functions given by (3-52). For a bit error probability $p=10^{-6}$, the PSK signal with the phase function of (3-52) requires 0.8 dB more energy than that of the constant phase PSK signal. If the definition of bandwidth is changed to imply a 99.9% spectral energy containment, PSK with the time-varying phase is inferior by 0.5 dB to constant phase PSK. Since filtering of PSK signals prior to their transmission is equivalent to selecting non-rectangular amplitude pulse shapes g(t) in (3-49a) or h(t) in (3-51), Prabhu's results suggest that appropriately selected amplitude pulse shapes with constant phase PSK is not easily improved upon by utilizing time-varying phase functions.

Improvement of the bandwidth properties of FSK by use of non-rectangular pulse shapes g(t) in (3-46) has been demonstrated for MSK signals [S5]. As Fig. 3-11 shows, use of a pulse yielding a sinusoidal frequency deviation results in a transmitted binary FSK spectrum which, for levels less than -30 dB below the spectrum's centre frequency falls of much faster than binary MSK. However bit error probabilities for non-conventional FSK pulses are unavailable. The selection of arbitrary pulse shapes to yield desired FSK spectra and favourable error-rate performance appears to be a very difficult problem.

III-4 Summary and Concluding Remarks

By selecting PAM, PSK and FSK signals whose power spectral densities fall tightly within the emission constraints in Figs. 3-9 and 3-10, data rates of 36 Kbs are obtainable for binary FSK and rates of between 20 Kbs and 31 Kbs are obtainable for binary PAM and PSK. These rates are for double sideband signalling. Use of single sideband (SSB) signalling doubles these rates. However SSB introduces serious practical difficulties which include potentially devastating carrier phase and symbol timing errors [L1, L2, F3, G5]. Vestigital sideband modulation results in data rates between those obtainable using SSB and DSB, and also provides for more tolerance to timing errors than is inherent in SSB [D1, H2, R5, L1]. The way in which excess bandwidth, modulator pulse shapes, synchronization algorithms, pilot tone power allocation, fading rates and levels, and error rates interact is a complicated matter which is currently under study by several institutions.

Land mobile channel constraints currently in effect require at least 99% of the transmitted energy to be within ±8 kHz of the carrier frequency. This constraint limits the data rates to 16.2 Kb/s for PAM signals with no excess bandwidth, 8.1 Kb/s for PAM signals having raised cosine spectra with maximum excess bandwidth, 1.8 Kb/s for conventional rectangular PAM pulses and 13.3 Kbs for MSK signals. Again, these rates apply to binary DSB signal formats; use of multilevel signalling (for PSK), SSB or vestigital sideband would increase the allowable data rates as well as the associated bit-error rates. These existing spectral constraints also require a much smaller value of rms bandwidth B for voice modulated FM signals than B ≈ 2.9 kHz as permitted by the constraints in Fig. 2-2.

In considering FM transmission of speech it is natural to ask whether or not a pre-emphasis characteristic different from the one proposed in Fig. 3-1 would result in transmitted signals with increased values of rms bandwidth B and a spectral shape which still meets the spectral emission constraints in Fig. 2-2. Preliminary work based on existing analysis [B2, V1, C3] for large signal-to-noise (linearized) FM systems suggests that use of a pre-emphasis filter different from that in Fig. 3-1 may be advantageous. It remains to determine the extent to which B could be increased, the additional complexity inherent in the new pre-emphasis filter, the non-linear effects near signal-to-noise threshold and the effects on the subjective quality of the received speech.

An alternative, indirect way to restructure the spectra of a transmitted signal derived from speech is to use digital transmission. Use of binary MSK at 36 Kbs is more than sufficient for transmission of adaptive, differentially encoded speech [C5, H3, F5]. In fact fewer than 20 Kbs may be adequate, with the remaining 16 Kbs being available for forward error correction. Whether digital transmission is more cost-effective than analog transmission of speech is apparently unresolved at this time.

At the maximum bit rates determined for PAM, PSK and FSK, decoded bit-error probabilities may be excessive. Some alternatives for combatting these errors, most of which occur in bursts, are as follows:

1. Increase the transmitter power
2. Reduce the noise and/or interference
3. Reduce the data rate
4. Employ spatial diversity
5. Reduce the radio coverage area of a region, thereby reducing loss due to transmitter-receiver separation
6. Employ forward error correction (FEC)
7. Employ some form of variable-rate information transmission.

Increasing the transmitter power is a "brute-force" approach which would be possible only if the existing transmitter power was below the lawful maximum.

Reduction of the noise implies a more expensive receiver or reduced co-channel and/or adjacent-channel interference. Reduced adjacent-channel interference implies tighter out-of-band emission standards or increased channel separation. Tightened emission standards require narrowed transmission spectra via reduced data rates. Increased channel separation implies fewer available channels. Reduced co-channel interference means fewer channels available to any cellular region [J1] which implies increased waiting time or blocking probability and reduced overall system information throughput.

Lowering the bit rate reduces bit error rate p since the output signal-to-noise ratio increases as R^{-1} (see Chapter 4). However, the channel bandwidth occupied is also reduced in proportion to R. Utilization of this available bandwidth can result in a further SNR increase - and a corresponding further improvement in bit-error rate. For example, the binary PAM baseband waveform could frequency modulate a carrier whose frequency deviation is low enough to meet spectral emission standards. Demodulation by an FM receiver operating above threshold in white Gaussian noise yields an output signal-to-noise ratio which is proportional to $(B/R)^2$ P/RNo where P/No is the carrier power-to-noise density ratio and B is the rms bandwidth of the transmitted signal [W1, L1]. Since B tends to change slowly with R, the bit-error probability at any given signal level above threshold varies as R^{-3}.

Even in the absence of vehicle motion spatial diversity increases P/RNo, by virtue of signal averaging. Various diversity combining schemes have been analysed and shown to yield large SNR improvements even with receiver noise correlations [J1, F4]. During vehicle motion spatial diversity greatly reduces SNR variations due to Rayleigh fading, shadowing and changes in transmitter-receiver separation, as noted in Chapter 2.

Reduction in radio coverage area reduces received signal level variations resulting from changes in receiver-transmitter separation. The penalty is more base stations. An added benefit is fewer mobiles per base station, which increases the potential information throughput over the entire coverage area. The selection of base station sites, coverage areas, and assignment of frequency channels to coverage areas is a spectrum management problem of considerable current interest and importance [J1, K4].

Forward error correction is a form of time diversity which has been shown to be useful in reducing decoded bit-error rates on fading channels when the raw bit-error rate is below a threshold (typically $\simeq 10^{-2}$) [D2, M2, C4, F4]. FEC is particularly appropriate when transmission of real-time information such as speech is required.

Variable rate transmission is often implemented via automatic repeat request (ARQ) with error detection as noted in Chapter 2, and is available for transmission of non-real-time information such as text [D2, M2, F4]. Both FEC and ARQ reduce the actual information transmission rate but in an efficient way [G6, L3].

Finally, we note that signalling at very high data rates (typically $R \gtrsim 50$ Kbs) results in a transmitted spectrum which meets the emission constraints in Fig. 2-2. However, at these high rates interference would, undoubtedly, result in such high bit-error rates that no channel encoding scheme would be effective in combatting these errors [D2]. However, repeated transmission of each bit with receiver averaging would reduce the bit error probability even in low signal-to-noise-plus-interference environments; this error control approach is inherent in spread spectrum systems referred to in Section 2-3.

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IV EFFECTS OF INTERFERENCE AND NOISE ON DIGITAL DATA TRANSMISSION

In this chapter the effects of co-channel interference, adjacent-channel interference (cross-talk) and noise on digital data transmission are considered. We restrict our attention to binary PAM with zero isi and to binary MSK and MSK-type signals. The PAM results apply with obvious minor modifications to PSK and DPSK, as explained in Chapter 3. All of these signals have attributes which make them suitable for use on land mobile radio channels, and as this chapter shows, these signals also have favourable cross-talk characteristics.

IV-1 Digital Data Demodulation

For coherent detection of PAM (or binary PSK) signals the receiver signal processor is normally as shown in Fig. 4-1 [L1, W1, L2]. When the only source of disturbance is white noise, the (baseband) filter which maximizes the output signal-to-noise ratio in the absence of intersymbol interference is one matched to the basic pulse shape g(t) defined in Section 3-2 [L1, W1, L2]. When the noise is Gaussian as well as white, matched filtering also minimizes bit-error probability p, which can in this case be calculated as the area under the "tails" of a Gaussian distribution [W1, L1]. Calculation of p is of considerable interest, since this measure is the one which is most widely used to characterize the performance of digital communication channels. Receiver optimization for DPSK or multilevel PSK with zero isi and white noise disturbance also involves matched filtering of the baseband waveform [L1, W1].

For conventional M-ary FSK signals, receiver design for optimum p is possible when the noise is white Gaussian [L1, L2, W1]. In this case

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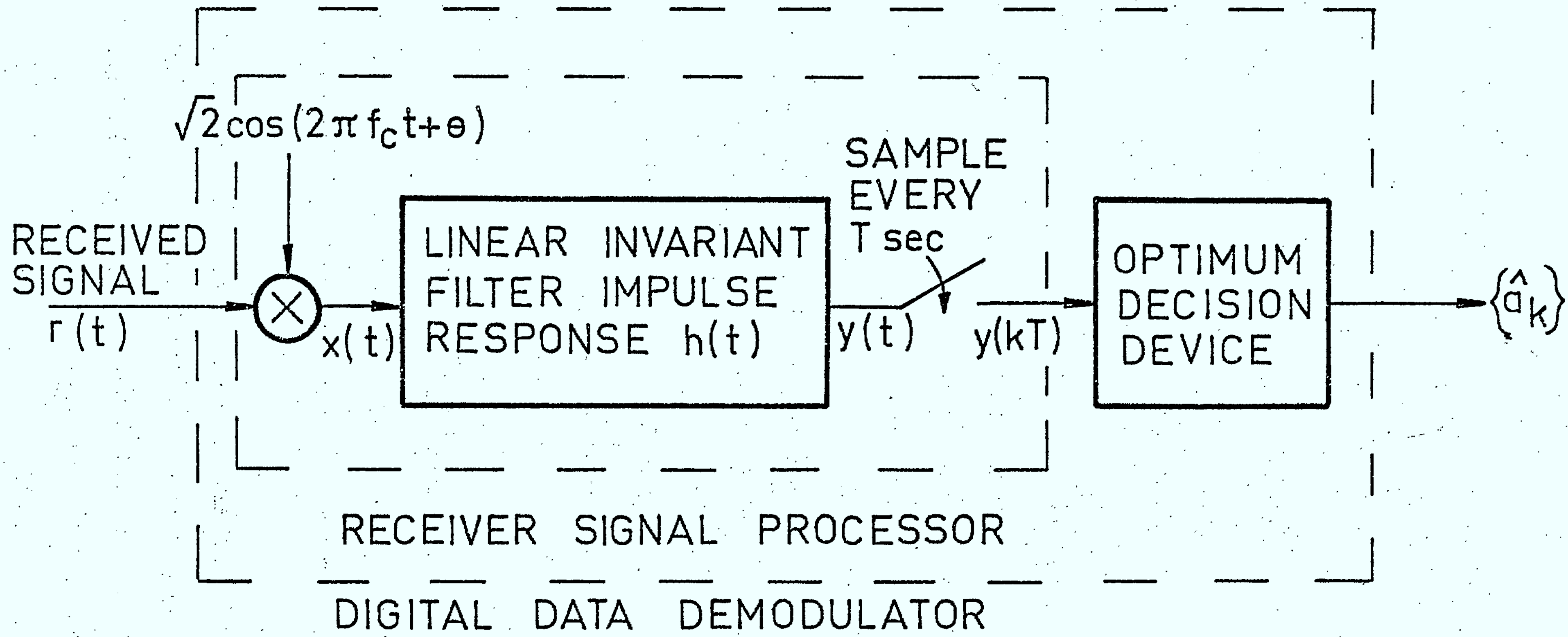

Fig. 4-1 Receiver for coherent demodulation of PAM signals with zero isi.

the receiver signal processor in Fig. 3-7 consists of parallel filters with each one matched to one of the M possible FSK pulses. Optimum incoherent reception of FSK signals is also possible using parallel filters, with a loss of up to 3 dB signal energy in comparison with coherent detection [L1, L2, W1]. When a conventional FM (analog) receiver consisting of a bandpass filter, limiter-discriminator and integrate-and-dump circuit is used as the receiver signal processor, optimization requires computer simulation and is possible only for the specific class of bandpass filters considered [T1].

As noted in Section 3-3, MSK signals and MSK-type signals are a special case of FSK signals which occur when the frequency deviation equals 1/4T where T is the basic symbol period. MSK signals can be regarded as consisting of in-phase and quadrature PAM signals each of data rate 1/2T and staggered with respect to each other by T sec. [K1, E1, G1]. Thus,

$$S(t) = \sqrt{P}\left[\sum_{i\ \mathrm{even}} a_i g(t-iT)\cos(2\pi f_c t+\theta) + \sum_{i\ \mathrm{odd}} a_i g(t-iT)\sin(2\pi f_c t+\theta)\right] \tag{4-1}$$

where s(t) is the transmitted MSK signal, θ is a random phase angle, P is the transmitted power and g(t) is the baseband pulse shape, where for MSK-type signals

$$g(t) = \begin{cases} \frac{1}{\sqrt{T}}\cos\phi(t) & |t| \leq T \\ 0 & |t| > T \end{cases} \tag{4-2}$$

and

$$\phi(t) = \begin{cases} \pm\pi/2 \\ \text{arbitrary} \\ 0 \\ \pm\pi/2 \pm \phi(t-T) \\ \pm\pi/2 \end{cases} \tag{4-3}$$

For MSK signals

$$\phi(t) = \pi t/2T \qquad |t| \leq T \tag{4-4}$$

From (4-2) one sees that both the in-phase and quadrature data signal in (4-1) is a PAM signal with basic pulse shape g(t) and zero isi. It follows that each data stream can be detected coherently as indicated in Fig. 4-2. If the only disturbance is white Gaussian noise, the receiver in Fig. 4-2 maximizes the output signal-to-noise ratio and also minimizes bit-error probability p.

For reasons of implementation simplicity it is convenient to replace the two filter branches in Fig. 4-2 by one, as in Fig. 4-1, and to sample its output every T sec. If this single filter is selected carefully, the performance degredation can be relatively small [K2].

IV-2 Approaches to Determining the Effects of Interference and Noise on Performance

Determination of the bit-error probability p for digital data transmission systems is based heavily on the assumption of Gaussian noise as the sole source of disturbance; specifically it is the fact that linearly filtered Gaussian signals are themselves Gaussian signals which allows the amplitude probability density of the baseband filter output samples y(kT) in Figs. 4-1 and 4-2 to be determined. When non-Gaussian interference is included as part of the input r(t), determination of the statistics of the baseband filter output is usually impossible, as is exact calculation of the bit-error probability p or the receiver which minimizes p. Determination of the output signal-to-noise ratio SNR is possible for most receivers

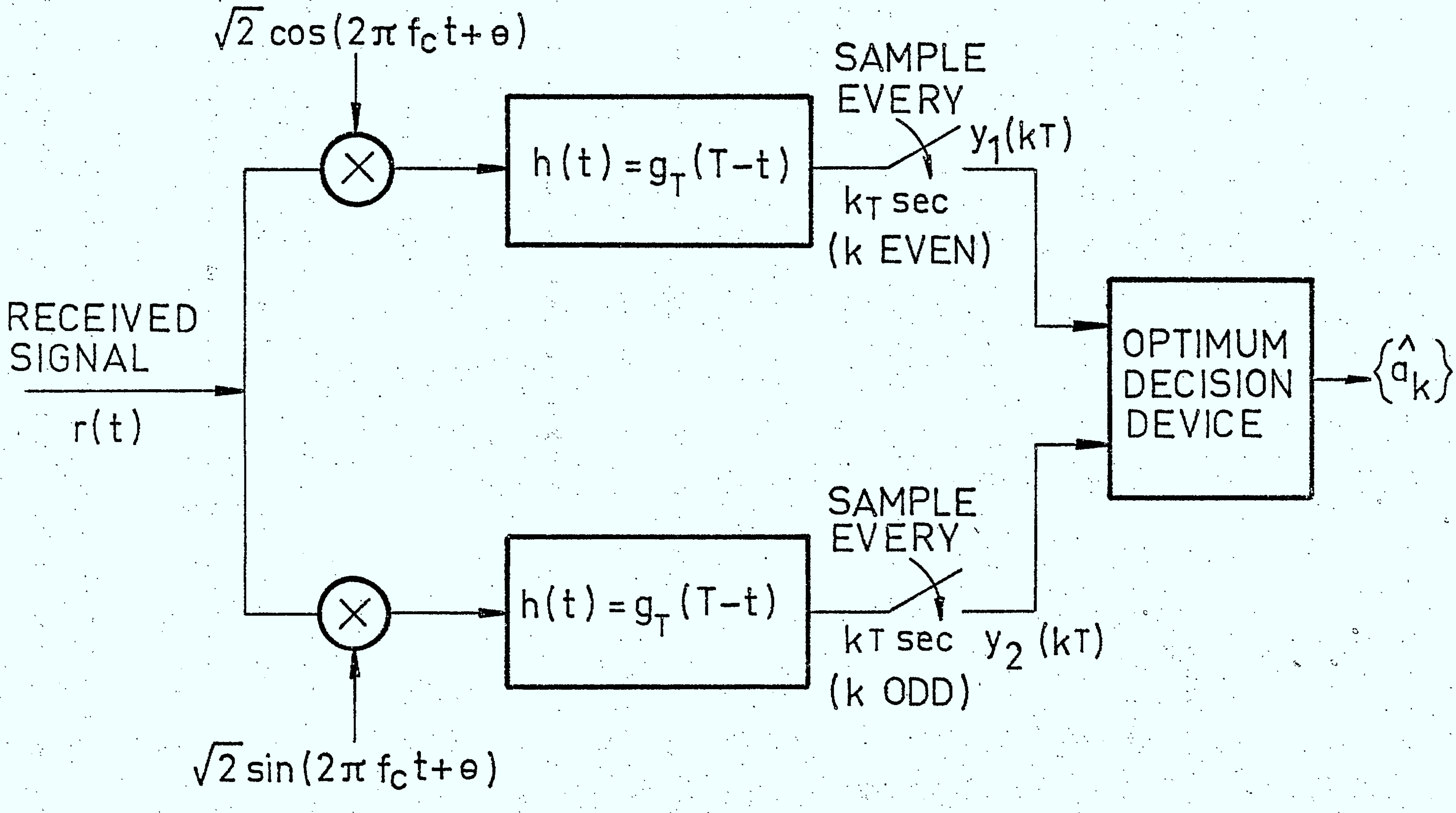

Fig. 4-2 Optimum MSK receiver; h(t) is the impulse response of the filters

including those in Figs. 4-1 and 4-2 as we show in the following sections. While one expects that a larger SNR should imply a lower error probability, it is not possible to prove that such is always the case.

As indicated earlier, the baseband filter which maximizes SNR can be determined, given the power spectral density of the received noise plus interference [W1, L1, L2]. In land mobile radio applications two difficulties arise: first, the levels of the different interfering signals vary continually, which causes the power spectral density of the noise plus interference to change continually; and second, the receiver baseband filter should be constrained to be one which maintains zero isi. For these reasons, a filter matched to the modulator pulse shape g(t) is often employed; this filter is optimum when the only disturbance is white Gaussian noise.

When many interfering signals are present, none of which are dominant, the central limit theorem shows that the totality of this interference tends to be Gaussian [W1] in which case error probability calculation is straightforward.

When noise plus interference is dominated by a single interfering signal, attempts are sometimes made to upper bound the effects of this interference by determining the phase difference between the desired and interfering signals which maximizes output interference-to-signal ratio [W2, K4]. Sometimes attempts are made to determine the error probability in this worst case. The difficulty with this approach is that the average error probability, which may be estimated by averaging over phase differences between the desired and interfering signal, is often not readily obtained.

An alternative approach is to determine the level of white Gaussian noise needed to give a SNR value identical to that resulting from the interfering signal. The error probability for the Gaussian noise case should be indicative of that caused by the interfering signal [K3].

In determining SNR or the error probability p for mobile channels in the 900 MHZ region the signal and interference level can be assumed constant during the reception of a single bit as explained in Chapter 2. Averaging p or SNR over the range of signal and interference levels which occur in a local region is then required, to give meaningful average performance measures [J1, F1]. The region normally selected is one where both signal and interference levels vary in accordance with a Rayleigh density with constant mean value (see Chapter 2).

IV-3 Expressions for Output Noise and Interference

In this and the following section we determine output signal-to-interference (SIR) and signal-to-noise (SNR) ratios for the receivers in Figs. 4-1 and 4-2. In considering Fig. 4-1 we assume PAM signalling with zero isi in the output samples $\{y(kT)\}$. In considering Fig. 4-2, we assume MSK signals represented as in (4-1).

In Fig. 4-1

$$r(t) = \sqrt{2P_s}\, p(t) \cos(2\pi f_c t+\theta) + n_c(t) + i_c(t) \qquad (4\text{-}5)$$

where P_s is the power in r(t) due to the transmitted signal, p(t) is the baseband data signal, θ is a random phase angle uniformly distributed between 0 and 2π, $n_c(t)$ is white Gaussian noise with power spectral density $N_o/2$ and $i_c(t)$ is interference consisting of other data signals and/or voice modulated FM. The input x(t) to the filter is

$$x(t) = \sqrt{P_s}\, p(t)[1+\cos(4\pi f_c t+\theta)] + \sqrt{2}(n_c(t)+i_c(t)) \cos(2\pi f_o t+\theta) \qquad (4\text{-}6)$$

The response y(t) in Fig. 4-1 is

$$y(t) = \int_{-\infty}^{\infty} x(t-\tau)h(\tau)d\tau \tag{4-7a}$$

The filter removes signals outside the data signal baseband, with the result that the response y(kT) due to the data signal alone at k=0 (there is no loss in generality in letting k=0) is

$$y(0) = \sqrt{P_s} \int_{-\infty}^{\infty} p(-\tau)h(\tau)d\tau \tag{4-7b}$$

If y(0) contains no intersymbol interference then

$$y(0) = \sqrt{P_s}\, a_k \int_{-\infty}^{\infty} g(-\tau)h(\tau)d\tau \tag{4-8a}$$

$$= \sqrt{P_s}\, a_k \int_{-\infty}^{\infty} G(f)H(f)df \tag{4-8b}$$

where G(f) and H(f) are Fourier transforms of g(t) and h(t) respectively.

For binary PAM, $a_k = \pm 1$ provided

$$\int_{-\infty}^{\infty} |G_T(f)|^2 df = 1 \tag{4-9a}$$

where

$$G_T(f) = G(f)/\sqrt{T} \tag{4-9b}$$

If the receiver filter is matched to G(f) then

$$G^*(f) = H(f) \tag{4-10a}$$

and

$$y(0) = \sqrt{P_s}\, a_k \int_{-\infty}^{\infty} |G(f)|^2 df \tag{4-10b}$$

When noise and interference are present y(0) contains, in addition to the signal term in (4-8), noise and interference components with powers

$$N = \int_{-\infty}^{\infty} S_n(f)|H(f)|^2 df \tag{4-11}$$

and

$$I_i = \int_{-\infty}^{\infty} S_i(f)|H(f)|^2 df \tag{4-12}$$

where $S_n(f)$ and $S_i(f)$ denote, respectively, the power spectral densities of the noise and interference at the input of the receiver filter.

If the noise is white Gaussian with two-sided power spectral density $S_{n_c}(f) = N_o/2$ then $S_n(f) = N_o/2$ and

$$N = (N_o/2) \int_{-\infty}^{\infty}|H(f)|^2 df \tag{4-13}$$

Consider now one interfering FM voice signal with normalized baseband power spectral density V(f) and separated from the transmitted data signal by Δ Hz as in Fig. 4-3. The spectrum of this interfering signal is (see Section III-1)

$$S(f) = (P_v/2)[V(f-f_c-\Delta) + V(f+f_c+\Delta)] \tag{4-14}$$

where P_v is the power in this signal at the receiver input, and $\int_{-\infty}^{\infty}V(f)df=1$. The power in y(0) due to this interfering signal is

$$I_v = (P_v/4)\int_{-\infty}^{\infty}|H(f)|^2[V(f-\Delta)+V(f+\Delta)]df \tag{4-15a}$$

$$= (P_v/2)\int_{-\infty}^{\infty}V(f-\Delta)|H(f)|^2 df \tag{4-15b}$$

$$= (P_v/2)\int_{-\infty}^{\infty}V(f)|H(f-\Delta)|^2 df \tag{4-15c}$$

where (4-15b) and (4-15c) follow from the fact that V(f) and $|H(f)|^2$ are even functions. If $\Delta=0$ co-channel interference results.

If the interfering signal is a data signal with normalized baseband power spectral density P(f) then the power in y(0) due to this interfering data signal is

$$I_d = (P_d/2)\int_{-\infty}^{\infty}P(f)|H(f-\Delta)|^2 df \tag{4-16}$$

where P_d is the power at the receiver input. For a PAM data signal of the

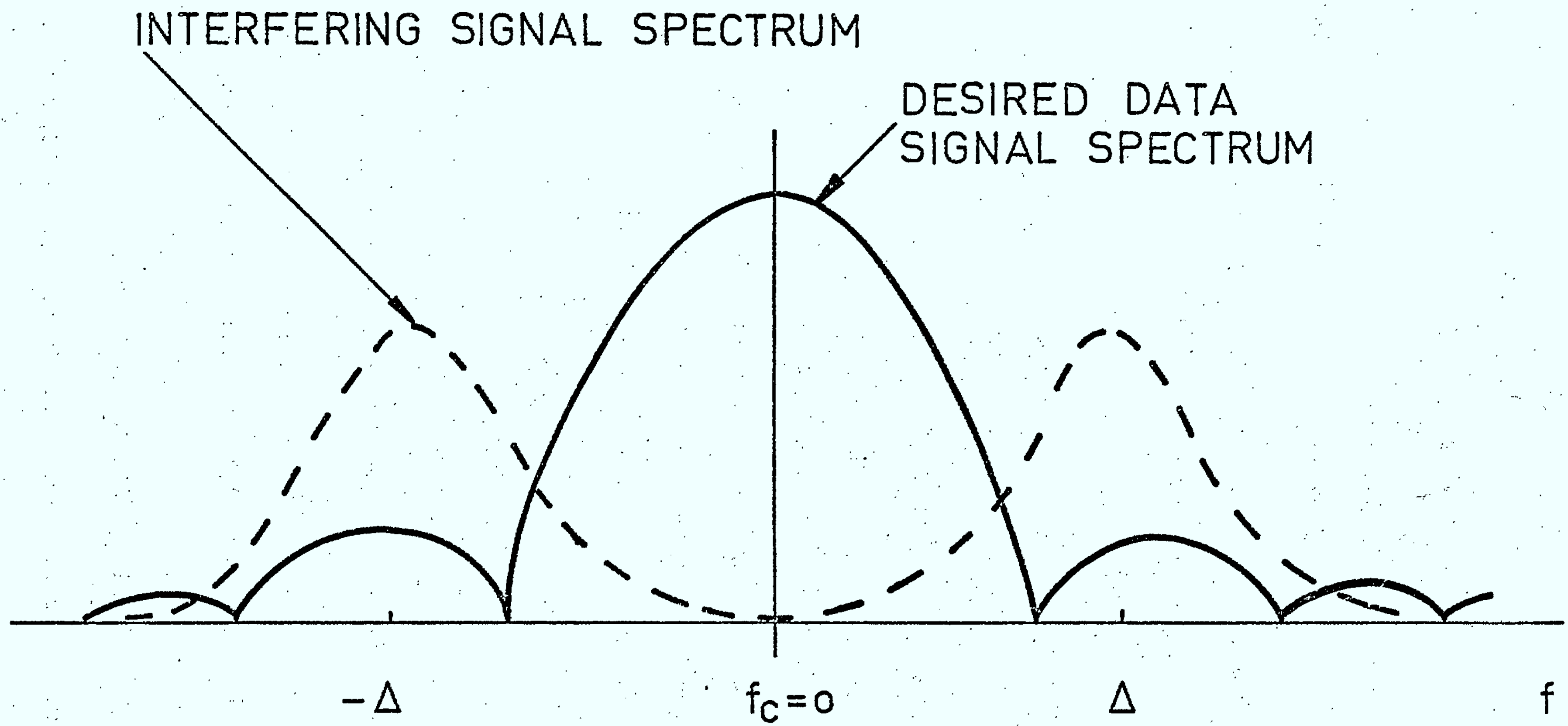

Fig. 4-3 - Spectra of the desired data signal and the interfering signal type described in Section III-2,

$$P(f) = |G_T(f)|^2 \tag{4-17}$$

where $|G_T(f)|^2$ is constrained as in (4-9).

If more than one interfering signal is present then the output noise-plus-interference power in y(0) is the sum of the powers of these component signals; this statement follows from the reasonable and usual assumption that the interfering signals are from different sources and are therefore uncorrelated.

IV-4 Signal-to-Interference and Signal-to-Noise Ratios

We can now calculate SNIR, the ratio of the signal energy in y(kT) in Fig. 4-1 to the total noise power N plus interference power I_v (for voice interference) and I_d (for data interference) in y(t). Thus,

$$SNIR = S/(N+\Sigma I_v+\Sigma I_d) \tag{4-18}$$

where sums Σ are over all voice and data interferers.

$$\begin{aligned} S &= y^2(0) \\ &= P_s[\int_{-\infty}^{\infty} G(f)H(f)\,df]^2 \\ &= P_s T[\int_{-\infty}^{\infty} G_T(f)H(f)\,df]^2 \end{aligned} \tag{4-19}$$

$$N = (N_o/2)\int_{-\infty}^{\infty} |H(f)|^2 df \tag{4-20}$$

$$I_v = (P_v/2)\int_{-\infty}^{\infty} V(f)\,|H(f-\Delta)|^2 df \tag{4-21}$$

$$I_d = (P_d/2)\int_{-\infty}^{\infty} |G_T(f)|^2 H|(f-\Delta)|^2 df \tag{4-22}$$

To evaluate (4-18) we first determine N/S, I_v/S and I_d/S as follows:

$$N/S = (N_o/2P_s)T^{-1}\int_{-\infty}^{\infty} |H(f)|^2 df/[\int_{-\infty}^{\infty} G_T(f)H(f)\,df]^2 \tag{4-23}$$

$$I_v/S = \frac{1}{2}(P_v/P_s)T^{-1}\int_{-\infty}^{\infty} V(f)\,|H(f-\Delta)|^2 df/[\int_{-\infty}^{\infty} G_T(f)H(f)df]^2 \tag{4-24}$$

$$I_d/S = \frac{1}{2}(P_d/P_s)T^{-1}\int_{-\infty}^{\infty} |G_T(f-\Delta)|^2 |H(f)|^2 df/[\int_{-\infty}^{\infty} G_T(f)H(f)df]^2 \tag{4-25}$$

The above equations simplify considerably when $H(f)=G_T^*(f)$ (matched filtering), in which case

$$\int_{-\infty}^{\infty} |H(f)|^2 df = 1 \tag{4-26}$$

$$\int_{-\infty}^{\infty} G_T(f)H(f)df = 1 \tag{4-27}$$

and

$$N/S = (N_o/2P_sT) \tag{4-28}$$

$$I_v/S = (P_v/P_s)\ C_v\ (\Delta T,\ B,\ g) \tag{4-29}$$

$$I_d/S = (P_d/P_s)\ C_d\ (\Delta T,\ g) \tag{4-30}$$

where, with $\lambda=fT$ and $x=\tau/T$

$$C_v(\Delta T,B,g) = (2T)^{-1}\int_{-\infty}^{\infty} V(f)\,|G_T(f-\Delta)|^2 df \tag{4-31a}$$

$$= (1/2T^2)\int_{-\infty}^{\infty} V(\lambda)\,|G_T(\lambda-\Delta T)|^2 d\lambda \tag{4-31b}$$

$$= \frac{1}{T}\int_o^{\infty} v(\tau)\ \rho_g(\tau)\ \cos 2\pi\Delta\tau d\tau \tag{4-31c}$$

$$= \int_o^{\infty} v(x)\rho_g(x)\ \cos 2\pi\Delta Txdx \tag{4-31d}$$

$$C_d(\Delta T,g) = (2T)^{-1}\int_{-\infty}^{\infty} |G_T(f-\Delta)|^2 |G_T(f)|^2 df \tag{4-32a}$$

$$= (1/2T^2)\int_{-\infty}^{\infty} |G_T(\lambda-\Delta T)|^2 |G_T(\lambda)|^2 d\lambda \tag{4-32b}$$

$$= \frac{1}{T}\int_o^{\infty} \rho_g^{\ 2}(\tau)\ \cos 2\pi\Delta\tau d\tau \tag{4-32c}$$

$$= \int_o^{\infty} \rho_g^{\ 2}(x)\ \cos 2\pi\Delta Txdx \tag{4-32d}$$

In (4-31) $v(\tau)$ is the inverse Fourier transform of $V(f)$ as defined in Section III-1 and

$$\rho_g(\tau) = \int_{-\infty}^{\infty} g_T(t)\ g_T(t-\tau)dt \tag{4-33}$$

where $g_T(t)$ is the inverse Fourier transform of $G_T(f)$. In obtaining (4-31c) and (4-32c) we have used

$$\int_{-\infty}^{\infty} A(f)B^*(f-y)\,df = \int_{-\infty}^{\infty} a(t)b(t)e^{j2\pi yt}\,dt \qquad (4\text{-}34)$$

where A(f) and B(f) are the Fourier transforms of a(t) and b(t) respectively, as well as the fact that v(x) and $\rho_g(x)$ are even functions.

For matched filter reception, the following statements are seen to apply:

1. N/S, I_v/S and I_d/S are all proportional to the ratio of the received noise or interference to the received signal power.
2. I_v/S and I_d/S are proportional to a cross-talk term which depends on the normalized channel spacing $\Delta T = \Delta/R$ and the time-normalized basic PAM pulse shape g(t/T).
3. I_v/S also depends on the rms bandwidth B which affects cross-talk term C_v, and on T.

These dependencies also apply when H(f) is a non-matched filter, but the expressions for C_v and C_d are more complex.

These results, obtained for Fig. 4-1, can be used when MSK signals (or zero-isi MSK-type signals) are detected using the system in Fig. 4-2. If the received data signal is represented by (4-1), the power is divided equally between the in-phase and quadrature data streams, which do not interfere with each other during detection. For interfering signals, however, both the in-phase and quadrature data streams do interfere with each of the desired data streams. An equivalent statement is that the power spectral density of the interfering signal is the sum of the (identical) power spectral densities of the in-phase and quadrature data streams, since these are uncorrelated. From these arguments it follows that C_v and C_d for MSK-type signals equal twice the values given by (4-31) and (4-32) provided T is replaced by 2T. As before P_s is the total power in the received signal, including in-phase and quadrature components, P_d the total received power in the interfering signal, and P_v the received power in the interfering FM voice signal. The same conclusion is reached by careful examination of Kalet's [K1] work, with appropriate changes in notation.

IV-5 Co-Channel and Adjacent-Channel Signal-to-Interference Ratios for Binary PAM and MSK Signalling

The output signal-to-noise-plus-interference ratio SNIR in (4-18) depends on T, ΔT, B, G(f) and H(f) although G(f) and H(f) should be selected for zero isi, as explained in Chapter 3.

In this section we determine I_v/S and I_d/S vs channel spacings Δ, including the co-channel interference case $\Delta=0$. We consider PAM signalling with rectangular pulses as well as pulses with raised cosine spectra, MSK signals, and MSK-type signals (SFSK). The receiver uses matched filtering, and in the case of MSK, a filter with an impulse response duration of T sec. is also considered in the receiver configuration of Fig. 4-2; such a filter leads to simplification of the receiver, as explained by Kalet and White [K2].

For raised cosine signalling with $\alpha=0$,

$$|G_T(\lambda)|^2 = \begin{cases} 0 & |\lambda| \geq 1/2 \\ T & |\lambda| < 1/2 \end{cases} \tag{4-35}$$

Thus, $C_d=0$ for $|\Delta T| \geq 1$, and for $|\Delta T| < 1$ (without loss in generality $\Delta>0$)

$$C_d = (1/2T^2) \int_{\Delta T-\frac{1}{2}}^{\frac{1}{2}} T^2 d\lambda$$

$$= (1-\Delta T)/2 \tag{4-36}$$

For raised cosine signalling with $\alpha=1$,

$$|G_T(\lambda)|^2 = \begin{cases} 0 & |\lambda| \geq 1 \\ T[\frac{1+\cos\pi\lambda}{2}] & |\lambda| < 1 \end{cases} \tag{4-37}$$

For $|\Delta T| \geq 2$, $C_d=0$. For $\Delta T < 2$

$$C_d = \frac{1}{8} \int_{\Delta T-1}^{1} (1+ \mathrm{cox}\ \pi\lambda)(1+ \cos \pi(\lambda-\Delta T)d\lambda$$

$$= \frac{1}{8}\ (1- \frac{\Delta T}{2})(2+ \cos \pi\Delta T)+ \frac{3}{16\pi} \sin \pi\Delta T \tag{4-38}$$

For rectangular-pulses (conventional PSK)

$$g_T(t) = \begin{cases} 1/\sqrt{T} & |t| \leq T/2 \\ 0 & |t| > T/2 \end{cases} \tag{4-39}$$

and

$$\rho_g(\tau) = \begin{cases} 1- \frac{|\tau|}{T} & |\tau| \leq T \\ 0 & |\tau| > T \end{cases} \tag{4-40}$$

$$C_d = \int_0^1 (1-x)^2 \cos (2\pi\Delta Tx)dx$$

$$= \frac{1}{2(\pi\Delta T)^2}\ [1- \frac{\sin 2\pi\Delta T}{2\pi\Delta T}] \tag{4-41}$$

For $\Delta T=0$, $C_d=1/3$.

With MSK regarded as staggered quadrature PAM signals,

$$g_{2T}(t) = \begin{cases} \frac{1}{\sqrt{T}} \cos \pi t/2T & |t| < T \\ 0 & |t| > T \end{cases} \tag{4-42}$$

as explained in Section IV-4 and as noted by Kalet [Kl] (our $\rho_g(\tau)$ equals 2T times Kalet's $R_S(\tau)$):

$$\rho_g(\tau) = \begin{cases} \frac{1}{\pi} \sin \frac{\pi|\tau|}{2T} + (1-\frac{|\tau|}{2T}) \cos \frac{\pi|\tau|}{2T} & |\tau| < 2T \\ 0 & |\tau| \geq 2T \end{cases} \quad (4\text{-}43)$$

In this case, $\rho_g(\tau)$ and $|G_{2T}(f)|^2$ are sufficiently complex that numerical integration of (4-32c) is the easiest way to obtain C_d.

Consider now a non-matched MSK receiver filter with impulse response $h(t) = \sqrt{2}\ g_{2T}(2t)$ where $g_{2T}(t)$ is given by (4-42), and $\rho_h(\tau) = \rho_g(2\tau)$ where $\rho_g(2\tau)$ is given by (4-43). Constant C_d is determined from twice the value given by (4-25) and (4-30) or from

$$C_d = \frac{1}{T} \int_0^T \rho_g(\tau)\rho_h(\tau) \cos 2\pi\Delta\tau d\tau / [\int_0^T g_{2T}(-\tau)h(\tau)dt]^2 \quad (4\text{-}44)$$

Numerical integration appears to be the easiest way to calculate C_d in (4-44).

In determining C_v, a value for both T and rms bandwidth B must be specified. We selected both B and T^{-1} to be the largest values consistent with the spectral emission constraints, in order to maximize data rate for digital transmission and output signal-to-noise ratio for voice transmission.

Fig. 4-4 shows C_d vs ΔT for optimum matched filter detection and $\Delta T \lesssim 5$. For $\Delta T \gtrsim 1.1$ conventional PSK is seen to most suseptible to adjacent-channel interference from other similar data signals. PAM signals with raised cosine spectra have zero adjacent-channel interference for $\Delta T \geq 1$ in the case of minimum bandwidth signals and for $\Delta T \geq 2$ for maximum excess bandwidth signals. MSK signals show less adjacent-channel interference than conventional PSK signals for $\Delta T \gtrsim 0.6$, less adjacent-channel

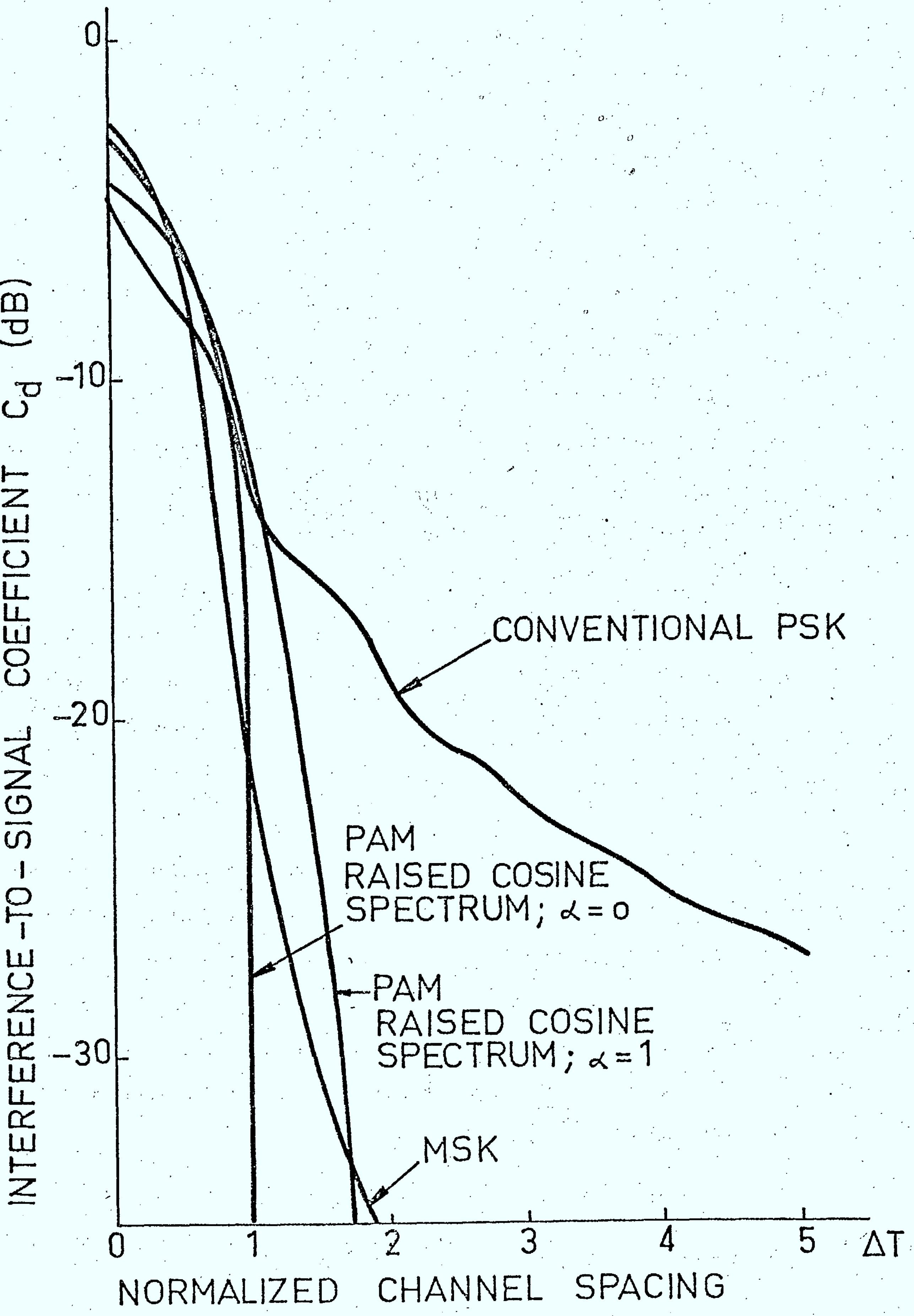

Fig. 4-4 - Crosstalk function C_d vs $\Delta/R = \Delta T$ for various data signals interference than minimum excess bandwidth PAM signals for $0.3 \lesssim \Delta T \lesssim 1.0$, and less adjacent-channel interference than maximum excess bandwidth PAM signals with raised cosine spectra for $0.4 \lesssim \Delta T \lesssim 1.70$. Regarding co-channel interference, differences do not exceed 2.1 dB for these four signalling schemes.

Fig. 4-5 shows C_v for $\Delta T \lesssim 5$. Again, adjacent-channel interference is largest for conventional PSK for $\Delta T \gtrsim 0.7$. For $\Delta T \lesssim 5$, the two PAM signalling schemes and the MSK scheme all show similar behaviour of C_v at the maximum data rates shown. The differences in co-channel interference in Fig. 4-5 are due largely to differences in data rates between the various signalling schemes, since C_v varies as T^{-1}.

Fig. 4-6 shows C_d and C_v vs T for $1 \lesssim \Delta T \lesssim 10$. From (4-41) one sees that C_d decreases as $(\Delta T)^{-2}$, whereas for MSK C_d decreases as $(\Delta T)^{-4}$ [R1, K1]. The rates of decrease of C_v are seen to be comparable, for $\Delta T \lesssim 10$, for the two PAM signalling schemes considered at the data rates shown. C_v for MSK begins to fall off more slowly, for $\Delta T \gtrsim 4$, than does C_v for the PAM schemes.

In comparing C_v and C_d for any of the signalling schemes considered, one sees that FM voice transmission creates more adjacent-channel interference than does another data signal of a type similar to the one being detected. Further, the difference between C_d and C_v for all schemes except PSK appears to increase as ΔT increases. The differences between C_v and C_d can be large; for example Fig. 4-6 shows an 11 dB difference for MSK at $\Delta T=3$, and 22.5 dB difference for $\Delta T=10$.

Fig. 4-7 shows differences in C_v and C_d which result when suboptimum filtering is employed for detection of MSK data. The increase in C_v is

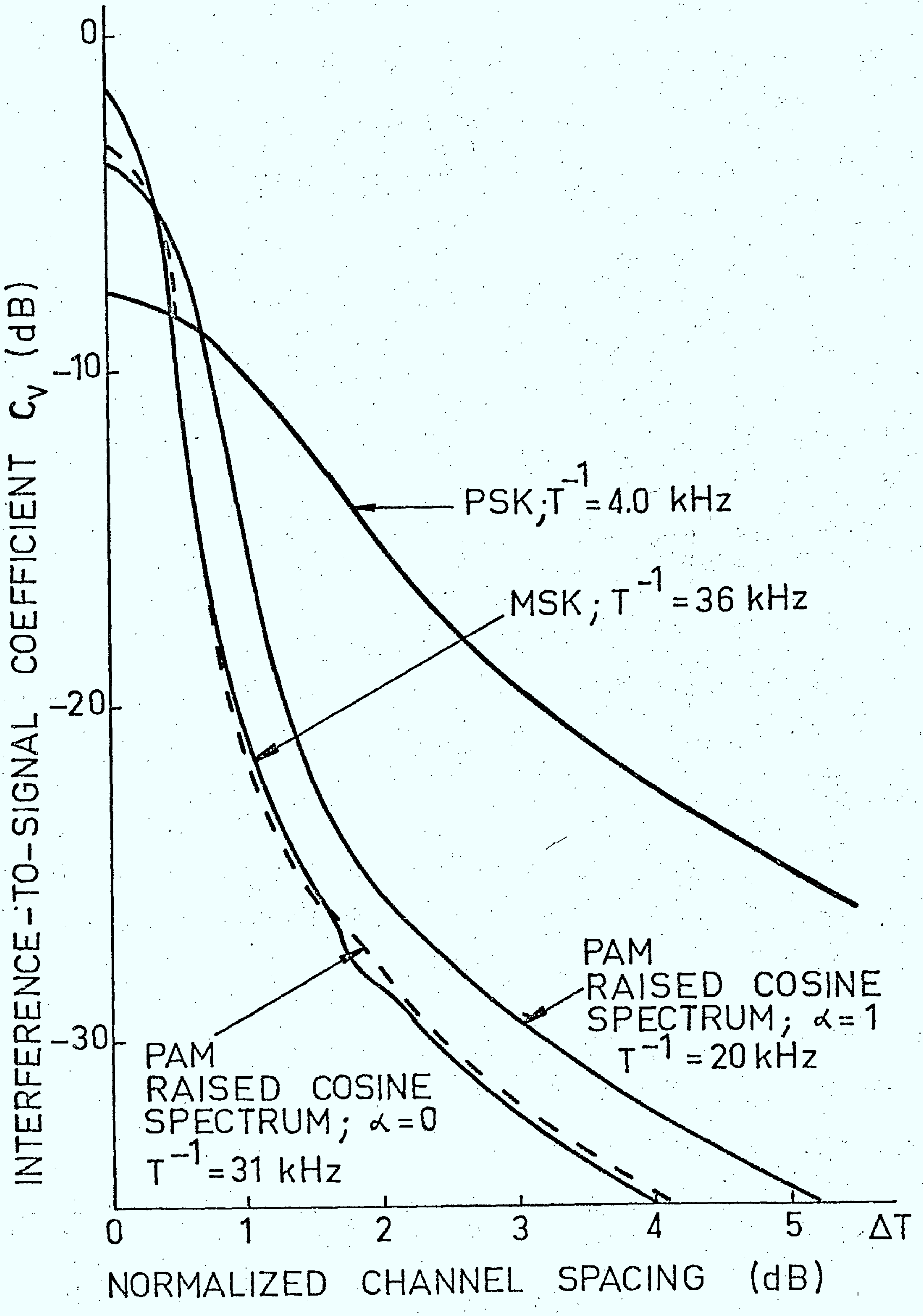

Fig. 4-5 - Crosstalk function C_v vs ΔT for various data receivers.

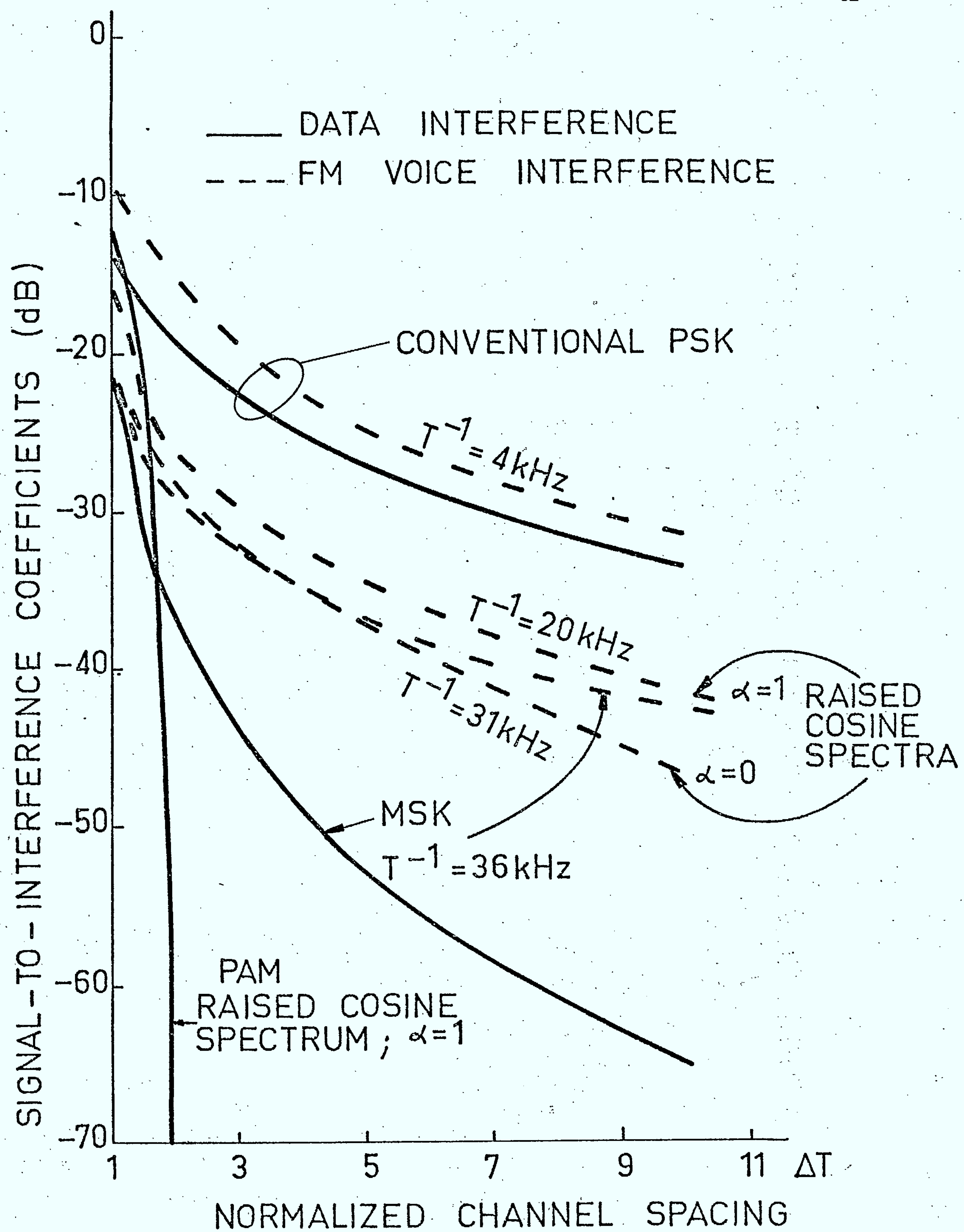

Fig. 4-6 – C_d and C_v vs ΔT for various data receivers

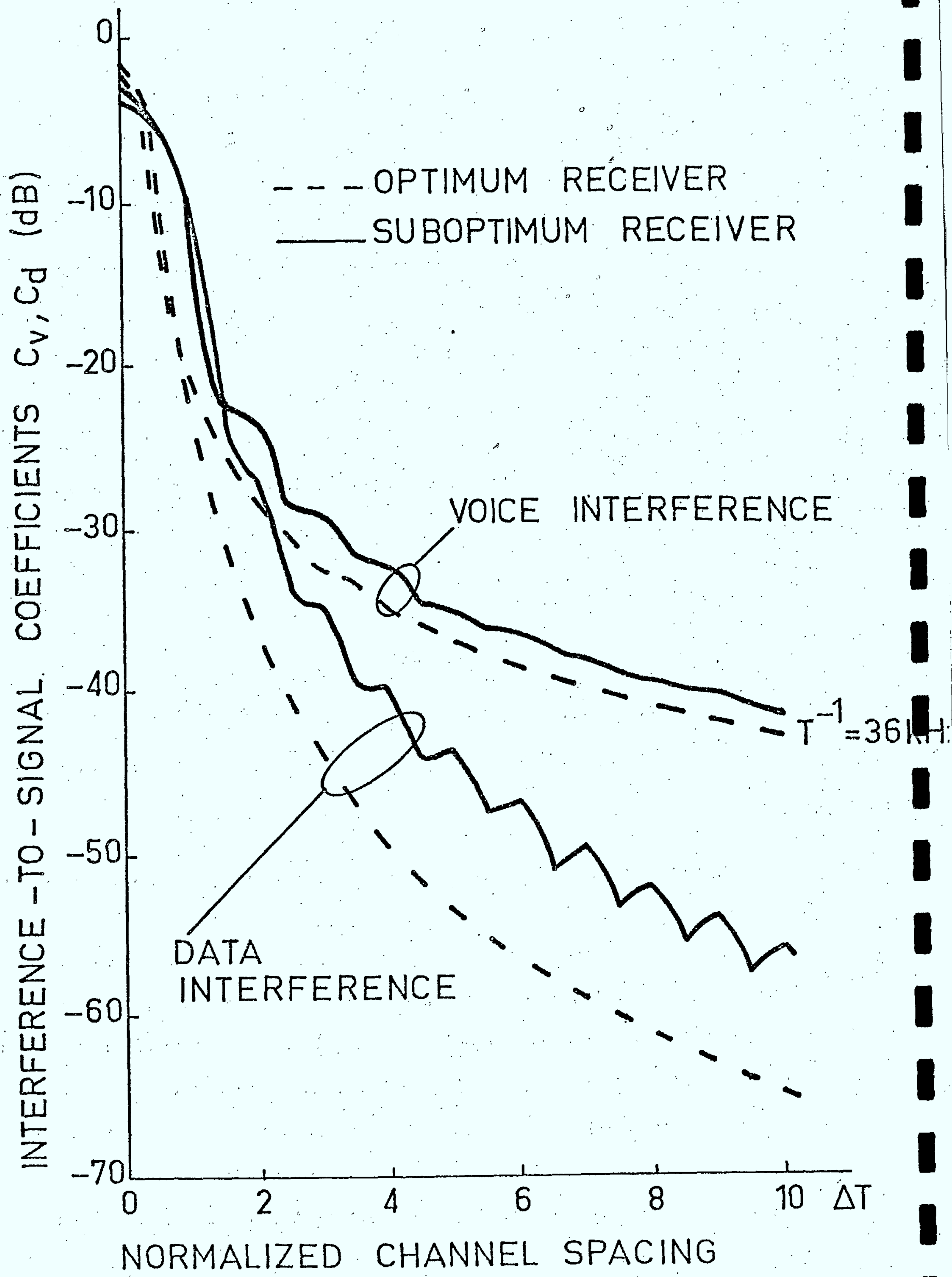

Fig. 4-7 - Comparison of C_d and C_v for optimum and suboptimum MSK receivers $T^{-1} = 36$ Kbs on the order of 2.2 dB, while an increase of approximately 8 dB occurs for C_d. In the case of data interference C_d does not decrease monotonically with ΔT and benefits of up to 2 dB are obtainable by selecting ΔT such that C_d shows a local minimum.

At this point it is natural to enquire whether or not other signalling schemes exist with lower values of C_v or C_d, even if the maximum data rate permitted by the spectral emission constraints is less than the 36 Kbs available for MSK. Sinusoidal FSK as proposed by Amoroso [A1] shows crosstalk behaviour C_d which decreases asymtotically as $(\Delta T)^{-8}$ [R1, K1]. For SFSK, $\phi(t)$ in (4-4) is of the form [A1, K1].

$$\phi(t) = \frac{\pi t}{2T} - U \sin \frac{2\pi t}{T} \tag{4-45}$$

where U = 1/4;

$$g_{2T}(t) = \begin{cases} \dfrac{1}{\sqrt{T}} \cos \left[\dfrac{\pi t}{2T} - U \sin \dfrac{2\pi t}{T}\right] & |t| < T \\ 0 & |t| \geq T \end{cases} \tag{4-46}$$

and,

$$\rho_g(\tau) = \left(1 - \frac{|\tau|}{2T}\right) \cos \frac{\pi|\tau|}{2T} + \frac{1}{T} \int_0^{T-\frac{|\tau|}{2}} \cos \left(\frac{\pi t}{T} - 2U \sin \frac{2\pi t}{T} \cos \frac{\pi|\tau|}{T}\right) dt$$

Numerical calculation of C_d and C_v for SFSK yields the results shown in Figs. 4-8 and 4-9. In calculating C_v for SFSK, the maximum data rate T^{-1} = 32 kHz was estimated using Fig. 3-11, notwithstanding the fact that the first sidelobe at $fT \simeq 2$ ($\simeq$64 kHz) may be one or two dB above the -60 dB spectral emission limit. Fig. 4-8 shows that SFSK is much less subject to adjacent-channel data interference than is MSK for $T \gtrsim 2.5$. Adjacent-channel voice interference coefficients C_v for SFSK and MSK are nearly equal.

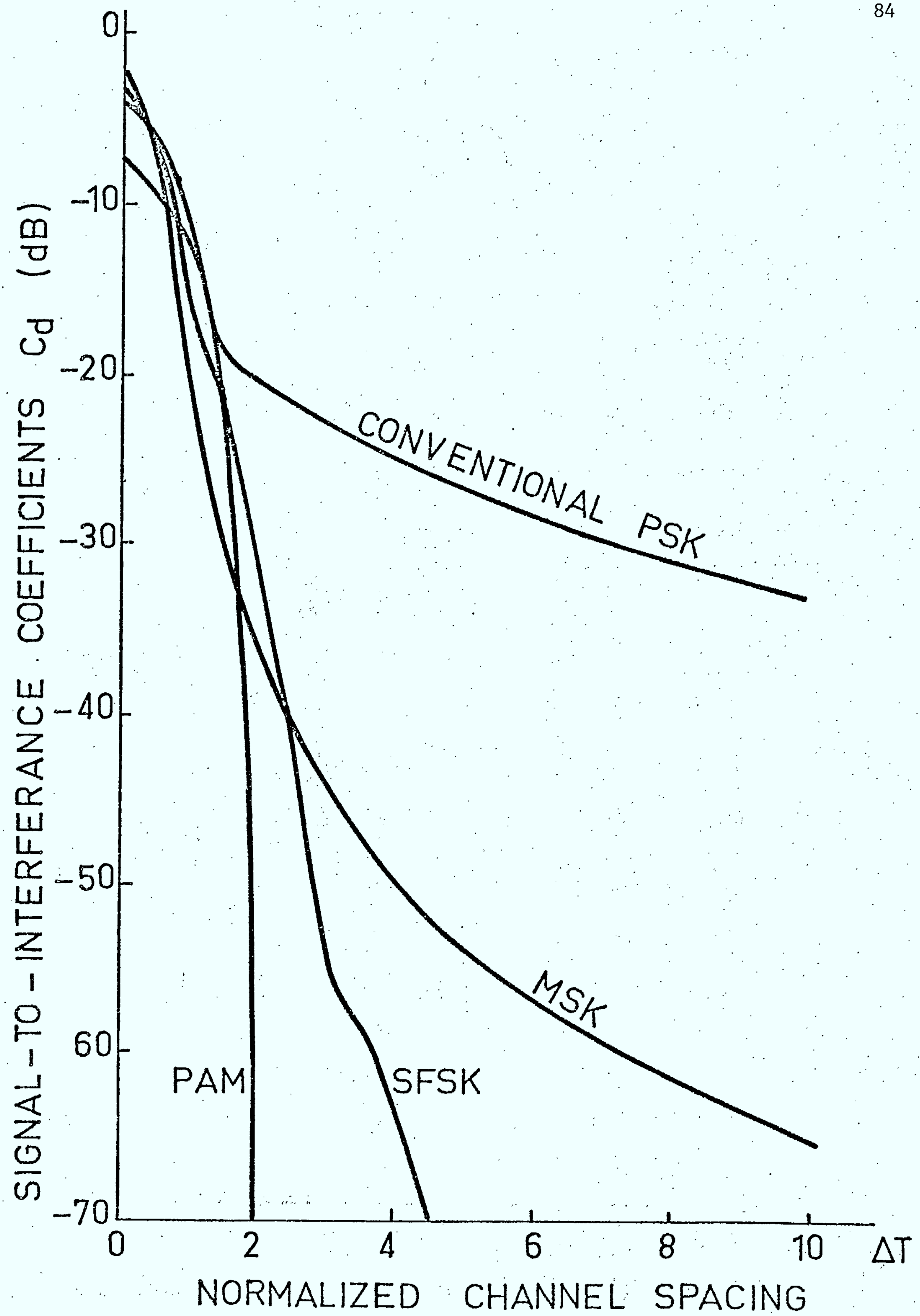

Fig. 4-8 - C_d for MSK , conventional PSK, PAM (raised cosine spectra and maximum excess bandwidth), and SFSK

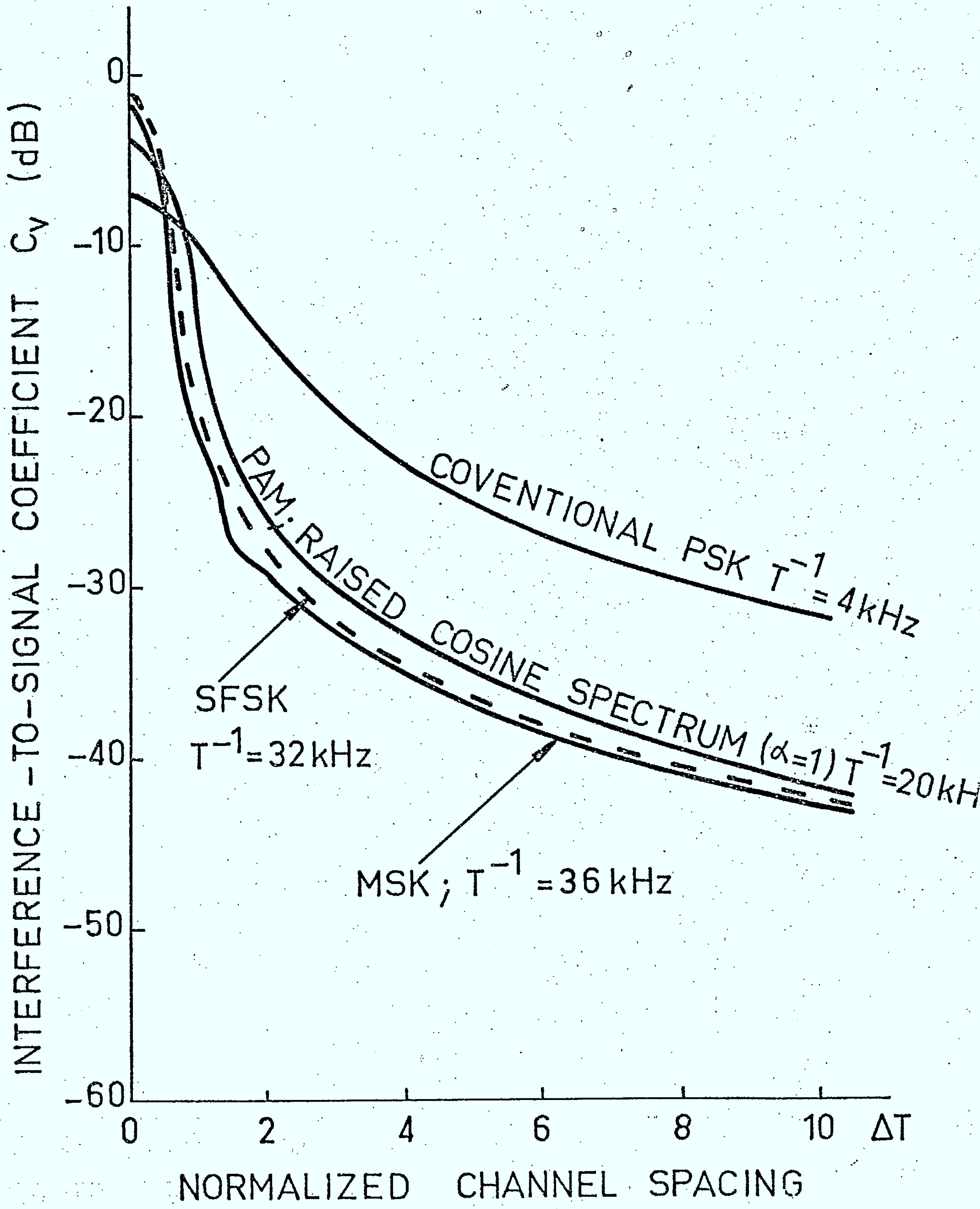

Fig. 4-9 - C_v for MSK, conventional PSK, PAM and SFSK; maximum data rates

Comparisons between C_v for various signalling schemes at the same data rates appear in Fig. 4-10 ($R = T^{-1} = 20$ kbs) and Fig. 4-11 ($R = 4$ Kbs). It is apparent that different signalling schemes operating at the same data rate show similar susceptibilities to FM voice interference. This fact suggests that the relatively slow rate of decrease of C_v for all signalling schemes considered results from a relatively slow roll-off of the FM spectrum, a fact confirmed by Fig. 4-12, which shows this spectrum V(f). From discussions in Rowe [R2] and Section 3-1 it follows that V(f) falls off as $1/f^2$ (6 dB/octave) and Fig. 4-12 confirms this behavior.

Reduction of the data rate for any signalling scheme does reduce C_v; for example reduction of the MSK rate from 20 Kbs to 4 Kbs results in reductions in C_v of 8dB at $\Delta T = 5$. Such reductions are due largely to the factor T^{-1} in the equation defining C_v. The available bandwidth created by such reductions can be used to reduce the bit error rate either by modulation of a bandwidth-expanding FM carrier signal, or by use of the original maximum data rate together with an error correcting code, as explained in Chapter 3.

The fact that V(f), the FM spectrum, decreases relatively slowly as the frequency increases suggests that reduction of rms bandwidth B would reduce C_v. Fig. 4-13 compares C_v for conventional PSK at 4 Kbs and MSK at 36 Kbs. In each case reducing B reduces C_v. In the case of MSK, the reduction is 8 dB at $\Delta T = 5$. Unfortunately, reduction of B decreases the output signal-to-voice ratio for the demodulated voice signal as explained in Chapter 5.

The two alternatives proposed in Section 3-4 for spectral compaction of the transmitted FM voice signal are also available to reduce C_v. Thus one possibility is a prefilter different from the one specified

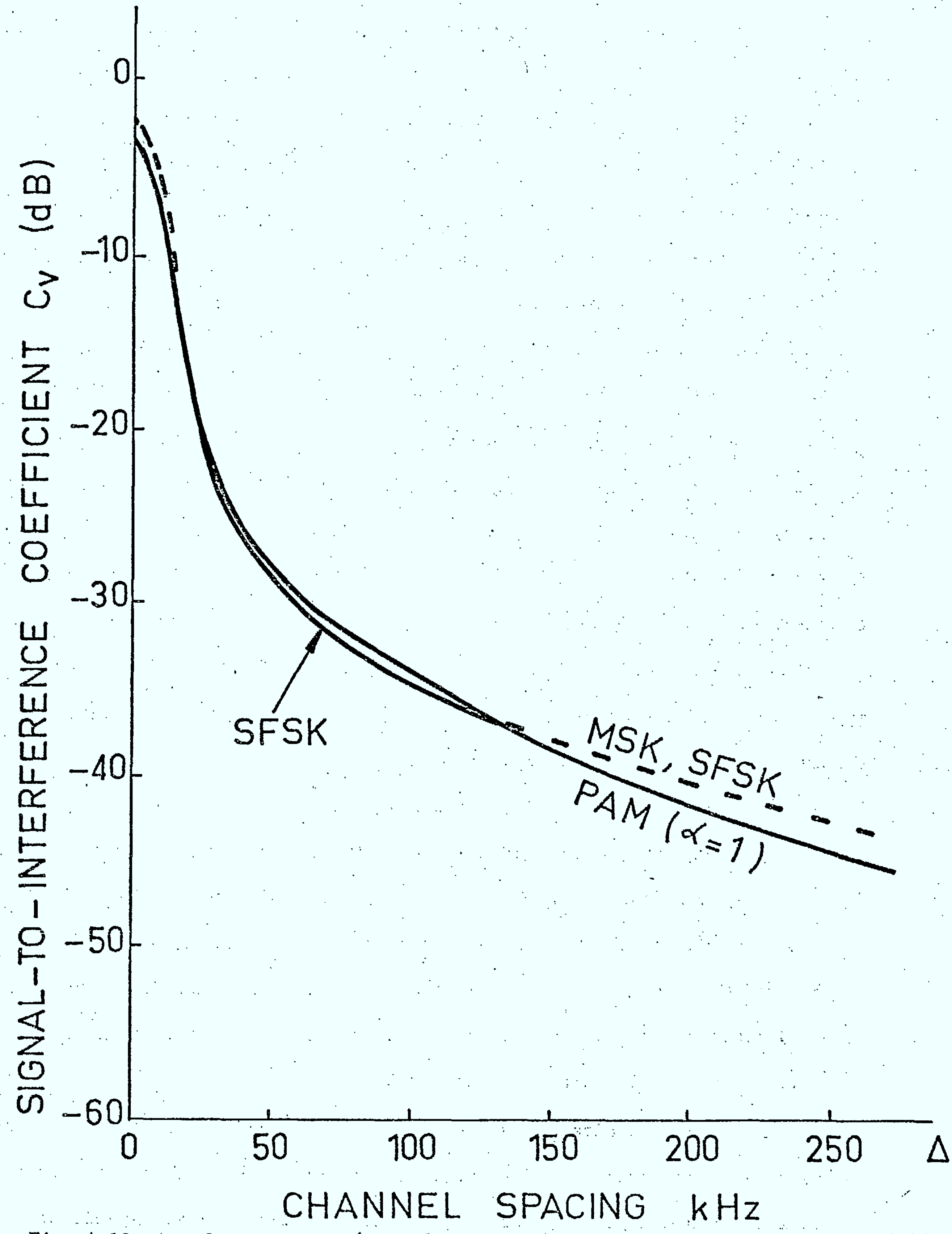

Fig. 4-10 – C_v for MSK, PAM (raised cosine spectra with maximum excess bandwidth) and SFSK; data rate T^{-1} = 20 kHz; B = 2.9 kHz

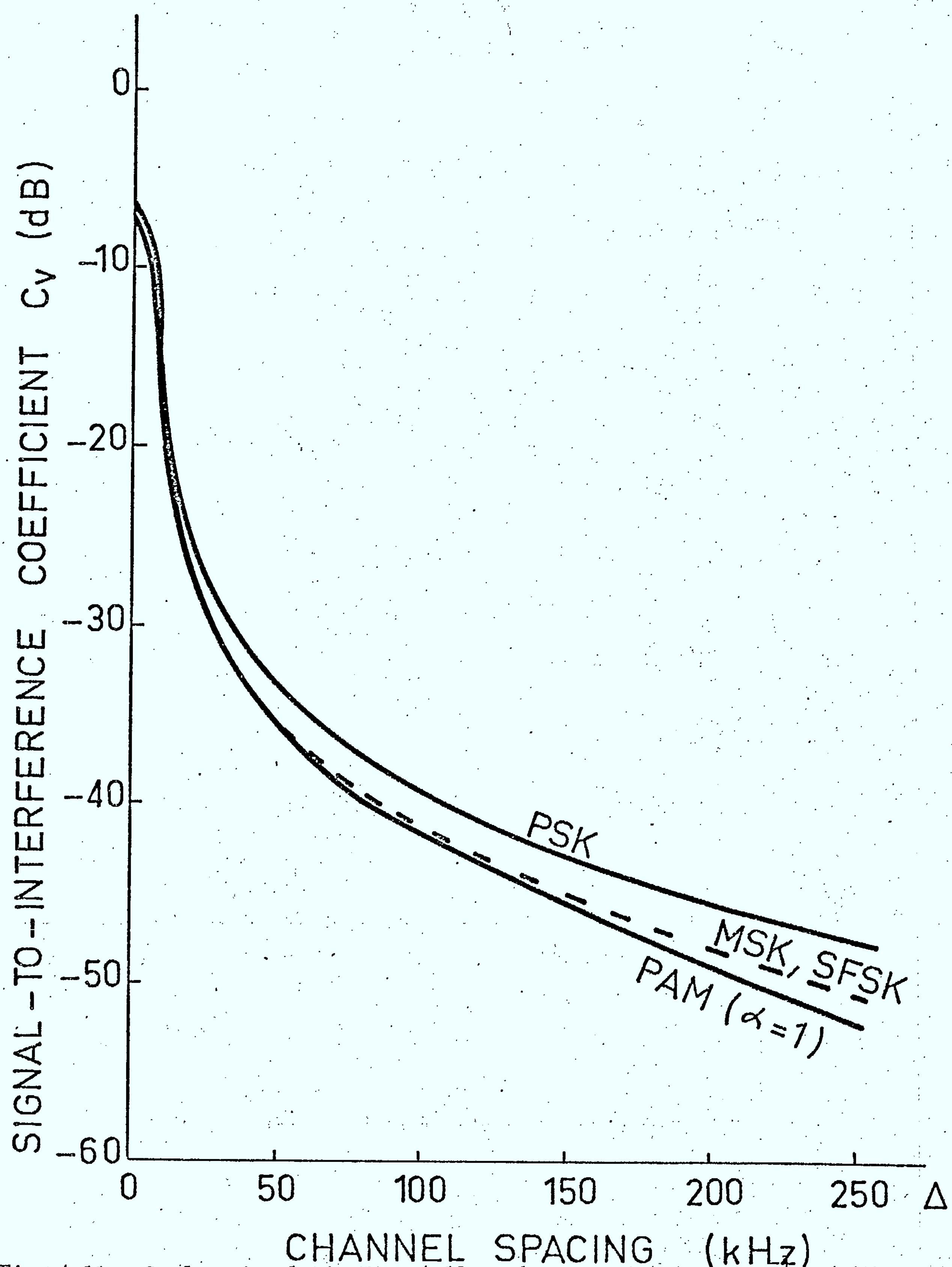

Fig. 4-11 - C_v for signals in Fig. 4-10, and conventional PSK; T^{-1} = 4 kHz

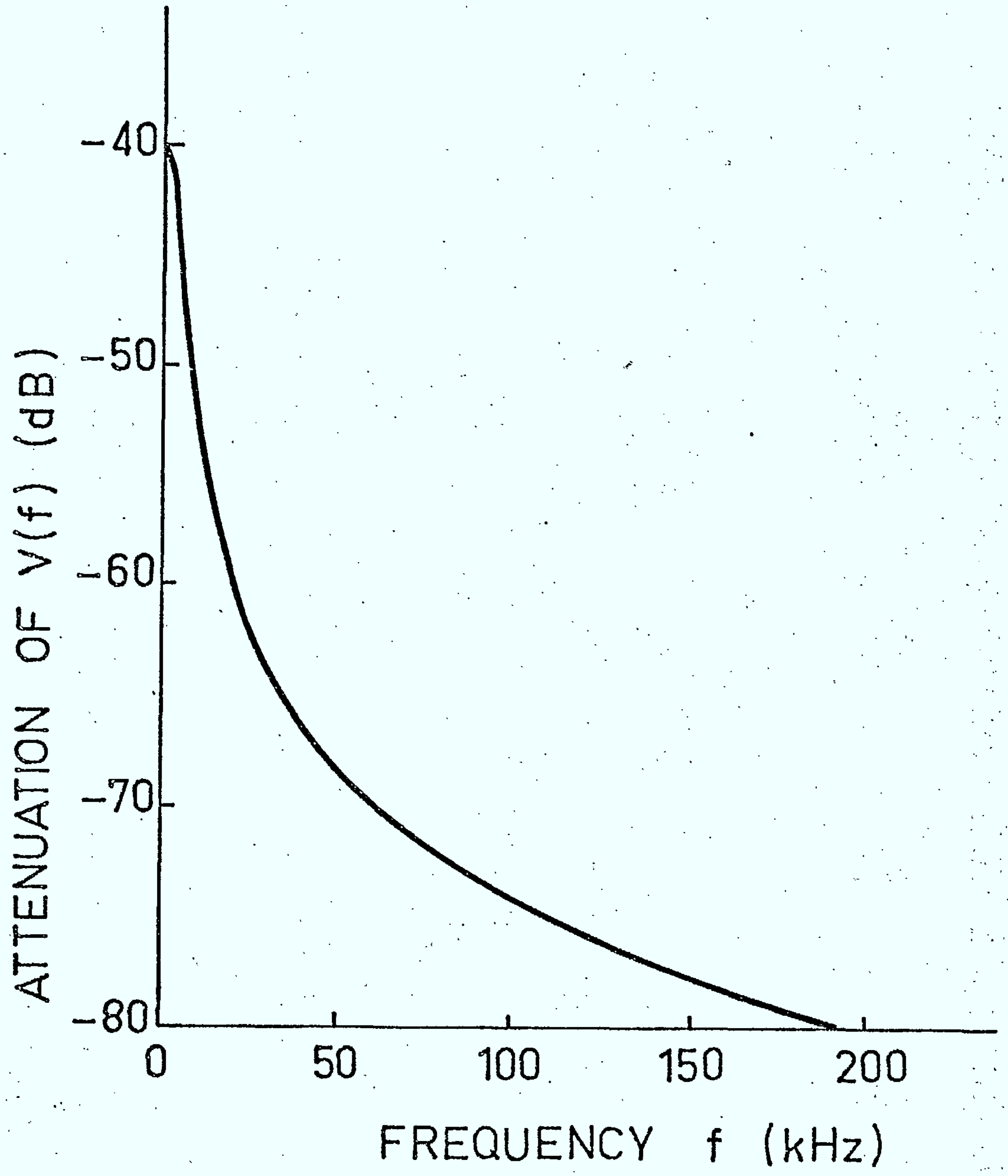

Fig. 4-12 - Spectrum V(f) of transmitted FM voice; B = 2.9 kHz

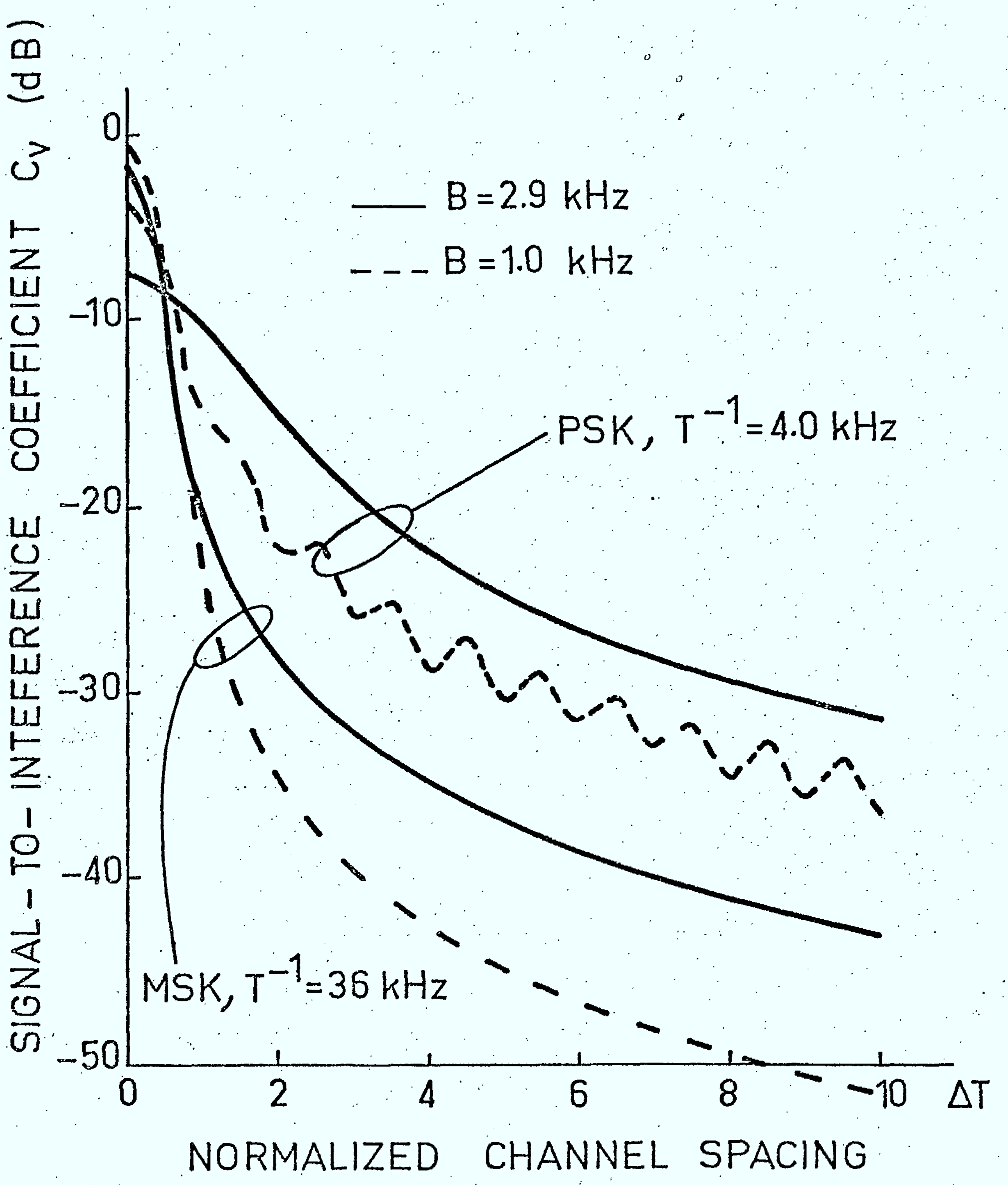

Fig. 4-13 - Comparison of C_v for B = 2.9 kHz and B = 1 kHz for MSK and conventional PSK in Fig. 3-1[B1, V1]. A second alternative is to employ digital transmission, using a source encoding scheme with bit-rates compatible with those permitted by the spectral emission contraints. Adaptive differential encoding permits good quality voice transmission at rates in the 20 Kbs range [H1, C1]. The effects of burst errors on subjective speech quality, as well as a subjective quality comparison with speech transmitted via FM appears to be unavailable.

Use of receivers other than those considered here can reduce adjacent-channel interference for MSK and SFSK signals. Eaves and Wheatly [E1] consider receivers with impulse responses given by (4-46); optimization of U in (4-46) permits a trade-off between increased adjacent-channel interference for large frequency separations and reduced interference at small separations [E1]. Because of the slow rate of decrease of V(f), use of filters different from those considered in this chapter would result in minimal reductions in interference from FM voice transmission. Receivers for reduction of interference are considered in Chapter 5.

IV-6 Bit-Error Probability Considerations

When binary antipodal signalling is used on white Gaussian noise channels, the decoded bit-error probability p is [L1, W1]:

$$p = \operatorname{erfc}\left(\sqrt{2P_S T/No}\right) \tag{4-48}$$

where

$$\operatorname{erfc}(x) = \frac{1}{\sqrt{2\pi}} \int_x^\infty \exp\left(-y^2/2\right) dy \tag{4-49}$$

For Rayleigh fading of the received signal [L1, W1],

$$p = \frac{1}{2+(\bar{P}_S T/No)} \tag{4-50}$$

where $\bar{P}_s$ is the received signal power, averaged over the exponential distribution in (2-2).

When interference as well as noise is present, p can not usually be calculated exactly, as explained earlier. If the noise plus interference is replaced by white Gaussian noise with output power equal to that of the noise plus interference, then (4-48) provides an estimate of the decoded bit-error probability of a single bit, since the signal and interference levels remain relatively constant during the reception of a bit, as explained in Chapter 2. In this case No/2 is replaced by

$$Ne/2T = (No/2T) + (\Sigma C_v P_v + \Sigma C_d P_d) \qquad (4\text{-}51)$$

where the summations are over all FM voice and data interferers.

To determine the average value of p, averaging over both the signal and interference levels is required. Averaging over the signal level yields (4-50), with No/2 replaced by Ne/2; thus

$$p = \overline{[\{(N_o/T) + I\}/\{(2N_o/T) + \bar{P}_s + I\}]} \qquad (4\text{-}52)$$

where the average is with respect to the interference

$$I = \Sigma C_v P_v + \Sigma C_d P_d \qquad (4\text{-}53)$$

Interference I is a sum of Rayleigh random variables; and I is therefore a random variable whose probability density $f_I(r)$ equals the convolution of the densities of the individual interfering signals [W1]. Thus,

$$f_I(r) = \underset{i}{\otimes} \frac{1}{r_i} \exp(-r/r_i)\, U(r) \qquad (4\text{-}54)$$

where

$$r_i = \begin{cases} C_v \bar{P}_v & \text{(voice interference)} \\ C_d \bar{P}_d & \text{(data interference)} \end{cases} \qquad (4\text{-}55)$$

$\bar{P}_v$ and $\bar{P}_d$ denote averaged received interference levels, $\otimes$ denotes convolution, and

$$U(r) = \begin{cases} 0 & r<0 \\ 1 & r>0 \end{cases} \qquad (4\text{-}56)$$

Define

$$F_I(s) = \int_{-\infty}^{\infty} f_I(r) e^{-sr} dr \qquad (4\text{-}57)$$

From (4-54) one obtains,

$$F_I(s) = \prod_i \frac{1}{r_i} \left[\frac{1}{s + (1/r_i)}\right]$$

$$= \left[\prod_i \frac{1}{r_i}\right] \sum_k \frac{R_k}{s+(1/r_k)} \qquad (4\text{-}58a)$$

where

$$R_k = \prod_{j, j \neq k} \frac{1}{(1/r_j) - (1/r_k)} \qquad (4\text{-}58b)$$

Inverse transformation of (4-57) yields,

$$f_I(r) = \left[\prod_i \frac{1}{r_i}\right] \sum_k R_k [\exp(-r/r_k)] U(r) \qquad (4\text{-}58a)$$

$$= \sum_k \frac{1}{r_k} \left[\prod_{i, i \neq k} \left[\frac{1}{1-(r_i/r_k)}\right] \exp(-r/r_k)\right] U(r)$$

(4-58b)

Averaging of p now requires calculation of integrals of the form.

$$\phi(a,b,r_k) = \int_0^{\infty} [(a + r) / (b + r)] \exp(-r/r_k)\, dr \qquad (4\text{-}59)$$

The error probability p can be expressed as follows:

$$p = \left[\prod_i \frac{1}{r_i}\right] \sum_k R_k \phi(a,b,r_k) \qquad (4\text{-}60)$$

where $a = No/T$ and $b = 2a + \bar{P}_s$.

From (4-52), (4-58) and (4-60) it follows that p will be in terms of $No/\bar{P}_s T$, $\bar{P}_s$, $\{C_v\}$, $\{C_d\}$, $\{\bar{P}_d/\bar{P}_s\}$ and $\{\bar{P}_v/\bar{P}_s\}$.

The above development assumes that the receiver remains synchronized to the desired data signal during those periods when the noise plus interference level exceeds the signal level. During such periods the receiver may loose synchronization or become synchronized to an interfering signal whose level temporarily exceeds that of the desired data signal. During loss of synchronization, the decoded bit error probability equals 1/2. Thus, the bit-error probability p_ℓ in the case of synchronization loss is as follows, where p is given by (4-60) and P_L denotes the probability of a synchronization loss:

$$p_\ell = p(1-P_L) + \frac{1}{2} P_L \qquad (4\text{-}61)$$

Various methods exist to determine P_L [L2].

IV-7 Implications for Cellular System Design

The ever increasing demand for electromagnetic spectrum for mobile communications particularly in large metropolitan areas, motivates cellular systems [J1, M1] designed for channel reuse. A typical cellular plan appears in Fig. 4-14. Groups of channels are assigned to cells, in such a way that co-channel interference from reused channels in other cells is below some tolerable limit. Adjacent-channel interference may also be considered in channel assignment schemes. Some authors contend that separate treatment of the two types of interference is unnecessarily restrictive. In particular Mikulski states [M1]:

> "Interference is basically an unwanted form of electromagnetic energy falling within the acceptance bandwidth of a receiver, whether it be caused by a cochannel transmitter or an adjacent channel transmitter. By accepting today's adjacent channel protection levels of about -75 dB while

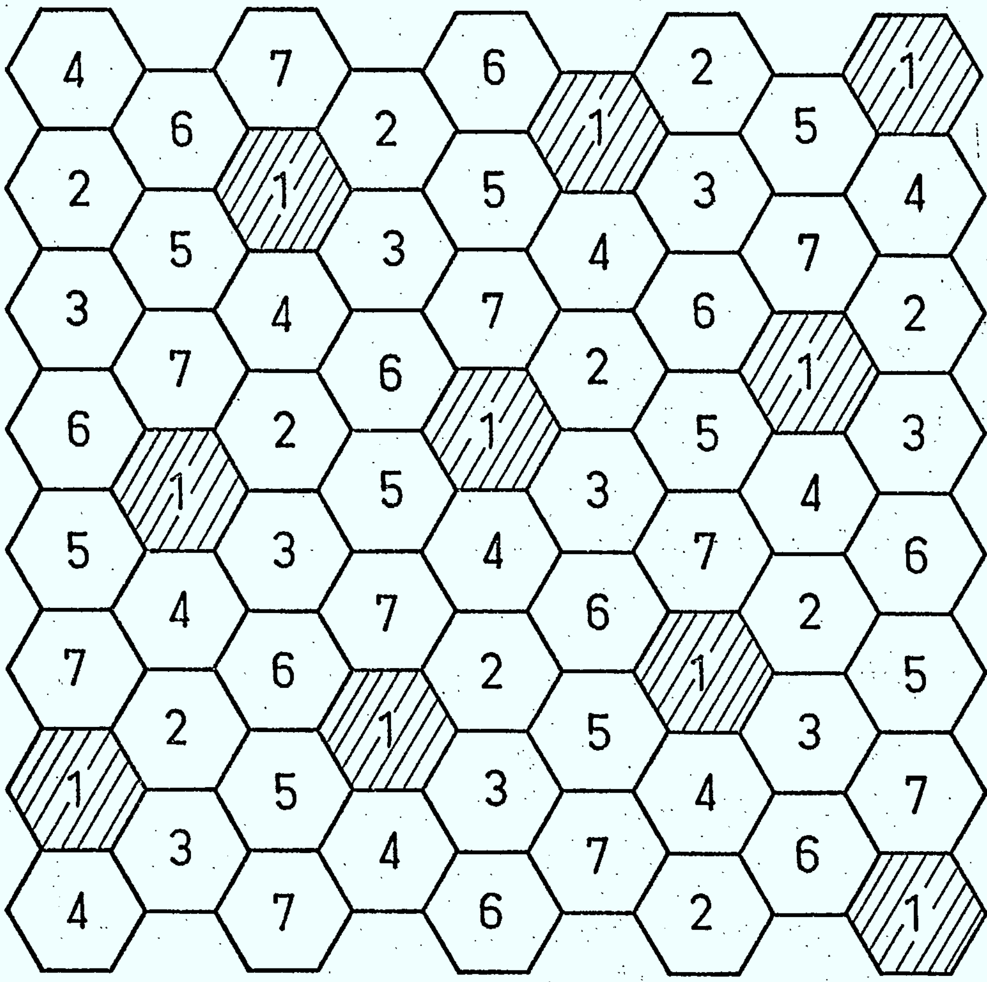

Fig. 4-14 - Cellular System Plan

striving to achieve the effect of about -30 dB protection levels for cochannel interference, system designers have created an unnecessary distinction between the two cases and inhibited the tradeoffs possible between the two cases."

Our results provide for the determination of both cochannel and adjacent-channel interference levels for cellular systems. In Fig. 4-14, for example, cochannel interference power ratios P_v/P_s in (4-29) and P_d/P_s in (4-30) are proportional to $(R/D-R)^n$ where R is the cell radius, D is the distance between the centres of two cells which reuse the same channel, and n is the propogation constant [J1]. Typically $n \simeq 3$ which implies, for equal transmitted power in each cell, that $\bar{P}_v/\bar{P}_s = \bar{P}_d/\bar{P}_s \simeq (1/3)^3 \simeq -14$ dB. This value of -14 dB is added to C_v and C_d in Table 4-1 to determine co-channel interference levels. This same approach can be used to determine adjacent-channel interference effects, and the approach in the previous section can then be used to estimate bit-error probability.

Regarding Table 4-1, cochannel interference coefficients C_v and C_d are within approximately $\pm$ 1.5 dB of -3 dB at maximum data rates, except that C_v for conventional PSK $\simeq$ -7.5dB. Thus, in assigning channels to cells on the basis of cochannel interference considerations alone differentiation between voice and data interference seems unnecessary. However, this same comment clearly does not apply to adjacent-channel interference, which depends not only on whether the voice or data interference is involved, but also on the data signalling format.

For bandwidth-efficient data signalling schemes, interference from FM voice transmission is most degrading. To obviate this diffi-

SIGNALLING SCHEME	BIT RATE (kHz)	CHANNEL SEPARATION (kHz)									
		0		25		50		100		200	
		C_v	C_d	C_v	C_d	C_v	C_d	C_v	C_d	C_v	C_d
PSK (conv.)	4.0	-7.49	-4.77	-27	-29	-33.5	-35	-39.5	-41	-45.5	-47
PAM (α = 1)	4.0	-7.17	-4.26	-29	0	-36	0	-42	0	-49	0
MSK	4.0	-6.97	-2.32	-29.5	-56.5	-35.5	-68.5	-41.5	-80.5	-47.5	-92.5
SFSK	4.0	-6.96	-2.48	-29	-83	-35.5	-107	-41.5	-131	-47.5	-155
PAM (α = 1)	20.0	-3.66	-4.26	-20	-18	-28	0	-34.5	0	-41.5	0
MSK	20.0	-2.17	-2.32	-20	-27	-28	-40.5	-34.5	-53	-40.5	-65
SFSK	20.0	-2.18	-2.48	-21	-19	-29	-41	-34.5	-77	-40	-101
MSK	36.0	-1.52	-2.32	-16	-11	-26	-28	-32	-43	-38	-55
SFSK	32.0	-1.62	-2.48	-18	-13	-24	-25	-31	-57	-37.5	-84

TABLE 4-1 - Comparison of co-channel and adjacent-channel interference levels (dB) for various digital signalling schemes.

culty, channels assigned for voice transmission could be placed in a frequency band separate from channels assigned for transmission of data. The disadvantage of this scheme is the increased delay in queuing for use of channels [J1, K5]; if all channels of one type are in use, those of the other type are unavailable to handle the traffic overload. In non-real-time transmission of data, erroneously received data would normally be retransmitted using an ARQ retransmission protocol [F2,B2], in which case increased levels of adjacent-channel interference from FM voice would as a primary effect reduce the information throughout.

Different users may employ different digital transmission formats, in which case the interference would again be calculated using (4-25) and (4-30). The rate of decrease of coefficient C_d vs frequency separation Δ would be governed by the data signalling format whose spectrum decreased most slowly [K1, R1]. Thus, interference of MSK by conventional PSK would be almost as harmful as interference by FM voice. However, data users would be motivated to employ formats with fast spectral roll-offs to achieve both high data rates and reduced vulnerability to interference of any type. The result would be a tendency towards bandwidth-efficient PAM or MSK-type signalling.

The fact that digital transmission of voice offers the possibility of less interference than conventional analog schemes creates interesting spectrum management alternatives, including the provision of particular incentives for suppliers and/or users of digital voice transmission facilities.

Finally, it must be noted that the results obtained here, together with the stated conclusions, are based on the speech model developed in Section 3-1. Implications of using alternative speech models are considered in Chapter 5.

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V. SPECTRAL EMISSION CONSTRAINTS, INTERFERENCE TO FM VOICE TRANSMISSION & RECEIVERS FOR INTERFERENCE REDUCTION

V-1 Introduction

The maximum data rates obtained in Chapter 3 are determined by the spectral emission constraints of Fig. 2-2. The maximum rms bandwidth B of the transmitted FM voice signal is determined by these emission constraints and by the spectral density in Section 3-1 for the speech signal. The FM spectra and digital data rates determine the interference coefficients C_v determined in Chapter 4.

In this chapter the implications of using different emission constraints as well as different spectral models for speech are considered. In particular, Communications Canada's current emission constraints (see Appendix) are examined.

Also considered are the effects of co-channel and adjacent-channel interference on FM voice transmission, as well as receivers for reduction of interference.

V-2 Spectral Emission Constraint Considerations

The following paragraphs, denoted by an asterisk on the appended Radio Standards Specifications, are relevant to the present discussion: 1.1, 5.1.3, 5.7, 5.8, 5.12.3, 7.2.1, 7.2.2, 7.2.3.

One sees from these sections that the maximum frequency deviation of $\pm$ 5 kHz provides the actual limitation on transmitted signal spectra. Adjacent-channel emission measurements (Section 7) are based on the amount of energy captured by a 16 kHz bandwidth filter, centered $\pm$ 25 kHz from the carrier frequency. The 16 dB signal level referred to in Section 7.2 would result in frequency modulation by a sinsusoidal signal subject to some amplitude limiting, following pre-emphasis in accordance with the characteristic in Fig. 3-1. The amplitude limiting would produce a signal resembling a square wave.

Difficulties arise in determining the transmitted spectrum V(f) which results when a speech signal FM modulates a carrier signal, under the $\pm$ 5 kHz maximum frequency deviation limitation. Since the instantaneous Hertzian frequency of the modulating signal $m(t)$ in (3-1) is $m(t)/2\pi$, this peak deviation constraint amplitude-limits the modulating signal which is no longer Gaussian. (Gaussian signals have unbounded peak amplitude). In practise, the speech signal's energy level would normally be adjusted to ensure that the probability of the signal amplitude reaching this clipping level would be small, in which case the spectrum V(f) would not be altered much by amplitude limiting. (However, companding the speech signal <u>would</u> alter V(f) as explained in Section V-3). If one set the clipping level to 2.5B where B is the rms bandwidth of V(f), the probability of $m(t)/2\pi$ reaching the $\pm$ 5 kHz maximum level is approximately 3%, since B in (3-26) equals the standard deviation of the pre-emphasized Gaussian speech signal $m(t)/2\pi$. At this clipping level, B = 2.0 kHz. For $|m(t)/2\pi| < 1.7B$, the probability of $m(t)/2\pi$ reaching the peak level is approximately 94%, and B = 2.9 kHz, the value used to obtain the results in Chapter 4.

For FSK data transmission the baseband data waveform should be de-emphasized in order to cancel the pre-emphasis which occurs prior to frequency modulation. The peak frequency deviation then limits the data rate, which determines the transmission spectrum. For example, binary MSK signals have carrier-frequency offsets of $\pm$ R/4 kHz where R is the data rate. A maximum frequency deviation of $\pm$ 5 kHz, therefore, implies a maximum data rate of 20 kHz, rather than 36 kHz as determined in Chapter 4. At this data rate, both C_v and C_d would be reduced from their values for 36 kHz, as shown in Table 3-1. What does not change is the fact that voice interference levels are considerably higher than data interference levels, because of the difference in spectral roll-off rates for FM voice and MSK signals.

For bandwidth-efficient PSK and PAM signalling the maximum frequency deviation cannot be related directly to the data rate, since it is not the instantaneous frequency which is modulated. Conventional binary PAM, (which is bandwidth-inefficient) can be generated by differentiating and then de-emphasizing the baseband waveform prior to pre-emphasis and frequency modulation of the carrier. Since the instantaneous frequency is the rate-of-change of the instantaneous phase, and for conventional PSK phase changes are discontinuous across phase boundaries, the instantaneous frequency is infinite at these points, and zero during the rest of the time when the phase is constant. To comply with the $\pm$ 5 kHz maximum deviation, either the phase-changes would have to be smoothed, or the maximum deviation criterion would have to be modified to become an average over the bit duration.

For bandwidth-efficient PAM or PSK, spectral emission constraints based on peak frequency deviations are not really suitable. Standards based on transmitted spectra, such as those used in Chapters 3 and 4 would be suitable for PAM and PSK as well as for other types of digital modulation and FM voice transmission. Such standards would be more difficult to administer, since emission spectra would have to be measured for signals actually transmitted, rather than for a standard test signal. However, emission levels would be placed under tighter control.

V-3 Alternative Models for Speech Spectra

Results obtained for interference coefficients C_v in Chapter 4 are based on the speech spectrum model in Section 3-1. The pre-emphasized speech spectrum could differ considerably from the one described in Chapter 3, for various reasons, including those detailed below:

1. The ±5 kHz maximum frequency deviation in the appended Radio Standards limits the speech signal prior to modulation. The result is to attenuate, in a non-linear way, the vowel and vowel-like sounds relative to the fricative and stop consonants; these consonants have lower amplitudes and a greater fraction of their energy at high frequencies than do the vowels [F1]. The result of such limiting would be to emphasize the high frequencies relative to the low ones.
2. Voice signals may undergo companding (compression) prior to modulation [F1, S1]. Fig. 5-1 shows a typical compressor characteristic which would have the same general effect on speech as would the limiting operation described in the above paragraph. At the receiver the speech signal would be processed by an expander, whose characteristic would be the inverse of the compressor characteristic.
3. The pre-emphasized and possibly compressed speech signal might be lowpass filtered by a relatively sharp cutoff filter at $W \approx 3$ kHz. The effect of such filtering is to strongly attenuate modulating signal frequencies above frequency W, which attenuation would tend to reduce the high frequency content of the transmitted FM spectrum V(f).

In order to assess the effects of lowpass filtering and companding, we consider phase modulation by baseband analog signals with the following power spectral density, which has been proposed [J1, P1, B1] as being acceptable as a speech spectrum model for some applications:

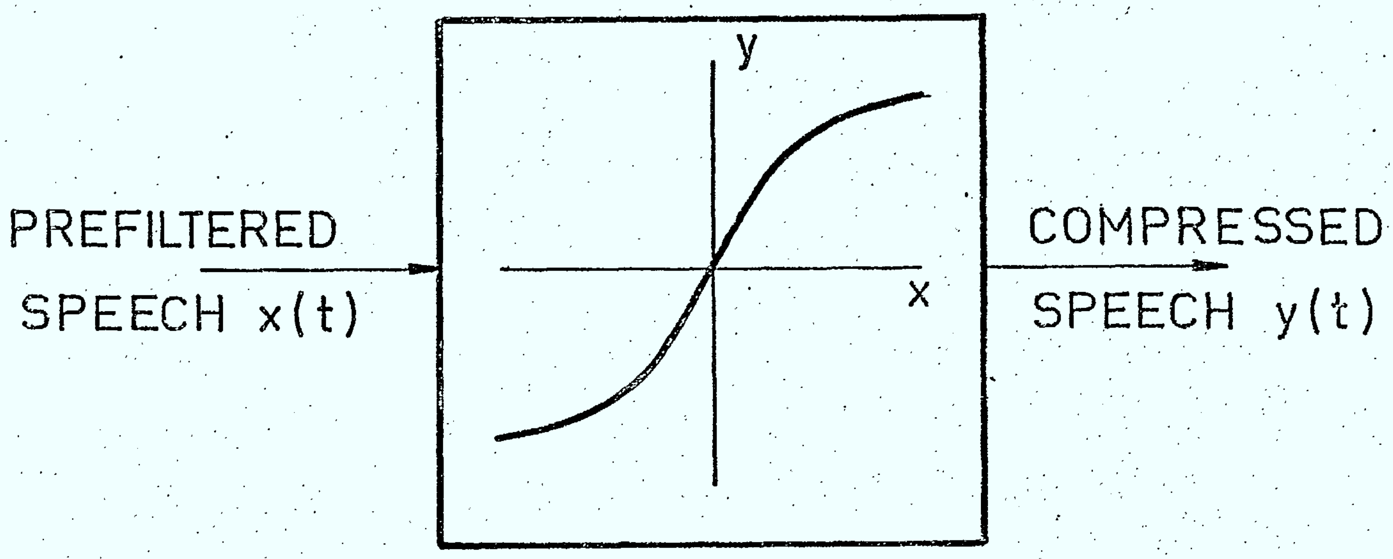

Fig. 5-1 Speech compression characteristic.

$$M(f) = \begin{cases} K^2 & |f| \leq W \\ 0 & |f| > W \end{cases} \qquad (5\text{-}1)$$

Since phase modulation by m(t) is equivalent to frequency modulation by dm/dt, the rms bandwidth B of the transmitted PM signal is, from (3-25),

$$B = [\int_{-W}^{W} f^2 \, M(f) \, df]^{\frac{1}{2}}$$

$$= KW\sqrt{2W/3} \qquad (5\text{-}2)$$

Solution of (5-2) for K in terms of B yields

$$K = [\sqrt{2W/3}]^{-1} (B/W) \qquad (5\text{-}3)$$

The autocorrelation function $\rho(\tau)$ of the speech process with spectrum M(f) in (5-1) is

$$\rho(\tau) = 3(B/W)^2 \sin 2\pi W\tau / 2\pi W\tau \qquad (5\text{-}4)$$

The autocorrelation function $v(\tau)$ of the transmitted PM signal is

$$v(\tau) = \exp [\rho(\tau) - \rho(0)] \qquad (5\text{-}5)$$

and the "baseband" power spectral density V(f) of the transmitted signal is given by [R1]:

$$V(f) = 2 \int_0^{\infty} v(\tau) \cos 2\pi f\tau d\tau \qquad (5\text{-}6)$$

Fig. 5-2 show $\rho(\tau)$ vs. τ. Fig. 5-3 shows $v(\tau)$ vs. τ for B = 2.9, 5.0 and 10.0 where B is in kHz.

Fig. 5-4 shows V(f) as obtained from (5-6) for various values of B. For comparison purposes V(f) as determined for the speech spectrum in Chapter 3 is also shown. It is seen that use of the emission constraints to determine the maximum value of B yields B = 7.0 kHz for the rectangular spectrum, as compared with B = 2.9 kHz for the spectrum in Chapter 3. The reason

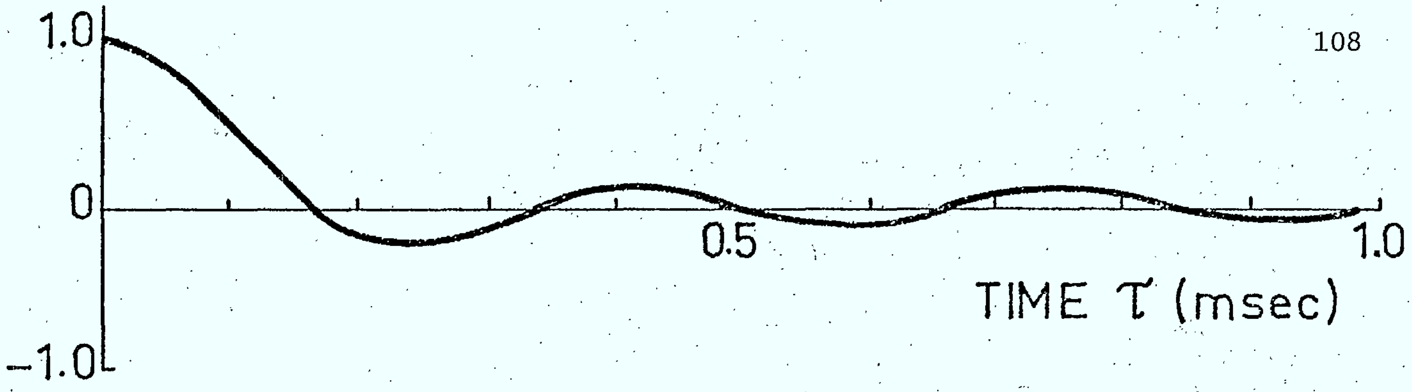

Fig. 5-2 Autocorrelation function $\rho(\tau)$ for rectangular speech spectrum model.

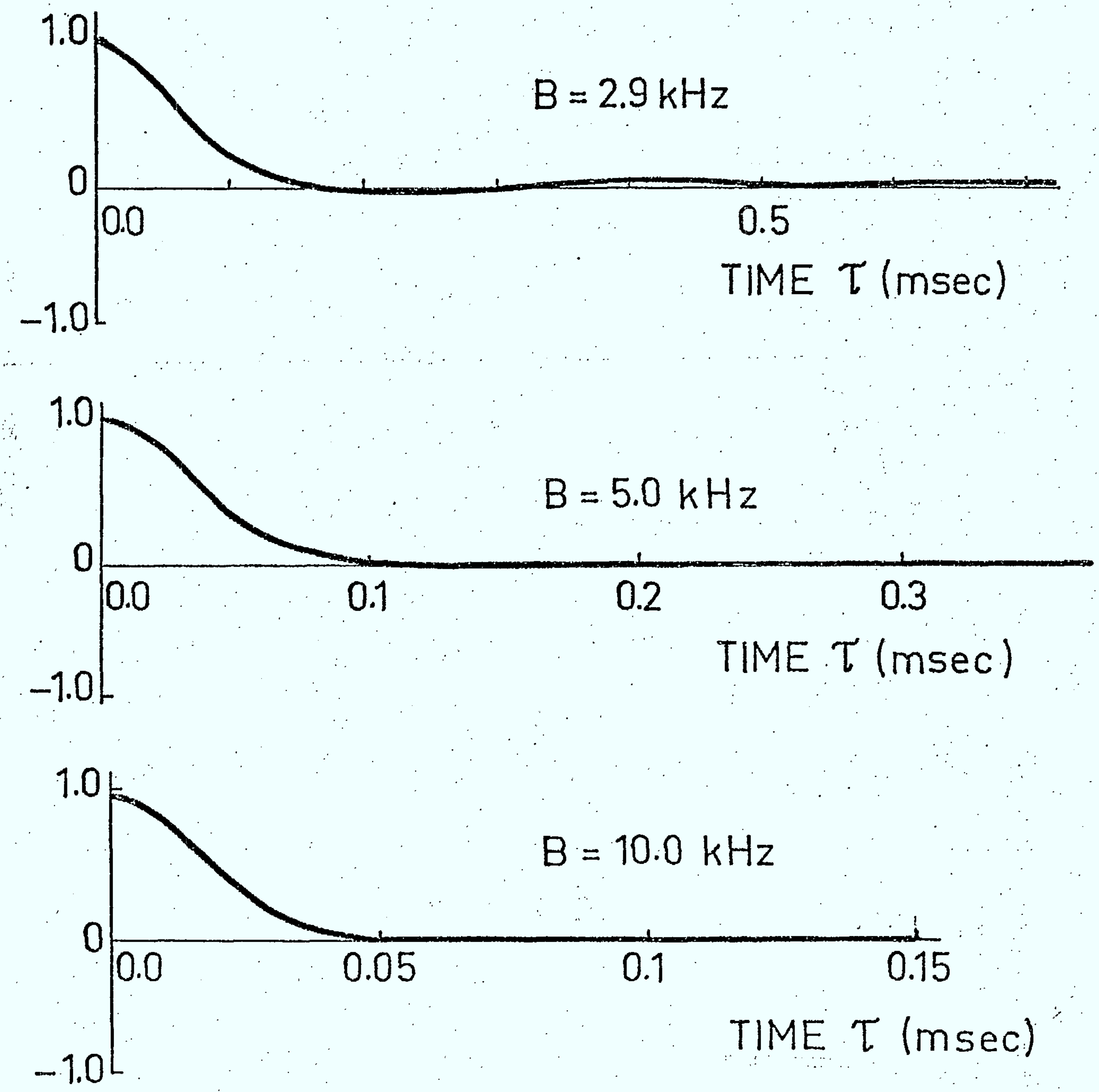

Fig. 5-3 Autocorrelation function $v(\tau)-v(\infty)$ for FM speech modulated signal.

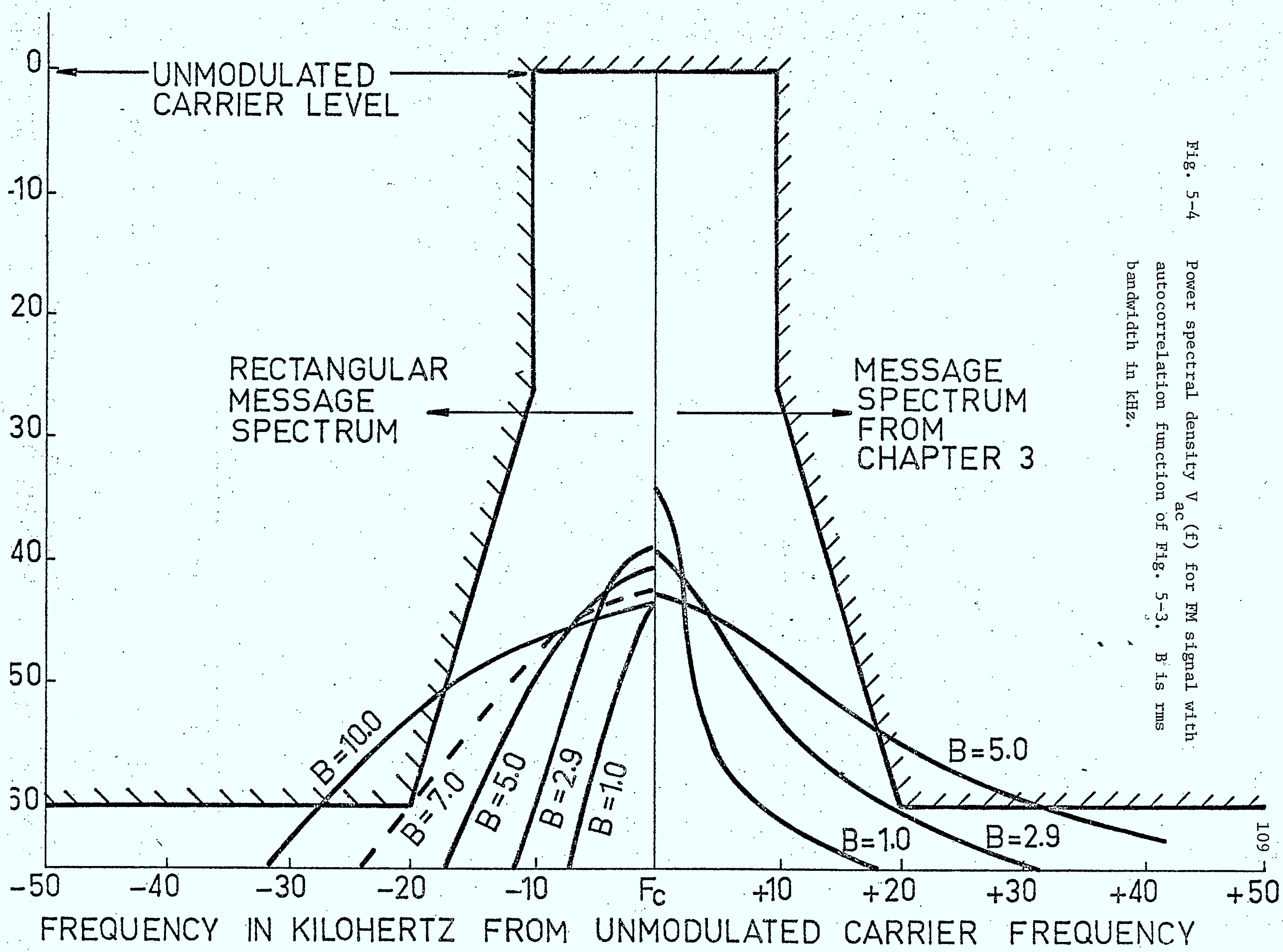

Fig. 5-4 Power spectral density $V_{ac}(f)$ for FM signal with autocorrelation function of Fig. 5-3. B is rms bandwidth in kHz.

for the difference in these maximum values is attributable in part to the constant factor which multiplies $\rho(\tau)-v(0)$ in the equations defining $v(\tau)$ and $V(f)$. For the rectangular spectrum, this constant factor equals $3(B/W)^2$ where $W = 3$ kHz. For the spectrum in Chapter 3, the constant equals $(2.5\ B)^2=6.25\ B^2$. On this basis one would expect a larger permissible value for B for the rectangular spectrum, in spite of the frequency modulation by a signal with a spectrum whose energy increases as f^2 to the cutoff frequency W.

Fig. 5-5 shows the interference coefficient C_v for a binary MSK data signal at bit rate $T^{-1} = 20$ kHz, for $B = 2.9$ and $B = 5.0$ kHz. Also shown for purposes of comparison is C_v from Fig. 4-10 for the same MSK data signal; this value of C_v is based on the speech spectrum described in Section 3-1 for $B = 2.9$. Table 5-1 compares C_v for these three cases with the data interference coefficient C_d, at specific channel spacings.

From Fig. 5-4 and Table 5-1 one sees that the rectangular speech spectrum model implies a much lower value of C_v than does the speech model in Chapter 3. It is clear that a good model of the speech process is required if the effects of adjacent-channel voice interference on data signals are to be calculated accurately. To obtain a good speech spectrum model, it is necessary to have reasonably detailed knowledge of any pre-modulation speech processing operations, including those inherent in the microphone, compandors and filters.

A Gaussian speech signal with $B = 7.0$ kHz would exceed the level ± 5 kHz with a relatively high probability (in excess of 10%); the standard deviation of dm/dt equals B, which actually exceeds 5 kHz in this case. For $B = 7.0$ kHz the Gaussian model for speech becomes sufficiently inaccurate that the calculated transmitted spectrum $V(f)$, which uses the Gaussian assump-

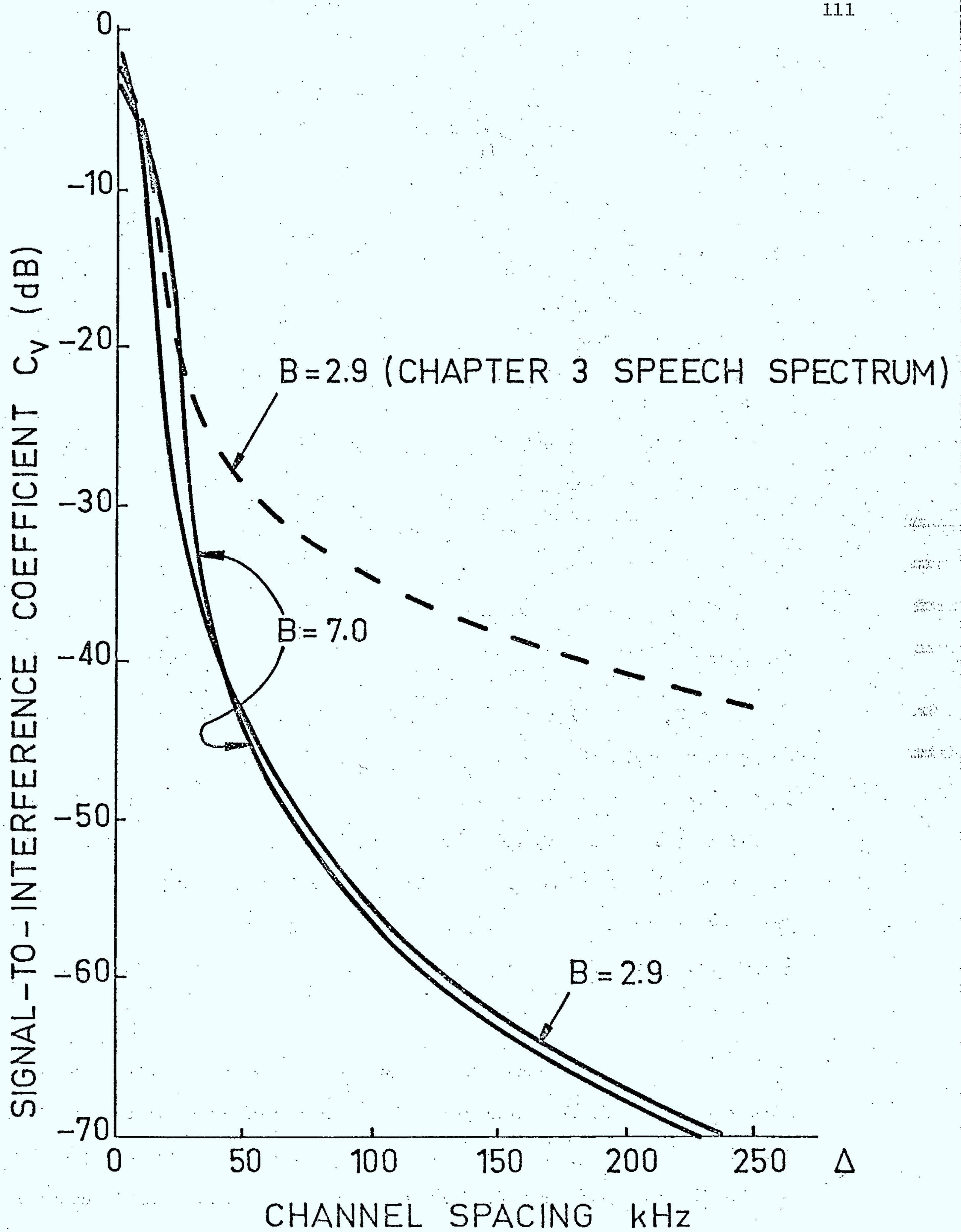

Fig. 5-5 Signal-to-interference coefficient C_v for MSK. Data rate T^{-1} = 20 kHz.

Voice Signal Model	RMS Bandwidth B (kHz)	CHANNEL SEPARATION (kHz)									
		0		25		50		100		200	
		C_v	C_d	C_v	C_d	C_v	C_d	C_v	C_d	C_v	C_d
Chapter 3	2.9	- 2.17	-2.32	-20	-27	-28	-40.5	-34.5	- 53	- 40.5	- 65
Rect. Spectrum	2.9	- 1.50	-2.32	-31.5	-27	-42.8	-40.5	-55.0	- 53	- 67.0	- 65
Rect. Spectrum	7.0	- 3.25	-2.32	-21.7	-27	-42.8	-40.5	-55.7	- 53	- 67.9	- 65

TABLE 5 - 1. Comparison of co-channel and adjacent-channel interference levels (dB) for two speech models with binary MSK at 20 kHz.

Voice Signal Model	RMS Bandwidth B (kHz)	CHANNEL SEPARATION (kHz)									
		0		25		50		100		200	
		TC_v	J_v	TC_v	J_v	TC_v	J_v	TC_v	J_v	TC_v	J_v
Chapter 3	2.9	-45.17	-44.8	-63	-62.0	-71	- 68.3	-77.5	- 74.3	- 83.5	- 79.2
Rect. Spectrum	2.9	-44.50	-43.3	-74.5	-92.5	-85.8	-105.6	-98.0	-107.6*	-110.0	-108.8*
Rect. Spectrum	7.0	-46.25	-47.0	-64.7	-60.8	-85.8	- 98.6	-98.7	-137.3	-110.9	-146 *

TABLE 5 - 2. Comparison between interference coefficients (dB) J_v and C_v for MSK at T^{-1} = 20 kHz (* denotes doubtful accuracy).

tion could deviate substantially from the actual spectrum. The appended Radio Standards would, of course, prevent B from exceeding 5 kHz.

From the results presented here and in Chapter 4, it is clear that final conclusions regarding levels of adjacent-channel interference resulting from transmitted FM (or PM) voice signals requires selection of an appropriate speech spectrum model, as well as measurements to confirm calculation of the transmitted spectrum V(f). Measured interference levels should also be obtained for purposes of comparison with those calculated. In many situations, results might be expected to fall between those obtained here on the basis of two quite different spectral models for speech.

V-4 Interference to FM Voice Transmission

Shown in Fig. 5-6 is a conventional receiver for demodulation of FM voice signals. The IF bandpass filter with transfer characteristic G(f) is selected to remove as much noise as possible without severely distorting the transmitted signal. The FM demodulator includes a limiter which amplifys and hard-limits its input signal, and a discriminator whose output is proportional to the instantaneous frequency of the signal from the limiter. The audio filter removes noise from the baseband signal and performs necessary post-processing including de-emphasis to generate a replica of the original speech signal. The above receiver can also be used for detection of phase modulated signals, provided the signal from the de-emphasis filter is integrated.

Optimization of the bandpass filter is probably an impossibility, because of the complicated way in which bandpass filtering affects the instantaneous frequency of the received signal. One reasonable approach is to

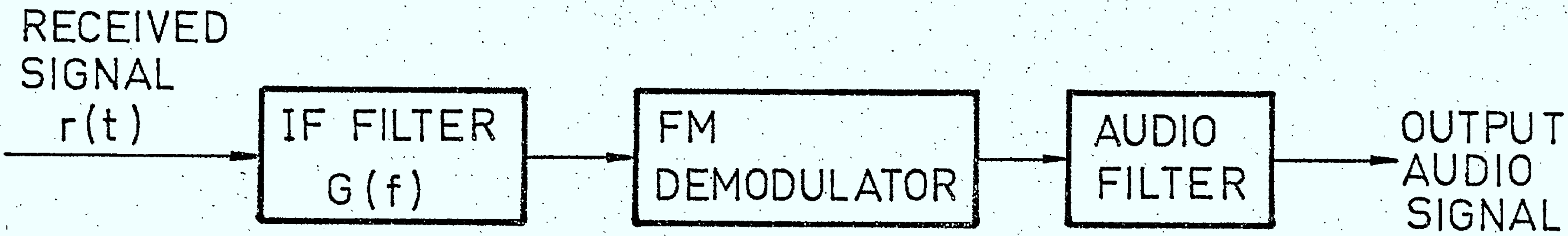

Fig. 5-6 Conventional FM receiver.

select $G(f)$ to maximize the filter's output signal-to-noise ratio Λ. Assuming that this filter does not significantly distort the message portion of the received signal, Λ is maximized as follows [T1, V1]:

$$G(f) = S(f)/[S(f) + S_n(f) + S_i(f)] \tag{5-7}$$

In (5-7) $S(f)$, $S_n(f)$ and $S_i(f)$ denote respectively the power spectral densities of the transmitted FM signal, the received noise and the received interference.

One sees from (5-7) that for those frequencies f where the signal level greatly exceeds the noise-plus-interference level

$$G(f) \simeq 1 \tag{5-8a}$$

and that when the converse is true

$$G(f) \simeq S(f)/[S_n(f) + S_i(f)] \tag{5-8b}$$

Thus, the optimum filter passes with virtually no attenuation those frequencies for which the signal-to-noise-plus-interference level is high, and severely attenuates those frequencies for which the converse is true.

If one assumes that $S_n(f) + S_i(f)$ is flat over that portion of the frequency band which contains most of the signal energy, then $|G(f)|^2 = S(f)$ approximates the criteria in (5-8), in which case

$$|G(f)|^2 = \tfrac{1}{2}[V_{ac}(f-f_c) + V_{ac}(f+f_c)] \tag{5-9}$$

In (5-9) $V_{ac}(f)$ is the power spectral density of the "baseband" transmitted FM signal, excluding any d.c. power. The power from the bandpass filter resulting from the transmitted FM signal is

$$S = \int_{-\infty}^{\infty} |G(f)|^2 \, S(f) \, df \tag{5-10a}$$

$$= P_s \, J_v(0) \tag{5-10b}$$

where

$$J_v(0) = \int_0^{\infty} V_{ac}(f) \, V(f) \, df \tag{5-11a}$$

$$= \int_0^\infty v_{ac}(\tau)\, v(\tau)\, d\tau \qquad (5\text{-}11b)$$

and P_s is the power in the received FM signal. In (5-10) and (5-11) $V_{ac}(f)$ and $V(f)$ are Fourier transforms of $v_{ac}(\tau)$ and $v(\tau)$, as in Section 3-1.

The procedures used in Chapter 4 to quantify interference effects can be employed here to determine the bandpass filter output due to noise and interference. For white noise, $S_n(f) = N_o/2$ which yields a noise output

$$N = \frac{N_o}{2} \int_{-\infty}^{\infty} |G(f)|^2\, df$$

$$= (N_o/2) \int_{-\infty}^{\infty} V_{ac}(f)\, df \qquad (5\text{-}12a)$$

$$= (N_o/2)(1-e^{-\alpha}) \qquad (5\text{-}12b)$$

In (5-12) $\int_{-\infty}^{\infty} V_{ac}(f)\, df$ and $1-e^{-\alpha}$ represent the fraction of the transmitted power not in the carrier signal. Normally, this fraction is close to unity. For the spectrum in Chapter 3, $\alpha = 52.4$ at $B = 2.9$ kHz and $e^{-\alpha}$, the fraction of the power at the carrier frequency equals 1.5×10^{-23}. For $B = 2.9$ kHz for the speech spectrum in this chapter $e^{-\alpha} = 0.06$; for $B = 7$ kHz $e^{-\alpha} = 8.1 \times 10^{-6}$.

The interference resulting from another FM (or PM) voice signal at carrier frequency separation Δ is

$$I_v = P_v \int_{-\infty}^{\infty} |G(f)|^2 S(f-\Delta)\, df$$

$$= P_v \int_{-\infty}^{\infty} \tfrac{1}{4}[V_{ac}(f+f_c)+V_{ac}(f-f_c)]\,[V(f+f_c+\Delta)+V(f-(f_c+\Delta))]\, df$$

$$= P_v\, J_v(\Delta) \qquad (5\text{-}13)$$

where

$$J_v(\Delta) = \int_0^\infty V_{ac}(f)\, V(f-\Delta)\, df \qquad (5\text{-}14a)$$

$$= \int_0^\infty v_{ac}(\tau)\, v(\tau) \cos 2\pi\Delta\tau\, d\tau \qquad (5\text{-}14b)$$

and P_v is the received power in the interfering voice signal.

The interference resulting from a data signal with zero-isi properties is

$$I_d = \frac{P_d}{2} \int_{-\infty}^{\infty} |G(f)|^2 [\,|G_T(f-(f_c+\Delta)|^2 + |G_T(f+f_c+\Delta)|^2]\ df$$

$$= P_d T C_{v_{ac}} \qquad (5\text{-}15a)$$

where $G_T(f)$ is the modulator pulse shape transfer function,

$$C_{v_{ac}} = \int_0^{\infty} V_{ac}(f)\,|G_T(f-\Delta)|^2 df \qquad (5\text{-}15b)$$

$$\int_0^{\infty} v_{ac}(\tau)\rho_g(\tau)\ \cos\ 2\pi\Delta\tau d\tau \qquad (5\text{-}15c)$$

and P_d is the received power in the interfering data signal.

From (4-31) one sees that

$$T(C_v - C_{v_{ac}}) = \int_0^{\infty} (v(\tau) - v_{ac}(\tau))\ \rho_g(\tau)\ \cos\ 2\pi\Delta\tau d\tau\ ,$$

$$= e^{-\alpha} \int_0^{\infty} \rho_g(\tau)\ \cos\ 2\pi\Delta\tau d\tau$$

$$= e^{-\alpha}\ |G_T(\Delta)|^2/2 \qquad (5\text{-}16)$$

From (5-16) one obtains $C_{v_{ac}}$ from $\alpha, |G_T(f)|^2$ and the interference coefficient C_v obtained in Chapter 4. In those cases where a small fraction of the power is in the carrier, $C_v \simeq C_{v_{ac}}$.

From (5-15) and the results in Chapter 4, one sees that, for matched filter detection of data signals, a kind of reciprocity applies. The more narrow the signal spectra the less subject these data signals tend to be to FM voice interference (as well as to data interference), and the less these data signals tend to interfere with voice signals (as well as with other data signals). Similarly, voice signals with narrow spectra tend to be less subject to data interference (as well as to voice interference) and also tend to cause less interference to data signals (as well as to other voice signals). It follows that users of FM voice transmission facilities are able to benefit themselves, as well as adjacent-channel users, by employing techniques which yield FM signals (and IF receivers) with compact spectra. These same comments apply even more forcefully to users of data transmission facilities.

Fig. 5-7 shows $J_v(\Delta)$ vs. channel separation Δ for FM transmission of the speech spectra in Chapter 3 and for PM transmission of the spectra in this present chapter. Table 5-2 compares $J_v(\Delta)$ and $TC_v(\Delta)$ for these two cases, with C_v being obtained from the results for binary MSK at $T^{-1} = 20$ kHz.

It is seen from Table 5-2 as well as from a comparison of Figs. 4-10, 5-5 and 5-7 that for the speech model in Chapter 3, the interference from other voice signals is considerably larger than the interference from MSK data signals, and that this difference increases with channel separation Δ. For the speech model in the present chapter, MSK data interference appears to be more detrimental then voice interference. The fact that C_v in Fig. 5-5 decreases as Δ^{-4} indicates that the rectangular speech spectrum falls off at least as fast as f^{-4}, since the roll-off rate of C_v is governed by the spectrum with the slowest roll-off rate, and MSK spectra decrease as f^{-4}.

Determination of the roll-off rate of V(f) at large values of f is difficult and requires use of approximate techniques or bounds [P1, P2]. Our objective here is not to precisely determine V(f), but rather to indicate the extent to which our interference calculations depend on the various models for speech spectra. In this context the comments in the last paragraph of the previous section are applicable.

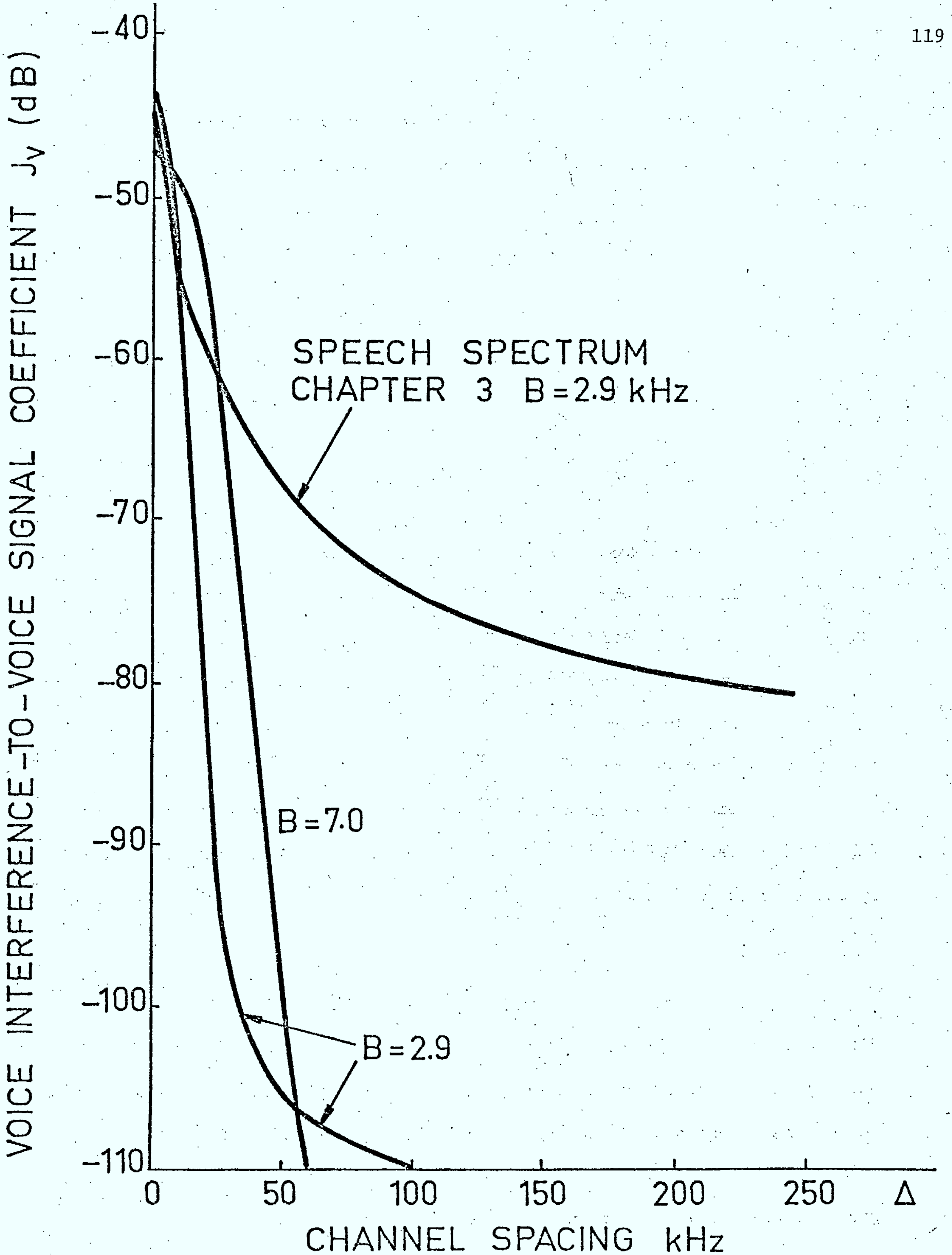

Fig. 5-7 Voice interference coefficient J_v.

Use of G(f) as given by (5-9) permits determination of the IF signal-to-noise ratio Λ for given received levels of P_s, P_d, P_v and $N_o/2$, as follows:

$$\Lambda^{-1} = \frac{(N_o/2P_s) + \sum_v (P_v/P_s)\, J_v(\Delta) + T_d \sum_d (P_d/P_s)\, C_v}{J_v(0)} \quad (5\text{-}17)$$

where in (5-17) the amount of carrier power is assumed negligible. The signal level fluctations described in Chapter 2 require that Λ be averaged over the fluctuations in P_s, P_v and P_d, to determine a meaningful average for Λ [J1].

Maximization of Λ is motivated by the fact that in non-fading environments where noise is the sole source of disturbance, Λ determines the threshold below which the bandwidth-expanding properties inherent in FM no longer manifest themselves in the output signal-to-noise ratio from the audio baseband filter in Fig. 5-7. The baseband output signal-to-noise ratio varies as the mean square bandwidth B^2 of the transmitted signal [R1], and is not strongly dependent on IF filter characteristic G(f) [J1]. In fading environments where interference as well as noise is present, the baseband output signal-to-noise ratio increases with both Λ and B, in a manner which depends on the number of interferers as well as on their spacing from the carrier signal frequency.

V-5 Receivers for Interference Reduction

As the channel separation Δ between the desired and interfering signal increases, the interference level eventually decreases at the spectral roll-off rate of either the receiver filter or interfering signal, whichever is less. For matched-filter detection of data signals, these two roll-off rates are equal, provided the desired and interfering signals use the same data formats. Thus, reduction of the rate of decrease of adjacent-channel interference with increasing Δ depends on the design of transmitted data signals with desirable spectral properties rather than on use of non-matched filter receivers.

As noted in Chapter 4, conventional PSK (and ASK) has a spectral roll-off rate of f^{-2}, which implies a high suseptibility to adjacent-channel interference. Voice signals as modelled by the spectrum in Chapter 3 generate an inherently high level of adjacent-channel FM interference. Raised cosine PAM signals have a very fast spectral roll-off rate, and reject all adjacent-channel interference from similar types of data signals beyond a certain value of Δ.

MSK-type signals as described by (4-46) possess desirable spectral properties. Recent work [E1] shows that U in (4-46) can be selected to reduce adjacent-channel interference from nearby channels below that generated by either MSK or SFSK signals, as shown in Fig. 5-8. Use of the MSK-type signals

$$g_{2T}(t) = \begin{cases} \frac{1}{\sqrt{T}} \cos\left(\frac{\pi t}{2T} + \theta(t)\right) & |t| < T \\ 0 & |t| \geq T \end{cases} \tag{5-18}$$

showed that by selecting $\theta(t)$ carefully, reduced interference from nearby

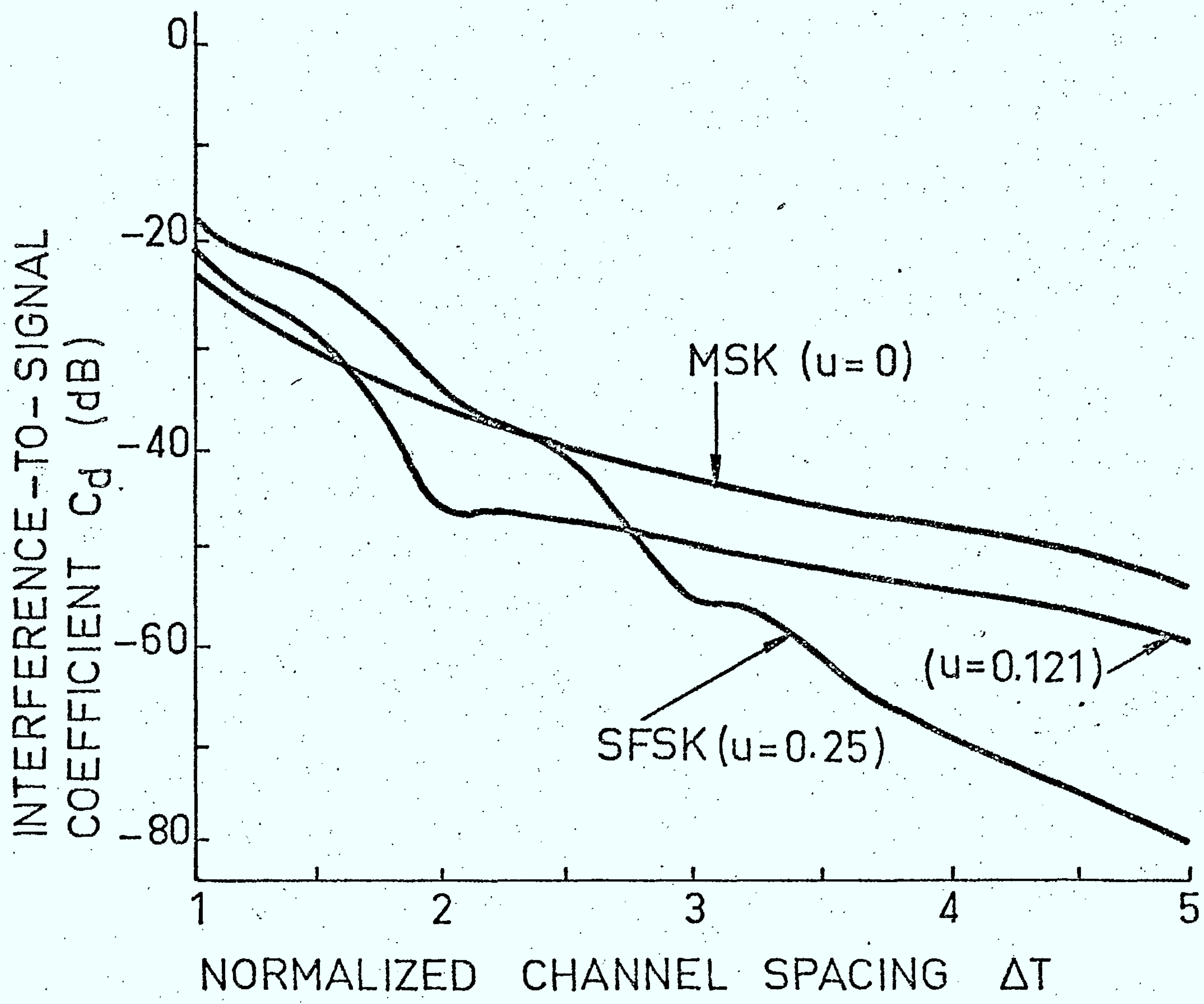

Fig. 5-8 Interference-to-signal coefficient C_y for MSK-type signal in (4-46) (after [E1]).

channels could be exchanged for increased interference from distant channels [E1].

MSK signals can be detected incoherently either by an optimum matched-filter incoherent receiver, or by a conventional FM receiver used for detection of analog FM voice signals [W1, L1, T2]. In the former case, the filters are matched to the MSK sinusoidal pulses [W1, L1] for optimum performance in white Gaussian noise. However, the $1/f^2$ roll-off rate of these filters implies poor adjacent-channel interference protection. Use of filters with impulse response $h(t) = \sqrt{2}\, g_{2T}(2t)$ where $g_{2T}(t)$ is given by (4-46) would have the same spectral roll-off rate as adjacent-channel MSK signals, and would result in a small penalty (typically less than 1 dB) in received signal power. This statement regarding signal power loss follows from the developments in Chapter 4, specifically (4-8) and (4-19).

In using the conventional FM receiver in Fig. 5-6 for MSK demodulation there is no known way to select the IF bandpass filter to minimize the bit-error probability, even if the only source of disturbance is white Gaussian noise [T2]. The best that can be done is to maximize the IF signal-to-noise-plus-interference ratio, which implies an IF filter characteristic with the same spectral roll-off rate as the interfering signal. These same comments regarding selection of the IF filter characteristic apply to demodulation of FM voice signals, and are in accordance with the discussions in Section 5-4. The selection of the bandwidth of the IF filter characteristic would depend on the average level of the interference; the higher this level the narrower the bandwidth which, however, does not affect the roll-off rate.

For incoherent detection of MSK-type signals, arguments similar to those in the above two paragraphs are applicable.

It may be possible to design data receiver filters to reduce adjacent-channel interference at small channel separations Δ by use of non-matched receiver filters. Matched filter designs are based on a flat noise spectrum, and spectra from interfering signals are not flat. Filters selected should not generate isi and should have spectral roll-off rates not less than those of the data signal. Existing studies [P3, G1] on this problem do not specify zero isi filters, do not constrain spectral roll-off rates and assume fixed signal and interference levels. It is unlikely that use of non-matched filters would reduce the interference levels from nearby channels by very much, without some corresponding change in the data signal format, as in the case of MSK-type signals discussed earlier (see Fig. 5-8).

Finally, we reiterate the earlier statement that co-channel interference-to-signal ratios are not much affected by the filter characteristics in either data or conventional FM receivers, because co-channel interference lies in the same band as the desired signal.

V-6 References for Chapter 5

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Government of Canada
Department of Communications

RSS-119
ISSUE 1

RADIO STANDARDS SPECIFICATION

LAND AND MOBILE STATIONS
PRIMARILY VOICE MODULATED FM OR PM
RADIOTELEPHONE TRANSMITTERS AND RECEIVERS
OPERATING IN THE ALLOCATED VHF/UHF BANDS
IN THE FREQUENCY RANGE 27.41 TO 470 MEGAHERTZ

EFFECTIVE DATE: 19 DECEMBER 1980
RELEASE DATE: 15 DECEMBER 1978

TELECOMMUNICATION REGULATORY SERVICE

LAND AND MOBILE STATION: PRIMARILY VOICE MODULATED FM OR PM RADIOTELEPHONE TRANSMITTERS AND RECEIVERS OPERATING IN THE ALLOCATED VHF/UHF BANDS IN THE FREQUENCY RANGE 27.41 TO 470 MEGAHERTZ.

1.0 INTENT

1.1 * This Specification prescribes the minimum requirements for the type-approval of radiotelephone transmitters and receivers operating primarily in a voice modulated mode with maximum frequency deviation of ±5.0 kilohertz in the allocated VHF/UHF bands in the frequency range 27.41 MHz to 470 MHz.

1.2 ALLOCATED BANDS AND STANDARD CHANNEL SPACING: Standard channel spacing in the allocated bands is as follows:

(a) 27.41 - 50 MHz band: 20 kilohertz
(b) 138 - 174 MHz Band: 30 kilohertz
(c) 410 - 420 and 450 - 470 MHz Bands: 25 kilohertz.

Radio Equipment operating in these frequency bands is often employed in an offset frequency environment, particularly near the U.S./Canada border.

1.3 RATED DUTY CYCLE: Application for type-approval shall be made under one or more of the following rated duty cycles:

(a) Continuous
(b) Semi-Continuous
(c) Intermittent
(d) Portable

1.3.1 Duty Cycle rating shall be designated by the suffix of the letter "C", "S", "I", or "P" to the Type-Approval Number corresponding to (a), (b), (c) or (d) above.

1.3.2 Standard test conditions for various duty cycle ratings are contained in paragraph 5.11.

1.4 Transmitters and receivers type-approved under this Specificaton are considered technically suitable for licensing in Canada pursuant to the Radio Regulations made under the Radio Act.

1.4.1 Transmitters with adjustable power output are considered technically suitable for licensing at any power level within the range for which they are approved and for the specific duty cycle for which they are approved.

2.0 GENERAL

2.1 Anyone seeking approval of radio equipment under this Specification shall satisfy the Department at his own expense that the equipment actually meets the requirements of this Specification.

2.2 Notwithstanding the fact that a particular radio equipment meets the requirements of this Specification, the Department reserves the right to require that adjustments be made to that equipment if it causes interference within the meaning of the Radio Act.

2.3 This Specification prescribes minimum requirements only for a type of equipment. Certain users may wish to specify additional requirments to meet specific applications using this type of equipment. Equipment submitted for type-approval must comply with the minimum requirements for all parameters contained herein, and all units of the type manufactured for sale shall comply for the entire period of validity of the type-approval. However, equipment utilizing in-band signalling or low data rate in-band information transfer as additional features to voice transmission will be considered under this Specification for type-approval. Equipment that does not meet the requirements of this standard will be considered for type-acceptance as appropriate, subject to the total occupied r.f. bandwidth and emission requirements stated in this Specification. The applicant, in such cases, will be required to prove, to the satisfaction of the department, the effectiveness of the design and the methods of tests employed to ensure the above.

2.4 The Department reserves the right to revise this Specification.

2.5 This Specification supersedes the existing specifications RSS 105, RSS 126 and RSS 139 in accordance with TRC-44.

3.0 RELATED DOCUMENTS

3.1 RADIO STANDARDS PROCEDURE 100: Procedure to obtain type-approval of Equipment.

3.2 RADIO STANDARDS PROCEDURE 104: Procedure to submit radio equipment for testing at the telecommunications engineering laboratory, Ottawa.

3.3 TRC-44: Implementation of Radio Standards Specification 119 and Amortization of Existing Land and Mobile VHF/UHF equipment type-approved in accordance with Radio Standards Specifications 105, 126 and 139.

4.0 EQUIPMENT REQUIREMENTS

4.1 EQUIPMENT IDENTIFICATION: The serial number of the transmitter, the name, the type of unit, the manufacturer's name, the country of manufacture, the type-approval number, as well as any other information required to identify the equipment shall be stamped on each transmitter chassis or on a nameplate permanently fastened to the chassis. The transmitter power output rating shall also be included, unless it appears on the nameplate of the overall equipment assembly of which the transmitter forms a part.

4.2 TRANSMITTER POWER OUTPUT CONTROLS: Operator controls shall not permit operation at power levels greater than that permitted by the license.

5.0 STANDARD TEST CONDITIONS AND DEFINITIONS

5.1 GENERAL: Standard test conditions are those conditions under which the equipment shall be operated while it is being tested for minimum requirements. These conditions apply, unless otherwise specified.

5.1.1 Standard Operating Conditions:

5.1.2 General: Except as specified below, the equipment shall be operated in accordance with the manufacturer's published instructions and in the case or cabinet supplied, where this is essential to the performance of the equipment.

5.1.3 Transmitter: Modulation limiting shall be initially adjusted in accordance with the manufacturer's published instructions for maximum frequency deviation of ±5 kilohertz and shall remain operative for all tests without re-adjustment.

*

5.1.4 Receiver: For all tests the receiver external control shall be unsquelched to the maximum extent and the r.f. external gain control shall be set to full gain.

5.1.5 Associated Equipment: Associated equipment shall be that normally used with the transmitter and/or receiver. Standard conditions shall apply also to the associated equipment.

5.2 STANDARD TEST VOLTAGE: Standard Test Voltage shall be within ±2% of the value stated by the manufacturer to be the test voltage. It shall be measured at the point of power connection to the equipment. For equipment powered by a battery, an appropriate substitute variable power source may be used for tests under environmental conditions.

5.3 WARM-UP PERIOD: The equipment may have a 15 minute warm-up period under standby conditions prior to all tests, except where otherwise specified.

5.4 STANDARD ATMOSPHERIC CONDITIONS: Tests shall be conducted under ambient conditions of atmospheric pressure and humidity at a temperature of 25°C ±5°C.

5.5 STANDARD TEST FREQUENCY: Except where otherwise specified, all tests shall be conducted on an assignable frequency which is near the middle of the frequency range within which the equipment is designed to operate and shall be specifically identified by the type-approval applicant.

5.6 STANDARD INPUT AND OUTPUT TERMINATIONS: Transmitter and receiver strandard terminations shall be resistive and numerically equal to the design impedances at the specific terminals concerned. The values shall be stated by the manufacturer or type-approval applicant and recorded in the test report. All input and output terminals shall be properly terminated under all test conditions.

5.7 STANDARD SYSTEM FREQUENCY DEVIATION: The standard system frequency deviation shall be ±5 kHz maximum. This is the maximum allowable system deviation and represents 100% modulation.

5.8 STANDARD TEST MODULATION: Standard test modulation shall be a 1000 hertz sinusoidal audio input signal having 1% or less total harmonic distortion applied at the level required to produce 60% of standard system frequency deviation (ie. ±3.0 kHz).

5.9 RECEIVER STANDARD INPUT SIGNAL SOURCE: Shall be a signal generator whose output impedance shall be matched to the nominal input impedance of the receiver (which shall be stated by the manufacturer) and whose calibrated output voltage shall be corrected to take into account impedance matching losses when terminated. Incidental modulation of the output signal voltage shall not affect the measurement accuracy.

5.10 INTERCONNECTION OF SIGNAL GENERATORS FOR MULTIPLE SIGNAL TESTS: When it is necessary to couple two or more signal generators to a receiver, a combining network shall be used such that each signal generator shall see a matched load when the output of the network is matched to the nominal input impedance of the receiver. Account shall be taken of losses in the combining and/or matching networks to establish the signal levels delivered to the receiver. The effect of intermodualtion products and noise generated within the signal generators shall not affect the measurement accuracy.

5.11 STANDARD TEST DUTY CYCLES: Transmitter rated duty cycle shall be designated by the manufacturer as either continuous, semi-continuous, intermittent or portable. The transmitter shall be tested in accordance with the following test conditions:

(a) Continous: The transmitter shall be operated continuously at full rated output power for a period of twenty-four (24) hours.

(b) Semi-Continuous: The transmitter shall be operated continuously at full rated output power for a period of eight (8) hours.

(c) Intermittent: The transmitter shall be operated at full rated output power under a cycle of one minute, carrier "on" (transmit), and four minutes, carrier "off" (standby), for a period of eight (8) hours, followed by three test cycles of five minutes "on", and fifteen minutes "off".

(d) Portable: As a minimum, the transmitter shall be operated at full rated power under a duty cycle of 0.25 minutes, carrier "on" (transmit), and 4.75 minutes, carrier "off" (standby), for a period of four (4) hours.

5.12 STANDARD TEST EQUIPMENT REQUIREMENTS

5.12.1 Standard Test Receiver: The standard test receiver shall consist of a configuration of test equipment complying with the following minimal requirements:

(a) It shall be tunable over the applicable range of radio frequencies.
(b) It shall be capable of measuring positive and negative peak values of carrier frequency deviations up to ±10 kilohertz with an accuracy of 5% or better.
(c) It shall incorporate a switchable de-emphasis network whose audio response does not vary more than ±1 dB from a 750 microsecond de-emphasis characteristic over the frequency range 50 to 6000 hertz. (6 dB per octave from 210 Hz to 6000 Hz)
(d) The nominal 3 dB audio band pass shall be from 50 to 50,000 hertz and the variation in the response shall not exceed ±0.5 dB within the range 300 to 3000 hertz.
(e) Distortion due to signal processing shall not exceed 2%.
(f) Inherent hum and noise output shall be at least 50 dB below the level of a signal deviated by 5 kHz.
(g) It shall be properly terminated during all

tests.

5.12.2 Spectrum Analyzer or Frequency Selective Voltmeter

(a) Shall have the capability of measuring each of two signals differing in frequency by 1000 Hz and in level by 60 dB, with an accuracy of ±2 dB.
(b) Shall have a dynamic range of at least 70 dB.
(c) Shall be capable of measuring relative levels of input signal components with an accuracy of ±1 dB or better.

5.12.3 Adjacent Channel Power Measuring Arrangement

* The adjacent channel power measuring arrangement shall comprise a mixer, a filter, a variable attenuator, a linear intermediate amplifier and a true RMS power meter connected in cascade, using a low noise signal generator as a local oscillator. The bandwidth requirements of the filter shall be as follows: (with a tolerance of ±5%)

Bandwidth between 6 dB attenuation points 16 KHz
Bandwidth between 70 dB attenuation points 35 KHz
Bandwidth between 90 dB attenuation points 50 KHz

The attenuator shall cover a range of at least 80 dB in 1 dB steps. The noise factor of the intermediate amplifier shall not be worse that 4 dB. Over the 6 dB bandwidth, the amplitude/frequency characteristics of the intermediate amplifier shall not vary by more than 1 dB.

The combined response of the filter amplifier outside the 90 dB bandwidth shall maintain an attenuation of at least 90 dB. The accuracy of the overall measuring arrangement over an input level range of 100 dB shall be better than 1.5 dB.

5.13 INSTRUMENTATION ACCURACY: The test report shall list all test equipment used and its date of calibration. The list shall identify instruments by manufacturer's type or model number. The following parameters shall be measured with instrumentation having at least the accuracy indicated below:

(a)	AC and DC power supply voltages	±1% (Full Scale)
(b)	Radio frequency	±0.00001%
(c)	Radio frequency power	±10%
(d)	Radio frequency output voltages	±2 dB
(e)	Radio frequency signal input voltages	±1 dB
(f)	Audio frequency	±1%
(g)	Audio frequency power	±5% (Full Scale)
(h)	Audio frequency noise and distortion	±5% (Full Scale)

(i) Ancillary load resistors, coupling devices, cables, attenuators, etc. shall have characteristics stated with a tolerance of ±5%.

6.0 ENVIRONMENTAL TEST CONDITIONS

6.1 ENVIRONMENTAL TEST CONDITIONS

(a) Self contained equipment.
Temperature: -10°C, +40°C.

(b) Equipment, including portable, intended for use with external power supplies:
Temperature: -30°C, +60°C.
Voltage: ±10% of standard test voltage.

6.1.1 Equipment with Self-Contained Batteries

(a) During environmental tests it shall be permissible to substitute an external voltage source. In these cases the batteries may be disconnected, but not removed.

(b) When tested with an external power supply, the equipment shall be operated under the conditions applicable to externally powered equipment.

6.2 TOLERANCES: Environmental test conditions shall be maintained within the following tolerances:

(a) Temperature: ±2°C.
(b) Voltage: ±2%.

6.3 TEMPERATURE STABILIZATION: For the purposes of these tests, equipment temperatures shall be considered stabilized when the temperature of the largest internal mass remains within +3°C of the specified value when the equipment is inoperative, or when the crest temperature of the largest internal mass does not vary by more than ±5°C, with the equipment operating.

7.0 TRANSMITTER TESTS UNDER STANDARD TEST CONDITIONS

7.1 POWER OUTPUT AND POWER OUTPUT RATING

7.1.1 Definitions: The transmitter power output is the r.f. power dissipated in the standard output termination when operating under the rated duty cycle selected by the applicant for approval. The power output rating of the transmitter is the value determined under the conditions of paragraph 7.1.2.

7.1.2 Method of Measurement: The transmitter shall be set up as specified under the standard test conditions and operated unmodulated at the maximum output power at the selected duty cycle designated in accordance with paragraph 5.11. Measurements shall be made to establish the radio frequency power delivered by the transmitter into the standard output termination. All details of the measurement shall be clearly stated in the test report. Power output shall be continuously monitored and recorded throughout the test period and no adjustment shall be made to the transmitter after the test has begun, except as noted below:

If the power output is adjustable, measurements shall be made for the highest and lowest power levels of the range for the appropriate duty cycle for which approval is requested. The results shall be recorded at the end of the test period.

7.1.3 Minimum Performance Standard: The maximum power output(s) measured under the conditions of paragraph 7.1.2 for each duty cycle for which the transmitter is rated shall be within +1 to 0 dB of the manufacturer's rating(s) of r.f. power output.

7.2 ADJACENT CHANNEL POWER AND MAXIMUM DEVIATION

7.2.1 Definition: Adjacent channel power is that part of the total power output of a transmitter, under defined conditions of modulation, which falls within a 16 kHz bandwidth, centred 25 kHz from the carrier frequency. The maximum deviation is the maximum value obtained under the defined conditions of modulation.

7.2.2 Method of Measurement: The transmitter shall be operated at full rated power output, and shall be modulated by a 2500 Hz sinusoidal signal at an input level 16 dB greater than that required to produce ±2.5 kHz deviation at 1000 Hz, and the deviation shall be measured. A portion of the r.f. output shall be coupled to the adjacent channel measuring arrangement. The adjacent channel power measuring arrangement is to be tuned to the carrier frequency and the attenuator shall be adjusted to provide a reference output reading of between 3 and 8 dB above the noise level. The adjacent channel power measuring arrangement is then retuned to a centre frequency 25 kHz higher than the carrier frequency, and the attenuator is re-adjusted to the reference output reading. The difference in the attenuator setting is the upper adjacent channel power expressed in dB relative to the on channel power. The measurement is to be repeated with the adjacent channel power measuring arrangement tuned to a centre frequency 25 kHz below the carrier frequency.

The above measurements are to be repeated for each power output rating.

7.2.3 Minimum Performance Standards:

* (a) The power in either adjacent channel shall be the lesser of 60 dB below the "on" channel power or 50 microwatts.

(b) A plot of the actual characteristics of the filter used in the measurement shall be provided along with the reference power output reading.

(c) The maximum deviation shall not be greater than ±5 kHz.

7.3 SPURIOUS EMISSIONS

7.3.1 Definition: Spurious emissions are emissions at any frequency outside the band necessary to ensure the adequate transmission of information. Spurious emissions include harmonic emissions, parasitic emissions, and intermodulation products which are remote from the immediate vicinity of this band.

7.3.2 Method of Measurement: The measurement set-up shall be specified in paragraph 7.2.2, except that standard test modulation shall be applied to the transmitter and a portion of the r.f. output shall be coupled to a spectrum analyzer or equivalent instrument. The output spectrum shall be carefully searched for spurious emissions over the frequency range from 1 megahertz to the higher of 1000 megahertz or the third harmonic of the standard test frequency (Fc), ±50 kilohertz from the standard test frequency. All spurious emissions greater than 2.5 microwatts shall be identified, measured, and recorded. The above measurement shall be repeated for each power output rating.

7.3.3 Minimum Performance Standard: No spurious emissions shall exceed 25 microwatts.

7.4 MODULATION DISTORTION

7.4.1 Definition: The modulation distortion of the transmitter is the minimum acceptable deviation of the carrier frequency without exceeding a specified level of distortion of the demodulated test signal.

7.4.2 Method of Measurement: The transmitter shall be operated under standard test conditions. The standard test receiver, with the standard de-emphasis network operative, shall be loosely coupled to the transmitter output and carefully tuned to the standard test frequency. Standard test modulation shall be applied to the transmitter. The frequency deviation of the carrier and the distortion of the demodulated test signal shall be measured. The above procedure shall be repeated with a modulating signal of 2000 hertz applied at a level to produce at least ±3.0 kilohertz deviation.

7.4.3 Minimum Performance Standard: The minimum modulation distortion shall be ±3.0 kilohertz deviation, without exceeding 10% distortion of the demodulated test signal.

7.5 AUDIO FREQUENCY RESPONSE

7.5.1 Definition: The transmitter audio frequency response is defined in terms of the degree of closeness to which the frequency deviation of the transmitter follows the prescribed 6 dB per octave pre-emphasis characteristic over a specified continuous audio frequency range, without exceeding flat response beyond that range, under conditions of a constant input level of modulating frequencies.

7.5.2 Method of Measurement: The measurement set-up shall be as specified in paragraph 7.4.2. The de-emphasis network of the standard test receiver shall be switched out for this test. The input level of the 1000 hertz modulating signal shall be adjusted for ±1.0 kilohertz deviation of the transmitter carrier frequency. The demodulated audio output level shall be measured and recorded as the reference level. With the audio input level held constant, the modulating frequency shall be varied from 300 to 10,000 hertz and corresponding demodulated output levels shall be measured and recorded. A sufficient number of measurements shall be recorded to plot a smooth curve on Figure 1. This plot shall be contained in the test report.

7.5.3 Minimum Performance Standard: The transmitter audio frequency response, when plotted, shall not lie anywhere in the hatched areas of Figure 1.

7.6 HUM AND NOISE LEVEL

7.6.1 Definition: The hum and noise level of the transmitter is defined in terms of the ratio of the level of the demodulated output under conditions of standard test modulation, to the level of the demodulated output with no modulation applied, both measured with the same bandwidth.

7.6.2 Method of Measurement: The measurement set-up shall be as specified in paragraph 7.4.2, with standard test modulation applied to the transmitter. The level of the demodulated output of the standard test receiver shall be measured and recorded. The modulation source shall be removed from the transmitter, and its audio input terminals shall be properly terminated. The hum and noise output level of the receiver shall be measured again and recorded. The difference in measured output levels, expressed in decibels, is the hum and noise level.

7.6.3 Minimum Performance Standard: The hum and noise level shall be at least 37 dB below the output level, with standard test modulation.

8.0 TRANSMITTER TESTS UNDER ENVIRONMENTAL CONDITIONS

8.1 OPERATIONAL PERFORMANCE STABILITY

8.1.1 Definition: Operational performance stability is the measured ability of the transmitter to function under environmental conditions of specified variations in power supply voltage, over a specified temperature range, without exceeding permissible carrier frequency stability tolerances, and with no more than a specified variation in power output.

8.1.2 Method of Measurement: The measurement procedure outlined below shall be followed:

Step 1 The transmitter shall be installed in an environmental test chamber whose temperature is controllable. Provision shall be made to vary the transmitter power supply voltage and to control the operation of the transmitter. Provision shall also be made to measure the following parameters:
(i) r.f. carrier frequency
(ii) r.f. power output.

Note 1 If the transmitter utilizes a self-contained battery power supply, new or fully charged batteries may be installed at the outset of environmental testing and Step 4 of this test procedure shall not apply, or alternatively an external power supply can be used.

Note 2 The operating frequency shall be set-up in accordance with the manufacturer's published operation and instruction manual prior to the commencement of these tests. No adjustment of any frequency determining circuit element shall be made subsequent to this initial set-up.

Step 2 With the transmitter inoperative (power switched "OFF"), the temperature of the test changer shall be adjusted to +25°C. After a one-hour temperature stabilization period at +25°C, the transmitter shall be switched "ON" with standard test voltage applied. The transmitter shall be operated in the "standby" mode for 15 minutes prior to keying the carrier "ON".

Step 3 The carrier shall be keyed "ON", and the transmitter shall be operated unmodulated at full r.f. power output for a continuous 5-minute period. The following parameters shall be measured in the order indicated:

(i) the r.f. carrier frequency shall be continuously monitored, and measurements shall be recorded at the outset and at 30 second intervals;

(ii) the r.f. power output shall be measured and recorded during the last minute of operation. This measurement shall be the reference level for minimum performance standards under environmental conditions.

Step 4 With the test chamber temperature maintained at +25°C, the measurement procedure specified in Step 3 shall be repeated with the power supply voltage adjusted to 110% and 90% of standard test voltage. In addition, for portable equipment tested under the duty cycle (d) frequency tolerance measurements shall be made with the external power supply set at 80% of the standard test voltage. All measurements shall be recorded. The transmitter shall be switched "OFF".

Note: For "intermittent" and "portable" duty cycle equipment, a maximum period of 15 minutes in "standby" mode may be allowed between periods of carrier keyed "ON" operation.

Step 5 The test procedures outlined in Steps 2, 3 and 4 shall be repeated after stabilizing the transmitter at the following environmental temperatures in the order indicated: self contained equipment, +40°C, -10°C; equipment, including portable, intended for use with external power supplies, +60°C, -30°C. All measurements shall be recorded and reported.

8.1.3 Minimum Performance Standards: Under all specified environmental test conditions:

(i) The r.f. carrier frequency shall not depart from the standard test frequency in excess of the following tolerances:

Frequency Band	Frequency Tolerances
(a) 27.41-50 MHz:	$\pm$545 to $\pm$1000 hertz (linear interpolation)
(b) 138-174 MHz:	$\pm$1740 hertz
(c) 410-420 and 450-470 MHz;	$\pm$2350 hertz

(ii) The r.f. power output shall not decrease by more than 3 dB, nor increase by more than 2 dB with reference to the power output measured under paragraph 8.1.2 (Step 3 (ii)).

9.0 RECEIVER TESTS UNDER STANDARD TEST CONDITIONS

9.1 REFERENCE AUDIO OUTPUT POWER

9.1.1 Definition: The reference audio output power of the receiver is the audio output power it will deliver to the standard output termination, without exceeding a specified level of distortion when responding to a specified RF signal input.

9.1.2 Method of Measurement: A standard input signal source of 100 microvolts, with standard modulation, shall be applied to the receiver input at its resonant frequency. The audio level control shall be adjusted so that the receiver delivers the maximum audio power output without exceeding 10% distortion, or the manufacturer's specified maximum power level, which-ever is less. The audio output power shall be measured and recorded.

9.1.3 Reference Level: The audio output power measured under paragraph 9.1.2 is the reference audio output power that will be used in subsequent sections of this specification.

9.2 AUDIO FREQUENCY RESPONSE

This test is to be conducted for all equipment, but the limits referenced in 9.2.3 apply only to equipment having an output characteristic suitable for a connection to a line.

9.2.1 Definition and Purpose: The audio frequency response at an available output is the amplitude/frequency envelope resulting from an r.f. input signal that is modulated at a specified constant frequency deviation over a specified range of frequencies.

This standard verifies that the audio frequency response complies, within specified limits, to a 6 dB per octave de-emphasis characteristic in order to control the overall audio response on a radio link between transmitter modulation input and receiver audio output.

9.2.2 Method of measurement: An r.f. input signal of 1000 microvolts, modulated with standard test modulation, shall be applied to the receiver.

The modulating signal shall be adjusted to provide 1 kHz deviation at 1000 Hz, and the new audio output level shall be measured and designated the 0 dB reference point.

The modulating frequency and input level shall be adjusted to provide a constant 1 kHz signal frequency deviation at the following modulating frequencies: 300 Hz, 500 Hz, 1000 Hz, 1500 Hz, 2000 Hz, 2500 Hz, 3000 Hz, 5000 Hz, 7500 Hz and 10,000 Hz. At each of these discrete steps the receiver audio output level shall be measured and referred in dB to the 0 dB reference point.

9.2.3 Minimum Performance Standard: In the range 300 to 3000 Hz, the audio frequency response limits shall be within +1/-3 dB of the 6 dB/octave de-emphasis characteristic. In the range 3000 Hz to 10000 Hz the level shall not exceed the 6 dB/octabe de-emphasis characteristic.

9.3 SENSITIVITY

9.3.1 Definition: The sensitivity of a receiver is the minimum level of standard modulated r.f. input signal at the receiver's resonant frequency, which will produce a minimum signal plus noise, plus distortion to noise, plus distortion ratio (SINAD) of 12 dB, and at least 50% of rated audio power output.

9.3.2 Method of Measurement: The initial measurement set-up shall be as specified in paragraph 9.1.2. With the receiver audio level control adjusted for rated audio power output, the r.f. input signal shall be reduced to the minimum level which will simultaneously satisfy both conditions of a minimum 12 dB SINAD and at least 50% of rated audio power output, without re-adjustment of the audio level control. In the event that 50% of rated audio power output is not attainable at a SINAD of 12 dB, the r.f. signal input level shall be increased to obtain the 50% value. The SINAD and minimum r.f. signal input required shall be measured and recorded.

9.3.3. Minimum Performance Standard: The r.f. signal input level measured across the receiver antenna terminals shall not be more than 0.75 microvolts across 50 ohms, or an equivalent input power (-109.5 dBm). For equipment using integral duplexers, the r.f. signal level measured across the duplexer antenna terminals shall not be more than 1.0 microvolt across 50 ohms or an equivalent input power (-107 dBm).

9.4 TWO-SIGNAL SELECTIVITY AND DESENSITIZATION CHARACTERISTIC

9.4.1 Definition: The two-signal selectivity and desensitization characteristic of a receiver is the measure of its ability to process an on-channel desired signal in the presence of an undesired signal on a nearby frequency, without exceeding a specified degradation of output.

9.4.2 Method of Measurement: Two similar generators shall be equally coupled to the receiver antenna input terminals through a suitable combining network (see para. 5.10). With the output of signal generator No. 2 adjusted to zero, the receiver and signal generator No. 1 shall be adjusted as outlined in paragraph 9.3.2. Signal generator No. 2 shall be set at the adjacent and offset (half channel spacing channel) frequencies and shall be modulated with a 400 hertz sinusoidal signal at a level of 60% of standard system frequency deviation (ie. $\pm$3.0 kilohertz deviation). At each frequency setting of signal generator No. 2, its output level shall be increased until the SINAD is reduced to 6 decibels, or until the total audio output power of the receiver is reduced by 3 decibels, whichever event occurs first. The corresponding output levels of signal generator No. 2 shall be measured and recorded, and their ratios, expressed in decibels with reference to the output level of signal generator No. 1, shall also be recorded.

9.4.3 Minimum Performance Standards:

(i) At adjacent channel frequencies the selectivity shall be at least 70 dB.

(ii) For use in 410-420, 450-470 MHz, selectivity shall be at least 30 dB at $\pm$12.5 kHz. For 140-174 MHz, selectivity shall be at least 30 dB at 15 kHz.

9.5 SPURIOUS RESPONSE ATTENUATION

9.5.1 Definition: Spurious response attenuation is the ratio of the response to any undesired signal to its response at its resonant frequency.

9.5.2 Method of Measurement: The method of measurement described in section 9.4.2 shall be used. Signal generator 2 shall be unmodulated. A careful search shall be made for spurious responses by slowly tuning the frequency of the input signal (2) source over the frequency range from the lowest intermediate frequency of the receiver to 1000 megahertz, excluding the band defined by the resonant frequency of the receiver plus and minus the standard channel spacing. Subharmonics of the resonant frequency of the receiver are also to be excluded in the search for spurious emissions. All spurious responses shall be identified by frequency and recorded. At each spurious response, signal generator No. 2 output level shall be increased until the SINAD is reduced to 6 decibels, or until the total audio output power of the receiver is reduced by 3 decibels, whichever event occurs first. The corresponding output levels of signal generator No. 2 shall be measured and recorded, and ratios of signal generator (1) & (2) levels shall be expressed in decibels, with reference to the output level of signal generator No. 1.

9.5.3 Minimum Performance Standard All spurious responses shall be attenuated by at least 70 dB.

9.6 INTERMODULATION SPURIOUS RESPONSE ATTENUATION

9.6.1 Definition: Intermodulation spurious response attenuation of a receiver is its measured ability to process a desired standard-modulated signal at its resonant frequency, without exceeding a specified degradation of output in the presence of two undesired signals, at freqencies which are so separated from the desired signal and from one another that nth order mixing in the non-linear elements of the receiver may generate spectral components which are within the pass band of the receiver. This standard is concerned only with third order mixing of frequencies which are separated from the desired frequency and from one another by integral multiples of the standard channel spacing.

9.6.2 Method of Measurement: The procedure outlined below shall be followed:

Step 1 With the output of the receiver terminated in the standard output termination, three equivalent standard signal generators shall be equally coupled to the receiver input (see para. 5.10). Signal generator No. 1 shall be modulated with standard test modulation; signal generator No. 2 shall be unmodulated; and signal generator No. 3 shall be modulated with a 400 hertz sinusoidal signal at a level of ± 3.0 kilohertz deviation.

Step 2 With the output levels of signal generators No. 2 and No. 3 adjusted to zero, the frequency of signal generator No. 1 shall be adjusted to the resonant frequency (fo) of the receiver and its output level shall be adjusted for a 12 dB SINAD at the receiver output.

Step 3 Signal generator No. 2 shall then be adjusted to the adjacent channel above the resonant frequency (ie. to fo + Δf, where Δf is the standard channel spacing), and signal generator No. 3 shall be adjusted to the alternate channel above the resonant frequency (fo + 2Δf).

Step 4 The output levels of signal generators No. 2 and No. 3 shall be maintained equal and increased simultaneously until the SINAD is reduced to 6 dB. The frequency of signal generator No. 3 shall be re-adjusted to produce a maximum degradation of SINAD before the required final output levels of signal generators No. 2 and No. 3 are established, measured, and recorded. The ratio of the output level of signal generator NO. 2 (or No. 3) to the output level of signal generator No. 1, expressed in decibels, is the intermodulation spurious response attenuation for the particular combination of undesired input signals present.

Step 5 The above procedure (Steps 2-4) shall be repeated with signal generators No. 2 and No. 3 adjusted to frequencies which are both above and both below the resonant fréquency of the receiver, as indicated in the following table:

	SIGNAL GENERATOR FREQUENCIES		
TEST NO.	S.G. No. 1	S.G. No. 2	S.G. No. 3
1 & 2 3 & 4	fo fo	fo $\pm\Delta f$ fo $\mp 2\Delta f$	fo $\pm 2\Delta f$ fo $\mp 4\Delta f$

Step 6 All measurements shall be recorded and reported.

9.6.3 Minimum Performance Standard: All intermodulation spurious responses shall be attenuated by at least 60 dB.

9.7 ANTENNA CONDUCTED RECEIVER SPURIOUS EMISSIONS

9.7.1 Definition: Antenna conducted receiver spurious emissions are radio frequency output voltages generated within the receiver, which appear at the antenna terminals.

9.7.2 Method of Measurement: The receiver antenna terminals shall be terminated in, or impedance matched to, a spectrum analyzer or frequency selective voltmeter whose nominal input impedance is 50 ohms resistive. The receiver shall be operated in the normal receiver mode on at least three test frequencies, one near the mid-point and the others approximately 10% inside the upper and lower extremities of the band over which the receiver is designed to operate. At each frequency of operation the output shall be searched by carefully tuning the spectrum analyzer or frequency selective voltmeter over the range from the lowest radio frequency generated in the receiver, to three times its operating frequency, or to 1000 megahertz, whichever is higher. All detected outputs shall be investigated and those within 20 dB of the permissible level shall be identified by frequency, measured, and recorded. If the receiver incorporates a scanning mode of operation, the above procedure shall be repeated with the receiver operating in the scanning mode.

9.7.3 Minimum Performance Standard: Radio frequency output power at the antenna terminals at any discrete frequency shall not exceed 800 picowatts (200 microvolts across 50 ohms).

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FIGURE I (PARAGRAPH 7.5.3)