Analysis and Simulation of a Digital Mobile Channel Using Orthogonal Frequency Division Multiplexing

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***Abstract*—This paper discusses the analysis and simulation of a technique for combating the effects of multipath propagation and cochannel interference on a narrow-band digital mobile channel. This system uses the discrete Fourier transform to orthogonally frequency multiplex many narrow subchannels, each signaling at a very low rate, into one high-rate channel. When this technique is used with pilot-based correction, the effects of flat Rayleigh fading can be reduced significantly. An improvement in signal-to-interference ratio of 6 dB can be obtained over the bursty Rayleigh channel. In addition, with each subchannel signaling at a low rate, this technique can provide added protection against delay spread. To enhance the behavior of the technique in a heavily frequency-selective environment, interpolated pilots are used. A frequency offset reference scheme is employed for the pilots to improve protection against cochannel interference.**

I. INTRODUCTION

SEVERE multipath propagation, arising from multiple scattering by buildings and other structures in the vicinity of a mobile unit, makes the design of a mobile communication channel very challenging [1]. This scattering produces rapid random amplitude and phase variations in the received signal as the vehicle moves in the multipath field. In addition, the vehicle motion introduces a Doppler shift, which causes a broadening of the signal spectrum. Measurements confirm that the short-term statistics of the resultant signal envelope approximate a Rayleigh distribution.

Multipath fading may also be frequency selective, that is, the complex fading envelope of the received signal at one frequency may be only partially correlated with the received envelope at a different frequency. This decorrelation is due to the difference in propagation time delays associated with the various scattered waves making up the total signal. The spread in arrival times, known as delay spread, causes transmitted data pulses to overlap, resulting in intersymbol interference. In a typical urban environment, a spread of several microseconds and greater can be occasionally expected.

There is an additional impairment in a *cellular* mobile system. The available radio channels are reused at different locations within the overall cellular service area in order to use the assigned spectrum more efficiently. Thus, mobiles simultaneously using the same channel in different locations interfere with each other. This is termed cochannel interference and is often the dominant impairment.

In addition, there is a long-term variation of the local mean of the received signal, called shadow fading. Shadow fading in a mobile radio environment is caused by large obstacles blocking the transmission path. This impairment is alleviated in cellular systems by using transmitted and received base-station signals at two different geographical locations [1], and will not be discussed in this paper.

Given the harsh mobile environment and the scarcity of available spectrum, it is desirable to look for channel designs which provide good performance for both speech and data transmission, and which are also bandwidth efficient. The channel designs presented in this paper could accommodate speech or data transmission. For the *narrow* channel assumed, a low-bit-rate speech coder would be required. For example, a 7.5 kHz channel using the system proposed in this paper can support 8.6 kbits/s. In what follows, the channel will be assumed to be transmitting data symbols.

In a conventional serial data system, the symbols are transmitted sequentially, with the frequency spectrum of each data symbol allowed to occupy the entire available bandwidth. Due to the bursty nature of the Rayleigh channel, several adjacent symbols may be completely destroyed during a fade. To illustrate the severity of the problem, consider the following example. Assume that there is a cochannel interferer with an average power level 17 dB below that of the desired signal. This condition occurs approximately 10 percent of the time in a cellular mobile system. A fade 17 dB below the average level will bury the desired signal in the interference. At a carrier frequency of 850 MHz and a vehicle speed of 60 mph, the average fade duration for a fade 17 dB below the local mean of the desired signal is 0.75 ms [1]. For a data rate of 10 kbits/s, 7 or 8 adjacent bits would be destroyed during such a fade.

In a serial system, higher data rates can be achieved, at the expense of a degradation in performance, by using higher order modulations or, at the expense of increased channel bandwidth, by decreasing the symbol interval. However, delay spread imposes a waiting period that determines when the next pulse can be transmitted. This waiting period requires that the signaling be reduced to a rate much less than the reciprocal of the delay spread to prevent intersymbol interference. Decreasing the symbol interval makes the system more susceptible to delay spread impairments.

A parallel or multiplexed data system offers possibilities for alleviating many of the problems encountered with serial systems. A parallel system is one in which several sequential streams of data are transmitted simultaneously, so that at any instant many data elements are being transmitted. In such a system, the spectrum of an individual data element normally occupies only a small part of the available bandwidth. In a classical parallel data system, the total signal frequency band is divided into N *nonoverlapping* frequency subchannels. Each subchannel is modulated with a separate symbol and, then, the N subchannels are frequency multiplexed. A more efficient use of bandwidth can be obtained with a parallel system if the spectra of the individual subchannels are permitted to overlap, with specific orthogonality constraints imposed to facilitate separation of the subchannels at the receiver.

A parallel approach has the advantage of spreading out a fade over many symbols. This effectively randomizes the burst errors caused by the Rayleigh fading, so that instead of several adjacent symbols being completely destroyed, many symbols are only slightly distorted. This allows precise reconstruction of a majority of them. A parallel approach has the additional advantage of spreading out the total signaling interval, thereby reducing the sensitivity of the system to delay spread.

Several systems have previously used orthogonal frequency

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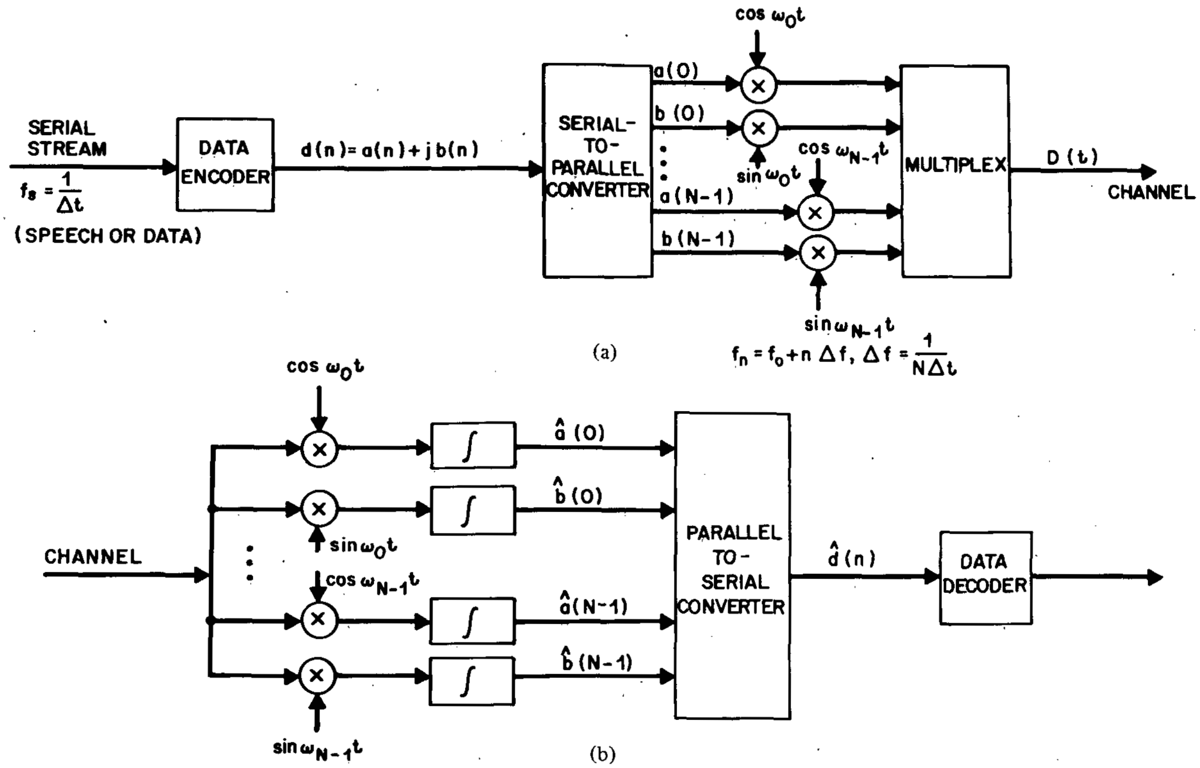

Fig. 1. Basic OFDM system. (a) Transmitter. (b) Receiver.

division multiplexing (OFDM) [3]-[8]. In particular, in the early 1960's, this technique was used in several high-frequency military systems (for example, KINEPLEX [9], ANDEFT [10], KATHRYN [11], [12]), where fast fading was not a problem. Similar modems have found applications in voice bandwidth data communications (for example, [13]) to alleviate the degradations caused by an impulsive noise environment.

In this paper, a parallel system which uses the OFDM technique is described. In Section II an analysis and simulation of the basic system, using pilot-based correction, is presented. In Section III a practical 7.5 kHz channel design is presented, along with a discussion of several of the problems encountered in reliably retrieving the pilots used in the data correction process. Several solutions to these problems are also presented.

This investigation is simplified by the assumption that the sole source of additive signal degradation is cochannel interference—thermal noise is assumed negligible. Man-made environmental noise, such as that caused by automotive ignitions or neon lights, is also ignored. However, these impairments are basically impulsive and their effect should be greatly reduced by this technique.

II. Basic Principles of Operation

A. Orthogonal Frequency Division Multiplexing (OFDM)

When an efficient use of bandwidth is not required, the most effective parallel system uses conventional frequency division multiplexing where the spectra of the different subchannels do not overlap. In such a system, there is sufficient guard space between adjacent subchannels to isolate them at the receiver using conventional filters. A much more efficient use of bandwidth can be obtained with a parallel system if the spectra of the individual subchannels are permitted to overlap. With the addition of coherent detection and the use of subcarrier tones separated by the reciprocal of the signaling element duration (orthogonal tones), independent separation of the multiplexed tones is possible.

tral shape is chosen so that interchannel interference does not occur; that is, the spectra of the individual subchannels are zero at the other subcarrier frequencies. The N serial data elements (spaced by $\Delta t = 1/f_s$ where f_s is the symbol rate) modulate N subcarrier frequencies, which are then frequency division multiplexed. The signaling interval T has been increased to $N\Delta t$, which makes the system less susceptible to delay spread impairments. In addition, the subcarrier frequencies are separated by multiples of $1/T$ so that, with no signal distortion in transmission, the coherent detection of a signal element in any one subchannel of the parallel system gives no output for a received element in any other subchannel. Using a two-dimensional digital modulation format, the data symbols $d(n)$ can be represented as $a(n) + jb(n)$ (where $a(n)$ and $b(n)$ are real sequences representing the in-phase and quadrature components, respectively) and the transmitted waveform can be represented as

$$D(t) = \sum_{n=0}^{N-1} \{a(n) \cos (\omega_n t) + b(n) \sin (\omega_n t)\} \tag{1}$$

where $f_n = f_0 + n\Delta f$ and $\Delta f = 1/N\Delta t$. This expression and the following analyses can be easily extended to include pulse shaping other than the assumed rectangular shape.

Theoretically, M-ary digital modulation schemes using OFDM can achieve a bandwidth efficiency, defined as bit rate per unit bandwidth, of $\log_2 M$ bits/s/Hz. This is easily shown as follows. Given that the symbol rate of the serial data stream is $1/\Delta t$, the bit rate for a corresponding M-ary system is $\log_2 M/\Delta t$. Each subchannel, however, transmits at a much lower rate, $\log_2 M/(N\Delta t)$. The total bandwidth of the OFDM system is

$$B = f_{N-1} - f_0 + 2\delta \tag{2}$$

where f_n is the nth subcarrier and δ is the one-sided bandwidth

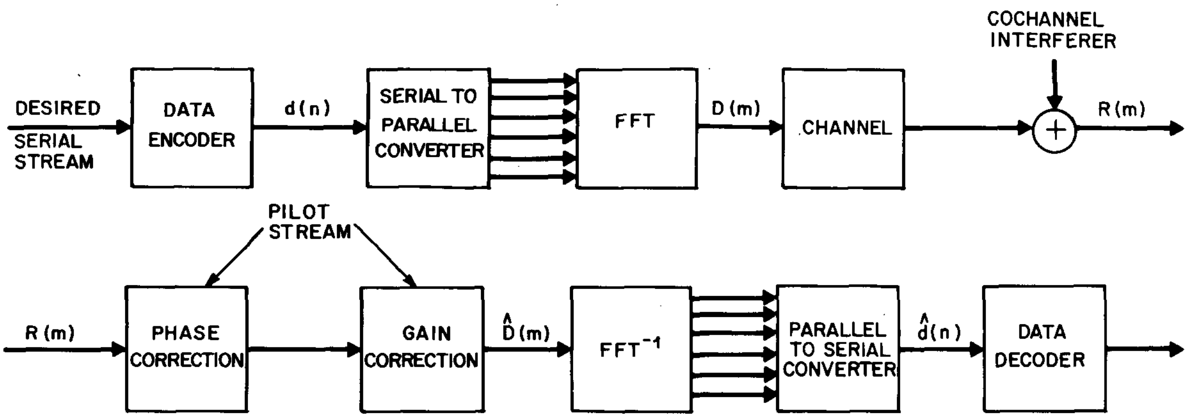

Fig. 2. OFDM system implemented with an FFT.

distance to the first null). The subcarriers are uniformly spaced so that $f_{N-1} - f_0 = (N-1)\Delta f$. Since $\Delta f = 1/N\Delta t$ due to the orthogonality constraint, $f_{N-1} - f_0 = (1 - (1/N))(1/\Delta t)$. Therefore, the bandwidth efficiency β becomes

$$\beta = \frac{\log_2 M}{\left(1 - \frac{1}{N}\right) + 2\delta\Delta t}. \quad (3)$$

For orthogonal frequency spacing and strictly band-limited spectra (bandwidth Δf) with $\delta = \frac{1}{2}\Delta f = 1/2N\Delta t$, $\beta = \log_2 M$ bits/s/Hz. In reality, however, the spectra overflow this minimum bandwidth by some factor α so that $\delta = (1 + \alpha)(1/2N\Delta t)$ and the efficiency (3) becomes

$$\beta = \frac{\log_2 M}{1 + \frac{\alpha}{N}} < \log_2 M. \quad (4)$$

To obtain the highest bandwidth efficiency in an OFDM system, N must be large and α must be small.

B. Implementation of OFDM Using the Discrete Fourier Transform

The principal objections to the use of parallel systems are the complexity of the equipment required to implement the system, and the possibility of severe mutual interference among subchannels when the transmission medium distorts the signal. The equipment complexity (filters, modulators, etc.) can be greatly reduced by eliminating any pulse shaping, and by using the discrete Fourier transform (DFT) to implement the modulation processes, as shown in [7], [8]. There it is shown that a multitone data signal is effectively the Fourier transform of the original data stream, and that a bank of coherent demodulators is effectively an inverse Fourier transform. This can be seen by writing (1) as

$$D(t) = \mathrm{Re}\left[\sum_{n=0}^{N-1} d(n)e^{-j\omega_n t}\right]. \quad (5)$$

Letting $t = m\Delta t$, the resulting sampled sequence $D(m)$ is seen as the real part of the DFT of the sequence $d(n)$.[1] The act of truncating the signal to the interval $(0, N\Delta t)$ imposes a $\sin x/x$ frequency response on each subchannel with zeros at multiples of $1/T$. This spectral shape has large sidelobes, and gives rise to significant interchannel interference in the presence of multipath. This point will be discussed in more detail in Section III.

Further reductions in complexity are possible by using the fast Fourier transform (FFT) algorithm to implement the DFT when N is large.

C. Pilot-Based Correction

If the transmission channel is distortionless, the orthogonality of the subcarriers allows the transmitted signals to be received without error at the receiver. Consider the system in Fig. 2 with the block of data represented by the sequence of N complex numbers $\{d(0), d(1), \cdots, d(N-1)\}$. These complex numbers are generated by the data encoder from a binary data sequence. A DFT is performed on this block of data, giving the transmitted symbols[2]

$$D(m) = \mathrm{DFT}\{d(n)\} = \sum_{n=0}^{N-1} d(n)\, e^{-j(2\pi/N)nm}. \quad (6)$$

Notice that this is a sampled version of (5) where the complex notation has been retained. All future analyses will be done in the complex domain. Under the assumption of a distortionless channel, the received data sequence (the output of the inverse DFT) will be exactly the transmitted sequence due to the orthogonality of the subcarrier tones (exponentials).

If the transmission channel distorts the signal, this orthogonality is impaired. In a flat Rayleigh fading environment (i.e., the environment is not frequency selective), the effects of the Rayleigh channel can be represented as a multiplicative noise process on the transmitted signal. This multiplicative process is characterized by a complex fading envelope with samples $Z(m) = A(m)e^{j\theta(m)}$ where the $A(m)$ are samples from a Rayleigh distribution and the $\theta(m)$ are samples from a uniform distribution [1]. These samples multiply the sequence of (6) to give

$$R(m) = Z(m)D(m). \quad (7)$$

The output data sequence $\hat{d}(k)$ is the inverse DFT of (7),

$$\hat{d}(k) = \frac{1}{N}\sum_{m=0}^{N-1} Z(m)D(m)\,e^{j(2\pi/N)km}$$

$$= \sum_{n=0}^{N-1} d(n)\left[\frac{1}{N}\sum_{m=0}^{N-1} Z(m)\,e^{j(2\pi/N)m(k-n)}\right]$$

$$= \sum_{n=0}^{N-1} d(n)z(k-n) \quad (8)$$

[1] It is convenient in this paper to think of $d(n)$ as being in the frequency domain and $D(m)$ as being in the time domain, contrary to the usual

[2] Throughout this paper, all indexes will be assumed to belong to the set {0,

where $z(n)$ is the inverse DFT of $Z(m)$. It can be seen from (8) that there is a complex-weighted averaging of the samples of the complex fading envelope. If $Z(m) = 1$ for all m (the distortionless channel), $z(k - n)$ is simply the Kronecker delta function δ_{kn} and $\hat{d}(k) = d(k)$. In the presence of fading, $z(k - n) \neq \delta_{kn}$ and

$$\hat{d}(k) = d(k)z(0) + \sum_{\substack{n=0 \\ n \neq k}}^{N-1} d(n)z(k - n). \tag{9}$$

The second term on the right represents the interchannel (intersymbol) interference caused by the loss of orthogonality. Without correction for the fading, the output sequence is corrupted by intersymbol interference *even if there is no cochannel interferer.*

Pilot-based correction provides an amplitude and phase reference which can be used to counteract the unwanted effects of multipath propagation. Similar considerations have been analyzed for single-sideband mobile radio systems [14], [15]. Coherent detection, by definition, requires a phase reference; however, gain correction is also needed in an OFDM system in a fading environment to remove intersymbol interference. If phase and gain correction is employed in the absence of cochannel interference, it is easily shown, in (9), that $\hat{d}(k) = d(k)$.

In a cellular mobile system, the dominant transmission impairment often comes from other users using the same carrier frequency. It is assumed here that the desired signal and a *single* undesired cochannel interferer are received simultaneously, and that both are digital signals modulated by different data sequences with identical signaling rates. It is also assumed that they are subject to mutually independent Rayleigh fading.

When a cochannel interferer is present in the received signal, it is not advantageous to do unlimited gain correction, due to the possibility of enhancing the energy of the interferer during deep fades of the desired signal. The detrimental effects of unlimited gain correction in the presence of a cochannel interferer can be seen as follows. Let $D(m)$ be the desired transmitted signal sequence and let $I(m)$ be the corresponding cochannel interferer sequence. With $Z_d(m) = A_d(m)e^{j\theta_d(m)}$ and $Z_i(m) = A_i(m)e^{j\theta_I(m)}$ the desired and interferer complex fading sequences, respectively, the sequence present at the receiver can be represented as

$$R(m) = Z_d(m)D(m) + \sqrt{\gamma}Z_i(m)I(m) \tag{10}$$

where γ is the interference-to-signal power ratio (SIR^{-1}). $R(m)$ is corrected by a complex correction sequence $Z_c(m) = Z_p(m)$, the complex pilot fading envelope, giving

$$\hat{D}(m) = \frac{R(m)}{Z_c(m)} = \frac{Z_d(m)}{Z_p(m)} D(m) + \sqrt{\gamma}\frac{Z_i(m)}{Z_p(m)} I(m). \tag{11}$$

Taking the inverse DFT of (11), the received data sequence becomes

$$\hat{d}(k) = \sum_{n=0}^{N-1} d(n)z(k - n) + \sqrt{\gamma} \sum_{m=0}^{N-1} \frac{1}{N} \frac{Z_i(m)}{Z_p(m)} \cdot I(m)e^{j(2\pi/N)mk} \tag{12}$$

where

$$z(k - n) = \frac{1}{N} \sum^{N-1} \frac{Z_d(m)}{\cdots} e^{j(2\pi/N)m(k-n)}.$$

If unlimited gain and phase correction is used [i.e., $Z_p(m) = Z_d(m)$], $z(k - n) = \delta_{kn}$, there is no intersymbol interference, and (12) becomes

$$\hat{d}(k) = d(k) + \sqrt{\gamma}\, \frac{1}{N} \sum_{m=0}^{N-1} I(m) \frac{Z_i(m)}{Z_d(m)}\, e^{j(2\pi/N)mk}. \tag{13}$$

The only distortion is caused by the cochannel interferer. However, since $Z_i(m)$ and $Z_d(m)$ are statistically independent, the desired signal may be in a fade when the interferer is not, and unlimited gain correction may boost the interferer average energy above that of the desired signal.

One alternative to unlimited gain and phase correction is to have a limit on the gain correction, so as not to follow the desired signal into deep fades [1]. This is done at the expense of increased intersymbol interference, due to imperfect correction of the desired signal. In this situation, the correction signal is of the form

$$Z_c(m) = \begin{cases} A_d(m)e^{j\theta_d(m)} & \text{when } A_d(m) > \epsilon \\ \epsilon e^{j\theta_d(m)} & \text{when } A_d(m) \leq \epsilon \end{cases} \tag{14}$$

where ϵ is the gain limit and is defined relative to the average value of the local field strength. Therefore, in (12), $z(k - n) \neq \delta_{kn}$, resulting in intersymbol interference. Consequently, there is a tradeoff between increasing the intersymbol interference and boosting the cochannel interference energy.

Another alternative is to develop an optimum gain correction factor which takes both distortion effects into account. An optimum gain correction factor $F(m)$ has been derived by minimizing the mean-square distortion between $D(m)$ and $\hat{D}(m)$. The derivation of $F(m)$ has been omitted for the sake of brevity. The correction sequence then becomes

$$Z_c(m) = Z_p(m)F(m)$$

$$= Z_d(m)\left[1 + \gamma\left(\frac{A_i(m)}{A_d(m)}\right)^2\right]. \tag{15}$$

This correction procedure would be more difficult to implement than the gain limiting procedure described above.

In addition to the impairments caused by intersymbol and cochannel interference, frequency-selective fading may also be present. This phenomenon causes a decorrelation of the received signal envelopes at different frequencies, lessening the effectiveness of the pilot-correction procedure, since a data point which is being corrected may be decorrelated from the corresponding pilot complex fading envelope.

Finally, one of the major advantages of the OFDM technique is its ability to "average" out impairments, making the bursty Rayleigh channel appear much less bursty. The extent to which this averaging approaches a Gaussian channel depends on the correlation between samples of the complex fading envelope. It can be seen that as N increases, more independent fades are averaged. This enables burst errors to be randomized and thereby aids in bit error correction. This property will be more evident in the simulation results, which indicate that the curves for the bit error rate fall between the linear Rayleigh channel curves and the exponential Gaussian channel curves. For large N and high vehicle speeds, the bit error curve approaches that for a Gaussian channel.

D. Distortion Analyses

Several mechanisms contribute to the overall distortion of the desired signal. In this section, emphasis is on the contributions due to gain limiting, evident in increased intersymbol tion resulting from decorrelation of the pilot due to frequency-selective fading or due to interference on the pilot is considered in Section II-F.

First, consider the case of gain-limited correction, where the amplitude correction is bounded to follow fades only as deep as ϵ. Assume that the random processes which produce the random sequences are ergodic, thereby permitting the equivalence of time and ensemble averages. The pilot complex fading envelope at a particular instant in time is $Z_p(m) = Z_d(m)$ and the correction sequence is

$$Z_c(m) = \max(A_d(m), \epsilon) e^{j\theta_d(m)}. \tag{16}$$

The corrected output samples become

$$\hat{D}(m) = \frac{R(m)}{Z_c(m)} = D(m)\frac{Z_d(m)}{Z_c(m)} + \sqrt{\gamma} I(m)\frac{Z_i(m)}{Z_c(m)}$$
$$= D(m) A_d(m) \min\left(\frac{1}{A_d(m)}, \frac{1}{\epsilon}\right)$$
$$+ \sqrt{\gamma} I(m) A_i(m) e^{j(\theta_i(m) - \theta_d(m))}$$
$$\cdot \min\left(\frac{1}{A_d(m)}, \frac{1}{\epsilon}\right). \tag{17}$$

The signal-to-distortion ratio (SDR) can be defined as in [14],

$$\mathrm{SDR} = \frac{\overline{|D(m)|^2}}{\overline{|\hat{D}(m) - D(m)|^2}} \tag{18}$$

where $\overline{X}$ denotes a time average of X. Assuming $\overline{|D(m)|^2} = \overline{|I(m)|^2} = 1$, the denominator in (18) reduces to

$$\overline{|\hat{D}(m) - D(m)|^2}$$
$$= \overline{A_d^2(m) \min^2\left(\frac{1}{A_d(m)}, \frac{1}{\epsilon}\right)}$$
$$+ \gamma \overline{A_i^2(m) \min^2\left(\frac{1}{A_d(m)}, \frac{1}{\epsilon}\right)}. \tag{19}$$

Assuming time averages can be replaced by expected values and assuming $A_d(m)$ and $A_i(m)$ are statistically independent and Rayleigh distributed, (18) becomes, after some manipulations,

$$\mathrm{SDR} = \left\{\left(\frac{\mathrm{SIR}+1}{\mathrm{SIR}}\right)\frac{1}{\epsilon^2}\left[1 - e^{-\epsilon^2}\right] + 1 - \frac{\sqrt{\pi}}{\epsilon}\,\mathrm{erf}(\epsilon) + \frac{E_1(\epsilon^2)}{\mathrm{SIR}}\right\}^{-1} \tag{20}$$

where $E_1(x) = -[\Psi + \ln(x) + (\Sigma_{n=1}^{\infty} (-1)^n x^n / nn!)]$ and Ψ is Euler's constant (=0.57721566 ⋯). The SDR in (20) is plotted in Fig. 3 for several values of SIR. Obviously, if SIR = ∞ (no cochannel interference), the results reduce to that in [14] and no gain limit should be used. However, for SIR < ∞ the curves clearly indicate the tradeoff between intersymbol interference, caused by gain limiting, and boosting of the cochannel interference average energy, caused by unlimited gain correction. If unlimited gain correction is used, SDR = −∞, indicating that the interferer completely distorts the desired signal. Notice, there is a definite maximum which is fairly flat. Although SDR as defined here is an analog transmission quality measure, it does indicate the degree to which intersymbol interference, caused by imperfect gain correction, and cochannel interference are problems. These factors are critically important in digital transmission. The SDR also clearly shows the tradeoffs which must be made when choosing the appropriate gain limit. This, in turn, directly affects the bit error rate (BER), as shown in the next section.

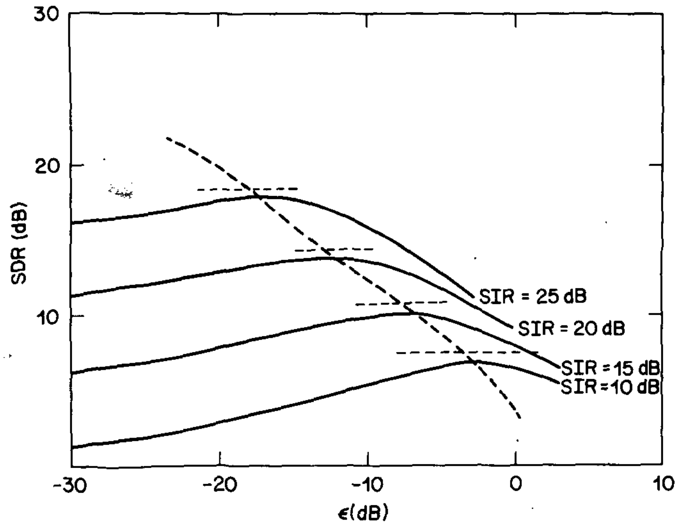

Fig. 3. Signal-to-distortion ratio for a flat Rayleigh fading environment when gain-limited correction is used.

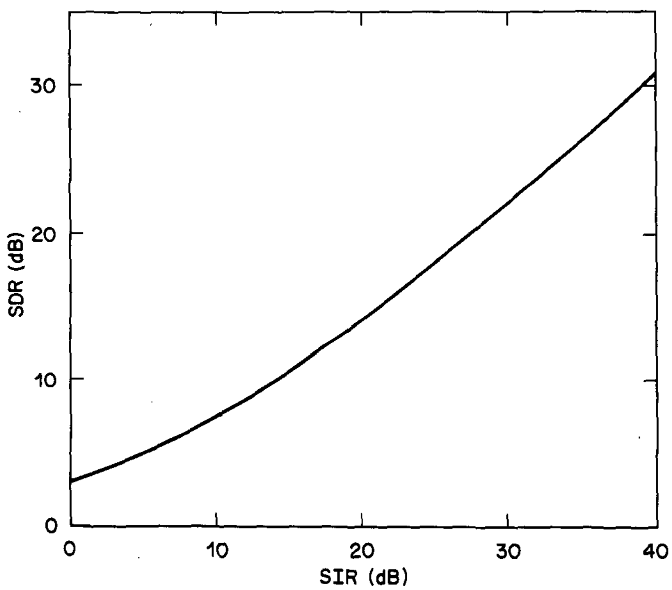

Fig. 4. Signal-to-distortion ratio for a flat Rayleigh fading environment when the optimum gain correction factor is used.

Similar results can be derived for optimum gain correction, as in (15), and the SDR can be shown to be

$$\mathrm{SDR} = \left[\left(\frac{1}{\mathrm{SIR}-1}\right)\left[\frac{\mathrm{SIR}}{\mathrm{SIR}-1}\ln(\mathrm{SIR}) - 1\right]\right]^{-1} \tag{21}$$

which is plotted versus SIR in Fig. 4. This curve indicates the *best* performance for a given SIR. Notice, by comparing Figs. 3 and 4, that using gain-limited correction does not sacrifice much if the gain limit is in the vicinity of the maximum. Both of these results could be used as an aid in determining the appropriate level for gain limiting for a given SIR.

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