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Adaptive Modulation Without Amplitude Estimation

Mr.Vijayananth.S[1]
Department of Telecommunication Engineering.
Asst professor (O.G)

Surendhar K N[2] Abinaya K[3]
Department of Telecommunication Engineering.
SRM UNIVERSITY
Chennai -603203

ABSTRACT

Channel estimation at the receiver side is essential to adaptive modulation schemes, prohibiting low complexity systems from using variable rate and/or variable power transmissions. Towards providing a solution to this problem, we introduce a variable-rate (*VR*) *M*-PSK modulation scheme, for communications over fading channels, in the absence of channel gain estimation at the receiver. The choice of the constellation size is based on the signal-plus-noise (*S+N*) sampling value rather than on the signal-to-noise ratio (*S/N*). It is analytically shown that *S+N* can serve as an attractive simpler criterion, alternative to *S/N*, for determining the modulation order in *VR* systems. In this way, low complexity transceivers can use *VR* transmissions in order to increase their spectral efficiency under an error performance constraint. As an application, we utilize the proposed *VR* modulation scheme in equal gain combining (EGC) diversity receivers.

Index Terms**—**Adaptive modulation, equal gain combining, fading channels..

1 INTRODUCTION

A common technique for dealing with fading in wireless communications, is transmission or reception diversity , If a feedback link is available, the fading can be mitigated by allowing the receiver to monitor the channel conditions and request compensatory changes in certain parameters of the transmitted signal. This technique is called adaptive transmission. The basic concept is the real-time balancing of the link budget through adaptive variation of the transmitted power level, symbol transmission rate, constellation size, coding rate/scheme, or any combination of these parameters.

In many practical applications, the *VR* communication systems, we adapt the transmission parameters according to the instantaneous channel condition system with no channel gain cannot benefit the adaptation technique.

2 PROPOSED SYSTEM

We propose a rate adaptation technique, which is based on the signal-plus-noise (*S+N*) samples, instead of the instantaneous *S/N*. They require no channel gain estimation. The *S+N* technique has been applied to selection diversity communication systems.

We introduce an *S+N*-based *VR* *M*-PSK modulation scheme that does not require any channel gain estimation. The *M*-PSK modulation schemes have some important advantages over QAM schemes

ADVANTAGES

- Adjacent channel interference and Inter symbol interference is reduced.
- Feedback channel is provided to avoid the channel estimation.
- Adaptive modulation is used.
- Data rate is increased and the data are transferred in a proper way with the total bandwidth.

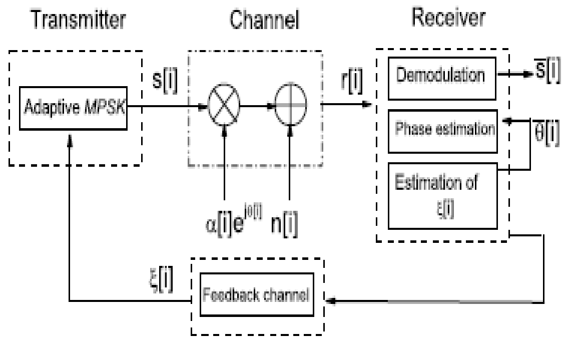
3. VARIABLE-RATE TECHNIQUES

In variable-rate modulation the data rate is varied relative to the channel gain. This process is done by fixing the modulation scheme or by fixing the particular modulation scheme, then change the constellation size or symbol rate of the signal;

Symbol rate variation is difficult to implement in practice since a varying signal bandwidth is impractical and complicates bandwidth sharing. They are currently used in systems such as EGPRS for data transmission in GSM based cellular systems varies between 8PSK and GMSK modulation thus for GPRS data transmission. In IS-136 TDMA cellular systems can use 4, 8, and 16level PSK modulation.

3.1 MODE OF OPERATION

We introduce a constant power, *VR M*-PSK modulation scheme, in which the decision on the modulation order is based on the *S*+*N* values



The *VR* M-PSK modulation scheme.

An ideal coherent phase detection is available at the receiver at time instant, while no channel gain estimation is necessary. The receiver estimates the *S*+*N* samples at time, the decision on the modulation order is sent back to the transmitter, transmitting bits through a feedback channel that does not introduce any errors. The elimination of the estimation errors can be assured by increasing its delay time and using an ARQ transmission protocol. Similarly to the *S/N* based adaptive modulation schemes, the estimation of the modulation order for the *S*+*N* based schemes; require a closed-form formula that relates, the *S*+*N* with the error probability. Because of the fact that such formulas are not available, *S/N* related formulas will be used instead. Moreover, formulas relating *S*+*N* with the SER would be useless for the adaptation scheme, since the instantaneous noise components between two time instances are different. Therefore, for the case of the *M*-PSK we can use the approximation for the SER, i.e.

$$P_M = erfc\left(\sqrt{\gamma} sin \frac{\pi}{M}\right)$$

Therefore, the selected order will

$$M = M_j, if \begin{cases} M_j \leq M < M_{j+1} \\ \gamma_j \leq \frac{\left(a\sqrt{E_s}+n_I\right)^2}{N_0} < \gamma_j \end{cases}$$

The proposed scheme gives the ability to communications systems with no channel gain estimation capabilities to adapt high transmission rate, so that their spectral efficiency can be increased.

4 PERFORMANCE EVALUATION

4.1 SPECTRAL EFFICIENCY

The normalized spectral efficiency of the *S*+*N* based *VR M*-PSK scheme is obtained as

$$S_{S+N} = \frac{R}{B} = \sum_{j=1}^{N} log_2(M_j) \mathrm{Pr}\{\gamma_j \leq \varepsilon < \gamma_{j+1}\}$$

Where $f_z(z)$ is the pdf of the random variable $z = a\sqrt{E_S} + n_I$, which is the sum of a Rayleigh and a Gaussian random variable, and is given by

$$\mathrm{Pr}\{\gamma_j \leq \varepsilon < \gamma_{j+1}\} = \mathrm{Pr}\{\sqrt{\gamma_j N_0} \leq a\sqrt{E_s} + n_I < \sqrt{\gamma_{j+1} N_0}\}$$

$$\mathrm{Pr}\{\gamma_j \leq \varepsilon < \gamma_{j+1}\} = \int_{\gamma_j}^{\gamma_{j+1}} f_z(z) dz$$

Thus

$$S_{S+N} = \sum_{j=1}^{N} log_2(M_j)[J(\gamma_{j+1}) - J(\gamma_j)]$$

The spectral efficiency of the *S/N* based *VR M*-PSK will be

$$S = \sum_{j=1}^{N} log_2(M_j)(e^{-\frac{\gamma_j}{J}} - e^{-\frac{\gamma_{j+1}}{J}})$$

4.2 SYMBOL ERROR PROBABILITY

Regarding the SER of a *S/N* based *VR M*-PSK scheme, it can be obtained by averaging over the SER for *M*-PSK in AWGN

$$P_s = \sum_i \int_{\gamma_j}^{\gamma_{j+1}} P_{AWGN}(M_j, \gamma) f_Y(\gamma) \gamma$$

For the case of the *S*+*N*, however, a similar approach cannot be followed, since a formula for the SER as a function of is required.

4.3APPLICATION OF EQUAL-GAIN COMBINING RECEIVERS

Consider a multichannel diversity reception system with L branches operating in a discrete-time channel, in which the receiver employs symbol-by-symbol detection. The signal received over the kth diversity branch, at the time instant i can be expressed as

$$u[i] = \sum_{k=1}^{L} w_k[i] u_k[i]$$

where $a_k[i]$ is the random magnitude, $\vartheta_k[i]$ is the random phase of the kth diversity branch gain and

$nk[i]$, represents the additive noise with $E\{nk[i]*nk[i]\} = Nk = 2\sigma 2 = N0$.Assuming that the random phase $\vartheta k[i]$ are known at the receiver, the received signals are co-phased and transferred to baseband so that the signal at the kth branch will be

$$r_k[i] = R\{|a_k[i]e^{j\theta_k[i]}s[i] + n_k[i]|\} = R\{R_k[i]\}, k = 1 \dots L$$

At the combination stage the signals $uk[i]$ are weighted and a summed to produce the decision variable

$$u[i] = \sum_{k=1}^{1} w_k[i]u_k[i]$$

Where $wk[i]$ is the weight of the kth branch. Then, applying Maximum Likelihood Detection (MLD), the combiner's output is compared with all the known possible transmitted symbols in order to extract the decision metric.

The coherent equal-gain-combining (EGC) receiver co phases and equally weights each branch before combining and therefore does not require estimation of the channel (path) fading amplitudes, but only knowledge of the channel phase, in order the demodulator to undo the random phase shifts introduced on the diversity channels.

4.4 SYSTEM MODEL

Consider a discrete-time channel, assuming that the fading amplitude $[i]$ follows a Rayleigh distribution with probability density function (pdf),

$$f_a(a) = \frac{2a}{\Omega} e^{-\frac{a^2}{\Omega}}$$

Where $\Omega = E\{a^2\}$

The received signal is:

$$r[i] = a[i]e^{j\theta[i]}\, s[i] + n[i]$$

Where r[i] = received signal,
a[i]= fading amplitude
Θ[i]= phase of the signal
s[i]= transmitted signal
n[i]= noise component
The received S/N is

$$\gamma[i] = E_S \frac{a^2[i]}{BN_0}$$

Where Es= average signal energy
B= bandwidth of the signal
No= variance of the signal

4.5 RELATION BETWEEN S/N AND S+N

The probability that both S/N and S+N determine the same constellation size, i.e.,

$$II_1 = P_r\left\{r_j \le E_s \frac{a^2}{No} < r_{j+1} \cap r_j \le \frac{(a\sqrt{E_s} + n_I)^2}{N_0} < r_{j+1}\right\}$$

Next deriving a closed-form solution for Π1.
By setting $y = a\sqrt{ES}$, can be equivalently rewritten as

$$II_1 = P_r\{\sqrt{N_0 r_j} \le y < \sqrt{N_0 r_{j+1}} \cup \sqrt{N_0 r_j} - y \le n_I < \sqrt{N_0 r_{j+1}} - y\}$$

$$\int_{\sqrt{N_0 r_j}}^{\sqrt{N_0 r_{j+1}}} P_r\{\sqrt{N_0 r_j} - y \le n_I < \sqrt{N_0 r_{j+1}} - y\}$$

$$f_y(y) = \frac{1}{\sqrt{E_s}} f_a \frac{y}{\sqrt{E_s}} = \frac{2y}{\Omega E_s} e^{-\frac{y^2}{\Omega E_s}}$$

Moreover, the probability

$$P_r\{\sqrt{N_0 r_j} - y \le n_I < \sqrt{N_0 r_{j+1}} - y\}$$

Involved in is the probability that the random Gaussian variable, nI, lies in a specific interval and it is directly related to the complementary error function erfc(·). Using the cumulative distribution function (cdf) of the zero-mean Gaussian random variable, x,

$$F_X(x) = \int_{-\infty}^{x} \frac{1}{\sqrt{2\pi\sigma}} e^{-\frac{z^2}{2\sigma^2}}\, dz$$

and the definition of the erfc(·),

$$\mathrm{erfc}(x) = \frac{2}{\sqrt{\pi}} \int_{x}^{\infty} e^{-z^2}\, dz$$

the probability can be calculated after some trivial manipulations as

$$P_r\{\sqrt{N_0 r_j} - y \le n_I < \sqrt{N_0 r_{j+1}} - y\}$$

$$\frac{1}{2}\left[erfc\left(\frac{\sqrt{N_0 r_j} - y}{N_0}\right) - erfc\left(\frac{\sqrt{N_0 r_{j+1}} - y}{N_0}\right)\right]$$

Therefore, the probability, Π1, can be written as

$$II_1 = \frac{1}{2}\int_{\sqrt{N_0 r_j}}^{\sqrt{N_0 r_{j+1}}} \left[erfc\left(\frac{\sqrt{N_0 r_j} - y}{N_0}\right) - erfc\left(\frac{\sqrt{N_0 r_{j+1}} - y}{N_0}\right)\right] F_y(y)dy.$$

Now, in order to derive a closed-form expression for we have to solve integrals of the form

$$I = \mathrm{D}\int_{\mathrm{t}}^{\mathrm{y}} \mathrm{erfc}\left(\frac{\mathrm{E}-\mathrm{x}}{\sqrt{2}\mathrm{c}}\right)\mathrm{x}\mathrm{e}^{-\mathrm{D}\mathrm{x}^2}\,\mathrm{dx}$$

Finally, combining a closed-form solution to Π1, is derived as

$$\begin{aligned} &II_1 \\ &= g\left(\sqrt{N_0 r_j}, \sqrt{N_0 r_{j+1}}\sqrt{\frac{N_0}{2}}, \frac{1}{E_s\Omega}, \sqrt{N_0 r_j}\right) \\ &- g\left(\sqrt{N_0 r_j}, \sqrt{N_0 r_{j+1}}\sqrt{\frac{N_0}{2}}, \frac{1}{E_s\Omega}, \sqrt{N_0 r_{j+1}}\right) \end{aligned}$$

As a general comment regarding criterion 1, one can observe that Π1 may not be always indicative of the efficiency of *S+N* to replace *S/N*, since it involves also the probability that *S/N* leads to a modulation order, *M*. In other words, if the probability that *S/N* leads to the modulation order, *Mj*, is too low, then Π1 will be also too low, independently of whether *S+N* led to the same order, *Mj* or not. Towards this, a supplemental criterion for the efficiency of *S+N* is presented in the following.

5 ADAPTIVE MODULATION

Adaptive modulation and coding enables robust and spectrally-efficient transmission over time-varying channels. The basic premise is to estimate the channel at the receiver and feed this estimate back to the transmitter, so that the transmission scheme can be adapted relative to the channel characteristics. Modulation and coding techniques that do not adapt to fading conditions require a fixed link margin to maintain acceptable performance when the channel quality is poor. Thus, these systems are effectively designed for the worst-case channel conditions. Since Rayleigh fading can cause a signal power loss of up to 30 dB, designing for the worst case channel conditions can result in very inefficient utilization of the channel. Adapting to the channel fading can increase average throughput, reduce required transmit power, or reduce average probability of bit error by taking advantage of favorable channel conditions to send at higher data rates or lower power, and reducing the data rate or increasing power as the channel degrades. In the optimal adaptive transmission scheme that achieves the Shannon capacity of a flat fading channel was derived. In this chapter we describe more practical adaptive modulation and coding techniques to maximize average spectral efficiency while maintaining a given average or instantaneous bit error probability. The same basic premise can be applied to MIMO channels, frequency-selective fading channels with equalization, OFDM, or CDMA, and cellular systems.

Adaptive modulation requires a feedback path between the transmitter and receiver, which may not be feasible for some systems. Moreover, if the channel is changing faster than it can be reliably estimated and fed back to the transmitter, adaptive techniques will perform poorly. Many wireless channels exhibit variations on different timescales, for example multipath fading, which can change very quickly, and shadowing, which changes more slowly. Often only the slow variations can be tracked and adapted to, in which case flat fading mitigation is needed to address the effects of multipath. Hardware constraints may dictate how often the transmitter can change its rate and/or power, and this may limit the performance gains possible with adaptive modulation. Finally, adaptive modulation typically varies the rate of data transmission relative to channel conditions. We will see that average spectral efficiency of adaptive modulation under an average power constraint is maximized by setting the data rate to be small or zero in poor channel conditions. However, with this scheme the quality of fixed-rate applications with hard delay constraints such as voice or video may be significantly compromised. Thus, in delay-constrained applications the adaptive modulation should be optimized to minimize outage probability for a fixed data rate.

By adapting channel fading:

- ➢ Increase average data rate
- ➢ Transmit power reduces
- ➢ Decreases average probability of bit error

5.1 ADAPTIVE TRANSMISSION SYSTEM

- Model is the flat fading channel as a discrete-time channel where each channel use corresponds to one symbol time Ts.
- The channel has stationary and ergodic time varying gain g[i] that follows a given distribution p(g) and AWGN n[i], with power spectral density N0/2.
- Assumptions:
    - ➢ Linear Modulation where the adaptation takes place at multiple symbol rate Rs=1/Ts.
    - ➢ Ideal Nyquist data pulses (sinc[t/Ts]), so signal bandwidth B=1/Ts.
- Denotation:
    - ➢ S=Average transmit signal power.
    - ➢ B=1/Ts, received signal power.
    - ➢ G=Average channel gain.
- The instantaneous received SNR is then γ[i]=Sg[i]/(N0B), 0≤γ≤∞, and its expected value γ=Sg/N0B.

- In adaptive modulation we estimate power gain or received SNR at time I and adapt the modulation and coding parameter accordingly.
- Parameters:
  - Data rate: R[i]
  - Transmit power: S[i]
  - Coding parameters: C[i]

M-ary Modulation data rate:

R[i]=log2M[i]/Ts=Blog2M[i] bps.

Spectralefficiency: R[i]/B=log2M[i] bps/Hz.

SNR estimate: $\hat{\gamma}[i]=\hat{S}g[i]/N0B$

Adaptive transmit power: $\hat{S}(\gamma[i])=S[i]$

Received power: $\gamma[i]S(\hat{\hat{\hat{\gamma}}}[i])S$

Data rate of Modulation: $R(\hat{\gamma}[i])=R[i]$

Coding parameters: $C(\hat{\gamma}[i])=C[i]$

We will omit [i] with respect to γ, S(γ), R(γ) and C(γ).

Channel information at transmitter allows it to adapt its transmission scheme relative to channel variation. Adaptive consider delay and error also. Assume feedback path does not introduce any errors. At low speed, shadowing is constant and multipath fading is slow so that it can be estimated and feedback to transmitter with an estimation error and delay that does not degrade performance. At high speeds the system cannot estimate effectively so shadowing is adapted.

5.1.1 Adjacent Channel Interference

Adjacent Channel Interference is caused by extraneous power from a signal in an adjacent channel. It is caused by inadequate filtering, improper tuning or poor frequency control.

5.1.2 Intersymbol Interference

Intersymbol Interference (**ISI**) is a form of distortion of a signal in which one symbol interferes with subsequent symbols. This is an unwanted phenomenon as the previous symbols have similar effect as noise, thus making the communication less reliable. ISI is usually caused by multipath propagation or the inherent non-linear frequency response of a channel causing successive symbols to "blur" together. The presence of ISI in the system introduces errors in the decision device at the receiver output. Therefore, in the design of the transmitting and receiving filters, the objective is to minimize the effects of ISI, and thereby deliver the digital data to its destination with the smallest error rate possible. Ways to fight intersymbol interference include adaptive equalization and error correcting codes.

5.2 DIVERSITY SCHEMES

Diversity improves transmission performance by using faded version of transmitted signal.
It is used because:

- The degradation of the transmission quality due to channel fading cannot simply be overcome by increasing the transmitted signal power. This is because, even with high transmitted power, when the channel is in deep fading, the instantaneously received SNR per bit can still be very low, resulting in a high probability of transmission error during the deep fading period.
- In wireless communications, the power available on the reverse link is severely limited by the battery capacity in user mobile units. With diversity, the required transmitted power can greatly be reduced. Wireless systems are mostly interference limited and once again, mitigation of the channel fading by diversity reception can translate into improved interference tolerance which, in turn, means greater ability to support additional users and therefore higher system capacity.

5.3DIVERSITY COMBINING TECHNIQUES

- Selection Combining(SC)
- Equal Gain Combining(EGC)
- Maximal Ratio Combining(MRC)

5.3.1 Selection Diversity

Among many branches of signal the received branch with highest instantaneous SNR will be fed to detector circuit.

$$g_k = \begin{cases} 1, if\ k = l \\ 0, if\ k \neq l \end{cases}$$

$$\gamma_{out} = max \frac{\alpha_{k/2}}{\bar{n}^2}$$

5.3.2 Maximal Ratio Combining

Each replica of signals are weighted according to SNR and co phased and distortions are cancelled out. It is assumed and it is like taking an average of weight and hence noise is reduced.

$$g_k = \frac{\alpha_k e^{-j\emptyset_k}}{\overline{n_k}^2}, for\ k = 1,2,\dots,l$$

$$\gamma_{out} = \sum_{k=1}^{L} \frac{\alpha_k^2/2}{\overline{n_k^2}} = \frac{\frac{1}{2}\sum_{k=1}^{L} \alpha_k^2}{\bar{n}^2}$$

5.3.3 Equal Gain Combining

Signals are co phased and are equally weighted by their amplitude. Branching weights are all set to unity.

$$g_k = e^{-j\,\emptyset_{ki}}, for\ k = 1,2,\ldots\ldots,L$$

$$\gamma_{out} = \frac{\frac{1}{2}\left(\sum_{k=1}^{L} \alpha_k\right)^2}{\overline{L_n^2}}$$

presented in the following

6 SIMULATION RESULTS

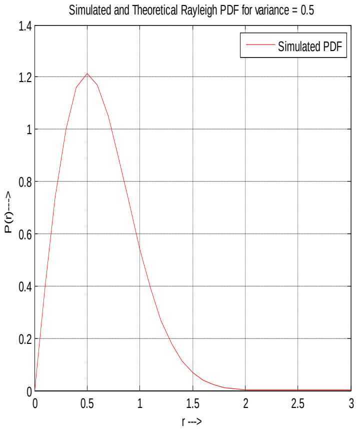
The evaluation of the probabilities, Π1 and Π2 is very important, since they imply the efficiency of S+N to estimate the quality of the instantaneous received signal.

For both these probabilities, a general conclusion derived from these figures is that if the S/N based VR M-PSK uses a specific modulation order, the probability that the S+N based VR M-PSK uses the same order is relatively high. Moreover, as expected, the efficiency of the S+N improves as the average S/N increases, since the instantaneous noise components become small, so that the divergence between $(a\sqrt{\ ES} + n)2$ and $a2ES$ decreases. It is seen that both schemes achieve the same spectral efficiency, while the difference in their SER is not significant. In general, it can be observed that the S+N can serve as an attractive criterion for determining the channel quality, when channel gain estimation is not available.

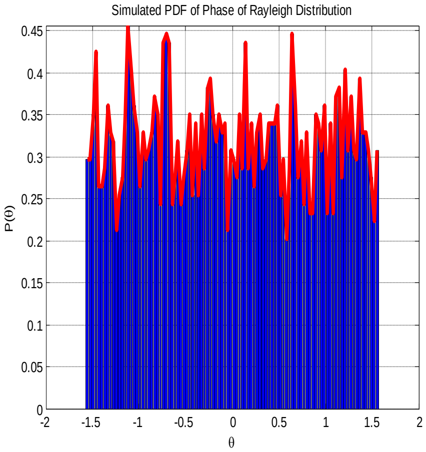
The explanation for the efficiency of the S+N criterion is as follows:

- In the conventional VR M-PSK, the decision on the modulation order is based on the instantaneous S/N, i.e. on $a2ES/N0$, which means that the instantaneous noise power is ignored, since only the average noise power is taken into account.
- On the contrary, the S+N samples, $(a\sqrt{\ ES}+n)2$ include the instantaneous noise component and therefore an indirect estimation of the instantaneous noise power is obtained. Similar results can be obtained also for the EGC employing VR M-PSK. The spectral efficiency is increased compared to a fixed rate M-PSK scheme, while the SER is maintained under the required levels, whenever this is possible.

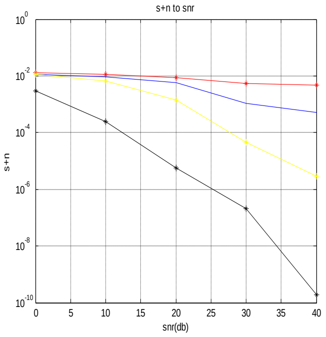
6.1 FADING AMPLITUDE



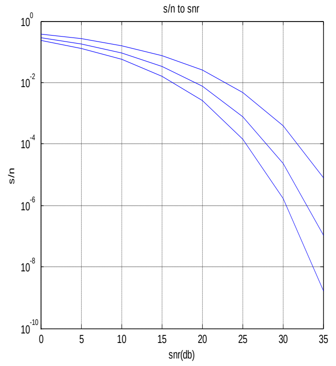
6.2 FADING PHASE



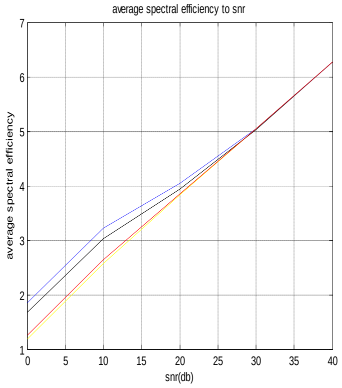
6.3 THE PROBABILITY, Π1, FOR A VR M-PSK MODULATION SCHEME. THEORETICAL (SOLID LINES) AND SIMULATION (SYMBOLS) RESULTS



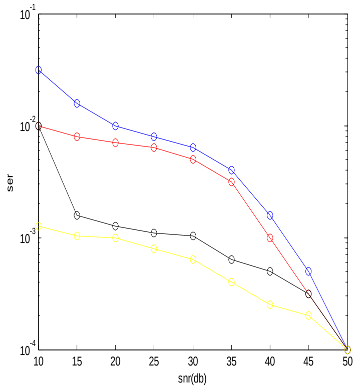
6.4 THE PROBABILITY, Π2, FOR A VR M-PSK MODULATION SCHEME. THEORETICAL (SOLID LINES) AND SIMULATION (SYMBOLS) RESULTS



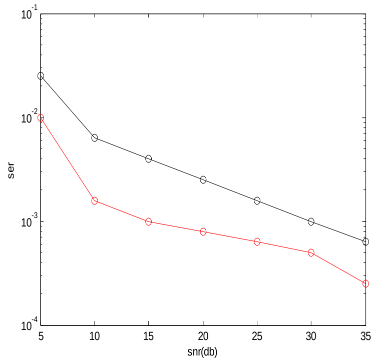
6.5 THE SPECTRAL EFFICIENCY OF VR M-PSK SCHEMES, FOR DIFFERENT SER RATES



6.6 THE SER OF VR M-PSK SCHEMES, FOR DIFFERENT SER TARGETS



6.8 THE SER OF EGC WITH VR M-PSK SCHEMES, FOR DIFFERENT SER TARGETS



7 CONCLUSION

- A VR M-PSK modulation scheme was introduced for wireless communication systems, where channel gain estimation is not available. VR M-PSK requires no channel gain estimation and increases the spectral efficiency of M-PSK systems under the instantaneous error rate requirement. The choice of the modulation order is not based on the S/N samples, but rather on the S+N ones.
- It was analytically shown that the S+N criterion is an attractive alternative to S/N for choosing the appropriate modulation order in VR communication systems, the same spectral efficiency can be achieved, when either S/N or S+N is used. Moreover, the proposed adaptive modulation scheme was applied to EGC receivers, enabling low-complexity diversity systems to increase their spectral efficiency.

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