ABSTRACT
The robustness of the proposed methods is illustrated by comparing the BER. Numerical simulations revealed good system performance. Finally, we consider experimental feasibility with both methods. Semianalytic defines an efficient procedure for computing exact semianalytic BER for modulation formats with circular constellations when the noise component of the decision variable has a circularly symmetric Gaussian distribution. The technique is demonstrated for 8PSK over the Digital Video Broadcasting-Satellite-Second Generation (DVB-S2) channel Semianalytic bit error rate (BER) estimation is a well-known method for evaluating the BER of a digital communication system. The main utility of the method is the significant time savings in computation relative to Monte Carlo simulation. Despite this advantage, no known reference defines the procedure for computing exact BER for M-ary phase shift keying (PSK) with ISI and AWGN using the semianalytic method. This letter defines an efficient procedure for computing exact semianalytic BER for modulation formats with circular constellations when the noise component of the decision variable has a circularly symmetric Gaussian distribution. In this thesis we do Comparative study of BER by both techniques and analysis the result by simulations. Also study of Monte Carlo simulation for digital modulation scheme and compare with Semianalytic technique. We concentrate our effort on semi-analytical error rate evaluation for digital transmission. Also study of delay for BPSK and QPSK due to ISI & AWGN environment through Monte Carlo simulation. Also do study of sensitivity of QPSK to phase jitter and impact of symbol jitter. Analysis of semianalytic techniques for M-ary PSK with Rayleigh channel, Rician channel.

KEYWORDS
Computer aided analysis, digital communication, error analysis, phase shift keying, simulation, Rayleigh and rician fading

1. INTRODUCTION
BER analysis is facing more and more challenges nowadays. Coding and modulation provide the means of mapping information into waveforms such that the receiver (with an appropriate demodulator and decoder) can recover the information in a reliable manner. Computer aided methods for determining the probability of bit error Pb or BER of a digital communication system are widely used when an accurate analytic expression for BER is not available. The most general technique for BER estimation is the Monte Carlo method. A simulation is performed by passing a known data sequence through a system model and counting the bit errors at the receiver decision device. A simple and unbiased estimator of the BER is the sample mean of the simulated BER. If N is the number of bits evaluated, then as N →∞ the estimate will converge to the true BER. In practice, the rule of thumb that N should be on the order of 10/Pb produces a 90% confidence interval of about two on the BER estimate. Simulations of systems operating at very low error probabilities therefore require prohibitively large run times to accurately estimate BER using the Monte Carlo method. There are several techniques for accelerating the estimation of BER. The method considered here is called the semianalytic or quasianalytic method. With this method the problem is broken into two parts, the first dealing with the signal component, and the second dealing with the noise component of the decision variable at the receiver. For a known data sequence and a static channel, the demodulated and sampled signal component at the decision device is deterministic and can be evaluated by simulation. The data sequence used in the simulation should have specific properties in order to emulate random data with equiprobable symbols.

2.1 System model: This model gives the general equation to compute BER with ISI and AWGN. It used to find BER for DVB-S2 channel.

PSK Constellation with ISI and awgn
The M signal waveforms for ideal M-ary PSK are represented as

\[ s_m(t) = g(t) \cos(\omega_c t + \theta_m), m = 0, 1, \ldots, M - 1 \]  

(1)

Where \( g(t) \) is a pulse waveform used to shape the spectrum of the transmitted signal and \( \theta_m \) is the information-bearing phase angle which takes M possible values

\[ \theta_m = \frac{(2m + 1)\pi}{M}, m = 0, 1, \ldots, M - 1 \]  

(2)
of both linear and non-linear time invariant distortions with memory. Assuming equiprobable symbols, then in order to completely characterize the ISI of a channel with L symbol periods of memory, it is sufficient to consider all possible sequences of L symbols. A maximal length pseudorandom \( M \) symbol sequence will satisfy this property. For \( M=2 \), linear feedback shift registers can be used to generate maximal length pseudorandom bit sequences. For \( M>2 \), efficient methods for generating maximal length pseudorandom symbol sequences have also been proposed. With the addition of cyclic prefixes and postfixes of L symbols each, a simulation using one cycle of a \( M \) length pseudorandom symbol sequence is sufficient to emulate equiprobable data symbols. Therefore, performing a simulation using \( M+2L \) symbols from a maximal length \( M \) symbol sequence, and discarding the first and last \( L \) demodulated and detected decision variable points, the resulting \( M \) decision variable points will completely characterize the effect of the system ISI on the signal. This set of decision variables can be defined in terms of their respective magnitudes and phases or in-phase and quadrature components 

\[
s_k = r_k e^{j\theta_k} = i_k + jq_k, \quad k = 0, 1, \ldots, \quad M - 1.
\]

(3)

When AWGN is present at the receiver input, the decision variables are 

\[
y_k = s_k + n_k, \quad k = 0, 1, \ldots, \quad M - 1
\]

(4)

A pseudorandom symbol sequence emulates equiprobable symbols, averaging the BER over all of the resulting decision variable points will determine the BER of the system. The bit error probability is therefore given by 

\[
P_b = \frac{1}{M} \sum_{k=0}^{M-1} P_b(s_k, m)
\]

(5)

Where \( P_b(s_k, m) \) is the bit error probability for decision variable \( s_k \) with correct decision region \( R_m \).

Expressing in vector (5) notation produces 

\[
P_b(s_k, m) = \frac{1}{\log_2(M)} \sum_{n=0}^{M-1} w_{mn} P_{n/m}(s_k)
\]

(6)

The term \( P_{n/m}(s_k) \) is the probability that the received decision variable \( s_k \) from region \( R_m \) would fall in decision region \( R_n \) under the influence of noise, and \( w_{mn} \) is the number of bit errors that occurs in this event. For an arbitrary bit mapping, the bit error weight matrix \( W \) with elements \( w_{mn} \) can be generated from the weight matrix for the binary bit mapping by swapping rows and then columns according to the desired symbol assignment. Upper equation can be simplified by noting that each received decision variable can be rotated to decision region \( R_m \) provided that the bit error weight matrix \( W \) is appropriately transformed as well. The rotated decision variable from decision region \( R_m \)

\[
s'_k = e^{-\frac{j2\pi m}{M}} s_k
\]

(7)

The required transformation for the \( W \) matrix consists of a cyclic left shift of row \( m \) by \( m \) positions. This operation can be expressed as 

\[
W_{mn} = W_{m(n+m)mod M}
\]

(8)

In terms of the transformed variables, (6)

\[
P_b(s_k, m) = \frac{1}{\log_2(M)} \sum_{n=0}^{M-1} w_{mn} P_{n/m}(s'_k)
\]

(9)

In this way, only the error event probabilities for decision variables from decision region need to be computed \( R_m \). Expressing (9) in vector

\[
P_b(s_k, m) = \frac{1}{\log_2(M)} W^T(k) p(k)
\]

(10)

Where \( p \) is a column vector with elements \( p(k) \) as shown in Fig.1 and circularly symmetric additive Gaussian distributed noise, \( M/2 \) independent half-plane probabilities and \( M/4 \) independent quarter-plane probabilities can be written

\[
h_i(k) = \frac{1}{2} \text{erfc} \left( \frac{r_i \sin(2\pi i M) - \theta_i}{\sigma} \right), \quad i = 1, 2, \ldots, \quad M/2
\]

(12)

\[
q_i(k) = h_i(k)(1 - h_i(M/4(K))), \quad i = 1, 2, \ldots, \quad M/4
\]

(13)

In addition to the \( 3M/4 \) probabilities specified by (11) and (12), the remaining required \( M/4-1 \) independent probabilities can be found using the following correction plane probabilities

\[
c_i(k) = \frac{1}{\pi\sigma^2} \int_0^{\infty} \int_{\tan^{-1}(x/M)}^{\tan^{-1}(y/M)} e^{-((x+i_k)^2+(y+q_i)^2)/\sigma^2} dy dx
\]

\[
i = 1, 2, \ldots, \quad M/4
\]

(14)
The expressions for the probabilities (11)-(14) are given by integrating the circularly symmetric Gaussian pdf over the specified regions. Evaluating the inner integral of (14) produces
\[ c_i(k) = \frac{1}{2\sqrt{2\pi \sigma^2}} \int_0^\infty e^{-\frac{(x+i)^2}{2\sigma^2}} f(x; i, M, q_i, \sigma) dx \]
\[ i = 1, 2, \ldots, \frac{M}{2} - 1 \]  
(15)
\[ f(x; i, M, q_i, \sigma) = \text{erfc}(\frac{\tan(\frac{(1-i)\pi}{M}) x + q_i}{\sigma}) - \text{erfc}(\frac{\tan(\frac{2\pi}{M}) x + q_i}{\sigma}) \]  
(16)

Where \( i_k = r_k \cos(\theta_k) \) and \( q_k = r_k \sin(\theta_k) \)

The values for the correction plane probabilities \( c_i \) specified by (15) and (16) can be evaluated by numerical integration. Expressions for the half-plane, quarter plane, and correction plane probabilities can also be written in terms of the desired probabilities \( P_{n/0}(s_k) \)

\[ h_i(k) = \sum_{n=1}^{\frac{M}{2}} P_{n/0}(s_k), i = 1, 2, \ldots, \frac{M}{2} \]  
(17)
\[ q_i(k) = \sum_{n=1}^{\frac{M}{2}} P_{n/0}(s_k), i = 1, 2, \ldots, \frac{M}{4} \]  
(18)
\[ c_i(k) = \sum_{n=1}^{\frac{M}{2}} P_{n/0}(s_k), i = 1, 2, \ldots, \frac{M}{4} - 1 \]  
(19)

The system of linear equations specified by (17)-(19) can be expressed compactly in matrix-vector notation as
\[ z(k) = Ap(k) \]  
(20)
Where \( z(k) \) is a known column vector of \( h(k), q(k), c(k) \) values, \( A \) is a matrix of known coefficients, and \( p(k) \) is an unknown column vector of \( P_{n/0}(s_k) \) values. The matrix \( A \) and the form of the vector \( z(k) \) are determined by the PSK modulation order \( M \). For 8PSK they are given by
\[ A = \begin{bmatrix} 1 & 1 & 1 & 1 & 0 & 0 & 0 & 0 \\ 0 & 1 & 1 & 1 & 1 & 0 & 0 & 0 \\ 0 & 0 & 1 & 1 & 1 & 1 & 0 & 0 \\ 1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 1 & 1 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 & 1 & 1 & 1 & 1 \\ 1 & 1 & 0 & 0 & 0 & 0 & 0 & 0 \\ 0 & 0 & 1 & 1 & 0 & 0 & 0 & 0 \end{bmatrix} \]
\[ z(k) = \begin{bmatrix} h_1(k) \\ h_2(k) \\ h_3(k) \\ h_4(k) \\ q_1(k) \\ q_2(k) \\ q_3(k) \\ q_4(k) \end{bmatrix} \]  
(21)

The system of equations specified by (20) can be solved for the unknown probabilities \( p(k) = A^{-1}z(k) \)  
(22)
Substituting (22) into (11) produces the desired expression for the bit error probability
\[ P_b = \frac{1}{M^2 \log_2(M)} \sum_{k=0}^{M^2-1} W^{T}(k) A^{-1} z(k) \]  
(23)

2.2. Semianalytic BER estimation for PSK: We now briefly consider the development of an algorithm for the determination of the BER in a binary PSK system using semianalytic simulation. We do this in a way that is easily extensible to QPSK. Consider the signal constellation. The transmitted signal points are denoted \( S_i \) and \( \tilde{S}_i \) is received. Difference \( (d_{ij}) \) between transmitted and received is due to inter symbol interference, nonlinear distortion, of other signal degrading effect.

The conditional error probability, conditioned of transmission \( S_i \) is
\[ P_r\{\text{error}|S_i\} = \int_{\tilde{S}_i \notin S_i} \frac{1}{\sqrt{2\pi \sigma^2}} \exp\left(-\frac{(n - \tilde{S}_i)^2}{2\sigma^2}\right) dn \]  
(24)
In terms of the Gaussian Q-function, the preceding equation becomes
\[ P_r\{\text{error}|S_i\} = Q\left(\frac{\tilde{S}_i}{\sigma}\right) \]  
(25)
The overall BER, obtained by averaging over the entire sequence of \( N \) bits, is given by
\[ P_b = \frac{1}{N} \sum_{k=1}^{N} Q\left(\frac{\tilde{S}_i}{\sigma}\right) \]  
(26)

2.3. Semianalytic BER estimation for QPSK: We now consider a semianalytic estimator for the symbol error probability \( P_S \) in a QPSK system. Since a QPSK signal constellation has four signal points rather than two, and since the signal space has two dimensions rather than one, the semianalytic estimator for QPSK is different from the estimator for PSK in that a dimension must be added for the quadrature channel.

Consider the signal constellation fig. The transmitted signal point are denoted \( S_j \), \( j=1,2,3,4 \), and the decision region are denoted \( D_j \); otherwise an error occurs.
Fig 4: Semianalytic BER estimation for QPSK

In Fig 4 it is assumed that $S_1$ is transmitted. As a result of intersymbol interference and distortion, $S_1 \neq \hat{S}_1$. We say that simulation will account for the effect of intersymbol interference but not the effect of noise.

The direct and quadrature component of $\hat{S}_1$ are denoted $\hat{S}_x$ and $\hat{S}_y$, respectively, where $\hat{S}_x = \text{Re}(\hat{S}_1)$ and $\hat{S}_y = \text{Im}(\hat{S}_1)$. When noise adding $n_x$ and $n_y$ to $\hat{S}_x$ and $\hat{S}_y$, respectively, a correct decision is made conditioned on $S_1$ transmitted, if

$$\hat{S}_x + n_x, \hat{S}_y + n_y \in D_1.$$  

An error is made if $\hat{S}_x + n_x, \hat{S}_y + n_y \notin D_1$. Since we are developing a semianalytic estimator, the impact of noise is treated analytically. Thus, given that $S_1$ is transmitted and $\hat{S}_1$ is received, an error is made if

$$\{ \hat{S}_1 = \hat{S} \}$$

it can be synthesized than final equation implement

$$P_e \{ \text{error} | S_1 \} = \int \int \frac{1}{2\pi \sigma_s \sigma_n} \exp\left(-\frac{(n_x - \hat{S}_x)^2}{2\sigma_n^2} - \frac{(n_y - \hat{S}_y)^2}{2\sigma_n^2}\right) dn_x dn_y$$

(27)

Than manipulate it $k$ transmitted symbol in a simulated sequence of $N$ symbols. Than conditional symbol Probability over the entire sequence of $N$ symbols, is given by

$$P_e < \frac{1}{N} \sum_{k=1}^{N} \left[ Q\left(\frac{\text{Re}\left(\hat{S}_{k}\right)}{\sigma_n}\right) + Q\left(\frac{\text{Im}\left(\hat{S}_{k}\right)}{\sigma_n}\right) \right]$$

(28)

2.4. DVB-S2 transponder models for simulations: For simulations, the “transparent” (i.e. non regenerative) satellite transponder model may be composed of an input filter (IMUX), a power amplifier (TWT or SSA) and an output filter (OMUX). Two amplifier models are here defined, the linearized TWTA (LTWTA) and the non-linearized TWTA. SSPAs have not been considered since they are less critical than TWTAs in terms of degradations.

The reference symbol rate with the specified IMUX/OMUX filter bandwidth is $R_s = 27.5$ Mbaud

Fig 5: Satellite transponder model

3 RESULTS AND DISCUSSION:

In this paper we have done the performance analysis of simulation techniques (Semianalytic and Monte Carlo Simulation) for different application with ISI and awgn environment. Shown in fig. simulation model of DVB-S2 channel. Mathematical model described on article for bit error rate calculation. Shown fig. simulation model of QAM and PSK. We have done QPSK and BPSK analysis to use of semianalytic and Monte Carlo simulation model. Shown in fig. simulation model of QPSK and BPSK. Mathematical model described on article for Semianalytic BER estimation for PSK and article for Semianalytic BER estimation for QPSK. We have done Rayleigh fading and Rician fading channel analysis for PSK by semianalytic simulation model and theoreothetical fading.

In this chapter we will discussed and show the following result

- Semianalytic and Monte Carlo simulation for DVB-S2 channel.
- BPSK and QPSK analysis by Semianalytic and Monte Carlo simulation with ISI and awgn.
- Semianalytic techniques use in 16-PSK to estimate fading performance.

3.1 Analysis of DVB-S2 Channel by semianalytic simulation and Monte Carlo simulation

Fig 3.1 constellation diagram of uncoded 8PSK with DVB-S2 channel

Fig 3.2 SA and MC BER as a function of SNR for encoded 8 PSK and DVB- S2 channel
The results of the semi analytic and Monte Carlo simulations are plotted in Fig. 3.2. The differences between the results are consistent with the expected estimation error for a Monte Carlo simulation of $10^9$ bits in length. While the Monte Carlo simulation produced no errors when the SNR was greater than 22dB, the semianalytic method produced accurate results at all signal-to-noise ratios.

**3.3 Analysis of BPSK, QPSK with ISI by semianalytic simulation and Monte Carlo simulation:**

Fig. 3.3 BPSK simulations for system with ISI

Fig. 3.3 Monte Carlo simulation produced error no close to theoretical result. The bandwidth of the transmitter filter, which gives rise to ISI, is equal to the bit rate. In MC simulation over a range in an estimated value of BER that is based on decreasing number of observed errors as Eb/No increases. Show MC result, the BER estimate at large values of Eb/No will be less reliable than the BER estimate at smaller values of Eb/No but SA result, more reliable than the MC. SA approach takes less time than MC to estimate the performance evaluation.

**QPSK ANALYSIS:**

Fig. 3.4 QPSK semianalytic simulation result

Fig. 3.5 Comparative study of MC and SA for QPSK

Fig. 3.4 show the SA simulation result of QPSK with the effect of the ISI resulting the transmitter filtering. The received signal constellation is show fig. 3.3. This scattering result from the fact that the system exhibits a memory length that exceeds two symbols, although the effect of this additional memory is small. The increase in the BER resulting from the ISI is show in fig. 3.4. Fig. 3.5 show the comparative study of MC and SA for QPSK. The increase is BER to increase ISI efficiently SA approach as compare to MC. SA take less time compare to MC for performance evaluation.

**Performance evaluation of 16-PSK with fading by semianalytic approach:**

**4. CONCLUSION**

We have seen that two important approaches to estimating BER. The Monte Carlo method is the least restrictive but costliest in term of computer time. The semianalytic is by far the most rapid and in a linear channel is exact as well, is so far as the effects of Gaussian noise are concerned. Because of speed it
also suitable sensitivity calculation of phase jitter and symbol jitter. Monte-Carlo method, which weights all errors equally and makes no assumption about the form of the decision metric, to more complex estimation schemes which do make assumptions about the decision metric. Recall that this decision allows one to expect a tradeoff between prior knowledge and computer execution time.

A fast and accurate method for computing semianalytic BER for Digital Modulation Techniques with ISI and AWGN has been presented. Two simplifications were applied relative to directly evaluating the semianalytic BER using numerical integration. First, the decision variable points were each rotated to decision region and the bit error weight matrix was appropriately transformed. The technique is applicable to ideally demodulated and matched filter detected Digital Modulation Techniques with a static channel distortion, and produces an exact BER result when the received noise has a circularly symmetric Gaussian distribution.

It should be pointed out that for extremely complex systems, it is usually desirable to start out with the simplest model that incorporates only the essential features of the system under study. Simulations based on simple models are easier to verify and errors are more easily identified. The simulation can then be enhanced to include other interesting and important features of the communication system under study. A criterion for the design of symbol sequences to be used in SA simulations was established which provides very good accuracy and reliability in the evaluation of performance parameters. It was shown that for an M-symbol modulating alphabet and an L symbol channel memory, M-ary maximum length complete sequences (MLCS) of length \( M^L \) can be used to accurately evaluate error probabilities and power spectra at the output of nonlinear devices. The extension of SA evaluation of interference effects in error probabilities was addressed. A very efficient technique was proposed which consists of combining the simulation of the main and interfering signals with Gaussian quadrature rule. Several results were presented which confirm both the accuracy and computing time efficiency of the methods proposed. In a similar way, the SA technique can also be extended to frequency non-selective slowly fading channels. An extension of the technique to include synchronization errors in the demodulator is described and is referred to as the statistical averaging method. The BER specified by (21) that accounts for time invariant distortions such as high power amplifier nonlinearity and non-ideal filtering is first expressed as a function of \( \alpha r_k,\alpha i_k,\alpha d_k,\alpha q_k \) where \( \alpha \) accounts for the effect of fading. The resulting expression is then averaged over the pdf of \( \alpha \).

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