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PERFORMANCE ANALYSIS OF 49-QPRS THROUGH NONLINEAR SATELLITE CHANNELS IN THE PRESENCE OF GAUSSIAN NOISE AND COCHANNEL INTERFERENCE

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ABSTRACT : This paper is concerned with the analytical performance analysis and evaluation of 49-ary Quadrature Partial Response Signaling when transmitted through nonlinear satellite channel in the presence of Additive White Gaussian Noise and Co-channel Interference ; in both uplink and downlink channels . The main source of the nonlinearities is the Traveling Wave Tube Amplifier on-board of the satellite. The transponder nonlinearities considered in this paper are due to : input amplitude -to-output amplitude conversion and input amplitude-to-output phase conversion . The cochannel interference results from interferers in the passband of the coherent receiver . The results in terms of the dependence of the average symbol error probability upon the uplink and downlink ; signal-to-noise ratio's and signal -to - cochannel interference signal ratio's at different values of Back-Off from saturation operation of the Traveling Wave Tube Amplifier on-board of the satellite are illustrated . The results showed that the Back-Off value is the dominant factor in determining the system performance .The appropriate values of the threshold levels ; d"s and compensation phases ; θ"s at the receiver are highly correlated and can only be arrived at by minimizing the average symbol error probability .

Key Words :

Quadrature Partial Response Signaling ; QPRS - Additive White Gaussian Noise ; AWGN Co-channel Interference ; CCI - Traveling Wave Tube Amplifier ; TWTA - Input amplitude -to-output amplitude conversion ; AM/AM -Input amplitude-to-output phase conversion ; AM/PM -The average symbol error probability; P_e - Signal-to-noise ratio's ; ρ_{nu} and ρ_{nd} - Signal -to - cochannel interference signal ratio's ; ρ_{cu} and ρ_{cd} - Back-Off from saturation operation of the TWTA ; BO - Threshold levels ; d"s -compensation phases ; θ"s at the receiver and Average symbol error probability; P_e

I-INTRODUCTION

The duobinary " correlative coded technique" signal introduces a controlled amount of Intersymbol Interference (ISI) in order to simplify the filter design ; particularly the phase-equalization problem ; and to enable the transmission at ; or slightly higher ; the Nyquist rate [1].

The 49-ary QPRS consists of two seven-level duobinary ; partial response baseband signals (PRS) which are modulating two orthogonal carriers. One of the main advantages of QPRS as compared to equivalent schemes is that they are speed tolerant, i.e., it is possible to transmit at rate which is higher than Nyquist rate without suffering significant degradation [1] and [2] . The 49-ary QPRS is expected to find increased applications in the future communication by satellite or microwave radio link due to its spectrum efficiency ; ≥4 bits / sec. / Hz of the IF-bandwidth , its relative simplicity of implementation and good error performance through linear Gaussian channels [3] - [6] . Performance of another schemes such as 16-ary QAM , 16-ary QAM/MSK , 16-ary CPSK and 9-ary QPRS through nonlinear satellite channel in the presence of AWGN are illustrated in [7] - [10] respectively .

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The purpose of this paper is to present an analytical performance evaluation of 49-ary QPRS in the presence of TWTA nonlinearities ; AWGN and CCI preceding and following the nonlinearities . In Section -II the average symbol error probability ; Pe analysis is presented . In section-III ; computation aspects , results and comments are presented . Section IV is concerned with the conclusion about the results .

II-AVERAGE SYMBOL ERROR PROBABILITY ANALYSIS

Assuming the system model shown in Fig.(1) ; the modulated QPRS ; $S_I(t)$ during any symbol duration ; T_s may be written as:

$$S_I(t) = \mu_I(t) \cos \omega_c t - \lambda_I(t) \sin\omega_c t \tag{1}$$

Where. $\omega_c = 2\pi f_c$, f_c is the carrier frequency , P(t) is the pulse shape defined by :

- $P(t) = A \qquad 0 \le t \le T_s$; and
  $\quad = 0 \qquad$ elsewhere
- $\mu_I(t) = a\,P(t)$ and $\lambda_I(t) = b\,P(t)$; (2)
- T_s is the symbol duration , a and b are two independent random variables and are $\in$ {-6,-4,-2,0,2,4,6} with probabilities $\in$ {1/16,2/16,3/16,4/16,3/16,2/16,1/16} respectively . It is assumed that ai and aj or bi and bj are independent for all values of $i \ne j$.

The uplink signal $S_I(t)$ is corrupted with the uplink narrow-band Gaussian noise $n_u(t)$. The resultant signal at the input of TWTA on-board the satellite's transponder may be written as :

$$S_i(t) = R(t) \cos [\omega_c t + \phi_i(t)] \tag{3}$$

Where . $R^2(t) = [X_I^2(t) + Y_I^2(t)]$ (4)

- $X_I(t) = \mu_I(t) + n_{uc}(t) + C_{uc}(t)$, and
- $Y_I(t) = \lambda_I(t) + n_{us}(t) + C_{us}(t)$ (5)
- $\phi_i(t) = \tan^{-1} [Y_I(t) / X_I(t)]$ (6)
- $n_{uc}(t)$ and $n_{us}(t)$ are the uplink ; inphase and quadrature components of the uplink AWGN , each are independent with zero mean and variance σ_{nu}^2 ; and
- $C_{uc}(t)$ and $C_{us}(t)$ are the uplink ; inphase and quadrature components of the interfering signal and are assumed to be originated independently of each other and of the transmitted signal or noise sources . They are represented in [9] as follows :

$$C_c(t) = \sum_i B_i \cos [(\omega_i + \omega_c)t + \gamma_i(t) + \varepsilon_i] \quad , \text{ and}$$

$$C_s(t) = \sum_i B_i \sin [(\omega_i + \omega_c)t + \gamma_i(t) + \varepsilon_i] \tag{7}$$

Where . $i = 1,2,3,\ldots\ldots,N_c$ represent the number of interferers (in this work $i = 1$);

- B_i represents the amplitude of ith interferer ;
- $\gamma_i(t)$ represents the digital modulated phase ; and
- ε_i represents the digital unmodulated phase which uniformly distributed (0 , 2π)

The signal $S_i(t)$ is degraded by the TWTA , the output signal $S_o(t)$ may be given as :

$$S_o(t) = F[R(t)] \cos \{ \omega_c t + \phi_i(t) - \psi[R(t)] + \theta \} \tag{8}$$

Where . F [R] denotes the AM/AM conversion function ;

. ψ [R] denotes the AM/PM conversion function ; and

. θ represents the compensation phase in radians at receiver to account for the average AM / PM conversion .

The signal S_o (t) is corrupted with the downlink Gaussian noise n_d (t) and cochannel interference C_d (t) to give the signal S_2 (t) at the input to the coherent receiver as:

$$S_2(t) = X_2(t) \cos(\omega_c t) - Y_2 \sin(\omega_c t) \tag{9}$$

Where .

$$X_2(t) = \mu_2(t) + n_{dc}(t) + C_{dc}(t) \; ;$$

$$Y_2(t) = \lambda_2(t) + n_{ds}(t) + C_{ds}(t) \; ;$$

$$\phi_o(t) = \phi_i(t) - \psi[R(t)] + \theta \; ;$$

$$\mu_2(t) = F[R(t)] \cos[\phi_o(t)] \; ; \text{ and}$$

$$\lambda_2(t) = F[R(t)] \sin[\phi_o(t)] \tag{10}$$

. n_{dc} (t) and n_{ds} (t) are the downlink ; inphase and quadrature components of the downlink AWGN , each of which with zero mean and variance σ^2_{nd} ; and

. C_{dc}(t) and C_{ds} (t) are the downlink ; inphase and quadrature components of the cochannel interference

The receiver coherently demodulate the input signal S_2 (t) to give the inphase and quadrature components of the baseband signal $X_2(t)$ and $Y_2(t)$. The later are sampled at $t = t_o + k T_s$, $0 \leq t_o \leq T_s$. A decision is made to estimate the receiving symbol corresponding to the transmitted symbol . The diagram of Fig.(2) illustrates all possible components of the baseband received samples in the absence of any degradation and the regions for correct decision ; R_i corresponding to the transmitted elements A_i , i = 1,2,3,4, 5,6,.. ...., 49.

Since each element A_i in the transmitted set has its specific probability of existence , $p(A_i)$ and conditional error probability ; $p_{e,\,A_i}$ it follows that the average symbol error probability P_e for 49-ary QPRS may be given as :

$$P_e = \sum_i p(A_i) \, . \, p_{e,\,A_i} \tag{11}$$

Where .

$$p(A_1) = p(A_7) = p(A_{43}) = p(A_{49}) = 1/256 \; ;$$
$$p(A_8) = p(A_{14}) = p(A_{36}) = p(A_{42}) = 2/256 \; ;$$
$$p(A_{15}) = p(A_{21}) = p(A_{29}) = p(A_{35}) = 3/256 \; ;$$
$$p(A_9) = p(A_{13}) = p(A_{37}) = p(A_{41}) = 4/256 \; ;$$
$$p(A_{16}) = p(A_{20}) = p(A_{30}) = p(A_{34}) = 6/256 \; ;$$
$$p(A_2) = p(A_6) = p(A_{44}) = p(A_{48}) = 2/256 \; ;$$
$$p(A_3) = p(A_5) = p(A_{45}) = p(A_{47}) = 3/256 \; ;$$
$$p(A_{10}) = p(A_{12}) = p(A_{38}) = p(A_{40}) = 6/256 \; ;$$
$$p(A_{17}) = p(A_{19}) = p(A_{31}) = p(A_{33}) = 9/256 \; ;$$
$$p(A_4) = p(A_{22}) = p(A_{28}) = p(A_{46}) = 4/256 \; ;$$
$$p(A_{11}) = p(A_{23}) = p(A_{27}) = p(A_{39}) = 8/256 \; ;$$
$$p(A_{18}) = p(A_{24}) = p(A_{26}) = p(A_{32}) = 12/256 \; ;$$
$$p(A_{25}) = 16/256 \tag{12}$$

. The conditional error probabilities $p_{e,Ai}$'s are also equal ; as given in (12); for example :

$$p_{e,A1} = p_{e,A7} = p_{e,A43} = p_{e,A49} \quad ; \quad \text{and so on for the other equalities} \tag{13}$$

Substituting from (12) and (13) into (11) yields :

$$p_e = (1/64)\,[\,4\,p_{e,A25} + 12\,p_{e,A32} + 8\,p_{e,A39} + 4\,p_{e,A48} + 9\,p_{e,A33} + 6\,p_{e,A40} + 3\,p_{e,A47}$$

$$+ 2\,p_{e,A48} + 4\,p_{e,A41} + 6\,p_{e,A34} + 3\,p_{e,A35} + 2\,p_{e,A42} + p_{e,A25}] \tag{14}$$

It is evident that the error probability assuming A_i is transmitted may be wr.tten as :

$$p_{e,Ai} = 1 - \iint_{R_i} p_{Ai}(X_2, Y_2)\, dX_2\, dY_2 \tag{15}$$

Where . $p_{Ai}(X_2, Y_2)$ denotes the joint probability density function ; pdf of X_2, Y_2 assuming that element A_i is transmitted .

Using Bayes rules for conditional probability ; (15) may be written as :

$$p_{e,Ai} = 1 - \iint_{R_i} [\iint_{-\infty}^{\infty} p(X_2, Y_2 / X_1, Y_1)\, p_{Ai}(X_1, Y_1)\, dX_1\, dY_1\,]\, dX_2\, dY_2 \tag{16}$$

Using the fact that : integration process are linear transformation or mapping , (16) may be written as :

$$p_{e,Ai} = 1 - \iint_{-\infty}^{\infty} [\iint_{R_i} p(X_2, Y_2 / X_1, Y_1)\, p_{Ai}(X_1, Y_1)\, dX_2\, dY_2\,]\, dX_1\, dY_1 \tag{17}$$

Where . $p(X_2, Y_2 / X_1, Y_1)$ denotes the joint pdf of X_2 and Y_2 conditioned upon X_1 and Y_1 assuming that the signal element Ai is transmitted .

. R_i defines the region for correct decision when Ai is transmitted .

. $p_{Ai}(X_1, Y_1)$ is the joint pdf of X_1 and Y_1 assuming A_i is transmitted .

Assuming sampling of the random processes defined by (5) ; both $n_{uc}(t)$ and $n_{us}(t)$ are Gaussian random processes , at specific time (t_o) , may be regarded as independent Gaussian random variables n_{uc} and n_{us} respectively ; each with zero mean and standard deviation σ_{nu} whilst C_{uc} and C_{us} are the inphase and quadrature components random variables associated with the uplink cochannel interference . The joint pdf of X_1 and Y_1 conditioned upon C_{uc} and C_{us} is given as follows :

$$p_{Ai}(X_1, Y_1 / C_{uc}, C_{us}) = (\,1/2\,\pi\,\sigma_{nu}^2\,)\,\exp - [\,(X_1 - \mu_1 - C_{uc})\,/\,(\sqrt{2}\,\sigma_{nu}\,)]^2 \;.$$

$$\exp - [\,(Y_1 - \lambda_1 - C_{us})\,/\,(\sqrt{2}\,\sigma_{nu}\,)]^2 \tag{18}$$

Assuming sampling of the random processes defined by (9) , both $n_{dc}(t)$ and $n_{ds}(t)$ are Gaussian random processes , at specific time (t_o) , may be regarded as Gaussian random variables n_{dc} and n_{ds} respectively ; each with zero mean and standard deviation σ_d . Thus the conditional pdf may be given as :

$$p\,(X_2, Y_2 / X_1, Y_1 \text{ and } C_{dc}, C_{ds}) = (1/2\pi\sigma_{nd}^2)\,\exp -[(X_2-\mu_2-C_{dc})/(\sqrt{2}\,\sigma_{nd})]^2 . \exp -[(Y_2-\lambda_2-C_{ds})/(\sqrt{2}\,\sigma_{nd})]^2 \quad (19)$$

Using Taylor series expansion about $[(X_1-\mu_1)/(\sqrt{2}\,\sigma_{nu})]$ and $[(Y_1-\lambda_1)/(\sqrt{2}\,\sigma_{nu})]$ for (18) and ; an$(X_2-\mu_2)/(\sqrt{2}\,\sigma_{nd})$ and $(Y2-\lambda 2)/(\sqrt{2}\,\sigma_{nd})$ for (19) . The conditional pdf in (18) and (19) transform to the following forms :

$$p_{Ai}(X_1, Y_1) = (1/2\pi\sigma_{nu}^2) . \sum_m^\infty \sum_n (-1)^{m+n} . \{1/(\sqrt{2}\sigma_{nu})\}^{2m+2n} . b_{2m,2n} . \exp -[(X_1-\mu_1)/(\sqrt{2}\sigma_{nu})]^2 . \exp -[(Y_1-\lambda_1)/(\sqrt{2}\,\sigma_{nu})]^2 . H_{2m}[(X_1-\mu_1)/(\sqrt{2}\sigma_{nu})] . H_{2n}[(Y_1-\lambda_1)/(\sqrt{2}\,\sigma_{nu})] \quad (20)$$

$$p\,(X_2, Y_2 / X_1, Y_1) = (1/2\pi\sigma_{nd}^2) . \sum_K^\infty \sum_L (-1)^{K+L} . \{1/(\sqrt{2}\sigma_{nu})\}^{2K+2L} . a_{2k,2l} . \exp -[(X_2-\mu_2)/(\sqrt{2}\sigma_{nd})]^2 . \exp -[(Y_2-\lambda_2)/(\sqrt{2}\,\sigma_{nd})]^2 . H_{2k}[(X_2-\mu_2)/(\sqrt{2}\sigma_{nd})] . H_{2l}[(Y_2-\lambda_2)/(\sqrt{2}\,\sigma_{nd})] \quad (21)$$

Where . $b_{2m,2n}$ and $a_{2k,2l}$ are the coefficients of infinite double power series expansion of the joint characteristic functions of the uplink random variables C_{uc} and C_{us} ; and downlink random variables C_{dc} and C_{ds} respectively for one interferer [9] and [10] .

. $H_{2m}(.)$ and $H_{2n}(.)$; and $H_{2k}(.)$ and $H_{2l}(.)$ are calculated from the recurrence relationship for the Hermit Polynomial defined by :
$H_{m+1}(X) = [2X . H_m(X) - 2m . H_{m-1}(X)]$, given that $H_0(X) = 1$ and $H_1(X) = 2X$ constitute the starting points for evaluating $H_m(X)$ for all values of m

Substitute (20) and (21) into (17) ; for element A_i , $i = 25$ as an example , we get:

$$p_{e, A25} = 1 - \iint_{-\infty}^{\infty} F(X_1, Y_1)\, p_{Ai}(X_1, Y_1)\, dX_1\, dY_1 \quad (22)$$

Where . $p_{Ai}(X_1, Y_1)$ given in (20) ; and

$$. F(X_1, Y_1) = \iint_{R_{25}} P(X_2, Y_2 / X_1, Y_1)\, dX_2\, dY_2$$

$$= (1/4) . [\operatorname{erf}(d-\mu_2)/(\sqrt{2}\sigma_{nd}) + \operatorname{erf}(d+\mu_2)/(\sqrt{2}\sigma_{nd})] . [\operatorname{erf}(d-\lambda_2)/(\sqrt{2}\sigma_{nd}) + \operatorname{erf}(d+\lambda_2)/(\sqrt{2}\sigma_{nd})] - (1/4\pi)[\operatorname{erf}(d-\lambda_2)/(\sqrt{2}\sigma_{nd}) + \operatorname{erf}(d+\lambda_2)/(\sqrt{2}\sigma_{nd})] .$$

$$\sum_{K=1}^{\infty} (-1)^K . \{1/(\sqrt{2}\sigma_{nd})\}^{2K} . a_{2k,0} \{ \exp -[(d-\mu_2)/(\sqrt{2}\sigma_{nd})]^2 . \text{(cont.)}$$

$$H_{2k-1}\,[(d-\mu_2)/(\sqrt{2}\sigma_{nd})] + \exp-[(d+\mu_2)/(\sqrt{2}\,\sigma_{nd})]^2 \,.\, H_{2k-1}\,[(d+\mu_2)/(\sqrt{2}\sigma_{nd})]\,\}$$

$$-\,(1/2\sqrt{\pi}).\,[\,\mathrm{erf}\,(d-\mu_2)/(\sqrt{2}\sigma_{nd}) + \mathrm{erf}\,(d+\mu_2)/(\sqrt{2}\sigma_{nd})\,]\,.\sum_{L=1}^{\infty}(-1)^L \,.\{1/(\sqrt{2}\sigma_{nd})\}^{2L}$$

$$.\,a_{0,2l}\,.\,\{\,\exp-[(d-\lambda_2)/(\sqrt{2}\sigma_{nd})]^2 \,.\, H_{2l-1}\,[(d-\lambda_2)/(\sqrt{2}\sigma_{nd})] + \exp-[(d+\lambda_2)/(\sqrt{2}\,\sigma_{nd})]^2.$$

$$H_{2l-1}\,[(d+\lambda_2)/(\sqrt{2}\sigma_{nd})]\,\} + (1/\pi)\,.\sum\sum_{K,L=1}^{\infty}(-1)^{K+L} \quad a_{2k,2l}\,\{1/(\sqrt{2}\sigma_{nd})\}^{2K+2L}\,.$$

$$\{\,\exp-[(d-\mu_2)/(\sqrt{2}\sigma_{nd})]^2 \,.H_{2k-1}\,[(d-\mu_2)/(\sqrt{2}\sigma_{nd})] + \exp-[(d+\mu_2)/(\sqrt{2}\,\sigma_{nd})]^2$$

$$.\,H_{2k-1}\,[(d+\mu_2)/(\sqrt{2}\sigma_{nd})]\,\} \,.\,\{\,\exp-[(d-\lambda_2)/(\sqrt{2}\sigma_{nd})]^2 \,.\, H_{2l-1}\,[(d-\lambda_2)/(\sqrt{2}\sigma_{nd})] +$$

$$\exp-[(d+\lambda_2)/(\sqrt{2}\,\sigma_{nd})]^2 \,.\, H_{2l-1}\,[(d+\lambda_2)/(\sqrt{2}\sigma_{nd})]\,\} \qquad (23)$$

Substituting (20) and (23) into (22) we get the error probability for one element $p_{e,A25}$. The $p_{e,Ai}$, $i \neq 25$ are calculated in the same way, substitute into (14) to get the average error probability ; p_e for 49-QPRS at specific conditions of back-off ; uplink and downlink signal-to-noise ratio and signal-to-interference ratio .

III-COMPUTATION , RESULTS AND COMMENTS

Each infinite-double integration defining conditional error probability is numerically evaluated using the Cartesian products of Gauss-Hermit quadrature formulas [11] . The amplitude-phase model of the TWTA nonlinearities represented in [7]- [9] by :

$$F(R) = 10^{(\alpha\{\cos[\log(R/\hat{R})/\beta]-1\})} \qquad R > \tilde{R}$$
$$= R \qquad R \le \tilde{R} \qquad \text{, and}$$

$$\psi(R) = K_1\,[\,1-\exp(-K_2\,R^2)\,] + K_3\,R^2 \qquad (24)$$

Where . α , β , $\hat{R}$, $\tilde{R}$, K_1 , K_2 and K_3 are constants chosen to fit the measured amplitude and phase characteristics of the TWTA of type : A-TRW-DSCS-II with constant parameters above given to be 0. 394 , 0. 475 , 2. 317 , 0. 355 , 0. 605 , 0. 66 and 1/102.4 respectively .

The TWTA average transmitted power is given by :

$$[\,\bar{P}_t\,]_{max.} = \hat{R}^2\,/\,2 \qquad \text{at full saturation mode , and}$$

$$\bar{P}_t = (\,\hat{R}^2/\,2\,)\,.\,10^{(-BO/10)} \qquad \text{at any other operation mode} \qquad (25)$$

Where . BO denotes the degree of input back-off in dB .

The average power transmitted for 49-ary QPRS is given as :

$$\overline{P}_t = \sum_i \sum_j [(a_i^2 + b_j^2)^{1/2} A]^2 p(a_i) . p(b_j) = 10 A^2 \quad (26)$$

From (25) and (26) we can get an expression for the amplification factor ; **A** given as :

$$A = (\hat{R} / \sqrt{20}) . 10^{(-BO / 20)} \quad (27)$$

The uplink σ_{nu} and σ_{cu} and the downlink σ_{nd} and σ_{nd} expressed in terms of uplink and downlink ρ_{nu} and ρ_{nd} and ρ_{cu} and ρ_{cd} respectively as follows :

$$\sigma_{nu}^2 = \overline{P}_t / \rho_{nu} \quad \text{and} \quad \sigma_{nd}^2 = \overline{P}_r / \rho_{nd} \quad ; \text{and}$$
$$\sigma_{cu}^2 = \overline{P}_t / \rho_{cu} \quad \text{and} \quad \sigma_{cd}^2 = \overline{P}_r / \rho_{cd} \quad (28)$$

Where . $\overline{P}_r$ is the average power at the TWTA output , assuming noise and interference signal powers are neglected compared with the uplink signal power P_t , is given by :

$$\overline{P}_r = \sum_i \sum_j [F\{(a_i^2 + b_j^2)^{1/2} A \}]^2 p(a_i) p(b_j) \quad (29)$$

It is found from the results that the later depend upon back-off , signal-to-noise ratio and signal-to-interference ratio but the back-off is dominated factor ; a sample of the computed results are shown in Fig. (3) and Fig. (4) . The appropriate values of threshold level ; **d** and compensation θ can only be arrived at by minimizing the average symbol error probabilities; $\mathbf{P_e}$. Fig. (5) shows the minimum average error probability ; $\mathbf{P_e}$ detection for different values of threshold levels **d** and different compensation phases θ.

IV-CONCLUSION

In this paper we presented a complete analysis for the performance of 49-ary QPRS through two link nonlinear satellite channels in the presence of AWGN and CCI preceding and following the TWTA nonlinearities . An expression of the average symbol error probability has been derived and evaluated using Gauss-Hermite quadrature techniques for infinite-double integration and Hermit Polynomial for double summation . It is found from the results that $\mathbf{P_e}$; as a measure for the system performance ; depends upon back-off , signal-to-noise ratio and signal-to-interference ratio but back-off is the dominated factor . The values of threshold levels and compensation phase are found to be highly correlated and can only be arrived at optimum by minimizing the average symbol error probability $\mathbf{P_e}$. The results are useful for : satellite ; data and information Networks and radio-relay communications . QPRS ; for more spectral efficient ; is candidatd to replace the other modulation techniques in the near future .

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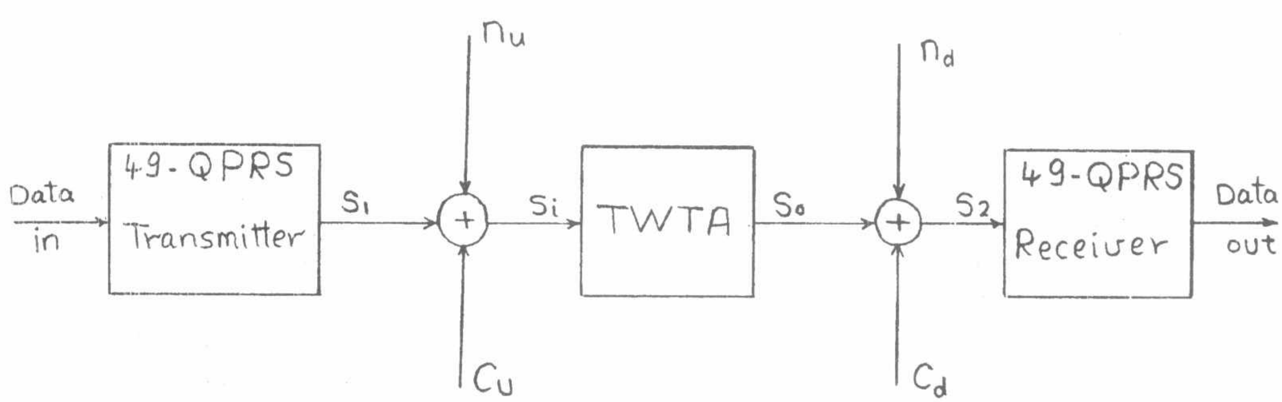

Fig. (1) System Model

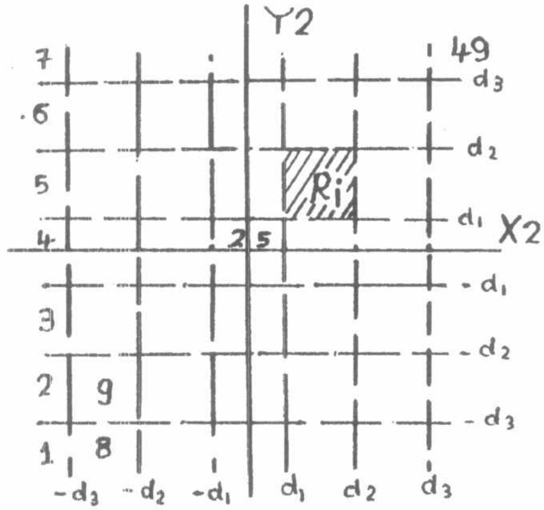

Fig. (2) Region for correct decision ; R_i.

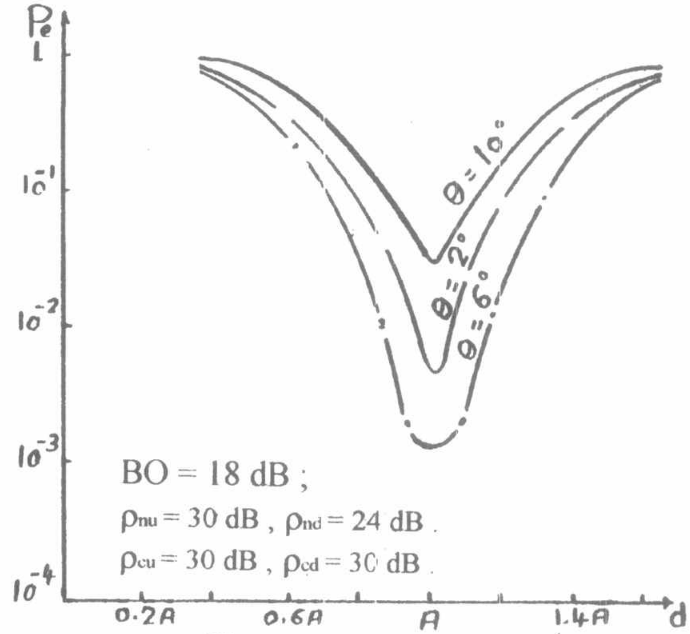

Fig. (5) Minimum Pe detection .

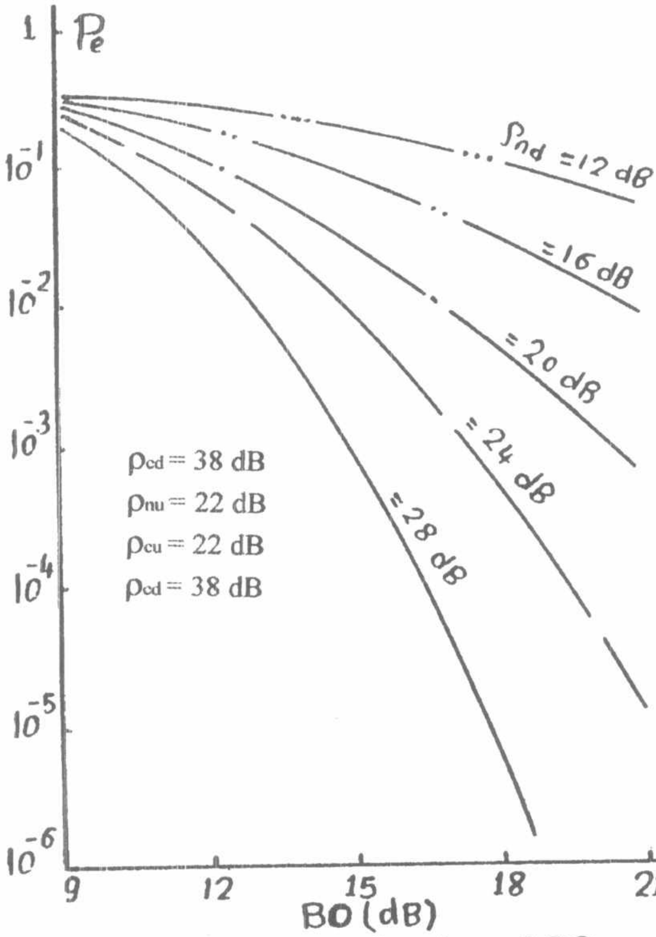

Fig. (3) Pe as function of BO.

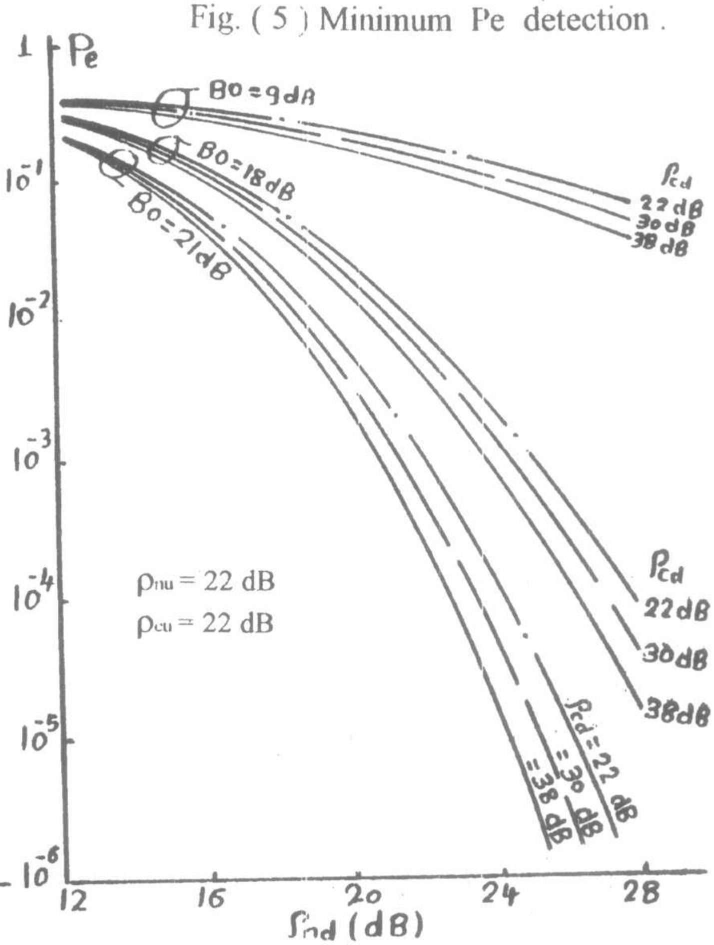

Fig. (4) Pe as function of ρ_{nd} .