무선통신에서의 Non-Linear Detector System 설계

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The System of Non-Linear Detector over Wireless Communication

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(Abstract)

Wireless communication systems, in particular, must operate in a crowded electro-magnetic environment where in-band undesired signals are treated as noise by the receiver. These interfering signals are often random but not Gaussian. Due to nongaussian noise, the distribution of the observables cannot be specified by a finite set of parameters; instead r-dimensioal sample space (pure noise samples) is equiprobably partitioned into a finite number of disjointed regions using quantiles and a vector quantizer based on training samples. If we assume that the detected symbols are correct , then we can observe the pure noise samples during the training and transmitting mode. The algorithm proposed is based on a piecewise approximation to a regression function based on quantities and conditional partition moments which are estimated by a RMSA (Robbins-Monro Stochastic Approximation) algorithm. In this paper, we develop a diversity combiner with modified detector, called Non-Linear Detector, and the receiver has a differential phase detector in each diversity branch and at the combiner each detector output is proportional to the second power of the envelope of branches. Monte-Carlo simulations were used as means of generating the system performance.

I. Introduction

For high bit rate digital transmission over mobile radio multipath channels, delay spread cannot be neglected. The particular characteristics of the multipath channel are generally unknown. Thus, the objective of the receiver is to identify the channel characteristics in order to reduce (or surpress) the ISI component and additive noise. This is accomplished by the adaptive operation of the receiver. The use of statistical methods of detection and estimation theory has provided powerful signal processing techniques for communication system design, image processing and many other system. In general, the statistical model most widely employed by system designers has been a gaussian noise model because the main justification for this model is the central limit theorem. Because the channel characteristics are not known beforehand, it is frequently erroneous to assume the gaussian noise model as a channel model. Despite its difficult mathematical structure, the nongaussian noise model is often more appropriate than the gaussian model in system design [12]. It has been recognized for a long time that the performance of detector suffers when inaccurate statistical models are used. Therefore, the system needs to be robustized so that departure from assumed conditions results in small degradation of performance and should be adaptive or self-learning so that almost optimal performance is available under all conditions. In this paper, non-linear (robust) detector which is based on piecewise approximations to a regression function involving only quantiles and partition moments which are estimated by a RMSA algorithm. For each partition. the coefficients of the non-linear detector are evaluated by using the conditional minimum mean square estimation criterion. In this system, QDPSK is employed as a modulation-detection scheme and postdetection diversity is chosen because the cophasing function nessary in predetection combining is not required and because switching noise due to an abrupt phase is not produced.

II. Description of System Model

In this paper, we employ a diversity reception providing a viable solution, because of the presence of a diversity path. In postdetection diversity, detector outputs are weighted according to each branch's channel condition before combining so that the contribution of the weaker signal branch is minimized [1]-[4]. The combiner output can be expressed

$$I+jQ=\sum_k X_k(nT)\cdot X_k^*((n-1)T) \quad (1),$$

where $X_k(t)$ is the k^{th} branch received faded signal. In this paper, we consider rician fading channel as a channel model and express three components as x_{dk}, x_{fk}, x_{nk} separately as the *DPD* (differential phase detection) detector input components. In general, $M-ary$ DPSK transmitted signal can be expressed as $S(t)=\sum_{n=-\infty}^{\infty} a_n\cdot p(t-nT)\cos(w_c t+\theta_n)$ (2)

The DPD detector input $x_k(t)$ of the k^{th} $(k=1,2)$ branch can be expressed as

$$x_k=x_{dk}(t)+x_{fk}(t)+x_{nk}(t)$$

$$x_{dk}(t)=\sum a_n q(t-nT)\cos(w_c t+\theta_n(t))$$

$$x_{fk}(t)=\frac{1}{\beta}\sum_{n=-\infty}^{\infty} a_n\, c(t)\; q(t-nT)\cdot \cos[w_c(t-\tau(t))+\theta_n(t)]$$

$$q(t)=h(t)*p(t),\quad h(t)\text{: channel filter} \quad (3)$$

and $x(t)$ can be expressed by in-phase and quadrature components representing the input to the detector of the signal

$$\begin{aligned} x_k(t) &= x_{ki}(t)\cos w_c t - x_{kq}(t)\sin w_c t \\ &= \sqrt{x_{ki}^2(t)+x_{kq}^2(t)}\cos[w_c t+\tan^{-1}(\frac{x_{kq}(t)}{x_{ki}(t)})] \\ &= R_k(t)\cos[w_c t+\tan^{-1}(\frac{x_{kq}(t)}{x_{ki}(t)})] \quad (4) \end{aligned}$$

We note that the DPD (differential phase detection) output is the phase difference between two consecutive signal samples, $x_k(nT), x_k((n-1)T)$.

The sampled phase $\psi_k(n)$ is reduced by the previous $\psi_k(n-1)$ to obtain the difference $\triangle\psi_k$, $(\psi_k(n)-\psi_k(n-1), \bmod 2\pi)$ of the received signal over an symbol duration

$$\triangle\psi_k=\arg[x_k(nT)\cdot x_k^*((n-1)T)] \quad (5)$$

The *DPD* detector output $\triangle\psi_k$ defined over $[-\pi,\pi)$ is then

$$\triangle\psi_k=\triangle\phi_s+\triangle\phi_{fk}+\triangle\phi_{nk}, \bmod 2\pi \quad (6)$$

As we mentioned before, we employ the dual diversity system, so the resultant combiner output is given by

$$\triangle\psi=w_1\triangle\psi_1+w_2\triangle\psi_2, \bmod 2\pi$$

$$w_1+w_2=1 \quad (7)$$

where w_k, $k=1,2$ are the weights. The output of DPD $\triangle\psi_1$, $\triangle\psi_2$ are distributed around $\triangle\phi_s$ and the probabilities of $\triangle\psi_1$, $\triangle\psi_2$ are monotonically reduced as they deviate from $\triangle\phi_s$. This suggests that the combiner output phase should be inside the region sandwitched between $\triangle\psi_1$ and $\triangle\psi_2$. When $|\triangle\psi_1-\triangle\psi_2|>\pi$ occurs, the combiner phase is outside the defined region. To avoid this problem, we use Adachi, Ohmo and Ikura [4]. In here, we use the weight proportional to the m^{th} power of $R_k(nT)$,

$$w_k=\frac{R_k^m(nT)}{R_1^m(nT)+R_2^m(nT)},\quad k=1,2 \quad (8)$$

Substituting (14) into (13), we have

$\triangle\psi$, which is consisted of transmitted phase difference and sum of the weighted phase noise. (9)

III. Structure of Non-Linear Detector

Using the training mode, we estimate quantiles which are used in VQ (vector quantization) procedure by applying RMSA algorithm and then partition the observation space into Q regions using VQ. For each partition, the partition moments which are required to calculate the coefficients of MMSE by using RMSA.
If we assume that the detected symbols are correct, then we can observe the pure noise samples during the training and transmitting mode, i.e.,

$w_{n-k} = x_{n-k} - a_{n-k}$, where a_{n-k} is detected or known symbol and $k = 1, 2, \ldots, r$. An r-dimensional noise sample space is equiprobably partitioned into a finite number of regions, using a VQ and quantiles obtained by RMSA algorithm in the training mode. The received signals will be grouped accorging to hypothesis which is based on carrier phase and equiprobably partitioned parts of past noise samples. Noise sample w_{n-r} is partitioned, continuing the process until sample w_{n-1} is partitioned. Using the RMSA procedure, we will get the estimated value of mean $\mu_{i,k}$ under the each hypothesis and partitioned cells of past noise samples

$$\mu_{i,k} = E[x_n | VQ(w_{n-1}, w_{n-2}, \ldots, w_{n-r}) \ \varepsilon \ \rho, H_i]$$

(10) where I=0,1 (for hypothesis) and k : level of partition. The average value of estimated threshold can be represented by $th_k = \frac{th_{1,k} + th_{2,k}}{2}$ where

$$th_{j,k} = \frac{\mu_{1,k} + \mu_{0,k}}{2} \quad , \ j = 1, 2 \quad (11).$$

IV. Discussion of Results

Here we use the dependent noise model generated by the auto-regressive (AR) model as a noise component in a rician fading channel, i.e.,

$$n(n) = \rho_1 n_{n-1} + \rho_2 n_{n-2} + \ldots + \rho_m n_{n-m} + v_n$$

(12)

and use the 6^{th} order Chebychev filter to generate ISI and use 125, total number of partitions, and 1000,000 training samples. Through the computer simulation, we demonstrate that the performance of this system is much better than conventional decision feedback equalizer and single sample detector in terms of *Log10(Pe)* as shown in Fig.1(k=22, m=2, rectangular pulse and dependent gaussian noise case). Table1-Table2 show that the amount of improvement in Pe compared to a single sample detector is from 0.5 to 11.2 db under any noise distribution condition, where the improvement (in db) is defined as compared to the single sample detector.

$$\textit{improvement of } P_e(db) = 10 \log_{10} \mathrm{x} \quad \frac{p_e(\textit{single sample detector})}{p_e(\textit{robust} - \textit{equalizer})} \quad (13)$$

We examine non-linear (robust) detector based on an estimated mean value for each hypothesis based on the past partitioned noise sample sets and show better performance as compared to a conventional decision feedback equalizer and to a single sample detector in terms of Pe.
Through the computer simulations, we can get the optimum m,(=2) weight power factor also.

(Note) : NL-D : non-linear detector, Con-EQ : Conventional Decision Feedback Equalizer
Sin : Single Sample Detector
Rec : Rectangular pulse shape

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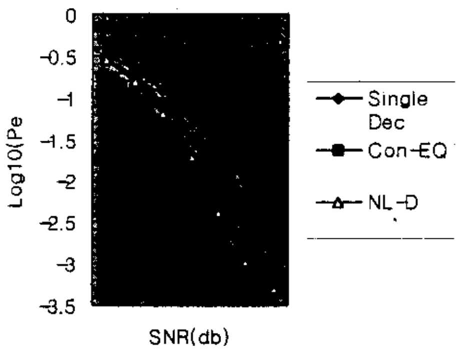
Fig1. Pe of NL-D, k=22, m=2



SNRd		Con. DFB	NL-D
4	Pe	0.5	0.8
7	Pe	0.9	1.4
10	Pe	1.5	2.7
13	Pe	2.6	4.8
16	Pe	4.4	7.3
19	Pe	6.3	10.5

Table 1. Improvement of Pe for Gaussian Noise Using NL-D (Rectangular Pulse, k=22, m=2)

SNRd		Con. DFB	NL-D
4	Pe	0.6	0.7
7	Pe	0.7	1.4
10	Pe	1.2	2.8
13	Pe	2.2	5.0
16	Pe	4.2	7.9
19	Pe	6.8	11.2

Table 2. Improvement of Pe for Riciian Noise Using NL-D (Rectangular Pulse, k=22, m=2)

$$\triangle \psi = \triangle \phi_s + \frac{(\triangle \phi_{f1} + \triangle \phi_{n1}) \cdot R_1^2(nT) + (\triangle \phi_{f2} + \triangle \phi_{n2}) \cdot R_2^2(nT)}{R_1^2(nT) + R_2^2(nT)} \mod 2\pi \quad (9)$$