A NOVEL DECISION-DIRECTED RECEIVER FOR DETECTION OF CPM SIGNALS

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ABSTRACT

A novel Decision-Directed Receiver (DDR) has been arrived at based on heuristic considerations for coherent demodulation of arbitrary Continuous Phase-Modulated (CPM) constructions received over AWGN channel. The complexity of the DDR is a linear function of the observed symbol intervals. However, repeated processing of the received (signal + noise) waveform segments is required. The error probability analysis of the DDR for a subclass of CPM known as CPFSK is presented. It is noted that the low-complexity DDRs are attractive by virtue of their superior error performance relative to the well known Average Matched Filter (AMF) receivers, at least for certain modulation indices, for example h = 0.8.

1. INTRODUCTION

Digital transmission using Continuous Phase-Modulated (CPM) signals is an important signaling technique, having wide spread applications in mobile communications, terrestrial digital radio, and satellite communications. Spectral and power saving properties of CPM are well known, as are demodulation techniques [1-3]. However, the successful application of CPM signaling requires the use of complex signal processing techniques. Furthermore, the CPM constructions are known to offer tradeoffs among power, bandwidth and a third commodity known as the "receiver complexity [4,5]." The optimum receiver structure for CPM is an exponential function of the observed symbol intervals. Thus, in the literature there have been various attempts to find low-complexity receiver structures for demodulation of CPM. Among the earliest efforts is the work of Pelchat and Adams [6], who used the concept of AMF receiver to examine the performance of noncoherent CPFSK in AWGN. Subsequently, this concept of AMF receiver has been employed for detection of a variety of CPM signals [7-10]. Also, the problem of finding low-complexity receivers for CPM has been dealt with in [10-12], where by utilizing the properties associated with the phase structures of CPM, reduced complexity receivers have been suggested. In all these works, four different types of receivers are considered of which two are general receivers that work for all CPM constructions, while the other two work for binary schemes with a modulation index of 0.5. In the paper, quite different to these contemporary approaches a structure based on heuristic considerations has been analyzed in an attempt to arrive at a low-complexity receiver for CPM.

The paper is organized as follows: The background required for the problem is given in Section 2. In Section 3 the decision-directed detection algorithm is given. The error rate analysis of the receiver is presented in Section 4. The associated numerical results with a discussion is given in Section 5. The paper is concluded in Section 6.

2. BACKGROUND

2.1 CPM Signal Description

The general expression for CPM signal over an n-bit interval is given by

$$S(t,\tilde{a}) = \sqrt{2S}\,\mathrm{Cos}(2\pi f_c t+\phi(t,\tilde{a})+\phi_o) \qquad (1)$$

where the information carrying phase is given by

$$\phi(t,\tilde{a}) = 2\pi h\int_0^t \sum_{i=1}^{n} a_i g(\tau-(i-1)T)d\tau \qquad 0 \le t \le nT \qquad (2)$$

and $\tilde{a} = (a_1, a_2,...,a_n)$ is an n-bit uncorrelated equally likely binary sequence with S the signal power per bit interval and f_c the carrier frequency. The modulation index employed is denoted by h, and ϕ_o is the arbitrary constant phase shift which, without loss of generality, can be set to zero for the case of coherent transmission. In eqn. 2 $g(\tau)$ is the frequency pulse lasting T seconds. Defining the baseband phase function by

$$q(t) = \int_0^t g(\tau)d\tau, \qquad 0 \le t \le nT \qquad (3)$$

the information carrying phase in eqn. 2 can be written as

$$\phi(t,\tilde{a}) = 2\pi h \sum_{i=1}^{n} a_i q(\tau-(i-1)T)d\tau$$

$$0 \le t \le nT \qquad (4)$$

As an example of phase function we cite the linear phase (LP) case:

$$q(t) = \begin{cases} 0, & t \le 0 \\ t/2T, & 0 \le t \le T \\ 1/2, & t \ge T \end{cases} \qquad (5)$$

In the case of LP, the frequency is held constant throughout the data interval. For MSK designs, h = 0.5, and for binary CPFSK, h = 0.715 maximize the detection efficiency. A detailed description of CPM signals is given in [1].

2.2 Minimum error rate receiver

The optimum coherent receiver which minimizes the bit error probability observes the received CPM signal contaminated with AWGN, over several bit intervals and makes a decision on the first bit in this interval [7]. The optimum receiver is complex and its precise analysis is too complicated to attempt analytically. The complexity of the receiver grows exponentially as the number of observed symbol intervals. The performance of the receiver is determined in terms of bounds tight at low and high SNR and are used in conjunction. These bounds are used for determination of optimum signal modulation parameters. The high SNR bounds (upper and lower) are determined on the receiver obtained by making high SNR approximation in the likelihood ratio dictating the optimum receiver structure, and the low SNR bound on the receiver obtained by making low SNR approximation in the likelihood ratio. The low SNR receiver structure is relatively far simpler to implement and offers an optimized performance equal to that of PSK (for h = 0.5 and n = 2). Furthermore, for n > 2 there seems to be no improvement in performance in the use of AMF receiver for any modulation index h. Questions such as "Are there modulation indices h in CPFSK designs that in conjunction with AMF receiver yield asymptotic performance improvements as a function of n?" have not been answered in the literature. In the Sections to follow we introduce the concept of decision feedback in the AMF-type of receiver and analyze it for error rate in an attempt to know whether asymptotic improvement in performance is possible with CPFSK designs for arbitrary h as a function of n. It is noted that the detection algorithm works for arbitrary CPM designs, although error analysis is presented for CPFSK only.

3. THE DECISION-DIRECTED ALGORITHM

The received (signal + noise) waveform over n-bit interval is modelled as:

$$r(t) = S(t,\tilde{a}) + n(t), \; 0 \le t \le nT \qquad (6)$$

where $S(t,\tilde{a})$ is the received information-contained CPM signal over n bit intervals with n(t) the AWGN with one-sided spectral density of N_o. The detection strategy consists of repeated processing of r(t) and arriving at an estimate of the data transmitted during the first bit interval. The strategy is clearly explained through the algorithm given below:

$i \leftarrow 1$;
observe r(t) over $0 \le t \le T$ using the receiver of Fig. 1 and obtain estimate $\hat{a}_{i1}$ of data a_i;
[determine $P_e(\hat{a}_{i1})$]

Step 1: $k \leftarrow i + 1$;
$j \leftarrow 1$;

Step 2: observe r(t) over $(k-1)T \le t \le kT$ using estimates $\hat{a}_{pj}$, p=i,i+1,...,k+1 and the receiver of Fig. 1 to obtain estimate of data a_k, i.e. $\hat{a}_{kj}$;
$k \leftarrow k + 1$;
if($k \le i+n-1$) go to Step 2
[we have $(\hat{a}_{ij}, \hat{a}_{i+1\,j}, \ldots, \hat{a}_{i+n-1\,j})$ and $P_e(\hat{a}_{ij})$]
$k \leftarrow j$;
$j \leftarrow j + 1$;
observe r(t) over $(i-1)T \le t \le (i+n-1)T$ using the estimates $\hat{a}_{qk}$, q = i + 1, i + 2, ..., i + n -1 and the receiver of Fig. 2 to obtain $\hat{a}_{ij}$;
[determine $P_e(\hat{a}_{ij})$]

if($P_e(\hat{a}_{i\,j-1}) \le P_e(\hat{a}_{ij})$)go to Step 4;

go to Step 2;

Step 4: $i \leftarrow i + 1$;

go to Step 1;

4. ERROR RATE ANALYSIS OF DDR

The received waveform given by eqn.6 is correlated during the 1st bit interval using the reference signal $y_{i1}(t)=S_1(t,b_1=+1,b_2,\ldots,b_n) - S_1(t,b_1=-1,b_2,\ldots,b_n)$ [i=1 for 1st iteration and $S_1(t,\ldots)$ denoting waveform during 1st bit interval] to generate

$$M_{i1} = \int_0^T r(t)y_{i1}(t)dt \qquad (7)$$

using which an estimate of a_1 i.e. $\hat{a}_{i1}$ is obtained.During the 2nd bit interval the receiver uses a_{i1} and $y_{i2}(t)=S_1(t,\hat{a}_{i1},b_2=1,b_3 \text{ to } b_n) - S_1(t,\hat{a}_{i1},b_2=-1,b_3,\ldots,b_n)$ generates M_{i2} and an estimate $\hat{a}_{i2}$ of data a_2. The process is continued until $(\hat{a}_{i1},\ldots,\hat{a}_{in})$ is obtained. The receiver of Fig.1 is used to obtain these. In general during the jth bit interval

$$M_{ij} = A_{ij} + N_{ij} \qquad (8)$$

is generated where

$$A_{ij} = \int_{(j-1)T}^{jT} S(t,\tilde{a})y_{ij}(t)dt \qquad (9)$$

and

$$N_{ij} = \int_{(j-1)T}^{jT} n(t)y_{ij}(t)dt \qquad (10)$$

in order to obtain estimate $\hat{a}_{ij}$ of data a_j in the ith iteration.

In order to refine the estimate of data a_1, the receiver of Fig.2 is employed and the n-1 estimates $(\hat{a}_{i2},\ldots,\hat{a}_{in})$. The refined estimate is $\hat{a}_{i+1,1}$(2nd iteration). To obtain this estimate the receiver generates

$$M_{i+1,1} = A_{i+1,1} + N_{i+1,1} \qquad (11)$$

where

$$A_{i+1,1} = \int_0^{nT} S(t,\tilde{a})y_{i+1}(t)dt \qquad (12)$$

and

$$N_{i+1,1} = \int_0^{nT} n(t)y_{i+1}(t)dt \qquad (13)$$

Using $\hat{a}_{i+1,1}$ and processing $S(t,\tilde{a})$ during 2nd bit interval (now using receiver of Fig.1) 2nd estimate of a_2, i.e. $\hat{a}_{i+1,2}$ is obtained. Likewise, $\hat{a}_{i+1,p}$;p=3,4,...,n are obtained. Again $(\hat{a}_{i+1,2},\ldots,\hat{a}_{i+1,n})$ in conjunction with receiver of Fig.2 next refined estimate of a_1 is obtained, i.e. $\hat{a}_{i+2,1}$. The processing is continued of stopped by checking at every refinement (using receiver of Fig.2) whether there is any improvement or not. Denoting $\tilde{a} = (a_1,A_k)$ and $C_q = (\hat{a}_{1,1},\ldots,\hat{a}_{1,n})$, the conditional probability of error of $\hat{a}_{21}$ is given by

$$P_e(\hat{a}_{21}/A_k,C_q) =$$

$$\Pr.[M_{21}<0/a_1=+1,A_k,C_q]P(a_1=+1,A_k,C_q)$$

$$+\Pr.[M_{21}>0/a_1=-1,A_k,C_q]P(a_1=-1,A_k,C_q) \qquad (14)$$

It can be shown that

$$P_e(\hat{a}_{21})= \sum_{A_k} \sum_{C_q} P(A_k,C_q/a_1=+1)$$

$$\times \Pr.[M_{21}<0/a_1=+1,A_k,C_q] \qquad (15)$$

Using likewise technique general expression $P_e(\hat{a}_{i1})$ can be derived [5]. In the next Section numerical results for the performance of DDR for CPFSK is presented and discussed.

5. NUMERICAL RESULTS

The performance of the Decision-Directed algorithm for detection of CPFSK at the second refinement in deciding about the data transmitted during the first interval was determined by programming expressions given in the previous Section. The error probability performance of the DDR, $P_e(\hat{a}_{12})$, is a function of: i) E_b/N_o, Signal-to-Noise Ratio; ii) h, the modulation index; and iii) n, observation length of DDR. The modulation index h that should be chosen for a given n and E_b/N_o, is the one that minimizes the $P_e(\hat{a}_{12})$. In Fig.3, plots of E_b/N_o vs. Optimum h are shown for n = 2,3,and 4. The performance of the DDR for observation lengths 2T and 4T sec. are shown in Figs. 4 and 5, respectively, for CPFSK with h=0.8. In these figures performances of PSK, AMF receiver for CPFSK with h=0.8, and DDR for CPFSK with optimum h (function of E_b/N_o) are also shown.

From Fig.4, it is observed that 2-bit decision-directed receiver (for CPFSK; h=0.8) provides an SNR gain of nearly 1.25 dB relative to the 2-bit AMF receiver (for CPFSK; h=0.8). However, with the latter receiver, decision about the data a_i is available after 2T secs, whereas with DDR decision is available only after 3T secs. The performance of the 2-bit DDR for CPFSK (h=0.8) is worse compared to PSK by nearly 0.5dB.

By going to 3T observation from 2T, the performance of the DDR becomes worse, for CPFSK with h=0.8, by nearly 1 dB. However, the performance of the 3-bit DDR remains superior to the performance of the 3-bit AMF receiver for CPFSK with h=0.8. The performance of the decision directed receiver deteriorates further for n=4, but again maintains superiority in performance compared to the performance of the 4-bit AMF receiver for CPFSK with h=0.8.

It is noted that the performance of the DDR for CPFSK with h=0.5, i.e. MSK, for 2T observation length is inferior to that of the performance of the 2-bit AMF receiver for CPFSK with h=0.5 by nearly 1.5 dB, and with increased observation lengths the performance of DDR deteriorates further. The 2-bit DDR is superior in terms of SNR by about 1.25 dB compared to the 2-bit DDR for CPFSK with h=0.5.

Finally, it noted that for large SNRs (≅ 18 dB) the 2-bit DDR for CPFSK (h=0.8) performs nearly as good as that of 2-bit AMF receiver for CPFSK with h=0.5.

6. CONCLUSIONS

In the paper a novel decision-directed receiver based on heuristic considerations has been presented, for coherent demodulation of arbitrary CPM constructions received over AWGN channel. From the performance analysis carried out for CPFSK, a subclass of CPM, it is clear that the DDR has certain interesting features. For example, the 2-bit DDR can outperform 2-bit AMF receiver, for CPFSK with h=0.8, by nearly 1.25 dB. Furthermore, our results show that CPFSK designs exist with h not equal to 0.5 that with only 2T observation provide performance equal to that of MSK, of course for large SNR.

While our experience with DDR is based on only second estimates, it is anticipated that the performance would improve further with higher order estimates, for at least certain CPFSK designs. Our conviction is that heuristic design of receivers for CPM could be highly attractive from the viewpoint of reduction in complexity and also to achieve comparable performance with that of the optimum.

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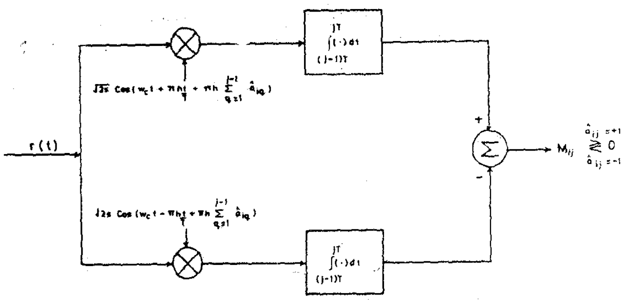

FIG. 1 DECISION DIRECTED RECEIVER FOR OBTAINING ESTIMATES $\hat{a}_{11}$ AND $\hat{a}_{ij}$; i = 1,2,.... ; j = 2,3,....,n

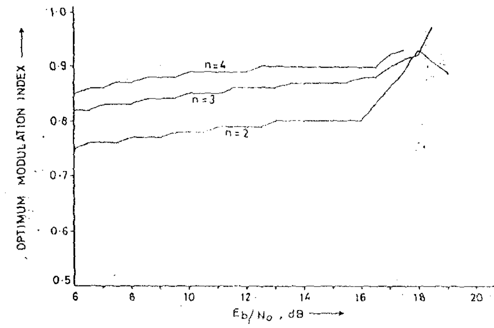

FIG. II-3 : OPTIMUM CPFSK SYSTEMS AS A FUNCTION OF SNR AND OBSERVATION LENGTH FOR DECISION DIRECTED DETECTION

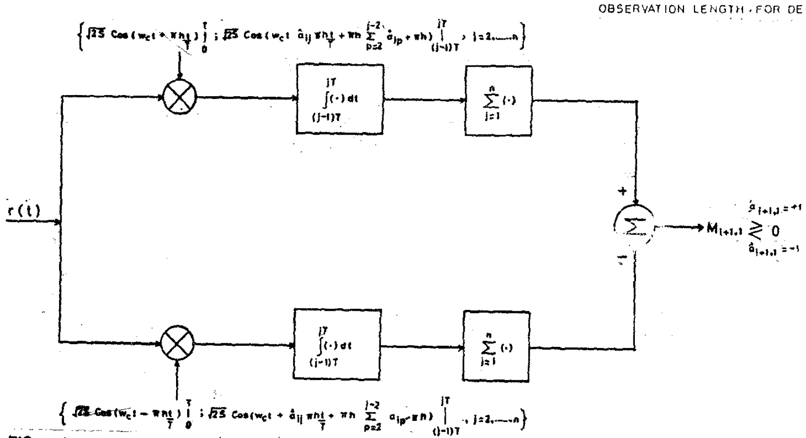

FIG. 2 DECISION DIRECTED RECEIVER OBTAINING ESTIMATES $\hat{a}_{ij}$; i = 2,3,......

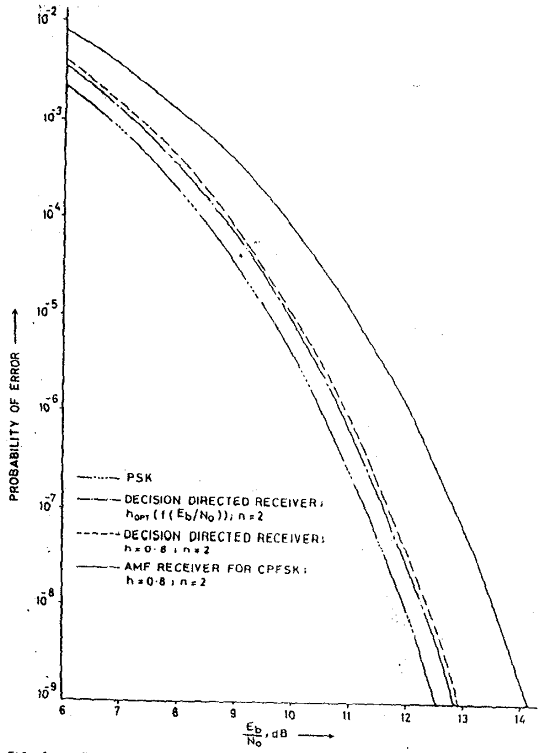

FIG. 4 : PERFORMANCE OF 2-BIT OBSERVATION DECISION DIRECTED RECEIVER ($P_e(\hat{a}_{21})$) FOR CPFSK (h = 0.8)

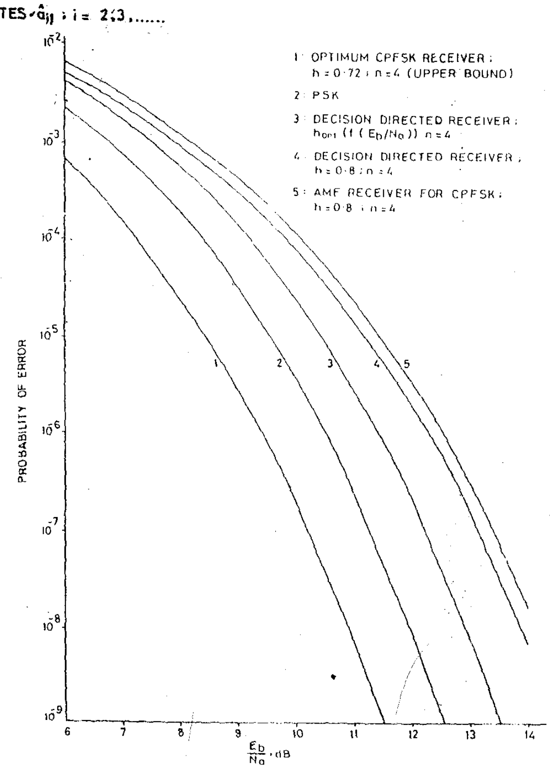

FIG. 5 : PERFORMANCE OF 4-BIT OBSERVATION DECISION DIRECTED RECEIVER ($P_e(\hat{a}_{21})$) FOR CPFSK (h = 0.8)