Efficient Preamble Design for Digital DMSK Packet Synchronization

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Abstract

In this paper, we study the preamble design for a fully digital feedforward receiver for minimum shift keying modulation and asynchronous burst mode transmission. Timing error and frequency offset are estimated simultaneously and corrected subsequently. Frame synchronization is based on tentative bit demodulation and bit correlation.
An optimal preamble has to be optimized with respect to best timing error and frequency offset estimation <u>and</u> to best performance of the frame synchronization algorithm. Optimal preambles with different lengths of the synchronization word and filter lengths of the estimation filter are given. The performance difference between optimally and sub-optimally designed preambles is assessed by computer simulations.

1 Introduction

Many communication systems operating in a mobile radio environment transmit packets of information. In such systems frame synchronization is a necessary issue. Much work has been devoted to study the performance of frame sync rules for linear modulations in additive white Gaussian noise [1,2,3], on channels with intersymbol interference [4] and on flat-fading channels [5]. In [6,7] bit sequences are presented which result in best frame sync performance if the synchronization word (unique word; UW) is preceded by a deterministic bit pattern used for carrier and timing recovery, for example. Those unique words yield the smallest number of lowest crosscorrelation sidelobes if the correlation rule is applied. The idea is extended to the derivation of maximum-likelihood frame synchronization rules for spontaneous packet transmission in [3].
However, all the work on frame sync mentioned so far assumes at least ideal timing and frequency synchronization prior to the frame synchronization unit. In this paper, a feedforward receiver is considered where the estimator for the timing error ($\hat{\varepsilon}$) and the frequency offset ($\Delta\hat{f}T$) also uses the symbols of the frame sync word. This calls for a preamble design with a joint optimization of the preamble symbol pattern with respect to

1. best performance of the estimation of $\hat{\varepsilon}$ and $\Delta\hat{f}T$ and
2. best performance of the frame synchronization rule.

The paper is organized as follows. First, the digital feedforward receiver structure is introduced. Then, a two-step optimization of the preamble design is presented. The paper closes with simulation results and a short summary.

2 Digital Feedforward Receiver

We consider binary transmission with MSK modulation where the signal in baseband is given by

$$\begin{aligned} s(t) &= e^{j(\phi(t-\varepsilon T)+\Delta\omega t+\theta)} \\ \phi(t) &= 2\pi h \sum_i b_i q(t-iT) \end{aligned} \quad (1)$$

ε denotes the fraction of a symbol duration T by which received symbols are time shifted with respect to the original symbol boundaries. $\Delta\omega$ is the frequency offset, h the modulation index ($h = 0.5$ for MSK), $b_i \in [-1, 1]$ the data symbols and $q(t)$ the phase response of the CPM modulator. The receiver structure is shown in fig. 1. The incoming signal is filtered and limited at an intermediate carrier frequency (IF) to perform automatic gain control (AGC). After mixing to baseband, the signal is sampled with a local free running clock. A deterministic symbol pattern at the beginning of each burst is exploited for a coarse automatic frequency control (CAFC). Behind the CAFC and predetection filtering the signal is given by

$$\begin{aligned} z_{k,i} &= z(t=(k+i/N)T) \\ &= e^{j\left\{\phi(kT+\frac{iT}{N}-\varepsilon T)+\Delta\omega(k+\frac{i}{N})T+\theta\right\}} + n_{k,i} \end{aligned} \quad (2)$$

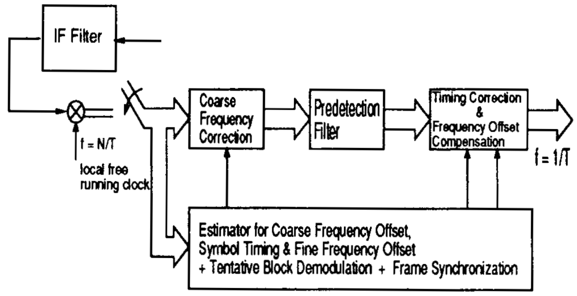

Figure 1: receiver structure

with $0 \leq i \leq N-1$. Here, we assume that the filtering effect of the predetection filter is negligible. θ denotes a constant phase offset and $n_{k,i}$ are noise samples. The expectation of the fourth-order nonlinear transformation of

$$c_{k,i}^2 = z_{k,i}^2 z_{k-1,i}^{*^2} \tag{3}$$

is given by [8,9]

$$\begin{aligned} \nu_{k,i} &= E\left\{c_{k,i}^2\right\} = E\left\{z_{k,i}^2 z_{k-1,i}^{*^2}\right\} \\ &= -\frac{1}{2}\left(1+\cos\left(2\pi\varepsilon - 2\pi\frac{i}{N}\right)\right) e^{j2\Delta\omega T} \end{aligned} \tag{4}$$

assuming white data symbols and white noise. Due to oversampling with rate $f_s = N/T$ there are N samples for each symbol. The signal sample with index $(k,i) = \hat{i}_k$ represents the k-th symbol with the smallest residual timing error if the decision variable $|\nu_{k,i}| > |\nu_{k,j}|$ for all $j : j \neq i$. Timing recovery is performed independently of offset estimation by decimating the signal $z_{k,i}$ to symbol rate $f_s = 1/T$, i. e. a decimator chooses the sample $z_{k,i} = z_{k,\hat{i}_k}$ from N possible symbol samples.
The investigated system uses a burst structure shown in fig. 2. The symbol pattern for CAFC is followed by the unique word of length L_W which is embedded in the preamble part of length $L+1$ used for timing and fine frequency offset estimation. The timing and frequency offset estimator is a filter of length $L = L_W + L_X - 1$ working on a block of $L+1$ incoming signal symbols. During each symbol interval such a block is coarsely corrected in frequency by the estimate of the CAFC unit. On the basis of the frequency corrected block an estimate for the frequency offset $(\Delta\hat{f}T)_k$ and the index $\hat{i}_k$ of the symbol sample with the smallest timing error are generated. With these estimates, L_W subsequent symbols $(z_{k,\hat{i}_k}, \ldots, z_{k+L_W-1,\hat{i}_k})$ are demodulated tentatively. The signal is demodulated differentially coherent, since in this case no carrier phase estimation unit is necessary. The demodulated bits $\tilde{b}_k$ are correlated with the frame sync word $(w_i \in [-1,1])$.

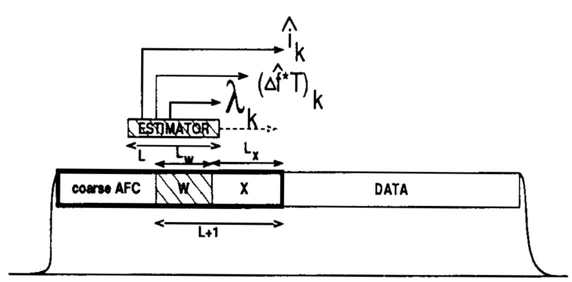

Figure 2: burst structure

$$\lambda_k = \sum_{i=0}^{L_W-1} \tilde{b}_{k+i} w_i \tag{5}$$

A full correlation is performed each symbol interval. If the correlation exceeds a given threshold, $\lambda_k > \delta$, frame synchronization is attained and the estimates for frequency offset and symbol timing are kept constant during the demodulation of the data portion of the packet. Tentative demodulation and full bit pattern correlation in each symbol interval is necessary, because there is only one time instance when the estimator filter covers the preamble part $L_W + L_X$ which is optimized for best offset and timing estimator performance. One set of estimates $\hat{i}_k, (\Delta\hat{f}T)_k$ should be used for the demodulation of the whole packet. The decision which estimates at which time instance k should be kept constant during the duration of a whole packet cannot be taken prior to successful frame synchronization. Therefore, frame sync with tentative symbol demodulation has to be performed in each symbol interval until the unique word has been found.

3 Optimal Preamble Design

In [6], binary frame synchronization sequences are given which provide best synchronization performance if the unique word is preceded by a specific bit pattern, and the synchronization rule is based on the correlation of bit patterns. Those binary sequences were found by an exhaustive computer search and are characterized by lowest absolute valued correlation sidelobes. In our case there are two optimization steps.

1. optimization of frame sync performance

2. optimization of timing recovery performance

As explained in the previous section frame synchronization is based on tentatively demodulated bits. Thus, the frame synchronization metric λ_k can take values from $[-L_W, L_W]$ for a unique word of length L_W. Following the idea of [6] the unique word is chosen according to conditions on the cross correlation between the unique word and the received demodulated bit sequence consisting of symbols from the start-up symbol pattern and the unique word. The optimal unique words for different startup symbol patterns are summarized in the next table. They have been found by an exhaustive computer search and are optimized with respect to the following conditions.

1) The maximum value of the cross correlation sidelobes becomes minimum. It is denoted by q_{max}.

2a) The number of sidelobes with value q_{max} becomes minimum.

2b) The maximum absolute value of the cross correlation sidelobes and their number becomes minimum. They are denoted by $|q_{max}|$ and $\#|q_{max}|$.

The following table lists optimal binary frame sync sequences with respect to a specific preceding bit pattern. ($a = q_{max}, b = \#q_{max}, c = |q_{max}|, d = \#|q_{max}|$)

start seq.	L_W	Hex sequence	a	b	c	d
1111	8	12	0	1	4	2
1111	8	A3	0	5	2	3
1010	8	25	2	1	2	3
1010	8	25	2	1	2	3
1100	8	D7	2	1	2	2
1100	8	D7	2	1	2	2
start seq.	L_W	Hex sequence	a	b	c	d
1111	16	0C95	0	1	6	2
1111	16	A86C	0	7	2	9
1010	16	B670	2	2	2	8
1010	16	B40C	2	3	2	5
1100	16	0EB3	2	2	4	1
1100	16	41D2	2	5	2	8
start seq.	L_W	Hex sequence	a	b	c	d
1111	20	0AC98	0	1	6	3
1111	20	A8793	0	12	4	2
1010	20	2197A	2	3	2	9
1010	20	2197A	2	3	2	9
1100	20	43CBB	2	3	4	2
1100	20	47892	2	6	2	11
start seq.	L_W	Hex sequence	a	b	c	d
1111	24	0B19A2	0	1	6	5
1111	24	33452F	0	6	4	9
1010	24	8F112D	2	5	2	15
1010	24	8F112D	2	5	2	15
1100	24	414F93	2	4	4	2
1100	24	4AFB3C	2	5	2	14
start seq.	L_W	Hex sequence	a	b	c	d
1111	28	06CA9CE	0	2	6	8
1111	28	622DA3C	0	7	6	1
1010	28	2671FA5	2	5	4	4
1010	28	2905CB9	2	8	2	18
1100	28	649F953	2	5	4	2
1100	28	649F953	2	5	4	2
start seq.	L_W	Hex sequence	a	b	c	d
1111	32	30C9E8AD	0	3	6	6
1111	32	5581CE49	0	9	6	2
1010	32	2670D2FD	2	8	4	4
1010	32	38D4B27F	2	10	4	2
1100	32	5C57B924	2	7	4	3
1100	32	5C57B924	2	7	4	3

The second optimization step focuses on the timing recovery process. Timing recovery is performed by decimating the oversampled symbol sequence to symbol rate with the smallest residual timing error. A residual timing error of $\varepsilon > 1/8$ significantly degrades the bit error rate performance. Fig. 3 shows the simulated BER for residual timing errors $\varepsilon_{res} = 0 - 1/4$ for a predetection filter bandwidth of $BT = 1.21$. Let

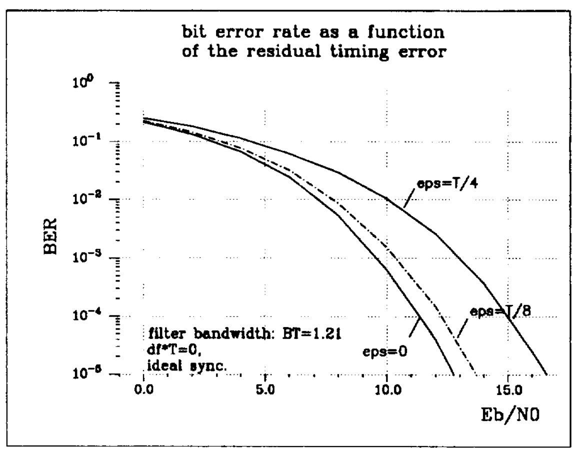

Figure 3: influence of a residual timing error

p_{dec} denote the probability that the decimator does not select the correct sample with the smallest residual timing error. Then, the average BER (P_b) for

$\varepsilon = 0$ is given by

$$\bar{P}_b \approx P_b(\varepsilon_{\text{res}} = 0)[1 - p_{\text{dec}}] + P_b(\varepsilon_{\text{res}} = 1/4)[p_{\text{dec}}] \quad (6)$$

There is only a small increase in the average BER for small values of p_{dec}. But the effect of incorrect decimation on packet transmission is more pronounced, since an estimated decimating point is fixed for a whole packet.

The second optimization step results in extending the unique word of length L_W by a bit sequence X of length L_X so that the error probability of the decimating process is minimized. Neglecting the influence of predetection filtering the magnitude of the output of the estimator filter is

$$\begin{aligned} |\nu_{k,i}| \;=\; & \frac{1}{L}|(L_{-1-1} + L_{11}) + \quad (7) \\ & (L_{-11} + L_{1-1})\cos(2\pi\varepsilon - 2\pi i/N) + \\ & j|L_{-11} - L_{1-1}|\sin(2\pi\varepsilon - 2\pi i/N)| \end{aligned}$$

L_{mn} denotes the number of bit pairs mn inside the estimator filter length L. If the bit pair consisting of the last bit of the startup sequence and the first bit of the unique word equals the last bit pair of the extended sequence X, we can assume $-1/(2N) \leq \varepsilon \leq 0$ without loss of generality. Then, $\hat{i}_k = 0$ is the index of the symbol sample with the smallest residual timing error. Defining the difference of the decision variables $\Delta_{ij} = |\nu_{k,i}| - |\nu_{k,j}|$, the symbol pattern inside the estimator filter has to be chosen according to the rule

$$\textbf{maximize: } \text{Minimum}(\Delta_{01}, \Delta_{02}, \ldots, \Delta_{0(N-1)}) \quad (8)$$

Let $L_{-1-1} + L_{11} = L_{even}$ and $L_{-11} + L_{1-1} = L_{odd}$. For $\varepsilon = 0$ the maximization is performed by setting $\Delta_{01} = \Delta_{0(N/2)}$ and deriving the following rule for L_{even}, L_{odd} and $L_{-11} = L_{1-1}$.

$$L_{even} = \frac{L_{odd}}{2}(1 - \cos\frac{2\pi}{N}) \quad (9)$$

This provides a condition for the length of the estimator filter which has to be

$$L = \left\lceil \left(1 + \frac{2}{1 + \cos(2\pi/N)}\right) \right\rceil L_{even} \quad (10)$$

$\lceil x \rceil$ gives the next integer value with $I \geq x$. For an oversampling factor of $N = 4$, the maximization results in

$$\text{Minimum}(\Delta_{01}, \Delta_{02}, \Delta_{03}) = 2/3 \quad (11)$$

for $L_{odd} = 2L_{even}$ and $L_{-11} = L_{1-1}$. If $\varepsilon = 1/8$ and $N = 4$, $\Delta_{01} = 0$ but $\Delta_{02} = \Delta_{03} = 2/3$. Thus, the difference between the decision variable of the index indicating the signal sample with the smallest residual timing error and the one with a residual timing error larger than 1/4 is always at least 2/3.

For example, the best frame sync word for a preceding 1111-pattern, 30C9E8AD, has to be extended by the pattern 2AA (Hex)=01010101010 to attain optimal estimator performance with an estimator filter length $L = 42$. If the CAFC unit uses a 1010 sequence and the optimal unique word fulfills the condition $L_{odd} < 2L_{even}$ the CAFC preamble symbols equal the pattern X necessary for best timing recovery performance. In this case the timing estimator can also work on the symbols preceding the unique word and thus allowing to shorten the complete preamble by L_X symbols.

4 Simulation Results

In the previous section the influence of a predetection filter was neglected in the derivation of the design criteria for an optimal preamble. Fig. 3 shows the func-

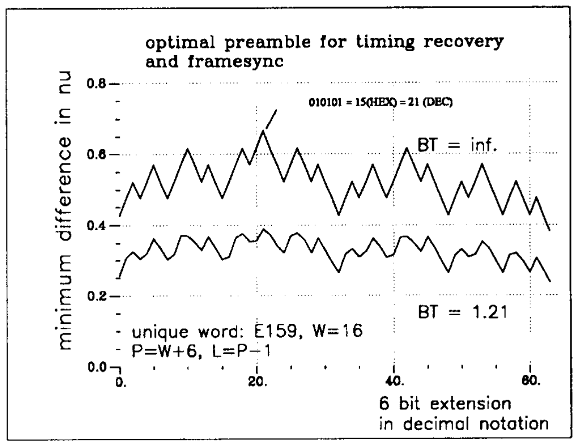

Figure 4: optimal extension

tion Minimum$(\Delta_{01}, \Delta_{02}, \Delta_{03})$ for the pattern E159 extended by a six bit sequence with and without the influence of a predetection filter. The x-axis covers all 64 possible six bit sequences in decimal notation. Due to predetection filtering $Min(\Delta_{0,i})$ is reduced. However, a carefully designed preamble according to section 3 is also a good choice in a system with narrow predetection filters. Fig. 5 reflects the influence of a frequency offset on the estimator performance. The investigated unique word is 30C9E8AD. The estimator filter length is $L = 42$, $\varepsilon = 0$ and $E_b/N_0 = 8$dB. The sequence 30C9E8AD_opt fulfills the condition $2L_{\text{even}} = L_{\text{odd}}$. In the sequence 30C9E8AD_white $L_{\text{even}} = L_{\text{odd}}$ holds and in the third pattern the unique word is extended by 11111111111. It can be observed that the advantage of the optimized pattern diminishes for increasing frequency offsets. However, this effect is mitigated by choosing a wider predetection filter.

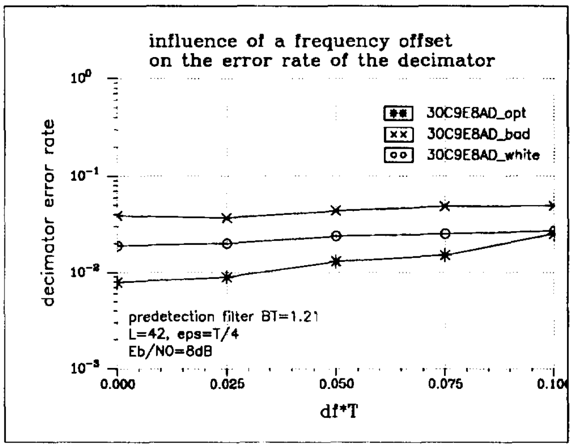

Figure 5: influence of a frequency offset

Fig. 6 compares three patterns based on the unique word 30C9E8AD over the E_b/N_0. The optimized pattern performs best in terms of decimator rate and BER.

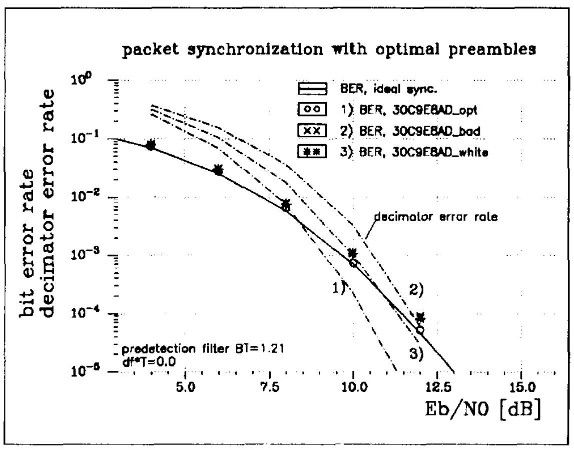

Figure 6: bit error rate, decimator error rate

The predetection filter bandwidth is $BT = 1.21$ [10], the initial frequency offset $\Delta fT = 0$ and the initial timing error $\varepsilon = 0$. From Fig. 6 it is obvious that a sub-optimally designed preamble degrades the performance of the estimator and thus raises the bit error rate and consequently also the error rate of the frame synchronization.

5 Summary

The preamble design for a digital receiver for burst mode transmission with joint estimation of symbol timing error and frequency offset and tentative frame synchronization follows two separate optimization rules. First, the unique word must have the lowest correlation sidelobes with respect to a given startup symbol sequence. Second, the symbol pattern inside the estimator filter must assure best estimator performance. It was shown that for a sampling rate $f_s = 4/T$ the number of bit transitions must be $L_{odd} = 2L_{even}$ and $L_{-11} = L_{1-1}$. Computer simulations showed the performance degradation for sub-optimal preambles.

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