

A HIGH-PERFORMANCE CPFSK PHASE-RECONSTRUCTION RECEIVER WITH VITERBI DECODING

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ABSTRACT

A number of modern wireless systems use modulation schemes like CPFSK, GMSK and DPSK, where the information to be transmitted is exclusively coded in the phase change of the RF-signal. Thus the detection of the received data can be performed by using only the zero crossings of the hard-limited IF-signal. With respect to implementation costs and power consumption this is a very attractive receiver concept. A superior performance compared to conventional zero crossing detectors can be obtained by reconstructing the phase trajectory from the hard-limited I- and Q-components. This can be achieved by methods of digital signal processing. Such a detector however is considerably more complex with respect to the implementation requirements for the DSP part. In this contribution it will be shown that the complexity can be drastically reduced by using proper approximation methods for phase reconstruction. A superior performance can be obtained by combining the phase reconstruction algorithm with a Viterbi detector. The performance is very close to that of an optimum detector based on MLSE.

1. INTRODUCTION

Continuous Phase Frequency Shift Keying and Differential PSK are quite often used for modulation in modern short-range wireless systems, e.g. according to the Bluetooth standard [1]. The reason for this is its inherent noise immunity and the possibility to use high efficiency non-linear power amplifiers on the transmitter side. Due to the advancement of digital signal processing receiver architectures with very low intermediate frequencies (IF) became more and more popular in the last years [2,3]. In the respective implementations the analogue IF-signal must be digitized by additional A/D-converters. FM demodulation and signal detection can be performed then by methods of digital signal processing [2]. Compared to analogue solutions performance improvements can thus be obtained, with the drawback of higher implementation costs and higher power consumption for the A/D-converters however.

For phase modulation schemes with constant envelope signals like GFSK or DPSK the A/D-converters can be replaced by simple limiters thus minimizing implementation costs and power consumption. It has been shown that for these modulation schemes no severe performance losses have to be taken into account due to the limiters, if sophisticated methods for digital post processing are used [4-6].

The received phase-modulated analogue IF-signal with modulation index η and symbol sequence $\{d_k\}$ can be described by:

$$r_{IF}(t) = A \cos \left(2\pi f_{IF} t + \phi_0 + \pi \eta \int_{-\infty}^t \sum_{k=-\infty}^{\infty} d_k \cdot g(t' - kT_{sym}) dt' \right)$$

with intermediate frequency f_{IF} , symbol duration T_{sym} , $g(t)$ the pulse shape of the filter, and ϕ_0 the initial phase. This signal is transformed into a square wave signal by the zero crossing detector.

Fig. 1 shows the respective receiver back-end architecture. The analogue front-end performs down-conversion of the broadband RF-signal to the low IF and generates the quadrature components I and Q. For Bluetooth typical values for the low IF are 1 or 2 MHz [2, 7]. Then channel-selection filtering is performed by an analogue band-pass filter. The 1-bit A/D-converters, operating with an over-sampling ratio of OSR with respect to the symbol rate, digitize the filtered IF signal to the binary I- and Q-components. These binary signals still contain the information about the phase $\phi(t)$ of the received signal in their zero transitions.

Using the zero crossings of these binary signals, the instantaneous phase of the base-band signal can be reconstructed by sophisticated methods of digital signal processing. After phase-reconstruction synchronization, down-sampling to the symbol rate, demodulation and detection can be performed in the base-band digitally.

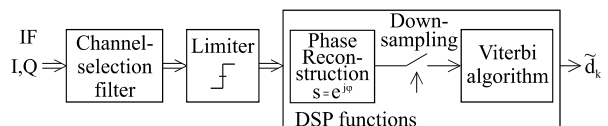


Fig. 1. Receiver for phase-modulated signals with hard-limiting of the IF-signal.

2. METHODS FOR PHASE RECONSTRUCTION

The input signal of the limiters can be described by:

$$\begin{aligned} s_{IF}(t) &= A(t) \cdot e^{j2\pi f_{IF}t} \\ &= |A(t)| \cdot e^{j\Phi(t)} \cdot e^{j2\pi f_{IF}t} \end{aligned}$$

Due to the limiter operation the information about the signal amplitude $|A(t)|$ gets lost. With perfect phase reconstruction the respective base-band signal is given by:

$$s_B(t) = e^{j\Phi(t)}$$

The phase reconstruction requires that the sampling rate of the limiters is considerably higher than the intermediate frequency. A reasonable value for the over-sampling ratio is e.g. 104. Based on the time instants t_j of the zero-crossings the respective instantaneous phase values $\Phi(t_j)$ have to be determined. This can be done e.g. for CPFSK by observing that the instantaneous phase of the inphase and quadrature components increases by π with each zero-crossing if the intermediate frequency f_{IF} is chosen appropriately. At the zero-crossing time instants t_j following relationships hold:

$$\begin{aligned} r_{IF}(t_j) &= \cos(2\pi f_{IF}t_j + \varphi_0 + \Phi(t_j)) = 0 \\ \Rightarrow 2\pi f_{IF} \cdot t_j + \varphi_0 + \Phi(t_j) &= (2 \cdot j - 1) \cdot \frac{\pi}{2} \end{aligned}$$

Thus the respective phase values of the base-band signal are given by:

$$\begin{aligned} \Phi(t_j) &= (2 \cdot j - 1) \cdot \frac{\pi}{2} - 2\pi f_{IF} \cdot t_j - \varphi_0 \\ &= \Phi(t_{j-1}) - 2\pi f_{IF} \cdot (t_j - t_{j-1}) + \pi \end{aligned}$$

where the intermediate frequency f_{IF} and the starting phase value φ_0 are known. Only one multiplication and two additions are required to determine the respective phase value from the previous one, each time a new zero crossing is detected. The time deltas $(t_j - t_{j-1})$ can be determined by a counter.

The instantaneous phase trajectory can be reconstructed from the values $\Phi(t_j)$ using an iterative algorithm [5, 6]. The respective block diagram is shown in fig. 2. This module reconstructs the instantaneous phase $\Phi(t)$ of the modulated base-band signal by methods of digital signal processing.

The problem of reconstructing the instantaneous phase $\Phi(t)$ can thus be viewed as a reconstruction problem from irregular samples [8]. If the instantaneous phase is band-limited with cut-off frequency Ω the iterative reconstruction algorithm can be described by following relationship:

$$\Phi_n(t) = \Phi_{n-1}(t) + \lambda \cdot \sum_{j=-\infty}^{\infty} [\Phi(t_j) - \Phi_{n-1}(t_j)] \cdot \frac{\Omega}{\pi} \cdot \text{sinc} \frac{\Omega}{\pi} (t - t_j)$$

Here, λ denotes a relaxation parameter and $\text{sinc}(u) = \sin(\pi u) / (\pi u)$. With

$$h(t) = \frac{\Omega}{\pi} \cdot \text{sinc} \frac{\Omega}{\pi} (t - t_j)$$

which is the impulse response of an ideal low-pass filter with cut-off frequency Ω .

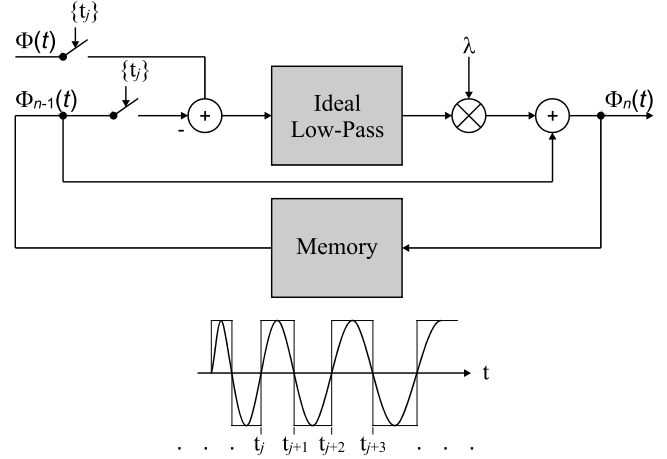


Fig. 2. Block diagram of the phase reconstruction module.

After phase reconstruction the IF-signal is down-converted into the base-band. Then sample rate reduction to the symbol rate and detection can be performed, as already described in section 1. Even though the iterative phase reconstruction process usually results in a good approximation within several iteration steps, the computational load is quite large. Since the impulse response of the sinc-filter decreases only slowly, a large number of coefficients and therewith computations are required for the low-pass filter in fig. 2. Furthermore the memory must be designed to store a long sequence of phase values.

In order to reduce the requirements we will consider thus simpler methods for phase reconstruction in the following. Based on the values $\Phi(t_j)$ a staircase function can be generated for the phase trajectory which can be smoothed then by a low-pass filter. If the parameters of this post-filter are properly chosen a sufficient phase approximation can be already obtained using a fairly simple FIR filter. Fig. 3 shows a phase trajectory for a GFSK-modulated signal and the approximation by a staircase function with and without post-filter. The filtered version is already in good agreement with the phase trajectory to be approximated when using an FIR-filter with 128 taps and a cut-off frequency of 0.8 times the bit-rate (see curve 4 in fig. 3). The time shift of the phase is due to the constant group delay of the post filter and can be compensated.

Even better results can be obtained when down-converting the binary I- and Q-signals into the base-band before phase-reconstruction. This can be achieved by a quadrature mixer. Behind the mixer the I- and Q-components are low-pass filtered by the FIR filter already described so that higher frequency components due to the limiter are suppressed. Thus the I- and Q-components of a linear receiver are approximated. The phase can be reconstructed from these quadrature components by well-established methods. The respective phase trajectory is also shown in fig. 3 by curve 3. In order to achieve a superior performance of the receiver, behind phase-reconstruction well-established methods of digital signal-processing like Maximum Likelihood Sequence Estimation can be applied, as described in the following section.

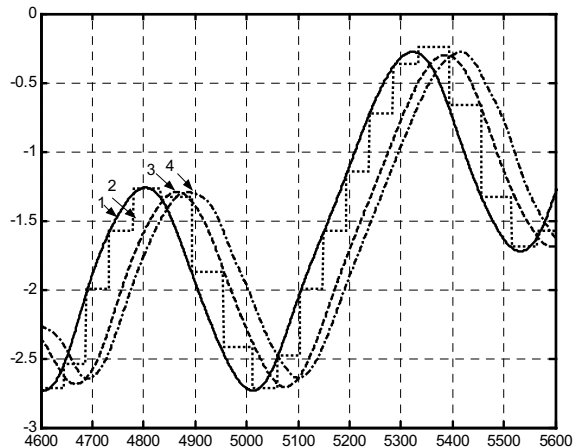


Fig. 3. Phase plot of a GFSK-modulated signal, ideal phase trajectory (curve 1), staircase approximation of phase obtained from zero-crossings (curve 2), filtered staircase function (curve 4), phase trajectory obtained from filtered binary I,Q-components (curve 3).

3. DETECTION BASED ON VITERBI ALGORITHM

The output data of the reconstruction module still suffer from Inter-Symbol Interference (ISI). Using an efficient search algorithm an optimal path through the trellis diagram can be determined. We have used the well-known Viterbi algorithm for our application [9, 10]. The Gaussian filter for Bluetooth has a length of about 3 bit periods, thus a Viterbi algorithm with 2 state variables d_{k-2} and d_{k-1} is sufficient. The state variables correspond to the bits to be detected and can be either +1 or -1. The respective four states and the state transitions are shown by the diagram in fig. 4.

Every bit period the Viterbi algorithm must calculate new branch metrics. These branch metrics BM_i are obtained from the Euclidian distance between the phase difference in a 3 bit interval based on the received data and the eight possible phase deltas corresponding to the 3 consecutive bits $\{d_{k-2}, d_{k-1}, d_k\}$. The values BM_i can be also obtained according to following relationship:

$$BM_i = \exp[-j\Phi_{k-2}^c] \cdot \exp[j\eta\pi(d_{k-2} + d_{k-1} + d_k)] \cdot \exp[j\Phi_k^c]$$

where Φ_k^c is the phase trajectory generated from the detected data bits 1 to k. If the detected data bit d_{k+1} is a -1, $\eta\pi$ is subtracted from the old phase trajectory otherwise $\eta\pi$ is added thus generating the new value Φ_{k+1}^c .

Adding these values BM_i to the 4 accumulated branch metrics ABM_i , 8 new accumulated branch metric values are generated. As shown in figure 4, two paths enter each state. According to the Viterbi algorithm only the path with the smallest accumulated metric has to be saved, whereas the other one can be discarded.

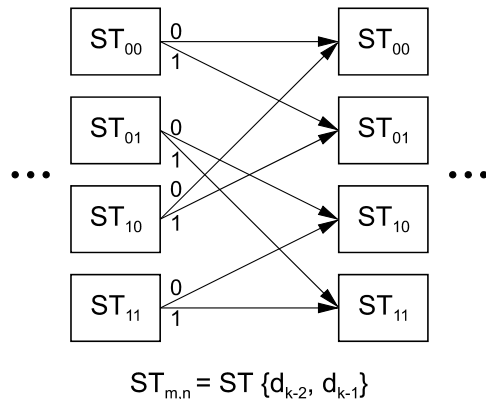


Fig. 4. State diagram for Viterbi algorithm.

Thus for each state an accumulated branch metric value is stored. In the next step the state with the smallest accumulated metric is determined and a decision about bit d_{k-2} is made. The algorithm is initialized by setting the first sample of the phase to a constant value and all states to zero. The algorithm starts at the fourth sample to determine the first bit.

4. PERFORMANCE RESULTS

In order to analyze the performance of the different receiver concepts simulation models based on MATLAB for a Bluetooth system with gross bit rate 1 Mbit/s and a nominal modulation index of $\eta=0.315$ have been used. The time bandwidth product is set to $BT=0.5$. The performance of a receiver with perfect phase reconstruction is compared with the two phase reconstruction receiver concepts already described. The limiters are operated with an over-sampling ratio of 104. For the post-filter a linear phase FIR-low-pass filter with corner frequency equal to 0.8 times the symbol rate and 50dB stop-band attenuation was designed. The respective suppression of the harmonics which are generated by the limiter operation turned out to be sufficient. For the implementation of the filter 128 coefficients are required.

Fig. 5 shows the performance results for an additive white gaussian noise (AWGN) channel. All architectures considered are based on a low IF receiver concept. After sample rate reduction to the symbol rate bit detection is performed using the Viterbi algorithm as described in section 3. In all cases perfect synchronization has been assumed. In a first step a linear phase band-pass filter with 1 MHz bandwidth was chosen for the channel selection filter. Curves 1 to 3 show the respective performance which is nearly identical for all three receiver concepts. This confirms that phase reconstruction can be sufficiently approximated by a simple post filter. For a bit error rate of 10^{-3} a E_b/N_0 of 10,1dB is obtained, which is very close to the MLSE bound [11].

In a second step the analog channel selection filter was implemented by a more realistic recursive structure. For the considered application we designed a band-pass filter derived from a 5th-order low-pass filter with chebycheff characteristic.

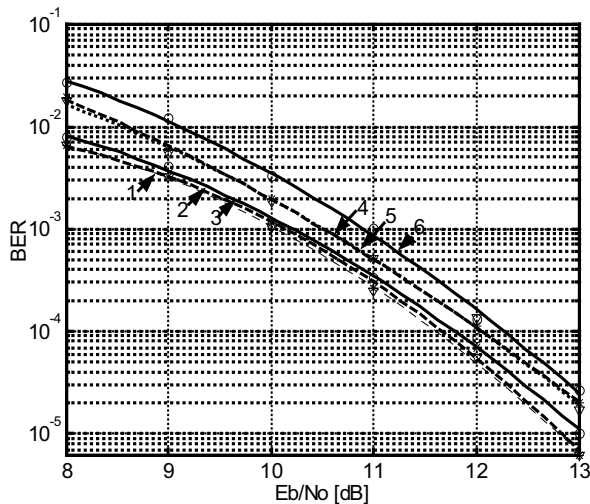


Fig. 5. E_b/N_0 versus bit error rate for the low IF-receiver in an AWGN-channel. Performance with a linear phase channel-selection filter (curves 1-3), and a recursive channel filter with chebycheff characteristic (curves 4-6), perfect phase reconstruction (curves 1&4), limiter with reconstruction of I- and Q-signals (2&5), limiter with reconstruction of base-band phase (3&6).

The corner frequency of the reference low-pass filter was set to 528 kHz to meet the Bluetooth requirements.

Due to the group delay distortion generated by this filter, the performance of the receiver with perfect phase reconstruction was degraded by about 0.3dB. The receiver with phase reconstruction behind down-conversion to the base-band showed nearly the same good performance. Direct phase reconstruction from the IF-signal however turned out to be a little bit more sensitive, the respective additional loss was 0.4dB.

Bit detection based on the Viterbi algorithm however is very sensitive with respect to deviations of the modulation index η , which was confirmed by respective simulations. A possible solution to this problem is to estimate the modulation index from the first part of the data packet which is often known since a fixed pattern is used for synchronization [9].

The simulation results confirmed the superior performance of the phase reconstruction receiver combined with Viterbi detection. Even if phase reconstruction is approximated by a simple FIR post-filter a performance close to the MLSE-bound can be obtained.

5. SUMMARY

In this paper we have described the optimization of a CPFSK-receiver based on a phase reconstruction structure. Combining the phase reconstruction module with a Viterbi detector a performance close to the MLSE-bound can be obtained.

Since in modulation schemes like CPFSK or PSK the information is exclusively coded in the phase of the transmitted signal, no performance losses have to be taken into account when using a limiter receiver with perfect phase reconstruction. Since phase reconstruction algorithm proposed up to now require sophisticated digital signal processing it has been shown in this paper that a performance close to the optimum can be also obtained when using simple post-filters for phase approximation behind the limiters. The performance of the proposed method has been verified by simulation models. It was confirmed that the concept has a superior performance compared to conventional limiter-discriminator demodulation and detection schemes.

For the digitization only comparators and no multi-bit A/D-converters are needed, thereby making the implementation simple and cost effective. The method can be completely implemented in the DSP-part of the receiver. The concept furthermore enables the complete integration of RF and base-band functionalities into a single integrated circuit in a pure CMOS technology.

6. REFERENCES

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