DECISION FEEDBACK EQUALIZATION FOR BLUETOOTH SYSTEMS

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ABSTRACT

Gaussian Frequency Shift Keying modulation has been chosen as the modulation technique for the physical layer of Bluetooth. Bluetooth is a standard for low cost and low power wireless communications between various mobile devices. The optimal demodulation of a GFSK signal involves an extremely complex Viterbi decoder. Therefore designers have opted for the noncoherent detection of GFSK which uses a frequency discriminator, followed by symbol by symbol detection. Here, we describe a decision feedback equalizer to be added after the discriminator. The DFE receiver gives gains in excess of 2 dBs. We also describe how to increase the current data rate of a Bluetooth system by increasing the symbol rate and not the alphabet size.

1. INTRODUCTION

A Bluetooth chip is supposed to have very low cost. Therefore, the modulation technique chosen for this system was Gaussian Frequency Shift Keying (GFSK) [1]. The envelope of the modulated waveform is constant, and therefore such a system has the advantage of not requiring a high cost linear power amplifier. Also, since the waveform produced by the GFSK modulator is continuous in phase even at the edges of symbols, this system has some containment in frequency, which allows for more channels in a given bandwidth.

Unfortunately, the optimum GFSK demodulator is a very complex Viterbi Decoder, where the number of states depends on the modulation index[1].. Small changes in the modulation index lead to a totally different receiver. This rendered the optimal receiver unpractical, and noncoherent suboptimal receivers are typically used to demodulate GFSK signals. Noncoherent modulation suffers from a significant loss compared to the optimal receivers, but they offer a very cheap implementation cost, which makes them very appealing to the likes of Bluetooth systems.

In this paper, we propose using a fractionally sampled decision feedback equalizer to enhance the performance of noncoherent GFSK receivers. We apply our receiver specifically to Bluetooth, since there are several characteristics there that make such a system unique [1]. We also propose increasing the data rate of the current Bluetooth system by increasing its symbol rate from 1 Mega symbols per second (MSPS) to x MSPS where $x > 1$. This increases the data throughput to x Mega bits per second. We show that this is preferred to increasing the alphabet size of Bluetooth from a binary alphabet to an alphabet of size 2^x.

Our paper is organized as follows. In the next section we present a brief overview of GFSK modulation. We explain the optimal Viterbi decoder. In section 3, we present an overview of the Bluetooth system. We then present our proposed DFE equalizer and present results showing its superiority to no equalization. In section 5, we show how to increase the data rate of a Bluetooth system and show how this might be attractive from an implementation point of view, and then we end with the conclusion.

[1] As will be explained later, the packet size in Bluetooth is short for a full adaptation, and the channel changes frequently

2. GFSK

Gaussian Frequency Shift Keying is a form of frequency modulation where the data bits are convolved with a rectangular signal and then with a Gaussian pulse before being used to frequency modulate a carrier. This leads to phase continuity at the edges of the symbols and hence containment in frequency. Mathematically we can write,

$$s(t) = cos(2\pi f_c t + 2\pi h \int_{-\inf}^{t} \sum_n [I_n * g(t - nT)]dt + \eta) \quad (1)$$

where I_n is the data bit at time instant nT , $g(t)$ is a Gaussian pulse convolved with a unit signal of duration T seconds, h is termed the modulation index, f_c is the carrier frequency and η is a random phase. The modulation index h, the bandwidth of the Gaussian pulse, and the symbol rate will specify the bandwidth of the transmitted signal. Fig. 1 shows a basic GFSK modulator.

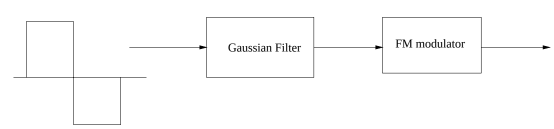

Fig. 1. GFSK Modulator

The optimal decoder for such a modulation schemes [1] employs first coherent down conversion to get the waveform down to baseband, followed by a Viterbi decoder is used. The number of states of the Viterbi decoder depend on the modulation index h and on the number of symbols the Gaussian pulse spans. if the pulse $g(t)$ has a width of L symbols, and h is a rational number than can be expressed as $\frac{m}{p}$, then the number of states of the Viterbi trellis is $p2^{L-1}$ or $p2^L$ for odd and even m respectively. After down conversion and phase correction, the received signal, assuming no noise, can be written as

$$\phi(t) = 2\pi h \int_{-\inf}^{t} \sum_n [I_n * g(t - nT)]dt \quad (2)$$

The states of the decoder are specified by this modulation phase.

3. BLUETOOTH

Bluetooth [2] is a short distance wireless connection, up to 10 meters, that is intended to be used as cable replacement and in ad hoc networking. A handsfree cellular phone will communicate to a wireless headset via Bluetooth, and meeting attendees will be able to freely exchange information on their laptops via a Bluetooth port on their computers. To make Bluetooth devices ubiquitous, the intention was to make it almost free to the consumers, and hence a very simple scheme was chosen, namely binary GFSK. Bluetooth operates in the ISM unlicensed band between 2.400 and 2.4835 GHZ. It divides the frequency spectrum into 78 channels each of width 1 MHZ, as per the FCC regulations [3], and hops onto these channels randomly at the rate of 1600 hops/sec. Bluetooth devices form a *piconet* where a master device controls the timing and chooses which device, termed slave, is to send in a particular time slot. Only communications to and from the master is allowed. Master and slave transmissions are time division duplexed every two slots. Two types of channels are defined, a packet switched asynchronous link where a slave cannot send a packet except if it was addressed by the master in the previous slot. The other link is a circuit switched synchronous link where the time between each transmission is agreed upon between the master and the slave at the beginning of the synchronous transmission. Typically, an asynchronous link is used for data transmission and a synchronous link is used for speech. The frequency hopping sequence of a piconet is determined by the 48 IEEE bit address and clock of the master. These are sent to a slave whenever it joins the piconet.

The maximum number of slaves allowed in a piconet is seven. Piconets can communicate with each other, by allowing a device to be a master in one piconet and a slave in another. This extends the range of Bluetooth communications. A Bluetooth packet starts with a synchronization word of length 64 bits. The synchronization word is essentially a BCH coded version [4] of the 48 IEEE bit address of the master. Both the master and slave use this word in their transmissions, and hence this serves to identify a particular piconet. The synchronization word has 8 header and trailer bits attached to it and is then followed by the packet header of length 54 bits. The packet header is used to identify the type of the packet, to piggy pack an acknowledgment bit, and to signal whether this packet is a new packet or a retransmission in response to a lost or negative acknowledgment. The header has error check bits attached to it and is coded using a 3 time repetition code. The payload then follows the header. The maximum length of the payload is 240 bits including a payload header and a 2 byte cyclic redundancy check in asynchronous link packets. This makes the maximum length of a packet 366 microseconds, leaving ample time for frequency synthesis for the next slot.

Two error correction mechanisms can be used, a rate 1/3 repetition code, and a rate 2/3 (15,10) shortened Hamming code.

To fit the Bluetooth signal within 1 MHZ, with -20 dB from the center of the band to its edges, the modulation index of Bluetooth is chosen to be in the range between .28 and .35. This means a maximum frequency deviation between 140 and 175 KHZ. A Gaussian pulse of a 3dB bandwidth of 500 Hz is used in the transmitter. To save power, a Bluetooth device typically transmits at 0 dBm, although there is an optional mode up to 20 dBm with power control.

4. DECISION FEEDBACK EQUALIZATION

A simple demodulator for a GFSK modulated signal is shown in Fig. 2. It consists, after downconversion and filtering, of a limiter and a discriminator. A discriminator is a frequency selective filter that is linear around the carrier frequency. After the low pass filter, the signal is compared with zero for a bit decision.

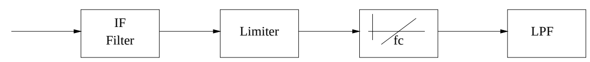

Fig. 2. GFSK demodulator

After the limiter and discriminator, the noise is no longer white. It is data dependent and only approximately Gaussian. Finding the optimal receiver is mathematically untractable. To try to converge to the optimal receiver, an adaptive architecture should be used. A decision feedback equalizer seems to present a good solution, due to its noise whitening characteristics and since by using that kind of a receiver we can increase our data rate as we will explain later.

Decision feedback equalization have been proposed before for GFSK demodulation. For example, in [5], the authors use a linear approximation to a continuous phase modulated signal [6] to derive a DFE equalizer. There coherent detection is used and the front end of the detector is linear. In [7], although a non-coherent detection is used, the equalizer is meant to combat the intersymbol interference caused by the channel and the Gaussian pulse in the transmitter. But since the Bluetooth data rate is 1 MBPS and it operates in an indoor channel whose delay spread is in the order of 50 ns [8], the channel can be modelled as a single path Rayleigh channel, and hence we suffer no ISI caused by the channel. Moreover, the Gaussian pulse used adds negligible ISI to the transmitted symbols. Nevertheless, significant performance enhancement can be expected by employing a DFE due to the noise spectrum at the output of the low pass filter of Fig. 2. As we will show shortly, the noise is colored at this point. The DFE is known to whiten the noise [9] while removing ISI, while having less complexity than the Viterbi decoder. Therefore, it is an ideal choice here.

Let us first derive the noise at the output of the low pass filters. We use a derivation similar to the one in [10] for analog FM. For simplicity of the derivation, we assume that we do not use a limiter. Before the discriminator, the signal can be written as

$$r(t) = s(t) + n(t) \tag{3}$$

where $n(t)$ is bandpass white noise centered around f_c and can be written as $n_c(t)cos(2\pi f_c t) + n_s(t)sin(2\pi f_c t)$. Therefore,

$$r(t) = R(t)cos(2\pi f_c t + \theta(t)) \tag{4}$$

where $R(t)$ is the envelope of $r(t)$, and

$$\theta(t) = tan^{-1}\frac{n_s + sin\phi}{n_c + cos\phi} \tag{5}$$

where we dropped the time dependency for brevity. We model the discriminator as a delay and multiply circuit [11] with the delay τ such that $f_c\tau = \frac{1}{4}$. Therefore after the LPF we get,

$$\begin{aligned} y(t) &= R(t)R(t+\tau)sin(\theta(t+\tau) - \theta(t)) \\ &= R(t)R(t+\tau)(sin(\theta(t+\tau))cos(\theta(t)) \\ &\quad -cos(\theta(t+\tau))sin(\theta(t))) \end{aligned} \tag{6}$$

But, $sin(\theta) = \frac{n_s + sin\phi}{R}$, and $cos\theta = \frac{n_c + cos\phi}{R}$. Therefore,

$$\begin{aligned} y(t) &= sin(\phi(t+\tau) - \phi(t)) \\ &+ n_s(t+\tau)n_c(t) + n_c(t+\tau)n_s(t) \\ &+ n_s(t+\tau)cos(\phi(t)) + n_c(t)sin(\phi(t+\tau)) \\ &- n_s(t)cos(\phi(t+\tau)) - n_c(t+\tau)sin(\phi(t)) \end{aligned} \quad (7)$$

Notice that, although the noise is not Gaussian, it can be approximated as being Gaussian at relatively high SNR.

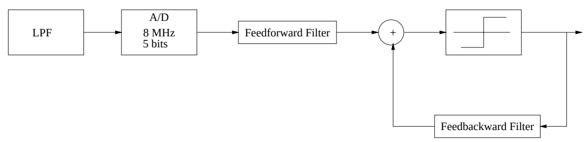

Fig. 3. Decision Feedback Equalizer

Fig. 3 shows a DFE added after the LPF of Fig. 2. After the LPF we add a 5-bit A/D operating at 8 MHz. After several simulation trials, a feedforward length of 3 symbols (24 samples) was used, and a feedbackward length of 1. This makes the implementation simply a convolution with a filter and then comparison to a variable threshold that depends on the previous detected bit. Fig. 4 and Fig. 5 compares the power spectrum density of the noise at the sampling instance after the LPF and after the DFE. The DFE is affecting the PSD of the noise. If the DFE were infinitely long, one would expect that the noise would be white if it can be approximated to be independent on the data, which is a reasonable approximation in high signal to noise ratio scenarios.

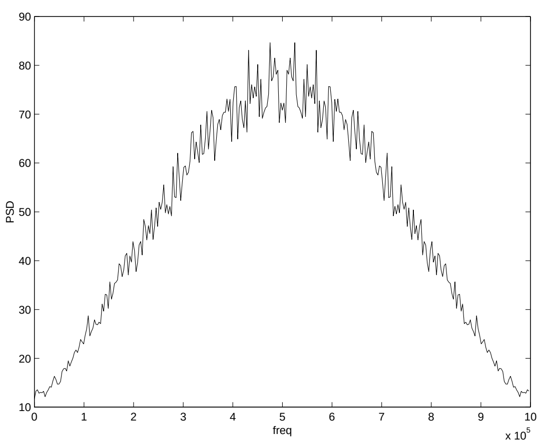

Fig. 4. Power Spectrum Density of Noise before DFE

An adaptive DFE could be used if one has enough training bits to adapt it on line. But since the synchronization bits of a Bluetooth packet is short (72) bits, it was decided that it is better to use a non adaptive version of the DFE. A DFE will be trained off line at a certain SNR and then this will be used in reception. We can also use this a starting point and adapt during the synchronization word, but our simulations show that this is not needed as the DFE equalizer that was obtained was very robust to changes in SNR and to changes in the modulation index, h.

Fig. 6 shows the DFE performance compared to a hard limiter after the LPF.

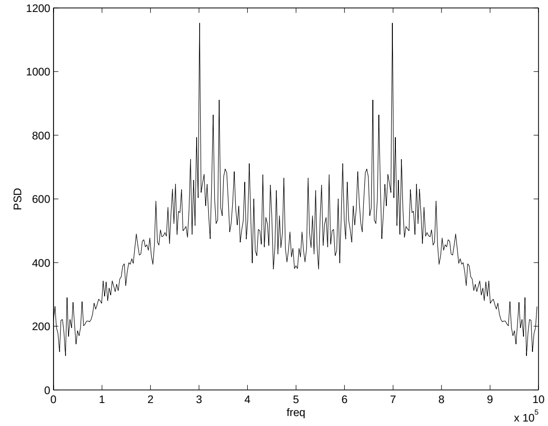

Fig. 5. Power Spectrum Density of Noise after DFE

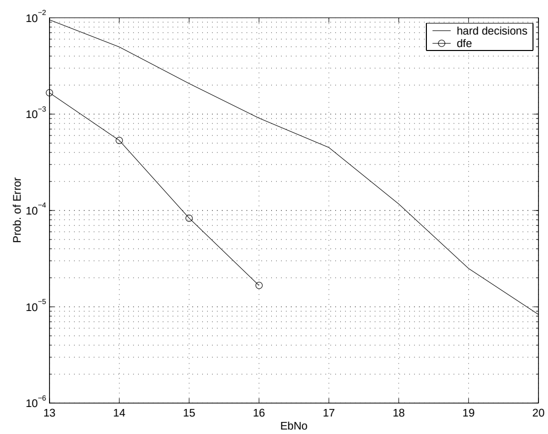

Fig. 6. DFE performance at 8 MHz

Fig. 7 shows that the performance is not affected by dropping the sampling rate down to 4 MHz. We started noticing performance degradation when we go down to 2 MHz.

These results were obtained for an AWGN channel. Since Bluetooth is a wireless link, the channel is better modelled as a single ray Rayleigh fading channel. Fig. 8 shows the packet error rate of a system employing DFE. The gain is around 4 dB's.

5. OVERSAMPLED FSK

To increase the data rate of a Bluetooth system, we can use an FSK of a larger alphabet than the binary. This limits the choices to an integer number of bits per second. Another more flexible option, would be to simply change the intersymbol period at the input of the Gaussian filter. So, for example, if we want to increase the data rate to 2 MBPS, we can use a 4-FSK system at the same rate as the original FSK system, or use a 2- FSK system at twice the rate, i.e. 2 MHz. This is in a sense an oversampled system as the bandwidth of the channel is only 1 MHz. But this is a nonlinear modulation scheme, and hence Nyquist criterion does not really apply here. Notice that twice the rate system adds considerable ISI, but it can be successfully removed using our proposed DFE equalizer. Fig. 9 shows the results. In these results a sampling rate of 48 MHz was used before the DFE. Fig. 10 shows a 4x system. At probability of

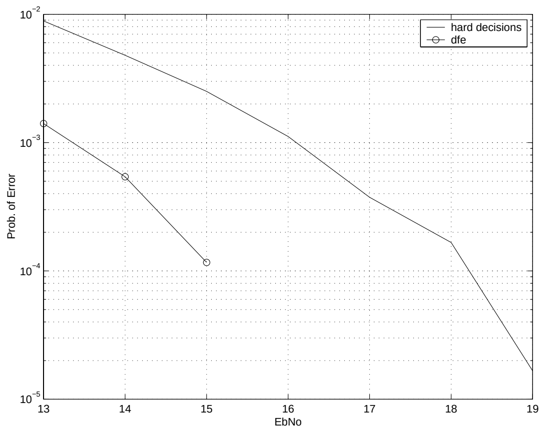

Fig. 7. DFE performance at 4 MHz

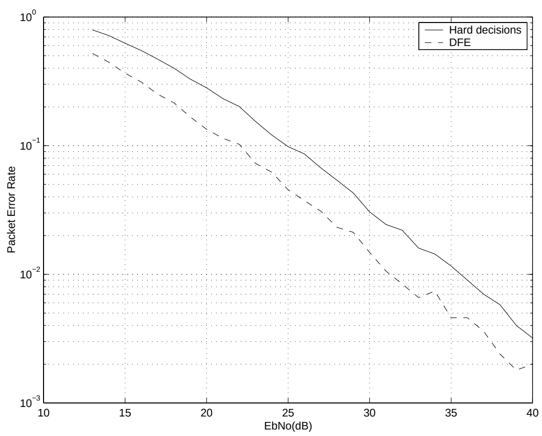

Fig. 8. Packet Error Rate

error of 10^{-3}, we loose around 3 dB over the 2x system. Notice that going from a 4QAM to a 16QAM we loose 6 dB's.

6. CONCLUSION

We presented a non-adaptive DFE for GFSK signals. Our receiver employing the DFE performs better than the hard decision receiver in both AWGN channels and fading channels. We have also shown that it is feasible to increase the data rate via increasing the symbol rate. Our results show that this is preferable to increasing the alphabet size. This offers a very attractive technique for increasing the throughput of Bluetooth devices.

7. REFERENCES

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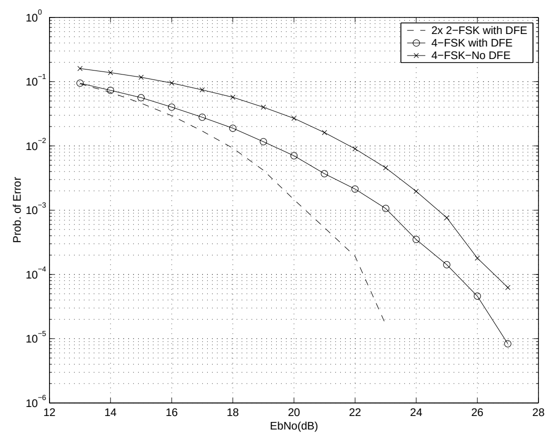

Fig. 9. Results for a 2-FSK system at twice the rate

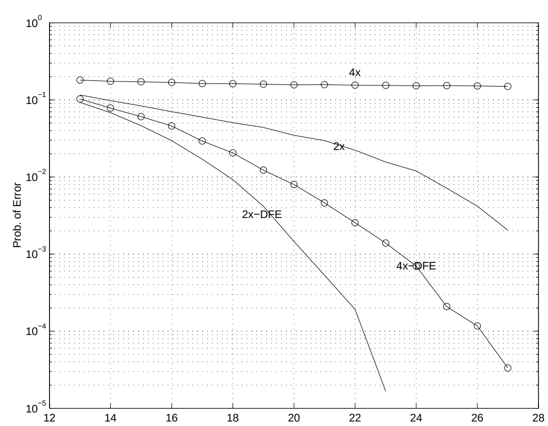

Fig. 10. A 2x system vs a 4x system