IEEE P802.15
Wireless Personal Area Networks

Project	IEEE P802.15 Working Group for Wireless Personal Area Networks (WPANs)		
Title	**XtremeSpectrum CFP Document**		
Date Submitted	March 2003		
Source	Rick Roberts 8133 Leesburg Pike, Suite 700 Vienna, Va 22182	Voice: Fax: E-mail:	[703.269.3043] [703.749.0248] [rroberts@xtremespectrum.com]
Re:	Response to CFP 02/372		
Abstract	[Detailed information for the XtremeSpectrum proposal. The summary detail is contained in document 03/153.]		
Purpose	[Description of what the author wants P802.15 to do with the information in the document.]		
Notice			
Release			

Table of Contents

Introduction and PDF Draft Text

This document supports the XtremeSpectrum proposal summary (document 02/153) and presents technical details of significant. This document will be periodically revised as details become available.

To see the XSI proposed draft standard text in PDF format, click on the yellow box below.

Coming Soon

1. Proposal Summary

- BPSK & QPSK in two PHY bands
- CDMA coding for multiple access
  - Supports up to 4 piconets per PHY band
- High Band: 8.2 GHz with BW=2.736 GHz
  - 57 Mbps to 800 Mbps
- Low Band: 4.1 GHz with BW=1.368 GHz
  - 28.5 Mbps to 400 Mbps
- Nyquist Filtering with 50% excess bandwidth
- Protection for UNII band
- Constant symbol rate per sub-band
  - High Band 114 Ms/s
  - Low Band 57 Ms/s
- 24 Chip/Symbol Ternary (3 state) Codewords
- CCA Supported
  - Each piconet uses an offset center frequency and chipping rate
- Three FEC options
  - Convolutional (low speeds, moderate coding gain)
  - Reed-Solomon (high speeds, moderate coding gain)
  - Concatenated (low speeds, high coding gain)
- Three preamble length options for QoS dependent PHY overhead
  - Long Preamble for long range and low rates
  - Medium Preamble for nominal range and rates
  - Short Preamble for short range and high rates
- RX link budget summary
  - Provide 114 Mb/s at 10 meters in the low band with 7.0 dB of margin
  - Provide 200 Mb/s at 10 meters in the low band with 5.1 dB of margin
  - Provide 600 Mb/s at 4 meters in the high band with 3.3 dB of margin
- RX uses DFE for ISI mitigation
- RX uses RAKE for ICI mitigation
- The 802.15.3 MAC is used with very little modification

- Ranging enhancement is proposed
- Simple “durable” radio design
  - Worldwide capable
  - Addresses future, potential regulatory differences
- Graceful coexistence with other services
  - Spectral stewardship
- Inexpensive
- Low power consumption
- High throughput with graceful path to higher data rates
  - Maintains graceful backwards compatibility

2. Performance Details

2.1. Power Consumption

Goal is <200 mW total for RF and PLCP baseband. Current estimates (circ. 2002) are shown below.

Chip	Power mW
RF Front End	260 mW (3.3 V) TX, 280 mW (2.5 V) RX
PLCP Baseband	Sleep: 5 mW, Idle: 74 mW, TX: 95 mW, RX: 300 mW

Idle is stand-by with clocks running. Idle is clocks off except for master.

2.2. Die Size Estimates

Process technology is circ. 2002.

Chip	Size Estimate
RF Front End	4.7 mm x 4.1 mm, 0.18 SiGe
PLCP Baseband	4.4 mm x 4.4 mm, 0.18 CMOS, 1.8V core, 3.3V I/O Gate Count Estimate: TBD

2.3. Time to Market

XtremeSpectrum, Inc. proposes a simple durable worldwide radio design via an inexpensive, low power consumption alternative PHY for a higher data rate amendment to standard 802.15.3. This proposal will include IEEE formatted draft amendment standard text for the Alt PHY as well as text for enhanced MAC subclauses, and informative annexes as required. If the IEEE down selection process yields a single proposal by the close of the Session #25/San Francisco 25Jul03 plenary meeting we estimate that P802.15.3a draft standard compliant silicon product is possible for integration by end of calendar year 2003.”

2.4. Antenna Size Estimates

The XSI UWB antennas are planar patch antennas with the following areas.

Band	Area
4.1 GHz	1.1 in. x 1.1 in.
8.2 GHz	0.6 in. x 0.6 in.

Note: the effective aperture is approximately ½ the real aperture and many practical antenna designs should be able to meet this principle. The antenna XtremeSpectrum is using in its demonstration is simply one existence proof showing the practicality of an antenna.

2.5. TX Peak-to-Average Value

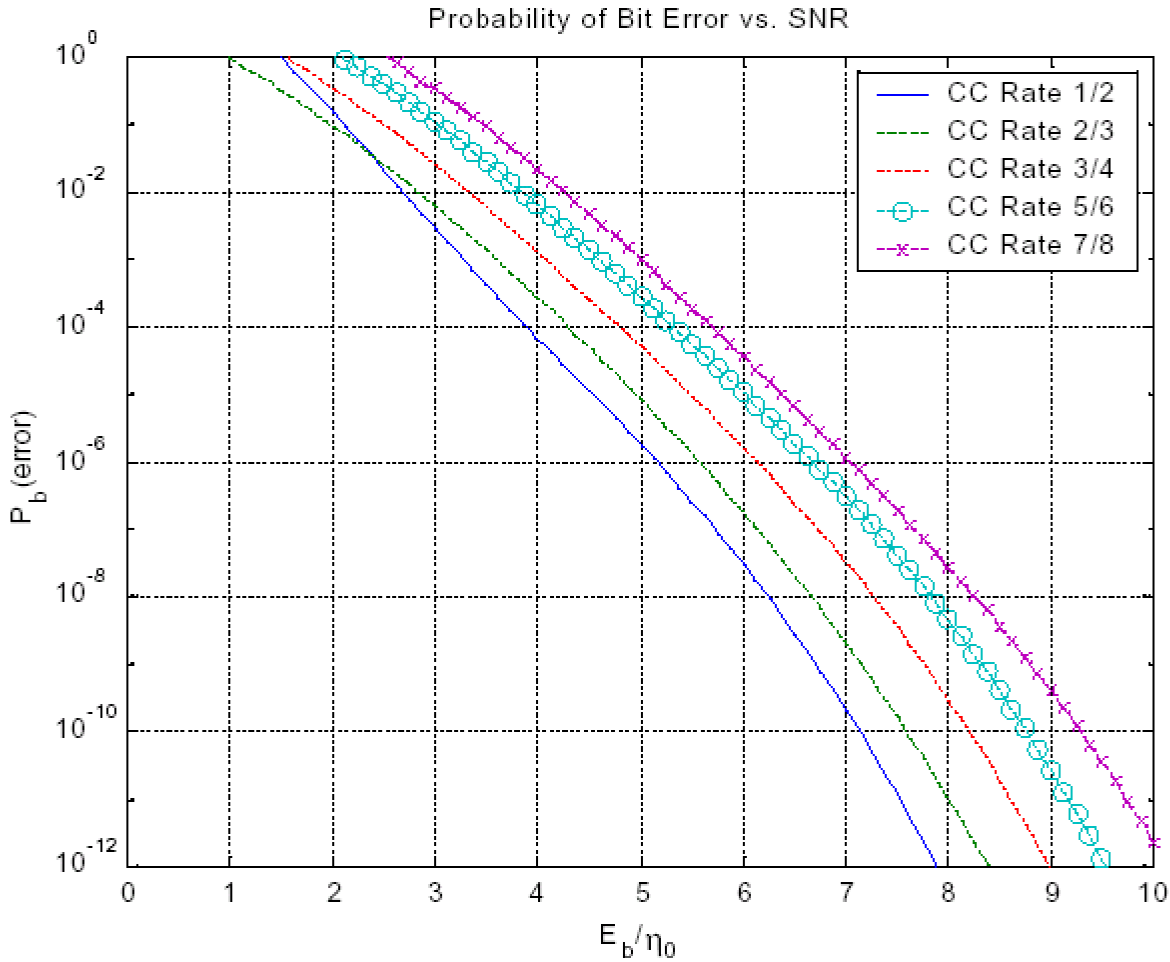
PAR (peak to average ratio) is 6.4 dB for either the low or high band.

2.6. Coding Gain

2.6.1 Convolutional FEC

Viterbi Decoder: K = 7 ($G_1 = 171_8$ $G_2 = 133_8$), 5.2 dB coding gain @ 10^{-5} BER (R=1/2)

Reference: STEL2060C Data Sheet, http://www.intel.com/design/digital/STEL-2060/index.htm



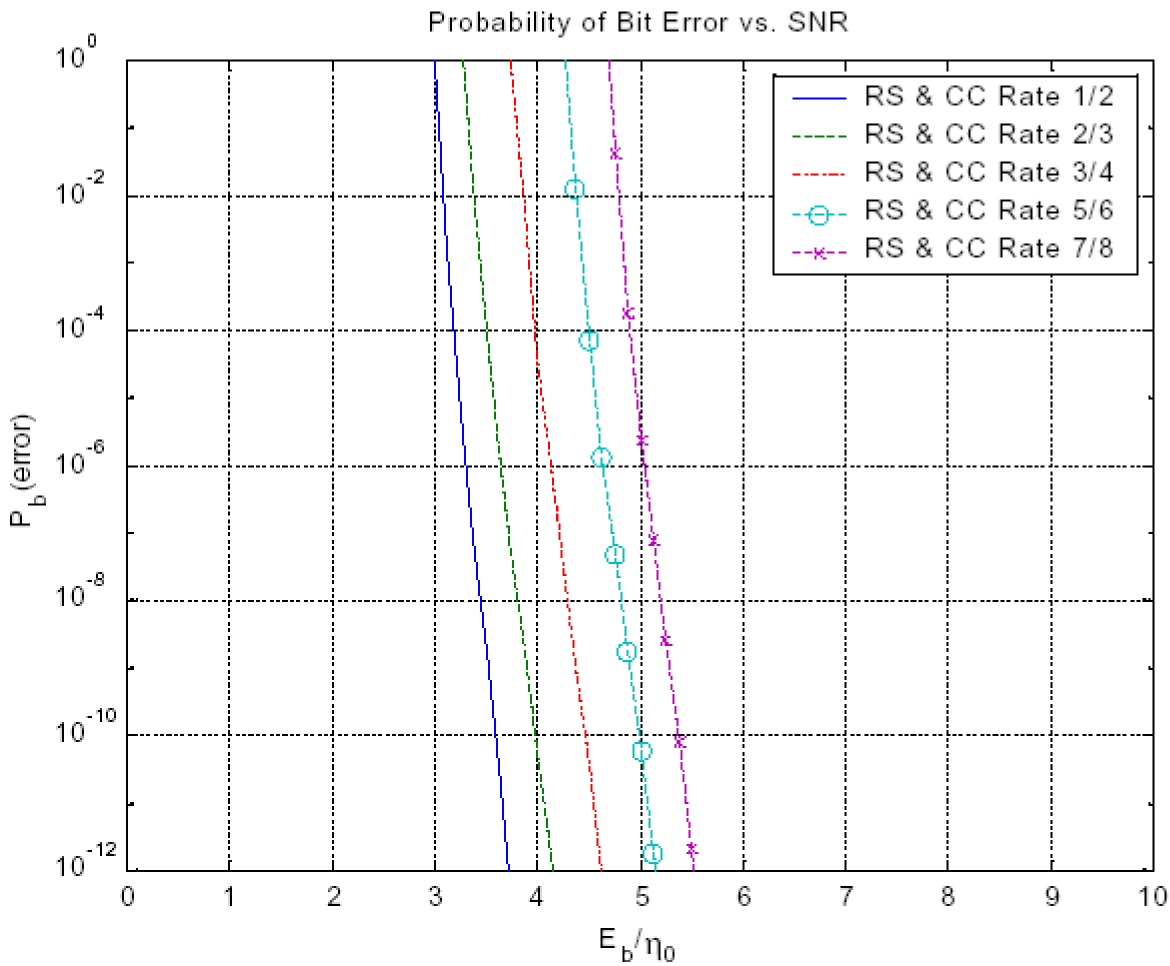
Regeneration for k=6 under progress

Source: Document 802161pc-00_33.pdf, contribution to IEEE802.16.1, "FEC Performance of Concatenated Reed-Solomon and Convolutional Coding with Interleaving", Foerster, Jeff and John Liebetreu, June 2000

2.6.2 Reed-Solomon FEC

R-S[223,255], Coding Gain in AWGN @10e-5, 3.4 dB (via Monte-Carlo Simulation)

2.6.3 Concatenated FEC



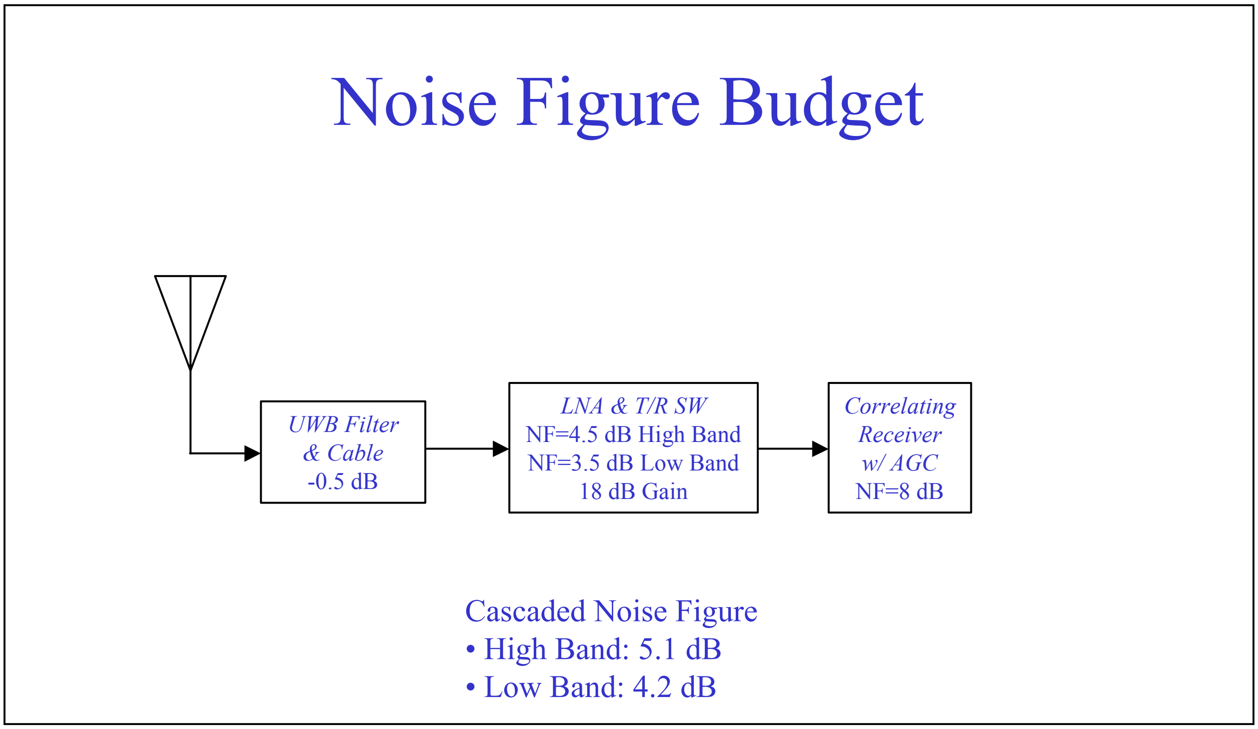
Regeneration for k=6 under progress

Figure [illegible] Performance of Concatenated Reed-Solomon with Convolutional Code

Source: Document 802161pc-00_33.pdf, contribution to IEEE802.16.1, “FEC Performance of Concatenated Reed-Solomon and Convolutional Coding with Interleaving”, Foerster, Jeff and John Liebetreu, June 2000

2.7. Typical Receiver Sensitivity Tables

2.7.1 Receiver Noise Figure



2.7.2 Low Band

Low Band Symbol Rates and Link Budget

Eb/No=9.6 dB, 3 dB implementation loss, 0 dB RAKE gain, NF=4.2 dB
½ rate code gain: 5.2 dB, 2/3 rate code gain: 4.7 dB, RS code gain: 3 dB
8-BOK coding gain: 1.4 dB, 16-BOK coding gain: 2.4 dB

Rate	Modulation	CDMA Code Type	FEC	Fc GHz[1]	Range AWGN	10 meter margin	RX Sensitivity[2]
28.5 Mbps	BPSK	2-BOK (1 bits/symbol)	½ rate convolutional	4.0	35.1 meters	10.9 dB	-84.8 dBm
57 Mbps	BPSK	4-BOK (2 bits/symbol)	½ rate convolutional	4.0	24.8 meters	7.9 dB	-81.8 dBm
75 Mbps	BPSK	8-BOK (3 bits/symbol)	Concatenated	4.0	27.0 meters	8.6 dB	-82.6 dBm
114 Mbps	BPSK	8-BOK (3 bits/symbol)	2/3 rate convolutional	4.0	22.4 meters	7.0 dB	-80.9 dBm
200 Mbps (199.4 Mbps)	BPSK	16-BOK (4 bits/symbol)	RS(255, 223)	4.0	18.1 meters	5.1 dB	-79.1 dBm
400 Mbps (398.8 Mbps)	QPSK	16-BOK (8 bits/symbol)	RS(255, 223)	4.0	12.8 meters	2.1 dB	-76.1 dBm

[1] Center frequency determined as geometric mean in accordance with 03/031r7, clause 5.6
[2] Based upon corrected Eb/No of 9.6 dB after application of all coding gain

Coding Gain References:
- http://www.intel.com/design/digital/STEL-2060/index.htm
- http://grouper.ieee.org/groups/802/16/tg1/phy/contrib/802161pc-00_33.pdf

Table is representative - there are about 22 logical rate combinations offering unique QoS in terms of Rate, BER and latency

2.7.3 High Band

High Band Symbol Rates and Link Budget

Eb/No=9.6 dB, 3 dB implementation loss, 0 dB RAKE gain, NF=5.1 dB
½ rate code gain: 5.2 dB, 2/3 rate code gain: 4.7 dB, RS code gain: 3 dB
8-BOK coding gain: 1.4 dB, 16-BOK coding gain: 2.4 dB

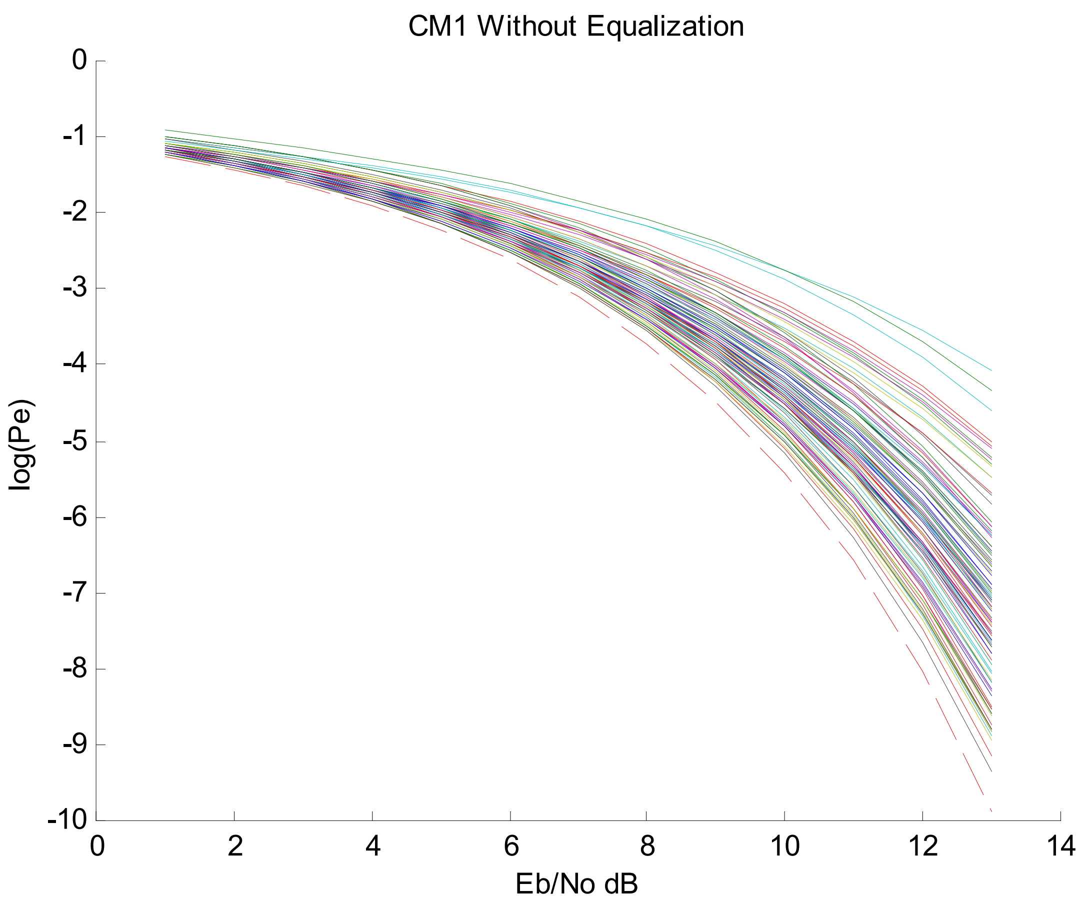
Rate	Modulation	CDMA Code Type	FEC	Fc GHz	Range AWGN	4 meter margin	RX Sensitivity
100 Mbps	BPSK	4-BOK (2 bits/symbol)	Concatenated	8.1	12.6 meters	10.0 dB	-79.0 dBm
114Mbps	BPSK	4-BOK (2 bits/symbol)	½ rate convolutional	8.1	11.1 meters	8.8 dB	-77.9 dBm
200 Mbps (199.4 Mbps)	BPSK	4-BOK (2 bits/symbol)	RS(255, 223)	8.1	8.6 meters	6.7 dB	-75.8 dBm
300 Mbps (299.1 Mbps)	BPSK	8-BOK (3 bits/symbol)	RS(255, 223)	8.1	8.2 meters	6.3 dB	-75.4 dBm
400 Mbps (398.8 Mbps)	BPSK	16-BOK (4 bits/symbol)	RS(255, 223)	8.1	8.1 meters	6.1 dB	-75.2 dBm
600 Mbps (598.2 Mbps)	QPSK	8-BOK (4 bits/symbol)	RS(255, 223)	8.1	5.8 meters	3.3 dB	-72.4 dBm
800 Mbps (797.6 Mbps)	QPSK	16-BOK (8 bits/symbol)	RS(255, 223)	8.1	5.7 meters	3.1 dB	-72.2 dBm

Table is representative - there are about 22 logical rate combinations offering unique QoS in terms of Rate, BER and latency

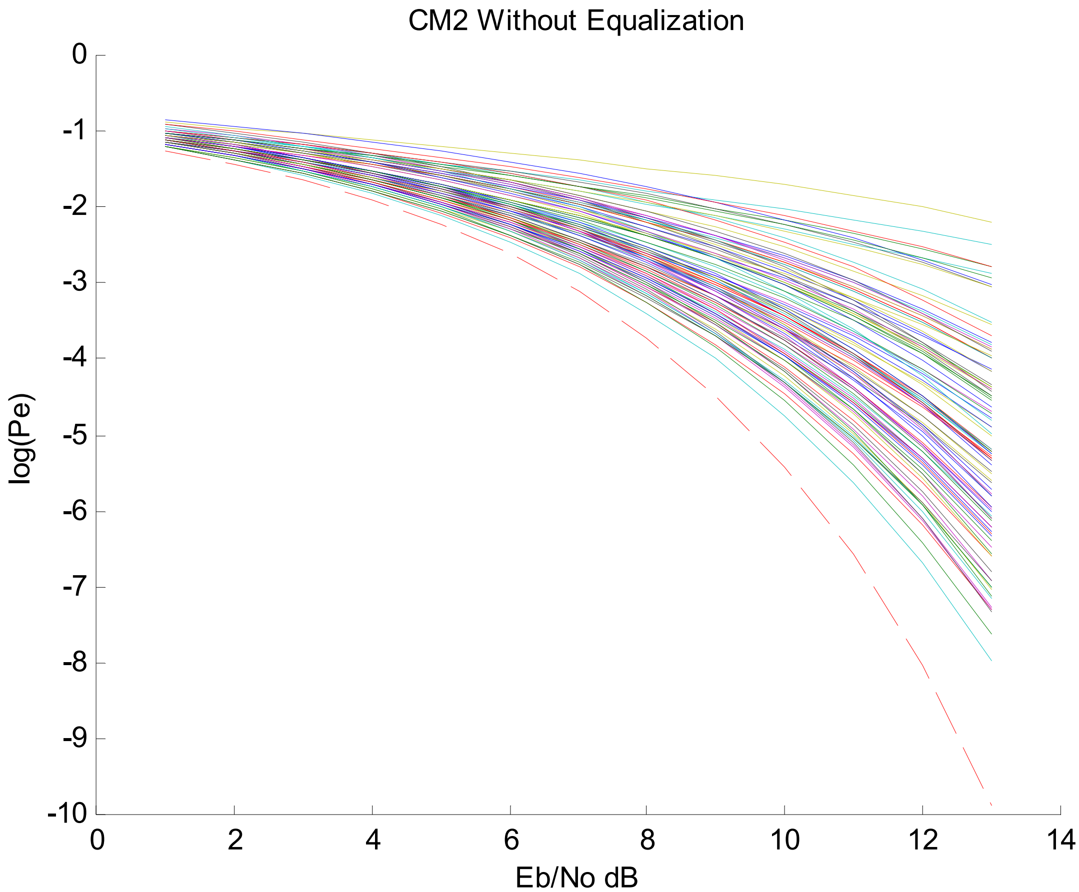
2.8. 2-BOK BER Curves with Multipath Channels

In this section we'll look at performance with various equalizer residual error. Obviously, an ideal equalizer would provide performance mimicking the theoretical curve. The equalizer used here is a modified DFE that provides two feed forward taps that are activated after the feedback taps have opened up the eye pattern. Each plot shows all 100 channels associated with the channel model in question.

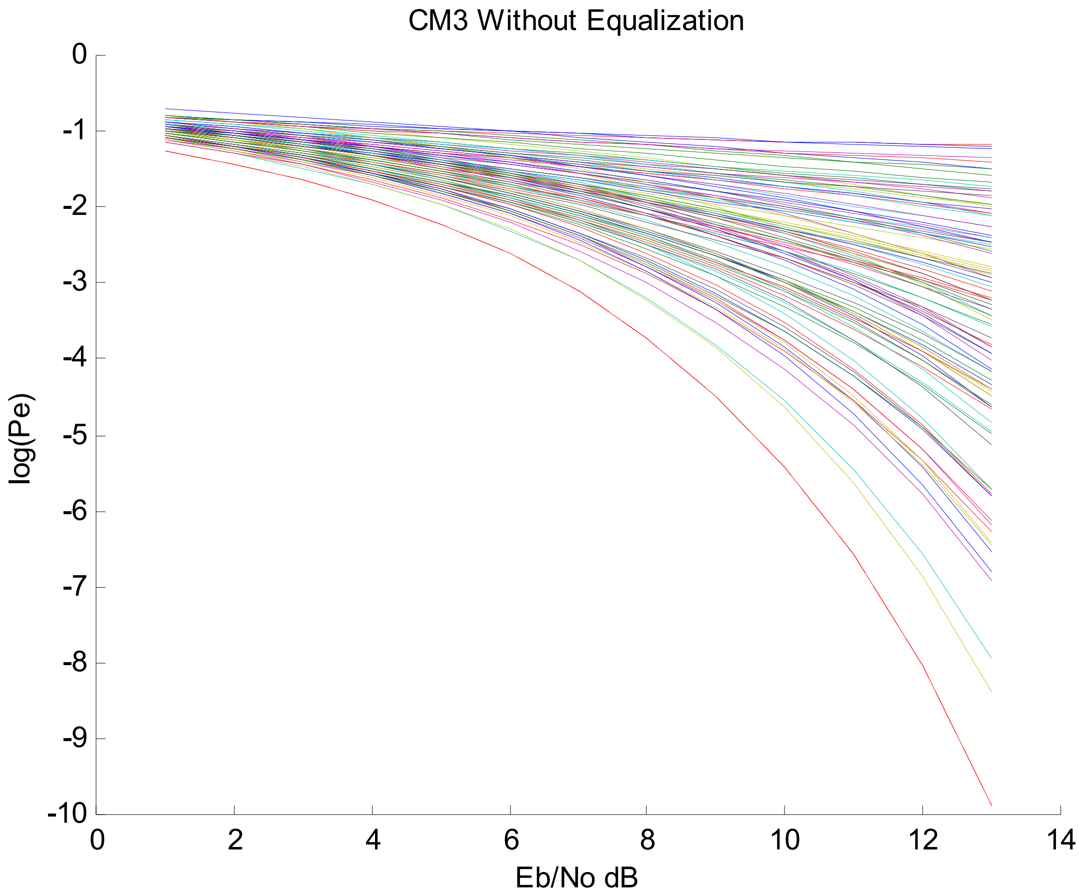
2.8.1 No Equalizer High Band (1/114e6 symbol duration)



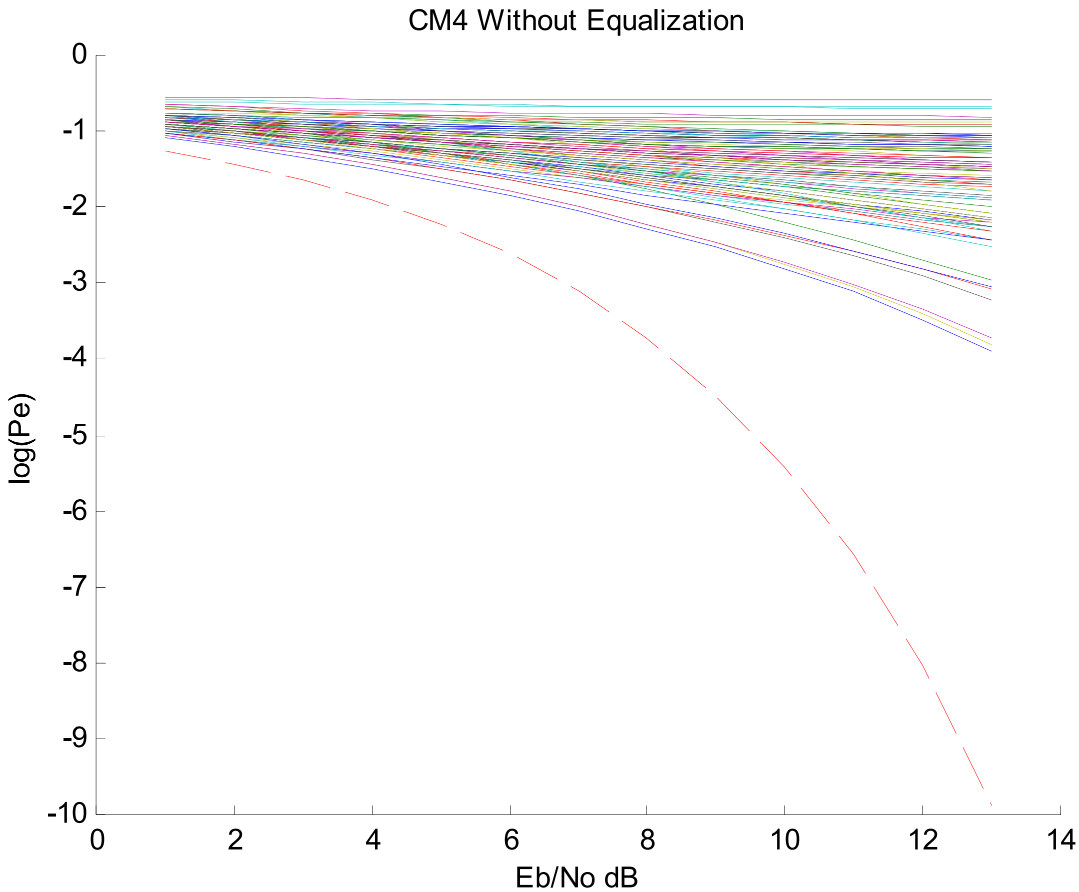
No Equalizer High Band (1/114e6 symbol duration)



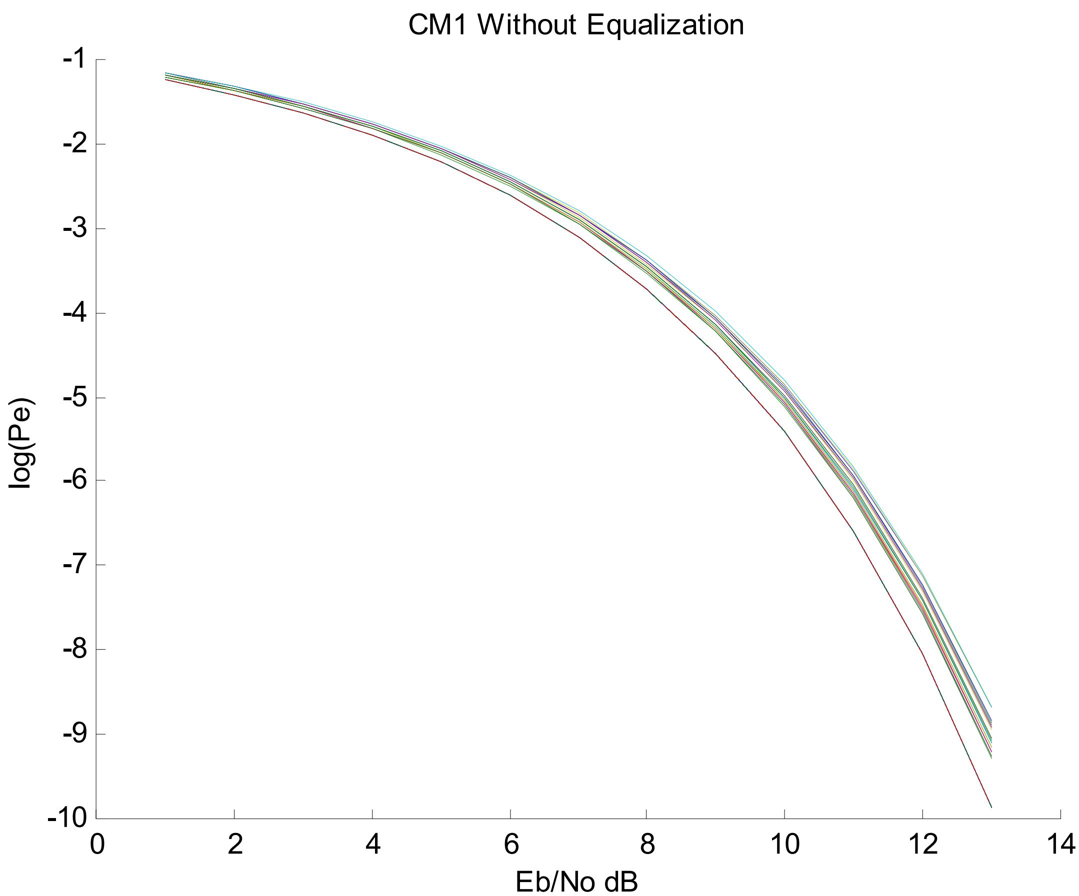
No Equalizer High Band (1/114e6 symbol duration)



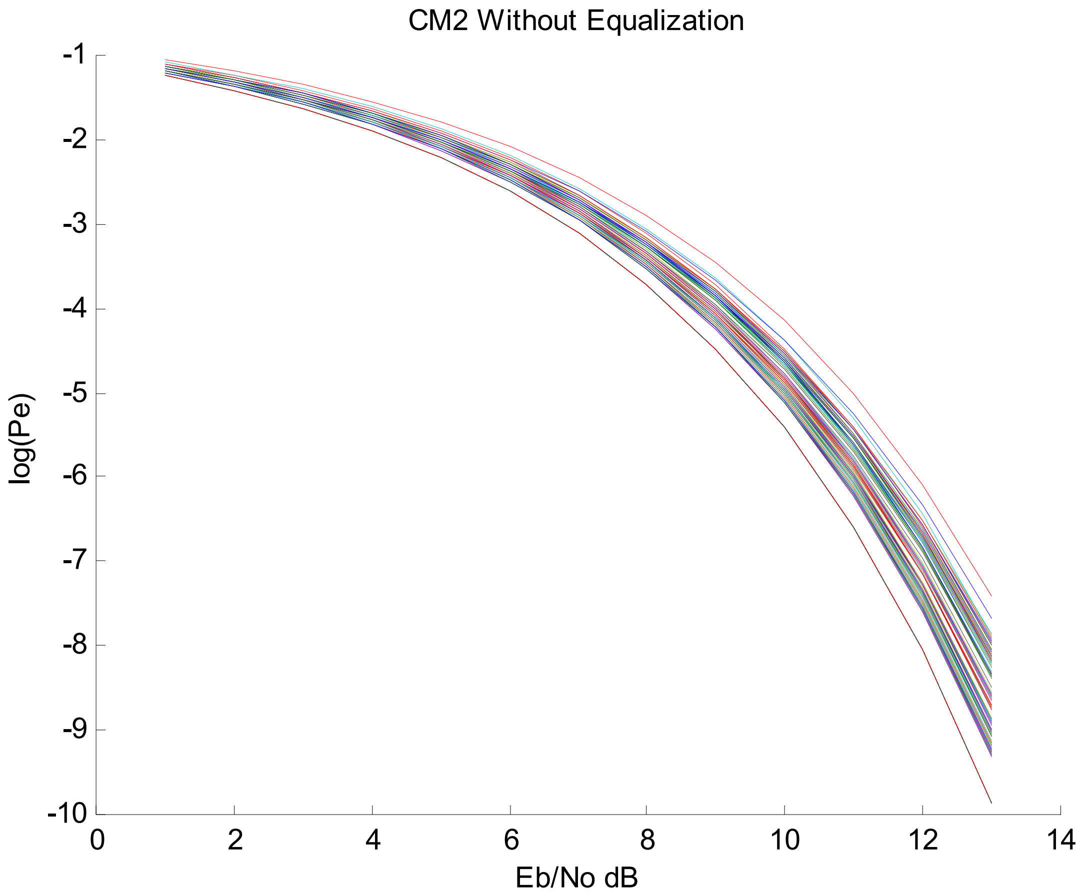
No Equalizer High Band (1/114e6 symbol duration)



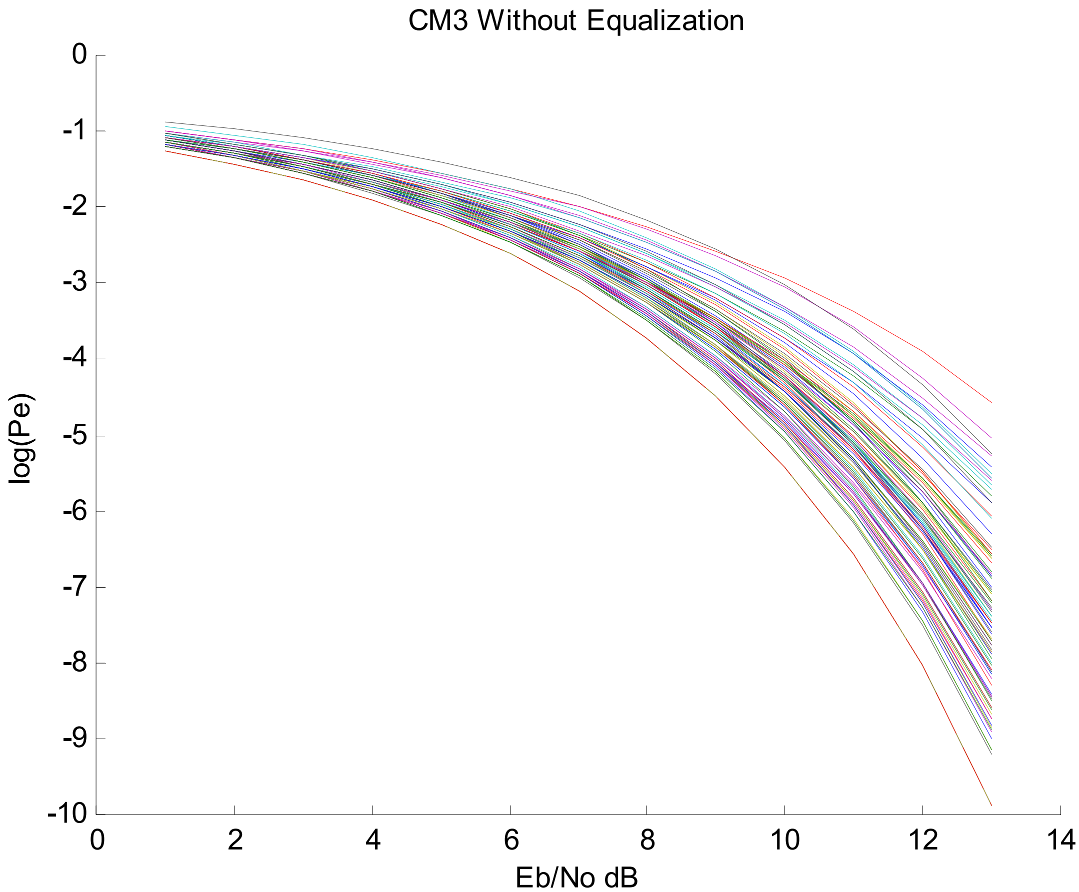
2.8.2 No Equalizer Low Band (1/57e6 symbol duration)



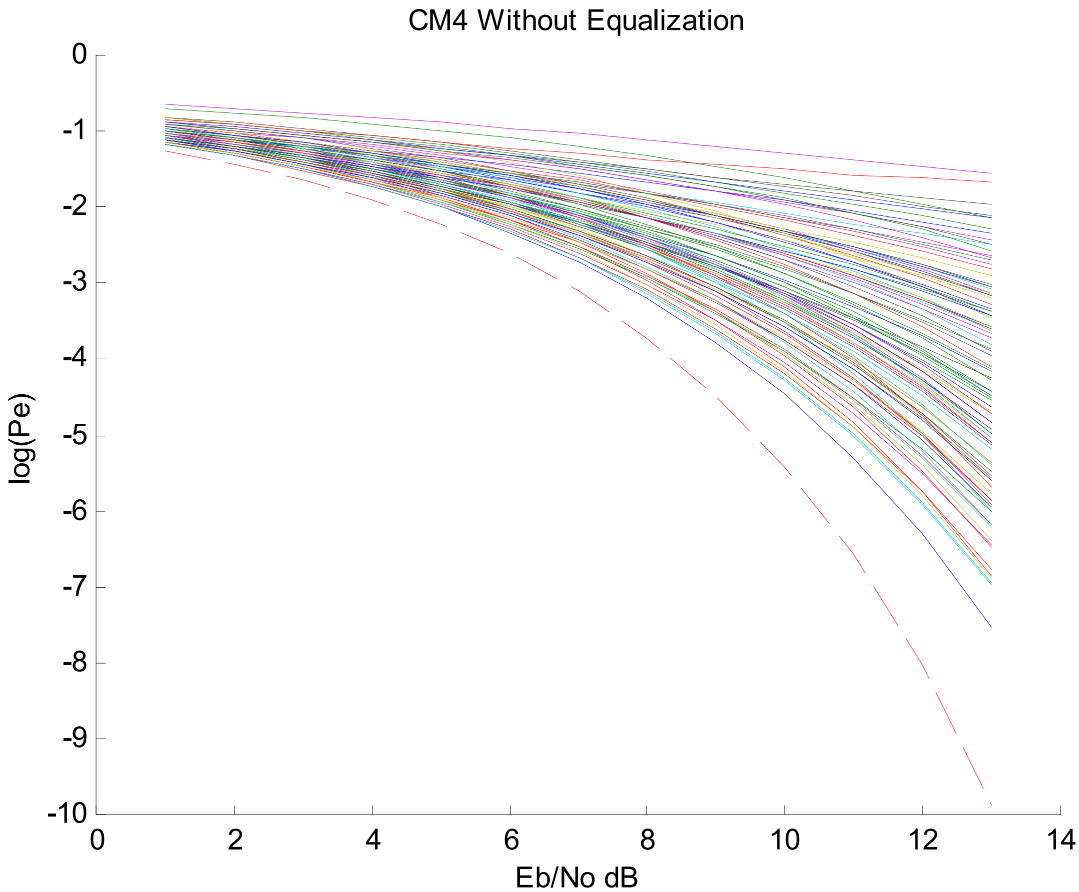
No Equalizer Low Band (1/57e6 symbol duration)



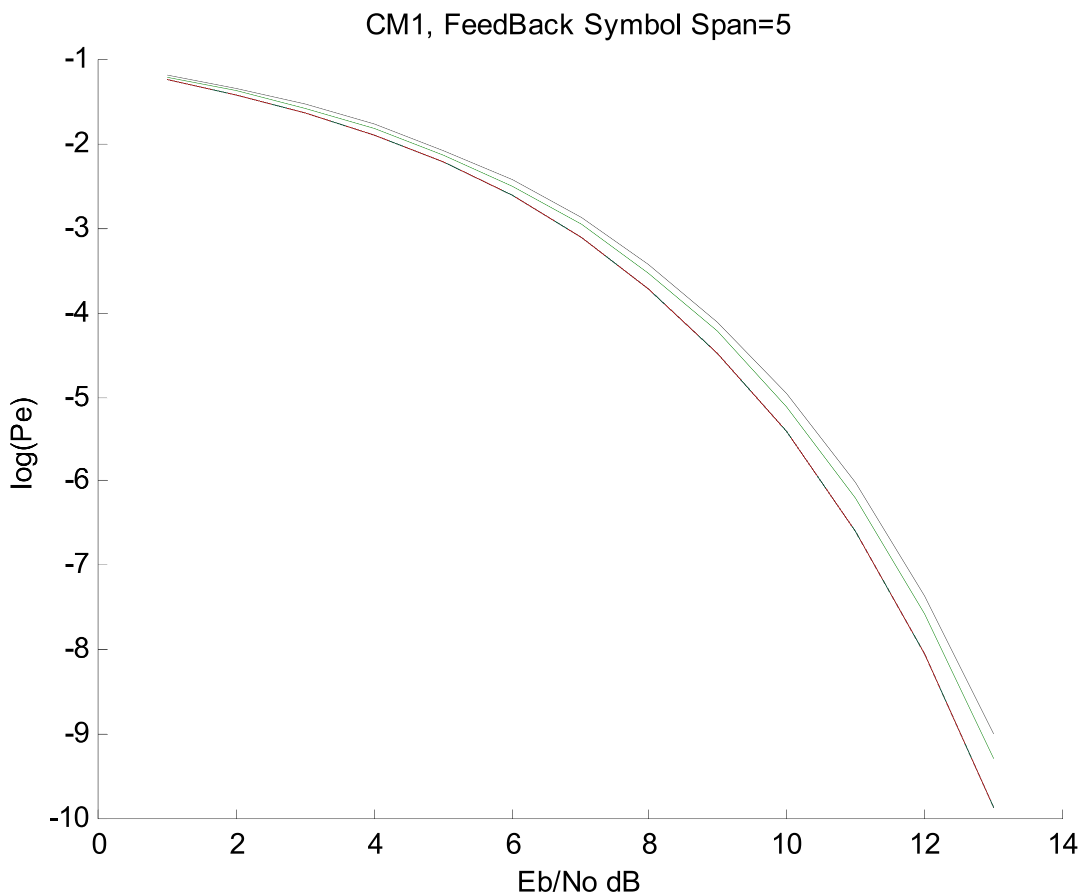
No Equalizer Low Band (1/57e6 symbol duration)



No Equalizer Low Band (1/57e6 symbol duration)



2.8.3 5 Symbol Span Equalizer High Band (1/114e6 symbol duration)



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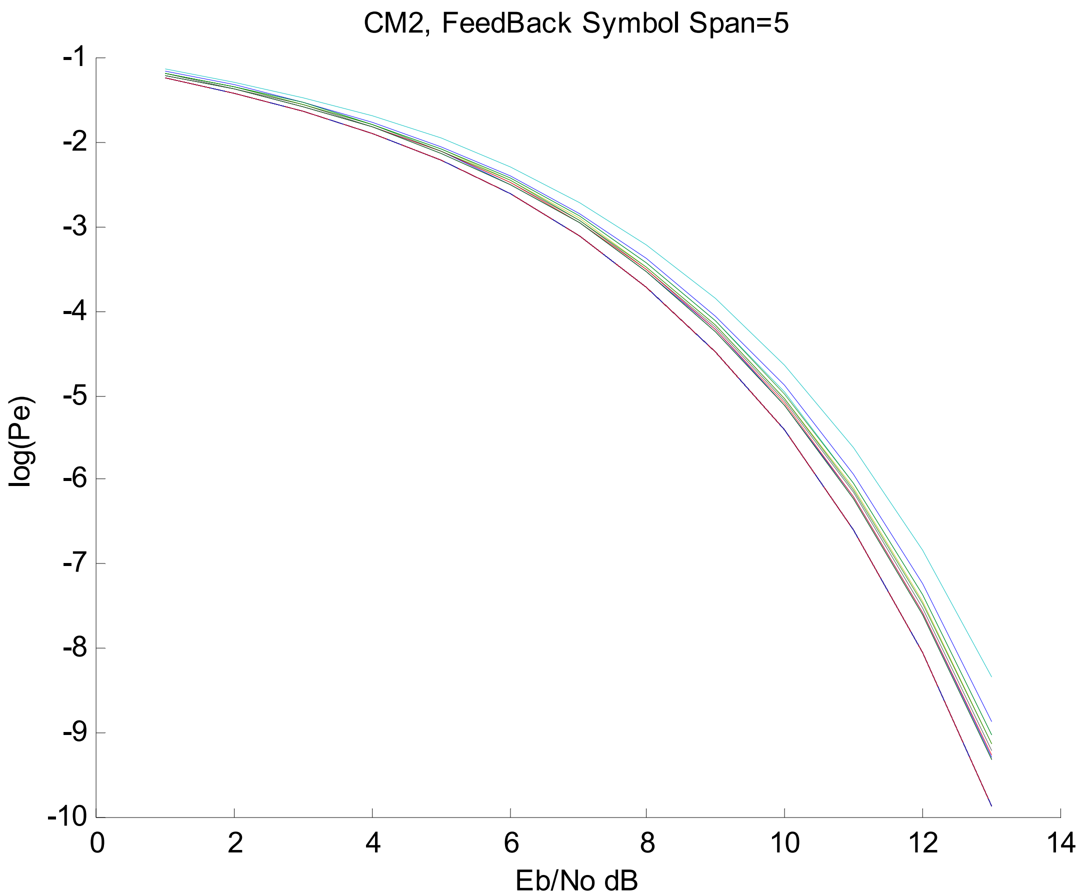
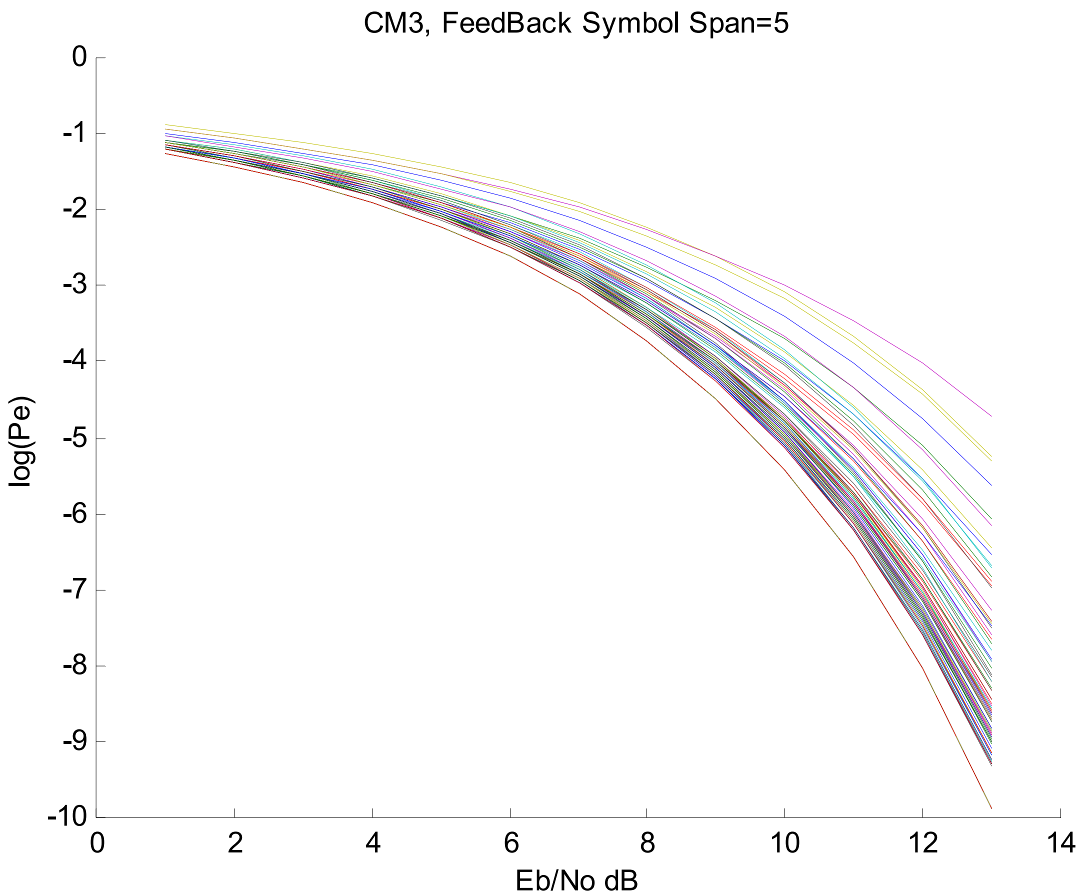
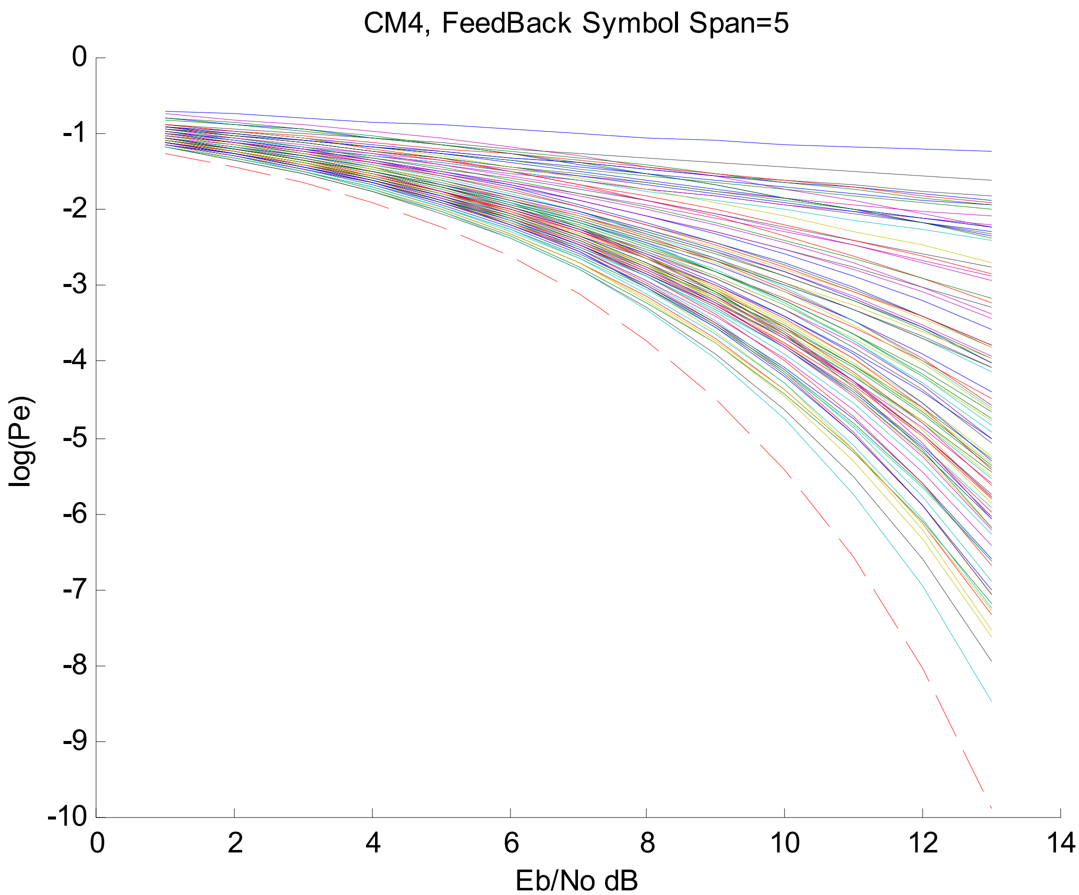


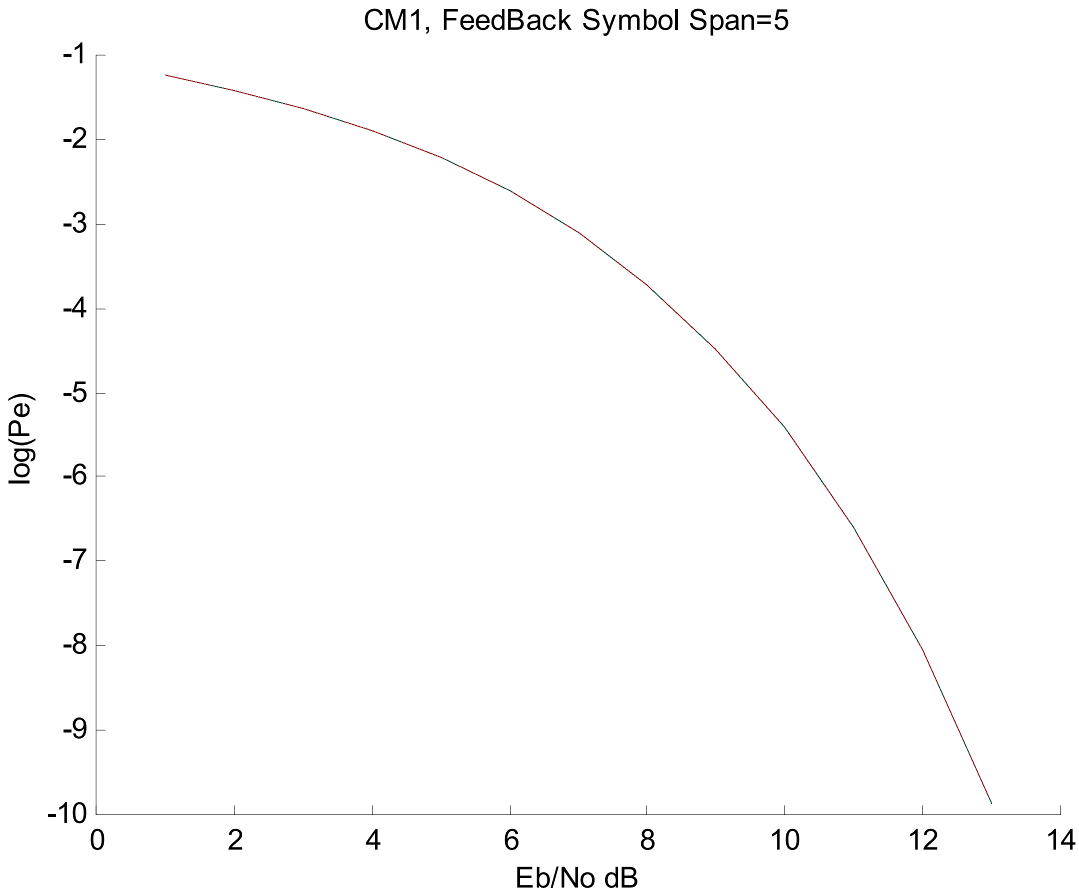
5 Symbol Span Equalizer High Band (1/114e6 symbol duration)



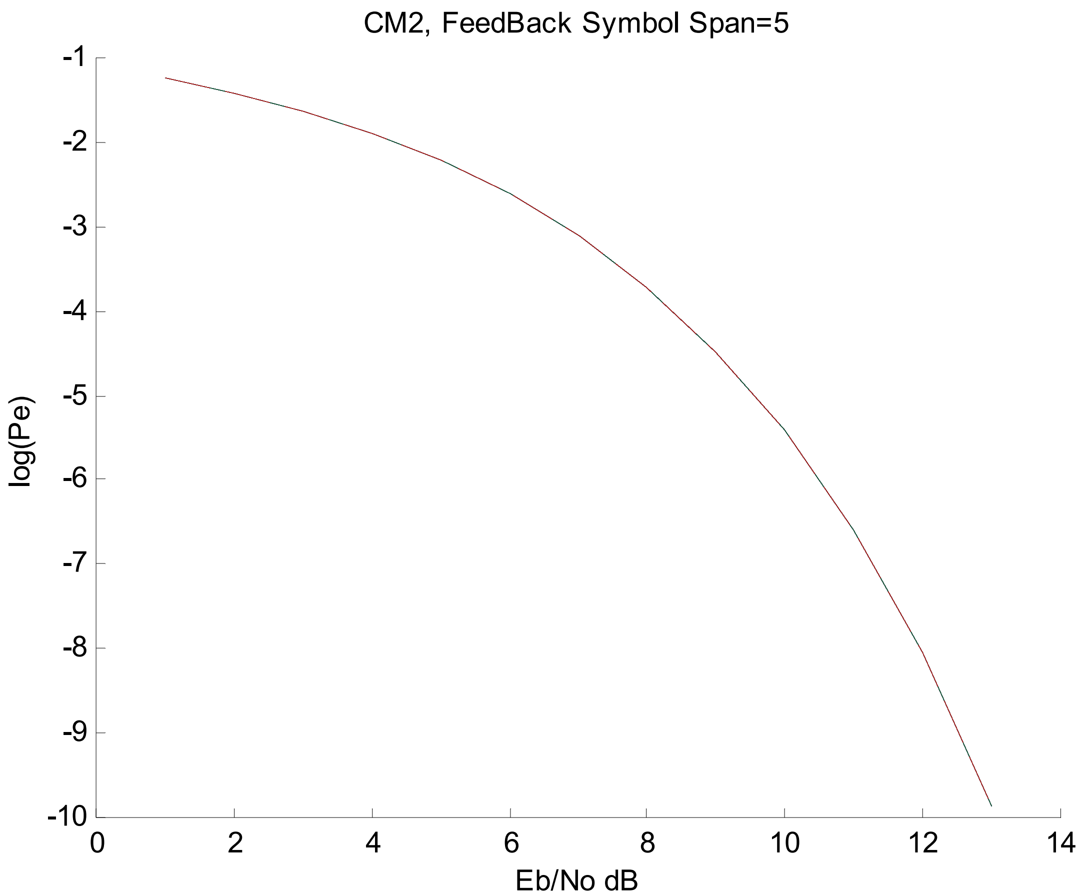
5 Symbol Span Equalizer High Band (1/114e6 symbol duration)



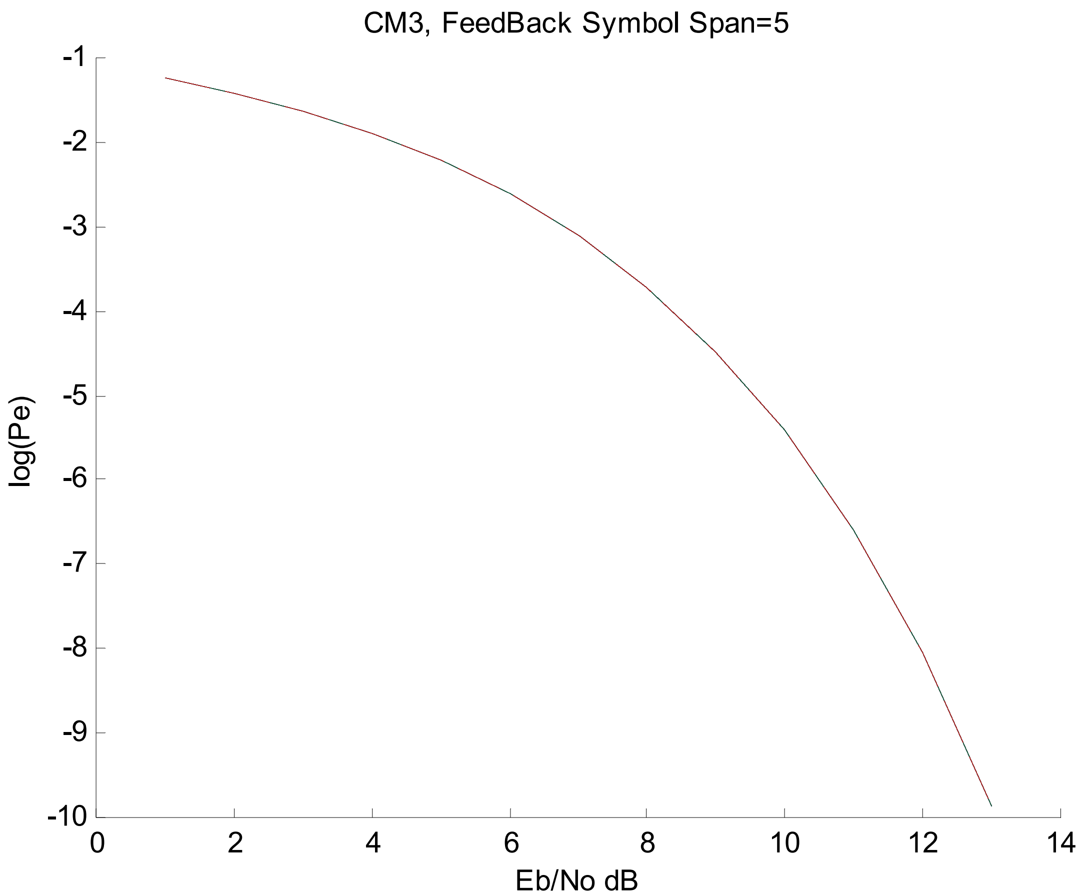
2.8.4 5 Symbol Span Equalizer Low Band (1/57e6 symbol duration)



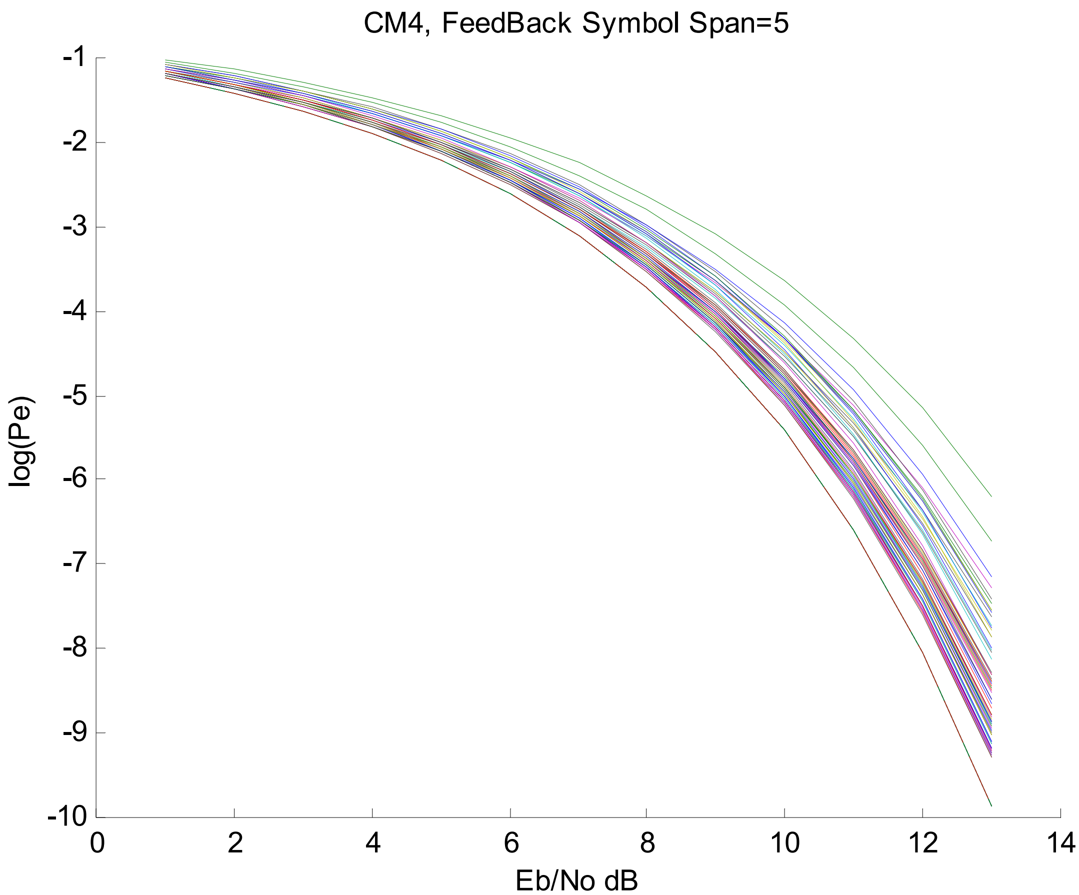
5 Symbol Span Equalizer Low Band (1/57e6 symbol duration)



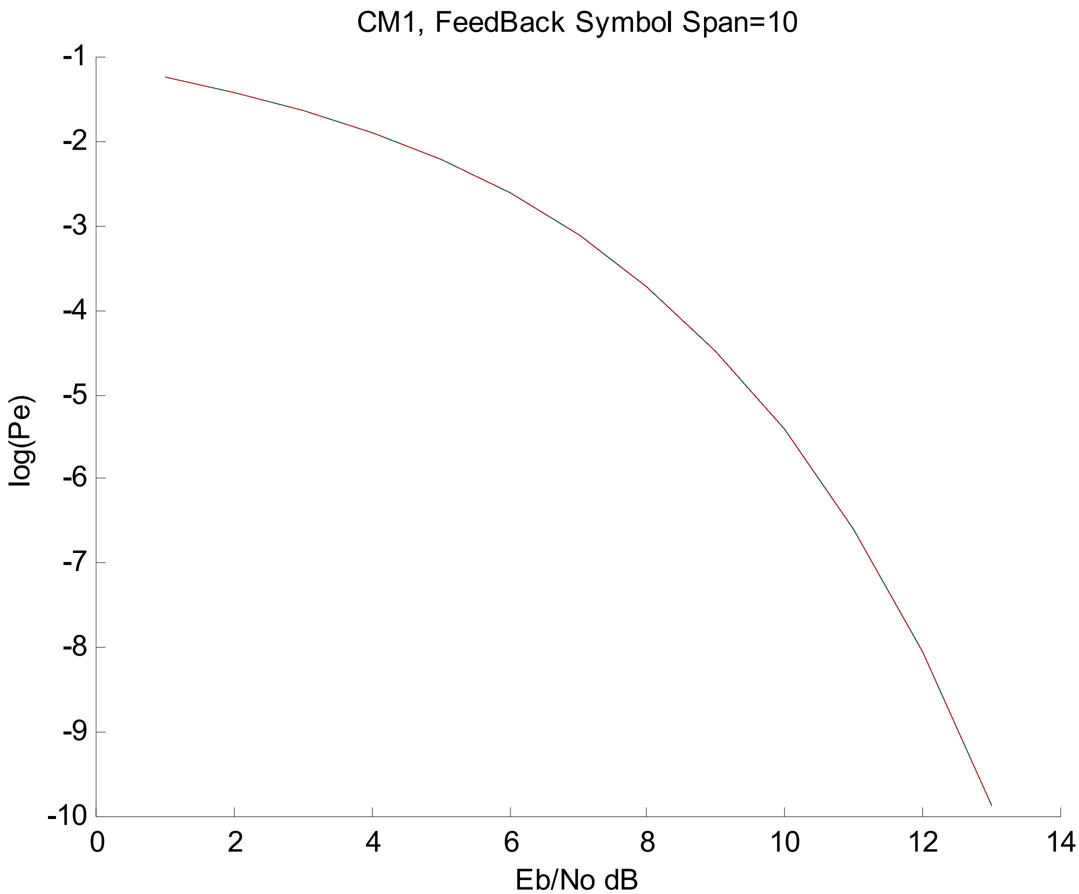
5 Symbol Span Equalizer Low Band (1/57e6 symbol duration)



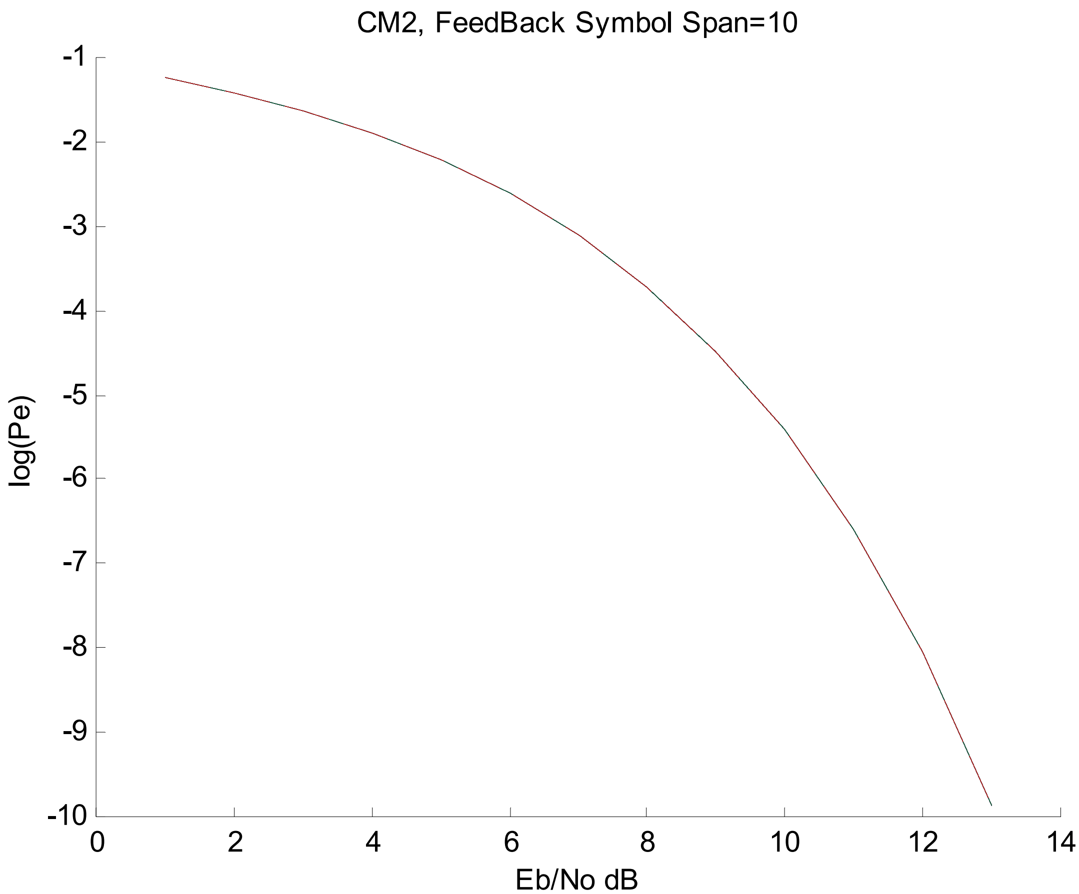
5 Symbol Span Equalizer Low Band (1/57e6 symbol duration)



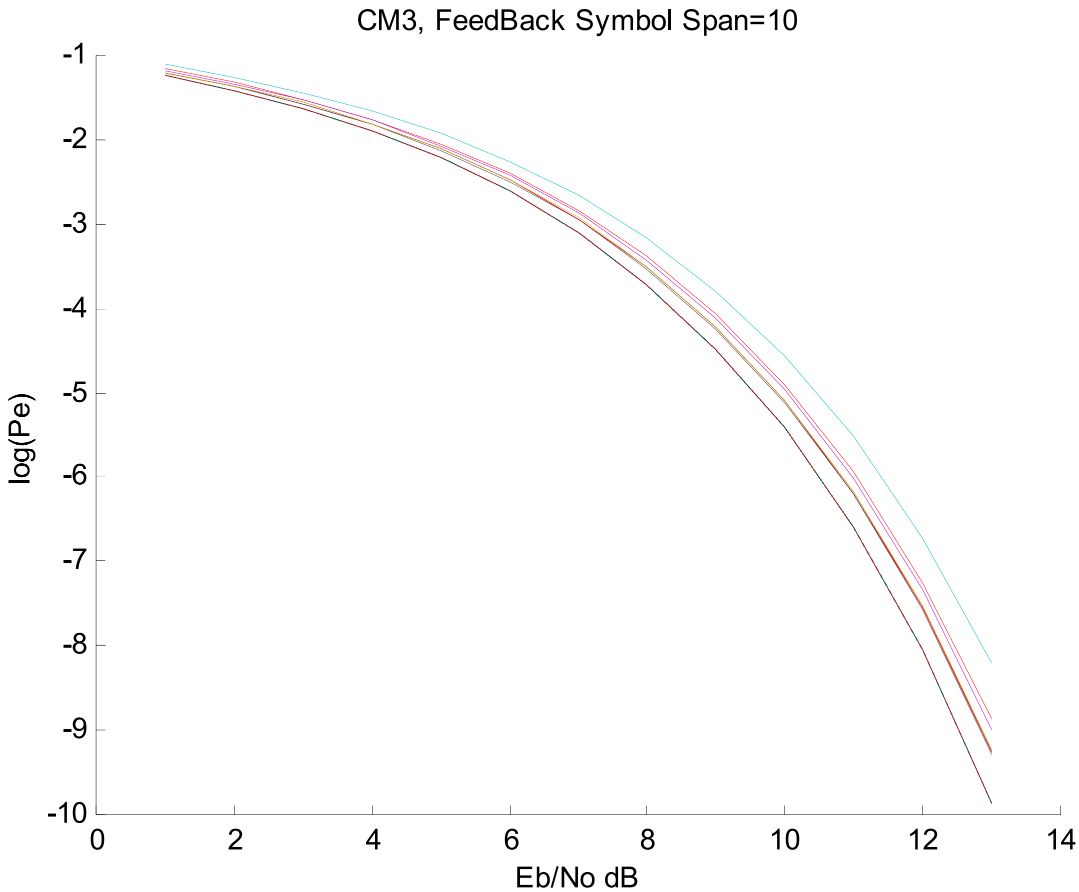
2.8.5 10 Symbol Span Equalizer High Band (1/114e6 symbol duration)



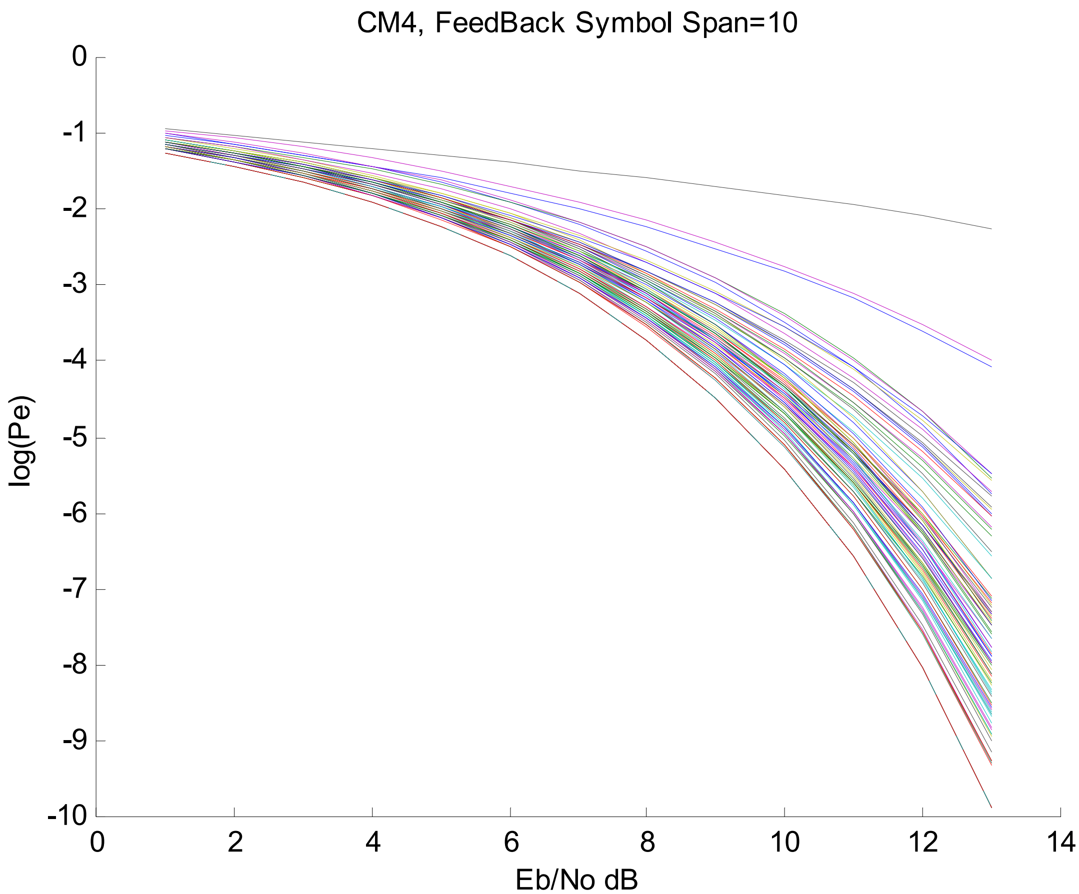
10 Symbol Span Equalizer High Band (1/114e6 symbol duration)



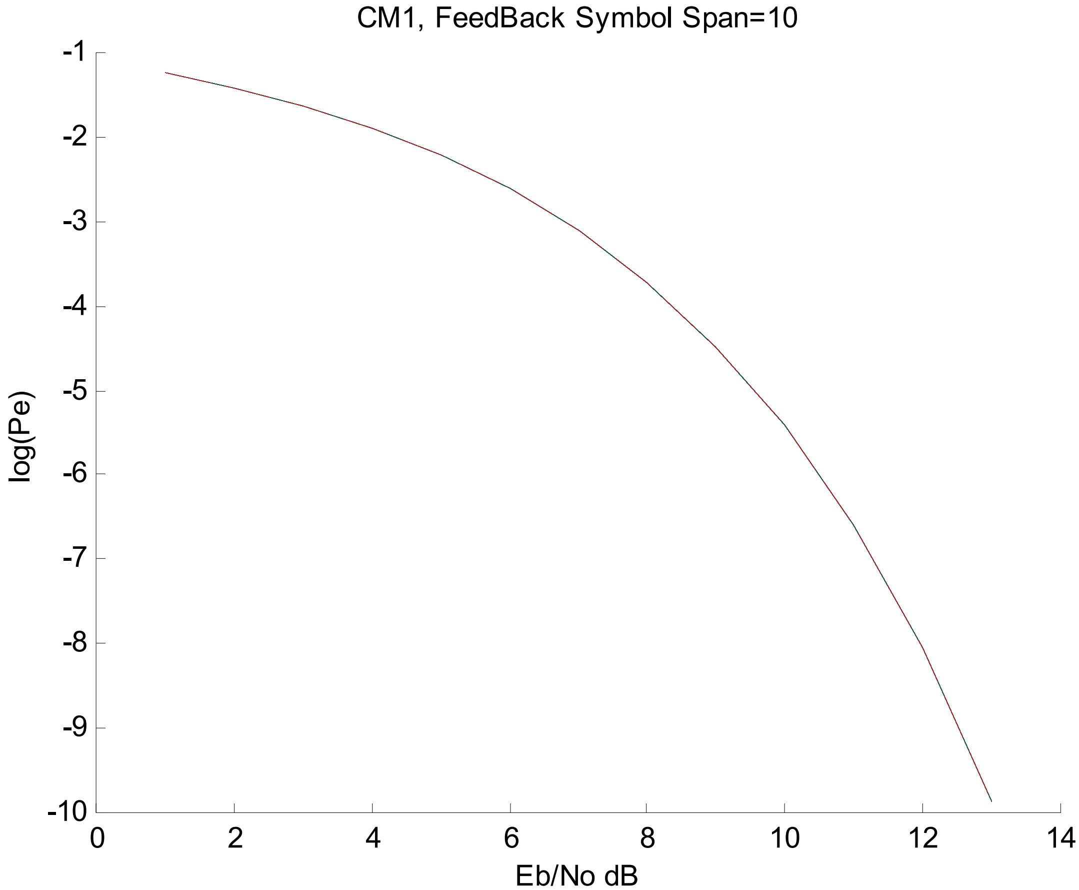
10 Symbol Span Equalizer High Band (1/114e6 symbol duration)



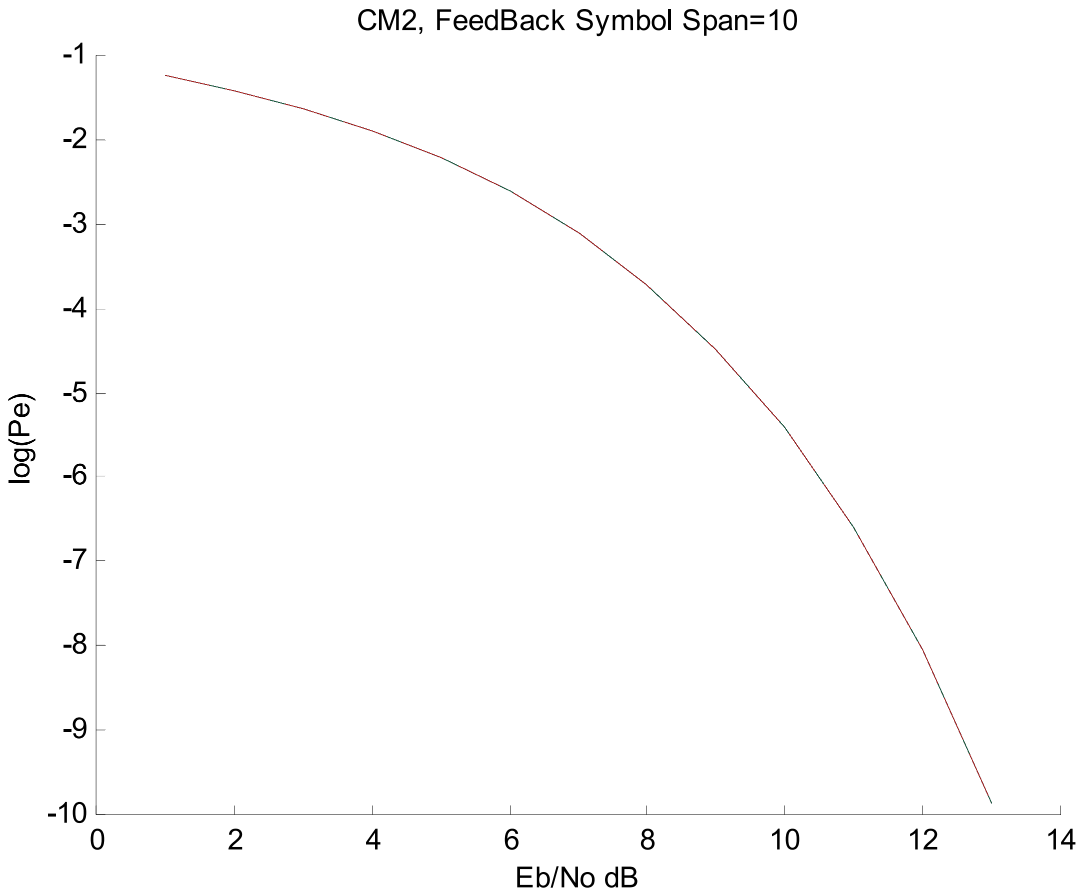
10 Symbol Span Equalizer High Band (1/114e6 symbol duration)



2.8.6 10 Symbol Span Equalizer Low Band (1/57e6 symbol duration)



10 Symbol Span Equalizer Low Band (1/57e6 symbol duration)



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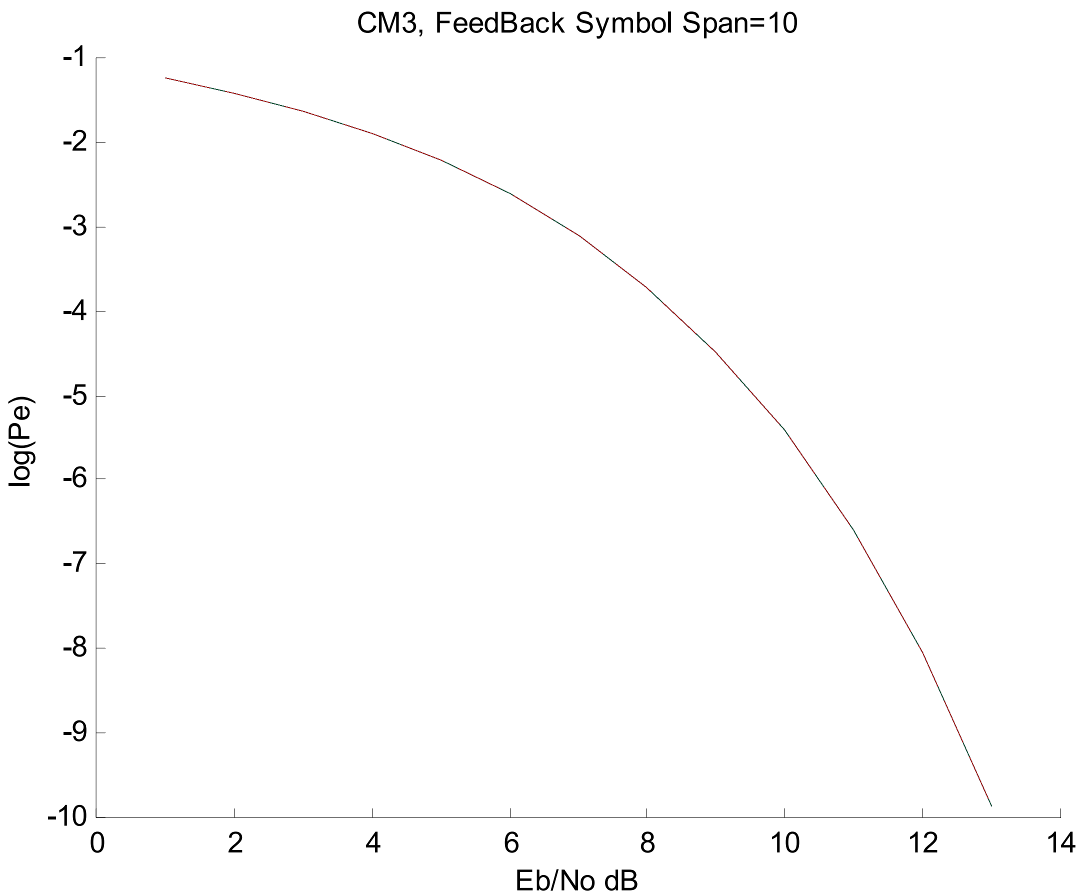

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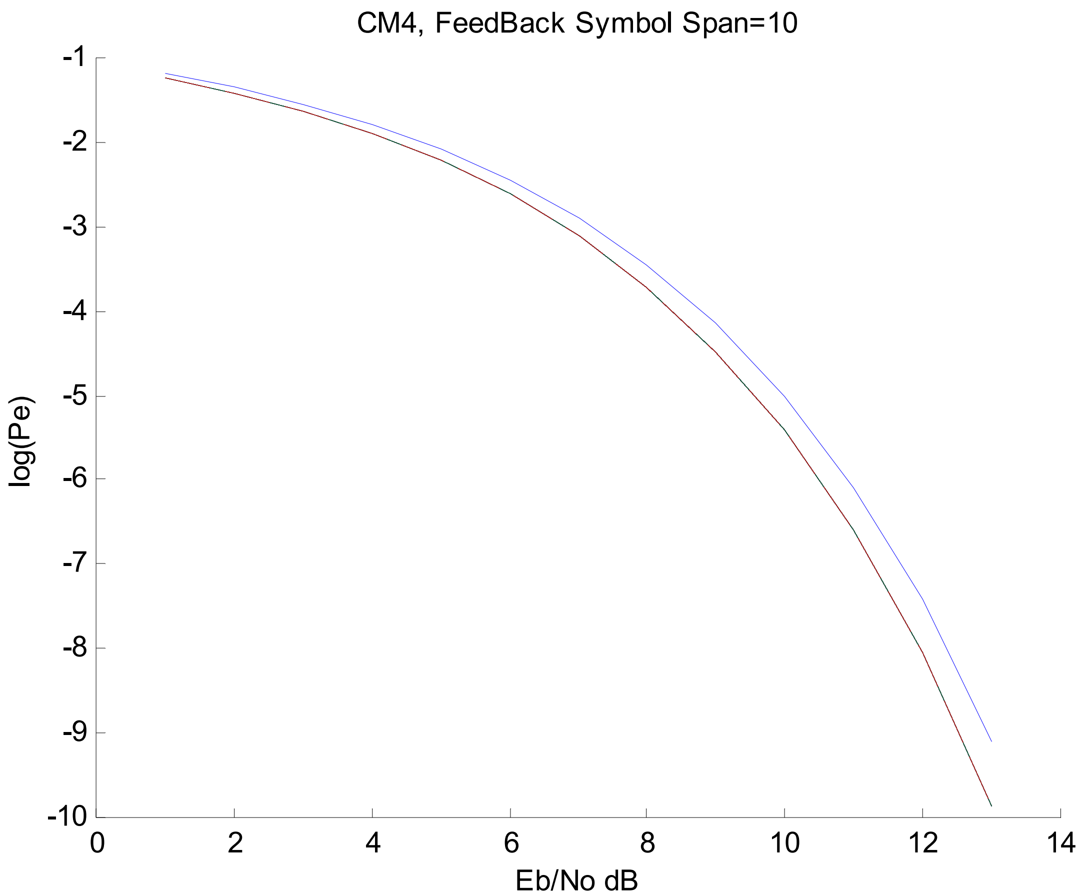


2.9. Acquisition ROC Curvers

Required Probability of Detection during Acquisition: 96%
- 4% Probability of Miss

Integration over two acquisition symbols for 3 dB gain
Acquisition Budget:
- Post-FEC BER=2.47e-6 (PER due to Payload Error=2%)
- PER due to Header Error=2%

Total Error Probability: 8%

2.9.1 ROC Curve for High Band Acqusition

2.9.2 ROC Curve for Low Band Acqusition

2.10. PER Curves (AWGN)

2.11. PER Performance with Multipath Channels

2.12. DFE Error Propagation

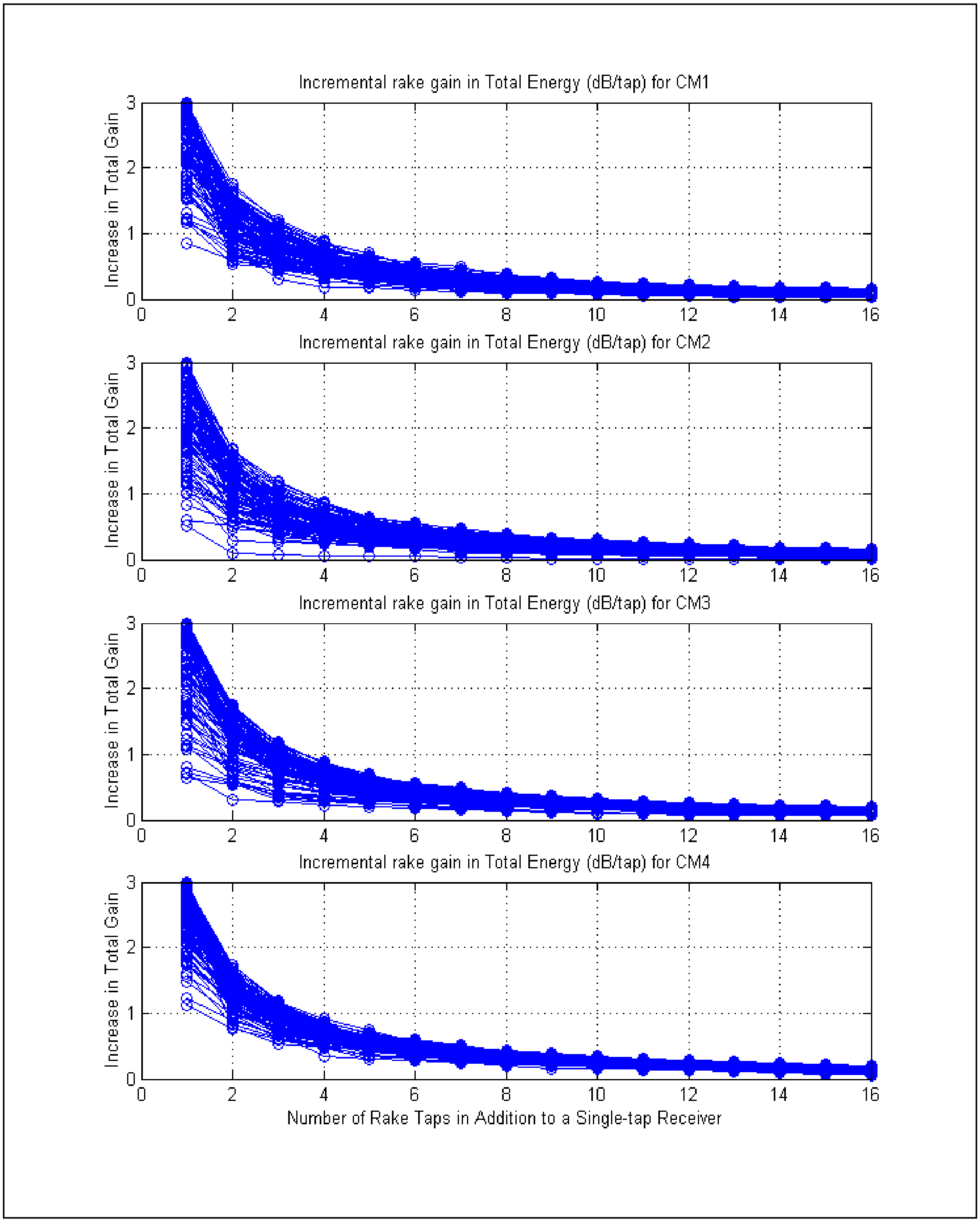
DFE burst error propagation doesn't appear to be a significant problem. A test was ran where a burst of 6 errors was introduced at two different Eb/No's, 9.6 dB and 12.6 dB. The number of resulting DFE errors was recorded after having trained the DFE against the TG3a channel. This was done for each of the 400 TG3a channels. Most of the time we just got 6 errors "out" for 6 errors "in", but occasionally we got 7 or 8 errors out given 6 input errors. The number of additional errors was proportional to the severity of the channel with CM4 giving the worst burst error expansion. On rare occasions CM4 would give more than 8 errors "out" for 6 errors "in". Typical results are shown below.

Number of output errors for an input of 6 burst errors

	Num Errors @ Eb/No=12.6 dB	**Num Errors @ Eb/No=9.6 dB**
CM1	6 on all 100 channels	6 on all 100 channels
CM2	6 on all 100 channels	7 on 26 channels, 6 on 74 channels
CM3	7 on 4 chans., 6 on 96 chans.	7 on 8, 8 on 3, 11 on 1, 6 on 88 channels
CM4	7 on 12, 8 on 1, 11 on 1, 6 on 86	7 on 16, 8 on 8, 9 on 1, 13 on 2, 6 on 73

2.13. RAKE Gain Considerations

The following figure shows the incremental benefits of additional RAKE fingers with respect to the TG3a multipath channels. Notice that regardless of the channel model, having more than 5 fingers in the RAKE yields a diminishing return of less than 1 dB as additional fingers are added. While an implementer can have as many fingers as they desire in their RAKE implementation, we generally feel that having more than 5 is unnecessary.

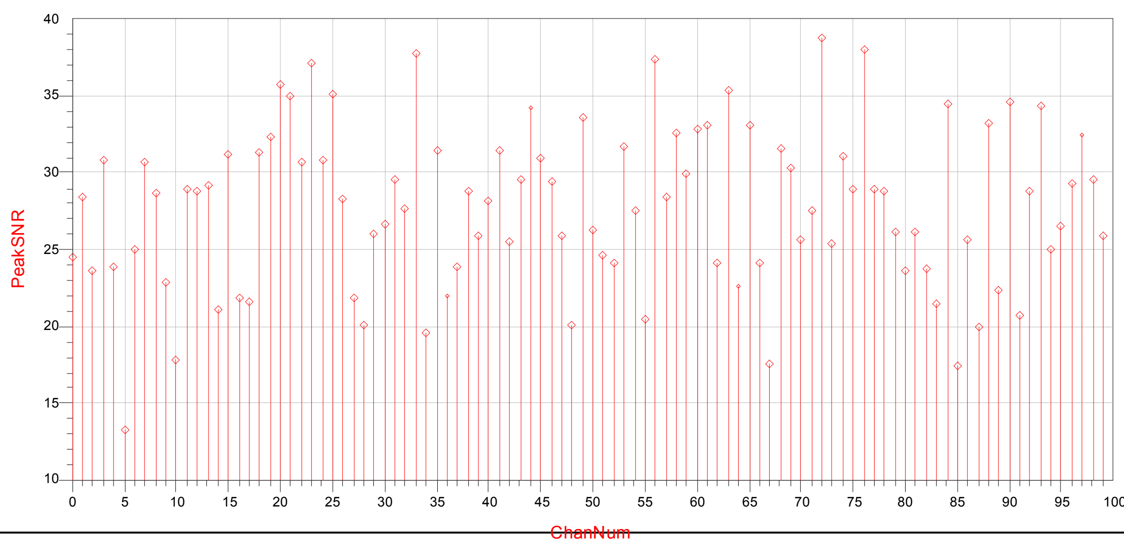

Incremental rake gain in Total Energy (dB/tap) for CM1
Increase in Total Gain
Incremental rake gain in Total Energy (dB/tap) for CM2
Increase in Total Gain
Incremental rake gain in Total Energy (dB/tap) for CM3
Increase in Total Gain
Incremental rake gain in Total Energy (dB/tap) for CM4
Increase in Total Gain
Number of Rake Taps in Addition to a Single-tap Receiver

2.14. *CCA Performance with Multipath Channels*

Clear Channel Assessment (CCA) is generally provided via either a sliding correlator or an energy detector. At this time the state-of-the-art makes a sliding correlator impractical at UWB data rates and bandwidths[1]. The immediate option is energy detection. Energy detection can be realized for BPSK via a squaring circuit (centered on the carrier frequency) and for QPSK via a quadrupling circuit.

2.14.1 BPSK Squaring Circuit with the CM1 channel model

The CM1 channel model was used to explore the performance of a squaring circuit working with BPSK in a multi-path environment. The same multi-path model was used for evaluation at both 4 meters and 10 meters. The range attenuation scaling is based upon 1/r^2 for the largest component of the channel impulse response. The performance metric was the SNR at the output of the squaring circuit in a 200 KHz detection bandwidth. The TX power is –41.3 dBm/MHz and the receiver noise figure is 7 dB.



Fc=4.1 GHz, D=4 meters … Average SNR=27.8 dB … Number of failures below 6 dB SNR: 0

[1] There are no fundamental limitations in our proposal that would prohibit the use of a sliding correlator for implementing CCA. As a matter of fact, future implementations of UWB will probably be DSP based.

2.14.2 BPSK Squaring Circuit with frequency offset detection

It is suggested that the frequency offsets be ±3MHz and ±9MHz and that the PNC attempts to pick a frequency offset that provides maximum distance from its' neighbors.

The following analysis was done for 7 piconets spaced at 1 MHz intervals, but the concepts are still applicable to 4 piconets spaced at 3 MHz intervals.

Multi-piconet Identification Analysis

The multi-piconet environment can be modeled as a vector of signals

$$V_s(t) = \left[S_{-3}(t) \quad S_{-2}(t) \quad S_{-1}(t) \quad S_0(t) \quad S_{+1}(t) \quad S_{+2}(t) \quad S_{+3}(t)\right]$$

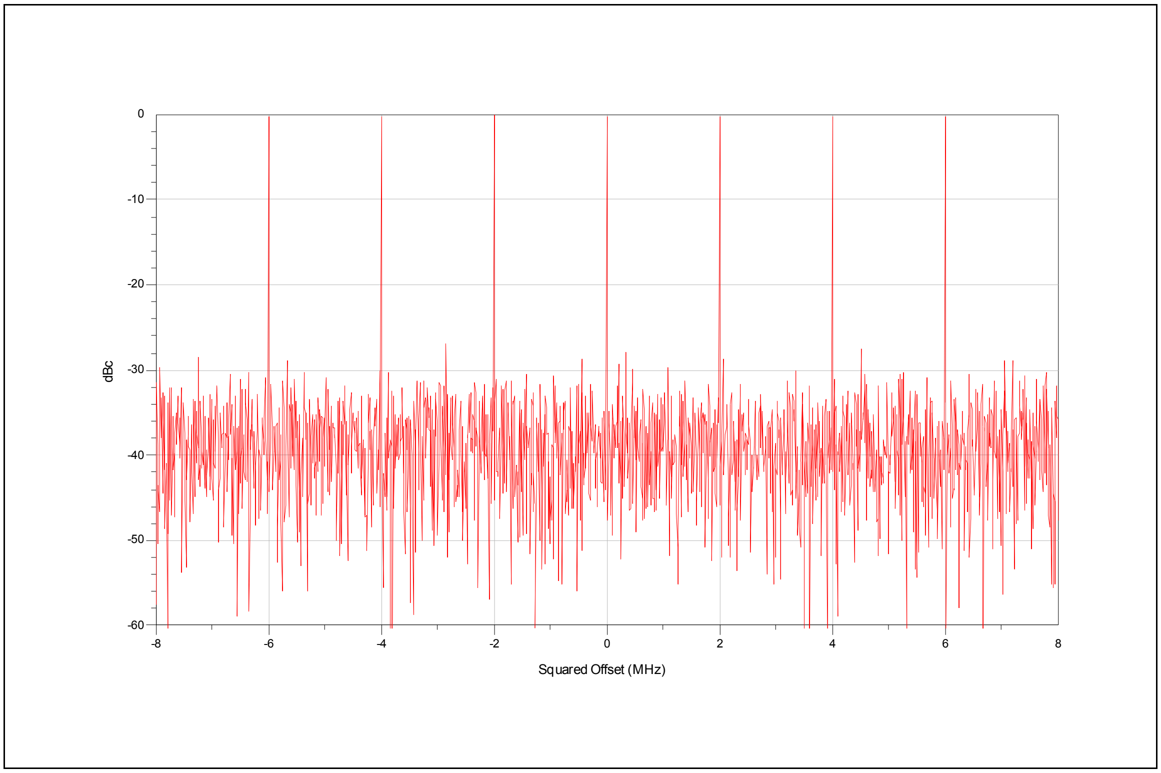
where $S_i(t) = m_i(t) * \cos\{(\omega_0 + \omega_i)t\}$ and ω_i is the frequency offset and m_i is the time dependent modulation. This vector will be processed by a square law device (squaring circuit).

The matrix product is given as

$$V_S^T(t) * V_S(t) = \begin{matrix}
S_{-3}^2(t) & S_{-3}(t)S_{-2}(t) & S_{-3}(t)S_{-1}(t) & S_{-3}(t)S_0(t) & S_{-3}(t)S_{+1}(t) & S_{-3}(t)S_{+2}(t) & S_{-3}(t)S_{+3}(t) \\
S_{-2}(t)S_{-3}(t) & S_{-2}^2(t) & S_{-2}(t)S_{-1}(t) & S_{-2}(t)S_0(t) & S_{-2}(t)S_{+1}(t) & S_{-2}(t)S_{+2}(t) & S_{-2}(t)S_{+3}(t) \\
S_{-1}(t)S_{-3}(t) & S_{-1}(t)S_{-2}(t) & S_{-1}^2(t) & S_{-1}(t)S_0(t) & S_{-1}(t)S_{+1}(t) & S_{-1}(t)S_{+2}(t) & S_{-1}(t)S_{+3}(t) \\
S_0(t)S_{-3}(t) & S_0(t)S_{-2}(t) & S_0(t)S_{-1}(t) & S_0^2(t) & S_0(t)S_{+1}(t) & S_0(t)S_{+2}(t) & S_0(t)S_{+3}(t) \\
S_{+1}(t)S_{-3}(t) & S_{+1}(t)S_{-2}(t) & S_{+1}(t)S_{-1}(t) & S_{+1}(t)S_0(t) & S_{+1}^2(t) & S_{+1}(t)S_{+2}(t) & S_{+1}(t)S_{+3}(t) \\
S_{+2}(t)S_{-3}(t) & S_{+2}(t)S_{-2}(t) & S_{+2}(t)S_{-1}(t) & S_{+2}(t)S_0(t) & S_{+2}(t)S_{+1}(t) & S_{+2}^2(t) & S_{+2}(t)S_{+3}(t) \\
S_{+3}(t)S_{-3}(t) & S_{+3}(t)S_{-2}(t) & S_{+3}(t)S_{-1}(t) & S_{+3}(t)S_0(t) & S_{+3}(t)S_{+1}(t) & S_{+3}(t)S_{+2}(t) & S_{+3}^2(t)
\end{matrix}$$

All the signals off the main diagonal represent the product of two uncorrelated spread spectrum signals which yields just another spread spectrum signal (represents an increase in the noise floor). However, the trace represents the square-law product sum of the signals $S_i^2(t) = m_i^2(t) * \cos^2\{(\omega_0 + \omega_i)t\}$. The expectation of each double frequency term is given by $\overline{S_i^2(t)} = \frac{1}{2} * \overline{m_i^2(t)} * \cos\{2(\omega_0 + \omega_i)t\}$ where we can assume that $\overline{m_i^2(t)} \approx 1$. Thus we see that the trace terms collapse to a double frequency component and the cross-product terms (off main diagonal terms) simply raise the noise floor. Assuming each piconet uses a unique chipping rate offset, the output of the squaring loop can be used for piconet identification.

An example output of the squaring circuit for an input of 7 piconets is shown below. These results readily scale to operation with less than 7 piconets, such as with 4 piconets.



Seven Term Squared Output

2.14.3 BPSK Squaring Circuit ROC Curves – 8.2 GHz, 2 meters

2.14.4 BPSK Squaring Circuit ROC Curves – 8.2 GHz, 4 meters

2.14.5 BPSK Squaring Circuit ROC Curves – 4.1 GHz, 2 and 4 meters

2.14.6 BPSK Squaring Circuit ROC Curves – 4.1 GHz, 10 meters

2.15. Simultaneously Operating Piconets

2.15.1 CDMA Code Set Characteristics

The CDMA code set contains 32 codes which are grouped into four groups of 8 codes as depicted below:

Code Set 1	**Code Set 1**	**Code Set 1**	**Code Set 1**
➢ $CW1_0$	➢ $CW2_0$	➢ $CW3_0$	➢ $CW4_0$
➢ $CW1_1$	➢ $CW2_1$	➢ $CW3_1$	➢ $CW4_1$
➢ $CW1_2$	➢ $CW2_2$	➢ $CW3_2$	➢ $CW4_2$
➢ $CW1_3$	➢ $CW2_3$	➢ $CW3_3$	➢ $CW4_3$
➢ $CW1_4$	➢ $CW2_4$	➢ $CW3_4$	➢ $CW4_4$
➢ $CW1_5$	➢ $CW2_5$	➢ $CW3_5$	➢ $CW4_5$
➢ $CW1_6$	➢ $CW2_6$	➢ $CW3_6$	➢ $CW4_6$
➢ $CW1_7$	➢ $CW2_7$	➢ $CW3_7$	➢ $CW4_7$

The rms cross-correlation between any code in one group and any code in a different group <-14 dB while the cross-correlation between any code within the same group <7.6 dB.

2.15.2 Single Co-channel separation distance test

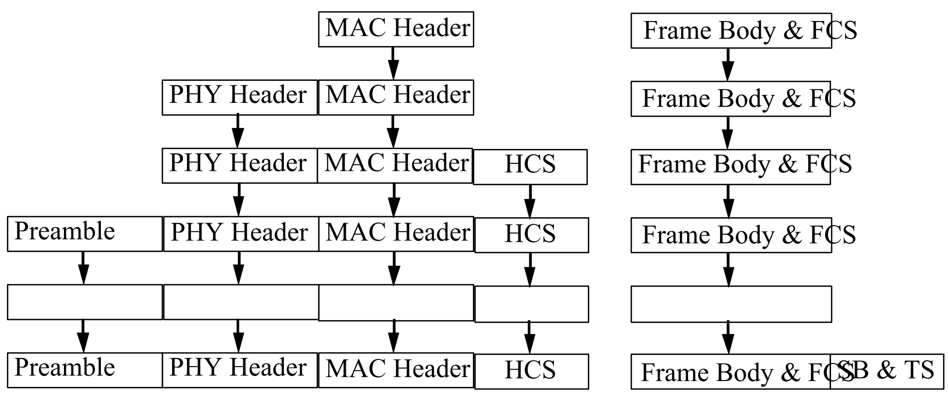
CM1 Testing
CM2 Testing
CM3 Testing
CM4 Testing

2.16. MAC Enhancements Estimates

3 PHY Technical Detail

3.1 Frame Format

The PHY frame format for all data rate modes is illustrated below. The UWB PHY prepends the PHY header to the MAC header, calculates the HCS, and appends this to the MAC header. If the size of the frame body plus FCS, in bits, is not an integer multiple of the bits/symbol, then stuff bits are added following the FCS. The PHY preamble is sent first in the packet, followed by the PHY and MAC header, followed by the MPDU and finally the tail symbols.



PHY frame formatting

3.2 Randomization

Randomization shall be employed to ensure an adequate number of bit transitions to support clock recovery. The stream of downlink packets shall be randomized by modulo-2 addition of the data with the output of the pseudo-random binary sequence (PRBS) generator. The randomizer shall be used for the MAC header and frame body. The PHY preamble and PHY header shall not be scrambled. The polynomial for the pseudo random binary sequence (PRBS) generator shall be

$g(D)=1+D^{14}+D^{15}$ (4).

The polynomial forms not only a maximal length sequence, but also is a primitive polynomial. By the given generator polynomial, the corresponding PRBS is generated as

$$Xn = Xn - 14 \oplus Xn - 15 .$$

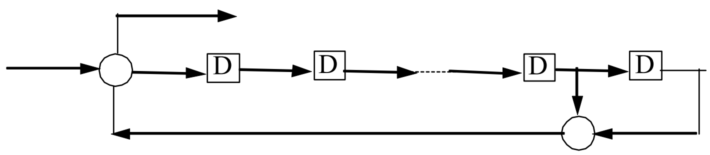


Figure 3 - Realization of the randomizer linear feedback shift registers

The following sequence defines the initialization sequence,

$x_{init}=[x^{i}n\text{-}1\ x^{i}n\text{-}2\ x^{i}n\text{-}3\ x^{i}n\text{-}4\ x^{i}n\text{-}5\ x^{i}n\text{-}6\ x^{i}n\text{-}7\ x^{i}n\text{-}8\ x^{i}n\text{-}9\ x^{i}n\text{-}10\ x^{i}n\text{-}11\ x^{i}n\text{-}12\ x^{i}n\text{-}13\ x^{i}n\text{-}14\ x^{i}n\text{-}15]$ (6)

where $x^{i}n\text{-}k$ represents the binary initial value at the output of the k^{th} delay element.

The scrambled data bits, s_n, are obtained as follows

$$s_n = b_n \oplus x_n$$

where b_n represents the unscrambled data bits. The side-stream de-scrambler at the receiver shall be initialized with the same initialization vector, x_{init}, used in the transmitter scrambler. The initialization vector is determined from the seed identifier contained in the PHY header of the received packet.

Randomizer Seed Selection

Seed Identifier	Seed Value
0,0	0011 1111 1111 111
0,1	0111 1111 1111 111
1,0	1011 1111 1111 111
1,1	1111 1111 1111 111

The 15 bit seed value chosen shall correspond to the seed identifier. The seed identifier value is set to 00 when the PHY is initialized and is incremented in a 2-bit rollover counter for each packet that is sent by the PHY. The value of the seed identifier that is used for the packet is sent in the PHY header. The 15-bit seed value is configured as follows. At the beginning of each PHY frame, the register is cleared, the seed value is loaded, and the first scrambler bit is calculated. The first bit of data of the MAC header is modulo-2 added with the first scrambler bit, followed by the rest of the bits in the MAC header and frame body.

3.3 FEC coding

The forward error correction (FEC) schemes are selectable from the types in Table 5. All FEC schemes shall be mandatory and supported by all DEVs.

FEC Code Types

Code Type	Outer Code	Inner Code
1 - Reed-Solomon	RS GF(255,223)	None
2 – Convolutional	None	Rate ½ or ¾, K=6, (65, 57) Convolutional Code
3 – Concatenated	RS GF(255,223)	Rate ½ or ¾, K=6, (65, 57) Convolutional Code

The following is a summary of the three Code Types:

a)***Code Type 1:*** *Reed-Solomon only:* Reed-Solomon decoders offer implementation advantages that extend practical bit rate limits far beyond what is possible with convolutional decoders (specifically the Viterbi algorithm decoder). Reed-Solomon codes offer high coding rate at high bit rates with moderate latency. The protection offered by the GF(255, 223) code is t =16. The code works on octets and not individual bits.

b)***Code Type 2:*** *Convolutional code:* This code is useful for low to moderate coding rates providing good coding gain with low latency. Implementation issues often limit the practical bit rate to less than 200 Mbps. The basic code is a ½ rate code that can be punctured to achieve a code rate of ¾ at slightly less coding gain. Typically an interleaver is used prior to the convolutional decoder to randomize burst errors.

b)***Code Type 3:****Reed-Solomon + Convolutional code:* A Reed-Solomon code concatenated with a Convolution code offers excellent coding gain for those applications that can tolerate the low to moderate code rate and longer latency. The bit rate is generally limited by the convolutional decoder. A symbol block interleaver is required between the convolutional decoder and the Reed-Solomon decoder to random burst errors caused by the convolutional decoder.

3.3.1 Reed-Solomon GF(255,223) Code

RS(255, 223) codes are a class of block codes that operate on non-binary symbols. The symbols are formed from 8 bits of a binary data stream and a code block is formed with 255 symbols. In each block 223 symbols are formed from the encoder input and $2t$ parity symbols are added where t=16 in this case. The code is thus a systematic code, with a code rate of 223/255, and the code is able to correct up to 16 symbol errors in a block. The RS generator polynomial can be reduced to a $2t$ order polynomial as shown below.

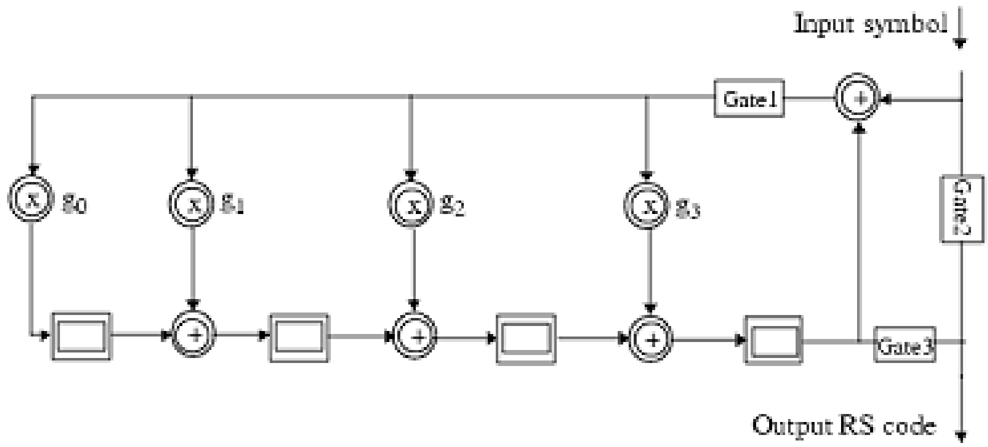
$g(x)=x^{2t} + g_{2t-1}x^{2t-1} + ... + g_0.$

The RS field generation polynomial is specified as below.

P(x) = 1 + x2 + x3 + x4 + x8

For last codeword or the MAC message less than K, shortened operation is performed. Let k' as the message length, and the shortened operation is described as:

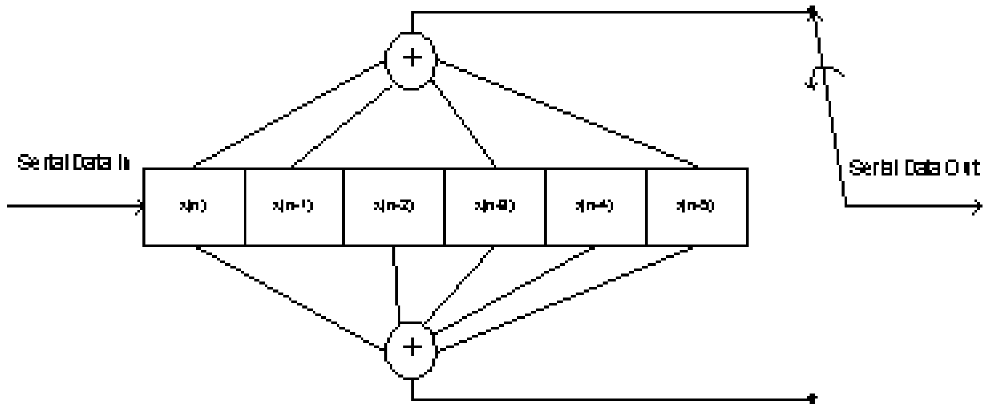
A1) Add (K-k') zero bytes to the block as a prefix.
A2) Encode the K bytes and append the R parity bytes.
A3) Discard all of the (K-k') zero symbols.
A4) Serialize the bytes and transmit them to the inner coder or the modulator MSB first.



Reed-Solomon Encoder

3.3.2 Convolutional Code

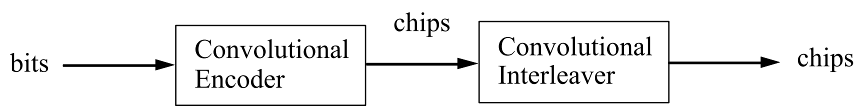
The convolutional encoder is used to encode data so that errors introduced due to noise in the channel can be corrected by the decoder. Two important characteristics of a convolutional encoder are its rate and constraint length. If k data bits are shifted in for every n encoded bits shifted out, the rate of the code equals k/n. If the maximum degree of the generator polynomials are m, then the constraint length of the code equals k(m+1). The half-rate convolutional encoder is a linear feed-forward shift register network in which, for every data bit that is shifted in, 2 encoded bits are shifted out. The structure of the encoder is specified by a set of generator polynomials go=65 and g1=57. The error correcting capabilities of a code are determined by these generator polynomials.



Rate ½ Encoder

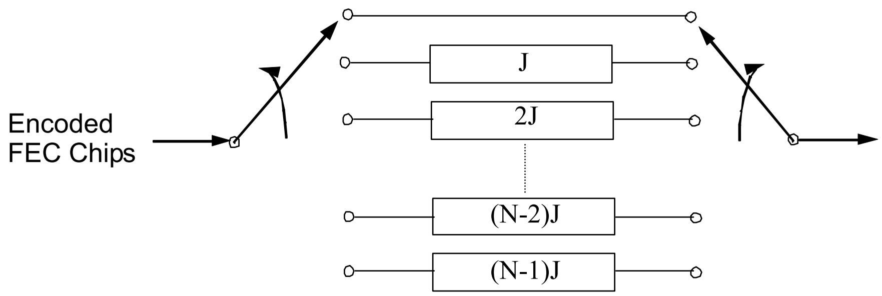
3.3.3 Chip Wise Convolutional Interleaver

The convolutional decoder is sensitive to burst errors; hence, interleaving is used to disperse burst errors as shown below.



Convolutional Encoder with Interleaving

Convolutional interleaving is used over block interleaving because of it has lower latency and memory requirements. The structure for a convolutional interleaver is shown below. The encoded chips are sequentially shifted in to the bank of N registers; each successive register provides J chips more storage than did the preceding. The zeroth register provides no storage. With each new code chip the commutator switches to a new register, and the new code chip is shifted in while the oldest code chip in that register is shifted out. After the (N-1)th register, the commutator returns to the zeroth register and starts again. The deinterleaver performs the inverse operation, and the input and output commutators for both interleaving and deinterleaving must be synchronized.

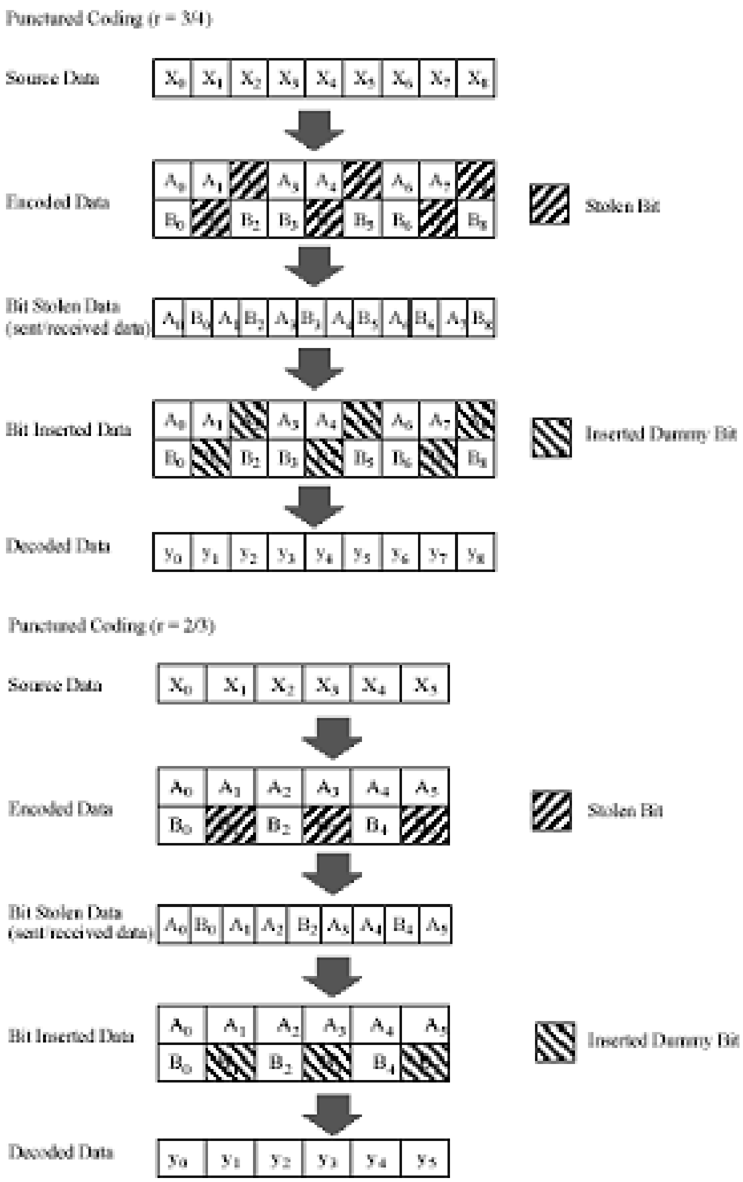


Convolutional Chip-wise Interleaver

The chip interleaver shall have the values of J=7 and N=10.

3.3.4 Puncturing

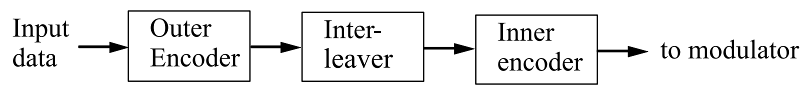
Higher data rates are derived from convolutional encoders by employing “puncturing.” Puncturing is a procedure for omitting some of the encoded bits in the transmitter (thus reducing the number of transmitted bits and increasing the coding rate) and inserting a dummy “zero” metric into the convolutional decoder on the receive side in place of the omitted bits. This allows a 1/2 rate code to be transformed into a 2/3 rate code or a 3/4 rate code. The puncturing patterns are illustrated in Figure X. Decoding by the Viterbi algorithm is recommended.



Puncturing Patterns

3.3.5 Concatenated Code

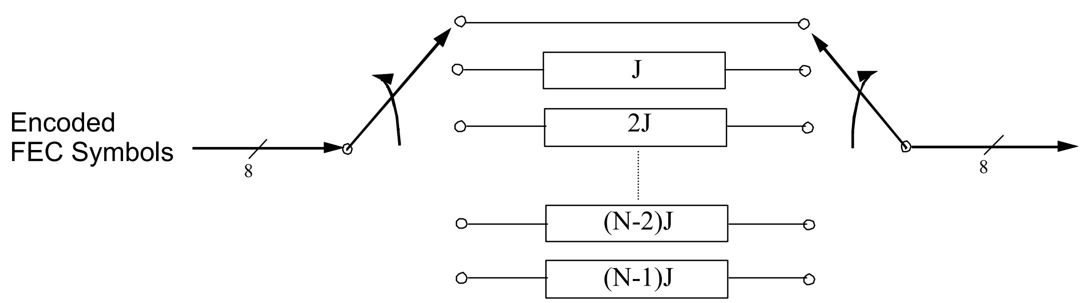
A concatenated code is one that uses two levels of coding, an inner code and an outer. Figure X shows a concatenated encoding scheme with an interleaver between the outer encode and the inner encoder. The concatenated code shall consist of a Reed-Solomon outer code (ref. 11.4.1) and a convolutional inner code (ref. 11.4.2). A bytewise interleaver shall be used between the outer encoder and the inner encoder.



Concatenated Coding Scheme

3.3.6 Convolutional Symbol Interleaver

Convolutional interleaving is used over block interleaving because of it has lower latency and memory requirements. The structure for a convolutional interleaver is shown below in Figure X. The code symbols are sequentially shifted into the bank of N registers; each successive register provides J symbols more storage than did the preceding. The zeroth register provides no storage. With each new code symbol the commutator switches to a new register, and the new code symbol is shifted in while the oldest code symbol in that register is shifted out. After the (N-1)th register, the commutator returns to the zeroth register and starts again. The deinterleaver performs the inverse operation, and the input and output commutators for both interleaving and deinterleaving must be synchronized.



Convolutional Symbol-wise Interleaver

The symbol interleaver shall have the values of J=21 and N=15.

3.4 Code Set Modulation

3.4.1 Base Symbol Rate

The IEEE 802.15.3a UWB physical layer standard specifies bi-phase modulated CDMA coded BPSK wavelet signalling. The base symbol rate for all modulations shall be 114 Msps (Mega Symbols per Second) for the high band and 57 Msps for the low band**.**

3.4.2 Code Sets and Code Set Modulation

This proposal enables higher data rates primarily through the use of bi-orthogonal keying (BOK). This modulation format uses a set of orthogonal codes and their inverses to send multiple bits of information in each symbol interval. This modulation allows higher data rates without the need to increase ADC clock rates and also provides more robust BER performance through lower Eb/No requirements.

There are 4 piconets with up to 8 code words per piconet. The code words are 24 chips long with each chip selected from the set {-1, 0, -1}. The following tables show half of the code space. The other half of the code space is just the inverse. The codes are used for 2-BOK (first row), 4-BOK (first and second rows), 8-BOK (first, second, third and forth rows) and 16-BOK (all rows).

Piconet 1 BOK Codes

symbol bit mapping	**codeword**
1111	
1110	
1101	
1100	
1011	
1010	
1001	
1000	

Piconet 2 BOK Codes

symbol bit mapping	codeword
1111	
1110	
1101	
1100	
1011	
1010	
1001	
1000	

Piconet 3 BOK Codes

symbol bit mapping	codeword
1111	
1110	
1101	
1100	
1011	
1010	
1001	
1000	

Piconet 4 BOK Codes

symbol bit mapping	codeword
1111	
1110	
1101	
1100	
1011	
1010	
1001	
1000	

3.4.3 Preamble and Header Modulation Rate

The PHY header and MAC header shall be modulated at the base rate using one of the 2-BOK PNC selected formats as shown in the tables with no FEC.

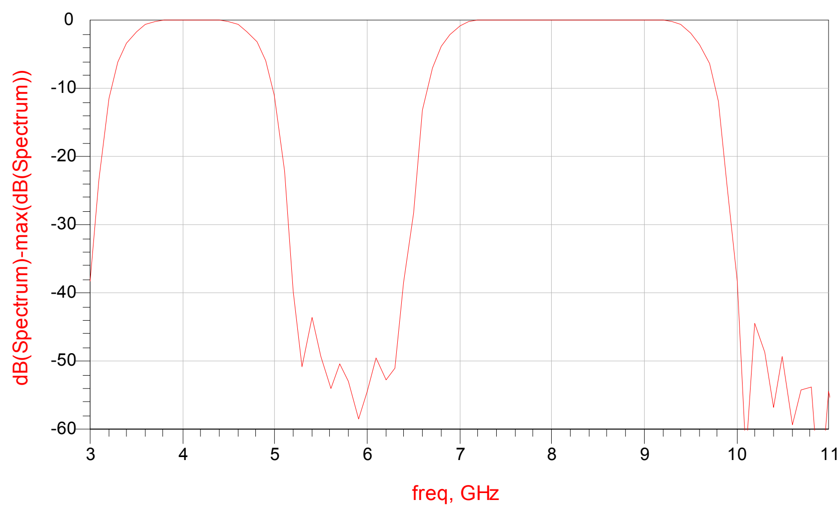
3.5 Pulse Shaping and Modulation

3.5.1 Impulse Response

The reference pulse is a root raised cosine low pass filter. For the low frequency band the filter cutoff frequency (-3 dB point) is 684 MHz and for the high frequency band the filter cutoff frequency (-3 dB point) is 1368 MHz. The implemented baseband impulse response must have a peak cross-correlation within 3 dB of the reference pulse. The excess bandwidth is 50%.

3.5.2 Reference Spectral Mask

The reference spectral mask is shown in figure below. Out-of-band emissions must meet regulatory domain requirements.



Super-impose Lower and Upper Band Reference Pulse Spectral Mask

3.5.3 Chip Rate Clock and Chip Carrier Alignment

The chip rate clock and the chip carrier shall be provided from the same source with the frequencies shown below. The accuracy is 25 ppm.

Chip Rate Clock and Chip Carrier Frequencies

Chip Rate	1.368 Gcps	2.736 Gcps
Carrier Frequency	4.104 GHz	8.208 GHz
RRC Excess BW	50%	50%

3.6 Power Management Modes

3.7 General Requirements

4. Preamble and Acquisition Details

4.1 General Preamble Structure

There are 3 preamble options that are structurally the same except for field durations:

1. A nominal preamble used for nominal data rates and channels
2. A long preamble used for low data rates and difficult channels
3. A short preamble used for high data rates and benign channels

Acq Seq 8 uS	TBD TBD uS	Training 4 uS	SFD 1 uS	PHY Header	MAC Header	data

Preamble Structure

The preamble structure is summarized below.

1. The TX MAC selects one of 4 piconet acquisition codes (PAC) and sets the corresponding carrier offset frequency.

2. The TX modulates the PAC code (one bit per PAC symbol) with random data to generate the acquisition sequence which is used by the receiver for initial acquisition (AGC and clock frequency lock)

3. Next is sent the training frame. The receiver uses this training frame to adjust the receiver.

4. The TX next sends the SFD (start frame delimiter) that indicates to the RX the next frame will be the PHY header, which contains rate information.

5. After the PHY header comes the MAC header.

6. Following the MAC header the TX starts sending data frames.

4.2 Acquisition Sequence

There are four PAC codewords with are given by the first row of tables 6 through 12, and the correpsonding frequency offsets are given in table 14. Each piconet uniquely uses one of these codes. The selection of the code is determined by the PNC during the initial scan prior to initiating the piconet (the PNC selects a PAC codeword that is not in use). Use of the PAC codewords provides a degree of "channel separation" between overlapping piconets during preamble acquisition, limited only by the rms cross-correlation properties of the PAC codeword set. The PAC codewords increases the probability that a DEV will train on the preamble associated with the "desired" piconet.

Piconet number to carrier offset mapping

PNC Number	Chipping Offset (MHz)	Carrier Offset (MHz)
1	-3	-9
2	-1	-3
3	+1	+3
4	+3	+9

The preamble starts with the acquisition sequence, which is used primarily by the receiver to set gains and achieve clock synchronization. The acquisition sequence is random bits, one bit per codeword.

4.3 Training Frame

The length of the training frame varies depending upon the preamble length. Table XYZ indicates the bits that shall be sent during the preamble. The high band bit time duration is 1/114e6 and the low band bit time duration is 1/57e6.

Training Frame Bit Sequence

Preamble Type	High Band Preamble Sequence (base 32)	Low Band Preamble Sequence (base 32)
Short	000000000000000000000000000000000 000	000000000000000000
Medium	000000000000000000000000000000000 000000000000000000000	000000000000000000000000000
Long	000000000000000000000000000000000 000000000000000000000000000000000 000000000000000000000000000000000 000000000	000000000000000000000000000000000 000000000000000000000

The notation for Base32 is: 0123456789ABCDEFGHJKMNPQRSTVWXYZ

4.4 SFD

The SFD consists of the 16-bit binary pattern 0000 1100 1011 1101 (transmitted leftmost bit first) as modulation on the selected 2-BOK code. The first bit of the SFD follows the last bit of the sync pattern. The SFD defines the frame timing in anticipation of the PHY header.

4.5 PHY Header

The PHY header consists of two octets that contain the number of octets in the frame body (including the FCS), the data rate of the frame body and seed identifier for the data scrambler. The fields for the PHY service field are shown in Table 13. Bit b0 is sent over the air first and the other bits follow sequentially.

PHY service field

Bits	Content	Description
b0-b1	Seed Identifier	2 bit field that selects the seed for the data scrambler, defined in Table 82
b2-b4	FEC Type	3 bit field that selects the FEC type 000 = no FEC 001 = 1/2 rate Convolutional 010 = 2/3 rate Convolutional 011 = 3/4 rate Convolutional 100 = Reed-Solomon(223, 255) 101 = Concatenated R-S, 1/2 rate 110 = Concatenated R-S, 2/3 rate 111 = Concatenated R-S, 3/4 rate
b5-b6	M-BOK	2 bit field that selects the M-BOK type 00 = 2-BOK 01 = 4-BOK 10 = 8-BOK 11 = 16-BOK
b7	PSK	1 bit field that selects the PSK type 0 = BPSK 1 = QPSK
b8-b9	Interleaver Type	2 bit field that selects the interleaver type 00 = no interleaver 01 = bit interleaver 10 = byte interleaver
b10-b23	Frame Body Length	An 14 bit field that contains the length of the frame body, in octets, MSB is b5, LSB is b15, e.g. 4 octets of data, is encoded as 0b00000000100. A zero length frame body is encoded as 0b000000000000 and there is no FCS for this packet.

4.6 MAC Header

TBD

4.7 HCS

The header check sequence is calculated on the combined PHY and MAC Headers. The header check sequence is appended after the MAC header and contains the 16 bit CRC for the combined PHY and MAC headers. The polynomial used is:

$x^{16} + x^{12} + x^{5} + 1$

This CRC is the same one used in IEEE Std 802.11b-1999.

5. Ranging Support

6. Comments on Receiver Performance

6.1 Maximum Input Level

The receiver maximum input level is the maximum power level of the incoming signal, in dBm, present at the input of the receiver for which the error rate criterion met. A compliant receiver shall have a receiver maximum input level of at least -20 dBm.

6.2 Coexistence

6.2.1 IEEE802.15.1

Our proposed UWB waveform does not intentionally emit power into the 2.4 GHz ISM band.

6.2.2 IEEE802.11b

Our proposed UWB waveform does not intentionally emit power into the 2.4 GHz ISM band.

6.2.3 IEEE802.15.3

Our proposed UWB waveform does not intentionally emit power into the 2.4 GHz ISM band.

6.2.4 IEEE802.11a

Our proposed UWB waveform does not intentionally emit power into the 5 GHz UNII band.

6.2.5 IEEE802.15.4

Our proposed UWB waveform does not intentionally emit power into neither the 2.4 GHz ISM band nor the 900 MHz ISM band.

6.3 Receiver Jamming Resistance

A general comment on receiver jamming resistance is the fact that the RX RF front end needs to have enough dynamic range to handle any signal overload condition until the detection bandwidth is set by the cascaded signal filtering, whether this be band pass filtering or low pass filtering.

6.3.1 Microwave Ovens

Our proposal does not require a receiver that is responsive to energy in the 2.4 GHz ISM band.

6.3.2 IEEE802.15.1

Our proposal does not require a receiver that is responsive to energy in the 2.4 GHz ISM band.

6.3.3 IEEE802.11b

Our proposal does not require a receiver that is responsive to energy in the 2.4 GHz ISM band.

6.3.4 IEEE802.15.3

Our proposal does not require a receiver that is responsive to energy in the 2.4 GHz ISM band.

6.3.5 IEEE802.11a

Our proposal does not require a receiver that is responsive to energy in the 5 GHz UNII band.

6.3.6 IEEE802.15.4

Our proposal does not require a receiver that is responsive to energy in neither the 2.4 GHz ISM band nor the 900 MHz ISM band.

6.3.7 In-Band Modulated and Unmodulated Interference

6.3.8 Out-of-Band Modulated and Unmodulated Interference

Our proposal does not require a receiver that is responsive to energy outside the specified low UWB band and the high UWB band. Receiver filtering can be of the root raised cosine type exhibiting very sharp cutoff frequency response while not causing inter-chip interference problems. The receiver front-end needs to be designed against anticipated out-of-band overload conditions, of which the exact specification is out-of-scope of the standard.

6.4 Link Quality Indication

The link quality indication (LQI) shall be reported using an SNR estimation. The SNR shall be measured at the decision point in the receiver. The SNR includes the thermal noise, distortion, uncorrected interference and other signal impairments at the decision point in the receiver. The receiver shall report the SNR as a 5 bit number that covers a range of TBD dB to TBD dB of SNR. The value 0x00000 shall correspond to less than or equal to TBD (lower limit) dB SNR and 0x11111 shall correspond to more than or equal to TBD dB (upper limit) SNR with equal steps in between.