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Re:	[]
Abstract	[Proposition of a high rate wireless system in the 60 GHz range, providing data rates ranging from 335 Mbps to 3 Gbps.]
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Notice	
Release	

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Acronyms

ACF	Autocorrelation function
CFO	Carrier frequency offset
FFT	Fast Fourier Transform
IFFT	Inverse Fast Fourier Transform
OFDM	Orthogonal Frequency Division Multiplexing
PRBS	Pseudo-Random Binary Sequence Generator
SF	Signal Field

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1. Introduction

This document presents a complete 60 GHz PHY Layer Proposal fulfilling the requirements of the IEEE 802.15 Task Group 3c. The proposed data rates range from 335 Mbps to 3 Gbps, with a high level of scalability, in terms of available data rates, channelization, PHY parameters and complexity. Up to nine channels of 1 GHz each are defined in the 57 – 66 GHz frequency band, with a possibility to use different channel sizes of 500 MHz and 2 GHz. The proposed system is based on the Orthogonal Frequency Division Multiplexing (OFDM) technology, providing a high spectrum efficiency. For the nominal channel bandwidth of 1 GHz, the system uses 1024 subcarriers, with 840 data sub-carriers per OFDM symbol. Convolutional coding is proposed with code rates ranging from 1/2 to 5/6, with a specific scheme of parallel coding for higher data rates. The decoding performance is increased using an optimized interleaving scheme, designed for both binary and tone interleaving. This proposal facilitates low power and cost-effective implementations.

2. PHY Layer Description

2.1. Frame Format

2.1.1. Basic Frame Structure

The basic frame format is similar to the 802.11a and g OFDM standards. The most important difference is that in our proposal for 60 GHz up to 20 MAC frames can be concatenated within one PHY train.

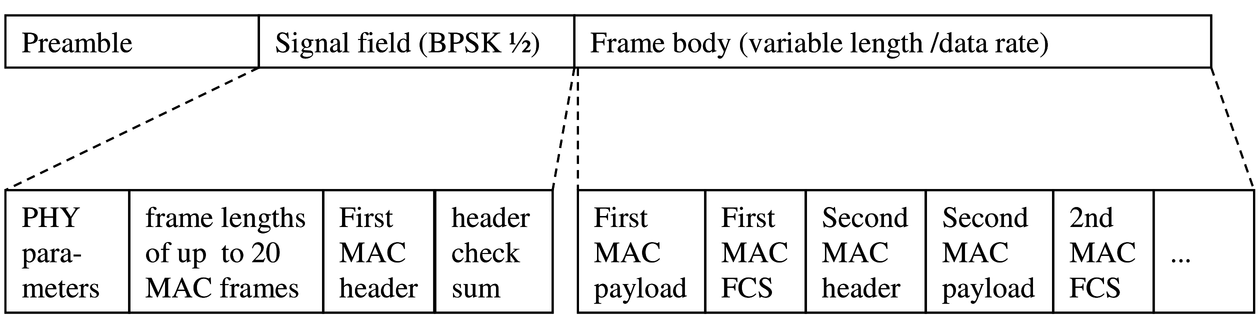

Figure 2-1. Basic Frame Structure

First, a preamble of about 6.6 µs duration is transmitted. The first OFDM symbol contains the PHY header, the sizes of all MAC frames and the first MAC header. It is always sent at the lowest data rate of 375 Mbit/s (BPSK ½). Then, a variable number of OFDM symbols follow containing the remaining MAC headers and all MAC payloads. They may be sent at a different data rate (modulation scheme), which shall be the same for all OFDM symbols, and which is indicated in the PHY header. If required, the last OFDM symbol is filled with arbitrary stuff bytes.

The PHY allows concatenating up to 20 MAC frames of up to 4078 bytes each into one PHY train, allowing for up to 80 kbyte MAC payload in one PHY frame. Each MAC frame has it's own MAC header and FCS. The first MAC header is part of the Signal Field (SF), the others are in the frame body.

The MAC header is fully 802.15.3 compliant (12 bytes including HCS). It is protected by the standard 802.15.3 header check sum (HCS, 16 bit CRC). The MAC payload is of variable length. It is protected by the 802.15.3 frame check sequence (FCS, 32 bit CRC). After each FCS, there is 1 byte space for Viterbi tail bits.

The length of the inter-frame spacing IFS between consecutive frames complies with the rules of the 802.15.3 standard (MIFS or SIFS time, depending on transmitter and receiver station).

2.1.2. Signal Field

The signal field (first OFDM symbol @ 375 Mbit/s = 52.5 byte) contains the following data:

Assigned bits	Length	logical name	Description and range
Byte 0, bits 0-2	3 bits	PHY_VERSION	PHY version (currently set to zero)
Byte 0, bits 3-7	5 bits	PD_MODE	Transmission mode for data payload (see table 2)
Byte 1, bits 0-2	3 bits	N_PERM	N_PERM+1 specifies the interleaver size in multiples of OFDM symbols, range = 1…8
Byte 1, bits 3-5	3 bits	N_STREAM	N_STREAM+1 = number of parallel coding streams
Byte 1, bit 6	1 bit	FRM_FOLLOW	1 = after the MIFS time, a subsequent frame is intended to be sent with the short preamble 0 = no subsequent frame is intended to be sent
Byte 1, bit 7	1 bit	-	Reserved (0)
Byte 2, bits 0-4	5 bits	N_MAC_FRM	N_MAC_FRM+1 = number of transmitted MAC frames in this physical frame, range = 1…20
Byte 2, bits 5-7	3 bits	-	Reserved (000)
Byte 3, bits 0-7 Byte 4, bits 0-6	15 bits	PD_SCR_INIT	Initial state for data scrambler
Byte 4, bit 7	1 bit	-	Reserved (0)
Byte 5, bits 0-7 Byte 6, bits 0-3	12 bits	NDATA_1	NDATA_1 = packet length in MAC frame 1
Byte 6, bits 4-7 Byte 7, bits 0-7	12 bits	NDATA_2	NDATA_2 = packet length in MAC frame 2
Byte 8, bits 0-7 Byte 9, bits 0-3	12 bits	NDATA_3	NDATA_3 = packet length in MAC frame 3
Byte 9, bits 4-7 Byte 10, bits 0-7	12 bits	NDATA_4	NDATA_4 = packet length in MAC frame 4

Byte 11, bits 0-7 Byte 12, bits 0-3	12 bits	NDATA_5	NDATA_5 = packet length in MAC frame 5
Byte 12, bits 4-7 Byte 13, bits 0-7	12 bits	NDATA_6	NDATA_6 = packet length in MAC frame 6
Byte 14, bits 0-7 Byte 15, bits 0-3	12 bits	NDATA_7	NDATA_7 = packet length in MAC frame 7
Byte 15, bits 4-7 Byte 16, bits 0-7	12 bits	NDATA_8	NDATA_8 = packet length in MAC frame 8
Byte 17, bits 0-7 Byte 18, bits 0-3	12 bits	NDATA_9	NDATA_9 = packet length in MAC frame 9
Byte 18, bits 4-7 Byte 19, bits 0-7	12 bits	NDATA_10	NDATA_10 = packet length in MAC frame 10
Byte 20, bits 0-7 Byte 21, bits 0-3	12 bits	NDATA_11	NDATA_11 = packet length in MAC frame 11
Byte 21, bits 4-7 Byte 22, bits 0-7	12 bits	NDATA_12	NDATA_12 = packet length in MAC frame 12
Byte 23, bits 0-7 Byte 24, bits 0-3	12 bits	NDATA_13	NDATA_13 = packet length in MAC frame 13
Byte 24, bits 4-7 Byte 25, bits 0-7	12 bits	NDATA_14	NDATA_14 = packet length in MAC frame 14
Byte 26, bits 0-7 Byte 27, bits 0-3	12 bits	NDATA_15	NDATA_15 = packet length in MAC frame 15
Byte 27, bits 4-7 Byte 28, bits 0-7	12 bits	NDATA_16	NDATA_16 = packet length in MAC frame 16
Byte 29, bits 0-7 Byte 30, bits 0-3	12 bits	NDATA_17	NDATA_17 = packet length in MAC frame 17
Byte 30, bits 4-7 Byte 31, bits 0-7	12 bits	NDATA_18	NDATA_18 = packet length in MAC frame 18
Byte 32, bits 0-7 Byte 33, bits 0-3	12 bits	NDATA_19	NDATA_19 = packet length in MAC frame 19
Byte 33, bits 4-7 Byte 34, bits 0-7	12 bits	NDATA_20	NDATA_20 = packet length in MAC frame 20
Byte 35-Byte 44	10 bytes	MHD_1	MAC header of packet 1
Byte 45-Byte 48	4 bytes	-	Reserved(0)
Byte 49, bits 0-7 Byte 50, bits 0-7	16 bits	CRC_HD	PHY header checksum (16-bit CRC)
Byte 51, bit 0-5	6 bits	-	Reserved (000000)
Byte 51, bits 6-7 Half byte 52, bits 0-3	6 bits		6 Viterbi tail bits

Table 2-1. Signal field specification

All multi-byte data are little endian.

The packet length values of the MAC frame as given in the signal field shall comprise the MAC header size (10 byte, except frame 1), the HCS (2 byte, except frame 1), the MAC payload size

(variable), Viterbi tail bits (1 byte) and, if applicable, the FCS size (4 byte). A value of 0 indicates that the respective entry is not used. The header and HCS of the first MAC frame is included in the signal field.
The available modulation schemes are specified in section 2.3.3. Scrambler details are provided in section 2.3.4.

2.2. Preamble Format and Frame Acquisition Scheme

2.2.1. Preamble format

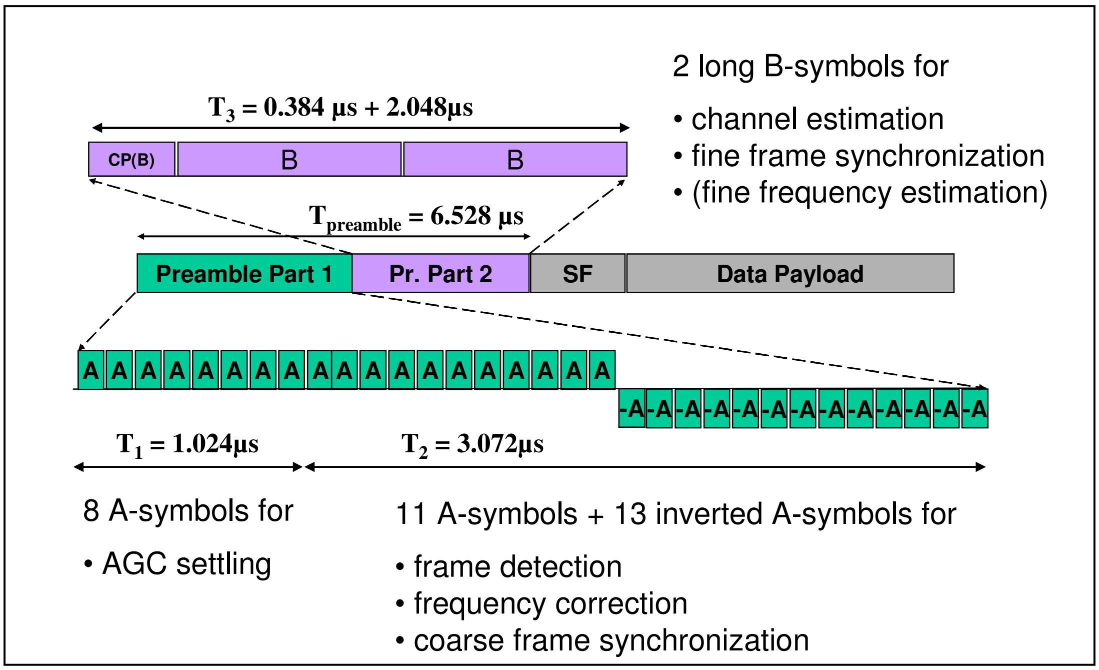

Figure 2-2. Preamble format and acquisition overview

The preamble structure is depicted in Figure 2-2 and consists of two different parts. The first part is composed of 8 short A-sequences, reserved for AGC settling, and another 11 A-sequences and 13 inverted A-sequences for frame detection, frequency correction and coarse time synchronization. Each A-sequence has a duration of $T_{FFT}/8$ = 128 ns. The second preamble part is composed of two long B-sequences, each with a duration of T_{FFT} = 1024ns, together with a extended cyclic prefix of 384 ns. It is used for channel estimation and fine time synchronization. The A-sequence and B-sequence are defined below. The A-sequence is generated by inverse FFT operation with a FFT of size N=128, whereas the B_sequence is generated with a FFT of size N=1024. In practice, both sequences of the full preamble can be stored as a fixed waveform in non volatile memory.

The A-sequence in time domain is defined as

$$z_A(t) = \sqrt{\frac{1}{112}} \sum_{k=-64}^{63} \frac{S_k^A}{\sqrt{2}} \exp(j2\pi \Delta f_{128} kt) \qquad \Delta f_{128} = \frac{1}{128} \cdot \frac{1}{ns}$$

The symbols S_k^A are rotated BPSK symbols taken from the alphabet [0;a;b] with a = (-1-j), b = (1+j).
The full vector $\left[S_k^A\right]_{k=-64\ldots63}$ is given by

$\left[S_k^A\right]_{k=-64\ldots63}$ =

[00000000bbaaabbbbbaababbabababababababbaaaaaaaaabbbbbbaabbbbbba0abaababaabbaabab
babbbaabaaaabaabaabbbaaaaaabaaabbaababba0000000]

The B-sequence in time domain is defined as

$$z_B(t) = \sqrt{\frac{1}{906}} \sum_{k=-512}^{511} \frac{S_k^B}{\sqrt{2}} \exp(j2\pi k \Delta f_{1024} t) \qquad \Delta f_{1024} = \frac{1}{1024} \cdot \frac{1}{ns}$$

The symbols S_k^B are BPSK symbols taken from the alphabet [0;a;b] with a = -1, b = +1.
The full vector $\left[S_k^B\right]_{k=-512\ldots511}$ is given by

$\left[S_k^B\right]_{k=-512\ldots511}$ =

[000000000000000000000000000000000000000000000000bbabaabbbbaabbaaabba
abaabbabbabaabbabaaaababababbbaababaabaaaabaaabbaabbaababbabbbabbabaababaaabbabaabaa
baababbbbbaabbbbbbaabaabbabbbabaaaaaabbaabaabbababaaaaababbbbababbaaababaaabababbaa
bbababbabaaaaabbabaaaabaaaaabbbabaaabaababbbaaaabbbababbababaabaaabbbabaababbaaabab
bbbbaaaabbabbbaabbbbaabaaaabbbababababbbbaaabbbaaaaaaaababbabbaabaabbababaabbabaabaaa
aaababbaababaaabbbaababbbbbbaabaabbaabbaababaabbbbbbabbaaaaaabbbababbbabbbbbbabab
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aaaabbaaaaabbbbaaabbababbaaababaaabaaaabbaaaababaababaaabaabbbbbaabababbbbbbabbbaab
baabbaabaaabaaababaabaabbabbaaaabbbbbbaabbbbaababababbbbbabbbbbaba0000000000000000
000000000000000000000000000000000000]

The preamble uses the same windowing transition time T_{TR} as used for the data frame specified in subsection 2.3.14. The preamble signal is defined as

$$r_{preamble}(t) = z_A(t)\cdot w_{T_{A1},T_{TR}}(t) - z_A(t-T_{A1})\cdot w_{T_{A2},T_{TR}}(t) + z_B(t-T_{GB}-T_{A1}-T_{A2})\cdot w_{T_B,T_{TR}}(t-T_{A1}-T_{A2})$$

$$T_{A1} = 19\cdot 128ns \qquad T_{A2} = 13\cdot 128ns \qquad T_{GB} = 3\cdot 128ns \qquad T_B = 3\cdot 128ns + 2\cdot 1024ns$$

The windowing function is defined in subsection 2.3.14. The A-sequence has a peak-to-power ratio of 3.6 dB, the B-sequence a peak-to-power ratio of 5.1 dB. It is possible to boost the A-sequence in order to increase acquisition performance.

2.2.2. Frame acquisition scheme

A normalized autocorrelation is related to a delayed version to exploit the sign flip of the A-sequence. All samples satisfying an "antiphase-condition" are marked. These marked samples are grouped in clusters, such that the distance of adjacent cluster samples is below some value d. The middle point in each cluster is defined as a peak at position x_k.Two peaks must be found in the frame with a distance of $|x_{k+1} - x_k| \in [D_{\min}, D_{\max}]$.

With the first peak as a time reference, a second ACF is evaluated for final frame detection and frequency offset correction. This concept is depicted in Figure 2-3.

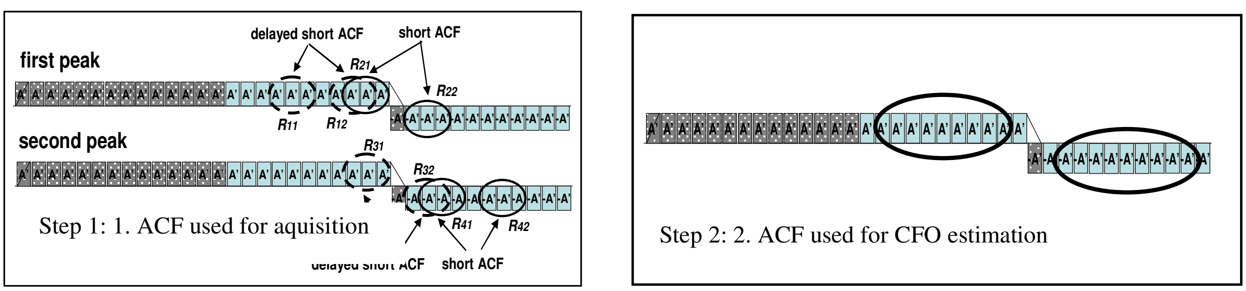


Figure 2-3. Frame acquisition scheme

Two FFTs are applied on the second preamble part for initial channel estimation. A phase-unwrapping method is used to estimate the position of the centroid of the channel impulse response in the frequency domain. The frame FFT start position is taken at a fixed offset position from the estimated centroid. The BPSK modulated OFDM symbol for the signal field is exploited after SF decoding to improve channel estimation (decision feedback estimation). This is depicted in Figure 2-4.

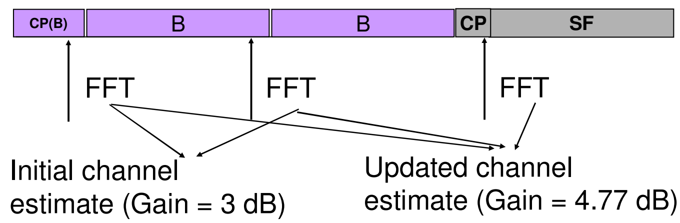


Figure 2-4. Illustration of the channel estimation technique

2.3. PHY Baseband Description

2.3.1. System overview

The 60 GHz system proposed by France Telecom R&D and IHP uses multi-carrier transmissions based on an OFDM scheme. The OFDM modulation corresponds to a parallel transmission of data symbols simultaneously modulated by *N* sub-carriers (Figure 2-5). The sinusoidal waveform of each sub-carrier ensures orthogonality between separate sub-carriers. It provides a high robustness to the effects of multipath channel propagation. The OFDM modulation is simply performed by an inverse FFT transform. In reception, the dual operation is an FFT.

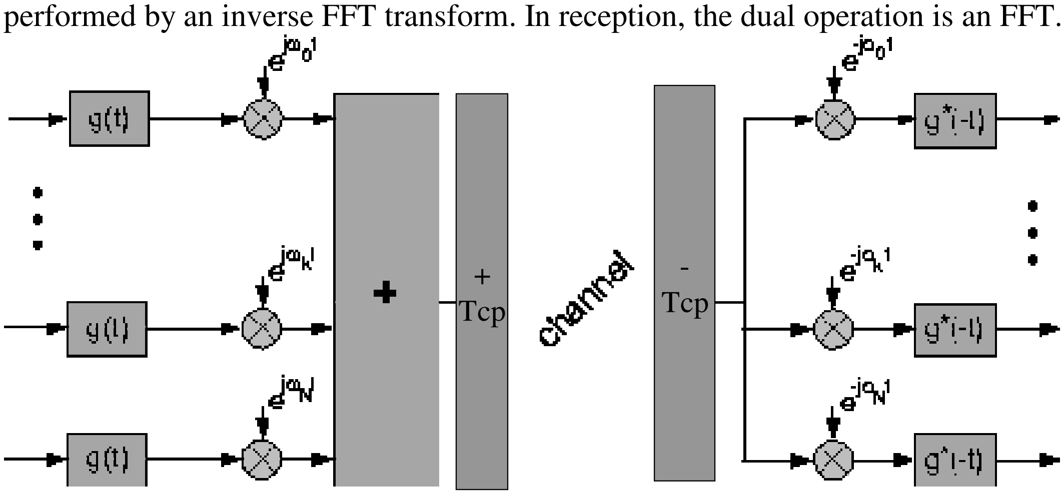


Figure 2-5. OFDM principle

Figure 2-6 presents an overview of the PHY layer transmission chain. The system parameters, as well as the different functions such as data scrambling, data encoding, puncturing, interleaving, constellation mapping and cyclic prefix insertion are detailed in the following paragraphs.

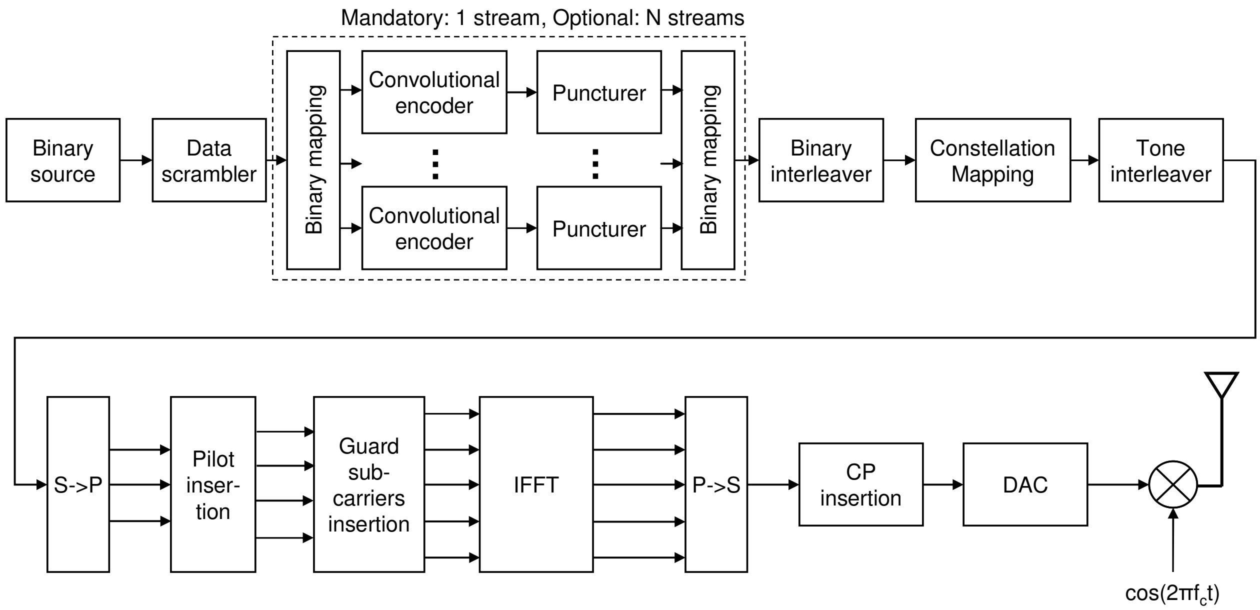


Figure 2-6. 60 GHz system overview

2.3.2. System parameters

As will be detailed further in section 3.1. , the mandatory operation mode uses a channel bandwidth of 1 GHz. Optional operation modes are defined using channel bandwidths of 500 MHz and 2 GHz respectively. Depending on the channel characteristics, three different values of the guard interval T_{CP} (or cyclic prefix, CP) may be used: 96 ns, 160 ns or 220 ns. A list of the system parameters associated with the OFDM PHY is listed in Table 2-3 for the mandatory operation mode using 1 GHz bandwidth. The detailed parameters for the operation modes using bandwidths of 500 MHz and 2 GHz are TBD.

The used notation is detailed in the next table.

Table 2-2. Used notation

Symbol	**Explanation**
R	Code rate
N_{BPSC}	Number of coded bits per sub-carrier
N_{SD}	Number of data sub-carriers
N_{SP}	Number of Pilot sub-carriers
N_{CBPS}	Number of coded bits per OFDM symbol
N_{DBPS}	Number of data bits per OFDM symbol
N_{ES}	Number of FEC encoders

Table 2-3. System parameters (1 GHz bandwidth mode)

Parameter	Value
Number of data subcarriers (N_{SD})	840
Number of pilot carriers (N_{SP})	66
Number of DC zero carriers (N_{DC})	5
Number of guard carriers	113
FFT size (N_{FFT})	1024
Channel bandwidth (B_{FFT})	1 GHz
Subcarrier frequency spacing	0.977 MHz
IFFT/FFT period	1024 ns
Cyclic prefix duration (T_{CP})	96 ns / 160 ns / 220 ns
Symbol interval (T_{SYM})	1120 ns / 1184 ns / 1244 ns

2.3.3. Modulation and Coding Schemes

The rate dependent modulation parameters are listed in Table 2-4 for the short CP (96 ns) and in Table 2-5 for longer CP (160 ns and 220 ns).

Table 2-4. Rate-dependent parameters for short CP (T_{CP} =96 ns) and B_{FFT} = 1 GHz

Data Rate (Mbps)	Modulation	Coding Rate (R)	Coded bits per OFDM symbol	Data bits per OFDM symbol	Data bytes per OFDM symbol
375 Mbps	BPSK	1/2	840	420	52.5
500 Mbps	BPSK	2/3	840	560	70
750 Mbps	QPSK	1/2	1680	840	105
1000 Mbps	QPSK	2/3	1680	1120	140
1500 Mbps	16-QAM	1/2	3360	1680	210
2000 Mbps	16-QAM	2/3	3360	2240	280
2500 Mbps	16-QAM	5/6	3360	2800	350
3000 Mbps	64-QAM	2/3	5040	3360	420

Table 2-5. Rate-dependent parameters for longer CP (160 ns or 220 ns) and B_{FFT} = 1 GHz

Minimum Data Rate (Mbps)	Modulation	Coding Rate (R)	Coded bits per OFDM symbol	Data bits per OFDM symbol	Data bytes per OFDM symbol
335 Mbps	BPSK	1/2	840	420	52.5
500 Mbps	BPSK	3/4	840	630	78.75
675 Mbps	QPSK	1/2	1680	840	105
1000 Mbps	QPSK	3/4	1680	1260	157.5
1350 Mbps	16-QAM	1/2	3360	1680	210
2000 Mbps	16-QAM	3/4	3360	2520	315
2700 Mbps	64-QAM	2/3	5040	3360	420
3000 Mbps	64-QAM	3/4	5040	3780	472.5

2.3.4. Data scrambler

A side-stream data scrambler shall be used to whiten the aggregated data frame, i.e. all of the MAC frames consisting of the MAC header, the HCS and the MAC payload, but excluding the eight zero tail bits at the end of each MAC frame. The scrambler shall be initialized to a random seed value specified in the signal field (SF). The scrambling process starts with the first bit in the aggregated data frame following the SF. Virtually, the scrambling is done in two steps. At first, the complete aggregated data frame including pad bits is subject to scrambling without re-initialization at any position. Then the scrambled eight tail bits after each MAC frame are replaced with zeros. The scrambling process is described below.

The polynomial generator, $g(D)$, for the pseudo-random binary sequence generator (PRBS) shall be : $g(D) = 1 + D^{14} + D^{15}$, where D is a single bit delay element. Using this generator polynomial, the corresponding PRBS, $x[n]$, is generated as

$$x[n] = x[n-14] \oplus x[n-15]$$

where "⊕" denotes modulo-2 addition. The following sequence defines the initialization vector, x_{init}, which is specified by 15 bits in the signal field.

$$x_{init} = \left[x_i[-1]x_i[-2]...x_i[-14]x_i[-15]\right],$$

where $x_i[-k]$ represents the binary initial value at the output of the k^{th} delay element.

The scrambled data bits, $v[m]$, are obtained as follows:

$$v[m] = s[m] \oplus x[m],$$

where $s[m]$ 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 PHY header.

Scrambling is applied also to the signal field data except for the header checksum and tail bits. This is done in order to avoid possible long sequences of zeros within the packet length table and for reserved bits. The same scrambler polynomial is used, but the scrambler is initialized to a fixed state for every transmission, which is the all-ones-state as shown below:

$$x_{init,SF} = \left[x_i[-1]x_i[-2]\ldots x_i[-14]x_i[-15]\right] = [1111\ldots111]$$

The PHY header checksum shall be performed over the scrambled SF data.

2.3.5. Tail bits

The tail bit fields are required to return the convolutional encoder to the "zero state". This procedure improves the error probability of the convolutional decoder, which relies on the future bits when decoding the message stream. Six tail bits are defined and sufficient for the SF, whereas eight tail bits shall be added after each MAC FCS to ease implementation. The tail bits are not counted in the length of each MAC frame. Using eight tail bits, the MAC frame length in source bits with added tail bits is always a multiple of octets (bytes).

2.3.6. Convolutional Encoder

The scrambled data are encoded using a convolutional code with a constraint length set to K=7. The convolutional encoder shall use the rate R = 1/2 polynomials, $g_0 = 133_8$ and $g_1 = 171_8$ as presented in Figure 2-7.

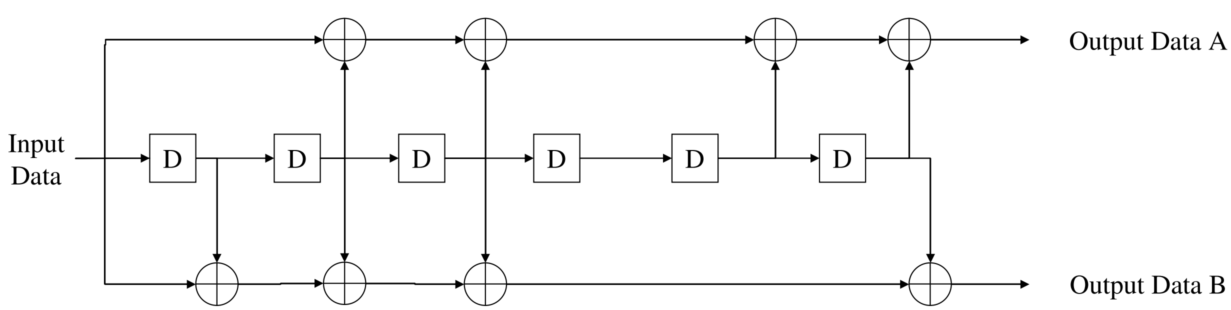


Figure 2-7. Convolutional encoder: rate R = 1/2, constraint length K = 7

The bit denoted as "A" shall be the first bit generated by the encoder, followed by the bit denoted as "B".

Decoding by the Viterbi algorithm is recommended.

Optionally, in order to reduce the implementation constraints, data encoding may be organised using several scrambled bit streams in parallel. In this mode of operation, the scrambled bits are first separated in N_{ES} streams using a serial to parallel conversion. Each parallel bit stream is then encoded using the convolutional encoder described above, and the different coded bit streams are parallel to serial converted. This mode proposes two streams (N_{ES}=2) in the case of 1GHz bandwidth and four streams (N_{ES}=4) in the case of 2 GHz bandwidth size. The operation mode is depicted in Figure 2-8 with N_{ES} = 4 parallel streams. The main advantage of this mode of operation is to reduce the required processing rate of each parallel coder / decoder.

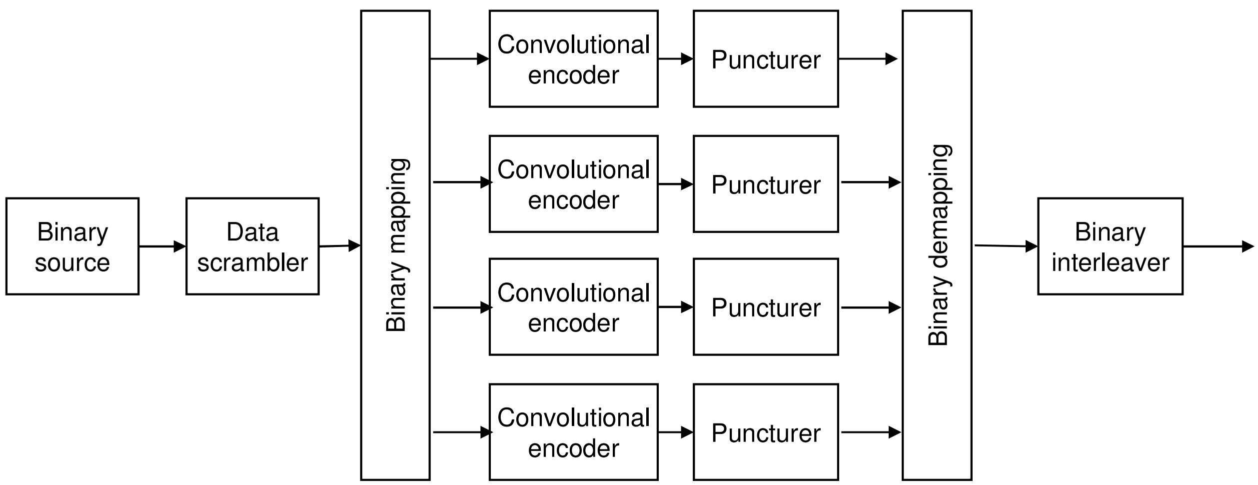


Figure 2-8. Parallel bit encoding mode, with *N* = 4 streams

2.3.7. Puncturer

The various coding rates are derived from the rate R = 1/2 convolutional code 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. The puncturing matrices for the mother code rates R = 1/2 and different output code rates is given in Table 2-6. Four punctured code rates {1/2, 2/3, 3/4 and 5/6} are proposed to provide a high net bit rate granularity.

Table 2-6. Puncturing patterns

Code rate	Puncturing pattern
1/2	N/A
2/3	$\begin{pmatrix} 11 \\ 10 \end{pmatrix}$
3/4	$\begin{pmatrix} 011 \\ 110 \end{pmatrix}$
5/6	$\begin{pmatrix} 11010 \\ 10101 \end{pmatrix}$

The frame header shall be encoded with a rate R = 1/2. The encoder shall be reset to the all-zero state after this. Next, the frame payload shall be coded with a rate R = 1/2, 2/3, 3/4 or 5/6, corresponding to the desired data rate.

2.3.8. Bit interleaving

The coded bit stream is interleaved prior to modulation. Bit interleaving provides robustness against burst errors. The interleaving is performed upon encoded bits included within an interleaving depth covering either 1 or 4 OFDM symbols.
The FEC- interleaving mode 1, considers the case where the interleaving processing is performed after the FEC structure as illustrated on the Figure 2-8. The second FEC interleaving mode, considered as an optional mode, implements interleaving processing upon each branch after the the puncturing as illustrated on Figure 2-9.

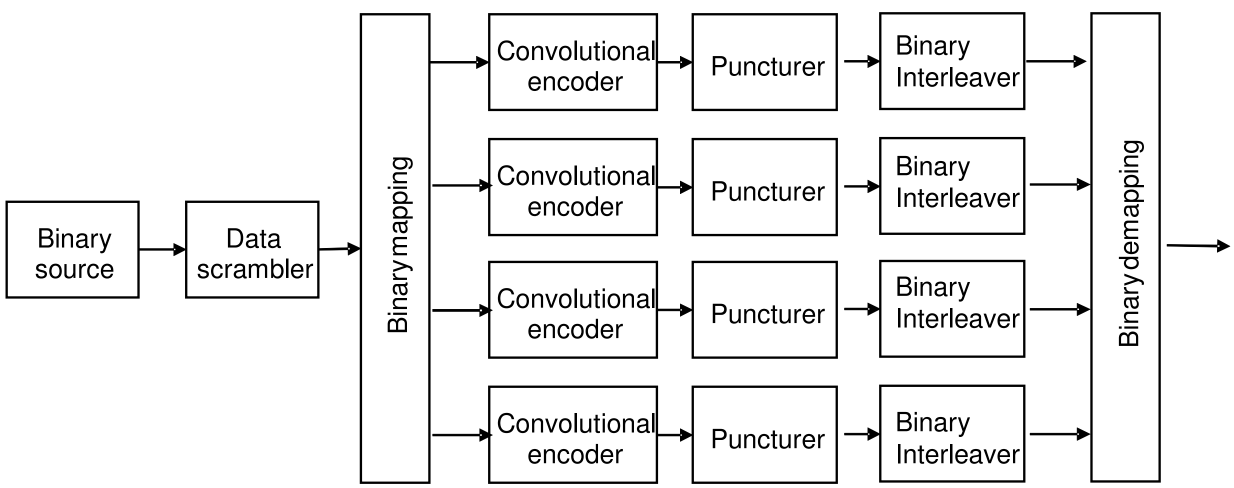


Figure 2-9: FEC interleaving mode 2

The FEC interleaving mode 2 provides additional equivalent interleaving depths *Keq* given by the following equation. In this case, different permutation rules may be considered on each branch. This configuration is equivalent to a single stream interleaving processing with a depth equal to:

$$Keq = \sum_{m=1}^{N_{ES}} K_i^m$$

K_i^m *:binary interleaving size for the m-thbranch*

$$K_i = i \cdot K, i = \{1,4\}$$

$$K = N_{BPSC} \cdot N_{SD}$$

Depending on the constellation index, the corresponding binary interleaver sizes are given in Table 2-7.

Table 2-7. Binary interleaver size with 1 GHz band

	BPSK		QPSK		16-QAM		64-QAM	
Number of OFDM symbols per interleaving frame	1	4	1	4	1	4	1	4
Binary interleaver size *Ki*	840	3360	1680	6720	3360	13440	5040	20160

2.3.8.1. Interleaving algorithm

The block interleaving process is performed using a permutation rule $L(k)$. $L(k)$ indicates the order to read samples from the input sequence S and writes it to the output sequence S' with a block size K_i corresponding to the interleaving block size. Sample index k at both output and input of the interleaving is ranged from 0 to K_i-1. That is, the k-th output, written to location k in the output vector, is read from location $L(k)$ in the input vector.

The block interleaving algorithm $L(k) = I^{(j)}{}_{p,q}(k)$ is described by four parameters: the block size K, an integer parameter p setting the partition size, an integer parameter q and the iteration j governing the interleaving spreading [3]. The relationship between the block of K coded bits $b_{coded}(k)_{\{k=0\ldots K-1\}}$ and the block of K intereleaved bits $b_{interleaved}(k)_{\{k=0\ldots K-1\}}$ is given by:

$$b_{\text{int}\,erleaved}(k) = b_{coded}\left(I_{p,q}^{(j)}(k)\right)$$

To realize the interleaver stage, it is recommended to implement a lookup table, where the interleaving rule is memorized.

The interleaving rule is based on an iterative structure in order to increase the scalability of the interleaver, as described in Figure 2-10.

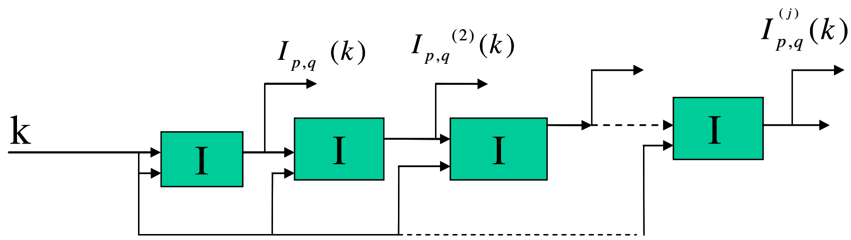


Figure 2-10. The turbo-based interleaving structure

The interleaving rule is described as follows for the 0th and the *j*th iteration :

$$I_{p,q}^{(0)}(k) = \left[K - p + k + q \cdot p \cdot [-k - p \cdot k]_K\right]_K$$
$$I_{p,q}^{(j)}(k) = \left[K - p + k + q \cdot p \cdot \left[-k - p \cdot I_{p,q}^{(j-1)}\right]_K\right]_K$$

The operation $[x]_K$ is the x modulo-K arithmetic operation, i.e. the rest of the division of x by K.

$$[x]_K = x - \left\{\left\lfloor \frac{x}{K} \right\rfloor \cdot K\right\}$$

where $\lfloor \cdot \rfloor$ represents the floor function.

The interleaver parameters are selected in order to optimize the interleaving spreading between successive samples. The interleaving spreading $\Delta L(s)$ is defined as the minimum distance between interleaved bits separated by a distance $s - 1$, and is expressed as follows:

$$\Delta L(s) = \underset{k}{Min}\left\{\left| I_{p,q}^{(j)}(k+s) - I_{p,q}^{(j)}(k) \right|\right\}$$

The interleaving spreading is calculated in an algebraic way and allows the selection of interleaving parameters $\{p,q,j\}$ for each interleaving block size K_i. The optimized values of the parameters K, p, q and j to be used to set up the binary interleaver are given in Table 2-8 [1].

[1] Note that the present interleaver parameters are optimized for the case where one single encoder stream is used. In the optional mode using several encoder streams in parallel, the optimized interleaver parameters are TBD.

Table 2-8. Binary interleaver parameters (for B_{FFT} = 1 GHz)

	BPSK		QPSK		16-QAM		64-QAM	
Number of OFDM symbols per interleaving frame	1	4	1	4	1	4	1	4
Binary interleaver size *K*	840	3360	TBD	TBD	3360	13440	5040	20160
p	12	96	TBD	TBD	96	21	36	288
q	2	2	TBD	TBD	2	2	2	2
j	3	1	TBD	TBD	1	3	3	3

2.3.9. Subcarrier constellation mapping

Bits are mapped over either BPSK, QPSK, 16-QAM or 64-QAM data symbols depending on the desired data rate. The generated complex symbols are then and forwarded to the OFDM multiplex. For this purpose, the encoded and interleaved binary serial input data shall be divided into groups of 1, 2, 4 or 6 bits and converted into complex numbers respectively representing BPSK, QPSK, 16-QAM or 64-QAM constellation points. The conversion shall be performed according to the Gray-coded constellation mappings, illustrated in Figure 2-11 to Figure 2-14, with the input bit, b_0, being the earliest in the stream[2]. The output values, *d*, are formed by multiplying the resulting (I + jQ) value by a normalization factor of K_{MOD}, as described in the following equation:

$$d = (I + jQ) \times K_{MOD}.$$

The normalization factor, K_{MOD}, depends on the base modulation mode, as prescribed in Table 2-9.

[2] These figures are from IEEE 802.11a-1999 standard [2].

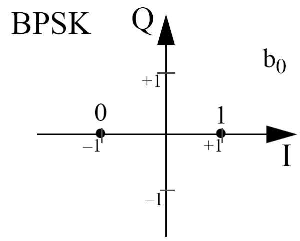


Figure 2-11 – BPSK constellation bit encoding

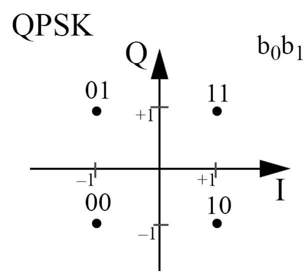


Figure 2-12 – QPSK constellation bit encoding

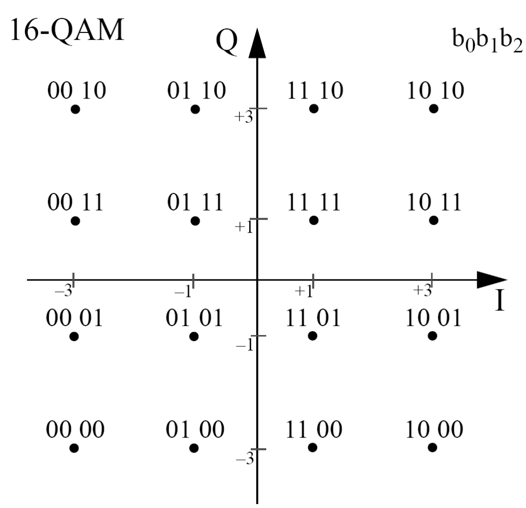


Figure 2-13 – 16-QAM constellation bit encoding

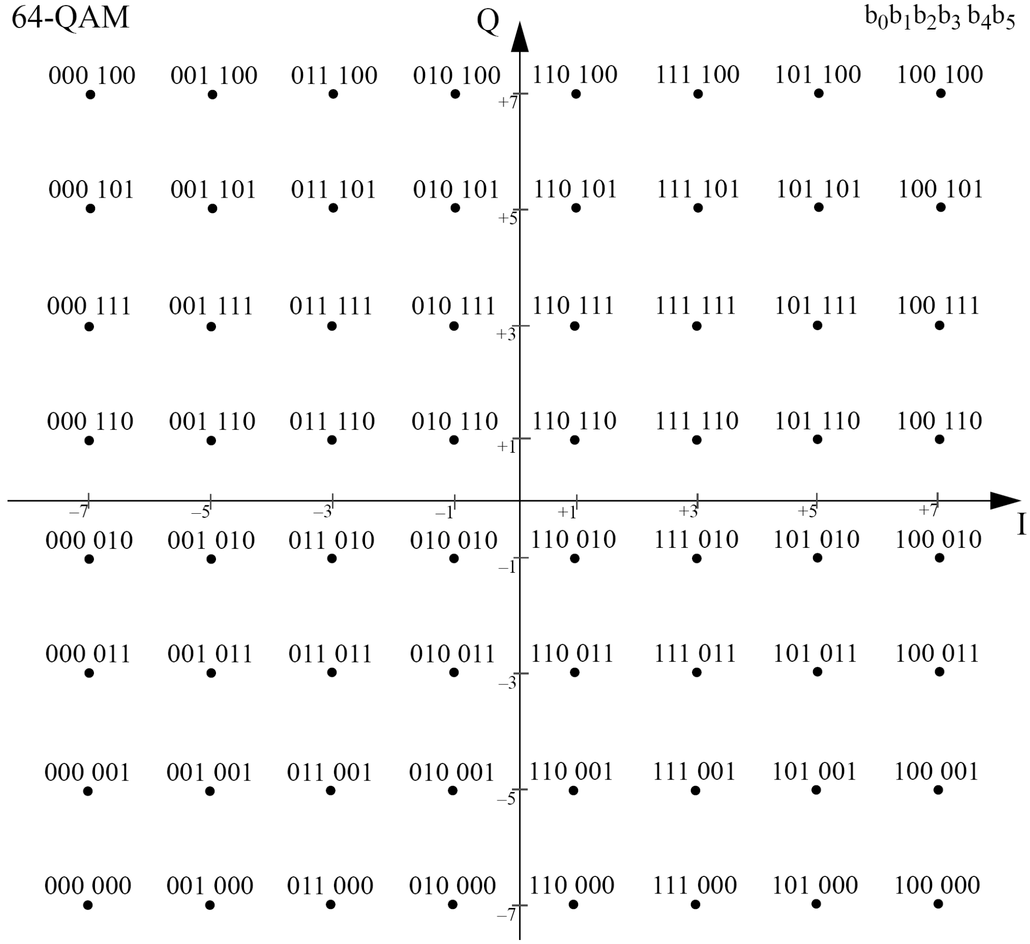


Figure 2-14 – 64-QAM constellation bit encoding

Table 2-9. Modulation-dependent normalization factor K_{MOD}

Modulation	K_{MOD}
BPSK	1
QPSK	1/√2
16-QAM	1/√10
64-QAM	1/√42

2.3.10. Tone interleaving

Optionally, tone interleaving may be used after sub-carrier mapping, in order to increase the system frequency diversity and reduce the co-channel interference.

For operation modes using tone interleaving, the stream of complex numbers is divided into groups of 840 complex numbers. These complex numbers are denoted $c[n, k]$ which corresponds to subcarrier n of OFDM symbol k, as follows:

$$c[n,k] = d[n + 840k], \qquad n = 0,1,...,839, \;\; k = 0,1,..., N_{SYM} - 1$$

where N_{SYM} denotes the number of OFDM symbols in the MAC frame body, tail bits, and pad bits.

The interleaving algorithm is the same as the binary interleaving (cf. Section 2.3.8.) with different sizes and permutation rules. The implementation is carried out in a dynamic way, meaning that two different permutation rules are successively applied to two data sub-carrier blocks associated with two successive OFDM symbols. These two rules shall be described either in a matrix format (lookup table) or thanks to a generic equation and two interleaving parameter subsets.

Let $\{p1, q1, j1\}$ and $\{p2, q2, j2\}$ be the two parameter sets associated with the two different permutation rules used at the tone interleaver. The output $c'[n, k]$ of the sub-carrier interleaving process is defined by:

$$c'[n,k] = c[I_{p1,q1}^{(j1)}(n), k], \qquad n = 0,1,...,839, \;\; k = 0,2,4,...$$

$$c'[n,k] = c[I_{p2,q2}^{(j2)}(n), k], \qquad n = 0,1,...,839, \;\; k = 1,3,5,...$$

The values of the parameters K, $p1$, $q1$, $j1$, $p2$, $q2$, $j2$ to be used to set up the tone interleaver are given in Table 2-10.

Table 2-10. Sub-carrier interleaver parameters

Sub-carrier interleaver size *K*	$N_{SD} = 840$	$N_{SD} = 4 \times 840$
***p*1**	12	96
***q*1**	2	2
***j*1**	3	1
***p*2**	[TBD]	[TBD]
***q*2**	[TBD]	[TBD]
***j*2**	[TBD]	[TBD]

2.3.11. OFDM modulation

An OFDM data symbol $r_{data,k}(t)$ is defined as

$$r_{data,k}(t) = \sum_{n=0}^{N_{SD}-1} c'[n,k]\exp(j2\pi M(n)\Delta_F(t-T_{CP}))$$

where N_{SD} is the number of data subcarriers (N_{SD} = 840), and the function $M(n)$ defines a mapping from the indices 0 to N_{SD}-1 to the logical frequency offset indices $-N_{ST}/2$ to $N_{ST}/2$-1, excluding the locations reserved for the pilot tones, the guard subcarriers and the DC subcarriers (as described below), with N_{ST} = 1024 being the number of total subcarriers:

$$M(n) = \begin{cases} n - 454 + \left\lfloor \dfrac{n}{13} \right\rfloor & 0 \le n \le 419 \\ n - 449 + \left\lfloor \dfrac{n+5}{13} \right\rfloor & 420 \le n \le 839 \end{cases}$$

where $\lfloor \cdot \rfloor$ represents the floor function.

The subcarrier frequency allocation is shown in Figure 2-15. To avoid difficulties in DAC and ADC offsets and carrier feed-through in the RF system, the subcarriers falling at the logical frequency offsets -2, -1, 0, 1 and 2 are not used.

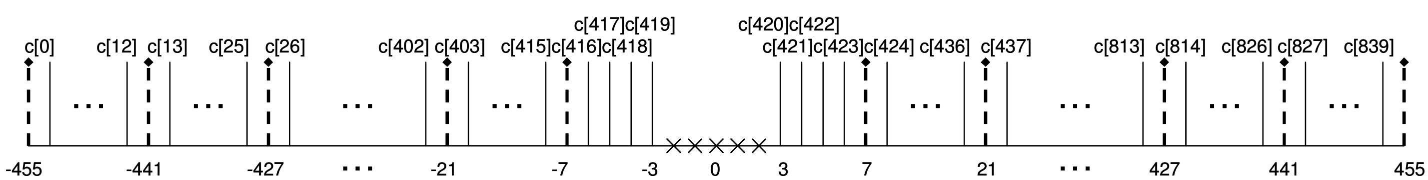


Figure 2-15 – Subcarrier frequency allocation

2.3.12. Pilot sub-carriers

In each OFDM symbol, 66 of the subcarriers are dedicated to pilot symbols in order to make coherent detection robust against frequency offsets and phase noise. These pilot symbols shall be put in subcarriers numbered -455 + 14 j, j = [0…65]. The OFDM signal can be viewed as a superposition of the data signal and the pilot signal. The pilot signal of the k-th OFDM symbol is defined as :

$$r_{pilot,k}(t) = \sum_{n=0}^{65} c_p[n,k] \exp(j2\pi M_p(n) \cdot \Delta_F \cdot (t - T_{CP}))$$

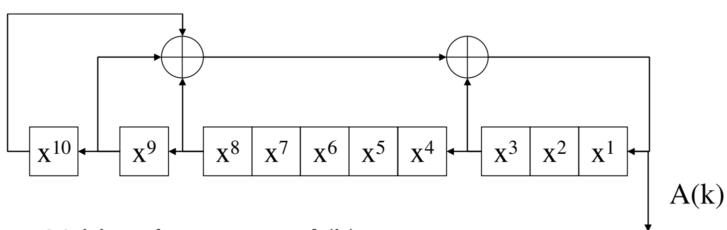
$$M_p(n) = -455 + 14n$$

$$\Delta_F = 1/T_{FFT}$$

The pilot subcarriers are BPSK-modulated and the pilot symbols defined as follows. Let the symbol index k=0 denote the signal field, k=1…N_d stand for the first up to the last OFDM data symbol. A modulation sequence shall be defined as the output of a feedback shift register which acts as a scrambler producing a modulation sequence A(k). The generator polynom, initial state and structure are given below.

Generator Polynom: $P(x) = x^{10} + x^3 + 1$

Initial state of scrambler: 1 1 1 0 0 0 0 1 0 0



First 30 bits of sequence A(k):

0 1 1 0 1 1 0 0 1 0 0 0 1 0 1 0 0 1 1 0 1 1 1 1 0 1 1 1 0 1 ..

The sequence A(k) modulates two different pilot patterns $P_1(n)$ and $P_2(n)$. $P_1(n)$ is used for every even OFDM symbol starting with the SF, whereas $P_2(n)$ is used for every odd symbol. Finally, the pilot symbols are defined as

$c_p[n,k] = (2 \cdot A(k) - 1) \cdot P_1(n)$ for n = 0,2,4,6 …

$c_p[n,k] = (2 \cdot A(k) - 1) \cdot P_2(n)$ for n = 1,3,5,7, …

The two pilot patterns are defined as

$P_1(n)$ = [1;-1;1;-1;-1;-1;1;-1;1;-1;-1;1;1;-1;1;1;-1;1;-1;-1;-1;1;1;1;-1;-1;1;-1;-1;-1;1;1;-1;-1;-1;-1;-1;-1;-1;1;-1;-1;1;1;1;-1;-1;-1;-1;1;-1;-1;1;1;1;1;1;-1;1;1;1;1;-1;1;-1]

$P_2(n)$ = [-1;1;-1;1;1;1;-1;1;-1;1;1;-1;-1;1;-1;-1;1;-1;1;1;1;-1;-1;-1;1;1;-1;1;1;1;-1;-1;1;-1;-1;-1;-1;-1;1;-1;-1;1;1;1;-1;-1;-1;-1;1;-1;-1;1;1;1;1;1;-1;1;1;1;1;-1;1;-1]

It can be noted that the two OFDM symbols used in each frame for channel estimation as well as the 66 pilot tones inserted in each OFDM symbol for phase noise mitigation convey highly sensitive information for an efficient compensation of channel impairments. For this reason, it is optionally recommended to allocate more power to these signals. The power ratio between these channel estimation signals and the data signal should be in the range of 1 dB to 3 dB for an optimum operation.

2.3.13. Guard subcarriers

113 guard subcarriers are used to limit the signal spectrum to nearly 90% of the Nyquist bandwidth for a system operating at 1 GHz sampling rate and using a 1024-FFT. This bandwidth limitation is necessary to facilitate signal filtering either in the digital or analog domain or both. Since 113 guard subcarriers leave a transition band of about 110 MHz, a channel filter with low to moderate complexity is sufficient. A filter with high complexity may significantly reduce the effective guard time of the OFDM system. The principle of using the full amount of subcarriers (complete FFT) together with a transition frequency band in the analog domain cannot be used, because the generated image band through the D/A-conversion process just next to the useful band will not be sufficiently attenuated and will cause interference of the outer subcarriers at the receiver for a carrier frequency offset within a practical range.

2.3.14. Cyclic prefix insertion and time windowing

In the following common definition of the waveform for the k-th OFDM symbol of the data part, a cyclic extension is implicitly given. With a normalization factor N_u, the synthesized OFDM symbol has a power level of P=1 on average.

$$r_{symbol,k}(t) = w_{T,T_{TR}}(t) \cdot \left\lfloor r_{data,k}(t) + r_{pilot,k}(t) \right\rfloor \cdot N_u$$

$$N_u = \sqrt{\frac{1024}{840+66}}$$

The windowing function of pulse duration T and transition time T_{TR} is defined as

$$w_{T,T_{TR}}(t) = \begin{cases} \sin^2\left(\frac{\pi}{2}(0.5 + t/T_{TR})\right) & (-T_{TR}/2 < t < T_{TR}/2) \\ 1 & (T_{TR}/2 \le t < T - T_{TR}/2) \\ \sin^2\left(\frac{\pi}{2}(0.5 - (t-T)/T_{TR})\right) & (T - T_{TR}/2 \le t < T + T_{TR}/2) \end{cases}$$

$$T = T_{FFT} + T_G$$

The pulse duration T is equal to the FFT time plus guard time and yields a cyclic prefix of T_g. Selecting a transition time $T_{TR} > 0$ leads to faster decay of the out-of-band emissions. A combination of time windowing and channel filtering is recommended to meet the requirements of the spectral mask. Within a digital system, the sampled window function can be applied. N_{ov}=3 overlap samples are recommended corresponding to a transition time of 5 nanoseconds. Note that in this case, for each OFDM symbol, the implementer has to extend the prefix by one additional sample and also add a suffix of 2 samples to realize the transitions.

2.3.15. Frame synthesis

The final frame is composed of the preamble signal, the signal field signal and the data frame signal as follows:

$$r_{frame}(t) = r_{preamble}(t) + r_{SF}(t - T_{preamble}) + r_{data}(t - T_{preamble} - T_{SF})$$

$T_{preamble}$ = 6.528 µs
T_{SF} = 1.244µs

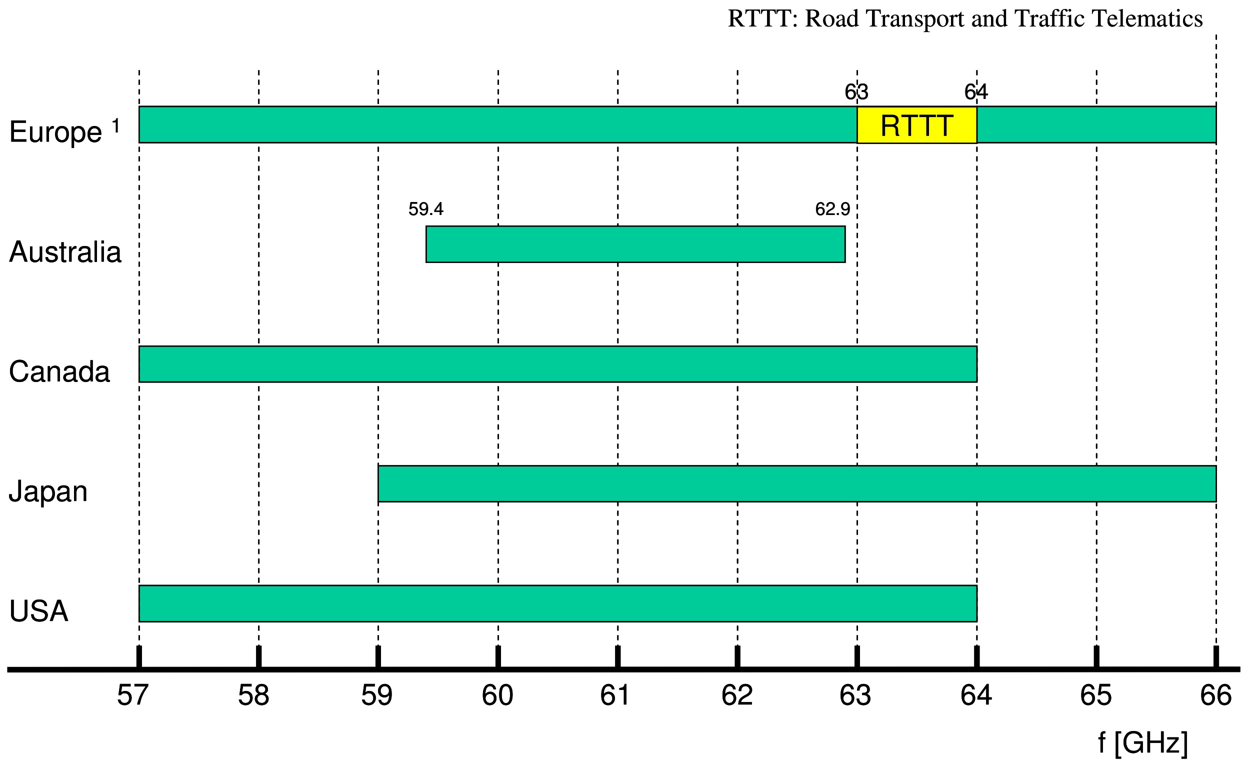
The data part is defined as

$$r_{data}(t) = \sum_{k=0}^{N_{sym}-1} r_{symbol,k}(t - k(T_{FFT} + T_G))$$

3. General requirements

3.1. Operating Frequency Range and Channelization

The proposed frequency band plan provides 9 channels of 1 GHz each, situated in the 57 GHz – 66 GHz frequency range. This allows an efficient use of the international "frequency grid" provided by the different regulation bodies worldwide, as shown in Figure 3-1.



[1] Frequency regulation in Europe is in progress.

Figure 3-1 – 60 GHz regulations in different countries

The band numbering is given in the following table. Optionnally, channels of 500 MHz or 2 GHz could be defined by respectively splitting one 1 GHz channel or aggregating two adjacent 1 GHz channels.

Table 3-1. Band allocation

Band number	Lower frequency	Center frequency	Upper frequency
1	57 GHz	57.5 GHz	58 GHz
2	58 GHz	58.5 GHz	59 GHz
3	59 GHz	59.5 GHz	60 GHz
4	60 GHz	60.5 GHz	61 GHz
5	61 GHz	61.5 GHz	62 GHz
6	62 GHz	62.5 GHz	63 GHz
7	63 GHz	63.5 GHz	64 GHz
8	64 GHz	64.5 GHz	65 GHz
9	65 GHz	65.5 GHz	66 GHz

3.2. Transmitter specifications

3.2.1. Transmit PSD mask

Figure 3-2 presents the transmitted spectral density of the transmitted signal mask. The 0 dBr (dB relative to the maximum spectral density of the signal) bandwidth is of 890 MHz. The 30 dBr bandwidth is 1 GHz, and the spectral density shall fall at – 36 dBr at a bandwidth of 1.11 GHz.

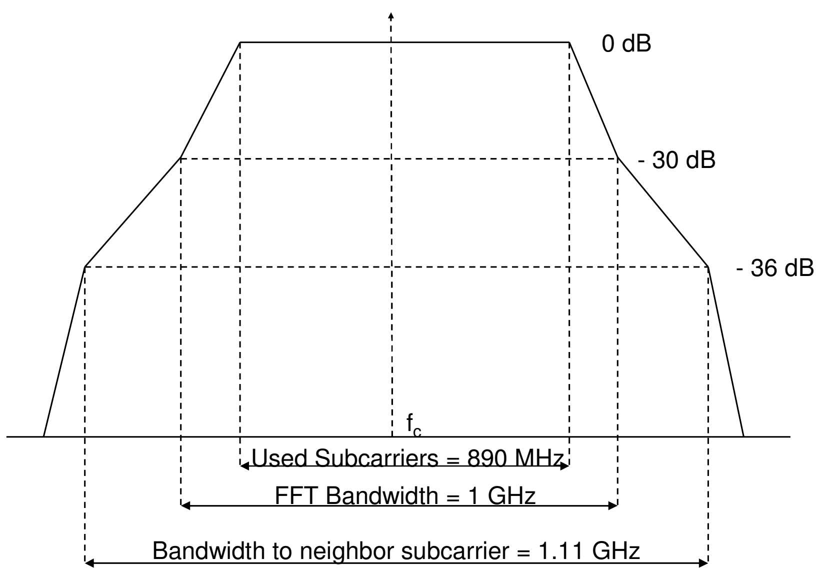


Figure 3-2 – Transmit PSD Mask

3.3. Receiver specifications

3.3.1. Receive sensitivity

For a packet error rate (PER) of less than 8% with a frame size of 2048 bytes, the minimum receiver sensitivity for the different data rates are listed in

Table 3-2. Receiver performance requirements

Data rate for CP = 96 ns	Minimum sensitivity (dBm)
375 Mbps	[TBD]
500 Mbps	[TBD]
750 Mbps	[TBD]
1000 Mbps	[TBD]
1500 Mbps	[TBD]
2000 Mbps	-62.9
2500 Mbps	[TBD]
3000 Mbps	-57.5

4. General Solution Criteria

4.1. Unit Manufacturing Cost/Complexity (UMC)

On the basis of a current FPGA implementation of the OFDM baseband processor, the chip area can be estimated. A 60 GHz frontend demonstrator in 0.25 µm SiGe BiCMOS technology is used for estimating and scaling the chip area of the analog frontend. Furthermore, we us the following assumptions:

- 4 parallel data streams,
- 500 MHz digital CLK,
- 65 nm digital CMOS process for MAC and baseband processor,
- 130 nm analog SiGe-BiCMOS process for analog frontend and data converters:

MAC Processor:	10 mm² (ca. 10 Mio Gates)
Baseband Processor:	15 mm² (ca. 15 Mio Gates)
Data Converters:	10 mm²
Analog Frontend (incl. PA)	6 mm²

Size Complete Transceiver PCB:	5 cm x 4 cm x 3 cm
Patch Array Antenna (3x4 elements):	30 mm x 40 mm x 2 mm (LTCC)
Vivaldi Antenna (10 dBi gain, 30 grd):	30 mm x 30 mm x 0.5 mm (PCB-material)

4.2. Signal Robustness

4.2.1. Interference and Susceptibility

The impact that other co-located radiators may have on the proposed PHY should be completed for the next TG3c meetings.

4.2.2. Coexistence

The distance at which the interfering power generated by the proposed system is at the minimum sensitivity level of different victim systems should be completed for the next TG3c meetings.

4.3. Technical Feasibility

4.3.1. Manufacturability

For verification of the manufacturability, a 60 GHz demonstrator was developed at IHP. This demonstrator comprises the analog frontend, the baseband processor and the MAC processor. With respect to this proposal, the demonstrator operates at approximately half of the signal bandwidth, i.e. 500 MHz channel bandwidth and 400 MHz FFT bandwidth. The maximum data rate achieved so far is 720 Mbit/s. The analog frontend of this demonstrator is shown in Figure 4-1 and the FPGA for the baseband processor with data converters is shown in Figure 4-2. The MAC processor (not shown) is based on a LEON-2 processor with an additional hardware accelerator for time critical operations such as CRC-check, timers and acknowledgement generation.

Even though the demonstrator does not reach the performance of the system proposed in this document, it does proof the manufacturability in general.

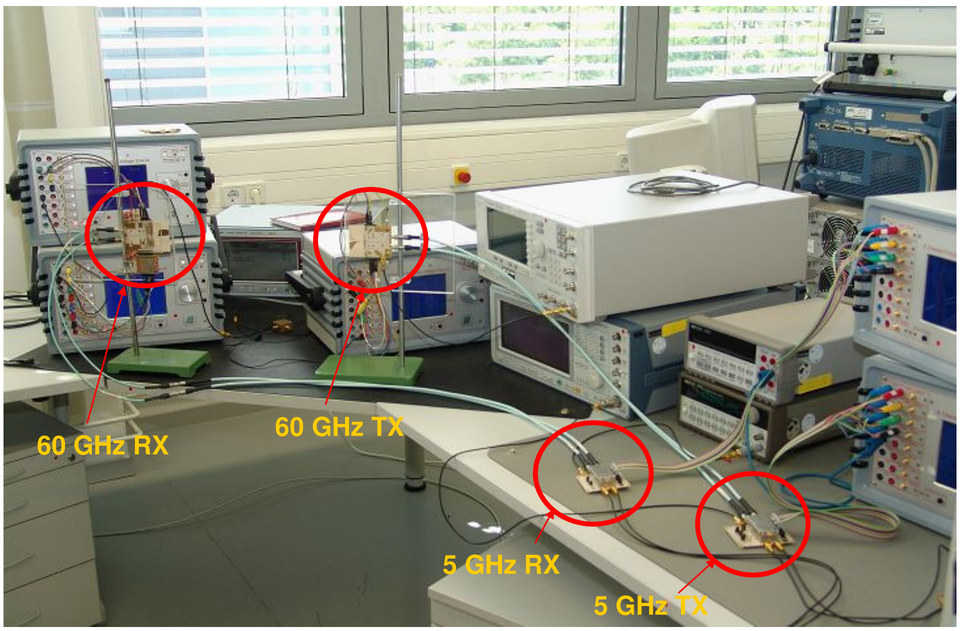


Figure 4-1. Demonstration of 60 GHz Analog Frontend

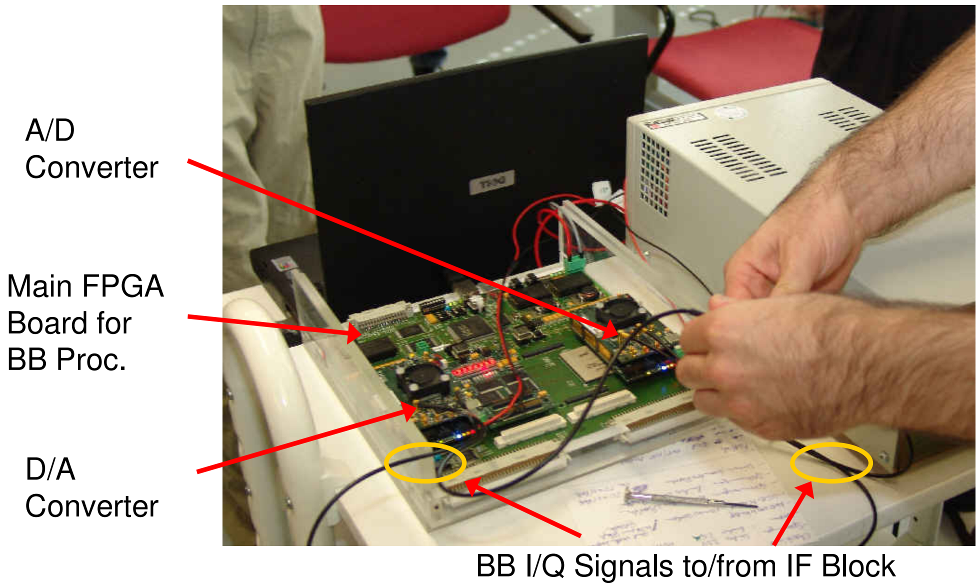


Figure 4-2. FPGA implementation of baseband processor

4.3.2. Time to Market

Based on already existing IP cores and on developed components, we assume the following timeline for products based on IEEE802.15.3c:

- 2 years after standard: first product prototypes
- 3 years after standard: Selling first products in volume
- 5 years after standard: Second generation of high performance products (low power, small form factor, low cost)

4.3.3. Regulatory Impact

The defined frequency channels are given in Table 3-1, while the current status of the regulation in Europe, Australia, Canada, Japan, and USA is summarized in Figure 3-1. Out of the 9 defined channels, 2 will be available in Australia, 7 in Canada, 7 in Japan and 7 in the USA. According to the current discussions regarding 60 GHz regulations in Europe, 8 channels could be defined in this region. Furthermore, smaller channels of 500 MHz can be defined, which allows for an even more flexible compliance with most regulatory specifications.

4.4. Scalability

A high degree of scalability was targeted during the design stage of the proposed system, which can be demonstrated for the following items :

- PHY-SAP Payload bit Rate and Data Throughput :

Nine different payload bitrates have been defined for different modes of operation, as specified in Table 2-4 and Table 2-5. Additional, intermediate bit rates may be defined using additional code rates. The data throughput efficiency may also be scaled by appropriately selecting the number of MAC frames incorporated within one PHY frame (up to 20).

- Channelization and frequencies of operation :

Up to nine channels of 1 GHz each have been defined in the 57 – 66 GHz range (Table 3-1), allowing for an optimum use of the international regulations frequency grid. 3 to 4 common channels may be defined between 59 and 63 GHz for most of the regions of interest in the world, while other channels may be used locally depending on the local regulatory status.

- Occupied bandwidth of operation :

The occupied bandwidth of operation may be scaled from 500 MHz to 2 GHz by respectively splitting one 1 GHz channel or merging two adjacent 1 GHz channels.

- PHY system parameters :

Several baseband system parameters present a high degree of scalability. Three values of Cyclic Prefix may be used to adapt to the channel conditions. The coding scheme may be split into several streams of parallel coders in order to relieve the constraints put on the Viterbi decoder. Finally, different settings for the binary and tone interleavers are provided to scale the system performance according to the desired complexity.

- Power consumption :

An estimate of the total Power Dissipation at a data rate of 2 Gb/s, for implementations in 2008, is given is section 6.9. The power dissipation of the digital circuitry should be reduced to 50% with the transition to the next technology node within two years. Due to advances in circuit design techniques and in process technology, the power dissipation of the analog modules will reduce to approximately 75% every 2 years.

5. MAC Protocol Supplements

5.1. Alternate PHY Required MAC Enhancements and Modifications

5.1.1. PHY-dependent MAC Parameters

The values for some MAC parameters in the IEEE 802.15.3 standard are specified as "PHY-dependent" (see clause 8.15, table 61). For the 2.4 GHz physical layer, these are specified in clauses 11.2.7, 11.2.8 (table 71) and 11.5.5.
For the 60 GHz PHY we propose the following values (for comparison, the 2.4 GHz values are also given):

Parameter name	proposed value – 60 GHz	value – 2.4 GHz
pPHYMIFSTime	1 µs	2 µs
pPHYSIFSTime	8 µs	10 µs
pCCADetectTime	2 µs	5*11/16 µs
pPHYChannelSwitchTime	500 µs	500 µs
pPHYClockAccuracy	+/- 15 ppm	+/- 25 ppm
pMaxFrameBodySize	4082 octets	2048 octets
pMaxTransferUnitSize	4066 octets	2044 octets
pMinFragmentSize	128 octets	64 octets

Table 5-1. PHY-dependent MAC parameters

pPHYMIFSTime (MIFS = Minimum Inter-Frame Spacing)

This time is the inter-frame-distance between two frames, when no station is required to switch from receive to transmit direction or vice versa (e.g. frames which have no immediate ACK or which are sent from different stations).
The proposed value is 1 µs, which corresponds to a few multiples of the longest propagation delay within a Piconet cell.

pPHYSIFSTime (SIFS = Short Inter-Frame Spacing)

This time is the inter-frame-distance between two frames, when the transmitter and/or receiver is required to switch from receive to transmit direction or vice versa (e.g. frames with immediate ACK).
The proposed value is 8 µs, which is the longest propagation delay (MIFSTime) plus estimated processing delays in PHY receiver (4 µs), PHY transmitter (1 µs) and MAC for ACK generation (2 µs).

pCCADetectTime (CCA = Clear Channel Assessment)

This is the time a station needs to safely detect the channel becoming busy or idle.
The proposed value is 2 µs, which is about that part of a frame's preamble that shall be used to settle the receiver's automatic gain control.

pPHYChannelSwitchTime

This is the time required by the PHY to switch to another RF channel.
The proposed value is 500 µs, which is the estimated settling time of the PHY's frequency oscillators (Phase Locked Loop, PLL). If required, the value could be further reduced to 200 µs.

BIFS, RIFS, pBackoffSlot

The values are specified in the same way as for the 2.4 GHz PHY (see table 72).

pPHYClockAccuracy

This is the required accuracy of the clock, which defines frame timing on the MAC level.
The proposed value is 15 ppm, which roughly corresponds to the ratio of 1 MIFS time (1 µs) to the longest possible superframe duration (65535 µs).

pMaxFrameBodySize

This is maximum size of the MAC payload (excluding PHY and MAC headers but including frame check sum FCS [4 byte]).
The proposed value is 4082 octets, which is the largest possible 12 bit number (= 4095) minus 1 octet for Viterbi tail bits minus 12 octets for MAC header including HCS. The value 4082 includes the FCS (4 byte).
The PHY allows concatenating up to 20 MAC frames in one PHY train, allowing for up to 80 kbyte MAC payload in one PHY frame.

pMaxTransferUnitSize

This is maximum size of a data unit passed from or to higher layers.
The proposed value is 4066 octets, which is 12 octets (for security overhead) plus 4 octets (for FCS) less than pMaxFrameBodySize.

pMinimumFragmentSize

The proposed value is 128 octets.

5.1.2. PHY-management Parameters

These parameters are used to describe PHY parameters in MAC information elements, see clause 7.4.11. Their values for the 2.4 GHz PHY are specified in clause 11.7 (tables 89 and 90).
For the 60 GHz PHY, we propose the following values:

SupportedDataRates

This is a 5 bit parameter. The following values are assigned:

00000 bin.	only lowest rate (375 Mbit/s) is supported
others	to be defined later

PreferredFragmentSize

This is a 3 bit parameter. The following values are assigned:

000 bin.	pMaxFrameBodySize (= 4082)
others	to be defined later
111 bin.	pMinFragmentSize (= 128)

The specification of PHY dependent PIB values (tables 92 and 93 in the 802.15.3 standard) is straightforward and can be done later.

6. PHY Layer Criteria

6.1. Size and Form Factor

The form factor of a complete 60 GHz module for different applications can estimated as follows:

1. Access point with Gbit/s connection to wired infrastructure with sector antennae:
   10 cm x 6 cm x 4 cm

2. High performance module for stationary or nomadic applications (single or multi antenna):
   6 cm x 4 cm x 3 cm (also PCMCIA or Card bus)

3. Low cost reduced capability modules (single antenna):
   2 cm x 2 cm x 1 cm

6.2. PHY-SAP Payload Bit Rate and Data Throughput

6.2.1. PHY-SAP Payload Bit Rate

The nominal PHY payload bit rates range from 335 Mbit/s to 3 Gbit/s. In total, 24 data rates (i.e. modulation schemes) are defined, which differ in modulation (BPSK, QPSK, QAM16, QAM64), code rate of convolutional coder (1/2, 2/3, 3/4, 5/6), and guard interval (96 ns, 160 ns, 220 ns). Details are provided in section 2.3.3.

6.2.2. Packet Overhead

Typical overhead times are:

- Frame preamble duration: 6.656 µs
- OFDM guard time: 96 ns, 160 ns, or 220 ns for 1024 ns net symbol duration
- Signal field duration: 1.12 µs (1 OFDM symbol with 96 ns guard)
- Minimum inter-frame spacing MIFS: 1 µs
- Short inter-frame spacing SIFS: 8 µs

- PHY header: 52.5 byte (includes 1st MAC header)
- MAC header: 12 byte (including 2 byte header checksum)
- Frame checksum (FCS): 4 byte
- Viterbi tail bits: 1 byte

6.2.3. Data Throughput

The throughput through the MAC protocol, which can be achieved at a certain PHY data rate, depends on the MAC and PHY overhead and the MAC payload length. The following figures show simulation results for two typical cases.
The goodput is calculated as

goodput (in Mbit/s) = MAC payload size (in bit) / frame transfer time (in µs)

where

frame transfer time = PHY overhead time + payload transfer time

and

payload transfer time = MAC payload size / nominal PHY data rate

The payload transfer time is rounded up to the next multiple of the OFDM symbol duration (1.12 µs, i.e. 96 ns guard). The simulations do not consider backoff nor MAC overhead due to beacons or other management frames.

Frame transfer without acknowledgement

This case is the basic case, which represents the losses due to the overhead of every single frame transmission itself. It directly applies to broadcast frames and other non-acknowledged data (e.g. video streams). In this case, the PHY overhead time is calculated as follows:

PHY overhead time = preamble duration (6.656 µs)
+ 1 OFDM symbol (signal field, 1.12 µs)
+ 1 MIFS time (1 µs)

The OFDM symbol duration includes 96 ns guard interval. The resulting PHY overhead time is 8.776 µs.

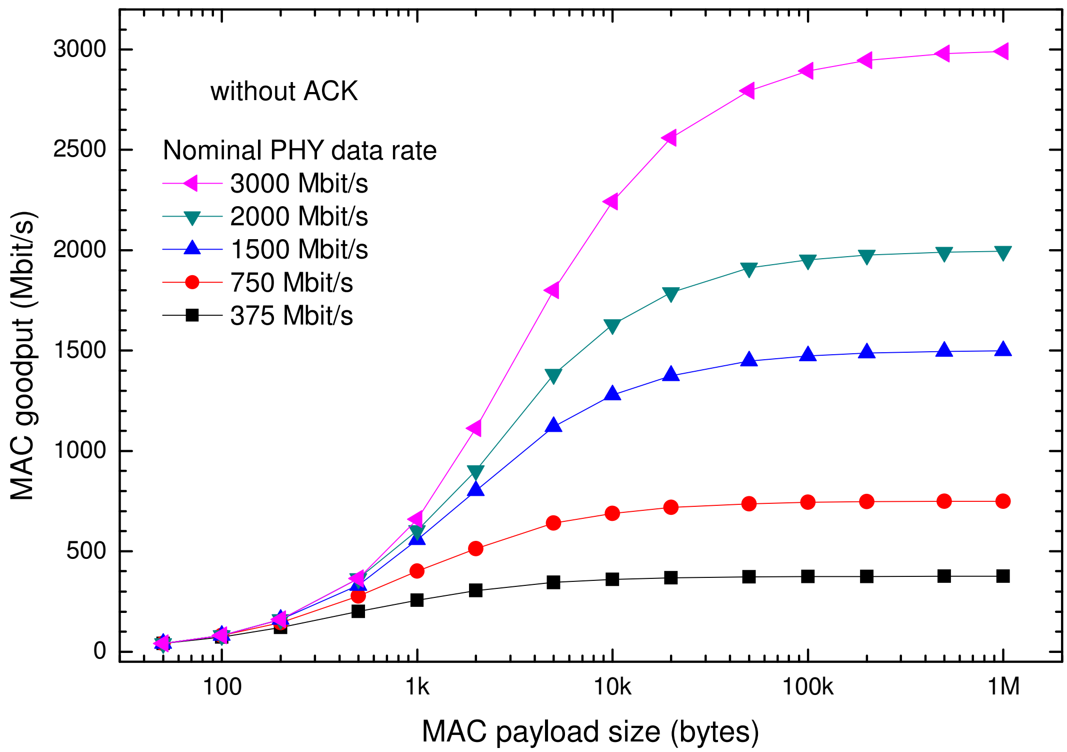


Figure 6-1. MAC goodput without acknowledgement

Numerical values for some representative cases are provided in the following table. A payload of 2048 byte is the reference value set in the downselection criteria. 4078 bytes is the largest MAC payload of one single MAC frame in our proposal, 81 560 bytes the largest accumulated MAC payload size (20 MAC frames) within on PHY train.
For any high-datarate system, the reference frame size of 2 kbyte from the downselection criteria is much too small to achieve satisfactory throughput. Therefore, our frame sizes may reach about 80 kbyte.

Our proposal does not define frames with short preambles as suggested in Figure 2 of the Selection Criteria document.

Nominal PHY data rate	Goodput for MAC Payload size of		
	2048 byte	**4078 byte**	**81 560 byte**
375 Mbit/s	306 Mbit/s = 82 %	339 Mbit/s = 90 %	373 Mbit/s = 99 %
750 Mbit/s	526 Mbit/s = 70 %	622 Mbit/s = 83 %	742 Mbit/s = 99 %
1500 Mbit/s	820 Mbit/s = 55 %	1046 Mbit/s = 70 %	1468 Mbit/s = 98 %
2000 Mbit/s	924 Mbit/s = 46 %	1276 Mbit/s = 64 %	1943 Mbit/s = 97 %
3000 Mbit/s	1140 Mbit/s = 38 %	1633 Mbit/s = 54 %	2872 Mbit/s = 96 %

Table 6-1. MAC goodput without acknowledgement

Frame transfer including group acknowledgement

This case includes the acknowledgement. Since our proposal allows for multiple MAC frames sent within one PHY train, we assume a group acknowledgement as defined in the 802.15.3 MAC standard that acknowledges all of them. Individual (immediate) acknowledgements would severely degrade the performance to roughly the value for a packet size of 4 kbyte. With group acknowledgement, the PHY overhead time is calculated as follows:

PHY overhead time = frame's overhead excl. MIFS as in last section

+ frame duration of the group ACK frame

+ 2 SIFS time (2 x 8 μs)

The group ACK frame consists of the preamble, the signal field, and one OFDM symbol. This is sufficient to acknowledge up to 20 frames, which could be sent within one PHY train. The ACK's duration is 6.656 + 2 x 1.12 μs = 8.896 μs. The resulting total PHY overhead time is 32.672 μs.

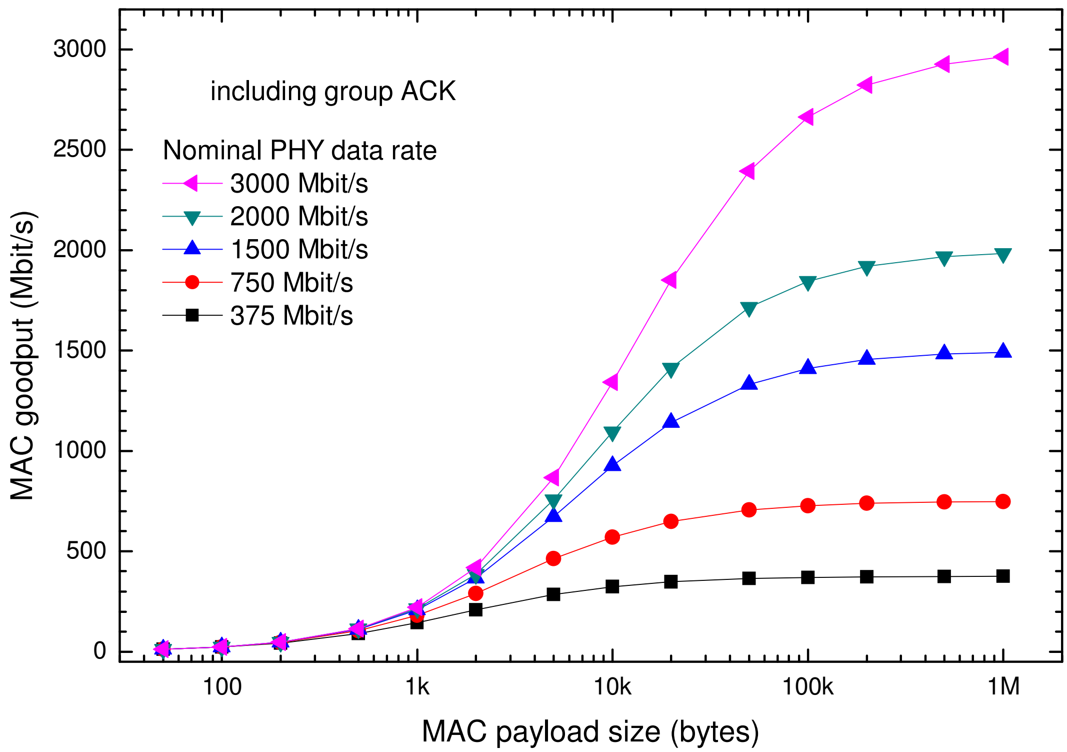


Figure 6-2. MAC goodput with group acknowledgement

Nominal PHY data rate	Goodput for MAC Payload size of		
	2048 byte	**4078 byte**	**81 560 byte**
375 Mbit/s	211 Mbit/s = 56 %	272 Mbit/s = 72 %	368 Mbit/s = 98 %
750 Mbit/s	298 Mbit/s = 40 %	427 Mbit/s = 57 %	722 Mbit/s = 96 %
1500 Mbit/s	373 Mbit/s = 25 %	592 Mbit/s = 39 %	1393 Mbit/s = 93 %
2000 Mbit/s	394 Mbit/s = 20 %	659 Mbit/s = 33 %	1814 Mbit/s = 91 %
3000 Mbit/s	428 Mbit/s = 14 %	743 Mbit/s = 25 %	2599 Mbit/s = 87 %

Table 6-2. MAC goodput with group acknowledgement

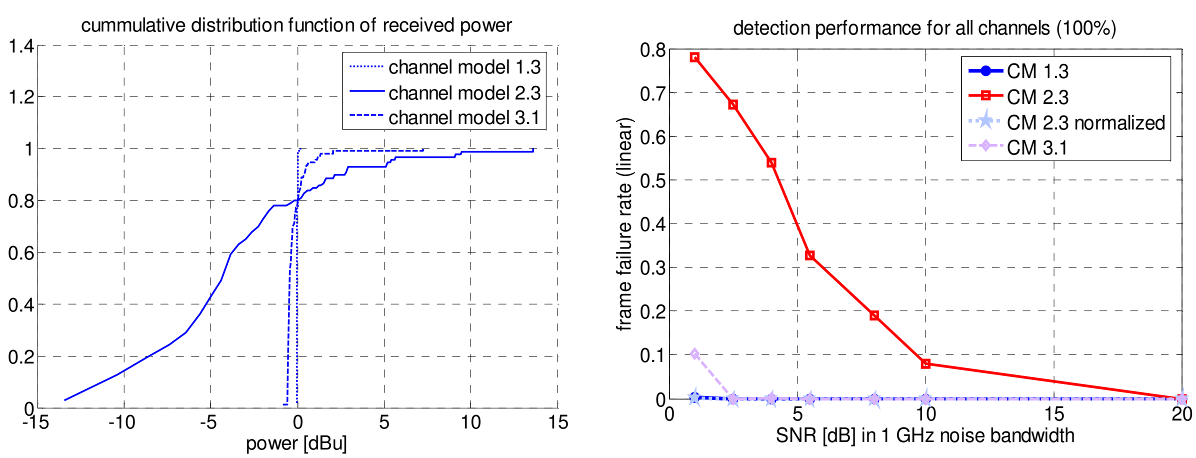
6.3. Co-Channel and Cross-Channel Interference

The simulations assessing the performance of the proposed PHY in the presence of multiple uncoordinated piconets should be completed for the next TG3c meetings.

6.4. Signal Acquisition

6.4.1. Signal acquisition performance

The signal acquisition performance is given for the defined channel models CM1.3, CM3.1 and CM2.3. and also for CM2.3n, which shall denote the case when every channel response taken from CM2.3 is normalized in time domain to have unity gain. For the channel models CM1.3, CM2.3 and CM3.1, this normalization was not applied. The cumulative distribution function shows a very high spread for the received power for CM2.3. As a result, the frame detection rate reaches a failure rate of below 1% only at high signal-to-noise ratios above 18 dB. For channel model CM2.3n, the missing rate is 0 down to SNR = 1 dB. Taking all channel models including CM2.3n but excluding CM2.3 into account, the simulations showed a failure rate below 1% for SNR >= 2 dB (number of simulated frames = 5000). A false alarm was never observed. The simulation includes phase noise with the following parameters: L_{ssb}=-92 dBc/Hz @ 1 MHz, cutoff (pole) = 100 kHz, noise floor = -130 dBc/Hz. A Rapp model with p=2 and 10 dB output backoff was applied to model the power amplifier nonlinearities.



(a) (b)

Figure 6-3. CDF of the received power for different channel models (a), and frame failure rate (b)

6.4.2. Frequency synchronization performance

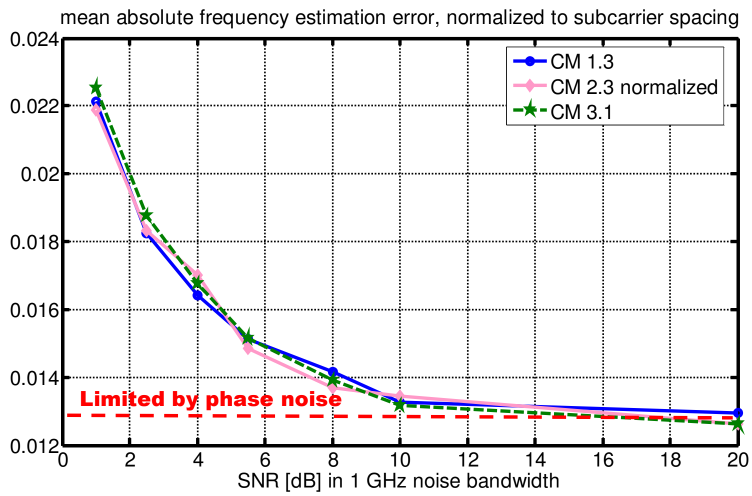


Figure 6-4. Normalized mean absolute frequency estimation error

The frequency synchronization performance shows that the mean remaining carrier frequency offset (CFO) is 1.6% for an SNR of 5 dB. The relative high phase noise limits the mean absolute frequency estimation error to 1.3%. Lower phase noise values allow better CFO estimation.

6.4.3. Timing performance

During coarse frame synchronization, a suitable FFT position is estimated for the second part of the preamble for the purpose of channel estimation and fine time synchronization. The guard time is 384 ns for the two B-sequences. After fine time synchronization, a suitable FFT reference position is estimated for the signal field and the successive data symbols
As a figure of merit for timing performance we chose the mean signal-to-interference ratio (SIR). We define the SIR with respect to the channel power-delay profile as the ratio of the useful received energy inside the guard interval to the remaining energy outside the guard interval. The location of the guard is determined by the chosen FFT position. This is illustrated in figure Figure 6-5. The accumulated power of the channel profile inside the guard interval is equivalent to the useful signal, whereas the channel echoes outside the guard constitute interference. Note that the bound for the signal-to-noise ratio seen by the subcarriers resulting from this inter-symbol interference (ISI) is higher than the SIR, because the interference signals only partly overlap with the FFT window.

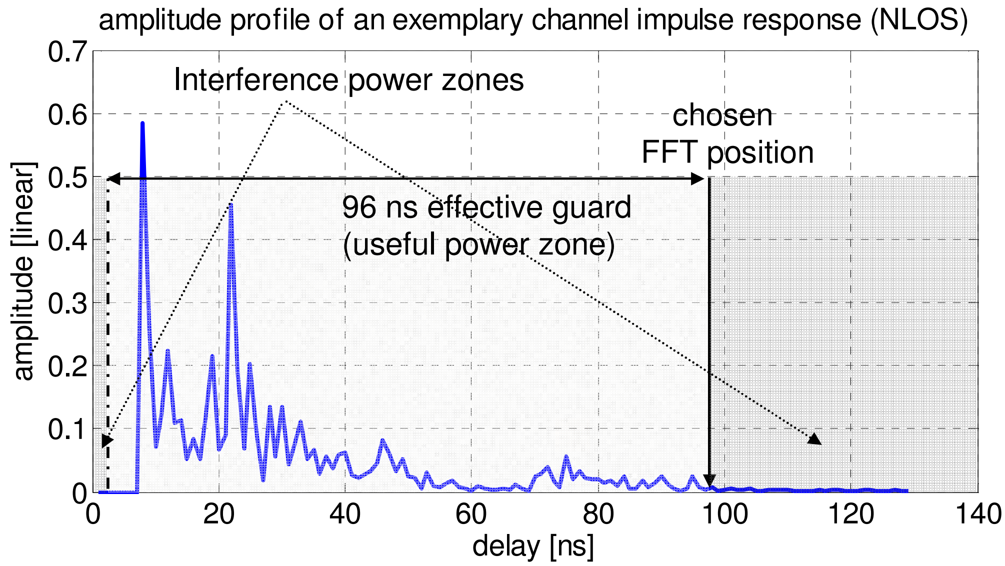
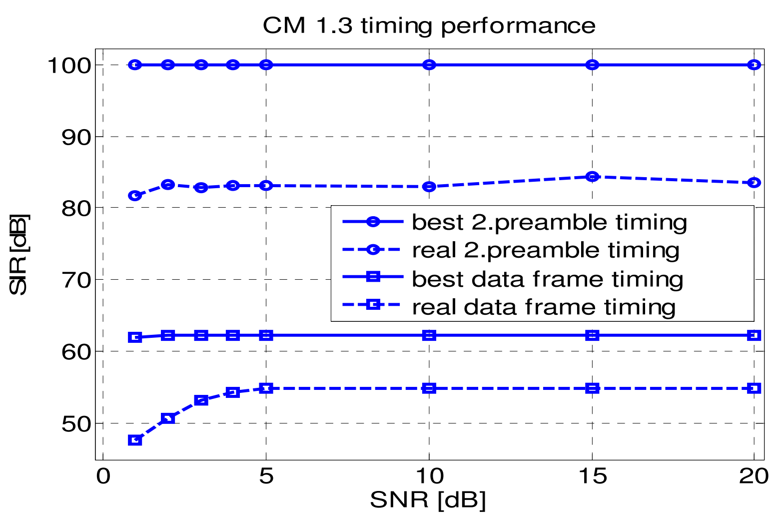
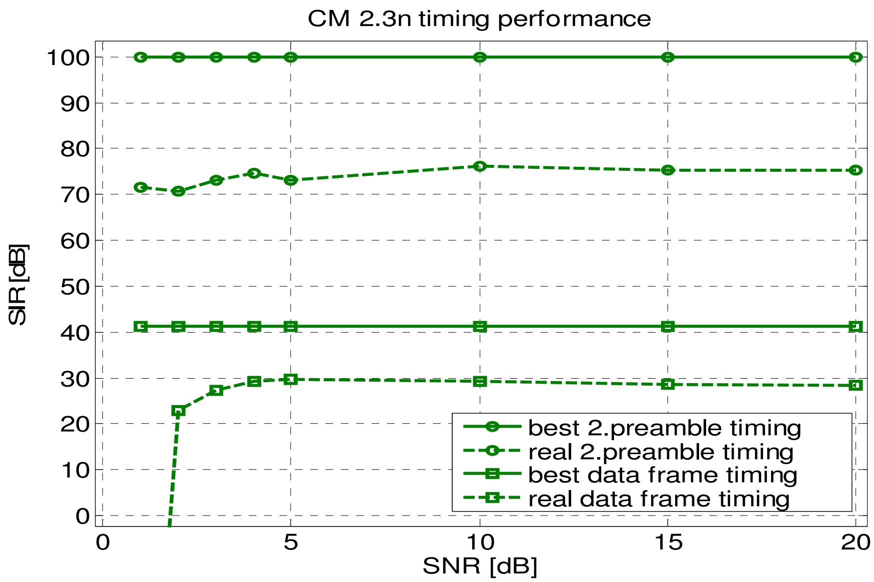


Figure 6-5. Illustration of useful power and interference power

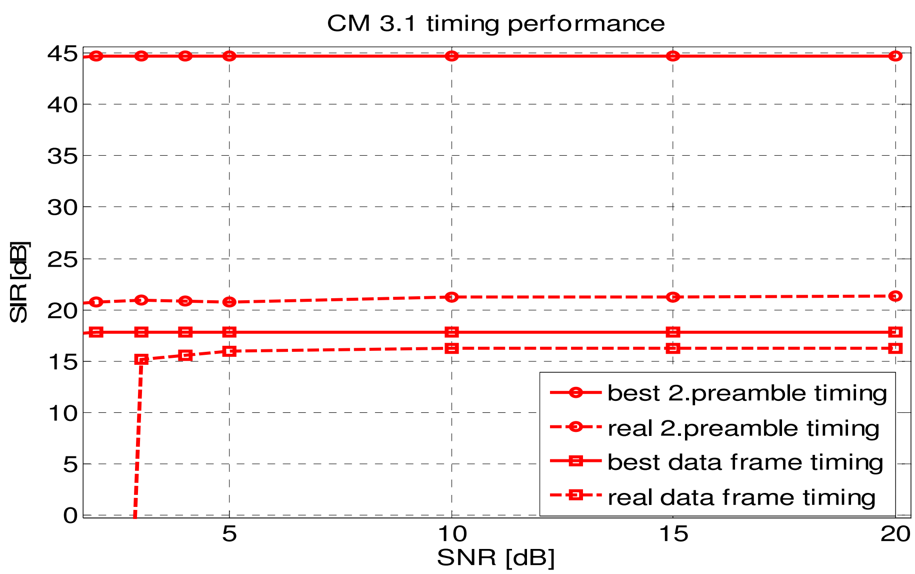
The next three figures show the obtained mean SIR for coarse and fine frame synchronization for channel models CM1.3, CM2.3 and CM3.1. The guard time for OFDM data symbols was chosen as T_g = 96 ns. SIR values above 100 dB are limited to 100 dB for convenient presentation. A SIR of at least 29 dB is obtained for CM1.3 and CM2.3. Channel model CM3.1 shows a very large delay spread, which leads to a performance degradation for 64-QAM. With imperfect timing, this effect is even more severe. For 16-QAM, synchronization performance is very good for all channel models (not shown here).



(a)



(b)



(c)

Figure 6-6. Timing performance for CM1.3 (a), CM2.3 (b), CM3.1 (c)

6.5. System Performance

The proposed system has been assessed through link level simulations for the mandatory mode providing a PHY payload bitrate of 2 Gbps and the optional mode providing a PHY payload bitrate of 3 Gbps, both over a channel bandwidth of 1 GHz. For theses two modes, a Cyclic Prefix of duration 96 ns was used and a binary interleaver depth of 4 OFDM symbols was used. The other system parameters are given in Table 2-4.

The simulations assumptions are as follows :

- Channel models assessed: AWGN, Residential LOS (CM1.3), Residential NLOS (CM2.3) and Office LOS (CM3.1).
- Spatial filtering: a 30° HPBW antenna is assumed at the receiver.
- Simulation scenario: 64 packets (payload 2048 bytes) are transmitted for 100 different channel realizations, corresponding to a total of about 1E8 transmitted bits.
- Computed metric is the mean 90% BER/PER link success probability as defined in [1].

It should be noted that for each channel model, the normalization is done at the Power Delay Profile level, so that realistic power fluctuations are considered between successive Channel Impulse Responses (CIR), with a unit power on average. For the sake of comparison, channel model CM2.3 was also normalized at the CIR level, i.e. each CIR is separately normalized to a unit power. This theoretical channel is denoted CM2.3n.

The following implementation impairments were included in the simulations :

- Channel estimation: realistic channel estimation for CM1.3, CM2.3 and CM3.1, no channel estimation for AWGN.
- Amplifier nonlinearities: the Rapp model was used, with an output backoff (OBO) of 10 dB and p=2. AM-PM distortion was not simulated.
- Phase noise: a single-zero, single pole model was used as defined in [1], with the parameters f_p = 1 MHz, f_z = 100 MHz, PSD(0) = -87 dBc/Hz for the 2 Gbps mode. For operation at 3 Gbps, we assume that more efficient RF components should be used, and we used the parameters f_p = 1 MHz, f_z = 100 MHz, PSD(0) = -92 dBc/Hz.

The resulting curves are given in Figure 6-7 to Figure 6-10.

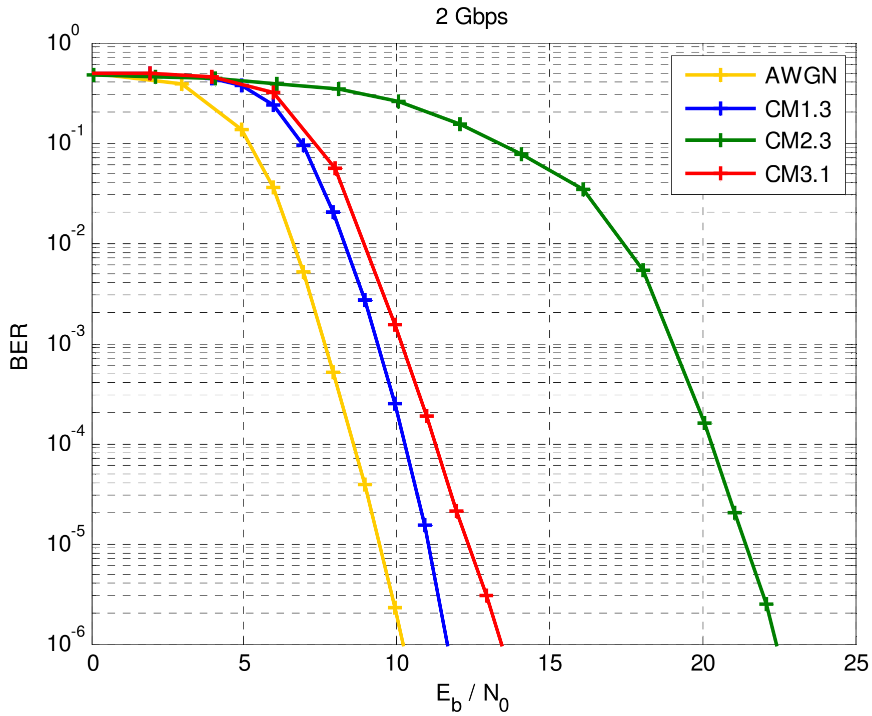


Figure 6-7. Mean 90% BER link success probability for different channels (2Gbps)

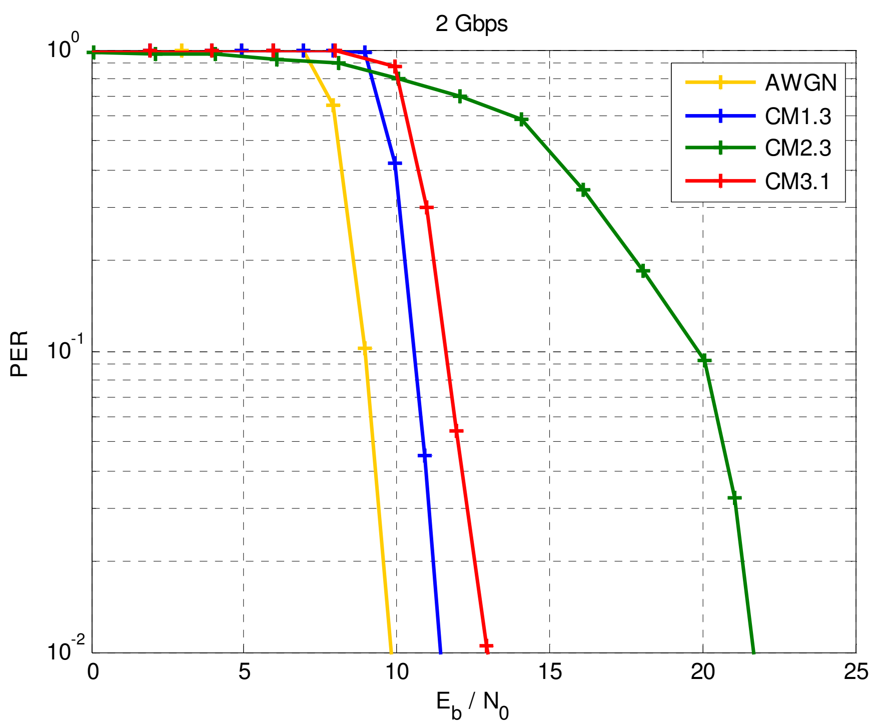


Figure 6-8. Mean 90% PER link success probability for different channels (2Gbps)

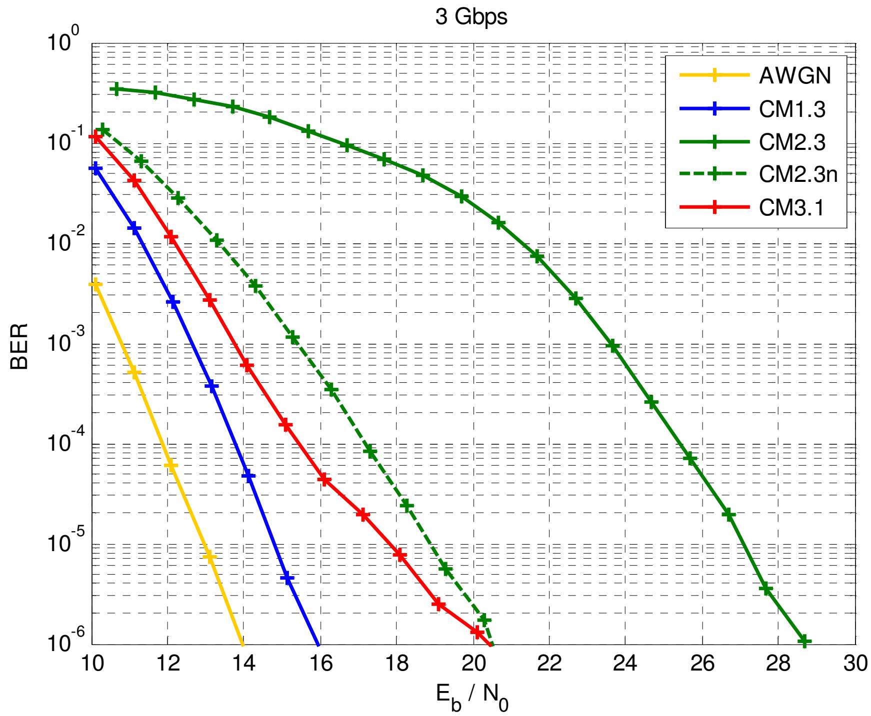


Figure 6-9. Mean 90% BER link success probability for different channels (3Gbps)

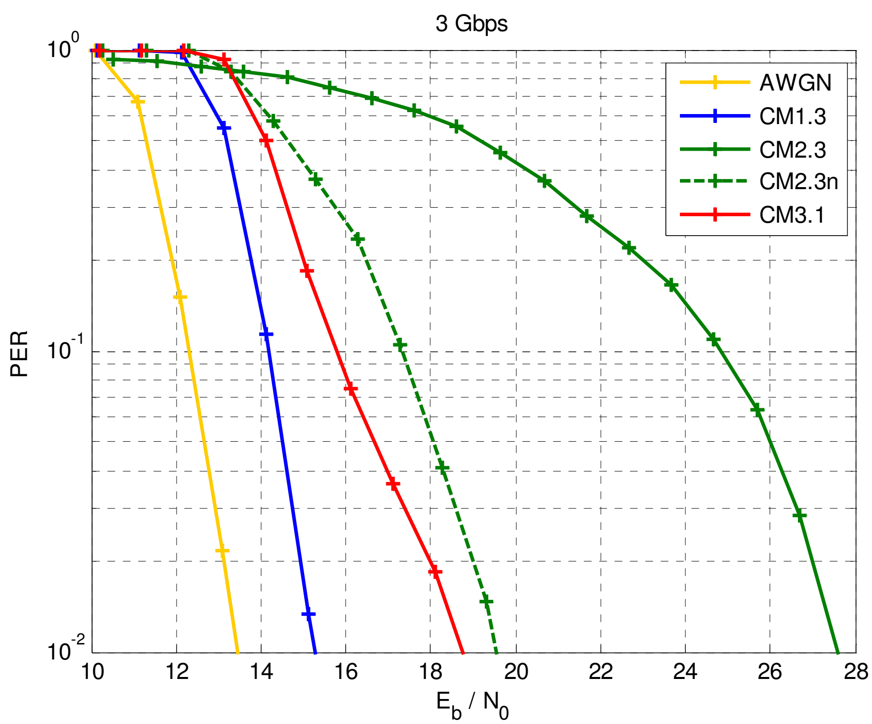


Figure 6-10. Mean 90% PER link success probability for different channels (3Gbps)

From these curves and the link budget tables in the next section, the mean 90% link success probability distance are reported below. The BER criterion corresponds to a 90% BER link success probability of 1E-6, and the PER criterion corresponds to a 90% PER link success probability of 8%

Table 6-3. Mean 90% link success probability distance are reported below

		AWGN		CM1.3		CM2.3		CM3.1	
		2 Gbps	3Gpbs	2 Gbps	3Gpbs	2 Gbps	3Gpbs	2 Gbps	3Gpbs
90% mean link success probability distance(m)	BER criterion	6,92	3,63	5,89	2,88	3,5	0,72	1,7	1,74
	PER criterion	7,85	4,22	6,46	3,43	4,06	0,99	2,16	2,88

6.6. Link Budget

The following table gives the link budget computed for AWGN, CM1.3, CM2.3 and CM3.1 for the two modes of operation (2 Gbps and 3 Gbps) defined in the previous section.

Table 6-4. Link budget table (2 Gbps)

Parameter	**Value**	**Value**	**Value**	**Value**	**Unit**
Channel model	AWGN	CM1.3	CM2.3	CM3.1	
PHY-SAP Payload Bit Rate (R_b)	2	2	2	2	Gb/s
Average Tx power (P_T)	10	10	10	10	dBm
Tx antenna gain (G_T)	9	9	9	9	dBi
Center frequency (f_c)	60	60	60	60	60GHz
Path loss at 1 meter ($PL_0 = 20\log_{10}(4*PI*f_c/c)$), $c = 3*10^8$ m/s	68	68	68	68	68.00dB
Rx antenna gain (G_R)	9	9	9	9	dBi
Average noise power per bit (N=-174+10*log$_{10}$(R_b)))	-81	-81	-81	-81	dBm
Rx Noise Figure Referred to the Antenna Terminal (N_F)	6	6	6	6	dB
Average noise power per bit ($P_N = N + N_F$)	-75	-75	-75	-75	dBm
Minimum E_b/N_0 for BER criterion	10,2	11,6	13,4	22,4	dB
Minimum Eb/N0 for PER criterion	9,1	10,8	11,8	20,3	dB
Shadowing link margin ($M_{shadowing}$)	5	5	5	5	dB
Implementation Loss (I)	3	3	3	3	dB
Tolerable path loss ($PL = P_T+G_T+G_R-P_N-S-M_{shadowing}-I-PL0$) (BER criterion)	16,8	15,4	13,6	4,6	dB
Tolerable path loss ($PL = P_T+G_T+G_R-P_N-S-M_{shadowing}-I-PL0$) (PER criterion)	17,9	16,2	15,2	6,7	dB
Maximum operating range ($d = 10^{PL/10n}$) (BER criterion)	6,92	5,89	3,5	1,7	m
Maximum operating range ($d = 10^{PL/10n}$) (PER criterion)	7,85	6,46	4,06	2,16	m

Table 6-5. Link budget table (3 Gbps)

Parameter	Value	Value	Value	Value	Unit
Channel model	AWGN	CM1.3	CM2.3	CM3.1	
PHY-SAP Payload Bit Rate (R_b)	3	3	3	3	Gb/s
Average Tx power (P_T)	10	10	10	10	dBm
Tx antenna gain (G_T)	9	9	9	9	dBi
Center frequency (f_c)	60	60	60	60	60GHz
Path loss at 1 meter ($PL_0 = 20\log_{10}(4{*}PI{*}f_c/c)$), $c = 3{*}10^8$ m/s	68	68	68	68	68.00dB
Rx antenna gain (G_R)	9	9	9	9	dBi
Average noise power per bit (N=-174+10*log$_{10}$(R_b)))	-79,2	-79,2	-79,2	-79,2	dBm
Rx Noise Figure Referred to the Antenna Terminal (N_F)	6	6	6	6	dB
Average noise power per bit ($P_N = N + N_F$)	-73,2	-73,2	-73,2	-73,2	dBm
Minimum E_b/N_0 for BER criterion (S)	14	16	28,7	20,4	dB
Minimum Eb/N0 for PER criterion (S)	12,7	14,5	25,3	16	dB
Shadowing link margin ($M_{shadowing}$)	5	5	5	5	dB
Implementation Loss (I)	3	3	3	3	dB
Tolerable path loss ($PL = P_T + G_T + G_R - P_N - S - M_{shadowing} - I - PL0$) (BER crit.)	11,2	9,2	-3,5	4,8	dB
Tolerable path loss ($PL = P_T + G_T + G_R - P_N - S - M_{shadowing} - I - PL0$) (PER crit.)	12,5	10,7	-0,1	9,2	dB
Maximum operating range ($d = 10^{PL/10n}$) (BER criterion)	3,63	2,88	0,72	1,74	m
Maximum operating range ($d = 10^{PL/10n}$) (PER criterion)	4,22	3,43	0,99	2,88	m

6.7. Sensitivity

The following table gives the minimum receiver sensitivity for the two modes of operation (2 Gbps and 3 Gbps) defined in the previous section, for the PER criterion defined in section 6.5.

Table 6-6. Sensitivity

PHY Bit Rate (Gbps)	Sensitivity (dBm)
2	- 62.9
3	- 57.5

6.8. Power Management Modes

The proposed PHY should at least support the Power Management Modes sleep, wakeup and poll as defined in the IEEE 802.15.3-2003 standard.

6.9. Power Consumption

6.9.1.1. Transmit and Receive

The following list indicates is an estimate of the total Power Dissipation at a data rate of 2 Gb/s (65 nm CMOS digital; 130 nm analog SiGe) for implementations in 2008. It is assumed that the power dissipation of the digital circuitry is reduced to 50% with the transition to the next technology node within two years. Due to advances in circuit design techniques and in process technology, the power dissipation of the analog modules will reduce to approximately 75% every 2 years.

2008:	*Transmit*	*Receive*
MAC Processor:	200 mW	200 mW
Baseband Processor:	200 mW	350 mW
Data Converters:	100 mW	150 mW
Analog Frontend	200 mW	200 mW
Power Amplifier	150 mW	20 mW
Total (continuous):	**850 mW**	**920 mW**

For 2010 the following power dissipation figures for 60 GHz transceivers can be achieved:

2010:	*Transmit*	*Receive*
MAC Processor:	100 mW	100 mW
Baseband Processor:	100 mW	175 mW
Data Converters:	75 mW	100 mW
Analog Frontend	150 mW	150 mW
Power Amplifier	125 mW	10 mW
Total (continuous):	**550 mW**	**535 mW**

6.9.1.2. Clear Channel Assessment
CCA Power dissipation (2008): ca. 250 mW
CCA Power dissipation (2010): ca. 200 mW

6.9.1.3. Power Save

TBD

6.10. Antenna Practicality

A number of antenna types were presented in the past. The most promising solutions are:

- Patch array antennae (also suitable for beam steering as phased array antenna)
- Patch antenna (low cost)
- Vivaldi antenna (low cost)
- Dipole antenna (low performance in-package and on-chip applications)

Due to the short wavelength of 5 mm, the antenna can be very small. This is a considerable advantage of 60 GHz systems.

7. References

[1] Seyedi, A., "TG3c Selection Criteria", *IEEE 802.15 Working Group for WPANs,* no. IEEE 802.15-05-493-26-003c, Jan. 2007.

[2] IEEE 802.11a-1999 standard, "Supplement to IEEE standard for information technology telecommunications and information exchange between systems - local andmetropolitan area networks - specific requirements. Part 11: wirelessLAN Medium Access Control (MAC) and Physical Layer (PHY) specifications:high-speed physical layer in the 5 GHz band", 1999.

[3] Siaud.I, Ulmer-Moll A.M, "A Novel Adaptive sub-carrier Interleaving : application to millimeter-wave WPAN OFDM Systems (IST MAGNET project)", *IEEE portable 2007 conf, 25-29* March 2007, Orlando (USA).

[4] Siaud.I, Ulmer-Moll, "Advanced Interleaving algorithms for OFDM based millimeter wave WPAN transmissions", SCEE Seminar, 8 February 2007, France.

[5] Pagani, P., Siaud, I., Ulmer-Moll, A. & Li, W., "High rate OFDM system for 60 GHz WPAN", *IEEE 802.15 Working Group for WPANs,* no. IEEE 802.15-07/539, Jan. 2007.

[6] Pagani, P., Siaud, I., Ulmer-Moll, A. & Li, W., "Advanced interleaving for high data rate 60 GHz communications", *IEEE 802.15 Working Group for WPANs,* no. IEEE 802.15-07/627, March 2007.

[7] E. Grass, M. Piz, F. Herzel, R. Kraemer ‘Draft PHY Proposal for 60 GHz WPAN’ IEEE 802.15 Meeting, Document Number: IEEE 802.15-05/0634r1, Vancouver (Can), Nov. 2005.

[8] E. Grass, M. Piz, F. Herzel, K. Schmalz, Y. Sun, S. Glisic, K. Tittelbach-Helmrich ‘60 GHz Demonstrator in 0.25 µm SiGe:C BiCMOS Technology’, IEEE 802.15 Meeting, Document Number: IEEE 802.15-06/0320r0, San Diego (CA), July 2006.

[9] E. Grass, F. Herzel, M. Piz, Y. Sun, R. Kraemer, 'Implementation Aspects of Gbit/s Communication Systems in the 60 GHz Band' Wireless World Research Forum (WWRF) / WG5, San Diego (CA), July 07-08, 2005.

[10] F. Herzel, S. Glisic, W. Winkler ‘Integrated Frequency Synthesizer in SiGe BiCMOS Technology for 60 GHz and 24 GHz Wireless Applications’, Electronics Letters 43(3), 154 (2007)