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Title	**IEEE 802.16b PHY Strawman**
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Re:	IEEE 802.16b (TG4) PHY Strawman
Abstract	This is the strawman PHY document for: Local and Metropolitan Area Networks Amendment to Standard Air Interface for Fixed Broadband Wireless Access Systems Additional Physical Layer for License-Exempt Frequencies
Purpose	The document reflects the work of the TG4 PHY editing team as May 07, 2001 and shall be subject to comments and review during Session #13.
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1. Notations

Comments and TBD issues are marked using *italics*.

2. Introduction (Octavian)

The PHY layer described in this clause is designated for operation in the unlicensed frequency bands of 5-6 GHz. In order to allow different deployment scenarios from dense populated areas (where sectorized environments and the high interference levels may require narrow channels) to sparse populated areas (where wider channels can help delivering better services at the same system cost), the PHY layer is allowed to operate with channel bandwidths of 10 MHz or 20 MHz and optionally 5 MHz. An 802.16b device must implement 10 MHz and/or 20 MHz channelization and may implement 5MHz channelization (see clause 6.2).

The PHY operation is based on OFDM (Orthogonal Frequency Division Multiplex) modulation and Time Division Duplexing (TDD). In OFDM the information is imposed onto the medium by modulating multiple carriers transmitted in parallel. As the modulation is often implemented using the FFT algorithm, the modes are designated by the minimum FFT size (called hereby FFT size), i.e. the next power of 2 above the number of used carriers. The PHY defines two mandatory modes 64 and 256 —FFT and one optional mode 1024-FFT. The mandatory modes employ Time Division Multiple Access (TDMA) while the optional mode employs a combination of TDMA and Orthogonal Frequency Division Multiple Access (OFDMA). The modes are summarized in Table 1.

Mode	Access method	FFT size	Status
64-FFT	TDMA	64	Mandatory
256-FFT	TDMA	256	Mandatory
1024-FFT	OFDMA	1024	Optional

Table 1: Mandatory and optional modes

To be standard compliant, a system must implement both the 64-FFT and the 256-FFT modes.

Notes: changes to this sections where made to reflect the latest group decisions on channel bandwidth and FFT sizes.

3. Time and frequency description (Nico — Editing team)

3.1. Frequency description

An OFDM symbol in the frequency domain consists of orthogonal carriers. An Inverse Fourier Transform is used to transform this signal into the time domain. The following three types of carriers are distinguished:

- Data carriers — Carriers used to modulate data symbols on.
- Pilot carriers — Carriers used for estimation purposes
- Null carriers — Carriers that remain unused to provide guard bands and DC carrier.

The purpose of the guard bands is to enable the signal to naturally decay within the desired spectral mask. By convention, carrier 0 is the DC carrier and it is not used. Other carriers are numbered in their relative position to the DC carrier, using positive incrementation for the adjacent carrier with higher frequency.

F_{clock} = clock frequency at output of IDFT

BW = channel bandwidth (20, 10, and optionally 5MHZ) (see clause 6.2)

DFT sizes: 64, 256 and optionally 1024

3.1.1. Carrier allocation for 64-DFT

In Figure 1, the carrier allocation of 64-DFT is shown. There are 4 pilot carriers at positions –21, -7, 7 and 21. On both positive and negative side of the spectrum, there are 6, 13 and 5 data carries allocated between DC and first pilot, between pilots and at the band edge, respectively. For all channelizations, F_{clock}/BW= 1.

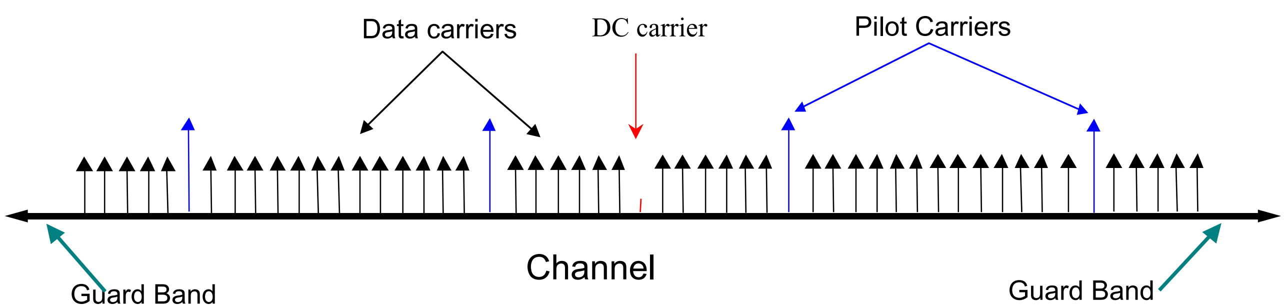

Figure 1: Carrier allocation for 64-DFT

3.1.2. Carrier allocation for 256-DFT

In Figure 2, the carrier allocation of 256-DFT is shown. There are 12 pilot carriers at positions -94, -77, -60, -43, -26, -9, 9, 26, 43, 60, 77 and 94. Between pilots, there are blocks of 16 data carriers. Blocks of 8 data carriers are allocated at both DC sides and both band edges. This allows for a total of 192 data sub-carriers, 12 pilot carriers and 1 DC carrier, for a total of 205 carriers. The other 51 carriers (-103 and below, as well as 104 and above) serve as guard-band.

The carrier allocation of 256-DFT for the optional 5 MHz channelization is the same as the 64-DFT allocation for this channelization as shown in Figure 1. For all channelizations, $F_{clock}/BW= 16/15$.

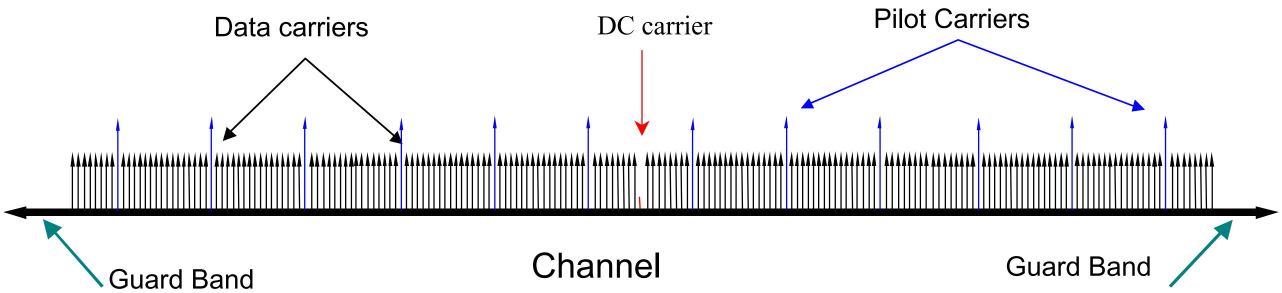

Figure 2: Carrier allocation for 256-DFT

3.1.3. Carrier allocation of 1024-DFT (optional)

In the 1024-DFT mode, a different allocation is used in the up- and down-stream. This is motivated by the intend to efficiently facilitate OFDMA. For all channelizations, $F_{clock}/BW= 8/7$.

3.1.3.1. Downstream

3.1.3.1.1 Frame structure

The transmission of the DS is performed on one or more subchannels of the OFDMA symbol. A DS sub-channel constitutes 48 (non-consecutive) data carriers. The mapping of the subchannels is performed in a two-dimensional grid, involving the subchannels in the frequency domain and OFDM symbols in the time domain. Figure 3 illustrates a possible two-dimensional transmission mapping (every color represents a different Modulation and coding scheme).

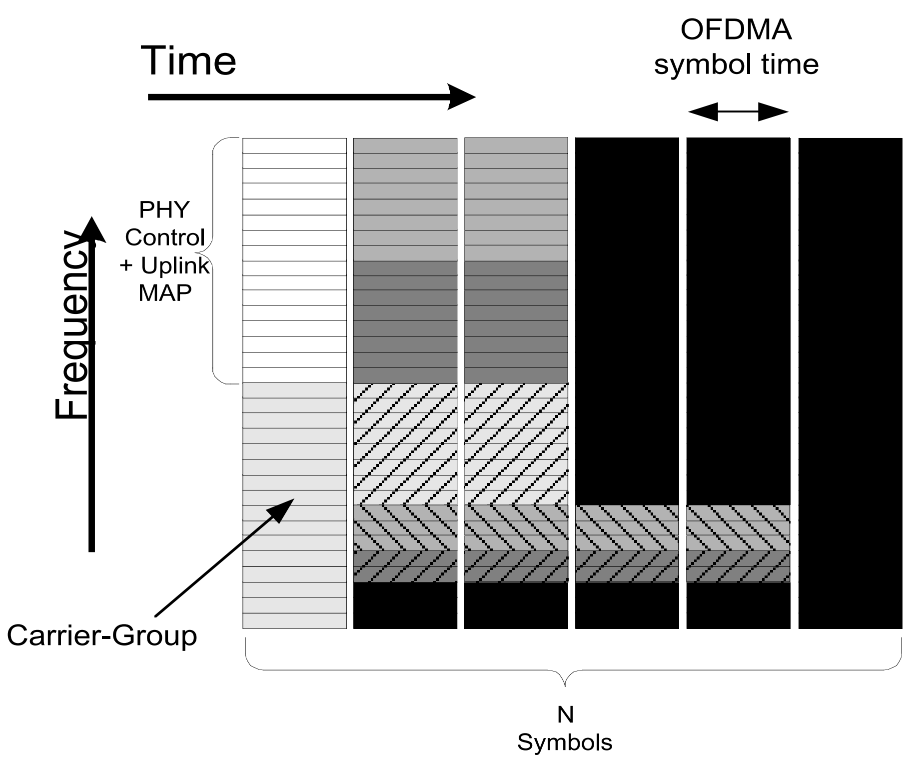

Figure 3: Map II, DS framing

3.1.3.1.2 Symbol Structure

The symbol structure made up of constant and variable location pilots, which are spread all over the symbol, and from data carriers, which are divided into subchannels.

First allocating the pilots and then mapping the rest of the carriers to Sub-Channels construct the OFDMA symbol. There are two kinds of pilots in the OFDM symbol:

- Constant location pilots - Transmitted every symbol
- Variable location pilots — Location shifted with a cyclic appearance of 4 OFDM symbols

The variable pilots are inserted in the locations defined by the next formula:

$$k = 3 * L + 12 * P_V$$

$k \in$ Indices from 0 to the number of Overall Usable Carriers minus 1

$L \in 0..3$ denotes the OFDM symbol number with a cyclic period of 4, with order: 0,2,1,3

$P_v \geq 0$ is an integer number

The pilot s locations are illustrated in Figure 4:

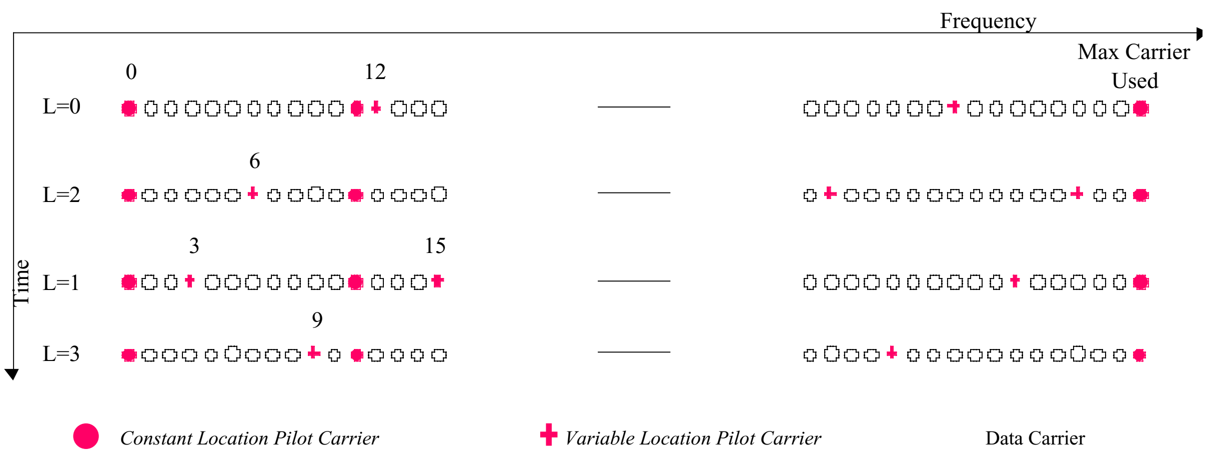

Figure 4: Pilots and data carrier location in the DS OFDMA symbol

After mapping the pilots, the rest of the carriers (not including the DC carrier, which is not used) are used as data carriers, which are grouped into sub-channels as described below.

In order to achieve the DS Sub-Channels, the data carriers are grouped into one space (in acceding order of their indices) and then divided into 48 basic groups ($\{N_{Groups}\}$=48), each containing 16 carriers. Note that as the variable pilots occupy different carriers modulo 4 with the OFDM symbol number, the space of data carriers also changes modulo 4 with the OFDM symbol.

Special permutations, as described in the 5 steps below, are used to extract the Sub-Channels. Each Sub-Channel is made up of 48 data carriers: $\{N_{Sub-Channel}\}$=48. $\{CellId\}$ is a MAC defined parameter, defining the current cell identification numbers, to support different cells.

1. The usable carrier space is divided into $\{N_{Groups}\}$ basic groups (section 3.1.3.1.2, 3.1.3.2.3)
2. We define a basic permutation $\{PermutationBase_0\}$, containing $\{N_{elements}\}$=16 elements
3. Different permutations ($\{PermutatedSeries\}$) are achieved by cyclically rotating the $\{PermutationBase_0\}$ to the left <= Nico:This requires some more specifics (rotation when, and by howmuch)

4. To get a $\{N_{Sub-Channel}\}$ length series we concatenate the permutated series:

$$\left\{\begin{array}{l}(PermutatedSeries + CellId);(PermutatedSeries + 2\cdot CellId);...\\ ...(PermutatedSeries + ceil(N_{Sub-Channel}/N_{elements})\cdot CellId);\end{array}\right\}\mathrm{mod}(N_{elements})$$

5. The last step achieves the carrier numbers allocated for the specific Sub-Channel with the current $CellId$. Using the next formula we achieve the 48 carriers of the current permutation in the cell:

$$Carrier\# = N_{elements} * n + Index(n)$$

where:
$Carrier\#$ - denotes the carrier number for this Sub-Channel using the $\{PermutatedSeries\}$
n - Indices 0.. $\{N_{Sub-Channel} - 1\}$
$Index(n)$ - denotes the number at index n of the $\{N_{Sub-Channel}\}$ length series

Parameter	Value	
N_{FFT}	1024 (1K)	
N_{used}	850	
Guard Carriers: Left, Right	87	87
Subchannels: nr, data carriers/subchannel	16	48
ConstantLocationPilots	{0,39,261,330,348,351,522,645,651,726,756,849,850}	
PermutationBase_0	{6, 14, 2, 3, 10, 8, 11, 15, 9, 1, 13, 12, 5, 7, 4, 0}	

Table 2: Downstream parameters Nico: do the adjacent constant pilots make sense?

3.1.3.2. Upstream

3.1.3.2.1 Subchannel description

The next section gives a description of the structure of a subchannel. A subchannel is made up of 48 usable carriers and 5 pilot carriers. The DS transmission for these modes is also made of subchannel transmissions, but the Sub-Channel is made up of 48 data carriers only, while pilot carriers are spread all over the OFDMA symbol, to be used for channel estimation. The US subchannel structure is shown in Figure 5.

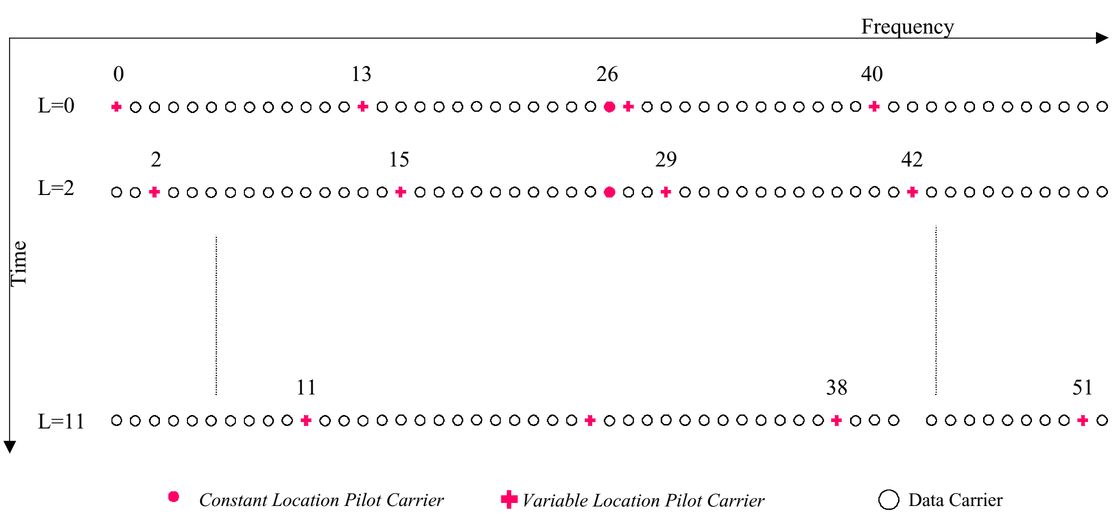

Figure 5: Allocation of data and pilot carriers for a US Sub-Channel

The US data symbol structure is comprised of data carriers and pilot carriers. The data symbols are produced with a modulo 13 repetition (L denotes the modulo 13 index of the symbol with indices 0..12), the location of the variable location pilots are shifted for every symbol produced, the first symbol (L=0) is produced after the all-pilot symbols (preamble). For L=0 the variable location pilots are positioned within the subchannel at indices: 0,13, 27,40 for other L these location vary by addition of L to those position, for example for L=5 variable pilots location are: 5,18, 32, 45. the US Sub-Channel is also comprised of a constant pilot at the index 26. all other carriers (48) are data carriers, their location changes for every L, the transmission ordering of L is 0,2,4,6,8,10,12,1,3,5,7,9,11.

3.1.3.2.2 Framing Structure

The basic allocation for a user US transmission is made up of subchannels, a basic user allocation is made up of one Sub-Channel over duration of 4 OFDMA symbols. The first is a preamble and remaining are used for data transmission, adding more data symbols or subchannels increases the amount of data sent by the user, this allocation is presented in Figure 6:

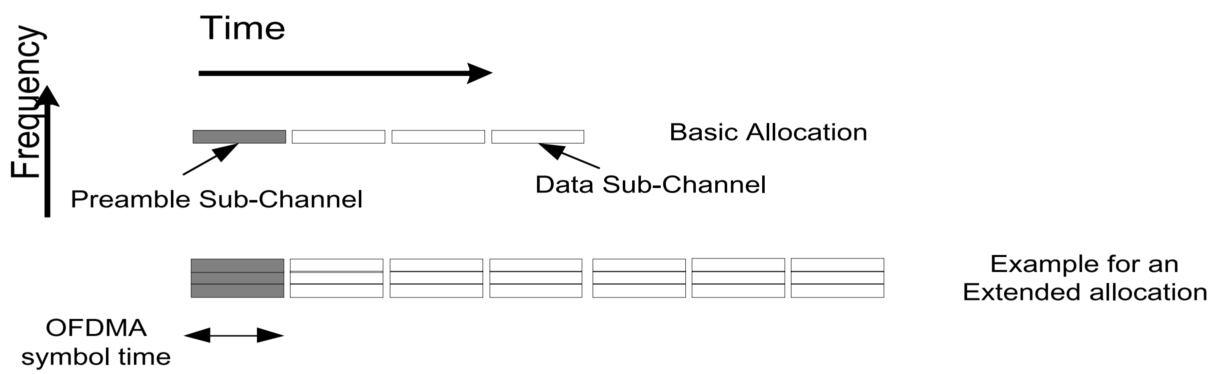

Figure 6: US User allocation

The framing structure used for the US includes the transmission of a possible symbol for Jamming monitoring, an allocation for Ranging and an allocation for data transmission. The MAC sets the length of the US framing, and the US mapping.

The framing for these modes involve the allocation of ranging Sub-Channels within the OFDMA symbols, while the rest of the Sub-channels are used for users transmission, the US mapping is illustrated in Figure 7:

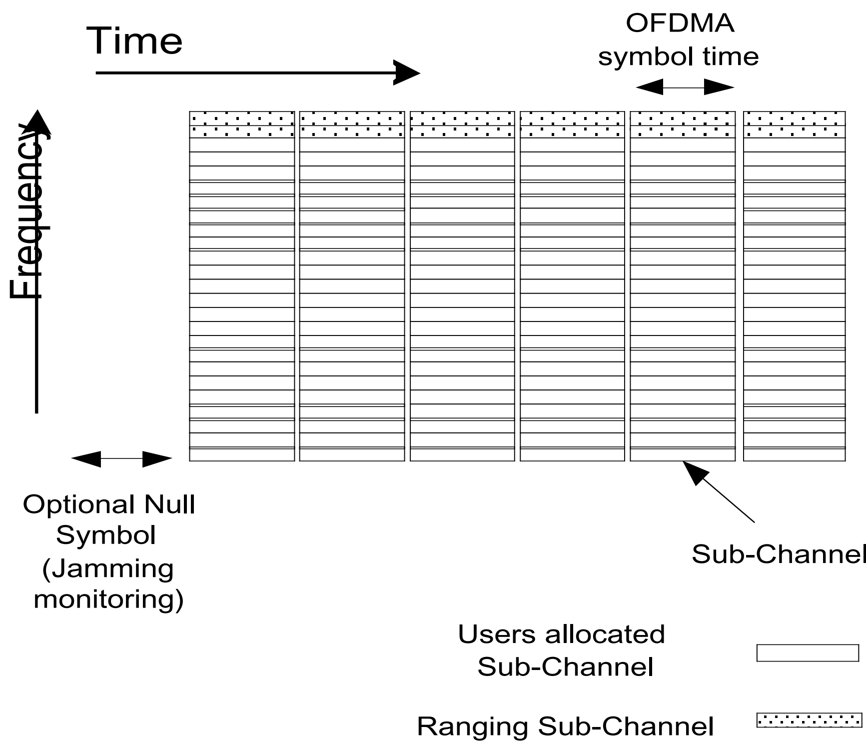

Figure 7: US framing for 1K FFT

3.1.3.2.3 Symbol Structure

The symbol structure is made up of Sub-Channels, by their basic structure described in section 3.1.3.2.1. There are several methods splitting the whole US OFDMA symbol into Sub-Channels, the first two methods are performed by first dividing the used carriers into basic groups (not including the DC carrier, which is not used), each containing a certain amount of carriers:

Then the following methods exist:

1. The number of basic groups is 53 ($\{N_{Groups}\}$=53) each comprising of 16 adjacent carriers, from the first usable carrier to the last. The sub-channels are extracted using the permutation method defined in 3.1.3.1.2 in which $\{N_{Sub-Channel}\}$=53.
2. The number of basic groups is 16 ($\{N_{Groups}\}$=16) each comprising of 53 adjacent carriers, from the first usable carrier to the last. Each basic group constitutes a sub-channel.

Parameter	Value	
N_{FFT}	1024 (1K)	
N_{used}	849	
Guard Carriers: Left, Right	88	87
Subchannels: nr, data carriers/subchannel	16	48
PermutationBase_0	{6, 14, 2, 3, 10, 8, 11, 15, 9, 1, 13, 12, 5, 7, 4, 0}	

Table 3: Upstream 1k parameters

Method 1 is the default method. Nico: Need some solution to switch between 1 and 2 here on network entry. They shouldn't be incompatible.

3.2. Time description

The Fourier transformed waveform is prepended with a circular extension of itself to form the guard-interval. The extension is expressed as a fraction of the transformed waveform. Fractions of 1/4, 1/8, 1/16, 1/32 must be provided with a minimum time-duration of 750 ns and a maximum time-duration of at most 6 μs. If a fraction provides a time-duration below 750ns or above 6 μs for a given FFT size and channelization, this fraction does not have to be implemented.

<table>
<tr><td>sampling</td><td colspan="2"></td><td>1</td><td>1 1/15</td><td>1 1/7</td></tr>
<tr><td>BW(MHz)</td><td colspan="2">FFT size</td><td>64</td><td>256</td><td>1024</td></tr>
<tr><td rowspan="7">20</td><td colspan="2">Δf (kHz)</td><td>312 1/2</td><td>83 1/3</td><td>22 9/28</td></tr>
<tr><td colspan="2">BWefficiency</td><td>82.81%</td><td>85.42%</td><td>94.87%</td></tr>
<tr><td colspan="2">Tb (us)</td><td>3 1/5</td><td>12</td><td>44 4/5</td></tr>
<tr><td rowspan="4">T_g / T_b</td><td>1/32</td><td></td><td></td><td>1 2/5</td></tr>
<tr><td>1/16</td><td></td><td>3/4</td><td>2 4/5</td></tr>
<tr><td>1/8</td><td></td><td>1 1/2</td><td>5 3/5</td></tr>
<tr><td>1/4</td><td>4/5</td><td>3</td><td></td></tr>
<tr><td rowspan="7">10</td><td colspan="2">Δf (kHz)</td><td>156 1/4</td><td>41 2/3</td><td>11 9/56</td></tr>
<tr><td colspan="2">BWefficiency</td><td>82.81%</td><td>85.42%</td><td>94.87%</td></tr>
<tr><td colspan="2">T_b (us)</td><td>6 2/5</td><td>24</td><td>90</td></tr>
<tr><td rowspan="4">T_g / T_b</td><td>1/32</td><td></td><td>3/4</td><td>2 4/5</td></tr>
<tr><td>1/16</td><td></td><td>1 1/2</td><td>5 3/5</td></tr>
<tr><td>1/8</td><td>4/5</td><td>3</td><td></td></tr>
<tr><td>1/4</td><td>1 3/5</td><td>6</td><td></td></tr>
<tr><td rowspan="7">5
(Optional)</td><td colspan="2">Δf (kHz)</td><td>78 1/8</td><td>20 5/6</td><td>5 47/81</td></tr>
<tr><td colspan="2">BWefficiency</td><td>82.81%</td><td>85.42%</td><td>94.87%</td></tr>
<tr><td colspan="2">T_b (us)</td><td>12 4/5</td><td>48</td><td>179 1/5</td></tr>
<tr><td rowspan="4">T_g / T_b</td><td>1/32</td><td></td><td>1 1/2</td><td>5 3/5</td></tr>
<tr><td>1/16</td><td>4/5</td><td>3</td><td></td></tr>
<tr><td>1/8</td><td>1 3/5</td><td>6</td><td></td></tr>
<tr><td>1/4</td><td>3 1/5</td><td></td><td></td></tr>
</table>

Table 4: Frequency and Time parameters

3.3. Frame and burst structure

In the PMP mode, the frame structure is build from BS (downstream) and SS (upstream) transmissions. Each burst transmission consists of one or more OFDM symbols. The cell radius is dependent on the time left open for random system access. This time should at least equal the maximum tolerable round trip delay plus the number of OFDM symbols necessary to transmit the ranging burst (see clause 3.3.2.1). Further, in each frame, the TX/RX transition gap (TTG) and RX/TX transition gap (RTG) need to be inserted between the downlink and uplink and at the end of each frame respectively to allow the BS to turn around (time plan for a single frame is shown in Figure 8).

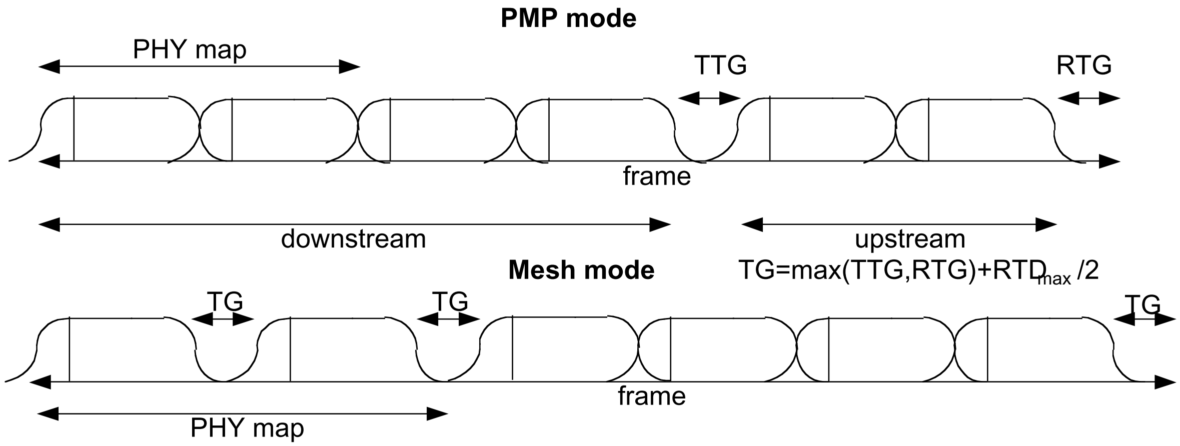

Figure 8: Frame structure

In the optional mesh mode, the frame structure is build from node transmissions. The cell radius is dependent on the Transition Gap (TG), which is inserted before every control packet and after the last control packet. Between data-transmissions of different nodes, no delay need be inserted. Random access uses the same method as the PMP mode. The number of control slots in the mesh mode shall be a network management variable.

3.3.1. Downlink and uplink bursts

The system has four different PHY bursts. These only apply to the mandatory FFT modes.

1) Downlink burst

   The downlink burst preamble consists of ten short OFDM symbols followed by a double duration cyclic prefix and two long OFDM symbols. The cyclic prefix is copied from the end of the second long OFDM symbol. In the downlink burst, between OFDM symbols of different modulation/coding, a midamble consisting of one long OFDM symbol is inserted.

2) Uplink burst with short preamble

   The uplink burst preamble consists of a double duration cyclic prefix followed by two long OFDM symbols. The cyclic prefix is copied from the end of the second long OFDM symbol.

3) Uplink burst with long preamble

   The downlink burst preamble consists of five short OFDM symbols followed by a double duration cyclic prefix and two long OFDM symbols. The cyclic prefix is copied from the end of the second long OFDM symbol.

4) Direct link burst (for optional mesh only)

   The downlink burst preamble consists of ten short OFDM symbols followed by a double duration cyclic prefix and two long OFDM symbols. The cyclic prefix is copied from the end of the second long OFDM symbol. In the downlink burst, between OFDM symbols of different modulation/coding, a midamble consisting of one long OFDM symbol is inserted.

3.3.2. OFDM Preambles — network entry, ranging, AGC, synchronization and equalization

3.3.2.1. Network entry

The base station shall allocate a number of symbols every few frames for network entry. This number of symbols shall be large enough to contain the maximum Round Trip Duration (RTD_{max}) plus a long preamble uplink burst with one OFDM symbol in data. A CPE attempting to enter the network shall listen to the base station until such a period is scheduled and send a network entry request using a long preamble uplink burst.

The power used for transmitting this burst shall be ($RSSI_{max}$ - $Rx_{sensitivity}$)/2 below de maximum possible output power, in which $RSSI_{max}$ is the maximum signal level received from the basestation and $Rx_{sensitivity}$ is the receiver sensitivity of the modulation to be used (BPSK, rate _). If this first attempt fails, the output power may be increased 2 dB per retry.

3.3.2.2. Time and Power Ranging of the users

During registration, a new subscriber registers during the random access channel and if successful is entered into a ranging process under control of the base station. The ranging process is cyclic in nature where default time and power parameters are used to initiate the process followed by cycles where (re)calculated parameters are used in succession until parameters meet acceptance criteria for the new subscriber. These parameters are monitored, measured and stored at the base station and transmitted to the subscriber unit for use during normal exchange of data. During normal exchange of data, the stored parameters are updated in a periodic manner based on configurable update intervals to ensure changes in the channel can be accommodated. The update intervals will vary in a controlled manner on a subscriber unit by subscriber unit basis.

Ranging on re-registration follows the same process as new registration. The purpose of the ranging parameter expiry is in support of portable applications capability. A portable subscriber unit s stored parameters will expire and are removed after the expiry intervals no longer consuming memory space and algorithm decision time.

This method is suitable for both OFDM and OFDMA.

3.3.2.2.1 Ranging using an OFDMA mapping

Time and Power ranging is performed by allocating several Sub-Channels to one Ranging Sub-Channel upon this Sub-Channels users are allowed to collide, each user randomly chooses a random code from a bank of codes. These codes are modulated by BPSK upon the contention Sub-Channel. The Base Station can then separate colliding codes and extract timing and power ranging information, in the process of user code detection the base station get the Channel Impulse Response (CIR) of the code, acquiring the base station vast information about the user channel and condition. The time and power ranging allows the system to compensate the far near user problems and the propagation delay caused by large cells.

The usage of the Sub-Channels for ranging is done by the transmission of a Pseudo Noise (PN) code on the Sub-Channel allocated for ranging transmission.The code is always BPSK modulated and is produced by the PRBS described in Figure 9 (the PRBS polynomial generator shall be $1+X^4+X^7+X^{15}$):

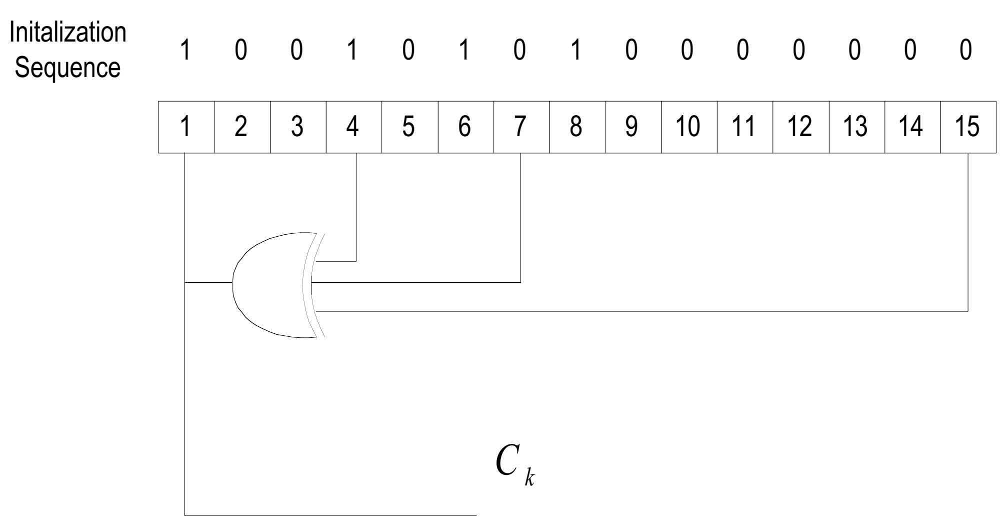

Figure 9: PRBS for ranging code production

Circulating through the PRBS (were each circulation produces one bit) produces the Ranging codes. The length of the ranging codes are multiples of 53 bits long (the default for the 1k mode is 1 Sub-Channels allocated as the ranging Sub-Channel therefore the ranging code length is 53), the codes produced are used for the next purposes:

- The first 16 codes produced are for First Ranging; it shall be used by a new user entering the system.
- The next 16 codes produced are used for maintenance Ranging for users that are already entered the system.
- The last 16 codes produced are for users, already connected to the system, issuing bandwidth requests.

These 48 codes are denoted as Ranging Codes and are numbered 0..47.
The MAC sets the number of Sub-Channels allocated for Ranging, these ranging Sub-Channels could be used concatenates as orders by the MAC in order to achieve a desired length.

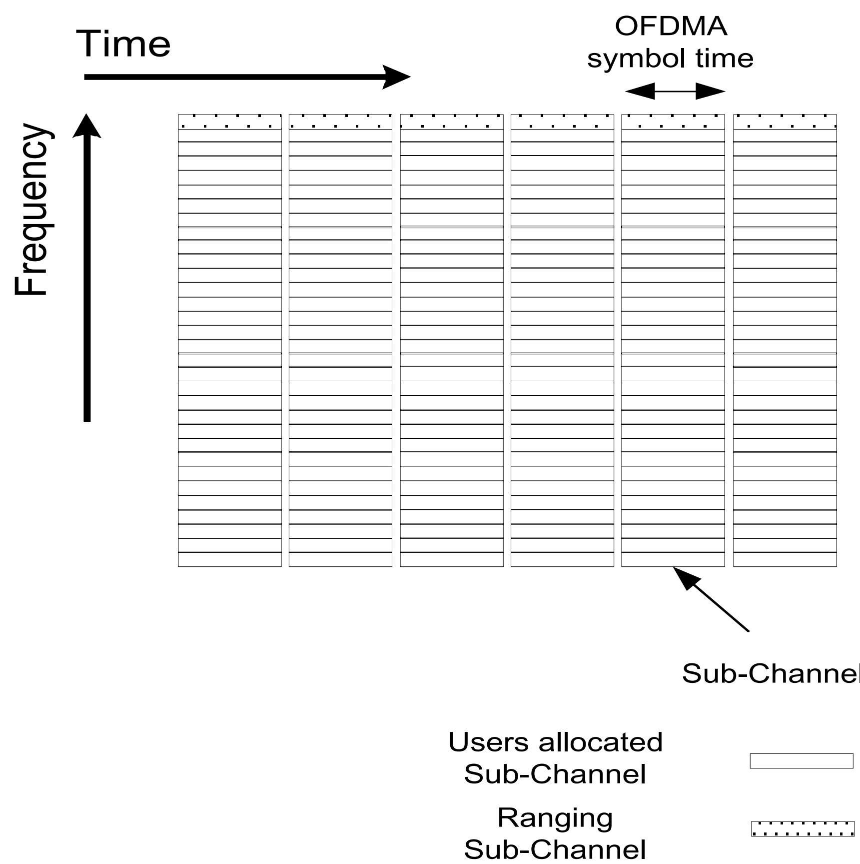

Figure 10: Ranging subchannel allocation for OFDMA mapping

3.3.2.2.1.1 Long Ranging transmission

The Long Ranging transmission shall be used by any SU that wants to synchronize to the system channel for the first time.
A Long Ranging transmission shall be performed during the two first consecutive symbols of the US frame.
Sending for a consecutive period of two OFDMA signals a preamble shall perform the long ranging transmission. The preamble structure is defined by modulating one Ranging Code, up on the Ranging Sub-Channel carriers. There shall not be any phase discontinuity on the Ranging Sub-Channel carriers during the period of the Long Ranging transmission.
This Long Ranging transmission is allowed only on the Ranging Sub-Channel resources defined by the MAC process in the Base Station.

3.3.2.2.1.2 Short Ranging transmission

The Short Ranging transmission shall be used only by a SU that has already synchronized to the system.
The Short Ranging transmission shall be used for system maintenance ranging or for fast bandwidth allocation requests.
To perform a Short Ranging transmission, the SU shall send a preamble for a period of one OFDM/OFDMA symbol in the duration of the ranging interval. The preamble structure is defined by modulating one Ranging Code on one Ranging Sub-Channel. This transmission may occur on any OFDM symbol out of the six available ranging symbols.
This Short Ranging transmission is allowed only on the Ranging Sub-Channel resources defined by the MAC process in the Base Station.

3.3.2.2.2 Ranging using an OFDM mapping

In the OFDM mapping regular uplink bursts shall be used for ranging. The only difference is that an extended header shall be used in order to allow resolving larger timing uncertainty, arising from the propagation delay in large cells.
Nico: I don't like the spreading code complexity for the OFDM modes. Given the number of CPE's per BS, the probability of colliding network entries is remote. Updating of the ranging can simply be done in the regular data symbols.

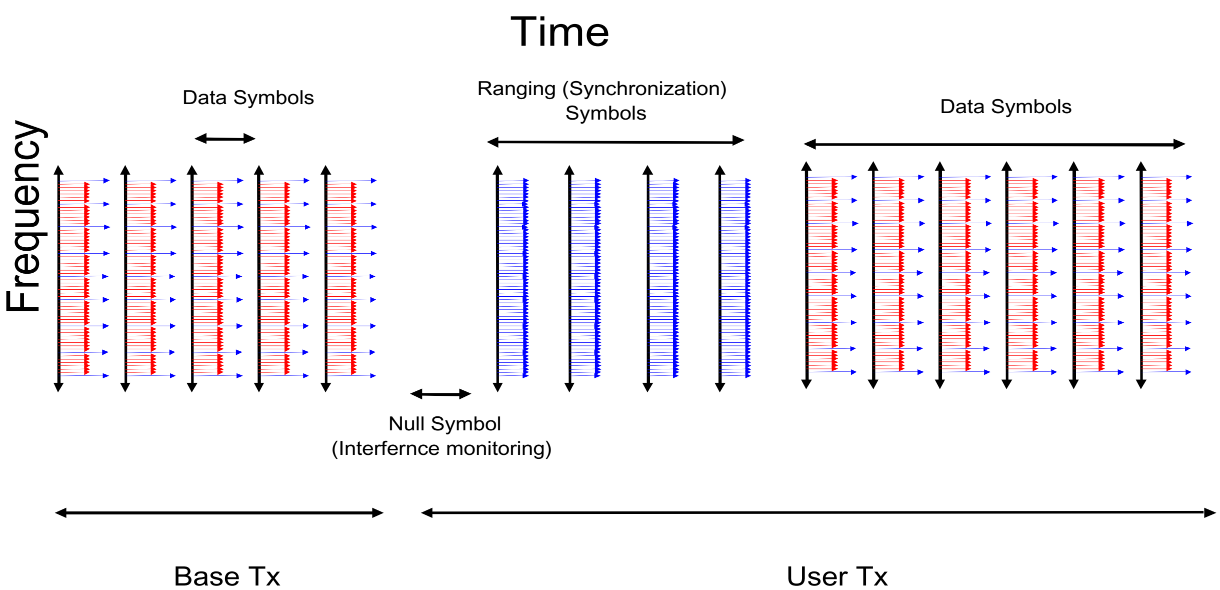

Figure 11: Ranging Symbol allocation for OFDM mapping

In the optional Mesh mode, the ranging and network entry shall be done in the control-slots.

3.3.2.3. Bandwidth request

The base station shall allocate a number of symbols every frame for bandwidth requests. This number of symbols shall be large enough to contain one or a multiple of long preamble uplink bursts with one OFDM symbol in data. CPEs requiring bandwidth may, using a backoff mechanism, use these slots to request bandwidth.

3.3.3. Bandwidth Requesting

3.3.3.1. Fast bandwidth requests using an OFDMA mapping

The usage of the Sub-Channels for fast bandwidth request is done by the transmission of a Pseudo Noise (PN) code on the Sub-Channel allocated for ranging transmission (see clause 3.3.2.2.1).

3.3.3.2. Bandwidth requests using an OFDM mapping

Bandwidth request in OFDM can take two forms.

- Contention based requests. In this mode regular uplink bursts shall be used for BW requests using an uplink burst with long preamble (see 3.3.1).
- Subcarrier based polling. In this mode, the BTS may poll a group of stations, and each station responds by issuing an energy on a small set of subcarriers determined by station ID, if it desires to respond to a poll. Nico: Don't like this, prefer to get rid of it.

4. **Data encoding** (Octavian- Brian, Editing Team)

Data encoding is composed of three steps: randomizer, FEC and interleaving. They shall be applied in this order at transmission. The complementary operations shall be applied in reverse order at reception.

Note: Two randomizer schemes and four FEC — interleaving combinations have been proposed. They are all described separately in this section until the group decides on main and optional encoding schemes.

4.1. Data randomizer (scrambler)

Notes:

- *Used word randomizer instead of scrambler to distinguish between whitening and security.*
- *Two methods were proposed: one as in 802.11a and one as in DVB*

4.1.1. Method 1 — 802.11a type scrambler

The data bytes to be transmitted are converted to a serial bit stream using the LSB first MSB last rule. The serial bit stream shall be passed through a data randomizer that uses the generator polynomial $R(X) = X^7 + X^4 + 1$ and is depicted in Figure 12. The same randomizer is used to scramble data at transmission and de-scramble received data.

In order to avoid retransmission of the same frame with the same initial state of the scrambler, the initial state of the scrambler will be set to a pseudo random non-zero state, for each transmission (burst). This can be easily achieved by preserving the last state from the previous transmission or by letting scrambler operate between transmissions. The first seven bits in the data field *(equivalent to first 7 bits in the SERVICE field in 802.11a)* will be set to all zeros prior to scrambling in order to enable estimation of the initial state of the de-scrambler in the receiver at reception.

Notes:

- *16 bits reserved for the SERVICE field in 802.11a were reduced to 7 used to initialize the scrambler.*
- *Random initialization of the scrambler avoids retransmission of the OFDM symbols that have high peak to average power ratio but very small probability to occur. This problem cannot be fixed using fixed initialization.*
- *Random initialization requires transmission of 7additional bits.*

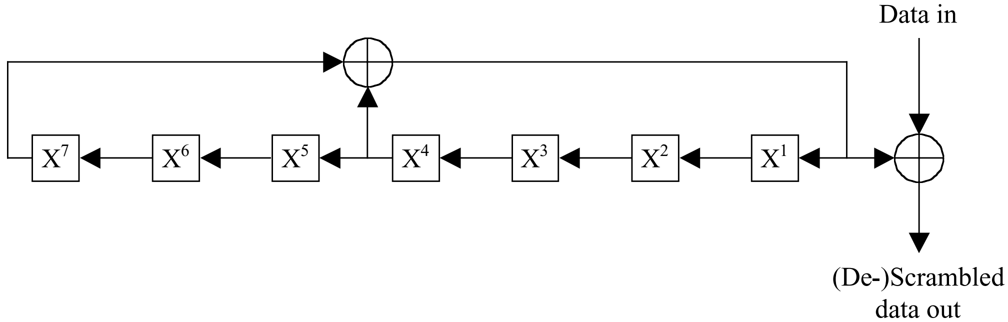

Figure 12: Data Randomizer

4.1.2. Method 2 — as in DVB

Data randomization is performed on data transmitted on the downlink (DL) and uplink (UL). The randomization is performed on each allocation (DL or UL), which means that for each allocation of a data block (Sub-Channels on the frequency domain and OFDM symbols on the time domain) the randomizer shall be used independently. If the amount of data to transmit those not fit exactly the amount of data allocated, padding of FFx (1 only) shall be added to the end of the transmission block, up to the amount of data allocated.

The shift-register of the randomizer shall be initialized for each new allocation or for every 1250 bytes passed through (if the allocation is larger then 1250 bytes).

The randomizer shall be initialized with the binary value: 100101010000000 (45200 in octal). Each data byte to be transmitted shall enter sequentially into the randomizer, MSB first.
The Pseudo Random Binary Sequence (PRBS) generator shall be $1+X^{14}+X^{15}$.

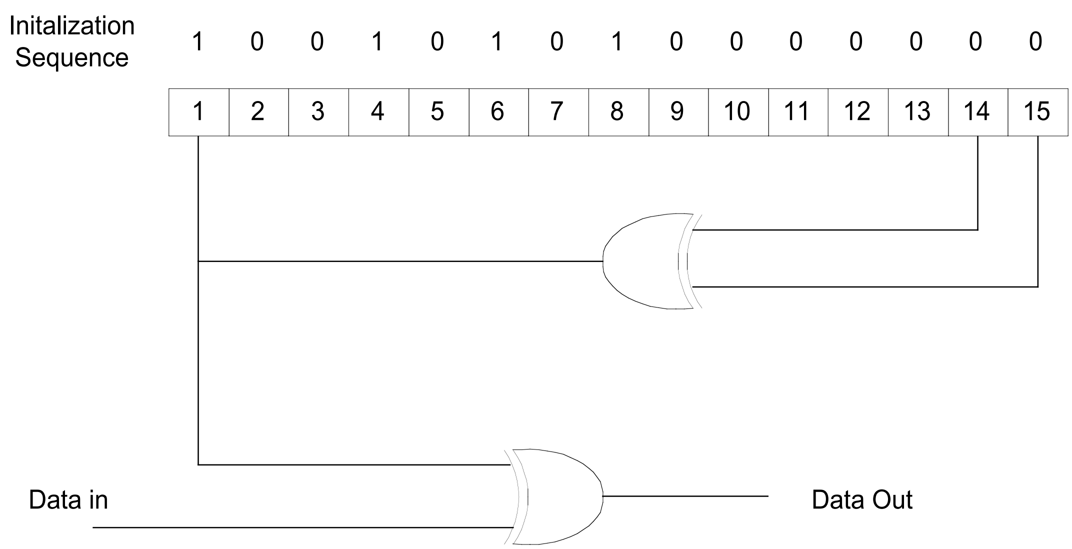

Figure 13: PRBS for data randomization

4.2. Forward error correction and interleaving

Note: There were four FEC and interleaving methods proposed. They are listed below.

4.2.1. Method 1: Convolutional encoder and symbol-size interleaving

This combination of FEC and interleaving consists in a standard convolutional encoder and a bit interleaver with a block size equal to one OFDM symbol.

4.2.1.1. Convolutional encoder

The serial data from randomizer shall be coded with a convolutional encoder of coding rate R=1/2, 2/3 or 3/4 corresponding to the desired data rate. The convolutional encoder shall use the industry standard generator polynomials $g_0=133_8$ and $g_1=171_8$ of rate R=1/2 as shown in Figure 14. Bit denoted as A shall output from the encoder before the bit denoted as B . Rates of 2/3 and 3/4 shall be derived by employing puncturing , i.e. by omitting some of the encoded bits in the transmitter based on known patterns. The puncturing patterns are illustrated in Figure 15 for R=3/4 and Figure 16 for R=2/3. The rate R=1/2 is illustrated in Figure 17. In the receiver, decoding by the Viterbi algorithm is recommended.

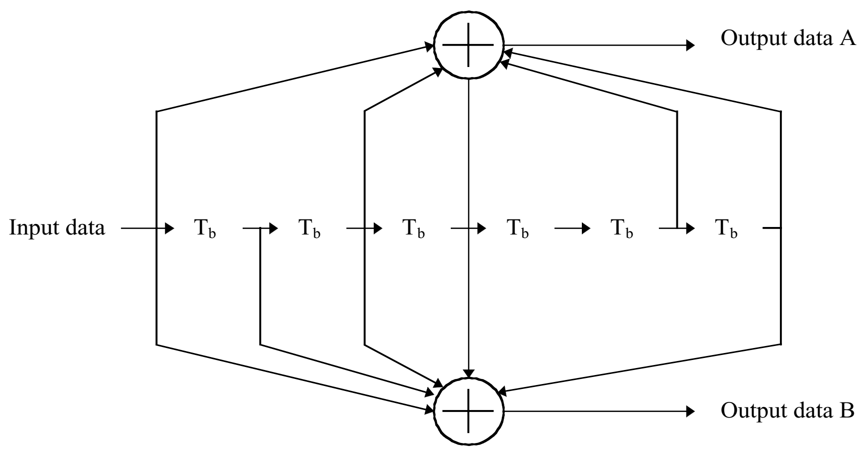

Figure 14: Convolutional encoder

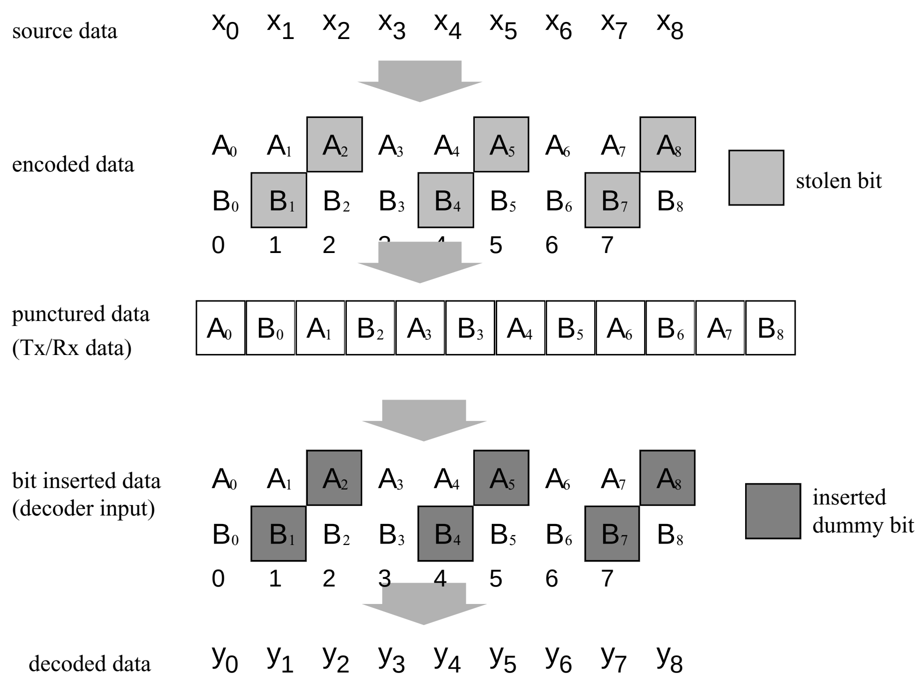

Figure 15: Puncturing for R=3/4, transmission and reception

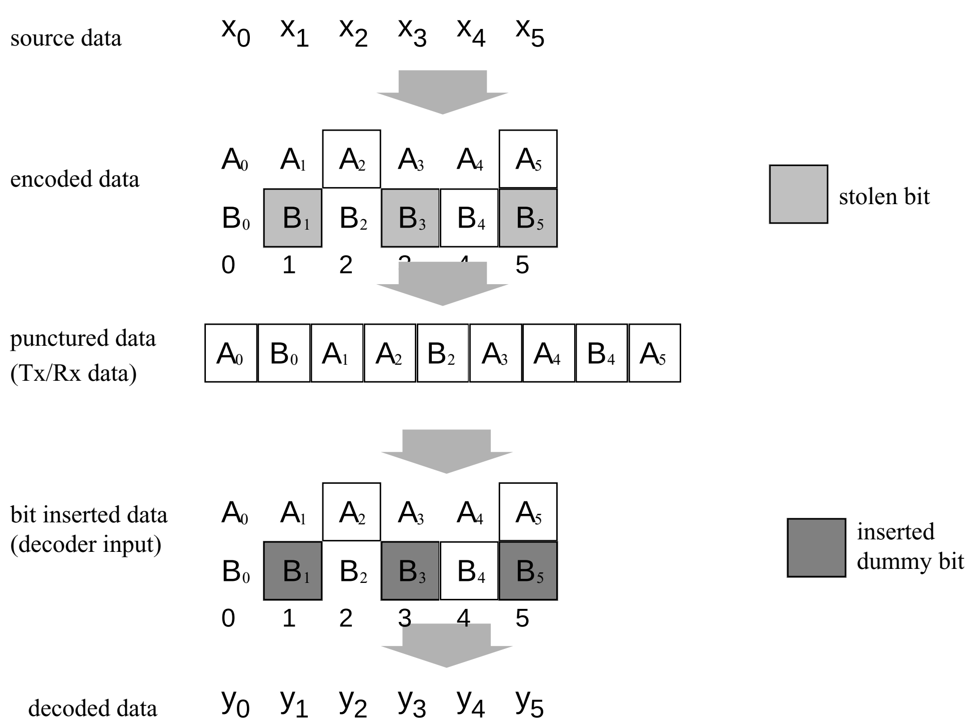

Figure 16: Puncturing for R=2/3, transmission and reception

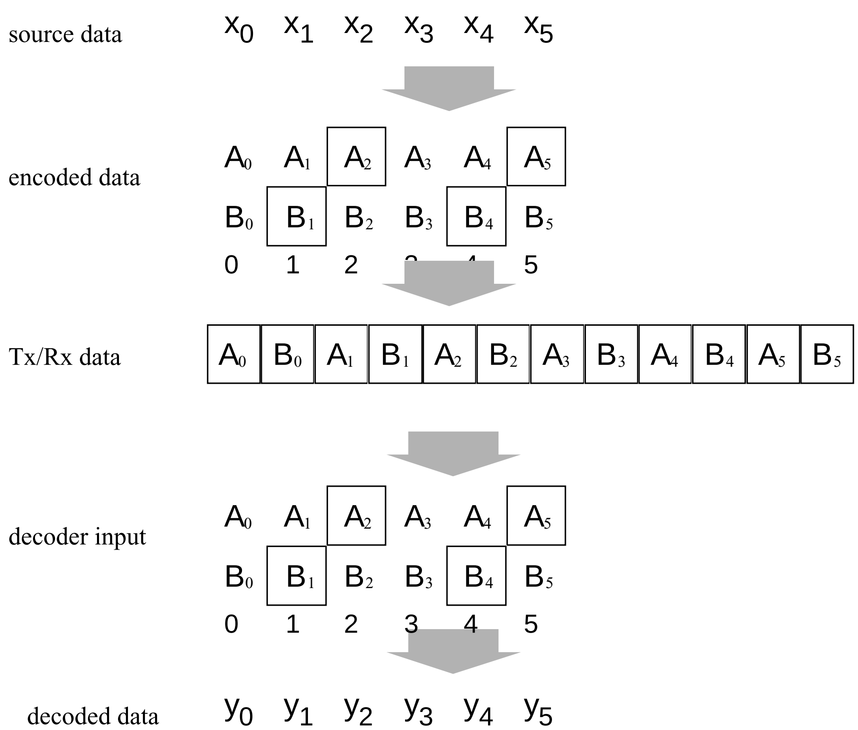

Figure 17: Bit ordering for R=1/2, transmission and reception

4.2.1.2. Symbol-size bit interleaving

All encoded data bits shall be interleaved by block interleaver with a block size corresponding to the number of coded bits per OFDM symbol, N_{CBPS}. The interleaver is defined by a two step permutation. The first ensures that adjacent coded bits are mapped onto nonadjacent sub-carriers. The second permutation insures that adjacent coded bits are mapped alternately onto less or more significant bits of the constellation, thus avoiding long runs of lowly reliable bits (LSB).

Let N_{BPSC} be the number of bits per sub-carrier, i.e. 1, 2, 4 or 6 for BPSK, QPSK, 16QAM ro 64QAM, respectively. Let $s = \max(N_{BPSC}/2,1)$. Let k be the index of the coded bit before the first permutation at transmission, m be the index after the first and before the second permutation and j be the index after the second permutation, just prior to modulation mapping.

The first permutation is defined by the rule:

$$m = (N_{CBPS}/16)\,(k \bmod 16) + \mathrm{floor}(k/16) \qquad k = 0, 1, \ , N_{CBPS}\text{-}1$$

The second permutation is defined by the rule:

$$j = s * \mathrm{floor}(m/s) + (m + N_{CBPS} — \mathrm{floor}(16 * m/N_{CBPS})) \bmod s \qquad m = 0, 1, \ , N_{CBPS}\text{-}1$$

The deinterleaver, which performs the inverse operation, is also defined by two permutations. Let j be the index of the received bit before the first permutation, m be the index after the first and before the second permutation and k be the index after the second permutation, just prior to delivering the coded bits to the convolutional decoder.

The first permutation is defined by the rule:

$$m = s * \mathrm{floor}(j/s) + (j + \mathrm{floor}(16 * j/N_{CBPS})\ ) \bmod s \qquad j = 0, 1, \ , N_{CBPS}\text{-}1$$

The second permutation is defined by the rule:

$$k = 16 * m - (N_{CBPS}-1) * \text{floor}(16 * m/N_{CBPS}) \qquad m = 0, 1, \ , N_{CBPS}-1$$

The first permutation in the deinterleaver is the inverse of the second permutation in the interleaver, and conversely.

Note: Both permutations can be applied as they are for any number of sub-carriers provided that N_{CBPS} is a multiple of 16. This shall be valid for all mandatory and optional modes.

4.2.2. Method 2 — Concatenated Reed-Solomon and bit-tailing convolutional

Code rates of 1/2, 3/4 for BPSK, QPSK and 16QAM, and 2/3, 3/4 for 64QAM are required. These coding rates shall be implemented using concatenated Reed Solomon and Convolutional codes as shown in Table 5

Modulation	Over-All Coding Rate	RS Code Rate	Convolutional Code Rate
BPSK,QPSK, 16QAM	1/2	3/4	2/3
64 QAM	2/3	8/9	3/4
BPSK,QPSK, 16QAM 64 QAM	3/4	9/10	5/6

Table 5: Coding modes

The Reed-Solomon-Convolutional coding rate 1/2 shall be used as the coding mode when requesting access to the network.
First passing the data through the RS encoder and then passing the data in block format to a tail biting convolutional encoder perform the encoding.

4.2.2.1. Reed Solomon encoding

Note: *Probably need to revise the RS(N,K,T) used, it doesn't seem to divide very well with multiples of 9 bytes data blocks. Also requires details as how to get the specified RS rates.*

The Reed Solomon encoding process shall use the systematic RS(255,239,8), with the possibility to make a variable RS(N,K,T), where:

- N - overall bytes, after encoding
- K - data bytes before encoding
- T - data bytes that can be fixed

The following polynomials are used for the systematic code:

- Code generator polynomial: $g(x) = (x+\lambda^0)(x+\lambda^1)(x+\lambda^2)...(x+\lambda^{2T-1}), \lambda = 02_{hex}$
- Field Generator polynomial: $p(x) = x^8 + x^4 + x^3 + x^2 + 1$

4.2.2.2. Convolutional encoding

Data bits issued from the Reed Solomon encoder, described in clause Figure 18, shall feed the convolutional encoder depicted in Figure 18.

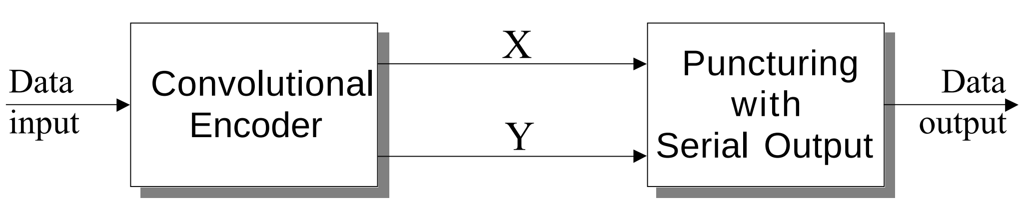

Figure 18: Convolutional encoder block diagram

The Convolutional encoder shall have a constraint length equal to k=7 and shall use the following mother codes:

$$G_1 = 171_{oct} \text{ For X}$$

$$G_1 = 133_{oct} \text{ For Y}$$

A basic convolutional encoding scheme, as depicted in Figure 19, shall be used.

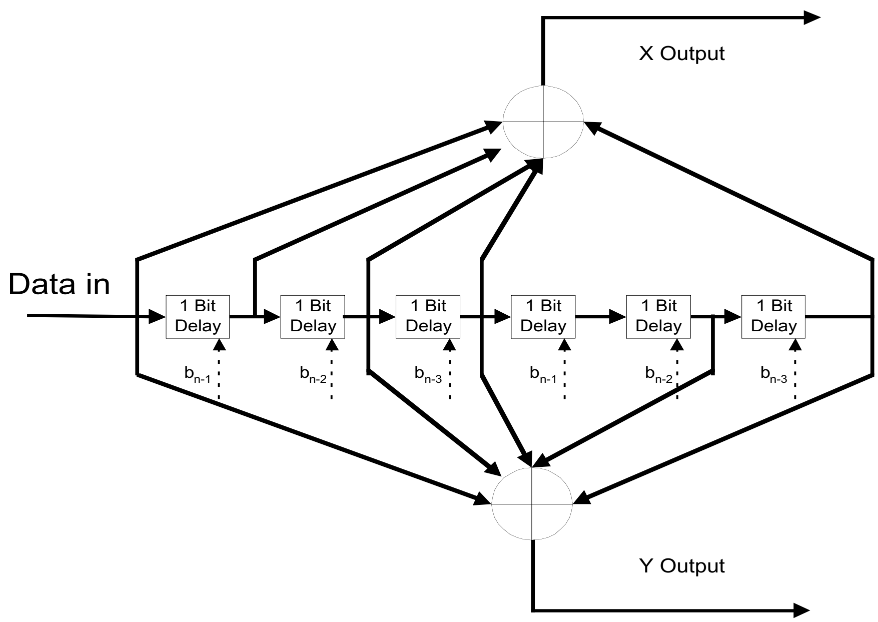

Figure 19: Convolutional encoder basic scheme

The puncturing pattern shall be as defined in Table 6.

CC Code Rate	Puncturing Pattern	Transmitted Sequence (after parallel to serial conversion)
$2/3$	X : 1 0 Y : 1 1	$X_1Y_1Y_2$
$3/4$	X : 1 0 1 Y : 1 1 0	$X_1Y_1Y_2X_3$
$5/6$	X : 1 0 1 0 1 Y : 1 1 0 1 0	$X_1Y_1Y_2X_3Y_4X_5$

Table 6: Puncturing patterns

4.2.2.2.1 Tail Biting Code Termination

In order to allow sharing of the ECC decoder, each of the multiple data streams subdivides its data into RS blocks. In this mode, each RS block is encoded by a tail-biting convolutional encoder. In order to achieve a tail biting convolutional encoding the memory of the convolutional encoder shall be initialized with the last data bits of the RS packet (the packet data bits are numbered $b_0..b_n$).

4.2.2.3. Block Lengths and Concatenated Code Rates

Table XX (reference to the next table) gives the block sizes and the code rates used for the different modulations and code rates:

Modulation	Block Size (Bytes)	Over-All Coding Rate	RS Coding	CC Coding Rate
QPSK	8	~1/2	(12,8,2)	2/3
QPSK	13	~3/4	(15,13,1)	5/6
QPSK	18	1/2	(24,18,3)	2/3
QPSK	26	~3/4	(30,26,2)	5/6
16QAM	36	1/2	(48,36,6)	2/3
16QAM	54	3/4	(60,54,3)	5/6
64QAM	72	2/3	(81,72,4)	3/4
64QAM	82	~3/4	(90,82,4)	5/6

Table 7: RS and CC coding rates as function of modulation

4.2.2.4. Bit interleaving

The encoded data encoded is passed through a bit interleaver with a nominal size of 3 Sub-Channel allocation units for 1024-FFT mode (OFDMA). When using the 256 FFT the interleaver size will be of 92 symbols, and it will be used twice for each OFDM symbol. . The interleaver size depends on the modulation used. Table 8 summarizes the bit-interleaver sizes as a function of the modulation and coding.

Modulation	OFMDA (FFT 1024)	OFDM (FFT 256)
BPSK	144	92
QPSK	288	184
QAM16	576	368
QAM64	864	552

Table 8: Bit-Interleaver size as a function of the Modulation

The interleaver scrambles the order of the input bits to produce the interleaved data, which is then fed to the mapper and the Sub-Channel allocation for OFDMA.

The PRBS generator depicted in Figure 20 is used to achieve the bit interleaver array, it is initialized with the binary value: 0001011010.
The PRBS generator produces an index value, which shall correspond to the new position of the input bit into the output interleaved data burst.
The interleaver shall use the following algorithm:

- The Interleaver indexes range from 1 to n (where n denotes the block size to be interleaved)
- For each input bit, the PRBS shall be rotated, the rotation produces a number, which is the value of the PRBS memory register.
- If the obtained number is bigger than n, it shall be discarded and the PRBS shall be rotated again. The rotation shall continue until an index between 1 to n is produced.
- The obtained index shall be used to address the position of the processed bit into the output interleaved data burst

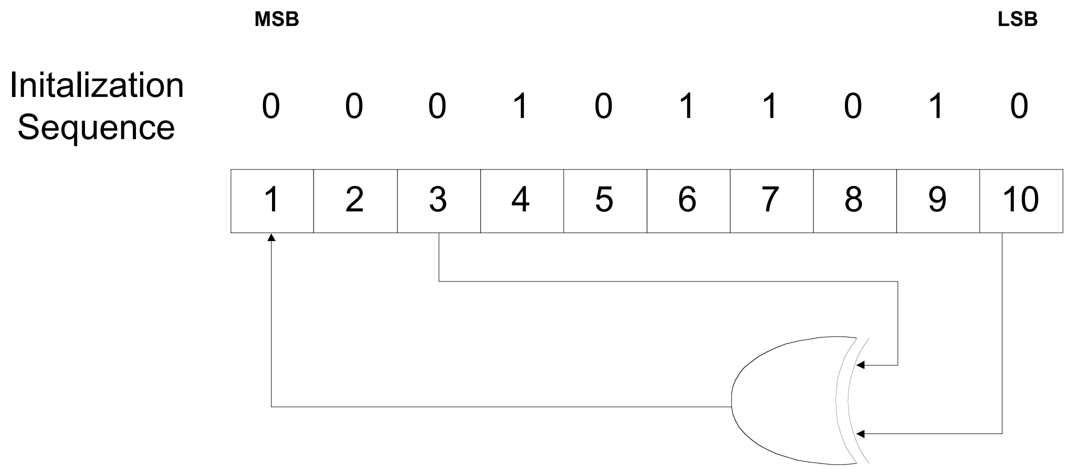

Figure 20: PRBS for Bit-Interleaver array

This method consists in three stages: Reed-Solomon encoder, convolutional encoder with tail biting and a bit interleaver with a size of 3 OFDM symbols (valid for 64 and 1024 —FFT modes; TBD for 256-FFT mode).

4.2.3. Method 3 - Block Turbo Coding

4.2.3.1. Proposed Block Turbo Codes

This type of coding based on the product of two or more simple component codes, is also called Turbo Product code, TPC. The decoding is based on the concept of Soft-in/Soft-out (SISO) iterative decoding (i.e, Turbo decoding). The component codes recommended for this proposal are binary extended Hamming codes or Parity check codes. The schemes supported follow the recommendation of the IEEE802.16.1 mode B. However, more flexibility in block size and code rates is enabled. The main benefits of using TPC mode are typically 2dB better performance over the Concatenated RS, and shorter decoding delays. A detailed description of Turbo Product Codes is included as Section 4.2.3.2. In this Section we present some particular turbo product codes that are perfectly matched for the proposed framing/modulation structure.

4.2.3.1.1 Block Turbo Constituent Codes

As mentioned in Section 4.2.3.2, TPCs are constructed as a product of simple component codes. The complete constituent code set is defined in Table 9.

(64, 57) Extended Hamming Code
(32,26) Extended Hamming Code
(16,11) Extended Hamming Code
(32,31) Parity Check Code
(16,15) Parity Check Code
(4,3) Parity Check Code

Table 9: Constituent Block Turbo Code List

4.2.3.1.2 Overall Turbo Product Codes

The defined Turbo Product Codes are all multiples of 48 bits to facilitate integration into the framing structure. Table 10 lists the possible codes, rates and block size in bits. The block sizes are achieved by bit shortening as described in Section 4.2.3.2.

X-Code	Y-Code	Z-Code	Rate	Block Size
56,49	55,48		0.76	3072
25,19	25,19	5,4	0.46	3072
48,41	48,41		0.72	2304
24,18	24,18	4,3	0.42	2304
64,57	21,20		0.85	1296
41,34	32,26		0.67	1296
64,57	12,11		0.82	768
28,22	28,22		0.62	768
54,47	9,8		0.77	480
30,24	16,11		0.55	480
32,26	6,5		0.68	192
14,9	14,9		0.41	192

Table 10: TPC Example Codes

The codes may be shortened using the method described in Section 4.2.3.2. Shortening should only be performed in multiples of 48 bits.

When using Turbo Product Codes additional Bit Interleaver sizes are defined as in Table 11.

X-Code	Y-Code	Z-Code	Bit Interleaver allocation
56,49	55,48		1024
25,19	25,19	5,4	1024
48,41	48,41		768
24,18	24,18	4,3	768
64,57	21,20		648
41,34	32,26		648
64,57	12,11		768
28,22	28,22		768
54,47	9,8		480
30,24	16,11		480
32,26	6,5		192
14,9	14,9		192

Table 11: Optimal Bit Interleaver Sizes

Termination of allocations does not have to be the multiplications of the above bit interleaver sizes. Sizes in any multiple of 48, up to a multiple of 12 can be used to terminate the allocation.

4.2.3.2. Turbo Code Description

The Block Turbo Code is a Turbo decoded Product Code (TPC). The idea of this coding scheme is to use well-known product codes in a matrix form for two-dimensional coding, or in a cubical form for three dimensions.
The matrix form of the two-dimensional code is depicted in Figure 21. The k_x information bits in the rows are encoded into n_x bits, by using a binary block (n_x, k_x) code. The binary block codes employed are based on extended Hamming codes.

The redundancy of the code is $r_x = n_x - k_x$ and d_x is the Hamming distance. After encoding the rows, the columns are encoded using another block code (n_y, k_y), where the check bits of the first code are also encoded. The overall block size of such a product code is $n = n_x \times n_y$, the total number of information bits $k = k_x$ « $k_y$and the code rate is $R = R_x \times R_y$, where $R_i = k_i/n_i$, i=x, y. The Hamming distance of the product code is $d = d_x \times d_y$.

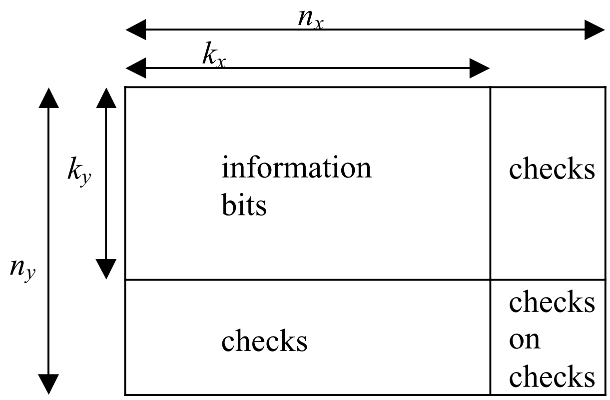

Figure 21: Two-dimensional product code matrix

4.2.3.2.1 Encoding of a Turbo Product Code

The encoder for TPCs has near zero latency, and is constructed of linear feedback shift registers (LFSRs), storage elements, and control logic. Encoding of a product code requires that each bit be encoded by 2 or 3 codes.
The constituent codes of TPCs are extended Hamming or parity only codes. Table 12 gives the generator polynomials of the Hamming codes used in TPCs. For extended Hamming codes, an overall even parity check bit is added at the end of each codeword.

N	k	Generator Polynomial
7	4	$x^3 + x + 1$
15	11	$x^4 + x + 1$
31	26	$x^5 + x^2 + 1$
63	57	$x^6 + x + 1$
127	120	$x^7 + x^3 + 1$
255	247	$x^8 + x + 1$

Table 12: Generators Polynomials of Hamming Codes:

In order to encode the product code, each data bit is input both into a row encoder and a column encoder. Only one row encoder is necessary for the entire block, since data is input in row order. However, each column of the array is encoded with a separate encoder. Each column encoder is clocked for only one bit of the row, thus a more efficient method of column encoding is to store the column encoder states in a $k_x \times (n_y - k_y)$ storage memory. A single encoder can then be used for all columns of the array. With each bit input, the appropriate column encoder state is read from the memory, clocked, and written back to the memory.

The encoding process will be demonstrated with an example.

4.2.3.2.2 Example of a 2-Dimesional Product Code

Assume a two-dimensional (8,4)×(8,4) extended Hamming Product code is to be encoded. This block has 16 data bits, and 64 total encoded bits. Figure 22 shows the original 16 data bits denoted by D_{yx}. Of course the usual way is to have a serial stream of data of 16 bits and then label them as D_{11}, D_{21}, D_{31}, D_{41}, D_{12}, , D_{44}.

D_{11}	D_{21}	D_{31}	D_{41}
D_{12}	D_{22}	D_{32}	D_{42}
D_{13}	D_{23}	D_{33}	D_{43}
D_{14}	D_{24}	D_{34}	D_{44}

Figure 22: Original Data for Encoding

The first four bits of the array are loaded into the row encoder in the order D_{11}, D_{21}, D_{31}, D_{41}. Each bit is also fed into a unique column encoder. Again, a single column encoder may be used, with the state of each column stored in a memory. After the fourth bit is input, the first row encoder error correction coding (ECC) bits are shifted out.

This process continues for all four rows of data. At this point, 32 bits have been output from the encoder, and the four column encoders are ready to shift out the column ECC bits. This data is also shifted out row-wise. This continues for the remaining 3 rows of the array. Figure 23 shows the final encoded block with the 48 generated ECC bits denoted by E_{yx}.

D_{11}	D_{21}	D_{31}	D_{41}	E_{51}	E_{61}	E_{71}	E_{81}
D_{12}	D_{22}	D_{32}	D_{42}	E_{52}	E_{62}	E_{72}	E_{82}
D_{13}	D_{23}	D_{33}	D_{43}	E_{53}	E_{63}	E_{73}	E_{83}
D_{14}	D_{24}	D_{34}	D_{44}	E_{54}	E_{64}	E_{74}	E_{84}
E_{15}	E_{25}	E_{35}	E_{45}	E_{55}	E_{65}	E_{75}	E_{85}
E_{16}	E_{26}	E_{36}	E_{46}	E_{56}	E_{66}	E_{76}	E_{86}
E_{17}	E_{27}	E_{37}	E_{47}	E_{57}	E_{67}	E_{77}	E_{87}
E_{18}	E_{28}	E_{38}	E_{48}	E_{58}	E_{68}	E_{78}	E_{88}

Figure 23: Encoded Block

Transmission of the block over the channel may occur in a linear fashion, for example with all bits of the first row transmitted left to right followed by the second row, etc. This allows for the construction of a near zero latency encoder, since the data bits can be sent immediately over the channel, with the ECC bits inserted as necessary. For the (8,4)«(8,4) example, the output order for the 64 encoded bits would be

D_{11}, D_{21}, D_{31}, D_{41}, E_{51}, E_{61}, E_{71}, E_{81}, D_{12}, D_{22}, , E_{88}.

Alternatively, a block-based interleaver may be inserted to further improve the performance of the system.

4.2.3.2.3 3-Dimensional TPC Encoding

For a three-dimensional TPC block, the element ordering for input/output for both encoding and decoding is usually in the order of rows, columns and then the z-axis. If we consider a serial stream of (i×j×k) data bits, labeled as:

$D_{1,1,1}$, $D_{2,1,1}$, $D_{3,1,1}$, , $D_{i,1,1}$, $D_{1,2,1}$, $D_{2,2,1}$, , $D_{,i,j,1}$,$D_{1,1,2}$, , $D_{i,j,k}$.

Note: this labeling is for convenience

Then the total size of the encoded block is ((i×j×k) + ECC bits), where there are p ECC bits for the x-axis, q ECC bits for the y-axis and r ECC bits for the z-axis, the bit order for input and output is:

$D_{1,1,1}$, $D_{2,1,1}$, $D_{3,1,1}$, , $D_{i,1,1}$, , $E_{p,1,1}$, $D_{1,2,1}$, $D_{2,2,1}$, , $E_{p,2,1}$, , $E_{p,q,1}$, $D_{1,1,2}$, $D_{2,1,2}$, , $E_{p,1,2}$, , $E_{p,q,2}$, , $E_{p,q,r}$

This is shown in Figure 24.

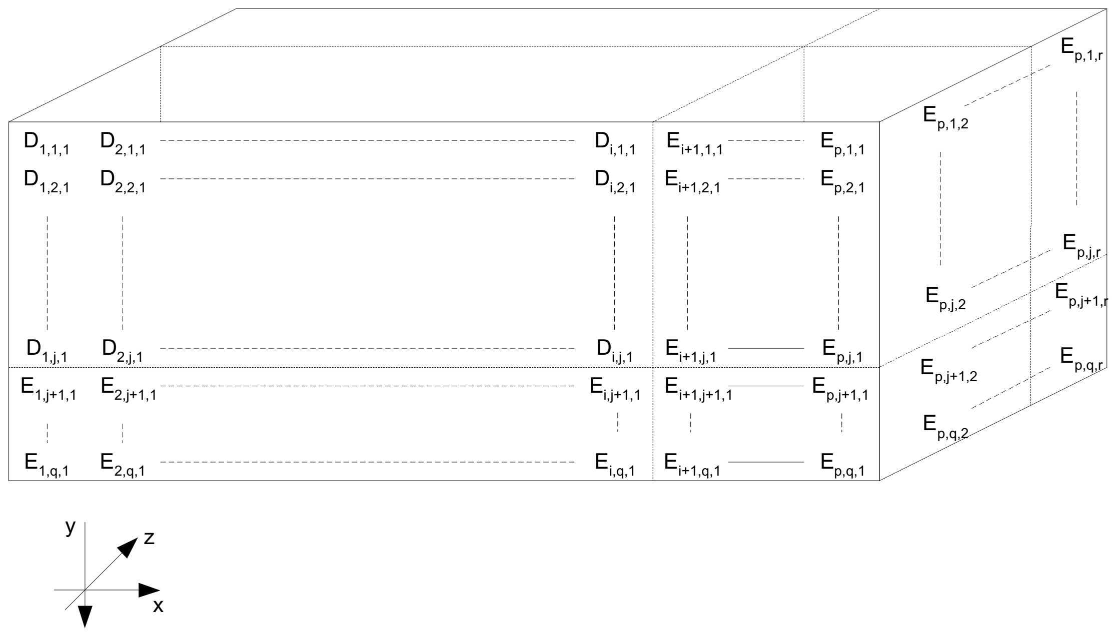

Figure 24: Structure of 3-Dimensional TPC

Notation:

- the codes defined for the rows (x-axis) are binary (n_x,k_x) block codes
- the codes defined for the columns (y-axis) are binary (n_y,k_y) block codes
- the codes defined for the z-dimension (z-axis) are binary (n_z,k_z) block codes
- data bits are noted $D_{y,x,z}$ and parity bits are noted $E_{y,x,z}$

4.2.3.2.4 Shortened TPCs

To match packet sizes, a product code may be shortened by removing symbols from the array. In the two-dimensional case rows, columns or parts thereof can be removed until the appropriate size is reached. Unlike one-dimensional codes (such as Reed-Solomon codes), parity bits are removed as part of shortening process, helping to keep the code rate high.

There are two steps in the process of shortening of product codes. The first is to remove an entire row or column from a 2-dimensional code, or an entire X, Y, or Z plane from a 3-dimensional code. This is equivalent to shortening the constituent codes that make up the product code. This method enables a coarse granularity on shortening, and at the same time maintaining the highest code rate possible by removing both data and parity symbols. Further shortening is obtained by removing individual bits from the first row of a 2-dimensional code, or from the top plane of a 3-dimensional code.

4.2.3.2.5 Example of a Shortened 2-Dimensional TPC

For example, assume a 456-bit block size is required with a code rate of approximately 0.6. The base code chosen before shortening is the (32,26)×(32,26) code which has a data size of 676 bits. Shortening all rows by 5 bits and all columns by 4 bits results in a (27,21)×(28,22) code, with a data size of 462 bits. To get the exact block size, the first row of the product is shortened by an additional 6 bits. The final code is a (750,456) code, with a code rate of 0.608. Figure 25 shows the structure of the resultant block.

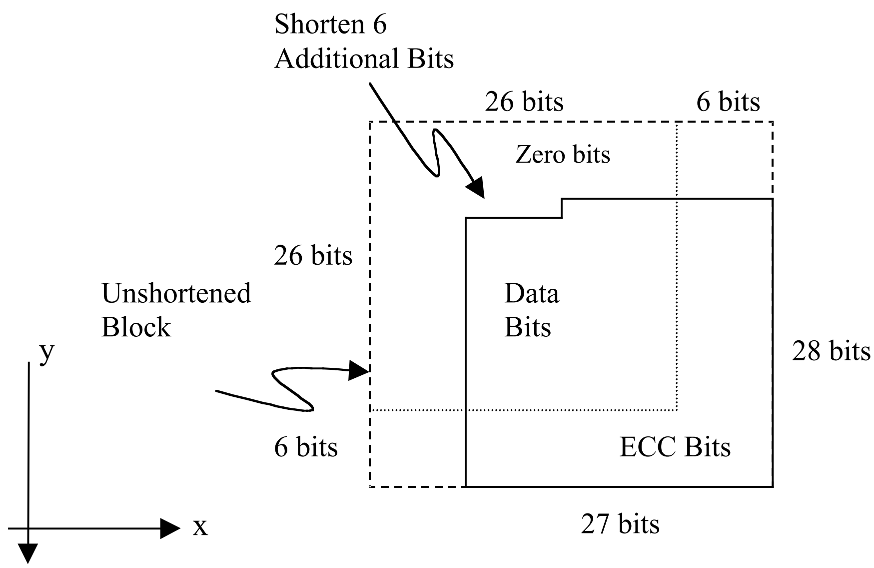

Figure 25: Structure of Shortened 2 D Block

Modifications to the encoder to support shortening are minimal. The shortening procedure is trivial, and yet an extremely powerful tool that enables construction of a very versatile code set.

4.2.3.2.6 Example of a Shortened 3-Dimensional TPC

Suppose a 0.4 - 0.45 rate code is required with a data block size of 1096 bits. The following shows one possible method to create this code.

Start with a (32,26)×(32,26)×(4,3) code. The optimum shortening for this code is to remove rows and columns, while leaving the already very short z-axis alone. Therefore, since a 1096 bit 3-Dimensional code is required, the desired vector data size can be found by taking the square root of 1096/3 and rounding up. This yields a row/column size of about 20. In fact, having a row size of 20, a column size of 19, and a z-column size of 3 gives the closest block size to 1096 bits.

The code size is now a (26,20)×(25,19)×(4,3) = (2600,1140). To get the exact data size, we further shorten the first plane of the code by 44 bits. This is accomplished by shortening 2 full rows from the first (xy)-plane, with each row removing 20 bits from the data block, and shortening another 4 bits from the next row. This results in a (2544,1096) code, with rate = 0.43. The following diagram shows the original code, along with the physical location of the shortened bits.
Figure 26 shows the original code along with the physical location of the shortened bits.

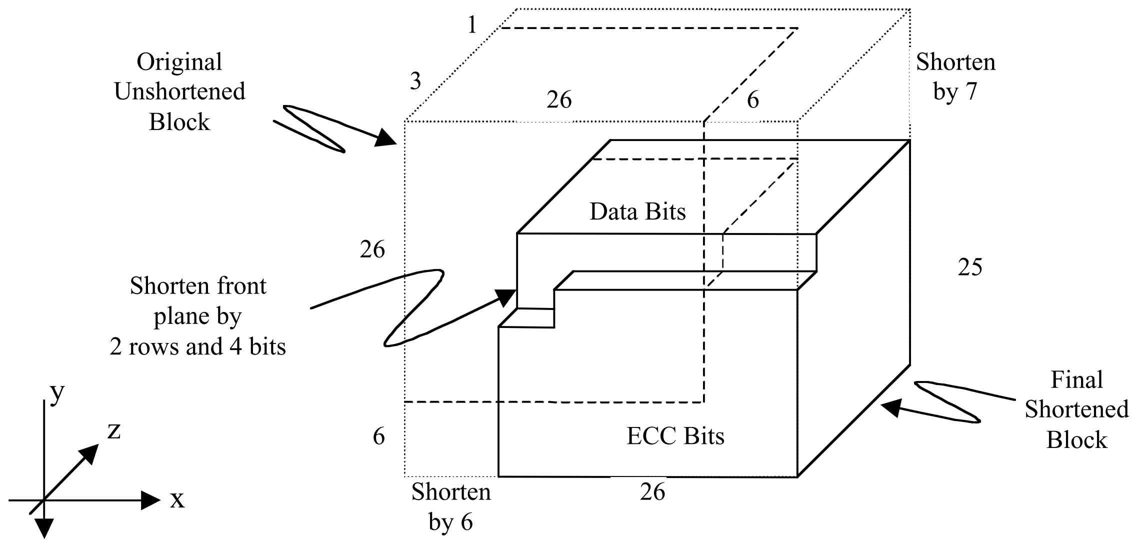

Figure 26: Structure of Shortened 3-D Block

4.2.3.2.7 Iterative Decoding

Huge performance advantages may be directly associated with the decoding mechanism for product codes. There are many different ways to decode product codes and each has its merits, however, the goal is maximum performance for a manageable level of complexity.

It is known that if it is possible to use unquantised information (so called soft information) from the demodulator to decode an error correcting code, then an additional gain of up to 2 dB over fully quantised (hard decision) information is achievable. It is therefore desirable to have soft information decision available to the TPC decoder.

Of course, we could in theory consider the decoding of this code a single linear code of size ($n_x \times n_y \times n_z$, $k_x \times k_y \times k_z$), using a soft decision decoder, but this will in general (apart from the smallest, and of course worst performing) be prohibitively complex.

It makes sense therefore, since these codes are constructed from (simple) constituent code that these soft decoders are used to decode the overall code. However until recently there have only been hard decision decoders for these constituent decoders. In recent years the computational power of devices has made it possible to consider (sub optimal) soft decision decoders for all linear codes. This is only half the solution as the main difficulty is with passing the information from one decoder to the next (i.e. when switching from decoding the rows to decoding the columns). For this, accuracy will need to be kept to a maximum, and so using soft input soft output (SISO) decoders will need to be considered. This is such that an estimate of the transmitted code word may be found and also an indication of the reliability. This new estimate may then be passed onto the next decoding cycle. Inevitably, there will be some degradation from optimal if we are to achieve our decoding using this method, but it does enable the complexity to be reduced to a level that can be implemented. Also, studies have shown that this degradation is very small, so this decoding system is very powerful.

What follows now is an explanation regarding the iterative nature of the decoding procedure. If we consider that, given 2-D TPC block, we define the first round of row and column decoding as a single iteration. We may then perform further iterations, if required. Thus, the main areas of investigation are that of the SISOs, and that of using some previously decoded information in subsequent decoding operations. These are both separate and yet connected areas of interest, as shall be explained.

With regards to the SISOs, there are many different methods including the following which have been described in detail in published academic papers:

1) Soft-Output Viterbi Algorithm (SOVA) [1]
2) The modified Chase algorithm [2]

3) The BCJR algorithm [3],

There have been many other papers explaining these algorithms both as independent algorithms for coding schemes and as part of turbo type decoding schemes. It must be noted that these are not the only algorithms that can achieve soft input soft output style decoding, but they are at present the most readily cited in academic literature.

Each block in a product code is decoded using the information from a previous block decoding. This is then repeated as many times as. In this way, each decoding iteration builds on the previous decoding performance.

Figure 27 illustrates the decoding of a 2-D TPC. Note here that prior to each decoding there needs to be a mathematical operation on all the data we have at that particular time, that is the current estimate of the decoded bits, the original estimate from the demodulator (this will not be used in the first decoding) and the channel information (where applicable).

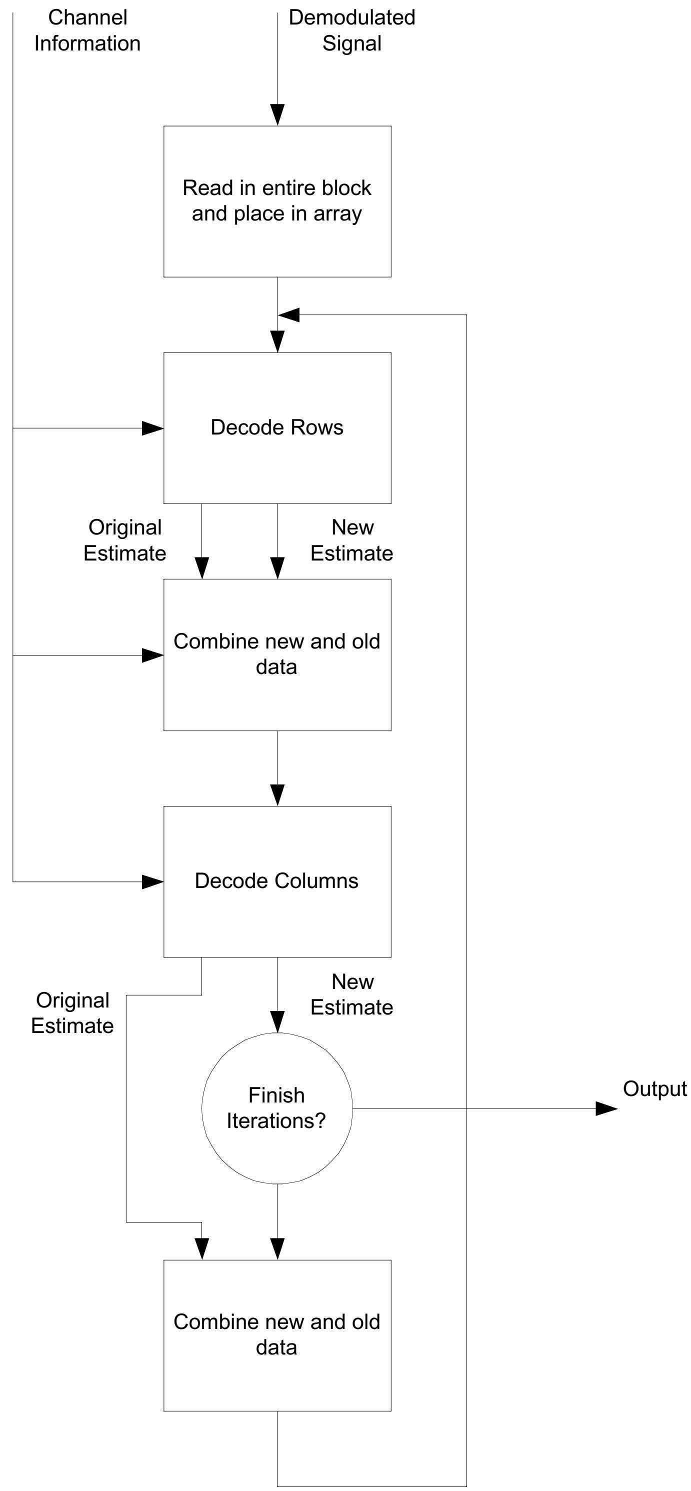

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Figure 27: Procedure for decoding of 2-D TPC

It can easily be seen from Figure 27 that the iteration idea is applicable to one complete decoding of the rows and one complete decoding of the columns.
There is an obvious question as to how the iteration procedure is terminated. This is a question only answerable by the system provider and depends on performance and delay; more iterations imply better performance as the expense of a

larger latency. Of course, over clocking the system in comparison can significantly reduce the latency. When considering hardware, the problem of varying delays may be encountered, thus it may be advantageous to fix the number of iterations performed.

4.2.3.3. Referecences

[1] G.Drury, G.Markarian, K.Pickavance. Coding and Modulation for Digital Television , KLUWER Academic Publishers, USA, 2001

[2] ETSI EN 301 021

[3] L.Bahl, J.Jelinek, J.Raviv, and F.Raviv, "Optimal Decoding of Linear Codes for minimising symbol error rate", *IEEE Transactions on Information Theory,* vol. IT-20, pp.284-287, March 1974.

4.2.4. Method 4 — Convolutional Turbo Codes

An optional convolutional turbo code may be employed to maximize range and throughput performance. The Turbo Code is based on Recursive Systematic Convolutional (RSC) code with a base rate of 1/2. The frames are encoded in blocks of N information bits (FRAME SIZE) listed in greater detail below. Figure 28 illustrates the overall encoding process and Figure 29 shows the RSC encoder.

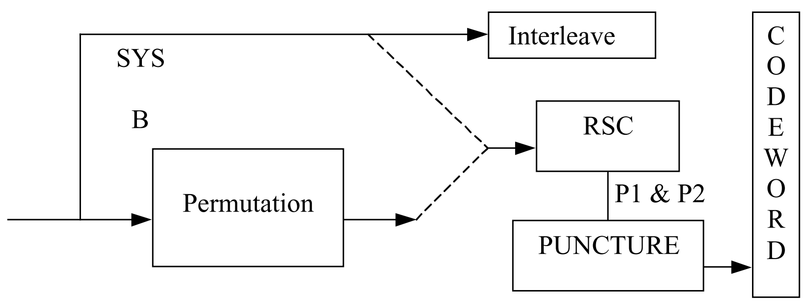

Figure 28: Turbo Encoder

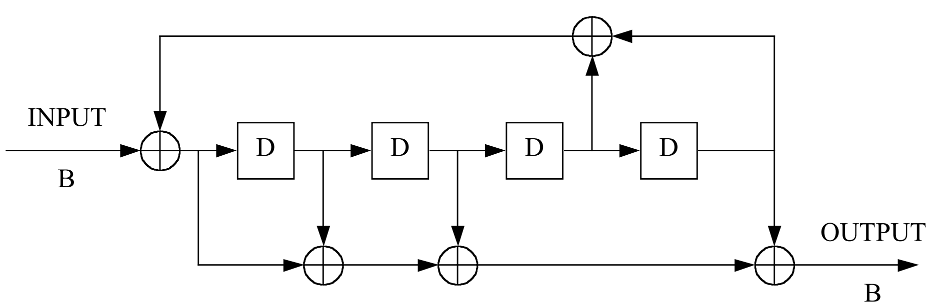

Figure 29: RSC Encoder

4.2.4.1. Termination

Each constituent code shall be terminated using a tail biting scheme.

4.2.4.2. Code Rates and Puncturing

Codes rates of 1/2, 2/3 and 3/4, are supported via puncturing. Table 13 lists the puncture patterns used to implement these rates.

Rate	Puncturing pattern						
1/2	P1	1	0				
	P2	0	1				
2/3	P1	1	0	0	0		
	P2	0	0	1	0		
3/4	P1	1	0	0	0	0	0
	P2	0	0	0	1	0	0

Table 13: Puncturing patterns

P1 are the parity check bits from the non-permuted encoding and P2 are the parity bits from the permuted encoding. All the systematic bits are transmitted for each rate.

Table 14 lists the standard set of frames sizes (N) and the number of 256 point symbols used for each coding rate. Each rate and frame size is designed to create an integer number of 256 point modulation symbols.

Frame Size (N)	Rate		
	1/2	2/3	3/4
96	1		
128		1	
144			1
288			2
384	4	3	
720			5
768	8	6	
1536	16	12	
1584			11

Table 14: Info bit frame sizes

4.2.4.3. Code Permuter

The size of the code permuter is equal to the information bits in the frame (N). To generate the interleaver of size N, two sets of values are stored in memory. The first set of values is size n and the second set of values is size m, where n*m = N. The n stored values are labeled n[0..n-1] and the m stored values are labeled m[0..m-1]. The interleaved addresses j[0..N-1] are generated as follows:

For first n addresses

j[0..n-1] = n[0..n-1]

Subsequent sets of n addresses are generated as follows

j[x] = (j[x-n]+m[x]) % N

Tables for n and m will be provided once frame sizes have been agreed upon.

4.2.4.4. Channel Interleaver

For each frame size a channel interleaver of size N is also defined. The channel interleaver is applied to the information bits after encoding by the first encoder. The interleaver I(x) is defined as follows:

I(0) = C, where c is a constant stored in memory.

I(x) = [I(x-1) + p] % N, where p is relatively prime to N.

4.2.4.5. CRC

A CRC-16 check sum block shall be added to each frame and is included in the information bits. The polynomial is $x^{16} + x^{15} + x^2 + 1$.

5. Constellation Mapping (Yossi — Editing Team)

The modulation used both for the US and DS data carrier is BPSK, QPSK, 16QAM and 64QAM. These modulations are used adaptively in the downlink and the uplink in order to achieve the maximum throughput for each link.

The modulation on the DS can be changed for each allocation, to best fit the modulation for a specific user/users. When using OFDMA the power of the modulated carrier can also vary by attenuation or busting of 6dB, this is used for the Forward APC.

For the up stream each user is allocated a modulation scheme, which is best suited for his needs.

The pilot carriers for the US and DS are mapped using a BPSK modulation.

5.1. Data Modulation

The data bits entering the mapper are after bit interleaving and they enter serially to the mapper, the mapping constellations are presented here after in Figure 30:

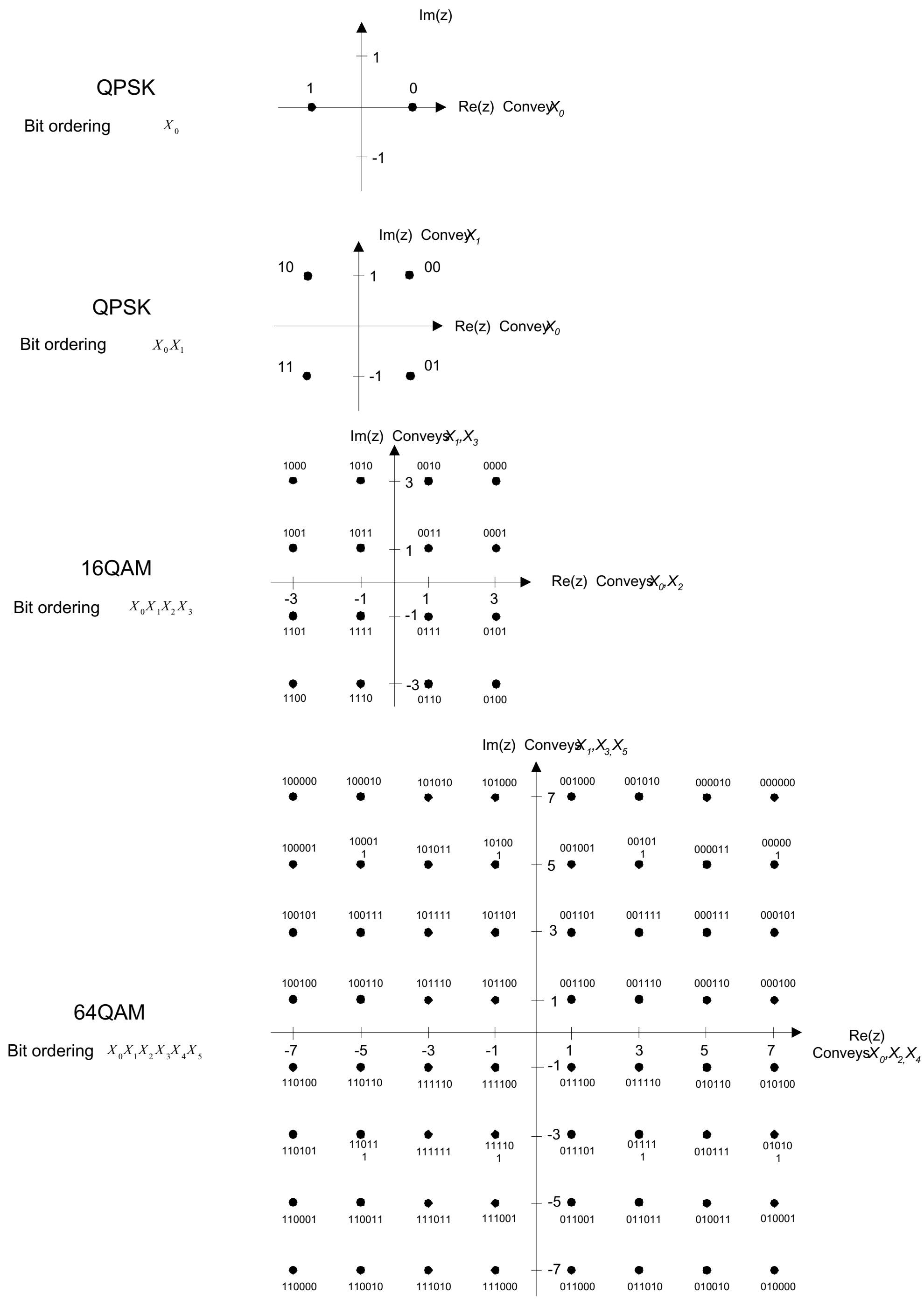

Figure 30: BPSK, QPSK, 16QAM and 64 QAM constellations

The complex number *z* shall be normalized by the value *c*, before mapping onto the carriers, by using the factor defined in the next table:

Modulation scheme	Normalization Factor 6 dB attenuation	Normalization Factor Reference 0dB	Normalization Factor 6 dB busting
BPSK	$c = z/2$	$c = z$	$c = 2 \cdot z$
QPSK	$c = z/\sqrt{8}$	$c = z/\sqrt{2}$	$c = z\sqrt{2}$
16QAM	$c = z/\sqrt{40}$	$c = z/\sqrt{10}$	$c = z\sqrt{2}/\sqrt{5}$
64QAM	$c = z/\sqrt{164}$	$c = z/\sqrt{42}$	$c = z\sqrt{2}/\sqrt{21}$

Table 15: Normalization factors

The complex number *c*, resulting from the normalization process, shall be modulated onto the allocated data carriers. The data mapping shall be done by sequentially modulating these complex values onto the relevant carriers. The reference-normalizing factor is used for the US, and the DS defined for 0dB busting or attenuation. The normalizing factors used for attenuation and busting are for DS use only, this is defined in the DS parameters for a specific burst type and is used for Forward APC.

5.2. Pilot Modulation

Pilot carriers shall be inserted into each data burst in order to constitute the Symbol Structure (see clause ???) and they shall be modulated according to their carrier location within the OFDM symbol.
The Pseudo Random Binary Sequence (PRBS) generator depicted hereafter, shall be used to produce a sequence, $w_{k.}$ .The polynomial for the PRBS generator shall be $X^{11} + X^2 + 1$.

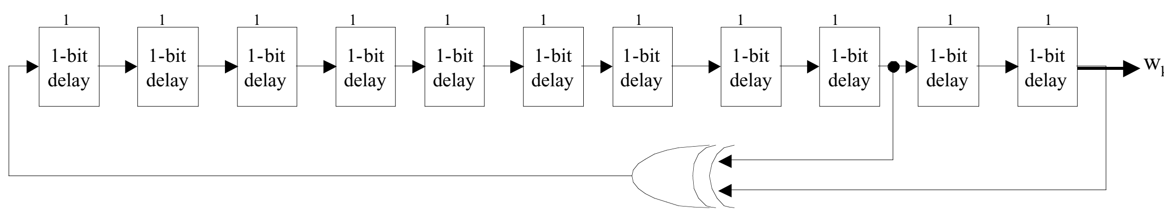

Figure 31: PRBS used for pilot modulation

The value of the pilot modulation, on carrier k, shall be derived from w_k.
When using data transmission on the DS the initialization vector of the PRBS will be: [11111111111]
When using data transmission on the US the initialization vector of the PRBS will be: [10101010101]

The PRBS shall be initialized so that its first output bit coincides with the first usable carrier. A new value shall be generated by the PRBS on every usable carrier. The DC carrier and the side-band carriers are not considered as usable carriers.
The pilots shall be sent with a boosting of 2.5 dB over the average energy of the data. The Pilot carriers shall be modulated according to the following formula:

$$Re\{C_k\} = 4/3 \times 2(\_ - w_k)$$
$$Im\{C_k\} = 0$$

When Sub-Channels are used for pilots transmission only (preamble or midamble) the pilots shall not be boosted. The Pilot carriers shall be modulated according to the following formula:

$$Re\{C_k\} = 2(\_ - w_k)$$
$$Im\{C_k\} = 0$$

5.3. Ranging Pilot Modulation

When using the ranging Sub-Channels the user shell modulate the pilots according to the following formula:

$Re\{C_k\} = (\_ - w_k) / 6$

$Im\{C_k\} = 0$

C_k is derived in clause XXX

6. RF Characteristics (Drayt, — Nico, Editing Team)

6.1. Regulatory Requirements

6.1.1. Introduction

The 802.16.4 PHY shall operate in the 5 GHz band as allocated by a regulatory body in its operational region. Spectrum allocation in the 5 GHz band is subject to authorities responsible for geographic specific regulatory domains e.g. global, regional, and national. The particular channelization to be used for this standard is dependent on such allocation as well as the associated regulations for use of the allocations. These regulations are subject to revision, or may be superseded. In the USA, the FCC is the agency responsible for the allocation of the 5 GHz U-NII bands.

In some regulatory domains several frequency bands may be available for 802.16.4 PHY based FWA devices. These bands may be contiguous or not, and different regulatory limits may be applicable. A compliant PHY shall support at least one frequency band in at least one regulatory domain. The support of specific regulatory domains and of bands within the domains shall be indicated by PLME attributes **RegDomainsSupported** and **FrequencyBandsSupported**.

6.1.2. FCC Regulatory issues

Equipment based on the 802.16.4 PHY is currently regulated in the USA by FCC document Title 47, Part 15. An interpretation of Subpart E, Section 15.407 is provided below for convenience. No rights may be derived from this text, and accuracy is not guaranteed.

The 5.15-5.25GHz lower U-NII band is restricted to indoor use only. As at least one device on an FWA link needs to be outdoors (typically the base station), this band is not available for FWA. As a result of this, the rule that the device must meet the maximum -27 dBm/MHz limit below 5250 MHz and above 5350 MHz applies. For the upper U-NII, the limits are maximum -27 dBm/MHz below 5715 MHz and above 5835 MHz, and maximum -17 dBm/MHz in the band 5715 - 5725 MHz and 5825 - 5835 MHz.

In the middle U-NII band, the transmit power is limited to the lesser of 24 dBm (250 mW) or $11+10\log(B_{26dB})$ dBm, with the peak power density (n.b. this is not the peak power) not exceeding 11 dBm/MHz. In the upper U-NII band, the transmit power is limited to the lesser of 30 dBm (1W) or $17+10\log(B_{26dB})$ dBm, with the peak power density not exceeding 17 dBm/MHz. For any multi-point system using directional antenna with gain over 6 dBi, limits for both bands are reduced by the antenna gain in excess of 6dBi.

6.1.3. ETSI Regulatory issues

Equipment based on the 802.16.4 PHY is currently regulated in Europe by ETSI document **TBD**. An interpretation of **TBD** is provided below for convenience. No rights may be derived from this text, and accuracy is not guaranteed.

6.2. Channelization

6.2.1. Channel Numbering

Channel center frequencies are defined at every integral multiple of 5 MHz above 5 GHz. The relationship between center frequency and channel number is given by the following equation:

Channel center frequency = 5000 + 2.5 * n_{ch} (MHz)

where n_{ch} = 0,1, 400. This definition provides a 9-bit unique numbering system of all channels with 2.5 MHz spacing from 5 GHz to 6 GHz to provide flexibility to define channelization sets for all current and future regulatory domains.

6.2.2. Valid operating channels

The set of valid operating channel numbers by regulatory domain is defined in Table 16. The channels in parenthesis are optional, the others are mandatory.

Regulatory domain	Band (GHz)	Channelization (MHz)		
		20	10	5
USA	U-NII-middle 5.25-5.35	(104), 112, 120, 128 (136)	(102), 106, 110, 114, 118, 122, 126, 130, 134, (138)	(103), (105), (107), (109), (111), (113), (115), (117), (119), (121), (123), (125), (127), (129), (131), (133), (135), (137)
USA	U-NII-upper 5.725-5.825	198, 306, 314, 322	(292), 296, 300, 304, 308, 312, 316, 320, 324, (328)	(293), (295), (297), (299), (301), (303), (305), (307), (309), (311), (313), (315), (317), (319), (321), (323), (325), (327)

Table 16: USA U-NII Operating channels

Figure 32 shows the channelization scheme for this standard, which shall be used with the FCC U-NII frequency allocation. The middle U-NII sub-band accommodates 3 channels of 20MHz, while the upper U-NII band supports 4 channels. Both U-NII bands accommodate 8 channels of 10 MHz. Additionally, two optional 20 MHz channels are defined in the middle U-NII band and one 10 MHz channel in each of the bands. The 5 MHz channelization, which is optional, provides 18 channels in each band.

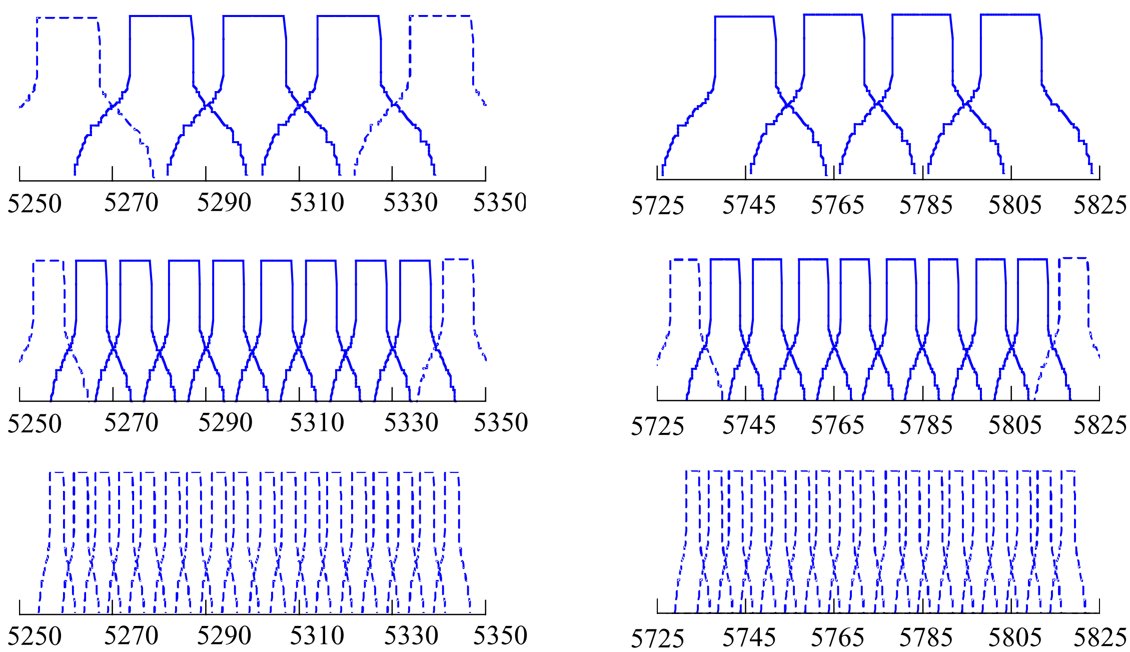

Figure 32: USA: U-NII frequency plan

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6.2.3. Channel Usage (Recommended Practice)

It is a recommended practice to use the channels in a Time Division Duplex (TDD) manner. If Frequency Division Duplexing (FDD) is used it will be a recommended practice to use the upper frequency band (5725-5825) for base station to subscriber unit transmission and the middle frequency band as the return link.

6.3. Transmitter Requirements

6.3.1. Transmit Spectrum Mask

The transmitted spectral density of the transmitted signal shall fall within the spectral mask as shown in Figure 33 and Table 17. The measurements shall be made using 100 kHz resolution bandwidth and a 30 kHz video bandwidth.

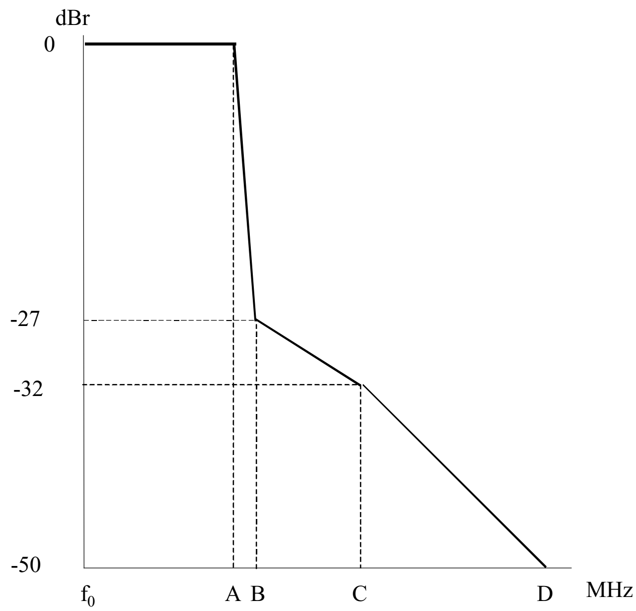

Figure 33: Transmit Spectral Mask (see Table 17)

Channelization (MHz)	A	B	C	D
20	9.5	10.5	19.5	29.5
10	4.25	TBD	TBD	TBD
5	2.25	TBD	TBD	TBD

Table 17: Transmit Spectral Mask Parameters

6.3.2. Transmit power levels

The maximum allowable output power according to FCC regulation is shown in Table 18.

Regulatory Domain	Band	Maximum Output Power			Comments
		20 MHz	10 MHz	5 MHz	
USA	U-NII middle	23 dBm	20 dBm	17 dBm	Up to 6 dBi antenna gain
USA	U-NII upper	29 dBm	26 dBm	23 dBm	Up to 6 dBi antenna gain

Table 18: Maximum output power

6.3.3. Transmit Power Level Control

The transmitter shall support power level control of 40dB minimum with resolution of 2dB maximum.

6.3.4. Transmitter Linearity

The transmitter linearity shall be sufficient to ensure minimal distortion of the transmitted OFDM signal. The test method will use a notch test method with the assistance of a test mode to create a notched OFDM signal. The average depth of the notch shall be **TBD** dB measured at the transmitter output. Use of an EVM may be used as an alternate test method, but all equipment must support all defined test modes.

6.3.5. Transmitter Spectral Flatness

The average energy of the constellations in each of the n spectral lines shall deviate no more than the following:

Spectral Lines	Spectral Flatness
Spectral lines from —n/2 to —1 and +1 to n/2:	+/-2dB from their average energy
Spectral lines from —n to —n/2 and +n/2 to n:	+2/-4dB from their average energy

Table 19: Maximum output power

This data will be taken from the channel estimation step.

6.3.6. Transmitter Constellation Error and Test Method

This requirement specifies limits for Error Vector Magnitude (EVM) measurements for the constellation elements.

The definition and test method are specified in 802.11a section 17.3.9.6.3. If this test is desired for 802.16.4, the definition can be copied straight from the 802.11a spec.

6.4. Receiver Requirements

6.4.1. Receiver Sensitivity

The bit error rate (BER) shall be less than 1E-6 at the power levels shown below for a standard message and test conditions. The measurement shall be taken at the antenna port or through a calibrated radiated test environment.

Channel Bandwidth (MHz)	Data Rate (Mbits/s)	Minimum Sensitivity (dBm)
20MHz		
20MHz		
20MHz		
10MHz		
10MHz		
10MHz		
5MHz		
5MHz		
5MHz		

Table 20: Receiver Sensitivity

Standard message: **TBD** bytes of **TBD** format
Test Conditions: room temp, no interference, conducted measurement if RF port available, radiated measurement in a calibrated test environment if antenna is integrated, FEC enabled

6.4.2. Receiver Adjacent and Alternate Channel Rejection

The adjacent channel rejection and alternate channel rejection shall be measured by setting the desired signal s strength 3dB above the rate dependent receiver sensitivity and raising the power level of the interfering signal until a 1E-6 bit error rate is obtained over **TBD** bytes. The power difference between the interfering signal and the desired channel is the the corresponding adjacent channel rejection. The interfering signal in the adjacent channel shall be a conformant OFDM signal, unsynchronized with the signal in the channel under test. For alternate channel testing the test method is identical except the interfering channel will be any channel other than the adjacent channel.

For the PHY to compliant, the minimum rejection shall exceed the following:

Channel Bandwidth (MHz)	Adjacent Channel Rejection (dB)	Alternate channel Rejection (dB)
20	40	50
10	40	50
5	40	50

Table 21: Adjacent and Alternate Channel Rejection

6.4.3. Receiver Interference Requirements

Interference may be mitigated through intelligent configuration and control, however it cannot be completely eliminated. For the receiver to be 802.16.4 compliant, it must not degrade more than **TBD** dB when operating in the presence of the following interference signals:

Signal Freq	User	Power level
#1	US govt radar	xx dBm
#2	ISM pt-pt	-10 dBm
#3	???	yy dBm
#4	???	zz dBm

Table 22: Interference Requirements

6.4.4. Receiver Maximum Input Signal

The receiver shall be capable of receiving a maximum on-channel signal of —20dBm, and must tolerate a maximum signal of 0dBm without damage.

6.4.5. Receiver Linearity

The receiver shall have a minimum Input Intercept Point (IIP3) of 0dBm.

6.4.6. Receiver Automatic Gain Control

The receiver shall support a minimum of 50dB of gain control with a minimum resolution of 2dB.

6.4.7. Receiver Group Delay

The receiver group delay variation across the desired receive channel shall not exceed **TBD** us.

6.4.8. Receiver Diversity (optional feature)

Receive diversity is recommended for enhanced performance and range.

6.5. Frequency Control Requirements

6.5.1. Transmit/Receive Center Frequency and Symbol Clock Frequency Tolerance

The transmitted center frequency, receive center frequency and the symbol clock frequency shall be derived from the same reference oscillator. At the BS the reference frequency tolerance shall be +/- 20ppm maximum for 64-FFT and 256-FFT modes and +/- 2ppm maximum for 1024-FFT mode. At the SS, both the transmitted center frequency and the symbol clock frequency shall be synchronized to the BS with a tolerance of maximum 1% of the inter-carrier spacing.

For mesh capable devices, all devices shall have a +/- 20ppm maximum frequency tolerance and achieve synchronization to its neighboring nodes with a tolerance of maximum 3% of the inter-carrier spacing.

During the synchronization period as described in the 802.16 MAC, the SS shall acquire frequency synchronization with the specified tolerance before attempting any uplink transmission. During normal operation, the SS shall track the frequency changes and shall defer any transmission if synchronization is lost.

6.5.2. Frequency Lock Detect

All modems will monitor the status of the frequency lock detect and prevent transmission if the oscillators loose synchronization to the base station clock.

6.5.3. Phase Noise

The phase noise of the RF oscillators must meet —80dBc/Hz at 10KHz offset from the carrier.

6.6. General Requirements

6.6.1. Temperature Range

Four temperature range classes are defined for compliance with this specification:
Class 1 0 to 40C for indoor office environments
Class 2 -20 to 50C for industrial environments
Class 3 -30 to 70C for industrial environments
Class 4 -40 to +85C for extreme outdoor environments

6.6.2. Antenna Interface

Any exposed transmit and receive antenna interface port shall be 50 ohms. It is permissible to integrate the antenna and eliminate any external RF port.

6.6.3. Power Connection Interface (Recommended Practice)

It is recommended practice for all outdoor 802.16.4 equipment to use 24VDC or 110VAC as the power source.

6.6.4. Indoor to Outdoor Unit Interface (Recommended Practice)

For systems deployed using an outdoor RF unit and an indoor modem the following interface is recommended:
Transmit IF = 479MHz??
Receive IF = 330MHz ?? or same as transmit if TDD mode is used
Reference frequency = 10MHz???
Control communication channel= TBD format using approx 10KHz carrier

6.6.5. Required Test Modes

To simplify the test and qualification of 802.16.4 compliant products, the following test modes shall be incorporated into all compliant equipment. Activation and deactivation of the test modes shall be performed via software control.

Test Mode	Purpose
Loopback	To allow loopback testing in the field
Notch test mode	Transmits a notched OFDM signal for transmit

	linearity testing
BER test mode	Performs functions required to perform a BER test

Table 23: Test Modes

7. Coexsitence in the middle UNII band - Interference and Mechanisms For Sharing the middle UNII band (Demosthenes Kostas —Editing Team)

7.1. General

The Wireless HUMAN Standard-based systems that will operate in the middle U-NII band (5.25-5.35 GHz) will have to share this band with a number of other systems (e.g., Earth Exploratory Satellite (active) Service (EESS) Synthetic Aperture Radars (SARs), Wireless HUMAN Standard-based systems, non-standard point-to-multipoint Broadband Fixed Wireless Access(BFWA) systems, terrestrial Radars, and IEEE 802.11a , 802.15 and Hiperlan/2 Wireless LANs). As this is a License-Exempt (LE) band these diverse systems will often be operated in the same geographical area by different operators. Moreover besides having to meet local Regulatory requirements (e.g., in the USA the FCC Subpart E Requirements) the Wireless HUMAN Standard-based systems will also be called to meet global agreements; e.g., from the World Radiocommunications Conference (WRC).

The goal of this Section is to identify (and where possible quantify) the interference requirements and identify mechanism that can be used by the Wireless HUMAN Standard-based systems to successfully share spectrum and geographical location with the likely to be encountered diverse systems.

7.2. EESS SAR interference requirement

This interference Wireless HUMAN interference requirement is considered first as it stems from the 1997 World Radiocommunications Conference (WRC '97) that allocated the 5250-5350MHz and 5350-5460 MHz bands on a world wide-primary basis to radiolocation services. These bands are currently also allocated on a world wide-primary basis to active space-borne sensors, including SARs (e.g., SAR 1-4). For the Characteristics of these SARs see Appendix 2 in ITU-R WP7C/126, "Analysis of Potential interference Between Spaceborne SARs and Wireless High speed Local Area Networks Around 5.3.GHz.". The "primary basis" classification means that it is a requirement that must be met by Wireless HUMAN Standard-based systems. The Subsections that follow identify mechanisms that can be used to enable the Wireless HUMAN Standard-based systems to meet the interference requirements of EESS SAR sensors.

7.2.1. Antenna directivity to mitigate interference to EESS

What follows gives an indication of the interference that Wireless HUMAN based BFWA systems can cause to SARs operating in Middle U-NII band, and identify means for minimization of this interference. In particular it has been shown by published results of ITU-R studies that BFWA antenna directivity is effective in minimizing interference to SAR-4, (e.g., USA ITU-R WP7C/24 Contribution). Table 24 shows that use of 6dB antenna directivity can decrease the SAR-4 interference by 4dB.

Note: The value of antenna directivity that should be specified requires trade-off studies with the other mechanism. SAR-4 is used because the SAR-4 system is more interference sensitive than SAR-3 and -4, and the SAR-4 center frequency is 5.3GHz.

The SAR-4 Synthetic Aperture Radar scans a path from 20^{o} to 55^{o} from Nadir. This corresponds to Earth incident angles of 21^{o} and 60^{o} which can be translated to angles of 69 ^{o} and 30^{o} with respect to the horizon. That is, any radiation from U-NII devices within that angular range could cause/contribute to satellite interference.

An approach that can be used in analyzing the interference potential from Middle U-NII BFWA systems into space-borne SAR-4 receiver is to determine the worst case signal power received from a single BFWA transmitter at the spaceborne SAR. Then, the single interferer margin can be calculated by comparing the single BFWA interferer level with the SAR-4 interference threshold. Knowing the SAR-4 footprint, the allowable density of active BFWA transmitters can then be calculated, if a positive margin results from a single BFWA interferer.

Parameter	System	Value	dB
Transmitted Power, Watts	BFWA1	0.25	-6.02
	BFWA2	0.25	-6.02
Building Loss, dB	-	0.00	0.00
Antenna High Elevation Gain, Xmit dB	BFWA1	0.00	0.00
	BFWA2	-4.00	-4.00
Antenna Gain, Rcv dB	-	44.52	44.52
Polarization Loss, dB	-	3.00	-3.00
Wavelength, m	-	5.65E-02	-24.96
(4*pi)-**2	-	6.33E-03	-21.98
Distance, km	-	425.67	-112.58
Power received, dBW	BFWA1	-	-124.03
	BFWA2	-	-128.03
Noise Figure, dB	-	4.62	4.62
k*T	-	4.00E-21	-203.98
Rcvr Bandwidth, MHz	-	46.00	76.63
Noise power, dBW	-	-	-122.73
SAR-4 Interference threshold (I/N=-6dB)	-	-	-128.73
Margin, dB	BFWA1	-	-4.71
	BFWA2	-	-0.71

Table 24: Interference from a Single U-NII BFWA Transmitter to SAR-4

Table 24 shows the signal power at the SAR-4 receiver from a transmitter with power output of —6 dBW (24 dBm) and an isotropic radiator with unity gain at all look angles.. The space loss at angles of 21^{o} and 60^{o}, receive antenna gain, polarization loss, scattering gain and satellite interference threshold are derived from ITU-R reports. The reference margin is the difference between the Signal Power at the Satellite Receiver and the Satellite Interference Threshold. The negative margin numbers indicate that radiating an EIRP of 24 dBm toward the satellite will exceed the interference threshold. Fortunately, real-world antennas do not exhibit unity gain at high elevation look angles, and this feature can be used to mitigate interference

A conclusion that can be drawn is that antenna directivity, if properly utilized, will provide interference margin for multiple transmitters. However, it should be noted that the satellite footprint is large (53 sq km at 20^{o} from Nadir and 208 sq km at 55^{o} from Nadir). Therefore, given the potential variables associated with the design, installation and maintenance of the various unlicensed transmitters, antenna directivity alone may not be sufficient to assure non-interference.

It should also be noted that antenna characteristics vary from antenna to antenna and from manufacturer to manufacturer. Also, the physical environment at the antenna mounting site can degrade the off-beam lobes of many antennas, especially omni-directional and directional antennas with no/small backplanes. Therefore, designers/installers should not assume published patterns and antenna range measurements are totally reliable.

7.2.2. DFS to mitigate interference

As frequency planning is not practical in licensed-exempt bands, Dynamic Frequency Selection (DFS) can be used to avoid assigning a channel to a channel occupied by another system. DFS action will be based on Subscriber Unit (SU) and/or Access Point (AP) received signal measurements (during idle channel conditions) and comparing the measurements to a threshold based on the required C/I parameter.

Considering that a Wireless HUMAN-based devices will cohabit the licensed exempt bands with other Wireless HUMAN-based devices and non-Wireless HUMAN-based devices, detailed DFS parameter description is needed for intra- and inter-system interference estimations. This inter-system interference includes coexistence considerations for such WRC 1997 globally primary assignments as Earth Exploratory Satellite Service SARs, and military radars. Wireless HUMAN-based intra-system interference estimates are also needed to assure that the desired level of performance is also possible.

7.2.3. Transmitted power control to mitigate interference

With power control, the EIRP of some transmitters will be below the permissible level. Therefore, an estimate of total interference within the footprint could be as much as n dB below the reference value. However, this assumption might well be negated as the subscriber population grows and interference between neighbors increase in which case, the transmitted EIRP could remain near the legal limit.

7.2.4. Antenna polarization to mitigate interference

The margin calculation in Table 24 includes 3 dB polarization loss. The fact that most P-MP systems rely on polarization for maximizing channelization, as many as half of the U-NII transmitters in a given area could be transmitting on each polarization. If so, the 3 dB polarization loss may not be fully realizable.

If the satellite were restricted to one linear polarization and the U-NII transmitters were restricted to the other linear polarization, greater polarization isolation could be achieved. Given the operational needs of both services, this is unlikely to happen.

7.2.5. Turning off Wireless HUMAN systems when a Spaceborne SAR is in the area

Earth Resources Satellites operate at very precise orbits, therefore, the Ephemeris (orbital position correlated to time) of the satellites can be calculated with great accuracy. If an Ephemeris calculator is included in the stations, and/or in the networks, the stations could be muted during the satellite pass. This would amount to a muting time of approximately 15 seconds per satellite pass typically 5-10 days per pass. Coupled with antenna directivity, this feature would allow virtually any number of stations to operate within the satellite footprint. The main concern with Ephemeris calculations the ability to maintain a reliable muting process over time. To do so requires as a minimum: 1) monitoring individual stations to assure that muting is taking place at the correct time, 2) maintenances process that assure prompt repair of faulty stations, 3) capability to update Ephemeris algorithms should orbital/time corrections be required, 4) adding new satellites as they are launched.

7.2.6. Other ways to mitigate interference

8. Optional features for operation in cellular and sectorized environments (Octavian - Editing Team)

8.1. Recommended practice for time synchronization in sectorized environments

In sectorized environments, where several base-stations are collocated in the same hub, the co-channel and adjacent channel interference can be significantly reduced if all base-stations are frame synchronized, i.e. they switch between transmit and receive and reverse at the same time. The reduction in co-channel and adjacent channel interference comes at the expense of the bandwidth usage because the partition between uplink and downlink cannot be optimized separately for each base station . For synchronization, the base-stations should use the same frame size and the same uplink/downlink partition. They should also be synchronized to a common time reference. A hub controller should be able to change the frame size and the uplink/downlink partition simultaneously in all base-stations that belong to that hub. Also, to preserve synchronization over a long period of time, the base-stations should use a common frequency reference.

If this feature is to be implemented, it is recommended that the BS should be able to synchronize the start of the MAC frame with 1 Hz pulses received on an external input. The same pulses should trigger changes in the frame size and uplink/downlink partition in all BS s at the same time. In other words, after the hub controller has updated these two parameters in each base-station, changes should take place simultaneously in all base-stations at the next synchronization pulse. To preserve synchronization over a long period of time, the base-station should be able to lock its reference frequency to an external 10 MHz reference frequency. It is recommended that the electrical specifications for the 1 Hz and 10 MHz inputs, should be compatible with the GPS. However, it is not required to use GPS for synchronization.

8.2. Recommended practice for time synchronization in cellular environments

Synchronization can be extended beyond a single hub, to the network level, if all hubs use a common time and frequency reference, like the GPS. Such synchronization makes the in-system interference deterministic, which can facilitate various interference mitigation techniques:

- Smart across-network UL/DL scheduling to avoid interference in sensitive areas
- Insertion of silence periods in some sectors/hubs to allow another sector/hub to accomplish difficult connections
- Synchronized Dynamic Frequency Selection (DFS) across the network

Depending on deployment conditions, different hubs may require different time offsets from the common time reference to reduce the overall interference. Thus, if synchronization between hubs is implemented it is recommended that each hub should be capable to use distinct delays from the 1 Hz pulse, with all base-stations in a hub having the same delay.

8.3. Recommended practice for power coordination in sectorized environments

For a single sector, automatic power control (APC) attempts to equalize the uplink received levels so that the automatic gain control (AGC) in the base station must not track large variations of the received signal level. The resulting received level will usually be a compromise between the near and far stations. In a sectorized environment, the APC may reach different compromise levels for different base stations. In such situations the base stations with lower received levels may suffer an additional bandwidth reduction due to co-channel and adjacent channel interference from the uplinks addressed to other base stations. It is possible to minimize the overall interference effects in a hub by equalizing the transmitted and received power levels (measured at the base-station) across the entire hub. However, a complete optimization should take in consideration also the network traffic for each separate station.

If this feature is to be implemented, the base station should be capable to adjust its APC transmit and receive levels upon request from the hub controller. The hub controller should adjust the levels so that it optimizes the bandwidth usage across the entire hub for the current network traffic.

8.4. Recommended practice for power coordination in cellular environments

Power coordination can be extended beyond a single hub, to minimize the interference between sectors belonging to different hubs. Such coordination, may be used to facilitate various interference mitigation techniques. However, to be efficient, power coordination across multiple hubs may require different transmit and receive power levels within the same hub. In other words, minimizing interference between hubs may cause an increase in the interference between sectors belonging to the same hub.

If such feature is to be implemented, the hub control should be capable to set distinct APC levels for different base stations. It is also recommended that, the power levels should be adjusted so that the overall network bandwidth is optimized for the current traffic.

8.5. Dynamic frequency selection in sectorized environments

8.6. Dynamic frequency selection in cellular environments

Note: The recent changes in this section reflect opinions debated on the TG4 reflector:

- *Four new subsection were added to cover another issues specific to sectorized and cellularenvironments*
- *The title says clearly that these are optional features*
- *The wording recommended practice applies now only for the implementation of the synchronization and coordination mechanisms. The synchronization and coordination themselves are optional.*
- *Synchronization drawbacks are clearly stated.*

Previous changes:

- *I modified this section to include opinions that were strongly debated in Hilton Head. I provided opportunities for synchronization between hubs (as suggested by Demos) but I made the delay between hubs adjustable (which should be better than random delay as suggested by Nico).*
- *I removed requirements for MAC since this is part of an upper layer and, as nicely argued by several people, they fall outside the scope of the standard.*
- *Anyway the whole section is optional and just recommended practice.*

9. Ranging and Access schemes (call for comments) (Tal —Editing Team)

The 802.16 MAC provides several mechanisms that can be used by an SS to request from BS additional bandwidth: piggyback, unicast polling and multicast polling. In the later case, all SS s matching the group address can place their bandwidth request in the BW Request Contention part of the uplink. The two proposed subcarrier-based polling methods provide an alternate way to answer the multicast polling without contention. The first method uses on-off keying while the second employs DBSK modulation.

9.1. Subcarrier-based polling using on-off keying

With subcarrier based polling each SS uses on-off keying over OFDMA to signal a bandwidth request. A well-known part in the uplink is dedicated to for the purpose of the polling. In this period, each SS is assigned a unique set of N_{BWR_SC} subcarriers spread in time and frequency. An SS requests bandwidth by energizing it s allocated subcarriers. When the BS detects the presence of energy in a specific set of allocated subcarriers, it sends a unicast polling addressed to the corresponding SS, at the earliest opportunity.

Let assume that N_{maxSS} is a multiple of the number of data subcarriers N_{SC} or that is increased to be such. The subcarrier based polling requires $N_{BWR\_SC}*N_{maxSS}/Nsc$ OFDMA symbols. In each allocation, N_{BWR_SC} subacrriers of the same symbol are allocated. Within a symbol, the subcarriers are assigned on a regular grid. The subcarriers are placed Nsc/N_{BWR_SC} apart. Thus to specify an allocation the symbol number in the range 0 $N_{BWR\_SC}*N_{maxSS}/Nsc$ —1 and the grid number in the range Nsc/N_{BWR_SC} are needed.

To avoid fades, the subcarrier assignment is randomly varied from poll to poll. This is performed as follows. At registration, the BS assigns to each SS a number *SSID* in the range 0 N_{maxSS}-1. Then, with each multicast polling, the BS broadcasts a random number *R* in the same range. An SS uses these values to determine the frequency allocation according to:

1. Let n=($R+SSID$) mod N_{maxSS}
2. Let k=floor (n/ (Nsc/N_{BWR_SC})) be the symbol number.
3. Let l = n mod (Nsc/N_{BWR_SC}) be the grid number.

9.2. Subcarrier-based polling using DBPSK

With subcarrier based polling each SS sends just one bit as opposed to sending a complete BW request packet (in contention window). If this bit is “1” then the SS needs additional BW and the BS must send a unicast polling addressed to the corresponding SS, at the earliest opportunity. If the bit is “0”, the BS knows it does not need to inquire further the SS. The subcarrier based polling response is transmitted using OFDMA with DBPSK modulation. Each SS needs one slot, where a slot occupies one carrier during two OFDMA symbols. The first OFDMA symbol is used as reference and can be the corresponding subcarrier of the long preamble. The second OFDMA symbol encodes differentially the BW request bit. The second OFDMA symbol preserves the phase from the first OFDMA symbol when the BW request bit is “0” and shifts the phase with 180 degrees when BW request bit is “0”.

To avoid channel fades, the slot positions are randomly permutated from polling to polling. Let the maximum number of SS s allowed in a system be N_{maxSS} (it could be useful to make N_{maxSS} a multiple of the number of data subcarriers N_{SC}). The subcarrier based polling requires N_{maxSS} OFDMA slots occupying 2*ceil(N_{maxSS} / N_{SC}) OFDMA symbols. Assume the slots indexed in the range 0 N_{SC} *ceil(N_{maxSS} / N_{SC})-1, first by the carrier and then by the OFDMA symbol. At registration, the BS assigns to each SS a number *SSID* in the range 0 N_{maxSS}-1. Then, with each multicast polling, the BS broadcasts a random number *R* in the range 0 N_{SC}*ceil(N_{maxSS} / N_{SC})-1. Each SS calculates the position of its BW request slot from its *SSID* and the random number *R* received from BS, by formula (*SSID*+*R*)mod(N_{SC}*ceil(N_{maxSS} / N_{SC})).

10. Additional Possible Features (T.J. — Editing Team)

10.1. Adaptive Arrays

10.2. Transmit diversity Alamouti’s Space-Time Coding

10.3. STC for FFT sizes 64 and 256

10.4. STC for FFT size 1k

10.5. Alamouti STC Encoding

11. Annex A — Channel and Interference Model (Tal — Editing Team)

11.1. Introduction

The purpose of this document is to present channel and interference models for 802.16.4 OFDM PHY. The models are targeted towards parameter optimization rather then for establishing the actual performance of the proposed system. The models include the following elements:

- A channel model, which captures effects of multipath.
- A Radio impairments models.
- An interference model, capturing the effects of typical interference, which exists at unlicensed bands.

The underlying guidelines for this work, are to try to define a mathematical framework for the elements under consideration, rather then to try to match the models to specific scenarios. This approach will result in an set of flexible models, which can tailored to specific situations and scenarios by a simple change of parameters.

The models proposed here are straightforward, simple to simulate and yet gives a realistic description of the system and the related impairments. The models are also mathematically tractable, and support a single parameter characterization.

The basic model is depicted in Figure 34. The specific blocks are described in subsequent sections.

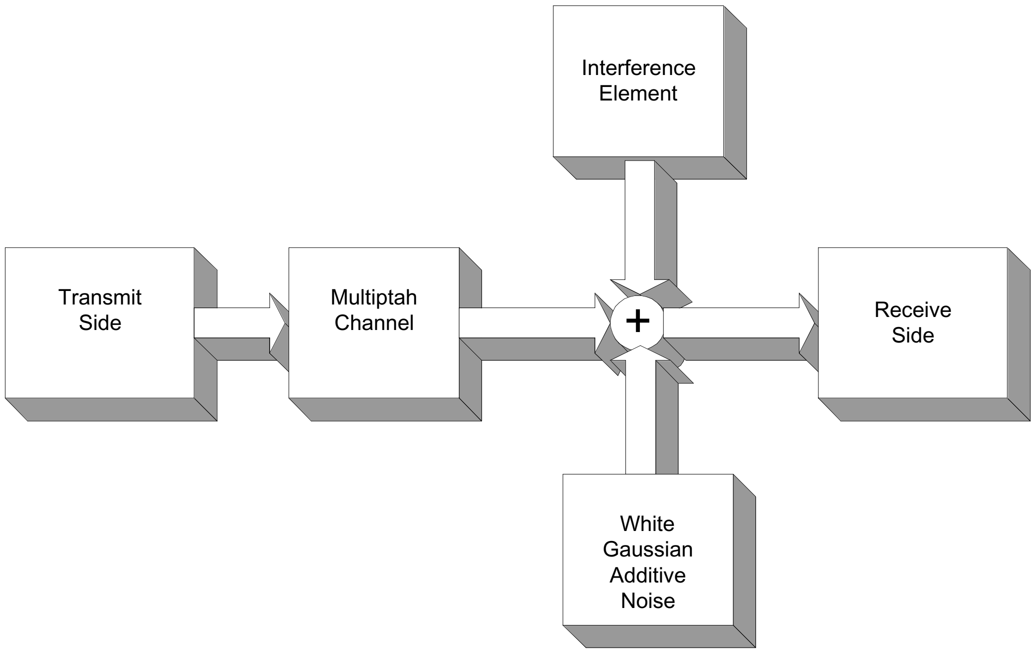

Figure 34: Basic model

11.2. Multipath channel

The multipath model is selected to be a Rayleigh fading model with an exponentially decaying power profile. The channel is specified by the RMS of the tap weights. This model is simple to analyze and simulate. With a proper choice of delay spread values it represents realistic conditions. For further discussion see [1] or [2].

The following, taken from [1], describes how to implement the multipath model in a discrete time simulation system.

Let $h_k = h(t)\big|_{t=kT_s}$ denote the sampled impulse response of the channel, where T_s is the sampling rate of the simulation system. The coefficients h_k are complex random numbers with random uniformly distributed phase and Rayleigh distributed magnitude. The average power decays exponentially. The RMS power average of the taps is given by the parameter T_{rms}. The coefficients are selected according to:

$$h_k = N(0, \tfrac{1}{2}\sigma_k^2) + jN(0, \tfrac{1}{2}\sigma_k^2)$$
$$\sigma_k^2 = \sigma_0^2 e^{-kT_s/T_{RMS}}$$
$$\sigma_0^2 = 1 - e^{-T_s/T_{RMS}}$$

where $N(0, \tfrac{1}{2}\sigma_k^2)$ is a zero mean Gaussian random variable with variance $\tfrac{1}{2}\sigma_k^2$ (produced by generating a $N(0,1)$ r.v. and multiplying it by $\sigma_k / \sqrt{2}$), and $\sigma_0^2 = 1 - e^{-T_s/T_{RMS}}$ is chosen so that the condition $\sum \sigma_k^2 = 1$ is satisfied to ensure same average received power.

In **Error! Reference source not found.**, the exponential power profile and a single realization of a channel are shown.

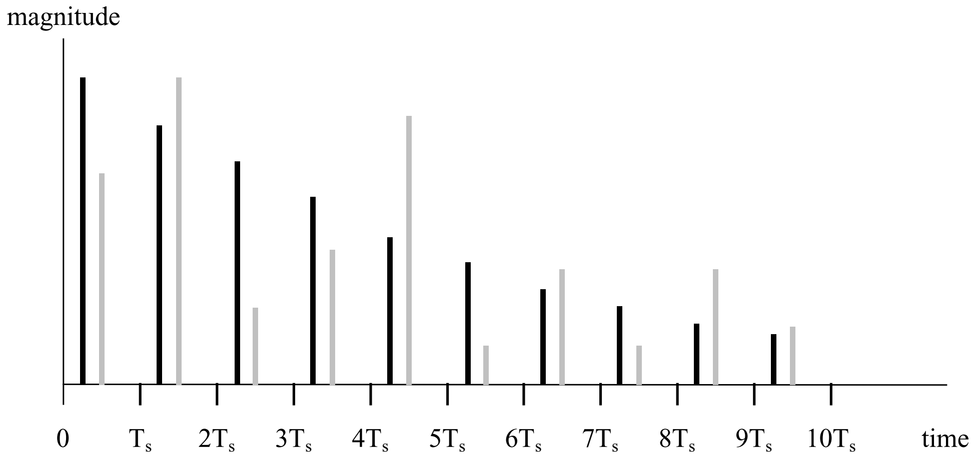

Figure 35: Power Profile (black) and a single realization (gray); the time positions are staggered for clarity only

The sampling time T_s in the simulation should not be longer than the smaller of 1/(signal bandwidth) or $T_{RMS}/2$. The number of samples to be taken in the impulse response should ensure sufficient decay of the impulse response tail, e.g. $k_{max}=10T_{RMS}/T_s$.

For each packet, a new channel response is generated. The channel is assumed to be static during a packet.

11.3. Interference models

Here the interference is assumed to be stemming from wide-band packetized transmissions (e.g. 802.11a HiperLAN/2). The model generates random interference bursts. For each burst the following parameters are selected at random:

1. Arrival time of burst.
2. Length of burst.
3. Center frequency of burst.
4. Power of burst relative to noise floor.

The focus in this section is to try to establish the mathematical framework for the interference model. Some crude assumptions are made with regard to the actual traffic parameters. These need to be refined.

The parameters are depicted schematically in Figure 36. Section 11.311.3.1 gives the underlying assumptions for the interference source. From those assumptions, the timing, power and signal descriptions are derived. They are described in section 3.2 and 3.3. Section 3.4 describes the procedure of generating the interference signal.

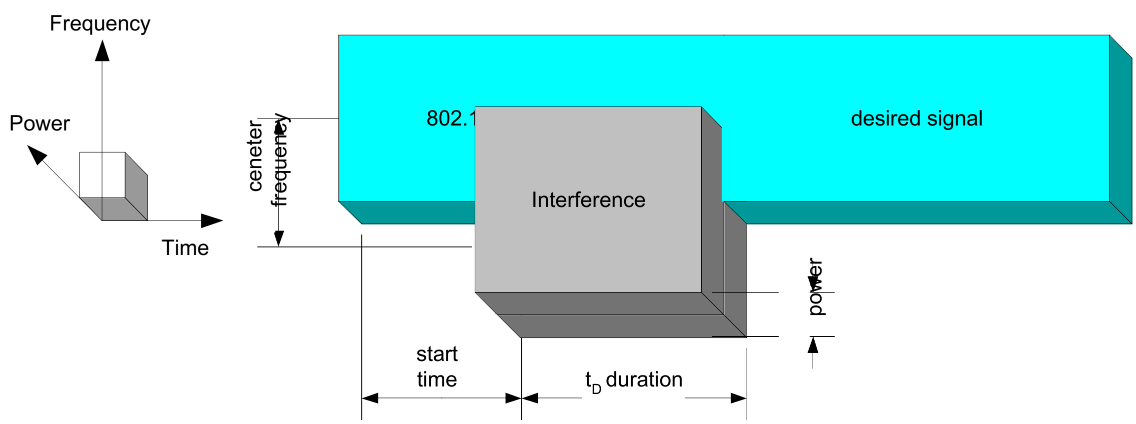

Figure 36: Interference parameters

11.3.1. Basic assumptions

Here we shall assume that the interferer is a 802.11a like signal (see [3]). The instantaneous transmission rate is 24Mb/s. The occupied bandwidth is about 17MHz. The PHY layer overhead per packet is assumed to be 20uSec.

For the interferer traffic, we shall use the results published in [4], where histograms of Ethernet packet sizes are shown. It is demonstrated that almost 75% of the packets are shorter than 522 bytes and nearly half the packets are 40-44bytes. In order to simplify and to reach round numbers, we shall assume that the packet size is uniformly distributed in the range 48 480bytes. This is equivalent to a packet duration in the range of 36 180 uSec.

The time between consecutive interference bursts is can be computed as follows. From [4], we can assume that most traffic, in bytes, is concentrated in long packets, say in 500bytes packets. Let us assume a channel utilization of 25%. Thus the average idle channel time is (1-0.25)/0.25*500*8/24Mb/s= 500uSec.
Here we shall assume that the average time from end of interference burst to beginning of next burst is Poisson distribution with a mean of 500uSec.

It is assumed that the power spectral density (PSD) of the interference is in the same order of magnitude as the thermal noise floor. More specifically, the PSD is in the range of N_0 to N_0+20dB, where N_0 is the thermal noise floor.(—174dBm/Hz).
The frequency offset between the interferer and the desired signal is uniformly distributed in the range -10Mhz +10MHz.

11.3.2. Signal Wave shape

The interference signal is generated by passing a white complex Gaussian process, through a raised cosine filter and amplifying it to the desired level. Then it is shifted in frequency in to a randomly selected center frequency.
The motivation for using this wave shape is as follows:

- The proposed signal is easy to generate
- The spectral signature is similar to that of many communication systems.
- It has roughly the same peak to average power ratio as that of an OFDM signal.
- This signal, can be easily modified to represent other interfering signals.

The parameters of the raised cosine filter are rolloff factor of β=0.25 and 6dB corner of f_c=10MHz. The filter is given by:

$$H(f) = \begin{cases} 1 & |f| < (1-\beta)f_c \\ 1/2\left[1 - \sin\frac{\pi}{2}\left(\frac{f}{f_c} - 1\right)\Big/\beta\right] & (1-\beta)f_c \le |f| < (1+\beta)f_c \end{cases}$$

11.3.3. Generating Procedure

The signal generation is depicted in Figure 37. The procedure for generating the interference is given below.

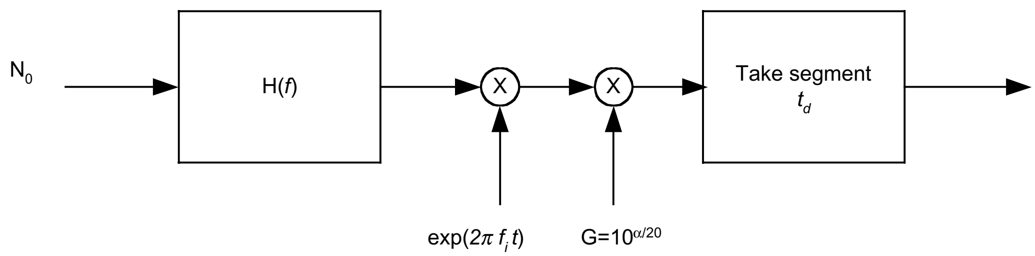

Figure 37: Generating the interference signal

1. Select the start time. Generate a poisson random variable t_0 with a mean of 1/λ= 500uSec according to the probability distribution function given by :
$$f_t(t) = t\lambda e^{-\lambda t}.$$
2. Add t_0 to the end of the last interference Burst. If this is the first interference burst, t_0 signifies start time from beginning of transmission burst.
3. Select the duration. Generate a uniformly distributed random t_D variable in the range 36uSec...180uSec.
4. Generate a white gaussian noise process with double sided PSD of N_0.
5. Filter the noise process with *H(f)* given above.
6. Select center frequency, f_i in the range —10MHz 10MHz.
7. Shift the signal in frequency by multiplying it by exp($j2\pi f$i t).
8. Select amplification. Generate a random variable α, uniformly distributed in the range 0 20. Set the amplification to G=$10^{\alpha/20.}$
9. Take a segment of t_D of the generated signal. Add it to the desired signal at the start time selected in steps 1 and 2.
10. Repeat with step 1.

11.4. Radio Impairments

The radio impairments models consist of phase noise models and power amplifier non-linearties.

11.4.1. Power amplifier non-linearity

The power amplifier model is based on the well-known Rapp's model ([5]) with knee parameter P=2. Besides its simplicity, the model well represents typical power amplifiers at the sub 10GHz range.

Consider using a complex baseband notation. Denote by v_{IN} and v_{OUT} the input and output complex signals, respectively. Let $P_{SAT} = |v_{SAT}|^2$ denote let the saturated power of the amplifier be $P_{SAT} = |v_{SAT}|^2$. Then the relation between v_{IN} and v_{OUT} is given by:

$$v_{OUT} = v_{IN} / \left(1 + \left(|v_{IN}| / v_{SAT}\right)^{2P}\right)^{1/(2P)}, P = 2 .$$

11.4.2. Phase noise

For the phase noise simplified phase noise model is selected. While maintaining a simple description the model adequately represent the behavior of typical microwave phase-locked loop oscillator.

The phase noise is presented as white gaussian noise process for which is driven through a single pole low pass filter. The 3dB corner of the low-pass should be set at 10KHz, which is a typical value for large step oscillators. A typical PSD is shown in **Error! Reference source not found.**.

The model ignores the contribution of the oscillator phase noise, which can be easily tracked and the effects of phase noise PSD flattening in high frequencies.

For simplicity it is recommended that the phase noise effects shall be simulated only on the transmitter side.

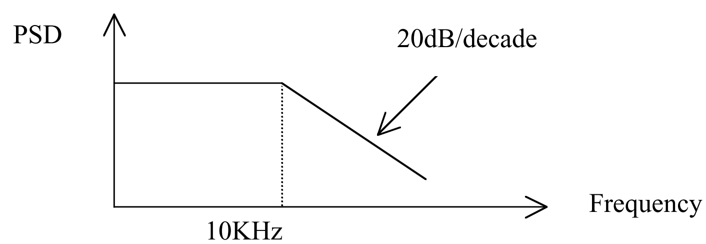

Figure 38: Phase noise PSD

11.5. Acknowledgements

It is a pleasure to acknowledge Dr. Vladimir Yanover for his assistance with the Caida database.

11.6. References

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[3] IEEE 802.11 Wireless LAN standard: high speed Physical layer in the 5Ghz Band.

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12. Annex B — MAC PHY interface (Radu — Itzik, Jori, + Editing Team)