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| Title | **PHY Layer System Proposal for Sub 11 GHz BWA** |
| Date Submitted | **2001-01-17** |
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| Re: | This contribution is submitted in response to "Call for Contributions: Session #11" by 802.16.3 Task Group chair on December 2nd, 2000 for submission of "PHY Proposals" for Sub 11 GHz BWA. |
| Abstract | This document provides team PHY System proposal of a low frequency (Sub 11 GHz) wireless access PHY for point-to-multipoint voice, video and data applications. The submission is for consideration of the Task Group to develop a PHY standard for BWA system. |
| Purpose | This contribution will be presented and discussed within the Task Group in Session #11 for possible adoption as baseline for a PHY standard Sub 11 GHz BWA. |
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# PHY Layer System Proposal for Single Carrier – Frequency Domain Equalizer for Sub 11 GHz BWA


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## 1 Scope

This document defines a proposed Physical Layer (PHY) for IEEE802.16.3 Broadband Wireless Access (BWA) systems in licensed frequency bands from 2-11GHz. Fixed BWA is a communication system that provides digital two-way voice, data, and video services. The BWA market targets wireless multimedia services to home offices, small and medium-sized businesses and residences. The BWA system shall be a point-to-multipoint architecture comprise of **Subscriber Stations** (SS) and **Base Stations** (BS, Hub station). Figure 1 illustrates a BWA reference model.

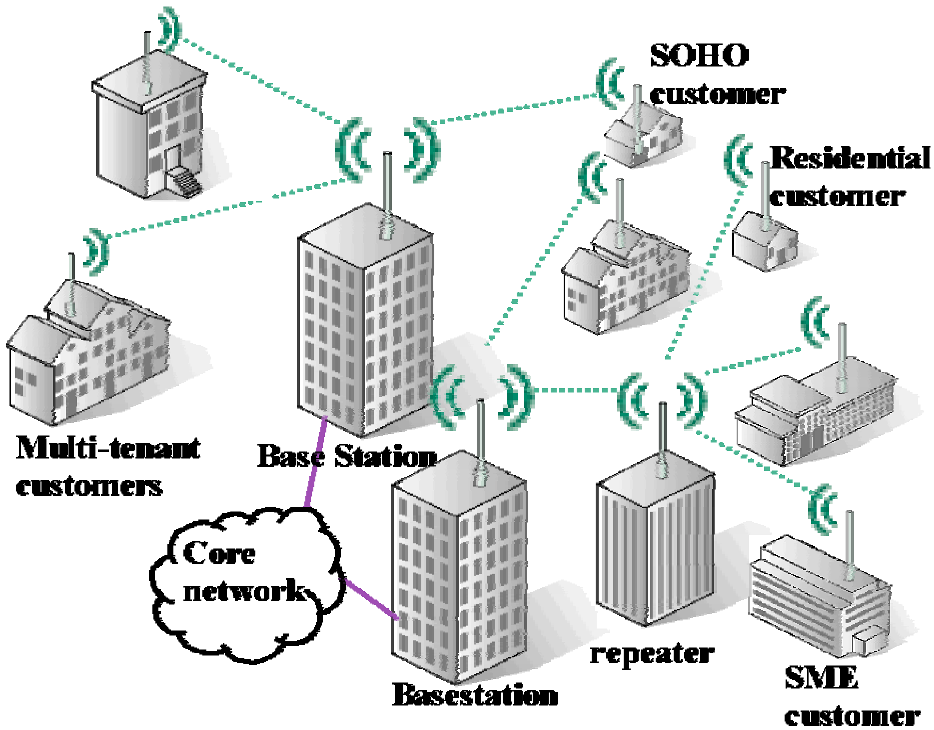



Figure 1: Wireless Access Reference Model

# 2 Introduction

## *2.1 General*

**The proposers believe that the 802.16.3 PHY standard should allow both Single Carrier (SC) and OFDM approaches to fully benefit from the features of each technology. This document will address the SC PHY.**

The proposed PHY system adopts TDM/TDMA bandwidth sharing scheme. The signal is transmitted downstream from the Base Station to all Subscriber Stations assigned to a carrier frequency in broadcast Time Division Multiplex (TDM) mode. The upstream signal is burst from the Subscriber Station sharing the same RF carrier with other Subscriber Stations to the Base Station in Time Division Multiple Access (TDMA) mode. This access scheme can be applied to either FDD or TDD. Both duplexing schemes have intrinsic advantages and disadvantages, so the optimum duplexing scheme to be applied depends on deployment-specific characteristics, i.e., bandwidth availability, Tx-to-Rx spacing, traffic models, and cost objectives.

Operating frequency band will be 2 to 11 GHz and the Base Station can use multiple sectors and will support smart antenna in future applications.

The proposed PHY layer is a Single Carrier (SC) modulation with a Frequency Domain Equalizer (FDE) (or SC–FDE). We will show that SC-FDE modulation can offer better performance than Orthogonal Frequency Division Modulation (OFDM) technology in solving the Non-Line of Sight (NLOS) problem that may arise in the 2 to 10.5 GHz bands.

In addition, this proposal introduces the concept of **coexistence** of the SC–FDE and OFDM modulation schemes for Sub 11 GHz BWA applications. Furthermore, the proposed frame structure for adaptive modulation and coding which is an ideal approach to make PHY almost independent from the MAC. **The PHY proposed here is based upon utilizing the structure of the 802.16 MAC.**

The **key benefits of the PHY proposed** here include:

1) Mature and well-proved
2) Adaptive Modulation and Coding
3) Flexible Asymmetry of Downstream and Upstream Paths
4) Scalability
5) Advanced Coding Schemes
6) Reduced System Delay
7) Easy Migration from simple SC to SC–FDE.
8) Straight forward migration to diversity receiver and multiple-input/multiple-output (MIMO technology).

## 2.2 Key Features

The PHY proposed here is a Broadband Wireless Access (BWA) **Point-to-Multipoint** communication system that can provide digital, two-way voice, data, Internet and video services. Proposed PHY offers an effective alternative to traditional wire line (cable or DSL) services.

Employing the functions of the 802.16 MAC such as QoS, the BWA system using the PHY proposed here will support services; such as packet data and Constant Bit Rate (CBR) as well as T1-E1, POTS, wide band audio and video services.

To maximize the utilization of limited spectrum resources in the low frequency bands (2 to 11 GHz), the air-interface supports statistical multiplexing over the air-interface using Time Division Multiple Access (TDMA) technology.

The main features of the proposal are the following:

- Full compatibility with the 802.16 MAC.
- Upstream multiple access scheme is based on TDMA.
- Downstream multiple access scheme is based on broadcast TDM.
- Duplex schemes are based on either TDD or FDD scheme.
- PHY uses a block adaptive modulation and FEC coding in both Upstream and Downstream paths.
- PHY proposal high capacity single carrier modulation with frequency Domain Equalization (SC-FDE) in addition to Decision Feedback Equalization in the time domain.
- The use of single carrier modulation techniques results in low cost Subscriber Stations (SS) and Base Stations (BS).
- The proposed modulation scheme is robust in multi-path and other channel impairments
- The PHY is flexible in terms of geographic coverage, in the use of frequency band, and capacity allocation.
- Base Station can use multiple sector antennas. Support for future use of smart antennas is implicit in the PHY design.
- The proposed PHY has an added feature of reconfigurability to support OFDM modulation.

# 3 PHY Proposal

As described in the Functional Requirement Document [1], this equipment employing this PHY and the 802.16 MAC have been designed to address the critical parameters for serving single family residential, SOHO, small businesses and multi-tenant dwellings customers--using **Broadband Wireless Access** technology. These critical parameters are combination of coverage, capacity and equipment cost factors that affect total cost per user. The deployability, maintainability, and product costs associated with the customer premise installation, and the spectrum efficiency and reuse for economically serving the required number of customer locations. Of particular importance to the proposed PHY presented here is the inherent versatility implicit in the Frequency Domain Equalizer (FDE) architecture. Conceptually, a dual mode receiver could be implemented in which the FDE configuration could be changed to receive an OFDM signal. The bases for this approach are shown in Figure 11.

## *3.1 Wireless Access System Model*

Figure 2 illustrates a top level block diagram of the proposed PHY layer system for BWA services:

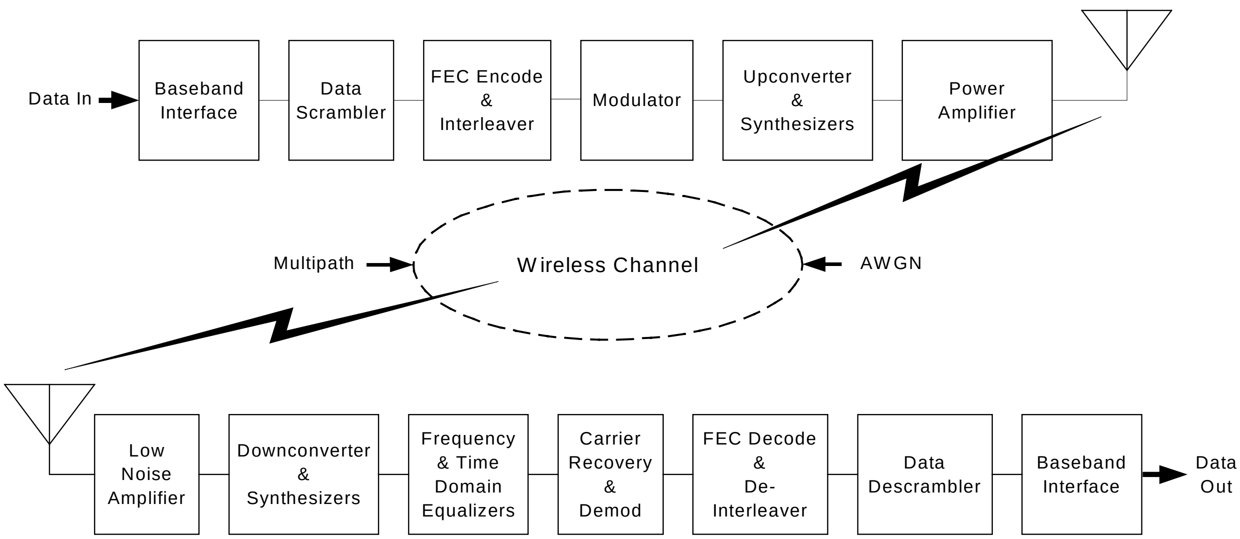



**Figure 2: The Proposed Single Carrier PHY Layer Block Diagram.**

## *3.2 Communications protocols*

As illustrated in Figure 3, the PHY Layer Protocol with upper layers stack is comprised of two sub-layers:

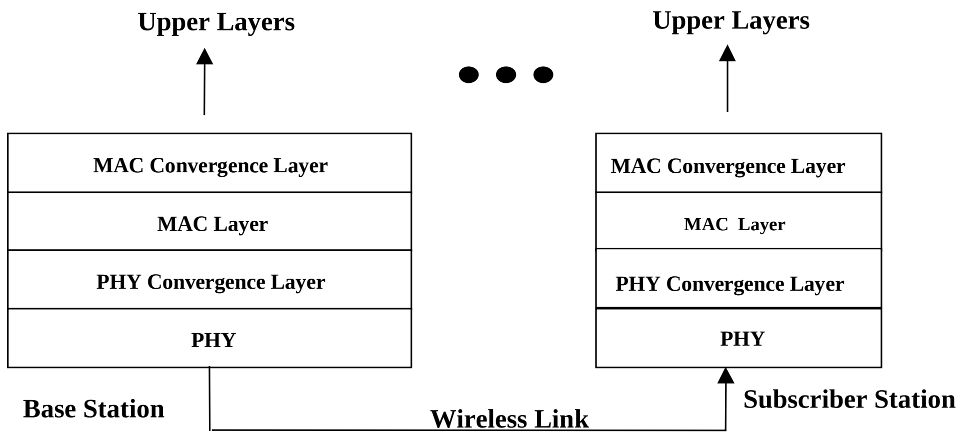



**Figure 3: The proposed PHY Layer with upper layers protocol stack.**

### 3.2.1 Physical Media Dependent (PMD) Sub-layer.

The PMD sub-layer involves the main processing parts of the PHY layer including: filtering and equalization, synchronization, randomizing, FEC encoding \ decoding and interleaving, baseband pulse shaping and other baseband processing units to enhance digitally modulated RF carriers over- the-air.

### 3.2.2 Transmission Convergence (TC) sub-layer.

This sub-layer is defined to adapt and map certain MAC services (such as changing resource allocations) to generic PMD services. These parts will be addressed briefly in the sequel and in more details at later stages of the standard development process.

## *3.3 The frequency range and the channel bandwidth*

The frequency range and the downstream and upstream channel bandwidth of the Proposed PHY system are given in Table 1.

**Table 1: Frequency Bands and Channel Bandwidth**

| Frequency Bands | Channel Bandwidth Options | Reference |
|---|---|---|
| 2.15- 2.162 GHz,<br>2.50- 2.690 GHz | 2 to 6 MHz downstream,<br>250 kHz to 6 MHz upstream | FCC 47 CFR 21.901 (MDS)<br>FCC 47 CFR 74.902 (ITFS, MMDS)<br>Industry Canada SRSP-302.5 (Fixed Services operating in the 2500 to 2686 MHz band) |
| b) 3.5 GHz | 1.75- 7 MHz downstream,<br>250 KHz to 7 MHz upstream | EN 301 021,<br>CEPT/ERC Rec. 14-03 E, CEPT/ERC Rec. 12-08 E, Others (TBD) |
| c) 10.5 GHz | 3.5, 5 and 7 MHz | EN 301 021, CEPT/ERC Rec. 12-05 E |

## 3.4 Duplex Schemes

In order to comply with the IEEE802.16.3 fuctional requirement [1], we propose to support both TDD and FDD systems and leave the selection of each system to the vendors /operators decision on implementation complexity, traffic scenario, cost objectives and spectrum availability.

### 3.4.1 TDD:

In **Time division duplex** (TDD) systems, the radio frame is divided into a downlink and an uplink section, offering flexibleand dynamic allocation of the upstream and downstream capacity. TDD enables the use of simpler antennas. In BWA system, where the delay between transmission and reception can consist of a few time slots, a guard time between the downlink and uplink sections of the frames has to be introduced in order to avoid collision between time slots. However, the guard time reduces system throughput, especially if the system is designed for low latency.

### 3.4.2 FDD:

In **Frequency division duplex** (FDD) systems, on the other hand, allocate a fixed proportion between uplink and downlink capacity. Residential users are likely to request asymmetrical uplink and downlink capacity, while in a business-user scenario, more symmetrical traffic behavior is likely to be the rule. FDD system can have full flexibility for instantaneous capacity allocation in the uplink and downlink for each access terminal and connection and it can address the business market segment easily.

## 3.5 Downstream Channel

### 3.5.1 Downstream Multiple Access Scheme

Each downstream RF channel (e.g., 6 MHz wide) is subdivided into fixed frames with which the RF carrier is suitability modulated (e.g., QPSK, 16 QAM, 64 QAM, etc) to provide a digital bit stream (e.g., 30 to 40 Mbps). Within each RF channel a frame structure is used to organize and schedule the transmission of voice, video and data traffic.

### 3.5.2 Modulation Scheme:

The applicable modulation schemes for the downstream are QPSK, 16 QAM or 64 QAM.

Adaptive **Modulation & Coding** shall be supported in the downstream. The upstream shall support different modulation schemes for each user based on the MAC burst configuration messages coming from the Base Station.

### 3.5.3 Downstream Randomization, Channel Coding & Interleaving, Symbol Mapping and Baseband Shaping

The downstream channel supports various modulation formats and FEC coding on the user data portion of the frame. Different modulation formats and FEC groups can be defined on a subscriber level basis. In this way the downstream channel supports adaptive modulation and coding. Note that each frame contains control portion with fixed modulation (QPSK) and FEC scheme. The details are described in Framing Section.

#### 3.5.3.1 Randomization for Spectrum Shaping

Prior to FEC encoding, the downstream channel will be randomized to ensure sufficient bit transitions to support clock recovery and to minimize occurrence of unmodulated carrier frequency. This process is done by modulo–2 addition (XORing) the data with the output of Linear-Feedback Shift Register (LFSR) with characteristic polynomial $1 + X^{14} + X^{15}$. The LFSR is cleared and preset at the beginning of each burst to a known value– 100101010000000.

The preambles are not randomized and only information bits are randomized. The LFSR sequence generator pauses while parity bits are being transmitted.

#### 3.5.3.2 Downstream Channel FEC definitions

Consistent with the structure of the 802.16 MAC, forward error correction code schemes which support both Block Turbo Coding and Concatenated Reed-Solomon+Convolutional coding will be employed. In addition, the provision of suppressing all FEC and operating using the ARQ mechanism in the 802.16 MAC for error control will be included.

Following is the summary of these coding schemes:

**Block Turbo Code**: This type of coding is based on the product of two or more simple component codes (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 BTC mode, are typically 2 dB better performance over the Concatenated RS, and shorter decoding delays. A detailed description of **Block Turbo Coding** is included as Appendix C.

**Concatenated Reed-Solomon+Convolutional code:** This case is based on concatenation of outer coding RS (204,188, t=8) and inner rate ½ Convolutional code with constraint length K=7. The Convolutional code is able to be configured to code rates 2/3, ¾, 5/6 and 7/8 using puncturing Convolutional interleaving with depth I=12 shall be implied as described in DVB-S spec [13]. The detailed description of Concatenated Reed-Solomon Coding is included as Appendix C.

#### 3.5.3.3 Symbol Mapping

The mapping of bits into I and Q axes will be Gray-coded and for Reed-Solomon codes is pragmatic that are described in Ref [26] for all constellations.

#### 3.5.3.4 Baseband Pulse Shaping

Prior to modulation, I and Q signals shall be filtered by square-root raised cosine. The roll-off factor α shall be either 0.15 or 0.25. The ideal square-root cosine is defined by the following transfer function H(f):

$$H(f) = 1 \qquad \textit{for} \quad |f| \; \langle \; f_N(1-\alpha)$$

$$H(f) = \left\{ \frac{1}{2} + \frac{1}{2}\sin\frac{\pi}{2f_N}\left[\frac{f_N - |f|}{\alpha}\right] \right\}^{1/2} \qquad \textit{for} \quad f_N(1-\alpha) \le |f| \le f_N(1+\alpha)$$

$$H(f) = 0 \qquad \textit{for} \quad |f| \ge f_N(1+\alpha)$$

Where:

$$f_N = \frac{1}{2T_S} = \frac{R_S}{2}$$

$f_N$ is the Nyquist frequency, and Ts is modulation symbol duration.

### *3.5.3.5 Frequency Domain Equalization (FDE) Scheme*

2-11 GHz systems may operate on NLOS conditions, in which severe multi-path is encountered. Multi-path delay spread is a major transmission problem, which affects the design of modulation and equalization. Delay spread varies with environment and characteristics of transmit and receive antennas. In typical MMDS operating conditions, average delay spread ~ 0.5 µs, but 2% of measured delay spreads > approx. 8-10 µs [15], [16], [17].

Single carrier modulation, with receiver **linear equalization** (LE) or **decision feedback equalization** (DFE) in frequency domain - approximately equal complexity to OFDM, without the power back-off penalty [16], [17], [18], [19].

An adaptive receiver based on frequency domain processing can handle both OFDM and single carrier modulation.

### *3.5.3.6 Single Carrier-Frequency Domain Equalization (SC-FDE) and OFDM*

OFDM transmits multiple modulated subcarriers in parallel. Each occupies only a very narrow bandwidth. Since only the amplitude and phase of each subcarrier is affected by the channel, compensation of frequency selective fading is done by compensating for each subchannel's amplitude and phase. OFDM signal processing is carried out relatively simple by using two **fast Fourier transforms** (FFT's), at the transmitter and the receiver, respectively.

The single carrier (SC) system transmits a single carrier, modulated at a high symbol rate. Frequency domain equalization in a SC system is simply the frequency domain analog of what is done by a conventional linear time domain equalizer. For channels with severe delay spread it is simpler than corresponding time domain equalization for the same reason that OFDM is simpler: because of the FFT operations and the simple channel inversion operation.

The main hardware difference between OFDM and SC-FDE is that the transmitter's inverse FFT block is moved to the receiver. The complexities are the same. A dual-mode system could be designed to handle either OFDM or SC-FDE by simply interchanging the IFFT block between the transmitter and receiver at each end (see Figure 11.)

Both systems can be enhanced by coding (which is in fact required for OFDM systems), adaptive modulation and space diversity. In addition, OFDM can incorporate peak-to-average reduction signal processing to partially (but not completely) alleviate its high sensitivity to power amplifier nonlinearities. SC-FDE can be enhanced by adding decision feedback equalization or maximum likelihood sequence estimation.

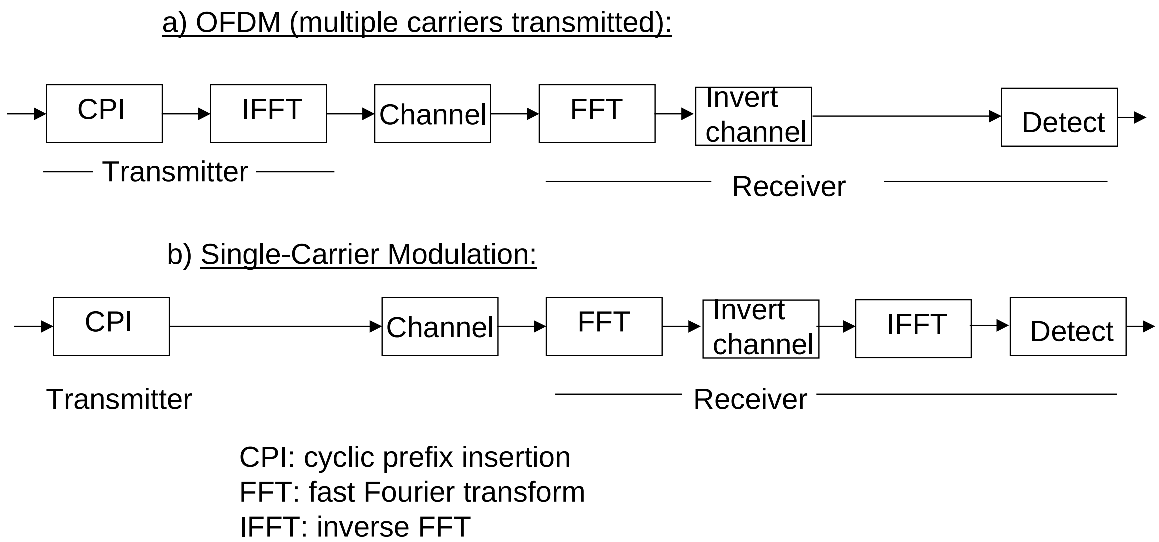



**Figure 11- OFDM and Single Carrier-Frequency Domain Equalization (SC-FDE).**

### *3.5.3.7 Compatibility of Single Carrier (SC-FDE) and OFDM*

Comparable SC-FDE and OFDM systems would have the same block length and cyclic prefix lengths. Since their main hardware difference is the location of the inverse FFT, a modem could be converted as required to handle both OFDM and single carrier signals by switching the location of the inverse FFT block between the transmitter and receiver. Therefore, the coexistence of OFDM and SC-FDE as a "convertible" modem can be feasible (see Figure 12).

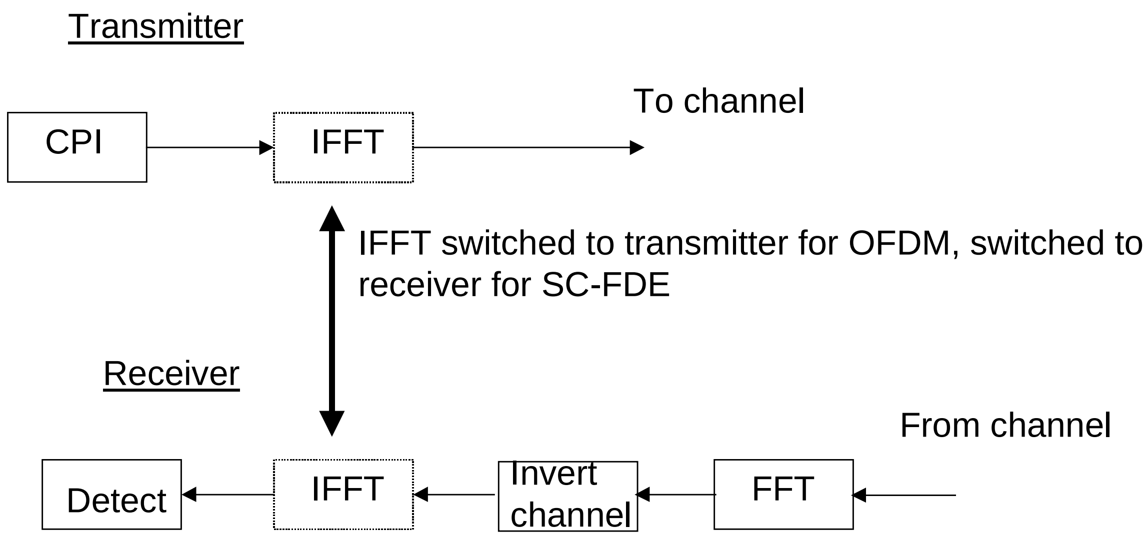



**Figure 12 - OFDM and SC-FDE "Convertible" Modem Approach.**

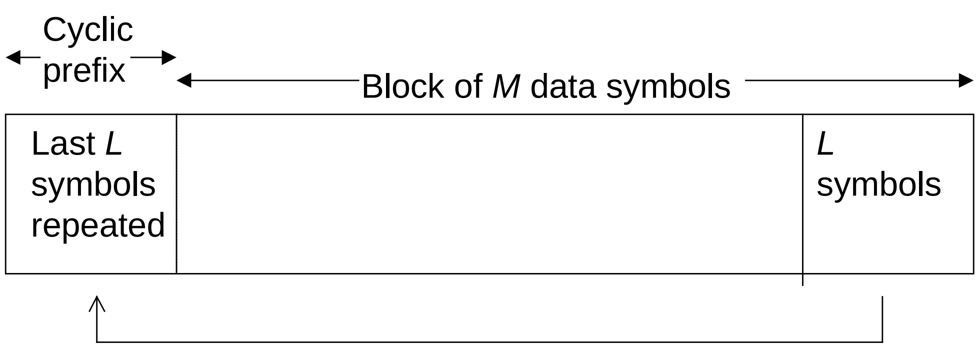



**Figure 13 - Block Processing in Frequency Domain Equalization**

As shown in Figure 13, that the cyclic prefix used in both SC-FDE and OFDM systems at the beginning of each block has two main functions:

- It prevents contamination of a block by intersymbol interference from the previous block.
- It makes the received block appear to be periodic with period $M$, which is essential to the proper functioning of the **fast Fourier transform** operation.

If the first $L$ and last $L$ symbols are identical sequences of training symbols, the overhead fraction is $L/(M+L)$.

For either OFDM or SC-FDE MMDS systems, typical values of $M$ could be 512 or 1024, and typical values of $L$ could be 64 or 128.

### 3.5.4 Framing Structure

The frame format is simple and block based, and accommodates both FDD and TDD operation. Figure 13 illustrates downstream FDD operation, while Figure 14 illustrates downstream TDD operation. Figure 15 illustrates upstream operation.

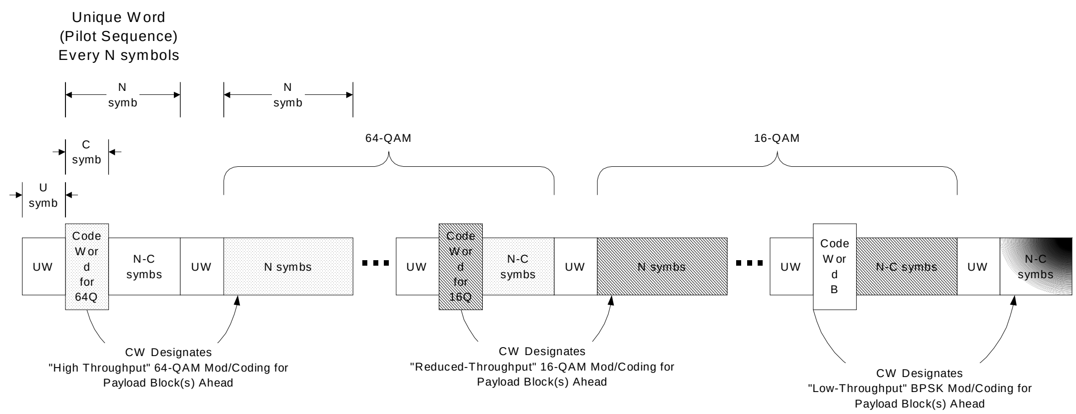



**Figure 13: Framing Structure for FDD system; Downstream.**

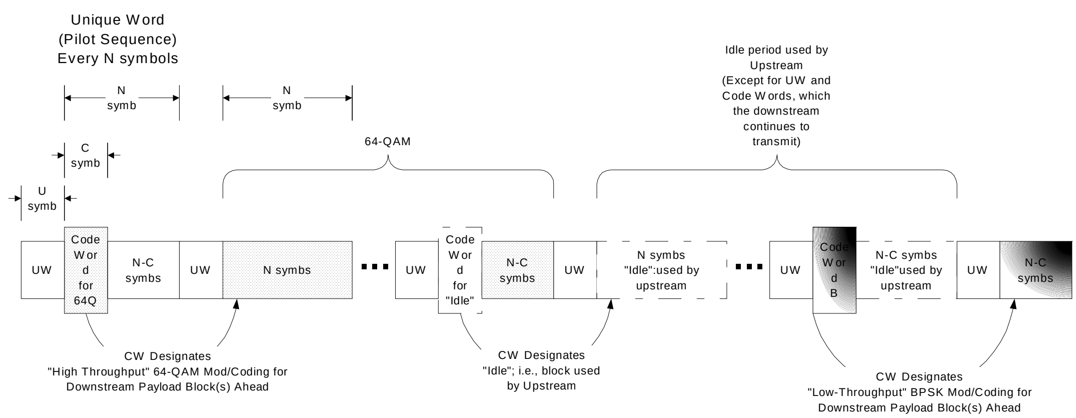



**Figure 14: Framing Structure for TDD system; Downstream.**

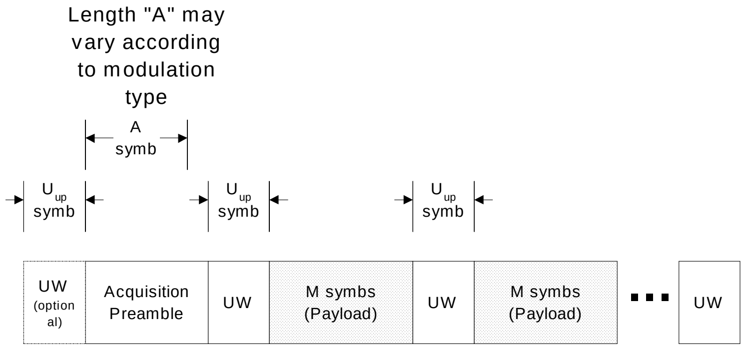



**Figure 15: Framing Structure for Upstream (for simplicity, gaps due to potential downstream TDD framing structure ommitted).**

Several features characterized the frame formatting. Chief among these is the use of **Unique Words**, which occur at intervals of N symbols (on the downstream). Every N symbols, one finds a sequence, called the Unique Word, of length U. This Unique Word serves as a pilot sequence; the symbols from which it is derived are known at both the transmitter and receiver. The Unique Word can be used for carrier phase tracking, channel tracking, and spectral inversion detection on the downstream---even when the payload data is of a modulation format beyond that detectable by a distant subscriber SS terminal. By its periodic nature, it also is very useful in expediting initial acquistion, and also re-acquisition, in the case of a deep fade, when the SS receiver might lose lock. Note that by this structure, the downstream is considered to be 'continuous' in nature, and the Unique Words aid all SS receivers in maintaining lock, even when tranmitted packets are not meant for them.

The symbol sequence within the Unique Word is selected to have good correlation properties (i.e., its autocorrelation resembles an impulse response), so that the transmission channel response can be identified with high accuracy. For easy demodulation at distant terminals (and optimal correlation sequence choice), the sequence is derived from a QPSK alphabet.

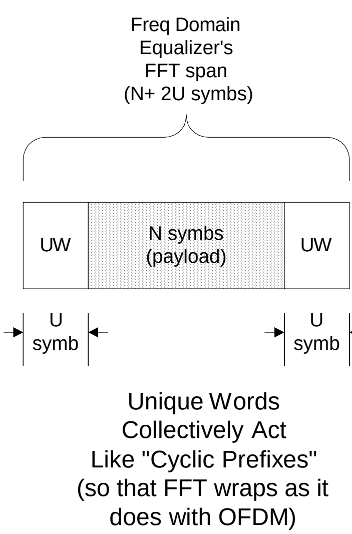



**Figure16: Structure of equalizer which can be exploited by frequency domain equalization.**

As Figure 16 illustrates, the Unique Word always 'sandwiches' data of length N on the downstream or length M on the upstream. This property is particularly useful for frequency domain equalizers, since these Unique Word segments on each end of the M point data serve as 'cyclic prefixes' so that the FFT used in the frequency domain equalizer 'wraps'. The FFT would be taken over the duration of the Unique Words on both ends of a N (or M) data point data payload. Note that this usage is very similar to the cyclic prefix used in OFDM. Note however, that here, the cyclic prefix is used as both pilot symbols and a cyclic prefix. This may result in higher spectral efficiencies than may be achieved with OFDM.

What is particularly attractive about the Unique Words is that, post-equalization, the known unique words can be cross-correlated against the Unique Word block regions---to to estimate the channel (delay spread) estimation error, and feed that back into channel response updates (used to equalize future blocks). A particular advantage over OFDM-type approaches is that these pilot symbols span the whole frequency spectrum, and thus are less affected by the notches due to multipath (frequency selectivity) of the channel.

Note however, that this approach does not presuppose use of a frequency domain equalizer.

A DFE or other time-domain equalizer can also exploit these known pilot sequences to improve both their channel estimation and equalization performance. In particular, the use of periodic known sequences can reduce the propagation of decision errors within a DFE.

Another feature of the proposed framing is that Code Words of length C are used at intervals to indicate the modulation type of the data payload, which will follow the next Unique Word. In other words, there would be a Code Word for 64 QAM, another for 16 QAM, another for QPSK, and another for BPSK, etc. Note that the intervals for a given modulation may span several Unique Words, and the minimum duration of a given modulation type, i.e., the spacing between Code Words, might be a system parameter.

Also, if the TDD option is exercised, then another Code Word, called 'Idle Channel' is used, to tell the SS terminal that the next data segment will be idle. The upstream may subsequently be utilized by one of the subscriber users, which have requested bandwidth. In the TDD mode, the Base Station will send idle slots, even when no upstream bandwidth grants have been made, so that SS users may signal grant requests.

## *3.6 UpStream Channel*

### 3.6.1 Upstream Multiple Access

The upstream multiple access method shall be TDMA.

### 3.6.2 Upstream Modulation Format

The upstream modulation shall be BPSK, QPSK, 16QAM or 64QAM.

### 3.6.3 Upstream Randomization, Channel Coding & Interleaving, Symbol Mapping And Baseband Shaping

The upstream channel has processing units similar to those described for the downstream. However, greater flexibility in packet transmission is allowed. The subscriber stations are transmitting only after receiving some configuration information from the base station through MAC messages. Several different configurations can be adjusted on the upstream channel on a burst-to-burst basis. The upstream payload is segmented into blocks of data designed to fit into the proper codeword size (including Transmission Convergence sublayer, TC, header). Note that payload length may vary from burst to burst.

#### *3.6.3.1 Randomization for Spectrum Shaping*

The upstream modulator uses a randomizer using LFSR with connection polynomial

$X^{15} + X^{14} + 1$, with a 15-bits programmable seed. At the beginning of each burst, the register is cleared and the seed value is loaded. The seed value is used to calculate the scrambler output bit, obtained as the XOR of the seed with first bit of DATA of each burst (which is the MSB of the first symbol following the last symbol of the preamble).

#### *3.6.3.2 FEC schemes for the upstream channel*

Consistent with the structure of the 802.16 MAC forward error correction code schemes which support both **Block Turbo Coding** (TPC) and Concatenated Reed-Solomon+Convolutional coding will be employed. In addition, the provision of suppressing all FEC and operating using the ARQ mechanism in the 802.16 MAC for error control will be included.

#### 3.6.3.3 Interleaving for the upstream channel

Interleaving is applied for the upstream channel only with BTC FEC scheme.

#### 3.6.3.4 Baseband Pulse Shaping

Prior to modulation, the I and Q signals shall be filtered by square-root raised cosine. The roll-off factor α shall be either 0.15 or 0.25. The ideal square-root cosine is defined by the following transfer function H(f):

$$H(f)=1 \qquad for \quad |f| \;\langle\; f_N(1-\alpha)$$

$$H(f)=\left\{\frac{1}{2}+\frac{1}{2}\sin\frac{\pi}{2f_N}\left[\frac{f_N-|f|}{\alpha}\right]\right\}^{1/2} \qquad for \quad f_N(1-\alpha)\le|f|\le f_N(1+\alpha)$$

$$H(f)=0 \qquad for \quad |f|\ge f_N(1+\alpha)$$

Where:

$$f_N=\frac{1}{2T_S}=\frac{R_S}{2}$$

$f_N$ is the Nyquist frequency, and Ts is modulation symbol duration.

## 3.7 System Capacity and Modulation Efficiency

Table 7 shows the BWA PHY with Downstream and Upstream modulation schemes and the corresponding system capacity and Bits per sec./ Hz. The aggregate transmission bit rate is optimized based on several constraints. These are:

- The allocated channel bandwidth;
- The modulation level;
- The spectrum shaping filter bandwidth with roll factor of α = %0.15 to %0.25;
- The FEC coding scheme (Reed-Solomon (n, k) over GF($2^8$));
- The requirement of upstream time tick for the Mini-slots burst duration; and
- Processing power limitation of available chips to be used.

Table 7 presents an example of achievable system capacity where all coding and FDE overhead budget is being included.

**Table 7: An Example Of System Capacity Objectives.**

| Channel Spacing | Downstream Transmission | | Upstream Transmission | |
|---|---|---|---|---|
| | Rate (Mb/s) | | Rate (Mb/s) | |
| | (16 QAM) 3.38 bps/Hz | (64 QAM) 5.07 bps/Hz | (QPSK) 1.46 bps/Hz | (16 QAM) 2.92 bps/Hz |
| 3.5 MHz | 11.02 | 16.54 | 4.77 | 9.54 |
| 5 MHz | 15.72 | 23.57 | 7.44 | 14.88 |
| 6 MHz | 18.82 | 28.21 | 8.93 | 17.86 |
| 7 MHz | 22.03 | 33.03 | 9.52 | 19,05 |

## *3.8 RF Propagation Characteristics*

The channel model is highly dependent upon the RF network topology, RF bands, terrain category and the various RF propagation impairments (see Appendix B)

### 3.8.1 RF Network Topology

The RF Network topology may include:

- **Mega-cell topology**: up to 50 km Tx\Rx separation,
- **LOS propagation** characteristics.
- **Directive antenna at both BS and SS** will result in negligible Co-Channel-Interference (CCI).
- **Multi-cell topology:** cell radius is typically less then 10 km.
- In a **Frequency re-use** cellular system,
- A cell may be subdivided into **multiple sectors**.

### 3.8.2 RF bands and Channelization

- Frequency range: 2 to 11 GHz
- Channelization: support 1.75, 3.5 and 7 MHZ using ETSI frequency masks (3.5 GHz systems) and 1.5, 3, and 6 using MDS mask (2.5 MHz systems).
- Supporting 0.25 to 7 MHz when using other masks and frequency plans.

### 3.8.3 Terrain category:

- **Urban Area.**
- **Suburban Area:** May be further divided into 3 types as proposed in [16b] –
  - Type A: Hilly/moderate-to-heavy tree density
  - Type B: Hilly/light tree density or flat/moderate-to-heavy density
  - Type C: Flat/Light tree density.

### 3.8.4 RF propagation impairments:

- Path Loss
- Fading (large scale – due to shadowing, small scale- due to multi-path).
- CCI and ACI
- Worst case fading bandwidth and maximum Doppler shift.

### 3.8.5 Minimum Performance Specifications

Based on the measurements given in Appendix B the channel model must meet the following requirements: Maximum time delay spread of **12 μsec**. The system should withstand a Doppler shift of more then **10 Hz**.

## *3.9 Antenna Systems*

### 3.9.1 Application of Smart Antenna

The PHY layer shall support future application of smart antenna for primary feature of providing the ability to track the line of sight target within a predetermined angle of uncertainty. Typically, one would expect 3 or more degrees of tracking. This active tracking capability of smart antenna will potentially provide better coexistence and will optimize the antenna pattern (transmit where the subscriber is located)

### 3.9.2 Antenna Diversity

Multiple antennas can be used at the transmitter and/or receiver to provide added dimension to the model.

When multiple antenna diversity (so called Multiple-Input/Multiple-Output; MIMO) is compared with a Single-Input/Single-Output (called SISO) technique, it is shown in performance that it can improve the capacity of the fading wireless channel regardless of the modulation techniques utilized. It is applicable to Single Carrier (SC) modulation. The benefits, however, of using space diversity should be examined against its implementation complexity and economic factors.

# 4 Comparative Analysis of OFDM vs SC- FDE:

Multi-carrier techniques like OFDM split a high-rate data-stream into a number of lower rate streams that are transmitted simultaneously over several sub-carriers. That is, creating several parallel narrow-band sub-channels. Therefore, while the symbol duration increases for the lower rate parallel subcarriers, the relative amount of dispersion in time caused by multipath delay spread is decreased. Intersymbol interference (ISI) is eliminated almost completely by introducing a guard time in every OFDM symbol. In the guard time, the OFDM symbol is cyclically extended to avoid Intercarrier interference (ICI) [3].

With OFDM a number of parameters are up for consideration, such as;

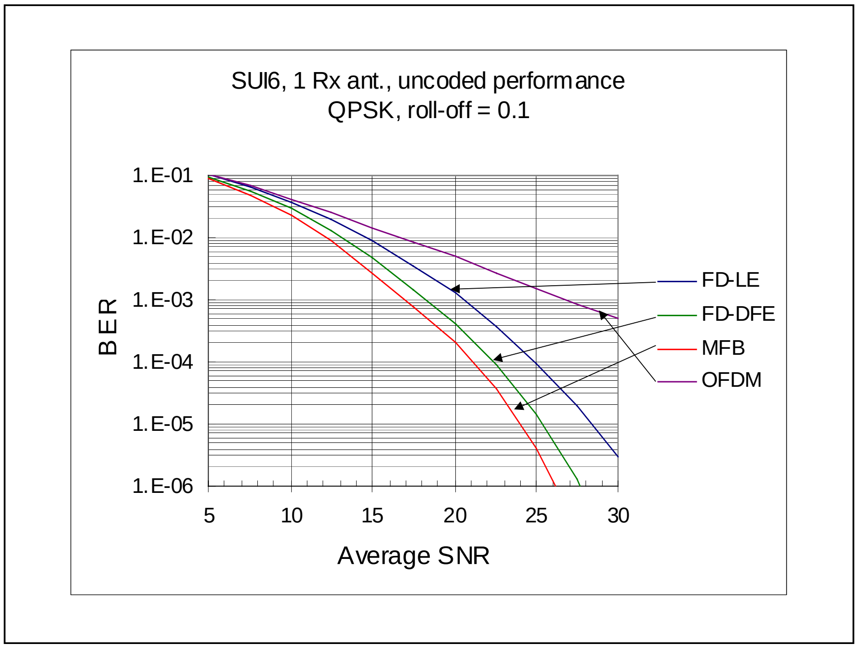
- the number of subcarriers,
- guard time,
- symbol duration, subcarrier spacing, modulation type per carrier, and
- the type of FEC coding.

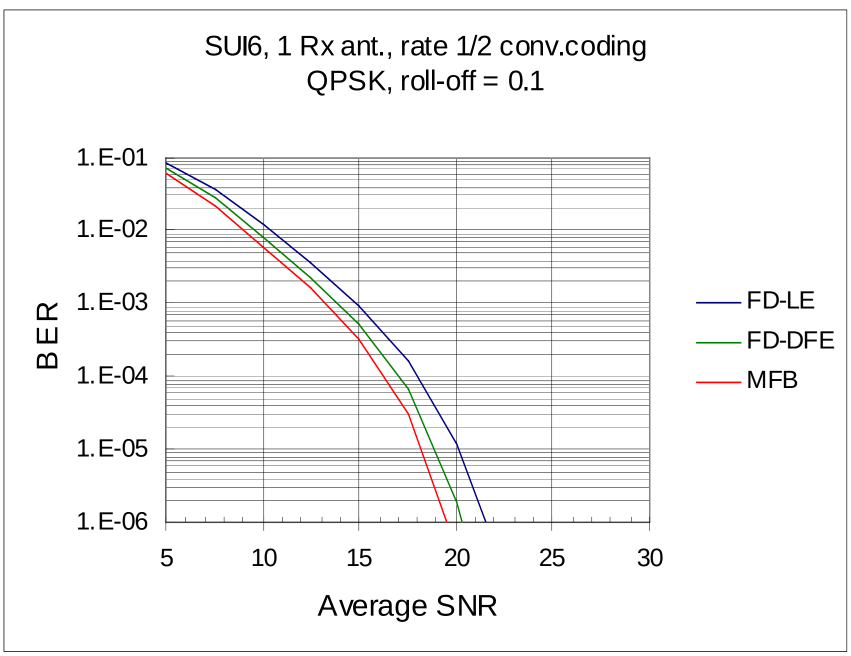
It has been shown that OFDM systems offer no advantages over Single Carrier modulation forms in severe multipath – see [21], [22], [23], and [24]. In particular [21] compares the performance of a coded OFDM signal in multipath environments and shows that single carrier systems offer superior performance in all but a very limited number of cases. In addition, in Section 4.1, we will present some of the results that will appear in [20] showing that Single Carrier with Frequency Domain Equalizer (SC–FDE) techniques can outperform OFDM in performance when the channels of communication suffer from deep multi-path fading in addition to the usual AWGN noise.

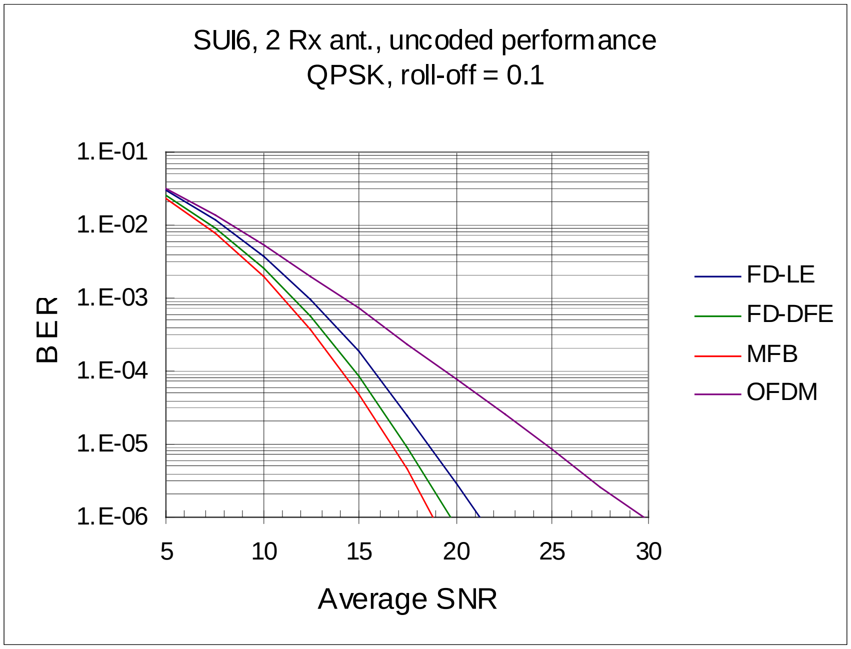
In a Single-Carrier system, the implementation complexity is dominated by the requirement of equalization, which is necessary when the delay spread is larger than about 10% of the symbol duration. In OFDM, the equalization is done by amplitude and phase-correction of each subchannel. The complexity of both modulation systems can largely be determined by FFT and inverse FFT requirements. This complexity in fact can be reduced by not requiring full multiplication, but rather phase rotations, which can be efficiently implemented by the CORDIC algorithm [3]. In fact, the technology is advancing rather rapidly in this area.

We should emphasize that OFDM imposes stricter constraints on the analog blocks due to its large **Peak-to-Average Power Ratio (PAPR)** characteristics and its sensitivity to carrier frequency offset and phase noise. Thus, to alleviate the time-varying frequency offsets between transmitter and receiver must use an accurate AFC circuitry, otherwise the sub-carriers will no longer be **orthogonal**. Synchronization of a multi-carrier scheme is much more difficult than a single carrier system. In addition, OFDM with a large number of sub-carriers, the combined signal has a very large PAPR and to maintain linearity over the range, the power amplifier will require back-off by as much as 10dB [7].

## 4.1 Performance in a Fading Channel




SUI6, 2 Rx ant., uncoded performance
QPSK, roll-off = 0.1
BER
1.E-01
1.E-02
1.E-03
1.E-04
1.E-05
1.E-06
5
10
15
20
25
30
Average SNR
FD-LE
FD-DFE
MFB
OFDM


SUI2, 2 Rx ant., uncoded performance
QPSK, roll-off = 0.1
BER
1.E-01
1.E-02
1.E-03
1.E-04
1.E-05
1.E-06
5
10
15
20
25
30
Average SNR
FD-LE
FD-DFE
MFB
OFDM

### *Summary of the above comparative results:*

Monte Carlo BER performances of SC frequency domain equalizers over multipath **Rayleigh fading**:

- Frequency domain linear equalizer (FD-LE)
- Frequency domain decision feedback equalizer (FD-DFE) without decision errors
- Matched filter bound (MFB), representing the ultimate (hypothetical) ideal performance assuming perfect capturing of multipath energy and no loss due to intersymbol interference (ISI)
- Uncoded OFDM performance also provided in the first 4 figures

**Channel Model:** SUI6 with Rician factor K=0 (i.e., Rayleigh) for all paths. No diversity.

**Simulation conditions:** QPSK with 0.1 roll-off, 10,000 fading channel realizations, 512 point FFT, quasi-static fading, no channel estimation errors. Sufficient number of feedback taps (feedback filter is as long as the channel span). The BER for each channel realization is computed using analytical formulae.

BER is given as a function of the per-branch SNR averaged over Rayleigh fading.

Uncoded OFDM results are obtained by averaging the BERs on individual tones. Since each tone is a complex Gaussian random process regardless of the channel model, the overall performance for any delay profile is the same as the BER averaged over flat Rayleigh fading (so, we can also regard the uncoded OFDM results as the flat fading performance for any system).

It is not surprising to see that all the single carrier frequency domain equalizers outperform OFDM in an uncoded case. The equalizers automatically exploit the so-called “inherent” or “built-in” multipath diversity (also called frequency diversity). OFDM can exploit multipath or frequency diversity only through coding across the tones.

**Performances for SUI-2 and SUI-6** are similar (slightly better for SUI-6 because of the higher degree of frequency selectivity).

**In the uncoded case**, FD-DFE outperforms FD-LE by about 2 to 4 dB (at BER below 0.001) without diversity. The gap increases with average SNR because of the “noise enhancement” effect. Conversely, when thermal noise is more dominant, the FD-LE will try less to invert the null, thereby causing less noise enhancement.

**With diversity**, the performance difference between FD-DFE and FD-LE is reduced to about 0.8 to 1.8 dB. Two effects interplay to reduce the difference: (i) MMSE diversity [Cla98] i.e., the receiver automatically sets antenna weights to either provide diversity gain or reduce ISI, depending on which gives the smaller mean-square error outcome. (ii) With diversity, the receiver achieves the desired BER at a lower average SNR, and therefore the noise enhancement effect described above is less significant.

From the above finding, it is suspected that the gap between FD-DFE and FD-LE performances should also be reduced when coding is used (again, because with coding, the receiver operates at a low SNR range and this should result in lower noise enhancement). The coded performance verify this point**. As you can see, even without diversity, FD-LE performs only about 1 dB worse than FD-DFE.**

## *4.2 Phase Noise Requirements*

An exact closed form expression relating the combined phase noise performance of a set of oscillators used for the frequency translation of a received QAM signal, and their effects on the BER of the demodulated product, is very hard to determine. One obstacle is the fact that an exact analytical representation of phase behavior that considers all of the noise processes in each oscillator is extremely complicated. An additional impediment is that there is no exact input/output relationship for the processed phase noise as it travels through amplifiers, PLLs, downconverters and other blocks. Therefore the results that follows are based on empirical data and basic analytical models of phase noise and frequency synthesis/translation.

Total RMS phase error for a dual-conversion, MMDS band receiver has been determined to be as shown in Figure 17. This receiver was designed utilizing low cost silicon technology with an inexpensive TCXO as the reference for the LO synthesizers.

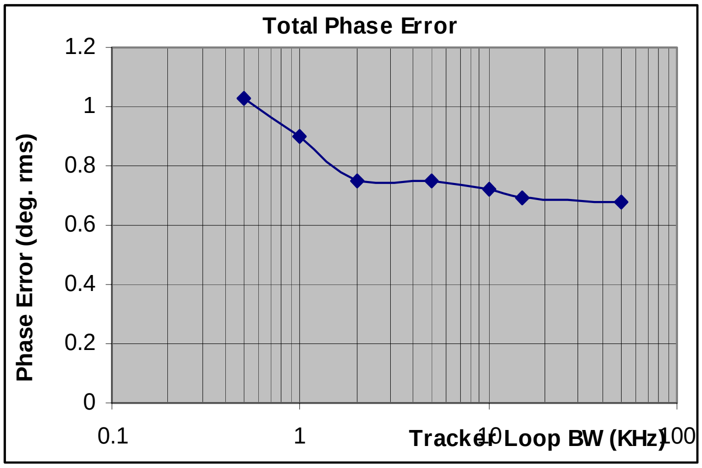



**Figure 17: Total RMS Receiver Phase Error**

Figure 18 illustrates the BER performance for a 64QAM demodulator as a function of residual SNR due to phase noise in the receiver.

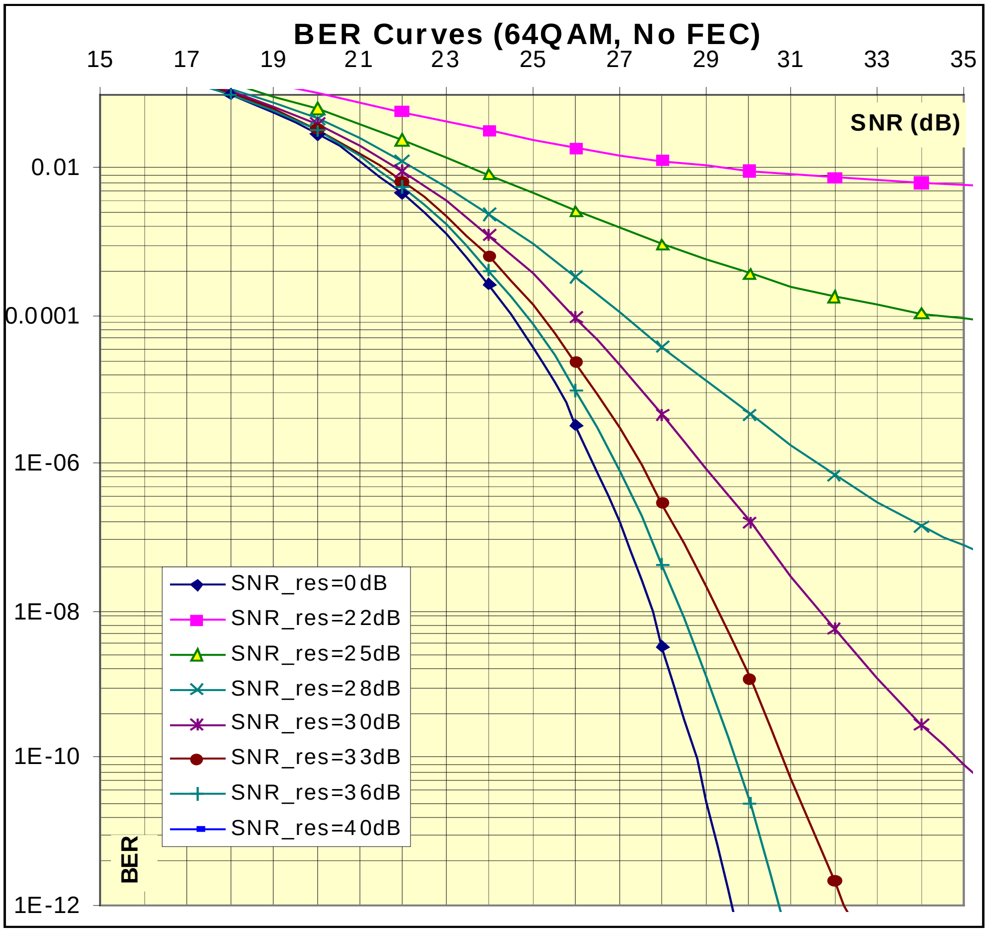



**Figure 18: BER vs SNR for various residual SNR**

The resulting BER for an OFDM receiver utilizing the same low-cost LO chain is shown in Figure 19. The performance degradation for the same LO phase error is approximately XX dB.

**Figure 19**

**Plot of OFDM BER vs. residual SNR**

**(to be created)**

### 4.3 Amplifier Linearity Requirements

One of the most significant contributors to the cost of a BWA radio is the **power amplifier** (PA). The following simulation data was generated to determine the linearity requirements for a PA with single carrier versus multi-carrier modulation **in order to better understand the impact of either modulation scheme on radio cost.**

## Model Assumptions

The data presented below was generated based on a PA model derived from empirical measurements of the AM-AM and AM-PM characteristics of a +30dBm P1dB, HBT PA operating at 2.5GHz. The model used in these simulations is a 5$^{th}$ order polynomial curve-fit of the empirical data.

Figure 20 shows the output of a GaAs power amplifier stage driven by a 64QAM signal. The output is compliant with the spectral mask defined in 47CFR21.908 that defines the characteristics of 2.5 GHz MMDS transmitters. The amplifier is operated at approximately 9 dB below its rated 1dB compressed output power.

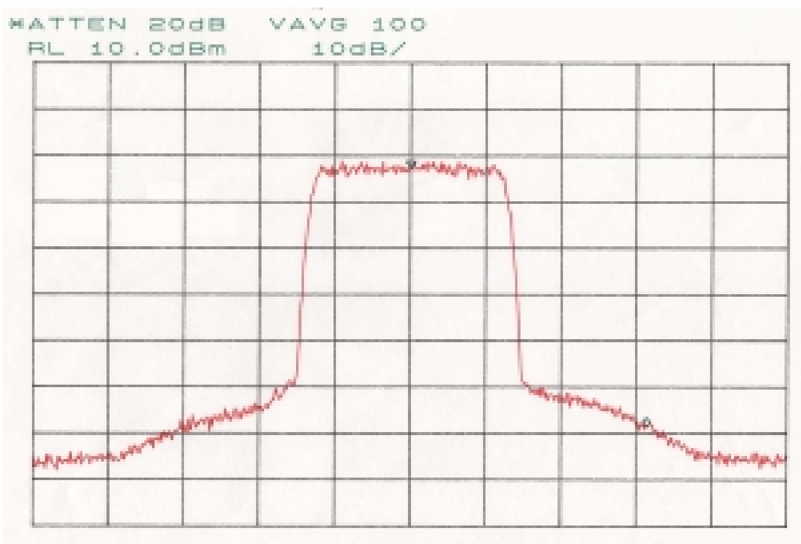



**Figure 20: 64 QAM at 9 dB back off**

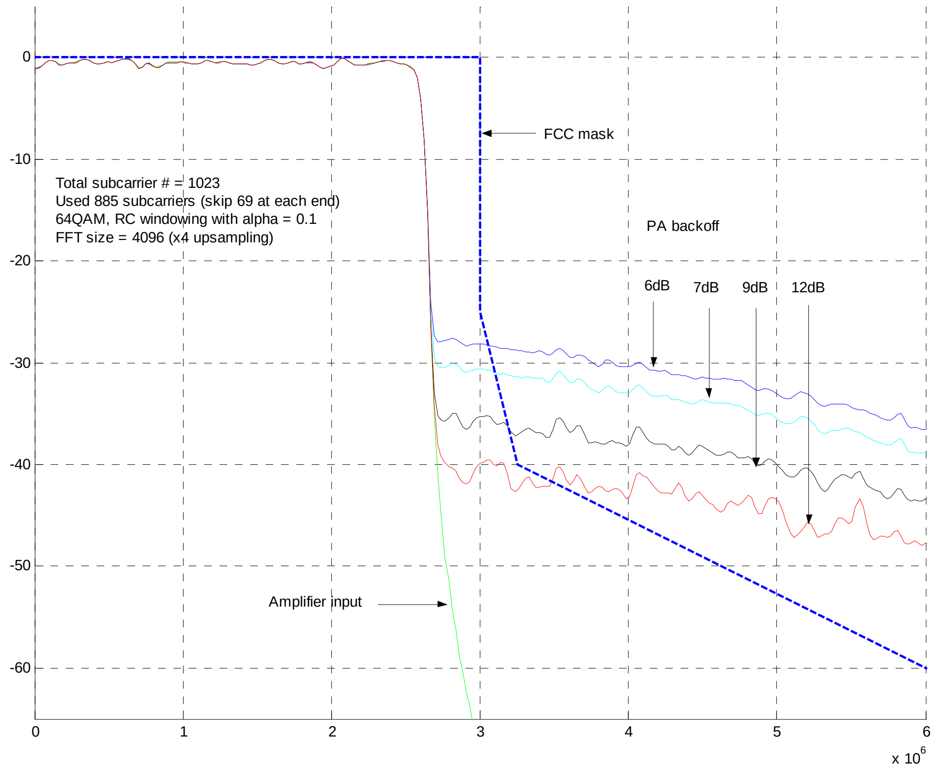



**Figure 21: OFDM at various back-ff levels**

Figure 21 shows the simulated output of a similar GaAs power amplifier driven with a 1023 tone OFDM signal with 64 QAM modulation. The FCC mask is shown by the dotted line and the inability to comply with the FCC mask at even 12 dB back-off is apparent. The accuracy of the commonly given value of 14 dB back-off for compliance with the FCC mask is evident in Figure 21.

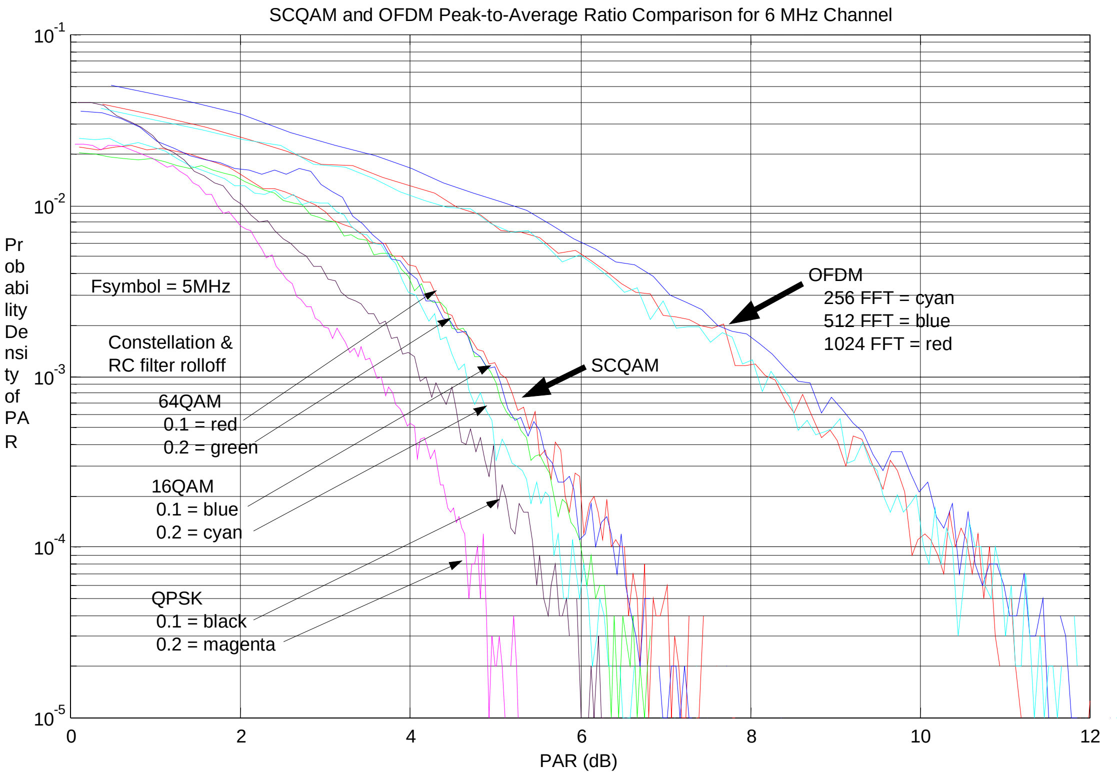



**Figure 22: Power Distribution PDF**

Figure 22 shows a PDF (probability distribution function) of the instantaneous power of the input signal to the PA. This plot shows that at $10^{-4}$ probability, there is approximately 4.5dB more energy with an OFDM waveform as opposed to a single carrier 16QAM waveform.

The OFDM waveform plot was windowed with a raised cosine roll-off factor of alpha = 0.1. For the 1024 subcarrier case we use a 4096 point FFT and eliminate 138 subcarriers to allow for some guard-band. The data is presented with the FCC MMDS spectral mask superimposed as a reference.

## 4.4 Summary and Conclusions:

- For severe multipath, Single Carrier QAM with simplified frequency-domain equalization performs at least as well as OFDM (better for uncoded systems).
- Frequency domain linear equalization has essentially the same complexity as uncoded OFDM, with better performance in frequency selective fading, and without OFDM's inherent backoff power penalty.
- A "Convertible" frequency domain receiver structure can be programmed to handle either OFDM or Single Carrier.
- Downlink OFDM / uplink single carrier may yield potential complexity reduction and uplink power efficiency gains relative to downlink OFDM / uplink OFDM.

| Advantages of SC and OFDM | |
|---|---|
| **Single Carrier** | **Multi-carrier (OFDM)** |
| Sensitivity (margin):<br>Less Affected by Freq Selective Fading (spectrum notches averaged)<br>Reduced overhead<br>Less pilots & No guard interval<br>'Lighter' coding possible | Simple Equalization<br>Tx diversity ostensibly easier |
| IC Complexity--Less Memory (data buffering) | Robustness at low SNR<br>Avoid DFE (use pilots)<br>PAPR unaffected by modulation order |
| Reduced RF expense:<br>Reduced Phase noise sensitivity<br>Reduced Freq Regist Reqments<br>Reduced PA Backoff<br>64QAM: 1e-3 env prob 3 dB less<br>QPSK: 1e-3 env prob 4-4.5 dB less<br>Important at edge of cell<br>Single Carrier can use Freq domain equalizer | IC Complexity--Less logic |
| Smaller packet granularities | Automatically integrates multipath<br>-- but not coherently |
| FIFO Advantages<br>Throughput (Queueing Theory)<br>Reduced MAC complexity (vs OFDMA) | Single Frequency Networks (OFDMA) |

# 5 Main Features and Benefits of the Proposal

This PHY proposal for the IEEE802.16.3 air interface standard presents basic features that meet all the requirements identified in [1], under the critical constraint of low-cost solution to the target markets. A migration approach that will enable an exploitation of current industry standards and systems is indicated. Further advanced features are recommended to improve the performance in a number of ways. Benefits of the proposed PHY and its unique features are outlined below:

1) borrowing key features from **well-established wireless standards**
2) **Adaptive Modulation and Coding -** allowing flexible bandwidth allocation to maximize spectral

3) **Mature and well-proved technology** - build on the footprint of the evolving cable modem technology and efficiency and overall system capacity. For example, near SS can use higher modulation scheme with high coding rate, while far SS or other SS experiencing severe interference profile can use more robust QPSK modulation. AMC exhibits more than 20dB gain relative to non-adaptive schemes (see 2000-10-30 IEEE 802.16.3c-00/39).

4) **Flexible Asymmetry –** supporting high degree of flexibility between deliver upstream and downstream via duplexing schemes; e.g., FDD and TDD.

5) **Scalability -** supporting IP, ATM and MPEG-2 packets with variable-length Packet Data Units (PDU). High immunity to RF impairments and radio equipment impairment. The proposal is based on Single Carrier M-QAM that is less sensitive than OFDM to RF impairments such as: linearity of power amplifier, frequency instability, phase noise, synchronization errors, Doppler spread etc.

6) **Advanced Coding Schemes** – based on Reed-Solomon concatenated with Convolutional codes or Block Turbo Coding (BTC). Both coding techniques provide a good solution for variable packet length with high code rates.

7) **Reduced System Delay** - using advanced Block Turbo Coding that can eliminate the need for a large interleaving. **Reduction in cost, complexity and network architecture simplification**. Advanced single carrier modulation based on M-QAM combined with adequate equalizing techniques and BTC reduces the overall system complexity. Note that a system using SC in uplink and OFDM in downlink is a possible avenue and might reduce subscriber unit complexity and also its power amplifier cost.

**8) Easy Migration from simple SC to SC –FDE:** that meet more demanding channel impairments and interference at increased spectrum efficiency.

**9) An easy migration path to diversity receiver and multiple-input/multiple-output (MIMO):** Improving the robustness to interference, channel impairments and radio equipment impairment for applications requiring additional link margin.

# 6 Similarity to other standards:

The proposed PHY is similar to some extend with TG1 PHY (supporting TDMA multiple access, both TDD and FDD, QPSK/m-QAM, and FEC coding), to some degree with DOCSIS (supporting TDMA multiple access, QPSK/m-QAM, and FEC coding).

# Statement on Intellectual Property Rights:

All team member companies have read this document and the IEEE patent policy and agree to abide by its terms.

# 7 References: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# APPENDIX A: Compliance with the Evaluation Criteria:

| | |
|---|---|
| **Meets system requirements** | This proposal is believed to meet the requirements described in the current version of TG3 FRD. |
| **Channel spectrum efficiency** | The average of bps /Hz in a typical deployment (TDD or FDD) is about 2.80 bps/Hz. In FDD mode, the spectrum efficiency of the system ranges from 1.36 for QPSK and to 2.7 bps/Hz for 16 QAM modulation for the Uplink. For the Downlink, 3.13 for 16-QAM and to 4.7 bps/Hz for 64 QAM modulation. |
| **Simplicity of implementation** | The major functions of the proposed PHY (i.e., QAM, FEC and OFDM) are well known or they are becoming available technologies and do require complex implementations. |
| **SS cost optimization** | Similarity that exists between this proposal with other standards mentioned above, will facilitate the availability of chip-sets to be used for the SS with lower cost. SC AT THE SUBSCRIBER UNIT TRANSMITTER REDUCES ITS POWER AMPLIFIER COST |
| **BS cost optimization** | The use of OFDM at the BS can be a drawback from the complexity and PA Back-off requirements, but this feature will be advantageous for future addition of Smart antenna capability to the system. |
| **Spectrum resource flexibility** | The proposed PHY can be scaled to any channel spacing. Modem bit rate can be easily modified to support 10 to 40 Mbps. |
| **Channel Rate Flexibility** | This data rate scalability can be obtained by changing FEC code rate and modulation scheme. The changes will have to meet the specified QoS in the FRD. |
| **System service flexibility** | The proposed PHY in conjunction with MAC layer will support various services defined within FRD that may require variable data rates and with different QoS requirements. |
| **Protocol interfacing complexity** | The proposed PHY will efficiently carry variable length packets and will comply with the delay and speed requirements by upper protocol layers. |
| **Reference system gain** | The system gain for 16QAM, 3.5GHz band, and 3.5 MHz BW Gain=103.5 and<br>The system gain for 16QAM, 10.5GHz band, and 3.5 MHz BW Gain=96.5. |
| **Robustness to interference** | The proposed PHY uses powerful coding scheme with interleaving and good interference rejection capabilities. |
| **Robustness to channel impairments** | The multi-path robustness of FDE an important capability of the system and it reduces (almost removes) the impact of small and large scale fading. |

| | |
|---|---|
| **Robustness to radio Impairments** | The proposed PHY has the capability to support multiple data rates, modulations, and power control circuitry. When the radio channel attenuation becomes severe, then through the MAC control loop, the PHY system can re-adjust the transmission level to the appropriate level to keep the good quality of service intact. |
| **Support of advanced antenna techniques** | The proposal supports the need for advanced antenna techniques such as smart antenna into the standard. This feature, in conjunction with OFDM can be powerful feature for the system. |
| **Compatibility with existing standards and regulations** | This proposal is compliant with ETSI, FCC, and other existing standards and regulations as provided in Table 2. |

# APPENDIX B: Channel Model For BWA PHY Systems

## B.1 Deployment Models

There are three models that describe the deployment of high speed, fixed wireless broadband Internet systems: the large line-of-sight (LOS) cell, the large non-line-of-sight (NLOS) cell, and clustered NLOS small cells. Large LOS cells are used in almost one hundred systems in a variety of terrains around the world. The other two cell models are under development but have not been commercially deployed yet.

The large LOS cell deployment is characterized by tall base station antennas with heights of 200 meters and more, and Subscriber Station (SS) antennae mounted on the roof or on poles on the roof at heights between 5 and 10 meters. In the United States, the licensed frequencies set the cell site radius to 35 miles, however SS sites have been operated out as far as 50 miles. In a number of cases, multiple cells are deployed to service a metropolitan area. The cells typically face inwards towards the market and are situated so that with directional SS antennae, there is little cell-to-cell interference.

The NLOS large cell deployments will be an extension of the LOS deployments. The two models share a common architecture because in many cases the large LOS cell will be adapted to accommodate NLOS customer sites. LOS and marginal LOS locations will continue to work out to the licensed limit of 35 miles. Because of link budget constraints, **NLOS locations will be restricted to distances of about 20 miles**. As in the LOS model, inter-cell interference is dependent on Base Station placement and SS antenna directivity. Systems are designed to minimize the interference.

Small cell site deployments are being considered for regions where the capacity of large cells is not enough to serve the market or where the terrain causes a lot of shadowed areas. Small cell Base Stations use PCS like towers with antenna heights that range from 15 to 40 meters. The target height of the SS antenna is 2 to 5 meters. Small cells are clustered together to provide service over a large area. The cells reuse the licensed frequencies to increase the overall system capacity. The reuse pattern, antenna patterns, and channel models are used to determine the inter-cell interference seen by small cells.

## B.2 RF Channel Models

### B.2.1 Large LOS Cells

The RF channel model for large LOS cells is well understood. The attenuation verses distance formula follows the free space equation of:

$PG_{los(dB)} = 20\log(4\pi d/\lambda)$

Where $d$ is the distance and $\lambda$ is the wave-length. This model applies to both the upstream and downstream paths. Seasonal variations are only a few dB. Multipath in LOS is less **than 5 mirco seconds** at power levels of –6 dB from the primary signal in a large percentage of installations using moderately directional antennae of $22^0$. Less directional antennae see stronger multipath. Rayleigh fading is not a significant factor in this environment. The fading is flat in channel widths that are 2 MHz and below.

### B.2.2 Large NLOS Cells

The RF channel model for large NOS cells is based on the LOS channel model. In both cases the bulk of the transmission path is characterized by the free space attenuation formula because of the height of the base station antenna. As a result the free space signal is delivered to a relatively small area near the customer location. The signal then under goes additional attenuation that comes not from distance but from the bulk absorption of few trees and refraction from buildings. This attenuation characterized in unpublished work can be modeled as a near neighborhood bulk attenuation of between 8 and 30 dB. Because of the influence of foliage, seasonal variations in the NLOS signal levels can vary by the amount of near neighborhood bulk attenuation allowed for in the formula. The modified NLOS attenuation formula is given by:

$PG_{nlos(dB)} = 20\log(4\pi d/\lambda) + n$

Where $n$ is the near neighborhood bulk attenuation factor. The relative power of multipath signals in NLOS cells can be greater because the primary signal may be attenuated more by the bulk attenuation than the reflected signal. However, the size of the multipath delay is similar to that of the LOS case. The characterization of large cell NLOS signals is an area where more field studies are needed. The attenuation model presented here applies to both directions of a two-way system. In the downstream, Rayleigh fading is dependant on SS antenna heights. At 4 meters and above the effects of Rayleigh fading are small. Upstream measurements are not available for this model but it is safe to assume that Rayleigh fading is the same or smaller than that seen in the downstream.

### B.2.3 NLOS Small Cells

The NLOS small cell channel model is based on the IEEE 802.16.3 paper (http://ieee802.org/16/tg3/contrib/802163c-oo_49r2.pdf). It describes the path for small cell with radius' out to 10 km, Base Station antenna heights up to 40 meters and SS antenna heights between 2 and 8 meters. The mean path loss is given by:

$$\mathbf{PL_{sc(dB)}} = A + 10\gamma\log(d/d_0) + S + \Delta PL_f + \Delta PL_h$$

Where $A = 20\log(4\pi d_0/\lambda)$

$\gamma = (a - bh_b + c/h_b)$ {$a$, $b$, and $c$ are constants given in the IEEE paper that depend on terrain and $h_b$ is the Base Station antenna height in meters}

$d_0$ = 100 meters

$S$ = lognormal shadow fading (typically set to 10 dB)

$\Delta PL_f$ = Frequency correction = $5.7\log(f/2000)$ $f$ in MHz

$\Delta PL_h$ = SS antenna correction = $-10.8\log(h_{cpe}/2)$ $h_{cpe}$ in meters.

At the SS antenna heights considered the Raleigh fading factor**, K, is found to be 0**, which means there are typically deep, fast fades from the mean path loss. Multipath is modeled as a delta function at time zero and an exponential drop off with time. In almost all cases, the multipath delays seen in low SS antenna height situations are **below 12 μsec**.

# Appendix C: Block Turbo and Reed-Solomon Coding

## *Turbo Code Description*

The Block Turbo Code (BTC) is a Turbo decoded Product Code (TPC). The general idea of BTC is to use simple component block codes (e.g., binary extended Hamming codes) for constructing large block codes that can be easily decodable by iterative Soft-In \ Soft-out (SISO) decoder. For the sake of this proposal, two-dimensional component codes are taken to construct a product-code [11]. The codes recommended for the current standard follows the lines of IEEE802.16.1 MODE B [9].

The matrix form of the two-dimensional code is depicted in Figure C1. The $k_1$ information bits in the rows are encoded into $n_1$ bits, by using a binary block $(n_1, k_1)$ code. The binary block codes employed are extended Hamming Codes or parity check codes. As product codes belong to a class of linear codes, the order of the encoding is not essential. In this proposal it is assumed that the encoding process is completed row-by-row, starting from the first row.

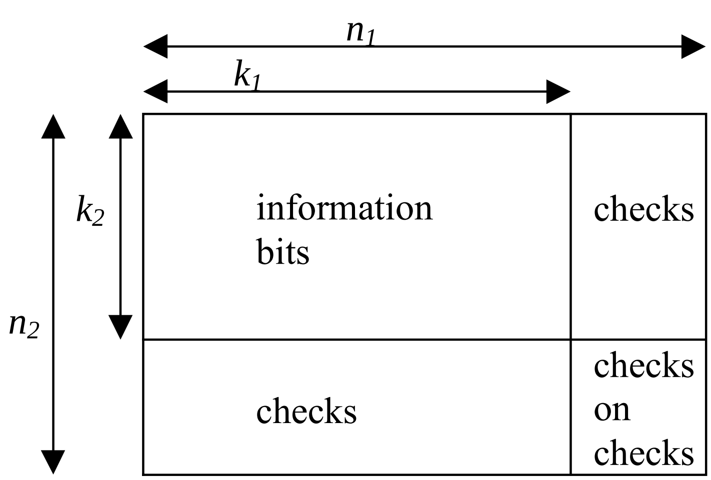


**Figure C1 - Two-dimensional product code matrix**

The redundancy of the code is $r_1 = n_1 - k_1$ and $d_1$ is the Hamming distance. After encoding the rows, the columns are encoded using another block code $(n_2, k_2)$, where the check bits of the first code are also encoded, producing checks on checks bits. The overall block size of such a product code is $n = n_1 \times n_2$, the total number of information bits $k = k_1 \times k_2$ and the code rate is $R = R_1 \times R_2$, where $R_i = k_i/n_i$, $i = 1, 2$. The Hamming distance of the product code is $d = d_1 \times d_2$.

### Encoding

The encoder for a TPCs has a latency of one row ($n_1$ bits) when employing interleaver type 1, and no more than one block of product code for general permutation interleaver. Encoders can be constructed of linear feedback shift registers (LFSRs), storage elements, and control logic. The constituent codes of TPCs are extended Hamming codes or parity check codes. Table C1 gives the generator polynomials of the Hamming codes used in TPCs. For extended Hamming codes an overall parity check bit is added at the end of each codeword.

| N | K | Generator |
|---|---|---|
| 15 | 11 | $x^4 + x^1 + 1$ |
| 31 | 26 | $x^5 + x^2 + 1$ |
| 63 | 57 | $x^6 + x + 1$ |

**Table C1 - Generators Polynomials of Hamming components Codes**

In order to encode the product code, each data bit is input both into a row encoder and a column encoder. Note that only one row encoder is necessary for the entire block, since data is input in row order. However, each column of the array must be encoded with separate encoders. Each column encoder is clocked for only one bit of the row, so a more efficient method of column encoding is to store the column encoder states in a $k_1 \times (n_2\text{-}k_2)$ 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. Assume a two-dimensional $(8,4) \times (8,4)$ extended Hamming Product code is to be encoded. This block has 16 data bits, and 64 total encoded bits. Figure C2 shows the original 16 data bits denoted by $D_{yx}$.

$$\begin{matrix} D_{11} & D_{21} & D_{31} & D_{41} \\ D_{12} & D_{22} & D_{32} & D_{42} \\ D_{13} & D_{23} & D_{33} & D_{43} \\ D_{14} & D_{24} & D_{34} & D_{44} \end{matrix}$$

**Figure C2 - Original Data for Encoding.**

The first four bits of the array are input to the row encoder in the order $D_{11}$, $D_{21}$, $D_{31}$, $D_{41}$. Each bit is also input to 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 shifted out at the end of the row. This continues from the remaining 3 rows of the array. Figure C3 shows the final encoded block with the 48 generated ECC bits denoted by $E_{yx}$.

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

**Figure C3 - Encoded Block.**

Transmission of the block over the channel occurs in a linear fashion, 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 $D_{11}$, $D_{21}$, $D_{31}$, $D_{41}$, $E_{51}$, $E_{61}$, $E_{71}$, $E_{81}$, $D_{12}$, $D_{22}$, … $E_{88}$.

**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
- data bits are noted $D_{y,x}$ and parity bits are noted $E_{y,x.}$

**Shortened BTCs**

To match packet sizes, a product code can be shortened by removing symbols from the array. In the two-dimensional case rows, rows 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 step is to remove S2 rows or S1 columns from a 2-dimensional code. This is equivalent to shortening the constituent codes that make up the product code, i.e, ($n_1$ –S1, $k_1$ –S1) and ($n_2$ –S2, $k_2$- S2). 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 could be obtained by removing individual S bits from the last row of a 2-dimensional code.

**Example**: To obtain 20 bytes payload based on (32,26)x(32,26) code, set S1=S2=13. The resulted product code has (19,13)x(19,13) structure which gives 169 payload bits. Then S=9 bits left over which are stuffed with zeros. Data input to the defined encoder is 160 bits (20 bytes) followed by 9 bits of zeros. The BTC codeword is transmitted starting with the bit in row 1 column 1 (LSB), then left to right, and then row by row.

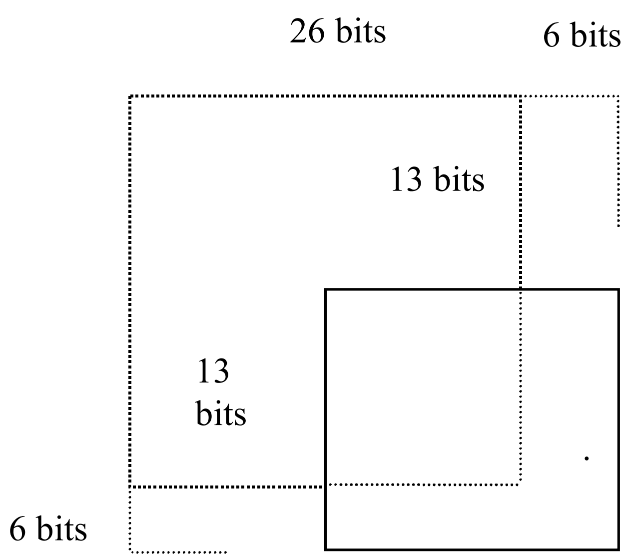


**Figure C4 – An Example of Encoded Block.**

**Block mapping to the signal constellation:** The first encoded bit out shall be the LSB, which is the first bit written into the decoder. When the row is not a multiple of the constelation log-size, then bits from next row are used to map bits into symbols.

**Shortened last codeword Mode:** This mechanism allows by shortening the last codeword a further flexibility to more closely match the block size of the BTC with the required message length. The following steps describe this mode.

Define a new codeword that has a minimum number of rows that will carry the required number of information bits. The number of columns should be kept unchanged.

If the number of positions for information in the resultant codeword, k', is greater than the number of information bits $k_1$ , then add k' – k stuff bits ( "1" ) to the end of the message.

Information bits and stuf bits k' are randomized.

### Examples of a Shortened Two-Dimensional BTC

For example, assume a 456-bit block size is required (53+4 bytes for payload), with 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 2 and all columns by 7 results in a (30,24) × (25,19) code, with a data size of 456 bits and the final code is a (750,456) code, with a code rate of 0.608.

The following shortened codes are given as examples:

- Product codes based on shortened binary Hamming code:

  $(2^m$ - S1, $2^m$-m-1-S1, 4)x($2^m$ - S2 ,$2^m$-m-1-S2, 4 ) where m is the encoder LFSR length

  and S, S1 and S2 are configurable shorthening parameters.

(19,13)x(19,13) m=5, S1=S2=13, S=9 (20 bytes payload)

(30,24)x(25,19) m=5, S1=2, S2=7 (53+4 bytes payload)

(30,24)x(24,18) m=5, S1=2, S2=8 (53+1 bytes payload)

(39,32)x(39,32) m=6, S1=S2=25 (128bytes payload)

(39,32)x(54,47) m= 6, S1= 25, S2=10 (188 bytes payload)

(63,56)x(63,56) m=6, S1=S2=1 (392 bytes payload).

- Product codes based on binary parity-check codes:

  (2k+1, 2k) X (2k+1, 2k) where k is configurable. (max k=32, min k =TBD).

### Iterative Decoding

Each block code in a product code is decoded independently. First, all the horizontal blocks are decoded then all the vertical received blocks are decoded (or vice versa). The decoding procedure is generally iterated several times to maximize the decoder performance. To achieve optimal performance, the block by block decoding must be done utilizing soft information. This soft decision decoder must also output a soft decision metric corresponding to the likelihood that the decoder output bit is correct. This is required so that the next decoding will have soft input information as well. In this way, each decoding iteration builds on the previous decoding performance.

The core of the decoding process is the **soft-in\soft-out** (SISO) constituent code decoder. High performance iterative decoding requires the constituent code decoders to not only determine a transmitted sequence, but to also yield a soft decision metric which is a measure of the likelihood or confidence of each bit in that sequence.

Since most algebraic block decoders don't operate with soft inputs or generate soft outputs, such block decoders have been primarily realized using the **Soft-Output Viterbi Algorithm** (SOVA) [12] or a soft-output variant of the modified Chase algorithm(s). However, this does not limit the choice of decoding algorithms as other SISO block decoding algorithms can be used [13], [14].

**Interleaving with BTC**

Three bit interleaver~~s~~ are recommended when using BTC case. The implementation of the interleaver is by writing the bits into the encoder/decoder memory and reading out as follows.

**Type 1 (no interleaver):** In this mode bits are written row-by-row and read row-by-row.

**Type 2 (block interleaver):** In this mode the encoded bits are read from the encoder, only after all first k2 rows were written into the encoder memory. The bits are read column-by-column from top position in the first column.

**Type 3 (permutation interleaver):** Reserved.

It is expected that other interleaving methods yield better performance in some cases, and especially when combined with M-QAM signaling.

**Typical performance with BTC**

The performance cited here are based on results given in IEEE802.16.1pc-00/35 [9]. That is, 5 iterations and quantization of soft metrics into sign + 4 bits per one dimensional modulation level. Interleave type 1 (no interleaver) was assumed.

**TABLE C2: Typical performance for BTC with large blocks (downstream \ upstream channels)**

| CODE | $(39, 32)^2$, S1=S2=25, s=0 | $(46, 39)^2$ S1=S2=17, s=17 | $(63, 56)^2$, S1=S2=1, s=0 |
|---|---|---|---|
| Rate | 0.673 | 0.711 | 0.790 |
| Eb/N0 dB @$10^{-6}$ 4/16/64 QAM | 3.5 / 6.5/ 10.7 | 3.6 / 6.6 / 10.5 | 3.5 / 6.6 / 10.6 |
| Eb/N0 dB @$10^{-9}$ 4/16/64 QAM | 4.3 / 7.5 / 11.7 | 4.3 / 7.8 / 11.5 | 4.3 /7.5 / 11.6 |
| Block size (information bytes) | 1024 (128 bytes) | 1504 (188 bytes) | 3136 (392 bytes) |
| Encoder Complexity | 10 Kgates | 10 Kgates | 10 Kgates |
| Decoder Complexity | Less than 150Kgates | Less than 150Kgates | Less than 150Kgates |

**TABLE C3: Typical performance for BTC with small blocks (upstream channel)**

| CODE | $(16, 11)^2$,<br>S1=S2=0, s=1 | (30, 24)x(25, 19)<br>S1=2, S2=7, s=0 |
|---|---|---|
| Rate | 0.469 | 0.608 |
| Eb/N0 dB @$10^{-6}$<br>4/16/64 QAM | 4.0 / 6.8/ 9.8 | 3.4 / 6.3 / 10 |
| Eb/N0 dB @$10^{-9}$<br>4/16/64 QAM | 5.8 / 8.8 / 11.8 | 4.7 / 7.5 / 11.5 |
| Block size<br>(information bytes) | 120<br>(15 bytes) | 456<br>(57 bytes) |
| Encoder Complexity | 10 Kgates | 10 Kgates |
| Decoder Complexity | Less than<br>150Kgates | Less than 150Kgates |

## 7.1 Reed – Solomon Coding

The processing will be as summarized in the following conceptual block diagram.

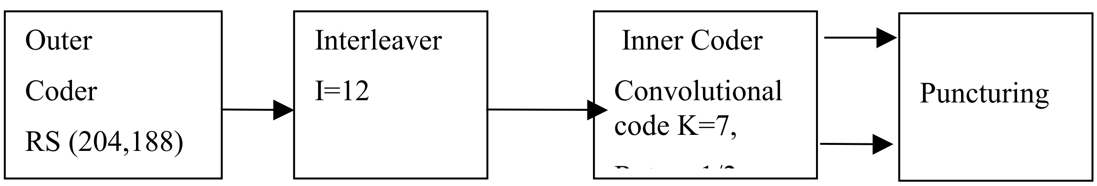


**Figure C5 – Concatenated Reed – Solomon and Convolutional Codes Encoder Block.**

The outer code is RS (204,188, T=8), shortened, systematic Reed - Solomon over GF(256), with information block length K=188 with 16 parity check bytes (i.e., correction capability of T=8 bytes).

$$p(x) = x^8 + x^4 + x^3 + x^2 + 1$$

*P(x)* is the Field generator polynomial, and the code generator polynomial is:

$$g(x) = (x + \lambda^0)(x + \lambda^1)(x + \lambda^2)\cdots(x + \lambda^{2T-1})$$

Where:

$$\lambda = 02HEX$$

$\lambda$ is a primitive root of *p(x)*.

The shortened RS is obtained from RS (255,239,T=8) code by adding 51 bytes, all set to zero, before the information bytes at the input of a RS (255,239) encoder. After encoding these nulls are discarded.

The convolutioal symbol interleaver is depth I=12 based on Forney approach [8].

**Inner Convolutional coding:** based on rate 1/2 Convolutional code with constraint length K=7, corresponding to 64 trellis states described by gererators G1=171 Octal and G2=133 Octal.

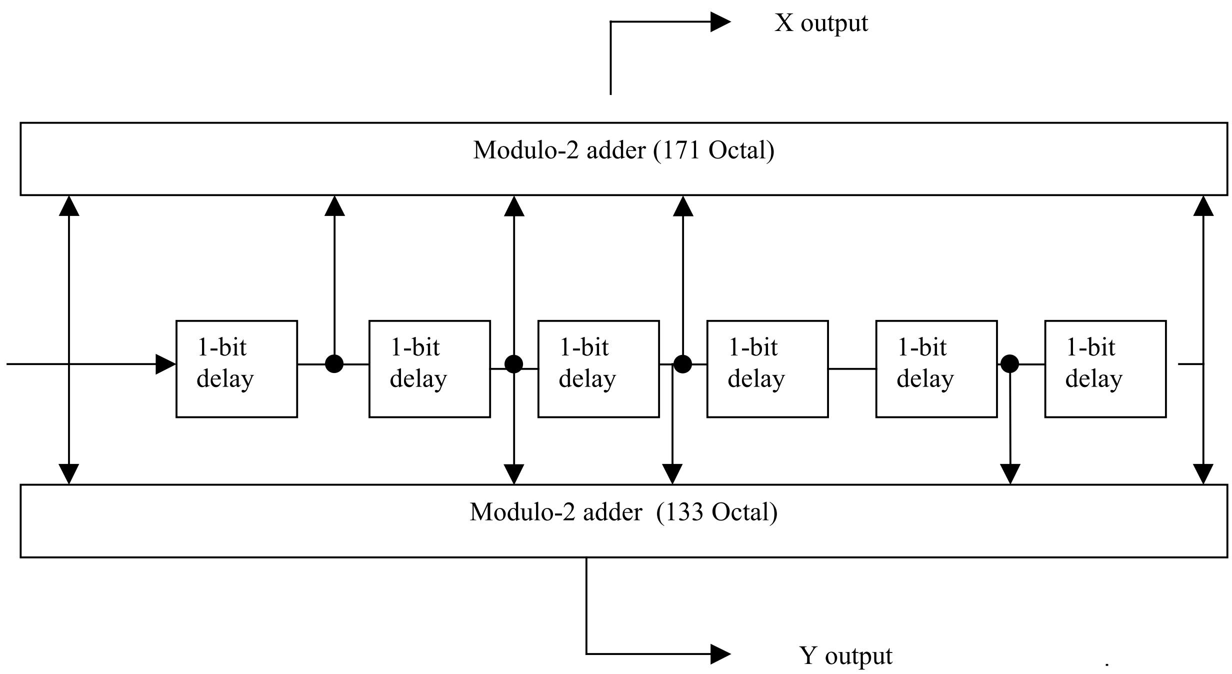


**Figure C6 – Convolutional Encoder Diagram.**

The inner Convolutional code has puncturing configuration defined in the following table. In this notations, (x, y) denotes a bit pairs at the output of the Convolutional encoder.

"1" in a puncture pattern means transmitted bit while "0" denotes non-transmitted bit. These bit pairs will be used for Gray coded (I, Q) mapping.

**TABLE C4: The inner Convolutional code with Puncturing Configuration**

| Original Code | | | Code rates | | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| | | | 1/2 | | 2/3 | | 3/4 | | 5/6 | | 7/8 | |
| K | G1 | G2 | P | $d_{free}$ | P | $d_{free}$ | P | $d_{free}$ | P | $d_{free}$ | P | $d_{free}$ |

| | | | | | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 7 | 17 1 | 13 3 | X: 1<br>Y: 1<br><br>I=$X_1$<br>Q=$Y_1$ | 10 | X: 10<br>Y: 11<br><br>I= $X_1Y_2Y_3$<br>Q= $Y_1X_3Y_4$ | 6 | X:101<br>Y:110<br><br>I= $X_1Y_2$<br>Q= $Y_1X_3$ | 5 | X :10101<br>Y :11010<br><br>I= $X_1Y_2Y_4$<br>Q= $Y_1X_3X_5$ | 4 | X:1000101<br>Y:1111010<br><br>I= $X_1Y_2Y_4Y_6$<br>Q= $Y_1Y_3X_5X_7$ | 3 |

Error Performance requirements with RS (204,188) + inner Convolutional coding (IF loop results, based on [9 Table 5] for AWGN @ BER =$10^{-8}$).

**TABLE C5: The inner Convolutional code with Puncturing Configuration**

| Modulation | Inner code rate | Spectral efficiency | Aggregate code rate | Eb/N0 dB * |
|---|---|---|---|---|
| QPSK | 1/2 | 0.92 | 0.46 | 4 |
| | 2/3 | 1.23 | 0.61 | 4.5 |
| | 3/4 | 1.38 | 0.69 | 5.0 |
| | 5/6 | 1.53 | 0.77 | 5.5 |
| | 7/8 | 1.61 | 0.81 | 5.9 |
| 16QAM | 3/4 | 2.76 | 0.69 | 8.5 |
| | 7/8 | 3.22 | 0.81 | 10.2 |
| * Notes: the figures cited here includes: 0.8 dB implementation loss for QPSK, and 1.5, 2.1 dB for the ¾, 7/8, 16QAM, respectively. | | | | |