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Re: This contribution is submitted to the IEEE 802.16a Task Group as it was requested by the Task Group chair on April 12th, 2001 for the SC-FDE PHY part of Draft Document submission for Sub 11 GHz BWA.

Abstract: This document provides team Draft Document for the SC-FDE part of the PHY System for a low frequency (Sub 11 GHz) wireless access PHY for point-to-multipoint voice, video and data applications

Purpose	The submission of the draft document is for review by the Task Group and it is expected to receive comments prior to be adopted by the Task Group as the SC-FDE part of PHY standard for Sub 11 GHz BWA during the Session#13.
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PHY Layer System for Single Carrier – Frequency Domain Equalizer for Sub 11 GHz BWA

(An OFDM Compatible Solution)

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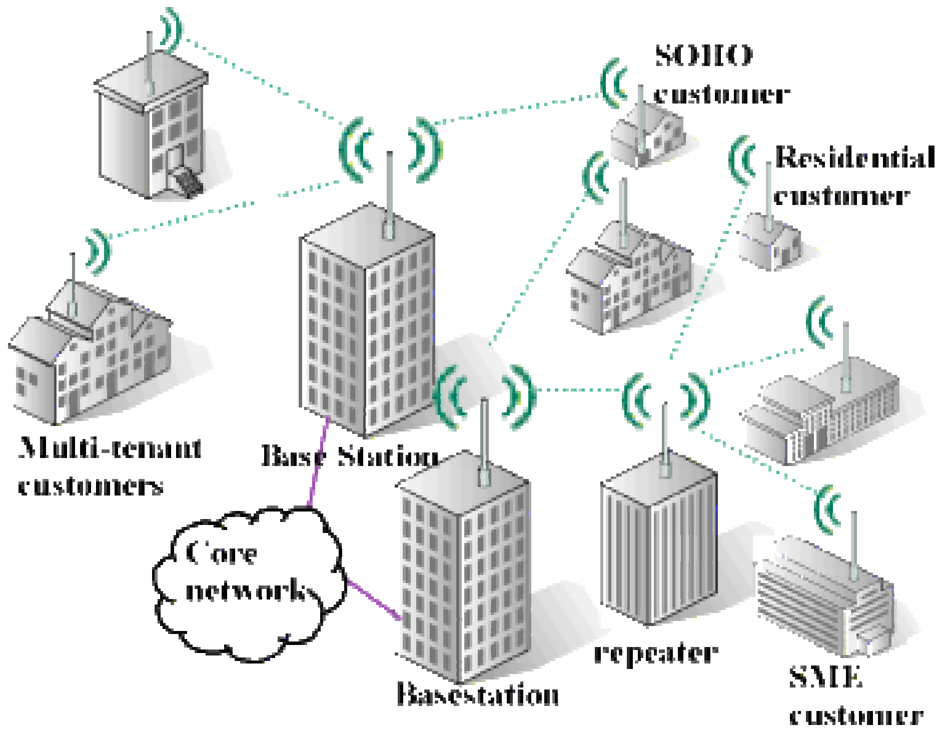
1 Scope

This document defines a Physical Layer (PHY) for IEEE802.16a 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 1.1 illustrates a BWA reference model.



1 **Figure 1.1: Wireless Access Reference Model**

2 Introduction

2.1 General

This document will address the SC — FDE PHY in detail and will highlight the added OFDM Compatibility Features.

The draft document describes the Single Carrier (SC) PHY system which adopts TDM/ TDMA bandwidth sharing scheme. The signal is transmitted downstream from the Base Station to all assigned Subscriber Stations using 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 assigned Subscriber Stations to the Base Station in Time Division Multiple Access (TDMA) mode. This access scheme can be either FDD or TDD. Both duplexing schemes have intrinsic advantages and disadvantages, so for a given application 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 from 2 to 11 GHz and the Base Station can use multiple sectors and will be capable of supporting smart antenna in the future.

The PHY layer uses a Single Carrier (SC) modulation with a Frequency Domain Equalizer (FDE) (or SC—FDE). We have shown that SC-FDE modulation can offer as good or better performance than Orthogonal Frequency Division Modulation (OFDM) technology in solving the Non-Line of Sight (NLOS) problem that may arise in the frequency bands between 2 and 11 GHz (See references [36 and 37]).

In addition, this document discusses the **compatibility** between of the SC—FDE and OFDM modulation schemes for Sub 11 GHz BWA applications. Furthermore, the frame structure presented in this draft document has all the capabilities for adaptive modulation and coding. The main objective of the frame structure has been to make the PHY almost independent from the MAC. The PHY here is based upon utilizing the structure of the 802.16 MAC.

2.2 Single Carrier PHY Features

The Single Carrier PHY is a Broadband Wireless Access (BWA) **Point-to-Multipoint** communication system that can provide digital, two-way voice, data, Internet and video services. This PHY shall offer 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 here will support services including 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 upstream statistical multiplexing over the air-interface using Time Division Multiple Access (TDMA) technology.

The key features of the PHY are the following:

- Full compatibility with the 802.16 MAC.
- Upstream multiple access is based on TDMA scheme.
- Downstream multiple access is based on broadcast TDM scheme.
- Duplexing is based on either TDD or FDD scheme.
- PHY uses a block adaptive modulation and FEC coding in both Upstream and Downstream paths.
- 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 can result in low cost Subscriber Stations (SS) and Base Stations (BS).
- The modulation scheme is robust and reduces multi-path and other channel impairments
- The PHY is flexible in terms of geographic coverage, in the frequency band used, and capacity allocation.
- Base Station can use multiple sector antennas. Support for future use of smart antennas is feasible and is implicit in the PHY design.
- The PHY can easily accommodate multi-beam and antenna diversity options; such as Multiple-In Multiple-Out (MIMO) and Delay diversity.
- The SC -FDE PHY has an added feature of re-configurability to support OFDM modulation.
- 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 Compatible 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.

3 Description of the SC-FDE PHY

As described in the Functional Requirement Document [1], 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 a combination of coverage, capacity and equipment cost factors that affect total cost per user. The PHY facilitates 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 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 3.22.

3.1 SC-FDE Wireless Access System Model

Figure 3.1 is a top level block diagram of the PHY layer system for BWA services. Figure 3.1a is an illustration for Single Carrier system and Figure 3.1b is the Compatible OFDM system.

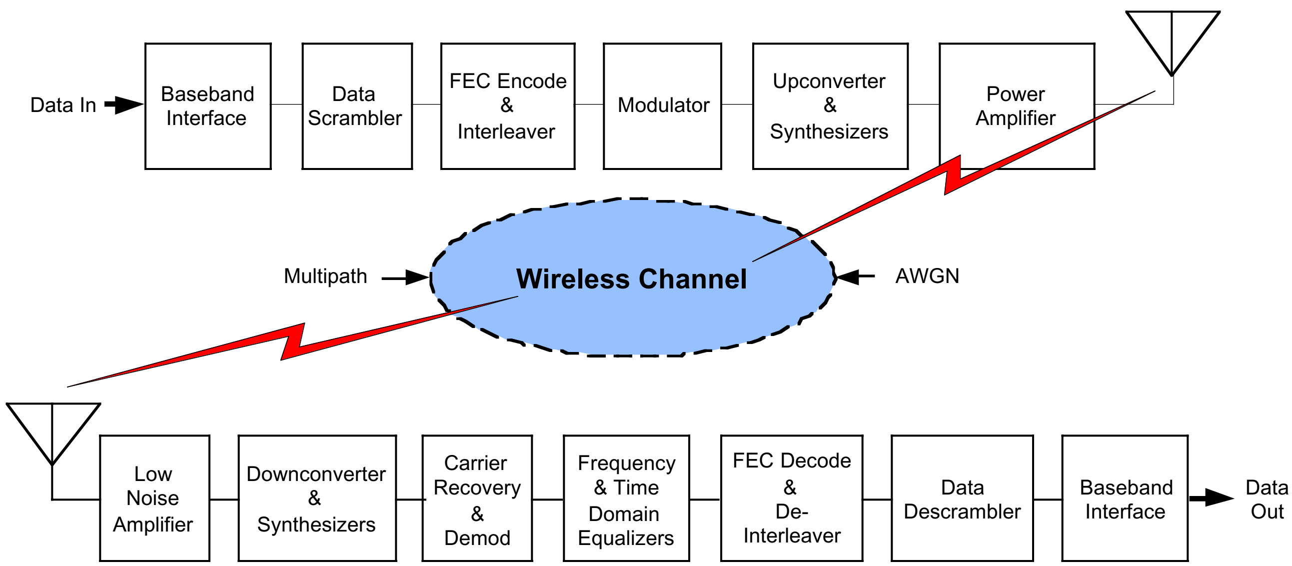


Figure 3.1a: The Single Carrier PHY Layer Block Diagram.

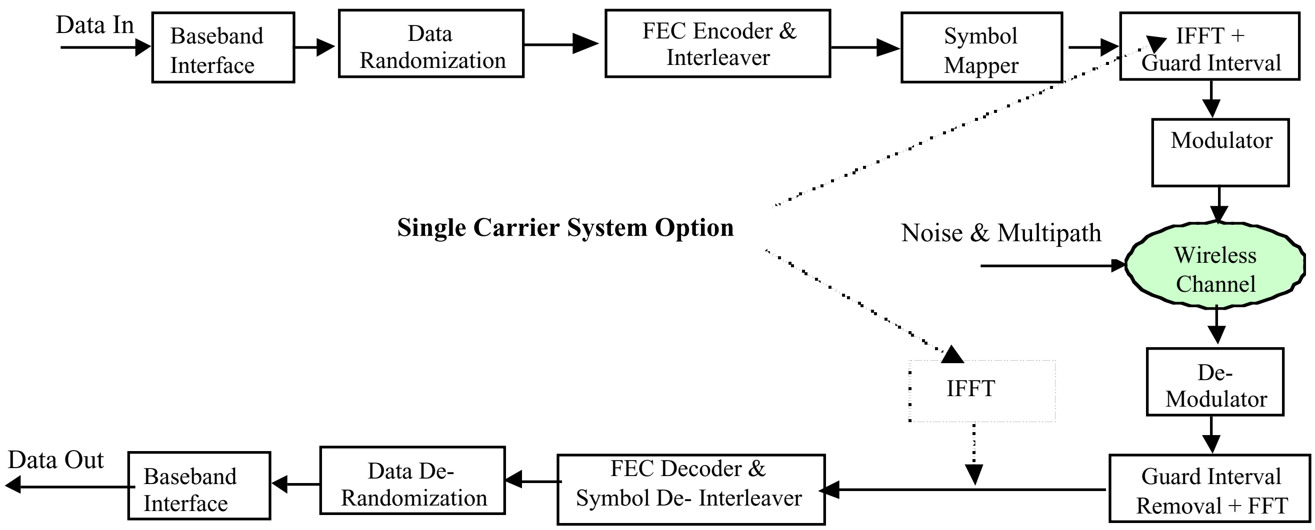


Figure 3.1b: The Compatible OFDM and Single Carrier PHY Proposal Block Diagram.

3.2 Multiple Access Formats and Framing

In this Section, we introduce multiple access formats, and the PHY framing and MAC/ PHY interface structures necessary to accommodate these formats.

Section 3.2.1 describes the PHY framing structures and sub-elements used to support various multiple access formats; Section 3.2.3 describes the MAC/PHY interface, and goes into details on the supported multiple access formats. In addition, Section 3.2.3 describes the use of adaptive antenna technology.

3.2.1 Physical Layer Framing Structures

3.2.1.1 Overview of Frame Formats and Their Application

Starting with a simple fundamental frame component and two formats, we construct PHY structures that may be applied to several multiple access techniques.

Two fundamental PHY block format options are available:

1. one used for continuous transmissions, and
2. another used for burst transmissions.

As we shall eventually demonstrate, the continuous transmission format might be used on the downstream of one type of a FDD system.

The burst format might be seen on

- the upstream of a FDD system; or
- the upstream and downstream of a TDD system; or
- the downstream of a burst-FDD system.

The burst format may be further categorized into two subformats:

1. TDMA burst, and
2. TDM burst.

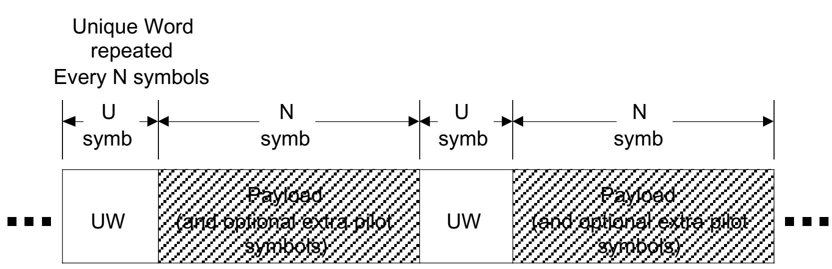
A TDMA burst contains information intended for one audience. This audience could be a single user, or a group of users receiving a broadcast message. In contrast a TDM burst generally contains multiplexed, concatenated information addressed to multiple audiences. A TDMA burst may, in fact, be interpreted as a type of TDM burst; however, because of differences related to usage of adaptive modulation, MAC messaging and multiple access, we shall in some sections choose to discuss these two burst types separately.

3.2.1.2 Continuous Transmission Format

As its name suggests, the continuous transmission format is utilized for a continuous channel, which may be monitored, for example, by all of the Subscriber Stations (SSs) within a Base Station (BS) cell sector. In particular, one might see this format applied in the operation of a (continuous) FDD downstream channel.

3.2.1.2.1 Unique Word: Interval Requirements and Usage

One characteristic of the continuous transmission format is illustrated in Figure 3.1: the continuous format has a fundamental pattern that repeats. This pattern consists of N-symbol payloads separated by U-symbol Unique Words .



2 Figure 3.1 Unique Word Intervals for Continuous Format

3.2.1.2.1.1 Unique Word Interval Requirements

A Unique Word is a contiguous, length-U sequence of known symbols, which are not FEC-encoded. A Unique Word is repeated at a regular interval, N. The interval between the Unique Words is typically chosen to accommodate receivers using frequency domain equalization, with F = N+ U symbols equaling the block length over which an FFT would be computed by a frequency domain equalizer.

To reduce computational requirements of FFTs, the length F = N + U should preferably be 2 raised to an integer power. Additional details on the composition of the symbols within Unique Words, and their accepted lengths, U, may be found in 3.2.1.4.2. Note that 0 is also an acceptable value for the length U. Additional details on accepted lengths for F may be found in 3.2.1.5.

Once the parameters for U and N (via F) are set, they should not be changed. In the event that these parameters must be changed, receiver resynchronization may be necessary.

3.2.1.2.1.2 Unique Word Usage

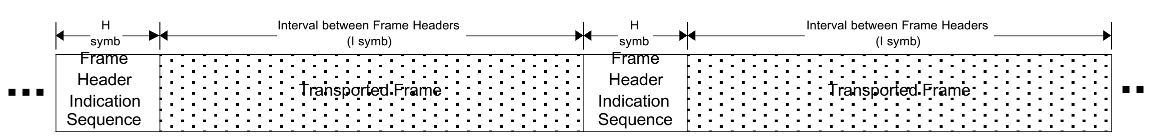
The Unique Word may be used as a cyclic prefix by a frequency domain equalizer, and/or as pilot symbols. When used as a cyclic prefix, a Unique Words should be at least be as long as the maximum delay spread of a channel. As pilot symbols, the Unique Words may assist in the estimation of demodulation parameters such as estimating equalizer channel coefficients, carrier phase and frequency offsets, symbol timing, and optimal FFT window timing (in a frequency domain equalizer). The Unique Words may also assist in the acquisition of a channel.

3.2.1.2.2 Frame Header Indication Sequence: Requirements and Usage

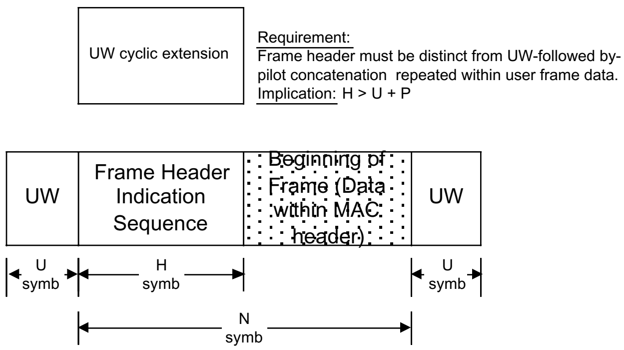
In addition to Unique Word intervals, continuous format data is further framed into MAC-based frames, and these MAC frame boundaries are delineated by a Frame Header Indication Sequence.

3.2.1.2.2.1 Frame Header Indication Sequence Requirements

As illustrated in Figure 3.2: Frame Header Indication Sequences of length H are periodically repeated, with repetition interval I. The repetition interval, I, for a Frame Header Indication Sequence must be an integer number of F = N+U Unique Word intervals. Furthermore, as Figure 3.3 illustrates, a Frame Header Indication Sequence must directly follow a Unique Word sequence.



3 **Figure 3.2 Frame Header Intervals for Continuous Format**



4 **Figure 3.3 Frame Header Indication Sequence Position within a Unique Word Interval**

The Frame Header Indication Sequence is a contiguous sequence of known symbols, which are not FEC-encoded. As Figure 3.3 further indicates, the Frame Header Indication Sequence is a cyclic extension of the Unique Word which implies that it is either a repetition of the Unique Word, or a partial replication of the first H symbols of the Unique Word, or a combination of a repetition and partial replication of the Unique Word. Moreover, the Frame Header Indication Sequence must be longer than a sequence of additional pilot symbols which may be contiguous to ensuing Unique Words within the data payload (see Figure 3.5 for details on the addition of extra pilot symbols).

Once the parameters for H and I are set, they should not be changed. In the event that these parameters are changed, receiver resynchronization may be necessary.

3.2.1.2.2.2 Frame Header Indication Sequence Usage

As indicated 3.2.1.2.2, MAC frame boundaries are delineated by the Frame Header Indication Sequence. Identification of the location of the MAC header is important during acquisition, because the MAC header contains much of the system and frame control information, including MAPs of user data, their lengths, modulation formats, and the FEC used to encode them. Therefore, once the MAC header is located and decoded, all ensuing user data that has the CINR to be decodable can be decoded. This begs that the Frame Header Indication Sequence be distinct, so that the location of the frame header may be easily identified, and distinguished from pilot symbols.

What s more, the Frame Header Indication Sequence has a role following initial acquisition. The structure (usage of Unique Word elements) and placement of the Frame Header Indication Sequence also enables re-acquisition and channel estimation before the outset of a subsequent MAC frame. This is important when per-user adaptive modulation is used, because, as indicated in 3.2.1.2.3.3, user data is sequenced in terms of modulation robustness. Therefore, receivers experiencing low CINRs may not be able to track completely through a MAC frame. The Frame Header Acquisition Sequence aids such a receiver in reacquiring, or getting a better, more solid channel estimate, before the appearance of data that it has the CINR to successfully decode.

3.2.1.2.3 Adaptive Modulation

3.2.1.2.3.1 Concept of Adaptive Modulation

Many SSs are intended to receive the continuous downstream channel. Due to differing conditions at the various SS sites (e.g., variable distances from the BS, presence of obstructions, local interference), SS receivers may observe significantly different CINRs. For this reason, some SSs may be capable of reliably detecting (non-pilot) payload data only when it is derived from certain lower-order modulation alphabets, such as QPSK. Similarly, CINR-disadvantaged SSs may require more powerful and redundant FEC schemes. On the other hand, CINR-advantaged stations may be capable of receiving very high order modulations (e.g., 64-QAM), with high code rates. Obviously, to maximize the overall capacity of the system, the modulation and coding format should be adapted to each class of SS, based on what the SS can receive reliably. Define the adaptation of modulation type and FEC to a particular SS (or group of SSs) as 'adaptive modulation', and the choice of a particular modulation and FEC as an 'adaptive modulation type.' The continuous transmission mode (as does the burst transmission mode) supports adaptive modulation and the use of adaptive modulation types.

3.2.1.2.3.2 Frame Control Header Information and Adaptive Modulation

As Figure 3.4 illustrates, Frame Control MAC messages are periodically transmitted over the continuous channel, using the most robust adaptive modulation type supported. Among other information, these Frame headers provide adaptive modulation type formatting instructions.

As 3.2.1.2.2 describes, in order that Frame headers may be easily recognized during initial channel acquisition or re-acquisition, the transmitter PHY inserts an uncoded Frame Header Indication Sequence immediately before the Frame header, and immediately after a Unique Word. Figure 3.4 illustrates this point, as well.

3.2.1.2.3.3 Adaptive Modulation Sequencing

Within the MAC Frame header, a PHY control map (DL_MAP) is used to indicate the beginning location of each of adaptive modulation type payload that follows. However, the DL_MAP does not describe the beginning locations of the payload groups that immediately follow; it describes the payload distributions at some MAC-prescribed time in the future. This delay is necessary so that FEC decoding of MAC information (which could be iterative, in the case of turbo codes) may be completed, the adaptive data interpreted, and the demodulator scheduling set up for the proper sequencing.

As Figure 3.4 illustrates, following the MAC Frame header, <u>payload groups are sequenced in increasing order of robustness</u> (e.g., first QPSK, then 16-QAM, then 64-QAM). This robustness sequencing improves receiver performance, because it enables receivers experiencing lower CINRs to track only through the modulation types that they can reliably receive. This sequencing also facilitates changes of modulation type at locations that are not contiguous to Unique Word boundaries.

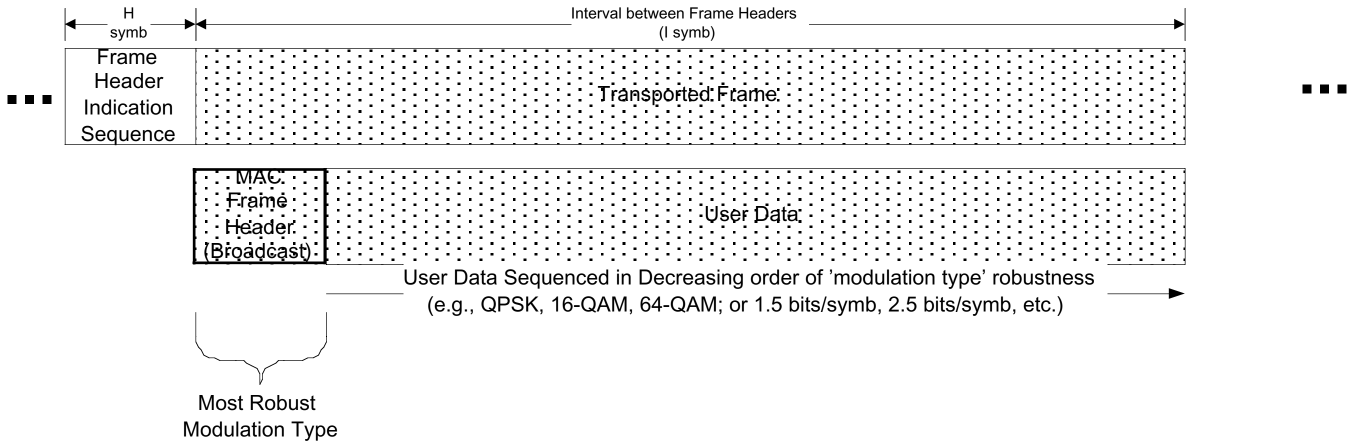


Figure 3.4 Adaptive Modulation Sequencing within a Continuous Mode Frame

3.2.1.2.3.4 UW Boundary-free Transitions Between Modulation Types

Note also that adaptive modulation type-to-other-modulation type changes are not restricted to occur only at Unique Word boundaries. They may change anywhere that the DL_MAP message indicates that they should change.

3.2.1.2.3.5 Per-Adaptive-Modulation-Type FEC Encapsulation

So that disadvantaged-SNR SSs are not adversely affected by transmissions intended for other advantaged-CINR users, FEC blocks end when a particular adaptive modulation type ends. Among other things, this implies that the FEC interleaver depth and code blocks are adapted to accommodate the span of a particular adaptive modulation type. Note, however, that data from several users could be concatenated by the MAC (and interleaved together by the PHY) within the span of a given adaptive modulation type.

3.2.1.2.3.6 MAC Header FEC Encapsulation

So that the MAC header data may be decoded by a receiver that has just acquired (and does not yet know the modulation lengths and distributions of user data), the MAC header data should be

1. a fixed, a priori-known block size; and
2. separately FEC-encoded (and interleaved) from all other user-specifically-addressed data.

3.2.1.2.4 Empty payloads

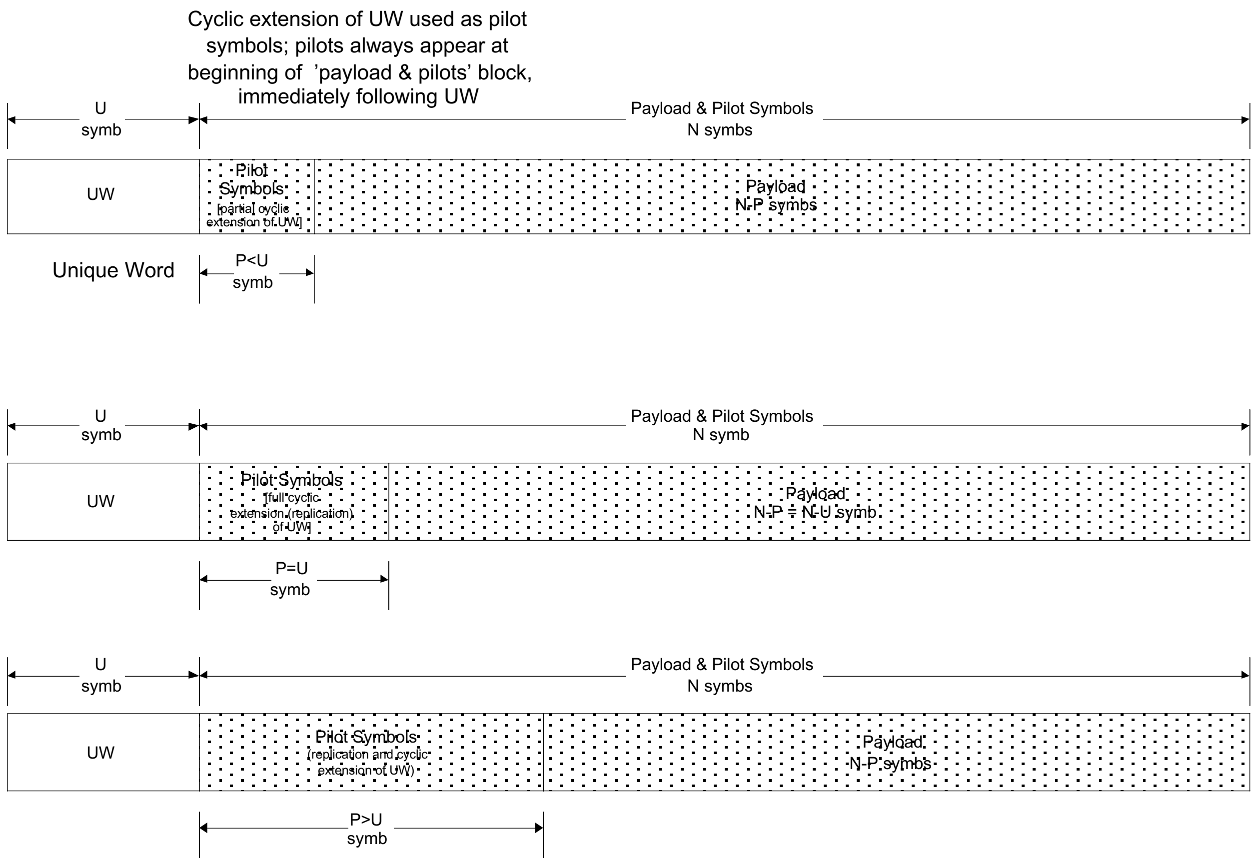
When data is not available for transmission, part of a payload may be stuffed with dummy data, or left empty, at the system operator s discretion. However, the transmitter cannot shut down completely. The unique words are always transmitted, so that all listening SSs may track the channel, and maintain synchronization.

3.2.1.2.5 Additional Pilot Symbols

When multipath delay spread spans almost the entire Unique Word interval, very little data remains that is not uncorrupted by delay spread from the arbitrary, a priori unknown payload symbols. In such an environment, non-decision aided channel (delay profile) estimation could become exceedingly difficult. One recourse is the increased utilization of decision-aided channel estimation.

To add an extra measure of robustness, many system operators may prefer, instead, to opt for the addition of P additional pilot symbols. For this reason, the addition of an extra P pilot symbols per Unique Word interval is an option, as contiguous cyclic extensions of the Unique Word. A contiguous cyclic extension of P symbols implies that it the pilots are either a repetition of the Unique Word, or a partial replication of the first P symbols of the Unique Word, or a combination of a repetition and partial replication of the Unique Word.

Figure 3.5 illustrates three cases where pilot symbols have been added: one for a case when only a few symbols have been added, a second where the number of added pilot symbols and Unique Word symbols are the same (i.e., the Unique Word has been replicated), and a third for a case where the number of pilot symbols is much greater than the number of Unique Word symbols (i.e., where the UW is replicated at least once, then cyclically extended.) Note that it is also possible to add zero extra pilot symbols. The range of limits for P are 0†P†N.



6 **Figure 3.5 Three examples where extra pilot symbols have been added, via cyclic extension of the Unique Word [UW]. (top) pilots symbols less than UW symbols; (middle) pilot symbols equal to UW symbols; (bottom) pilot symbols greater than UW symbols.**

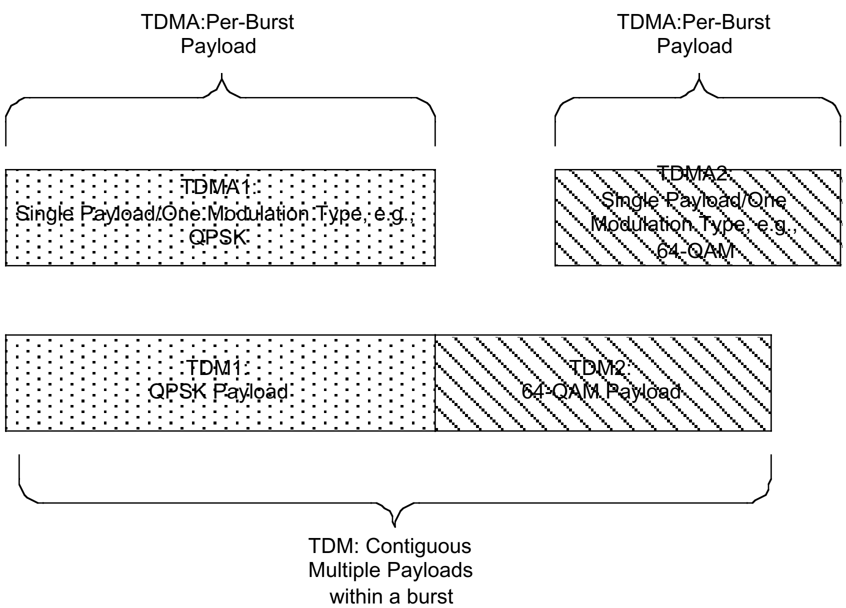
Note that the number of additional pilot symbols, P, used per N-symbol interval (in the continuous mode) may not change. In the event that these parameters must be changed, receiver resynchronization may be necessary.

Additional details on the composition of the symbols within Unique Words, and their accepted lengths, U, may be found in 3.2.1.4.2. Additional details related to accepted lengths for F = N + U (where N is the interval between Unique Word repetitions) may be found in 3.2.1.5.

3.2.1.3 Burst Transmission Format

In addition to the continuous transmission format, a second transmission format exists: the burst transmission format. As its name implies, the burst transmission format is utilized for burst transmissions, all of which may or may not be monitored by all SSs within a BS cell sector. In the broadband wireless application, one might see bursts on the (multiple-access) upstream, a TDD upstream and downstream, or a burst-FDD downstream. As described in the MAC/PHY Interface Layer Description in Section 3.2.2, half duplex burst FDD operation is also possible using this format.

Both TDMA and TDM burst options are supported. As Figure 3.8 demonstrates, a TDMA burst contains information intended for one audience. This audience could be a single user, or a group of users receiving a broadcast message. In contrast, as Figure 3.8 also demonstrates, a TDM burst generally contains multiplexed, concatenated information that is usually addressed to multiple audiences. A TDMA burst may, in fact, be interpreted as a type of TDM burst; however, due to their differences in areas such as adaptive modulation, multiple access, and MAC messaging, we shall treat these two burst types separately.



7 Figure 3.6 Example Comparison of TDMA and TDM bursts

3.2.1.3.1 Burst Ramping: Ramp-Up and Ramp-Down

Bursts begin with a ramp-up sequence, and end with a ramp-down sequence, each of length R symbols. The selection for R is left to the system operator. The selection for R may be based on several factors, such as regulatory requirements related to adjacent channel energy spillage, power amplifier considerations, antenna diversity sensing and switching delays, and the length of the startup transient of the transmit filter.

In creating the ramp-up sequence, the transmit filter is initially filled with zero-valued (null) symbols, and then desired transmit data symbols are pushed into the system to naturally ramp the system power up to its full value. If a ramp-up sequence length shorter than one-half the length of the impulse response of the transmit filter is desired, the transmit filter output samples related to first few symbols may be suppressed and a ramped power buildup achieved by windowing the ramp-up sequence, using a raised-cosine window of the desired length R, for example.

In creating the ramp-down sequence, zero-valued (null) symbols are pushed into the transmit filter data path following the last desired transmit symbol. If a ramp-down sequence length shorter than one-half the length of the impulse response of the transmit filter is desired, the transmit filter output samples in the ramp-down region may be windowed, using a raised-cosine window of the desired length R, for example.

3.2.1.3.2 Unique Word: Interval Requirements and Usage

A Unique Word is a contiguous, length-U sequence of known symbols, which are not FEC-encoded. In the middle portions of a burst , a Unique Word is generally repeated at a regular interval, N. The interval between the Unique Words is typically chosen to accommodate receivers using frequency domain equalization, with F = N+ U symbols equaling the block length over which an FFT would be computed by a frequency domain equalizer.

3.2.1.3.2.1 Unique Word: Interval Requirements

Like the continuous transmission format illustrated in Figure 3.1, the middle portions of a burst contain a fundamental pattern consisting of N-symbol Payloads separated by U-symbol Unique Words . In the case of a short burst, this fundamental pattern may be no more than a single Unique Word-Payload-Unique Word combination. For longer bursts, this may be a Unique Word-Payload-Unique Word-Payload (Unique Word-Payload) Unique Word-Payload-Unique Word combination. Note that the patterned middle portion of the block must always commence and conclude with a Unique Word.

All payload blocks within a burst except, potentially, the last payload block, must be of the same length, N. To reduce computational requirements of FFTs for frequency domain equalizers, the length F = N + U should preferably be 2 raised to an integer power.

Unlike the continuous transmission format, however, the final payload block need not be the same length as the other payload blocks. It may be shortened to a length N_{end}. This accommodates finer overall block length granularities. More details on shortening may be found in 3.2.1.3.2.1.1 and its subsections.

Unlike the continuous format, the parameters in use for U, F, and F_{end} used by the burst format can potentially be modified on a burst-by-burst basis---if the MAC messaging MAPs for burst profiles so allow.

Additional details on the composition of the symbols within Unique Words, and their accepted lengths, U, may be found in 3.2.1.4.2. Note that 0 is also an acceptable value for the length U. Additional details on accepted lengths for F and F_{end} may be found in 3.2.1.5.

3.2.1.3.2.1.1 Variable Burst Sizes

A characteristic of the burst format is that, for efficient operation, it may be necessary to accommodate many different burst sizes. These burst sizes could be different from some integer multiple of the nominal FFT size, F =N + U, of a frequency domain equalizer.

3.2.1.3.2.1.1.1 Variable Burst Sizes and Frequency Domain Equalizers

Even for implementations using a frequency domain equalizer, the single carrier burst PHY using frequency domain has some flexibility in this regard. For messages intended for a receiver with a frequency domain equalizer, the final payload block can be shortened to the length N_{end}, under the constraints $N_{end}+U = 2^n$, n is an integer, and $2^n \geq U$.

3.2.1.3.2.1.1.2 Variable Burst Sizes and Time Domain Equalizers

For receivers using time domain equalizers (such as decision feedback equalizers), the shortened length of the last block, N_{end}, can be completely arbitrary, and is only limited by MAC packet length granularity restrictions.

3.2.1.3.2.1.1.3 Variable Length Negotiation

Exchange of information regarding receiver capabilities during initial registration is one method to ensure that message granularities always conform to a burst receiver s capabilities to process them.

3.2.1.3.2.1.1.4 Broadcast/Multicast Messages

Broadcast or multicast messages would always be sent assuming a frequency domain equalizer s granularity limitations, since those limitations are more restrictive.

3.2.1.3.2.2 Unique Word Usage

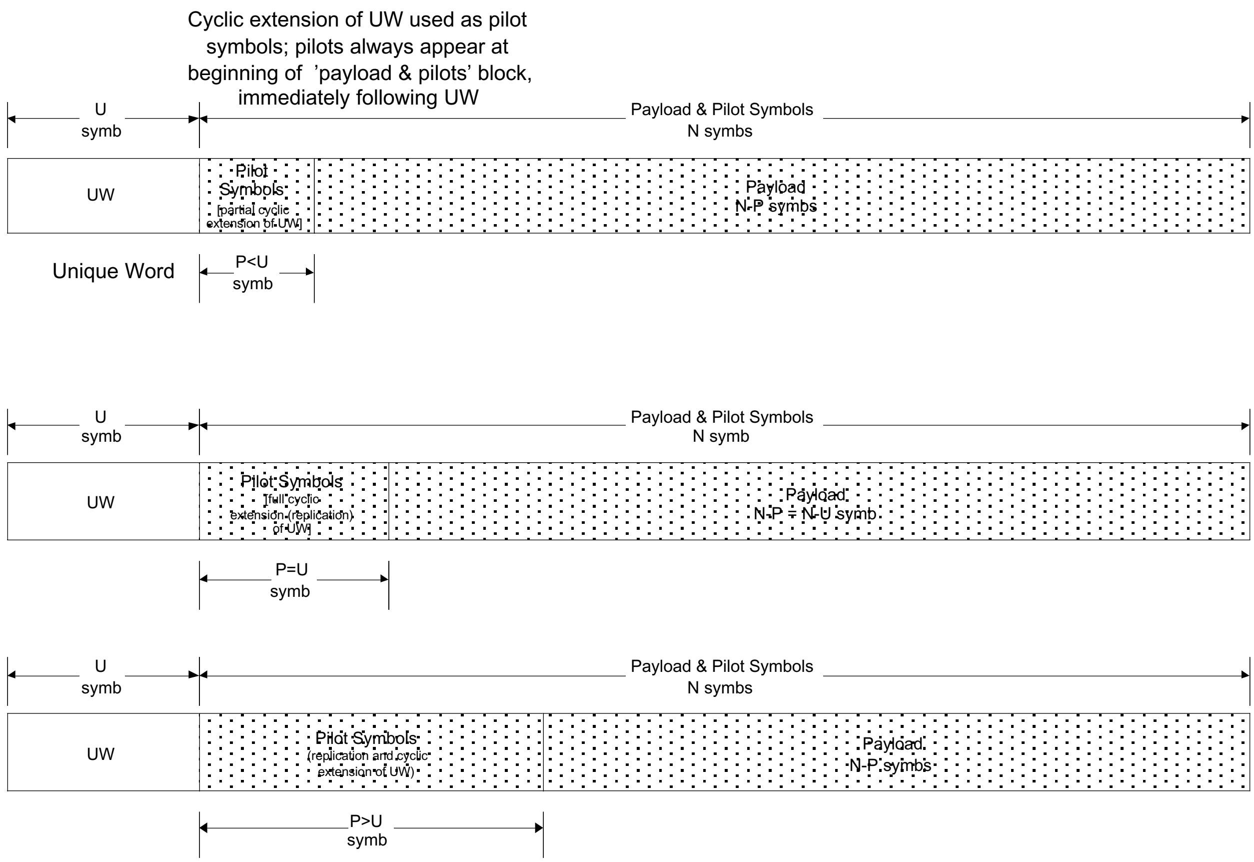
The Unique Word may be used as a cyclic prefix by a frequency domain equalizer, and/or as pilot symbols. When used as a cyclic prefix, a Unique Words should be at least be as long as the maximum delay spread of a channel. As pilot symbols, the Unique Words may assist in the estimation of demodulation parameters such as estimating equalizer channel coefficients, carrier phase and frequency offsets, symbol timing, and optimal FFT window timing (in a frequency domain equalizer).

3.2.1.3.3 Additional Pilot Symbols

When multipath delay spread spans almost the entire Unique Word interval, very little data remains that is not uncorrupted by delay spread from the arbitrary, a priori unknown payload symbols. In such an environment, non-decision aided channel (delay profile) estimation could become exceedingly difficult. One recourse is the increased utilization of decision-aided channel estimation.

To add an extra measure of robustness, many system operators may prefer, instead, to opt for the addition of P additional pilot symbols. For this reason, the addition of an extra P pilot symbols per Unique Word interval is an option, as contiguous cyclic extensions of the Unique Word. A contiguous cyclic extension of P symbols implies that it the pilots are either a repetition of the Unique Word, or a partial replication of the first P symbols of the Unique Word, or a combination of a repetition and partial replication of the Unique Word.

Figure 3.7 illustrates three cases where pilot symbols have been added: one for a case when only a few symbols have been added, another where the number of added pilot symbols and Unique Word symbols are the same (i.e., the Unique Word has been replicated), and one for a case where the number of pilot symbols is much greater than the number of Unique Word symbols (i.e., where the UW is replicated at least once, then cyclically extended.) Note that it is also possible to add zero extra pilot symbols. The range for the addition of pilot symbols, P, is 0†P†N.



8 **Figure 3.7 Three examples where extra pilot symbols have been added, via cyclic extension of the Unique Word [UW]. (top) pilots symbols less than UW symbols; (middle) pilot symbols equal to UW symbols; (bottom) pilot symbols greater than UW symbols.**

The number of additional pilot symbols, P, per N-symbol interval within a particular burst, shall be fixed for that burst. Moreover, many operators may fix P to be the same for all bursts. However, this does not have to be the case. P may change from burst to burst, at the discretion of, and by direction from, the MAC. This information would be contained in the MAC s MAP information, in the field that provides burst profiles.

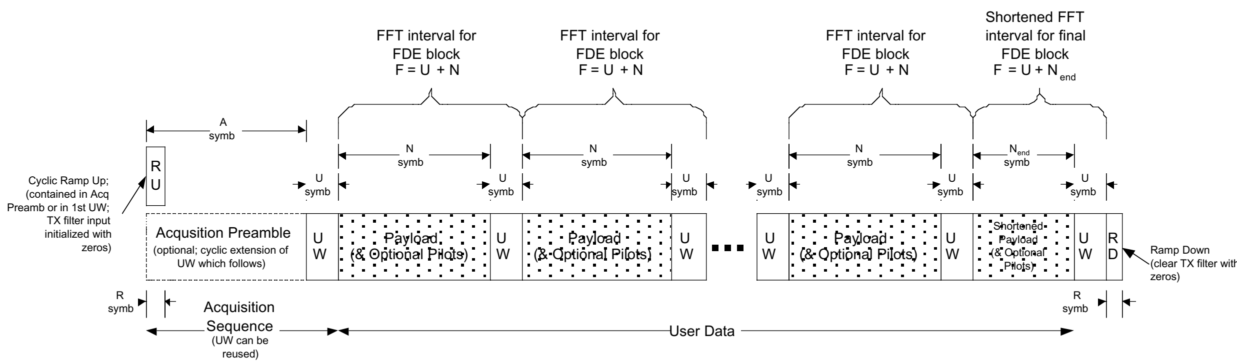
Additional details on the composition of the symbols within Unique Words, and their accepted lengths, U, may be found in 3.2.1.4.2. Additional details related to accepted lengths for F = N + U (where N is the interval between Unique Word repetitions) may be found in 3.2.1.5.

3.2.1.3.4 Burst Element Details Including the Acquisition Preamble

A burst conforming to the TDM or TDMA burst format is illustrated in Figure 3.8. Note that the burst may be long or short. In other words, it may consist of multiple Unique Word Intervals of length F = N + U (i.e, multiple FFTs for a frequency domain equalizer) plus one extra prefixing Unique Word, or a single Unique Word Interval of length F = N + U (single FFT for a frequency domain equalizer) plus one extra prefixing Unique Word. The final payload section may also be shortened. Additional details on the use and intervals of Unique Words within burst formats may be found in 3.2.1.3.2.

As Figure 3.8 illustrates, a burst may possess an optional acquisition preamble, of length A symbols. This optional preamble must be composed of Unique Word symbols that are a contiguous cyclic extension of the Unique Word which follows the preamble. In other words, this contiguous cyclic extension of A symbols implies that if A= U, then the preamble is a repetition of the Unique Word; if A<U, then the preamble is a partial replication of the last A symbols of the Unique Word; otherwise, the preamble is a combination of a repetition and partial replication of the last symbols in the Unique Word.

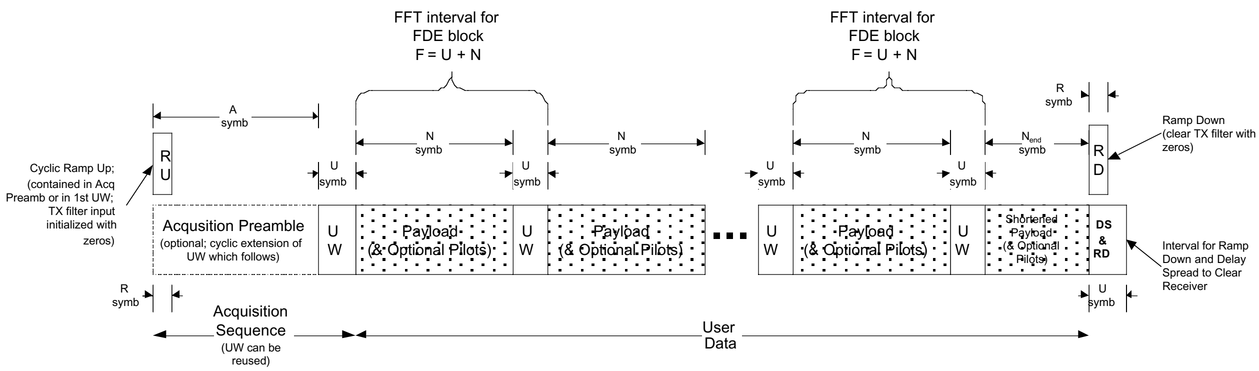
Note that the ramp-up sequence as described in 3.2.1.3.1 may be derived in part from this acquisition preamble, if all zero symbol (null) inputs are suppressed the ramp-up windowing sequence.
Burst profile parameters delivered by the MAC dictate the length of the acquisition preamble. The nominal length of A, used in broadcast system messages containing MAC MAP information, is TBD.



9 **Figure 3.8 TDM/TDMA burst format: element locations and details**

As Figure 3.8 illustrates, the ramp-down sequence is external to the final FFT interval which might be used by a frequency domain equalizer.

Figure 3.9 illustrates another burst approach (profile), which uses less overhead. In this approach, the burst-ending Unique Word is eliminated, and replaced by U-symbol idle data region, within which no data is transmitted. A part of this idle region is the ramp down;the ramp down directly follows the final data-bearing section. The idle interval is necessary for delay spread to clear the receiver. Note that the idle region may be used in receiver processing to integrate more delay spread energy into the decision-making (and equalization) processes. As was the case for Figure 3.8, the length N_{end} in Figure 3.9 should be chosen such that length N_{end} + U adheres to the length guidelines in 0.



10 Figure 3.9 Example of alternative TDM/TDMA burst format where final UW is eliminated, the ramp down interval is advanced, and, in place of the final UW, an idle region is inserted.

3.2.1.3.5 Details Pertinent to TDMA Bursts

A TDMA burst will support only a single modulation type in its data payload, since all bursts are addressed to a single audience. However, subsequent bursts may be transmitted using a different modulation type. What s more, the number of symbols in the acquisition preamble, if any, may be dependent on the burst profile assigned for either that modulation type, or the burst profile for the particular user being addressed by the TDMA burst. Such user profile data would be sent in control bursts, addressed either to a particular user, or to all users in the system, as a whole.

3.2.1.3.6 Details Pertinent to TDM Bursts

Some details, such as adaptive modulation, are only pertinent to the TDM burst format---since the TDMA format only supports a broadcast to one audience.

3.2.1.3.6.1 Adaptive Modulation Sequencing for TDM bursts

Within the MAC Frame header, a PHY control map (MAP) is used to indicate the beginning location of each of adaptive modulation type payload that follows. However, the MAP does not describe the beginning locations of the payload groups that immediately follow; it describes the payload distributions at some MAC-prescribed time in the future. This delay is necessary so that FEC decoding of MAC information (which could be iterative, in the case of turbo codes) may be completed, the adaptive data interpreted, and the demodulator scheduling set up for the proper sequencing. Note that this information containing the distribution of data within a particular burst may be contained in another burst.

Within a burst, payload groups are sequenced in decreasing order of robustness (e.g., first QPSK, then 16-QAM, then 64-QAM). This robustness sequencing improves receiver performance, because it enables receivers experiencing lower CINRs to track only through the modulation types that they can reliably receive. This sequencing also facilitates changes of modulation type at locations that are not contiguous to Unique Word boundaries.

3.2.1.3.6.2 UW Boundary-free Transitions Between Modulation Types (TDM)

Note also that adaptive modulation type-to-other-modulation type changes are not restricted to occur only at Unique Word boundaries. They may change anywhere that a MAC MAP message indicates that they should change.

3.2.1.3.6.3 Per-Adaptive-Modulation-Type FEC Encapsulation (TDM)

So that disadvantaged-SNR SSs are not adversely affected by transmissions intended for other advantaged-CINR users, FEC blocks end when a particular adaptive modulation type ends. Among other things, this implies that the FEC interleaver depth and code blocks are adapted to accommodate the span of a particular adaptive modulation type. Note, however, that data from several users could be concatenated by the MAC (and interleaved together by the PHY) within the span of a given adaptive modulation type.

3.2.1.3.6.4 MAC Header FEC Encapsulation (TDM)

So that the MAC header data may be decoded by a receiver that has just acquired (and does not yet know the modulation lengths and distributions of user data), the MAC header data should be

1. a fixed, a priori-known block size; and
2. separately FEC-encoded (and interleaved) from all other user-specifically-addressed data.

3.2.1.4 Unique Word Details

The Unique Word sequence is omnipresent, appearing in all frame structures, both in the continuous and burst formats.

3.2.1.4.1 Unique Word Sequence Design Criteria

The choice of Unique Word is critical, because it is used as both a Cyclic prefix for frequency domain equalizers, and also for channel estimation. Its cyclic prefix role imposes one constraint: the Unique

Word must be at least as long as the maximum delay spread to be experienced by an intended receiver. Its channel estimation role imposes another constraint: the Unique Word should have good correlation properties, and a broadband, un-notched frequency response. And lastly, since the Unique Word introduces overhead, it should be no longer than it need be; sectors/installations that experience less delay spread should not be burdened with the overhead of excessively long Unique Words. This implies that some flexibility in the choice (or construction) of Unique Words is required.

3.2.1.4.2 Unique Word Sequence Specification

One sequence class that seems to possess all of the desired properties is the modified PN sequence, as described by Milewski in Reference [26]. As the title suggests in Reference [26], this sequence class has optimal properties for channel estimation and fast start-up equalization. What s more, constructions for various sequence lengths are simple, due to their derivation from PN sequences.
The modified PN sequence is a complex-valued (I + jQ) sequence that might be described as quasi-BPSK. It possesses the following structure:

- The I channel component is derived from a PN-generator (linear feedback shift register) of period $U=2^n-1$ (where n is an integer), and
- The Q channel component is a small, but non-zero constant sequence, with value $\frac{1}{\sqrt{2^n-1}}$.

Table 3.1 lists the generator polynomials that must be used in generating the I component of the Unique Word, over a useful and practical range of sequence lengths, U. Support for the lengths U= 31, 63, and 127 is mandatory. Support of all other U lengths is optional.

11 Table 3.1 UW lengths and Generator Polynomials used to Generate PN —Sequence for I Channel (shaded must be supported)

Length, U (symbols)	PN Generator Polynomial (Binary, with 100101 <-> $x^5 + x^2 + 1$)
0	
7	1011
15	10011
31	100101
63	1000011
127	10000011
255	100011101
511	1000010001

3.2.1.5 FFT Interval (UW Interval) Specifications

In addition to the length, U, of the Unique Word sequence, another important framing parameter is the interval between Unique Words, N. However, rather than specifying N directly, we prefer to specify, F = N + U, where F would be the FFT length of a symbol-spaced frequency domain equalizers. The length F = N + U should preferably be 2 raised to an integer power n; in other words, $F = 2^n$.

A receiver and transmitter must support n = 1,2 ,10, which implies that the maximum FFT size which must be supported is $F = 2^{10} = 1024$. Other desirable FFT sizes, for longer delay spread channels, or higher data rates, are n = 11,12,13, i.e., F = 2048, 4096, 8092.

3.2.1.6 Framing Recommendations for Transmit Diversity

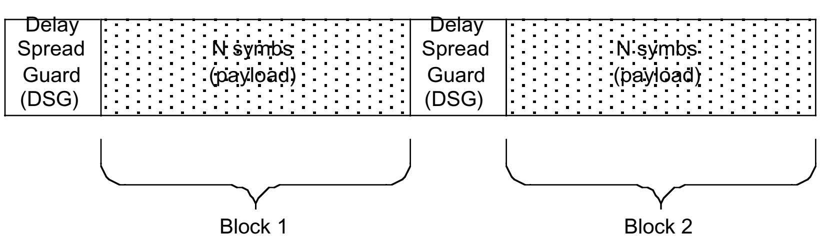
Diversity techniques are likely to find application in some broadband wireless installations. Non-invasive techniques such as receive diversity do not require any special considerations on the part of the air interface, or framing. For 2-way delay transmit diversity, where two transmit antennas are used and the output of the second antenna is delayed with respect to the first, the considerations are minor. Both receiver equalization and framing must be adequate to accommodate the extra delay spread introduced in the system due to the delayed output of the second transmitter.

However, the framing requires some thought when the Alamouti transmit diversity scheme [36], which achieves 2-way maximal ratio transmit diversity combining, is used.

The Alamouti Algorithm:

Alamouti diversity combining may be applied to either the continuous or burst formats, if two consecutive Unique Word Intervals (which we will denote here as blocks) are logically coupled, and are jointly processed at both the transmitter and receiver. Here we shall illustrate a technique which is particularly amenable to frequency domain equalization.

Figure 3.10 illustrates the aforesaid concept of block pairing, and also illustrates the necessity of separating the consecutively paired blocks with delay spread guard bands , so that no block leaks delayed information onto the other.



12 Figure 3.10 Two Blocks (Unique Word Intervals), to which Alamouti transmit diversity combiner processing are intended to be applied

Table 3.2 indicates the block signaling structure that must be used at the transmitter.
Note that Transmit Antenna 0 would transmit data according to burst or continuous format specifications, with no modifications. However, Transmit Antenna 1 must not only reverse the block order, and conjugate the transmitted complex symbols, but must also reverse the time sequence of data within each block before sending data over the air.

13 Table 3.2 Multiplexing arrangement to enable block Alamouti-like processing of delay-spreaded data with a Single Carrier System.

	Block 0	Block 1
Transmit Antenna 0	$s_0(t)$	$s_1(t)$
Transmit Antenna 1	$-s_1^*(-t)$	$s_0^*(-t)$

14

Let $S_0(e^{j\omega})$, $S_1(e^{j\omega})$,$H_0(e^{j\omega})$, $H_0(e^{j\omega})$, $H_1(e^{j\omega})$, $N_0(e^{j\omega})$ and $N_1(e^{j\omega})$ be the Discrete-time Fourier transforms of symbol sequences $s_0(t)$ and $s_1(t)$, channel responses $h_0(t)$ and $h_1(t)$, and additive noise sequences $n_0(t)$ and $n_1(t)$.

The received signals associated with each block, interpreted in the frequency domain, are:

$$\begin{aligned} R_0(e^{j\omega}) &= H_0(e^{j\omega})S_0(e^{j\omega}) - H_1(e^{j\omega})S_1(e^{j\omega}) + N_0(e^{j\omega}) \\ R_1(e^{j\omega}) &= H_0(e^{j\omega})S_1^*(e^{j\omega}) + H_1(e^{j\omega})S_0^*(e^{j\omega}) + N_1(e^{j\omega}) \end{aligned}$$

Equation 3.1

Assuming that the channel responses $H_0(e^{j\omega})$ and $H_1(e^{j\omega})$ are known, one can use the frequency domain combining scheme

$$\begin{aligned} C_0(e^{j\omega}) &= H_0^*(e^{j\omega})R_0(e^{j\omega}) + H_1(e^{j\omega})R_1^*(e^{j\omega}) \\ C_1(e^{j\omega}) &= -H_1(e^{j\omega})R_0^*(e^{j\omega}) + H_0^*(e^{j\omega})R_1(e^{j\omega}), \end{aligned}$$

to obtain the combiner outputs

$$\begin{aligned} C_0(e^{j\omega}) &= \left(\left|H_0(e^{j\omega})\right|^2 + \left|H_1(e^{j\omega})\right|^2\right)S_0(e^{j\omega}) + H_0^*(e^{j\omega})N_0(e^{j\omega}) + H_1(e^{j\omega})N_1^*(e^{j\omega}) \\ C_1(e^{j\omega}) &= \left(\left|H_0(e^{j\omega})\right|^2 + \left|H_1(e^{j\omega})\right|^2\right)S_1(e^{j\omega}) - H_1(e^{j\omega})N_0^*(e^{j\omega}) + H_0^*(e^{j\omega})N_1(e^{j\omega}) \,. \end{aligned}$$

These combiner outputs can be equalized using a frequency domain equalizer (see Reference[20], for example) to (eventually) obtain estimates for $s_0(t)$ and $s_1(t)$.

The channel responses can also be estimated using pilot symbols. Assume that corresponding pilot symbols are the same in the 0 and 1 blocks, i.e.,

$S_0\left(e^{j\omega_{pilot}}\right) = S_1\left(e^{j\omega_{pilot}}\right) \equiv S\left(e^{j\omega_{pilot}}\right)$, and that $S\left(e^{j\omega_{pilot}}\right)$ is known.

Using the expression from Equation 3.1, one can easily show that

$$\begin{aligned} S^*\left(e^{j\omega_{pilot}}\right)R_0\left(e^{j\omega_{pilot}}\right) + S\left(e^{j\omega_{pilot}}\right)R_1\left(e^{j\omega_{pilot}}\right) &= 2\left|S\left(e^{j\omega_{pilot}}\right)\right|^2 H_0\left(e^{j\omega_{pilot}}\right) \\ -S^*\left(e^{j\omega_{pilot}}\right)R_0\left(e^{j\omega_{pilot}}\right) + S\left(e^{j\omega_{pilot}}\right)R_1\left(e^{j\omega_{pilot}}\right) &= 2\left|S\left(e^{j\omega_{pilot}}\right)\right|^2 H_1\left(e^{j\omega_{pilot}}\right) \end{aligned}.$$

This suggests that one can estimate the channels $H_0\left(e^{j\omega}\right)$ and $H_1\left(e^{j\omega}\right)$ at the pilot locations, and thus identify the channels themselves (if the pilot sampling locations are selected properly) using the expressions

$$\begin{aligned} H_0\left(e^{j\omega_{pilot}}\right) &= \frac{S^*\left(e^{j\omega_{pilot}}\right)R_0\left(e^{j\omega_{pilot}}\right) + S\left(e^{j\omega_{pilot}}\right)R_1\left(e^{j\omega_{pilot}}\right)}{2\left|S\left(e^{j\omega_{pilot}}\right)\right|^2} \\ H_1\left(e^{j\omega_{pilot}}\right) &= \frac{-S^*\left(e^{j\omega_{pilot}}\right)R_0\left(e^{j\omega_{pilot}}\right) + S\left(e^{j\omega_{pilot}}\right)R_1\left(e^{j\omega_{pilot}}\right)}{2\left|S\left(e^{j\omega_{pilot}}\right)\right|^2} \end{aligned}.$$

Figure 3.11 illustrates a frame structure, with pilot symbols (Unique Word repetitions) which enables implementation of the aforesaid techniques, including simultaneous estimation (or channel updates) of the two channels arising from the use of two transmit antennas. Note that although the spacing between basic Unique Words is the same as previously, the intervals over which FFTs (for a frequency domain equalizer) are computed are reduced.

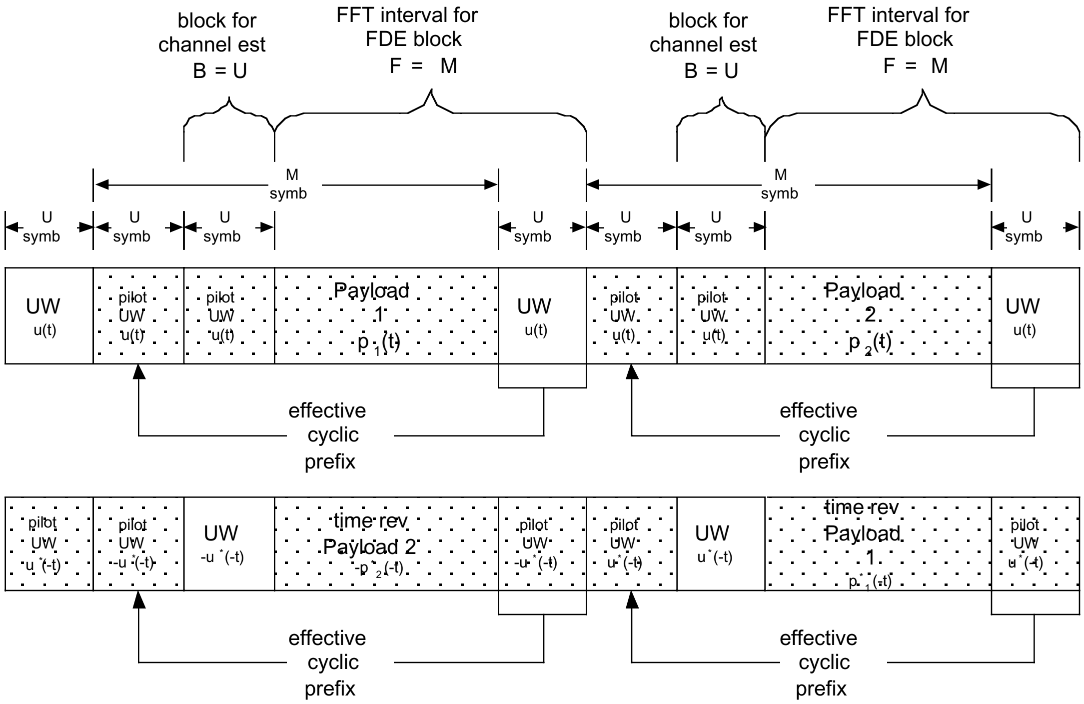
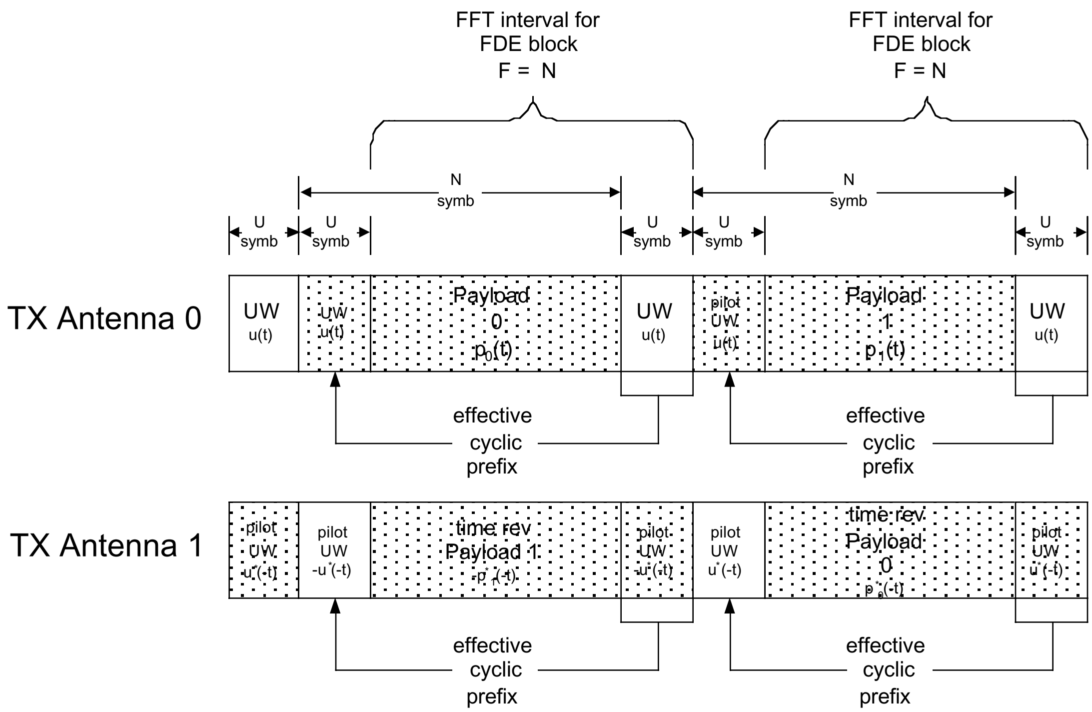


Figure 3.11 Frame structure suitable for Alamouti transmit diversity signaling and associated channel estimation, when channel estimation is required

Figure 3.12 illustrates a similar case as Figure 3.11, but where channel estimates and/or channel updates are not needed. This case might occur with in the burst format applications, where the channels might be estimated with sufficient accuracy using information in the acquisition preamble.



16 **Figure 3.12 Frame structure suitable for Alamouti transmit diversity signaling when channel estimation is not needed**

3.2.2 MAC and PHY Interface Layer

3.2.2.1 Overview

Two modes of operation have been defined for the point-to-multi-point downlink channel:

- One targeted to support a **Continuous transmission** stream format, and
- One targeted to support a **Burst transmission** stream format.

Having this separation allows each format to be optimized according to its respective design constraints, while resulting in a standard that supports various system requirements and deployment scenarios.

In contrast, only one mode of operation is defined for the **Upstream channel**:

- One targeted to support a **Burst transmission** stream format.

This single mode of operation is sufficient for the upstream, since the upstream transmissions are point-to-point burst transmissions between each transmitting Subscriber Station (SS) and each receiving Base Station (BS).

3.2.2.1.1 Downlink and Uplink Operation

Two different downlink modes of operation are defined: Mode A and Mode B. Mode A supports a continuous transmission format, while Mode B supports a burst transmission format. The continuous transmission format of Mode A is intended for use in an FDD-only configuration. The burst transmission format of Mode B supports burst-FDD as well as TDD configurations.

The Mode A and B options give service providers choice, so that they may tailor an installation to best meet a specific set of system requirements. Standards-compliant subscriber stations are required to support at least one (Mode A or Mode B) of the defined downlink modes of operation.

A single uplink mode of operation is also defined. This mode supports TDMA-based burst uplink transmissions. Standards-compliant subscriber stations are required to support this uplink mode of operation.

3.2.2.1.1.1 Mode A (Continuous Downlink)

Mode A is a downlink format intended for continuous transmission. The Mode A downlink physical layer first encapsulates MAC packets into a convergence layer frame as defined by the transmission convergence sublayer. Modulation and coding which is adaptive to the needs of various SS receivers is also supported within this framework.

Data bits derived from the transmission convergence layer are first randomized. Next, they are block FEC encoded. The resulting FEC-encoded bits are mapped to QPSK, 16-QAM, or 64-QAM signal constellations. Detailed descriptions of the FEC, modulation constellations, and symbol mapping formats can be found within the FEC and modulation sections. Following the symbol mapping process, the resulting symbols are modulated, and then transmitted over the channel.

In Mode A, the downstream channel is continuously received by many SSs. Due to differing conditions at the various SS sites (e.g., variable distances from the BS, presence of obstructions), SS receivers may observe significantly different SNRs. For this reason, some SSs may be capable of reliably detecting data only when it is derived from certain lower-order modulation alphabets, such as QPSK. Similarly, more powerful and redundant FEC schemes may also be required by such SNR-disadvantaged SSs. On the other hand, SNR-advantaged stations may be capable of receiving very high order modulations (e.g., 64-QAM) with high code rates. Collectively, let us define the adaptation of modulation type and FEC to a particular SS (or group of SSs) as 'adaptive modulation', and the choice of a particular modulation and FEC as an 'adaptive modulation type.' Mode A supports adaptive modulation and the use of adaptive modulation types.

A MAC Frame Control header is periodically transmitted over the continuous Mode A downstream channel, using the most robust supported adaptive modulation type. So that the start of this MAC header may be easily recognized during initial channel acquisition or re-acqusition, the PHY inserts an uncoded, TBD (but known) QPSK code word, of length TBD symbols, at a location immediately before the beginning of the MAC header, and immediately after a Unique Word. (See PHY framing section for more details on the Unique Word). Note that this implies the interval between Frame Control headers should be an integer multiple of F (the interval between Unique Words).

Within MAC Frame Control header, a PHY control map (DL_MAP) is used to indicate the beginning location of adaptive modulation type groups which follow. Following this header, adaptive modulation groups are sequenced in increasing order of robustness.
However, the DL_MAP does not describe the beginning locations of the payload groups that immediately follow; it describes the payload distributions some MAC-prescribed time in the future. This delay is necessary so that FEC decoding of MAC information (which could be iterative, in the case of turbo codes) may be completed, the adaptive data interpreted, and the demodulator scheduling set up for the proper sequencing.

Note that adaptive modulation groups or group memberships can change with time, in order to adjust to changing channel conditions.

In order that disadvantaged SNR users are not adversely affected by transmissions intended for other advantaged SNR users, FEC blocks end when a particular adaptive modulation type ends. Among other things, this implies that the FEC interleaver depth is adapted to accommodate the span of a particular adaptive modulation type.

3.2.2.1.1.2 Mode B (Burst Downlink)

Mode B is a downlink format intended for burst transmissions, with features that simplify the support for both TDD systems and half-duplex terminals. A Mode B compliant frame can be configured to support either TDM or TDMA transmission formats; i.e., a Mode B burst may consist of a single user's data, or a concatenation of several users' data. What's more, Mode B supports adaptive modulation and multiple adaptive modulation types within these TDMA and TDM formats.

A unique (acquisition) preamble is used to indicate the beginning of a frame, and assist burst demodulation. This preamble is followed by PHY/MAC control data. In the TDM mode, a PHY control map (DL_MAP) is used to indicate the beginning location of different adaptive modulation types. These adaptive modulation types are sequenced within the frame in increasing order of robustness (e.g., QPSK, 16-QAM, 64-QAM), and can change with time in order to adjust to the changing channel conditions.

In the TDMA mode, the DL_MAP is used to describe the adaptive modulation type in individual bursts. Since a TDMA burst would contain a payload of only one adaptive modulation type, no adaptive modulation type sequencing is required. All TDMA format payload data is FEC block encoded, with an allowance made for shortening the last codeword (e.g., Reed Solomon codeword) within a burst.

The Mode B downlink physical layer goes through a transmission convergence sublayer that inserts a pointer byte at the beginning of the payload information bytes to help the receiver identify the beginning of a MAC packet.

Payload data bits coming from the transmission convergence layer are first randomized. Next, they are block FEC encoded. The resulting FEC-encoded bits are mapped to QPSK, 16-QAM, or 64-QAM signal constellations. Detailed descriptions of the FEC, modulation constellations, and symbol mapping formats can be found within the FEC and modulation sections. Following the symbol mapping process, the resulting symbols are modulated, and then transmitted over the channel.

3.2.2.1.1.3 Uplink Access

The uplink mode supports TDMA burst transmissions from an individual SS to a BS. This is functionally similar (at the PHY level) to Mode B downlink TDMA operation. As such, for a brief description of the Physical Layer protocol used for this mode, please read the previous section on Mode B TDMA operation.

Of note, however, is that many of the specific uplink channel parameters can be programmed by MAC layer messaging coming from the base station in downstream messages. Also, several parameters can be left unspecified and configured by the base station during the registration process in order to optimize performance for a particular deployment scenario. In the upstream mode of operation, each burst may carry MAC messages of variable lengths.

3.2.2.2 Multiplexing and Multiple Access Technique

The uplink physical layer is based on the combined use of time division multiple access (TDMA) and demand assigned multiple access (DAMA). In particular, the uplink channel is divided into a number of 'time slots'. The number of slots assigned for various uses (registration, contention, guard, or user traffic) is controlled by the MAC layer in the base station and can vary over time for optimal performance.

As previously indicated, the downlink channel can be in either a continuous (Mode A) or burst (Mode B) format. Within Mode A, user data is transported via time division multiplexing (TDM), i.e., the information for each subscriber station is multiplexed onto the same stream of data and is received by all subscriber stations located within the same sector. Within Mode B, the user data is bursty and may be transported via TDM or TDMA, depending on the number of users that are to be serviced within the burst.

3.2.2.2.1 Duplexing Techniques

Several duplexing techniques are supported, in order to provide greater flexibility in spectrum usage. The continuous transmission downlink mode (Mode A) supports Frequency Division Duplexing (FDD) with adaptive modulation; the burst mode of operation (Mode B) supports FDD with adaptive modulation or Time Division Duplexing (TDD) with adaptive modulation. Furthermore, Mode B in the FDD case can handle (half duplex) subscribers incapable of transmitting and receiving at the same instant, due to their specific transceiver implementation.

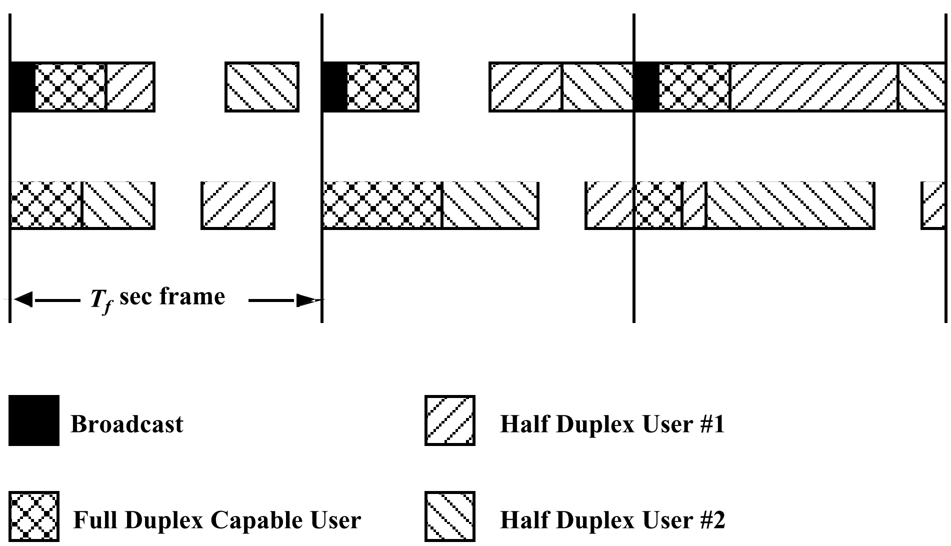
3.2.2.2.1.1 Mode A: Continuous Downstream Transmission for FDD Systems

In a system employing FDD, the uplink and downlink channels are located on separate frequencies and all subscriber stations can transmit and receive simultaneously. The frequency separation between carriers is set either according to the target spectrum regulations or to some value sufficient for complying with radio channel transmit/receive isolation and de-sensitization requirements. In this type of system, the downlink channel is (almost) always on and all subscriber stations are always listening to it. Therefore, traffic is sent in a broadcast manner using time division multiplexing (TDM) in the downlink channel, while the uplink channel is shared using time division multiple access (TDMA), where the allocation of uplink bandwidth is controlled by a centralized scheduler. The BS periodically transmits downlink and uplink MAP messages, which are used to synchronize the uplink burst transmissions with the downlink. The usage of the mini-slots is defined by the UL-MAP message, and can change according to the needs of the system. Mode A is capable of adaptive modulation.

3.2.2.2.1.2 Mode B: Burst Downstream Transmission for Burst FDD Systems

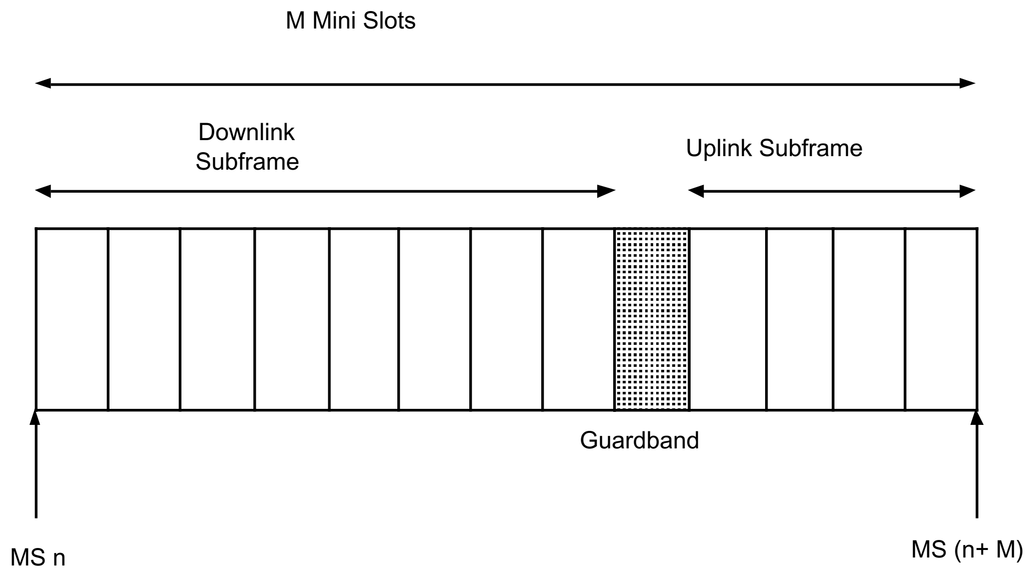
A burst FDD system refers to a system in which the uplink and downlink channels are located on separate frequencies but the downlink data is transmitted in bursts. This feature enables the system to simultaneously support full duplex subscriber stations (ones which can transmit and receive simultaneously) and, optionally, half duplex Subscriber Stations (ones which cannot transmit and receive simultaneously). If half duplex subscriber stations are supported, this mode of operation imposes a restriction on the bandwidth controller: it cannot allocate uplink bandwidth for a half duplex subscriber station at the same time that the subscriber station is expected to receive data on the downlink channel.

Frequency separation is as defined in 3.2.2.1.1.1 and Figure 3.13 illustrates the basics of the burst FDD mode of operation. In order to simplify the bandwidth allocation algorithms, the uplink and downlink channels are divided into fixed sized frames. A full duplex subscriber station must always attempt to listen to the downlink channel. A half duplex subscriber station must always attempt to listen to the downlink channel when it is not transmitting on the uplink channel.



17 **Figure 3.13: Example of Burst FDD bandwidth Allocation.**

3.2.2.2.1.3 Mode B: Burst Downstream for Time Division Duplexing (TDD) Systems



18 Figure 3.14: TDD Frame Structure

In the case of TDD, the uplink and downlink transmissions share the same frequency, but are separated in time (Figure 3.14). A TDD frame also has a fixed duration and contains one downlink and one uplink subframe. The frame is divided into an integer number of 'mini slots' (MS), which facilitate the partitioning of bandwidth. These mini slots are in turn made up of a finer unit of time called 'ticks', which are of duration 1 us each. TDD framing is adaptive in that the percentage of the bandwidth allocated to the downlink versus the uplink can dynamically vary. The split between uplink and downlink is a system parameter, and is controlled at higher layers within the system.

3.2.2.2.1.3.1 Tx /Rx Transition Gap (TTG)

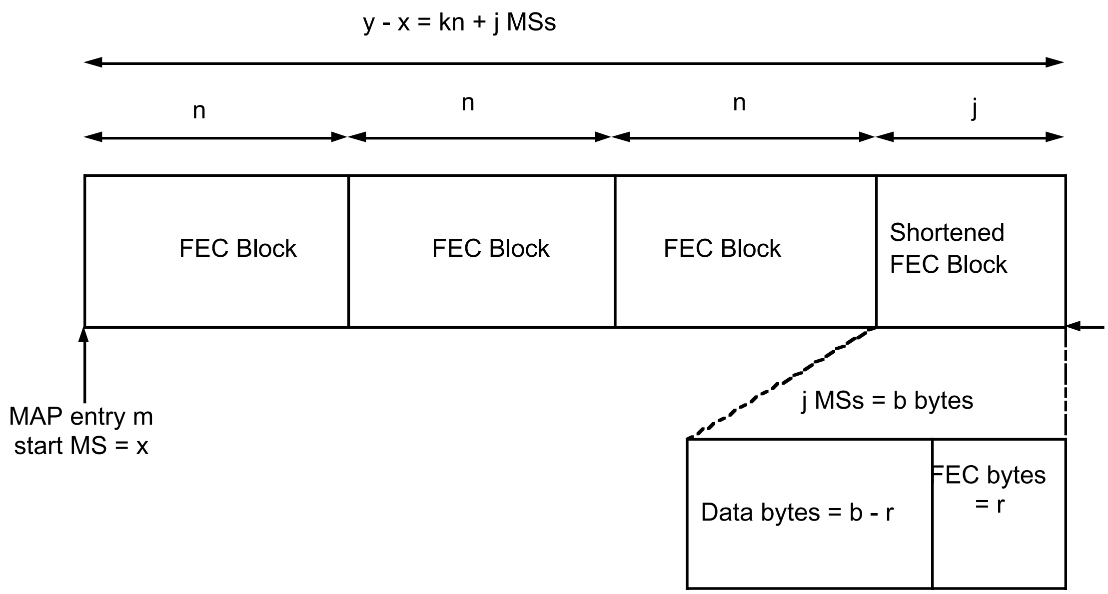
The TTG is a gap between the Downlink burst and the Uplink burst within a TDD system. The TTG allows time for the BS to switch from transmit mode to receive mode and SSs to switch from receive mode to transmit mode. During this interval, the BS and SS do not transmit modulated data. Therefore, the BS transmitter may ramp down, Tx / Rx antenna switches on both sides may actuate, the SS transmitter may ramp up, and the BS receiver section may activate. After the TTG, the BS receiver will look for the first symbols of uplink burst. The TTG has a variable duration, which is an integer number of mini slots. The TTG starts on a mini slot boundary.

3.2.2.2.1.3.2 Rx /Tx Transition Gap (RTG)

The RTG is a gap between the Uplink burst and the Downlink burst. The RTG allows time for the BS to switch from receive mode to transmit mode, and SSs to switch from transmit mode to receive mode. During this interval, the BS and SS do not transmit modulated data. Therefore, an SS transmitter may ramp down, delay spread may clear the BS receiver, the Tx / Rx antenna switch to actuate on both links, the BS transmitter may ramp up, and the SS receiver sections may activate. After the RTG, the SS receivers will look for the first symbols of modulated acquisition sequence data in the downlink burst. The RTG is an integer number of mini slots. The RTG starts on a mini slot boundary.

3.2.2.2.1.4 Mode B: Downlink Data

The downlink data sections are used for transmitting data and control messages to specific SSs. This data is always FEC coded and is transmitted at the current operating modulation of the individual SS. In the burst mode cases, data is transmitted in robustness order in the TDM portion. In a burst TDMA application, the data is grouped into separately delineated bursts, which do not need to be in modulation order. The DL-MAP message contains a map stating at which mini slot the burst profile change occurs. If the downlink data does not fill the entire downlink sub-frame and Mode B is in use, the transmitter is shut down. The DL-MAP provides implicit indication of shortened



19 Figure 3.15: Downlink MAP usage and Shortened FEC Blocks

FEC (and/or FFT) blocks in the downlink. Shortening the last FEC block of a burst is optional. The downlink map indicates the number of MS, p allocated to a particular burst and also indicates the burst type (modulation and FEC). Let n denote the number of MS required for one FEC block of the given burst profile. Then, $p= kn + j,$ where k is the number of integral FEC blocks that fit in the burst and j is the number of MS remaining after integral FEC blocks are allocated. Either k or j, but not both, may be zero. j denotes some number of bytes b. Assuming j is not 0, it must be large enough such that b is larger than the number of FEC bytes r, added by the FEC scheme for the burst. The number of bytes available to user data in the shortened FEC block is $b - r$. These points are illustrated in Figure 3.15. Note that a codeword may not possess less than 6 information bytes.

In the TDM mode of operation, SSs listen to all portions of the downlink burst to which they are capable of listening. For full-duplex SSs, this implies that a SS shall listen to all portions that have an adaptive modulation type (as defined by the DIUC) which is at least as robust as that which the SS negotiates with the BS. For half-duplex SSs, the aforesaid is also true, but under an additional condition: an SS shall not attempt to listen to portions of the downlink burst that are coincident---adjusted by the SS's Tx time advance---with the SS's allocated uplink transmission, if any.

In the burst TDMA mode of operation, bursts are individually identified in the DL_MAP. Hence, a SS is required to turn on its receiver only in time to receive those bursts addressed to it. Unlike the TDM mode, there is no requirement that the bursts be ordered in order of increasing robustness.

3.2.2.2.2 Uplink Burst Subframe Structure

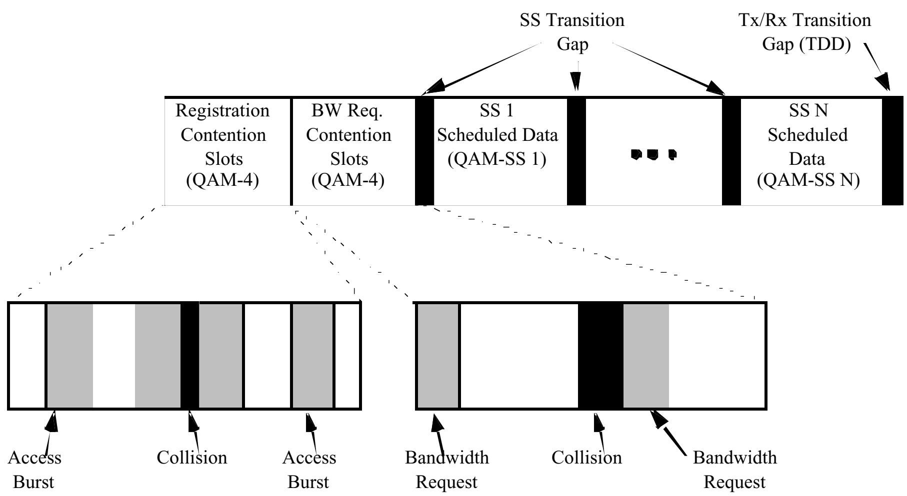
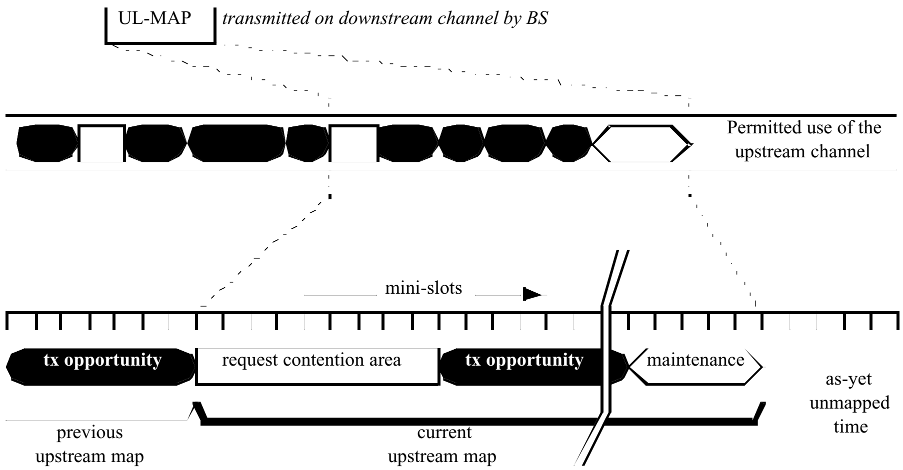


Figure 3.16: Uplink Subframe Structure.

The structure of an uplink subframe used by SSs to transmit with a BS is shown in Figure 3.16. Three main classes of bursts are transmitted by SSs during an uplink subframe:

a) Those that are transmitted in contention slots reserved for station registration.
b) Those that are transmitted in contention slots reserved for response to multicast and broadcast polls for bandwidth needs.
c) Those that are transmitted in bandwidth specifically allocated to individual SSs.

3.2.2.2.2.1 Mode A and Mode B: Uplink Burst Profile Modes



21 **Figure 3.17: Uplink Mapping in the Continuous Downstream FDD Case.**

The uplink uses adaptive burst profiles, in which the base station assigns different modulation types to different SSs. In the adaptive case, the bandwidth allocated for registration and request contention slots is grouped together and is always used with the parameters specified for Request Intervals (UIUC=1). (Remark: It is recommended that UIUC=1 will provide the most robust burst profile due to the extreme link budget and interference conditions of this case). The remaining transmission slots are grouped by SS. During its scheduled allocation, an SS transmits with the burst profile specified by the base station. Considerations which may influence this specification include the effects of distance, interference and environmental factors on transmission to and from that SS. SS Transition Gaps (STG) separate the transmissions of the various SSs during the uplink subframe. The STGs contain a gap to allow for ramping down of the previous burst, followed by a preamble allowing the BS to synchronize to the new SS. The preamble and gap lengths are broadcast periodically in a UCD message. Shortening of FEC and/or FFT blocks in the uplink is identical to the handling in the downlink, as described in 3.2.2.1.4.

3.2.3 Downlink Modes of Operation

This section describes the two different downlink modes of operation that have been adopted for use in this proposal. Mode A has been designed for continuous transmission formats, while Mode B has been designed to support burst transmission formats. Subscriber stations must support at least one of these downlink modes.

3.2.3.1.1 Physical layer type (PHY type) encodings

The value of the PHY type parameter as defined must be reported as shown in Table 3.3.

Table 3.3: Mode Selection Parameters.

Mode	Value	Comment
Mode B (TDD)	0	Burst Downlink in TDD Mode
Mode B (FDD)	1	Burst Downlink in FDD Mode
Mode A (FDD)	2	Continuous Downlink

3.2.3.1.2 Mode A: Continuous Downlink Transmission

This mode of operation has been designed for a continuous transmission stream in a FDD system. The physical media dependent sublayer has no explicit frame structure, other than the incorporation of regular pilot symbols. Adaptive modulation and multiple adaptive modulation types are supported.

3.2.3.1.3 Downlink Mode A: Message field definitions

3.2.3.1.3.1 Downlink Mode A: Required channel descriptor parameters

The following parameters shall be included in the UCD message:

TBD

3.2.3.1.3.2 Mode A:Required DCD parameters

The following parameters shall be included in the DCD message:

TBD

3.2.3.1.3.2.1 Downlink Mode A: DCD, Required burst descriptor parameters

TBD

3.2.3.1.3.3 Mode A: DL-MAP

For PHY Type = 2, a number of information elements follows the Base Station ID field. The MAP information elements must be in time order. Note that this is not necessarily IUC order or connection ID order.

3.2.3.1.3.3.1 Mode A: DL-MAP PHY Synchronization Field definition

Uplink Timestamp[31:16]
Uplink Timestamp[15:0]

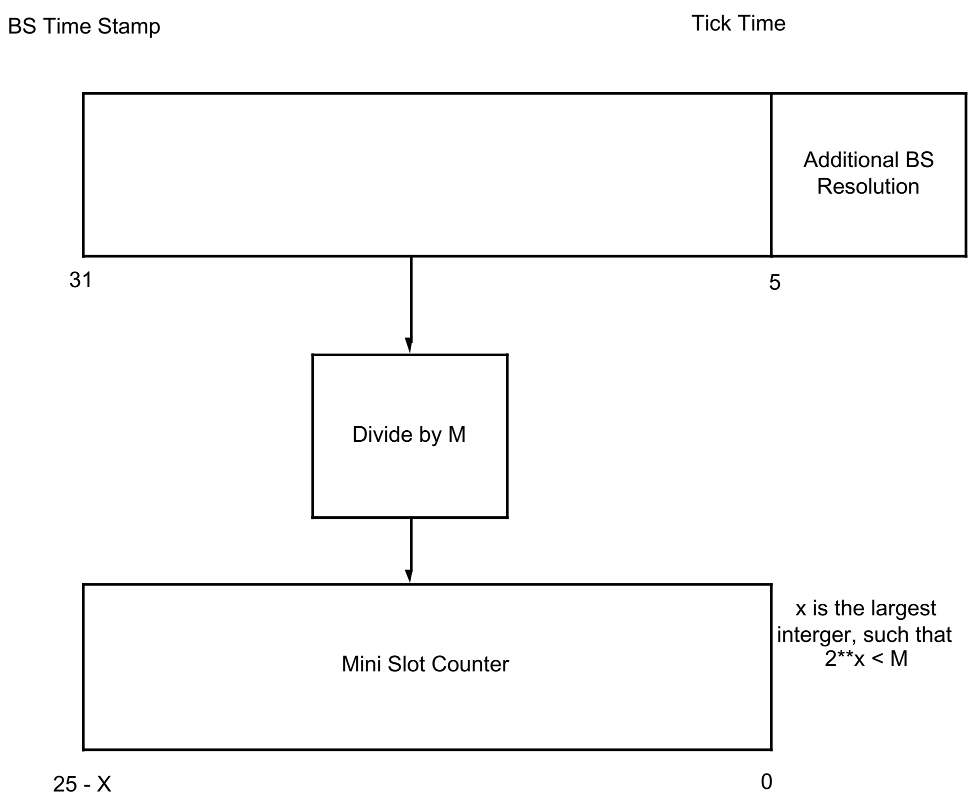
22 Figure 3.18: PHY Synchronization Field (PHY Type 2).

The format of the PHY Synchronization field is given in Figure 3.18. The Uplink Timestamp jitter must be less than 500 ns peak-to-peak at the output of the Downlink Transmission Convergence Sublayer. This jitter is relative to an ideal Downlink Transmission Convergence Sublayer that transfers the TC packet data to the Downlink Physical Media Dependent Sublayer with a perfectly continuous and smooth clock at symbol rate. Downlink Physical Media Dependent Sublayer processing shall not be considered in timestamp generation and transfer to the Downlink Physical Media Dependent Sub-layer. Thus, any two timestamps N1 and N2 ($N2 > N1$) which were transferred to the Downlink Physical Media Dependent Sublayer at times T1 and T2 respectively must satisfy the following relationship:

$$(N2 - N1)/(4 \times \text{Symbol Rate}) - (T2 - T1) < 500 \text{ ns.}$$

The jitter includes inaccuracy in timestamp value and the jitter in all clocks. The 500ns allocated for jitter at the Downlink Transmission Convergence Sublayer output must be reduced by any jitter that is introduced by the Downlink Physical Media Dependent Sublayer.

3.2.3.1.3.4 Mode A: UL-MAP Allocation Start Time definition



23 **Figure 3.19: Maintained Time Stamp Relation between the BS to the BS Mini-slot Counters.**

The Alloc Start Time is the effective start time of the uplink allocation defined by the UL-MAP or DL_MAP in units of mini-slots. The start time is relative to the time of BS initialization (PHY Type = 5). The UL-MAP/DL_MAP Allocation Start Time is given as an offset to the Time Stamp defined in 3.2.4.3.3.1. Figure 3.19 illustrates the relation of the Time Stamp maintained in the BS to the BS Mini-slot Counter. The base time unit is called a tick and is of duration 1 us, independent of the symbol rate, and is counted using a 26 bit counter. The additional BS resolution is of duration (1 tick/ 64) = 15.625 ns. The Mini-Slot count is derived from the tick count by means of a divide by M operation. Note that the **divisor M is not necessarily a power of 2**.

For arbitrary symbol rates, the main constraint in the definition of a mini slot, is that the number of symbols per mini slot be an integer. For example given a symbol rate of R Symbols/tick, and M ticks/mini-slot, the number of symbols per mini-slot N, is given by N = M*R. In this situation, M should be chosen such that N is an integer. In order to accommodate a wide range of symbol rates, it is important not to constrain M to be a power of 2. Since the additional BS resolution is independent of the symbol rate, the system can use a uniform time reference for distance ranging.

In order to show that the time base is applicable to single carrier and OFDM symbol rates, consider the following examples: (a) Single Carrier System - Given a symbol rate of 4.8 Msymbols/s (on a 6MHz channel), if the mini-slot duration is chosen to be 10 ticks (i.e., M = 10), then there are 48 symbols/mini-slot. Given 16QAM modulation this corresponds to a granularity of 24 bytes/mini-slot (b) OFDM System - Given an OFDM symbol time of 50 _s, the mini-slot duration is also chosen to be 50 ticks (i.e., M = 50). In this case there is only a single OFDM symbol per mini-slot.

3.2.3.1.3.5 UL-MAP Ack Time definition

The Ack Time is the latest time processed in uplink in units of mini-slots. This time is used by the SS for collision detection purposes. The Ack Time is given relative to the BS initialization time.

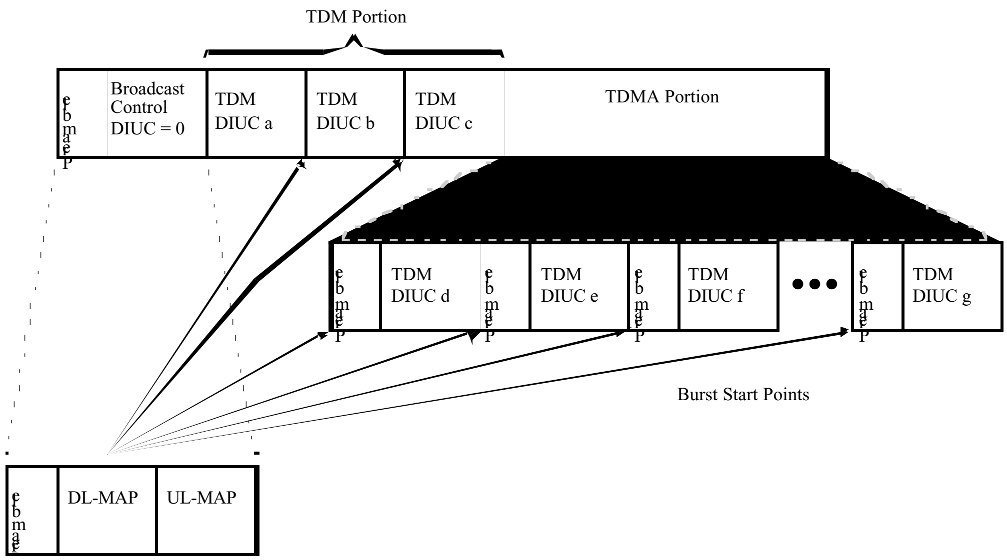
3.2.3.1.4 Mode B: Burst Downlink Transmission

Mode B supports burst transmission on the downlink channel. In particular, this mode is applicable for systems using TDD, which requires a burst capability in the downlink channel. In order to simplify phase recovery and channel tracking, a fixed frame time is used. At the beginning of every frame, an acquisition sequence/preamble is transmitted in order to allow for phase recovery and equalization training. A description of the framing mechanism and the structure of the frame is further described in 3.2.2.4.5.1.

3.2.3.1.4.1 Mode B: Downlink Framing

In the burst mode, the uplink and downlink can be multiplexed in a TDD fashion as described in subsection 3.2.2.2.1.3, or in an FDD fashion as described in 3.2.2.1.2. Each method uses a frame with duration as specified in subsection 3.2.2.5.1. Within this frame are a downlink subframe and an uplink subframe. In the TDD case, the downlink subframe comes first, followed by the uplink subframe. In the burst FDD case, uplink transmissions occur during the downlink frame. In both cases, the downlink subframe is prefixed with information necessary for frame synchronization.

The available bandwidth in both directions is defined with a granularity of one mini slot (MS). The number of mini slots within each frame is independent of the symbol rate. The frame size is selected in order to obtain an integral number of MS within each frame. For example, with a 10 us MS duration, there are 500 MS within a 5-ms frame, independent of the symbol rate.



24 Figure 3.20: Mode B Downlink Subframe Structure.

The structure of the downlink subframe used by the BS to transmit to the SSs, using Mode B, is shown in Figure 3.20. This burst structure defines the downlink physical channel. It starts with a Frame Control Header, that is always transmitted using the most robust set of PHY parameters. This frame header contains a preamble used by the PHY for synchronization and equalization. It also contains control sections for both the PHY and the MAC (DL_MAP and UL_MAP control messages) that is encoded with a fixed FEC scheme defined in this standard in order to ensure interoperability. The Frame Control Header also may periodically contain PHY Parameters as defined in the DCD and UCD.

There are two ways in which the downstream data may be organized for Mode B systems:

- Transmissions may be organized into different modulation and FEC groups, where the modulation type and FEC parameters are defined through MAC layer messaging. The PHY Control portion of the Frame Control Header contains a downlink map stating the MSs at which the different modulation/FEC groups begin. Data should be transmitted in order of decreasing robustness. For modulations this means QPSK followed by 16-QAM, followed by 64-QAM. If more than 1 FEC is defined (via DCD messages) for a given modulation, the more robust FEC / modulation combination appears first. Each SS receives and decodes the control information of the downstream and looks for MAC headers indicating data for that SS.
- Alternatively, transmissions need not be ordered by robustness. The PHY control portion contains a downlink map stating the MS (and modulation / FEC) of each of the TDMA sub-bursts. This allows an individual SS to decode a specific portion of the downlink without the need to decode the whole DS burst. In this particular case, each transmission associated with different burst types is required to start with a short preamble for phase re-synchronization.

There is a Tx / Rx Transition Gap (TTG) separating the downlink subframe from the uplink subframe in the case of TDD.

3.2.3.1.4.2 Frame Control

The first portion of the downlink frame is used for control information destined for all SS. This control information must not be encrypted. The information transmitted in this section is always transmitted using the well known DL Burst Type with UIUC=0. This control section must contain a DL-MAP message for the channel followed by one UL-MAP message for each associated uplink channel. In addition it may contain DCD and UCD messages following the last UL-MAP message. No other messages may be sent in the PHY/MAC Control portion of the frame.

3.2.3.1.4.3 Downlink Mode B: Required DCD parameters

TBD

3.2.3.1.4.3.1 Downlink Mode B: DCD, Required burst descriptor parameters

TBD

3.2.3.1.4.4 Downlink Mode B: Required UCD parameters

TBD

3.2.3.1.4.5 Downlink Mode B: DL-MAP elements

For PHY Type = {0, 1}, a number of information elements follows the Base Station ID field. The MAP information elements must be in time order. Note that this is not necessarily IUC order or connection ID order.

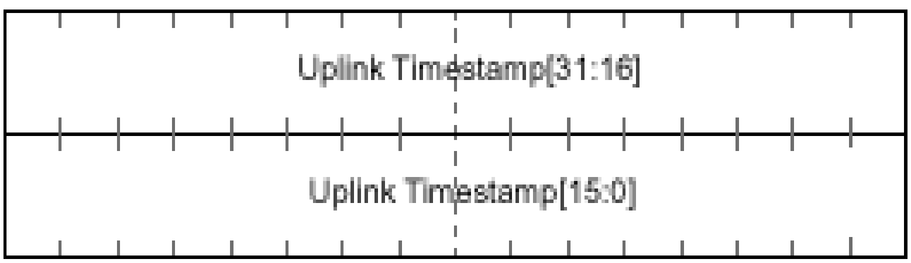
3.2.3.1.4.6 Allowable frame times

Table 3.4 indicates the various frame times that are allowed for the current downlink Mode B physical layer. The actual frame time used by the downlink channel can be determined by the periodicity of the frame start preambles.

25 **Table 3.4 - Allowable Frame Times**

Frame Length Code	Frame Time	Units
0x01	0.5	ms
0x02	1.0	ms
0x03	1.5	ms
0x04	2.0	ms
0x05	2.5	ms
ox06	3.0	ms
0x07	3.5	ms
0x08	4.0	ms
0x09	4.5	ms
0x0A	5.0	ms

3.2.3.1.4.6.1 Mode B: DL-MAP PHY Synchronization Field definition



26 **Figure 3.21: PHY Synchronization Field (PHY Type = {0,1})**

The format of the PHY Synchronization field is given in Figure 3.21. The Uplink Timestamp jitter must be less than 500 ns peak-to-peak at the output of the Downlink Transmission Convergence Sublayer. This jitter is relative to an ideal Downlink Transmission Convergence Sublayer that transfers the TC packet data to the Downlink Physical Media Dependent Sublayer with a perfectly continuous and smooth clock at symbol rate. Downlink Physical Media Dependent Sublayer processing shall not be considered in timestamp generation and transfer to the Downlink Physical Media Dependent Sub-layer. Thus, any two timestamps N1 and N2 (N2 > N1) which were transferred to the Downlink Physical Media Dependent Sublayer at times T1 and T2 respectively must satisfy the following relationship:

$$(N2 - N1)/(4 \times \text{Symbol Rate}) - (T2 - T1) < 500 \text{ ns}$$

The jitter includes inaccuracy in timestamp value and the jitter in all clocks. The 500ns allocated for jitter at the Downlink Transmission Convergence Sublayer output must be reduced by any jitter that is introduced by the Downlink Physical Media Dependent Sublayer.

3.2.3.1.4.7 Mode A: UL-MAP Allocation Start Time definition

The Allocation Start Time is the effective start time of the uplink allocation defined by the UL-MAP or DL_MAP in units of mini-slots. The start time is relative to the time of BS initialization (PHY Type = 5). The UL-MAP/DL_MAP Allocation Start Time is given as an offset to the Time Stamp defined in 3.2.4.3.3.1. Figure 3.19 illustrates the relation of the Time Stamp maintained in the BS to the BS Mini-slot Counter. The base time unit is called a tick and is of duration 1 us, independent of the symbol rate, and is counted using a 26 bit counter. The additional BS resolution is of duration (1 tick/ 64) = 15.625 ns. The Mini-Slot count is derived from the tick count by means of a divide by M operation. Note that the divisor M is not necessarily a power of 2.

For arbitrary symbol rates, the main constraint in the definition of a mini-slot, is that the number of symbols per mini-slot be an integer. For example given a symbol rate of R Symbols/ tick, and M ticks / mini-slot, the number of symbols per mini-slot N, is given by N = M*R. In this situation, M should be chosen such that N is an integer. In order to accommodate a wide range of symbol rates, it is important not to constrain M to be a power of 2. Since the additional BS resolution is independent of the symbol rate, the system can use a uniform time reference for distance ranging.

In order to demonstrate that the time base is applicable to single carrier and OFDM symbol rates, consider the following examples: (a) Single Carrier System - Given a symbol rate of 4.8 Msymbols/s (on a 6MHz channel), if the mini-slot duration is chosen to be 10 ticks (i.e., M = 10), then there are 48 symbols/mini-slot. Given 16QAM modulation this corresponds to a granularity of 24 bytes/mini-slot (b) OFDM System - Given an OFDM symbol time of 50 us, the mini-slot duration is also chosen to be 50 ticks (i.e., M = 50). In this case there is only a single symbol per mini-slot.

3.2.3.1.4.8 UL-MAP Ack Time definition

The Ack Time is the latest time processed in uplink in units of mini-slots. This time is used by the SS for collision detection purposes. The Ack Time is given relative to the BS initialization time.

3.2.4 MIMO Systems and Application of Beamforming Antenna Technology

In this Subsection, we will cover application of multiple input / multiple output (MIMO) and beamforming antenna technology.

3.2.4.1 Introduction

FWA system have a key requirement to operate in channels with large delay spreads and to provide a means of operation in line of sight, near line of sight (edge diffraction), an non-line of sight RF propagation channels.

Propagation loss will affect the energy level of the signal and ultimately the modulation complexity that can be supported. Multi-path and the resulting delay spread can result in distortions that make the signal impossible to demodulate regardless of received energy level unless some method to combat the multi-path is implemented. These methods include:

- Signal processing to perform channel equalization
- Directional antennas (limit sources of multi-path)
- Spatial diversity receivers (demodulation and coherent combining of one or more antenna/receiver sources)

In 1999 and 2000 two important paper provided detailed studies of the delay spread in 2 GHz and 2.5 GHz channels across a number of different line of sight (LOS) and non-line of sight (NLOS) channels.

Porter and Thweat provided a study of MMDS frequency propagation in a suburban environment [37]. Their results noted that a combination of directional transmit and receive antennas provided for RMS Delay Spread of less that 1 usec in 90% of the link cases. Also lower antenna heights resulted in lower delay spread but also greater propagation loss due to non-line-of-sight conditions. A summary of the test results is provided in the following table.

27 Table 3.5 - A Summary of the Test results for two Antenna Types

Visibility	Antenna type	RMS delay spread Min (usec)	RMS delay spread Max (usec)	RMS delay spread Mean (usec)
Line of Sight	directional	0.02	0.04	0.02
Line of Sight	omni	0.02	2.39	0.13
Non-Line of Sight	directional	0.02	5.26	0.14
Non-Line of Sight	omni	0.02	7.06	0.37

Erceg, Michelson, et. al. provided a similar study at 2 GHz [5]. As with the previously noted study, delay spread (full time span, not RMS delay spread) of up to 1 usec was detected for both omni and directional antennas.

More importantly, the use of diversity (multiple input) based on one or a combination of spatially separated antennas, polarization, and frequency/coding is considered a standard method of improving link fading performance. Jakes noted in his 1971 IEEE paper [38] that diversity improvement for 2 branches can provide nearly 20 dB of fade improvement while 3 branches can provide nearly 27 dB of improvement.

The use of diversity/MI (Multiple Input) techniques seek to improve signal performance in near/non-line of sight by combining the received energy of multiple diversity branches to reconstruct the receive signal.

In conjunction with Multiple Input technology, Multiple Output technology (e.g. Alamouti antenna diversity algorithm described in the previous section) seeks to create additional "artificial" diversity branch energy in the received signal.

While MIMO technologies improve link performance they do not reduce the C/I levels between cells or increase the frequency reuse factor in a wide scale system deployment. As systems are rolled out and subscriber densities within deployed cells increases, it is expected that advanced beam forming will be applied at the cell Base Ststion. Beamforming antennas will provide spatial reuse factors that will typically result in 2x to 4x increase in frequency efficiency.

The application of MIMO/beamforming technology must take into consideration the following requirements

- Compatibility with existing installed IEEE 802.16.3 subscribers meeting defined minimum requirements (MIMO/beamforming upgrade at a Base Station supports the current installed base)
- Consistent with industry requirements to reduce the cost of subscriber equipment. Add complexity at the Base Station and reduce complexity at the subscriber terminal
- Conform to all Regulatory requirements for EIRP, spurious emissions, and antenna beam restrictions (side lobe and front to back requirements)

The following sections will discuss the application of MIMO and beamforming. Mutiple input processing will be covered separately from mutiple output.

Note: These techniques can be applied to both FDD and TDD system. When specific processing for a type of duplexing method is required, this will be noted in the text

3.2.4.2 Antenna Regulatory Requirements

The TM4 group of ETSI has adopted a set of antenna beam shape requirements to limit C/I between cells and to improve frequency reuse. The requirements specify:

- Antenna beam roll off
- Antenna beam minimum side-lobe attenuation
- Antenna minimum front to back attenuation requirements

All IEEE 80216.3 antenna sub-systems must comply with these requirements [10].

Note that when Multiple Output techniques are applied, the additional transmissions cannot exceed the EIRP energy density (dB/Hz) limits imposed on the bandwidth of operation. That is, the sum of sources will be viewed by regulatory agencies responsible for Radio "type acceptance" as a composite single source.

3.2.4.3 Multiple Input (MI) Systems

Diversity (MI) processing is the most powerful way to minimize multi-path fading. Spatial techniques are applicable to any choice of modulation and represent a powerful enhancement that can improve the performance of any wireless access system. These techniques include:

- Directional antennas
- Multiple antennas with selection/scanning diversity
    - Spatial separation
    - Polarization separation
- Multiple antenna/receivers with signal combining.

As noted in the previous section, directional antennas at the subscriber greatly reduce multi-path and the resulting delay spread. The directional antenna acts as a spatial filter. Signals that are in the main beam (main-lobe) of the antenna are passed to the receiver while signals that reach the subscriber from the side-lobes are reduced by typically 25 to 40 dB. There are two other derivative benefits of directional antennas:

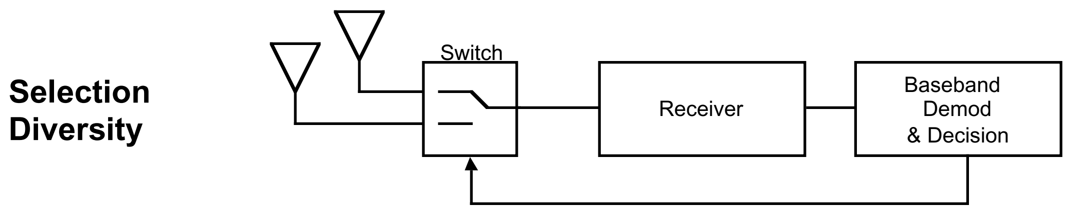
- Increased transmission gain and reduced cost and complexity of transmit PA (subscriber and basestation)
- Reduced interference from subscribers in adjacent cells. Increase in the overall capacity of a multiple cell system deployment.

Diversity is the use of separate distinct signal sources to enhance the received error rate and/or throughput of the system. Frequency diversity (use of 2 or more transmission paths to send the same data) has been used in microwave radio for decades. Likewise polarization diversity (use of horizontal and vertical antenna patterns) has been used in microwave radio and in 2nd generation cellular system Base Stations to enhance mobile subscriber reception. Time diversity (transmitting the same information at different times) has been used to mitigate periodic burst interference in military systems. These techniques can enhance reception by as much as 20 to 30 dB at the expense of reducing system capacity.

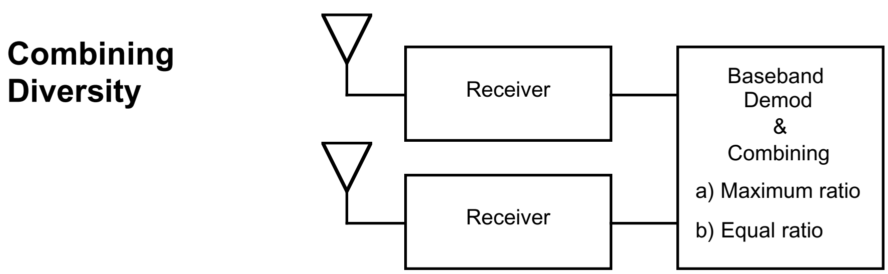
It is possible to have significant gains at the receiver by processing two or more receive paths using spatial or polarization diversity receivers. The implementation cost clearly increases with the number of paths/receivers that are processed. As a practical matter the use of a second receive source can improve resistance to fade and multi-path by as much as 20 dB. The addition of a 3rd receive source provides only an additional 6 to 7 dB under optimal conditions.

The simplest form of diversity is to sample the received SNR (lowest error rate) of one or more antennas and select the best source. The rate of scanning and selection must be performed at a rate much higher than the fade rate of the system. Selection diversity is shown Figure 3.21. This requires only a single receiver and represents one of the most cost effective methods to implement receive diversity. This technique maps well to TDD duplexing where the start of frame header can be used to perform diversity selection. The performance of selection diversity is inferior (by 3 to 6 dB) to combining of multiple receivers.

Combining Diversity is shown in Figure 3.22. An active receiver is required for each antenna. The signals of each receiver are combined (co-phased and summed) using either a maximum ratio criteria or by equal gain. The maximum ratio method weights signals based on their measured SNR (general S+N/N, i.e. signal & noise to noise ratio) and provides an output SNR that is the SUM of the input SNR (a gain in SNR). The benefit of maximum ratio combining is that the procedure can result in producing an acceptable output SNR even when the individual channels have marginal SNR.



28 Figure 3.21: Selection Diversity Receiver



29 **Figure 3.22: Combining Diversity Receiver**

Selection diversity can readily be applied to both TDD and FDD systems. TDD systems have the added benefit that the selected diversity branch can also be used to improve transmission.

Receiver diversity is not a requirement to conform with IEEE 802.16.3 FRD [1]. Given the inherent benefits, it is expected that Base Station will generally be deployed with 2 branches of diversity.

3.2.4.4 Multiple Output Systems

As previously discussed, the motivation for multiple output (MO) is to create and exploit a self-generated diversity branch in the processing of the system.

Given that the minimum requirements for Single Input Single Output (SISO) subscriber, the MO technique selected must be compatible with the minimum standards processing. It is anticipated that MO transmissions will, at the Base Station and that the subscriber, to minimize costs, be of either a SISO or MISO configuration.

At the current time two type of Multiple Output techniques are being investigated:

- The Alamouti antenna diversity algorithm
- Subscriber equalization delay diversity algorithm.

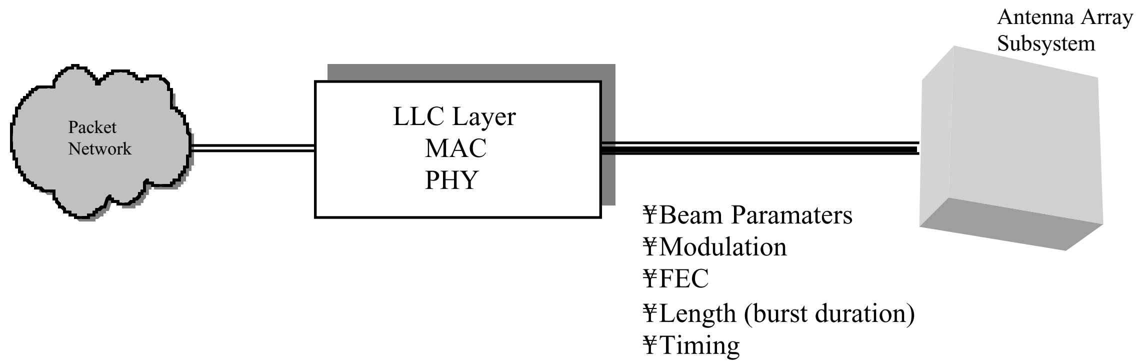
(Base Station would transmit the signal and a delayed replica at a delay that exceeds the channel delay spread Thus the equalizer would effectively act as a time domain combiner).

3.2.4.5 Application of Beam Forming Antenna Technology

The use of advanced antenna technology introduces an additional level of Media Access Control (MAC) complexity. The MAC/PHY has an added spatial/ beam component that must be factored into MAC coordination of the PHY. On a subscriber by subscriber (link by link) basis the MAC/PHY must coordinate the following parameters:

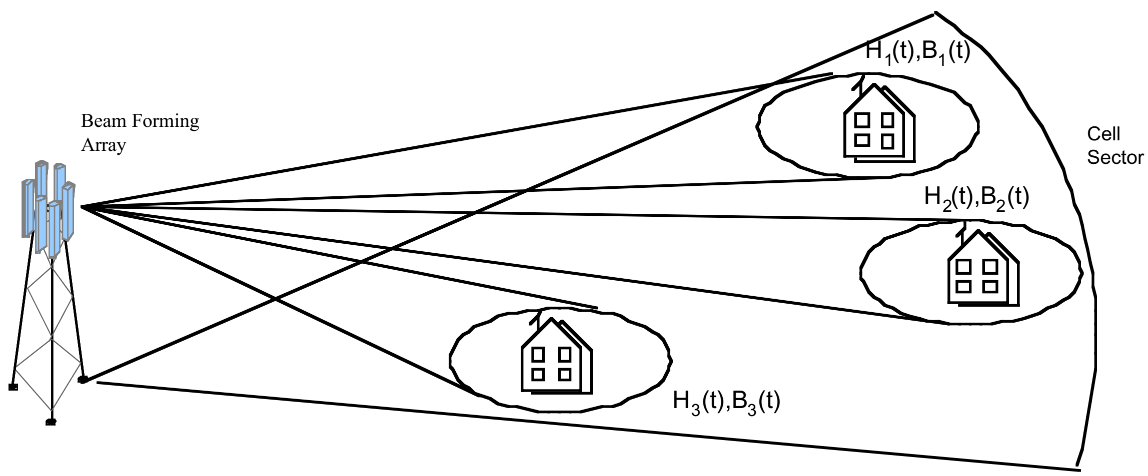
- Communications burst duration
  - Individual uplink or downlink for TDD
  - Joint up/down link for FDD
- Modulation Complexity
- FEC Rate
- Beam/Combining parameters.

The following figure illustrates the concept of coordinating MAC/PHY with the beam forming antenna element. While this standard does not attempt to define the specific technology or implementation of the beam forming technology, the design of the MAC and PHY must take into account that the beam forming subsystem places distinct restrictions on MAC/ PHY management and the coordination and passing of parameters necessary to support advanced beam forming.



30 Figure 3.23: The concept of Coordinating MAC/ PHY with the Beam Forming Antenna.

Beam forming and advanced antennas remove the basic paradigm that all subscribers have the capability of simultaneously receiving broadcast information from the Base Station. Beams are formed to optimize communications to a given subscriber with a channel response $H_n(t)$ and beam parameters $B_n(t)$. The following figure illustrates a sector of a base station that is communicating with 3 separate subscribers. Each subscriber is spatially distinct from the other subscribers. The transmission bursts sent to or from subscriber #1 would not be received by subscribers #2 or #3.

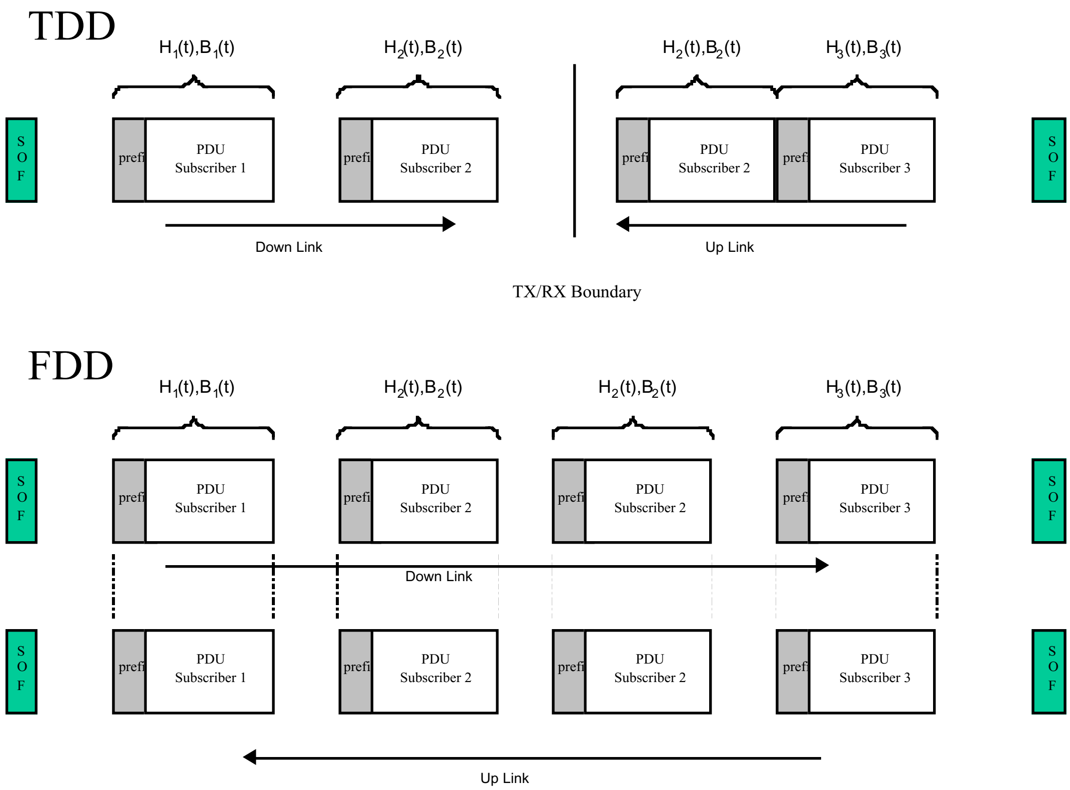


31 Figure 3.24: A Sector of a Base Station Communication with 3 Separate Subscribers.

In the described scenario, the Base Station is sequentially forming the beam and either sending or receiving from the subscribers in an order determined by the MAC.

To support advanced antenna systems both FDD and TDD links must be designed to provide transmissions based on self-contained bursts.

The following diagram illustrates the beam forming burst concept for both FDD and TDD.

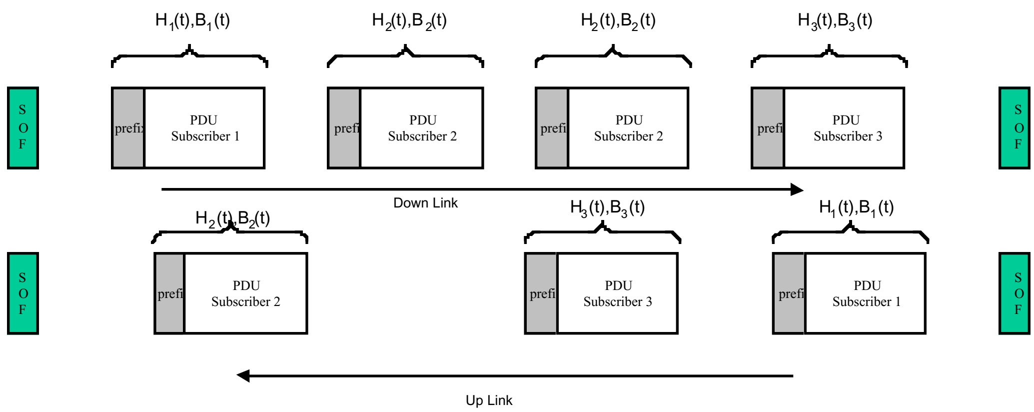


32 Figure 3.25: Beam forming Concept for TDD and FDD Cases.

Conceptually, TDD is easy to understand. A beam is formed for each transmitted burst in either the upstream or the down stream. The FDD solution can work one of two ways:

- Single beam forming for the Up/ Down Link
- Independent Up and Down Link beamforming. The system can support 2 independent formations of the beam on the up link frequency and the down link frequency (below)

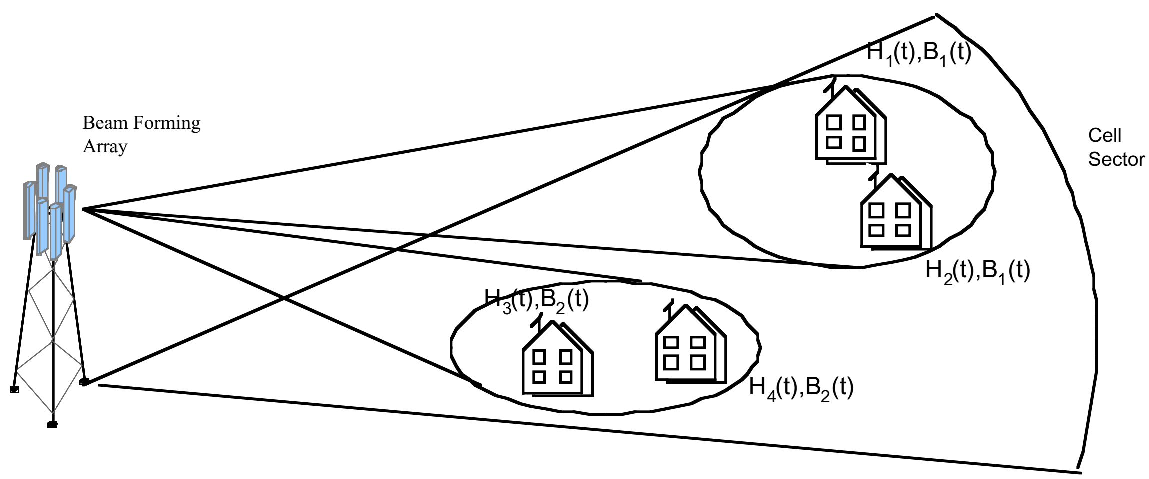
FDD with Independent beam forming



33 **Figure 3.26: An Example of FDD with Independent Up / Down Link Beam Forming**

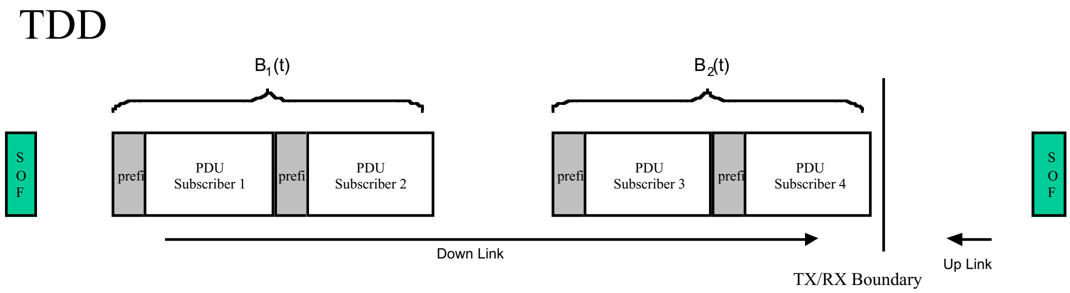
These simple sequential cases can be expanded to advanced beam forming techniques to provide simultaneous multiple access to spatially independent users. A beam-forming network can create 2 or more independent beams with low self-interference that allow simultaneous communications using the same frequency. While beam-forming complexity is increased, spectral reuse is also increased. The complexity of PHY hardware and MAC scheduling software also increase proportionally with the number of beams created.

The MAC and PHY also need to perform burst scheduling and transmission based on "spatial concatenation". One or more subscribers can be supported by a single set of beam-forming parameters due to close physical proximity as shown in the following figure. For this case, bursts to the subscribers that share the same beams.



34 **Figure 3.27: A Sectored Base Station Communicating with 4 Separate Subscribers in a Spatial Concatenated manner.**

The following figure illustrates how packets are grouped (concatenated) and transmitted based on physical proximity for a TDD Physical layer.



35 **Figure 3.28: An Example of TDD with Concatenated Packet transmissions.**

The PHY based on block processing and burst packet formats meets all the requirements to support advanced antenna processing techniques. As the standard progresses we must address the following issues in greater detail:

- Beam forming Transition/ Set-up time definition in the MAC (passing parameters to PHY)
- Method for broadcasting Uplink and Downlink MAP information
- Acquisition methods and beam scanning
- Cell to Cell interference and C/I issues
- Spatial multiplexing.

3.3 Single-Carrier with Frequency Domain Equalization (SC-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 [17], [18], [19] and [28].

Note that with an adaptive receiver based on Frequency Domain processing can handle both OFDM and Single Carrier modulation.

Further note that, Hikmet Sari in References [16 and 28 to 31] has significantly contributed to the development of Single Carrier modulation with Frequency Domain Equalization (SC-FDE). He also introduced the concept of Cyclic prefix to simplify the processing. However, there were few others (in the 70 s and 80 s) in the field who have introduced the concept of Frequency Domain Equalizer with overlap-add methods which can eliminate the need for a cyclic prefix but introduces added complexity in processing and adaptation. We should also mention that Sari [16] was the first to compare SC-FDE explicitly with OFDM.

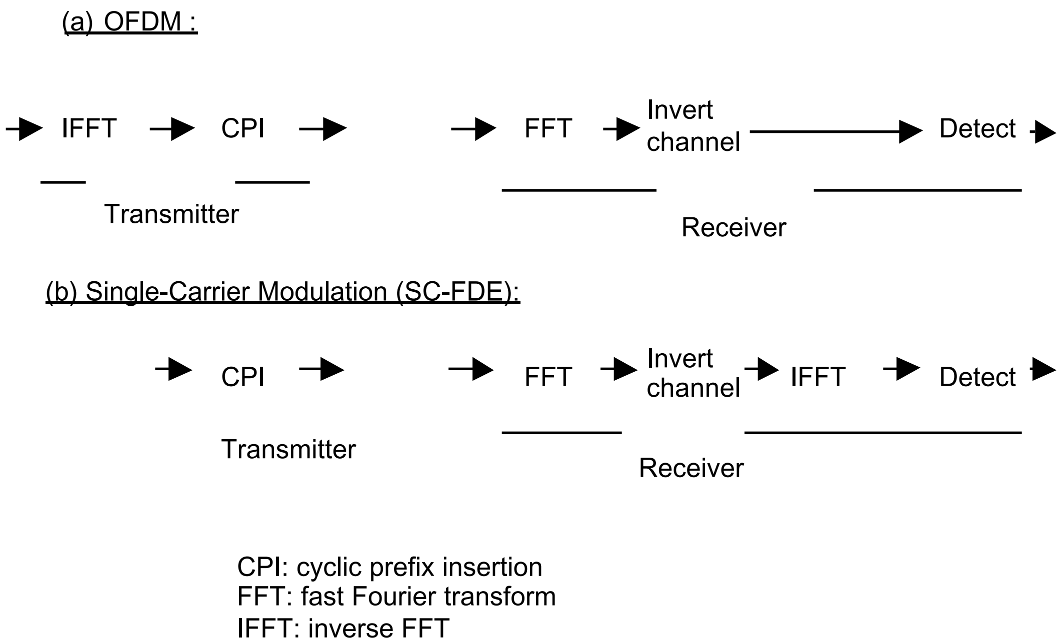
3.3.1 Single Carrier-Frequency Domain Equalization (SC-FDE)

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 3.29.)

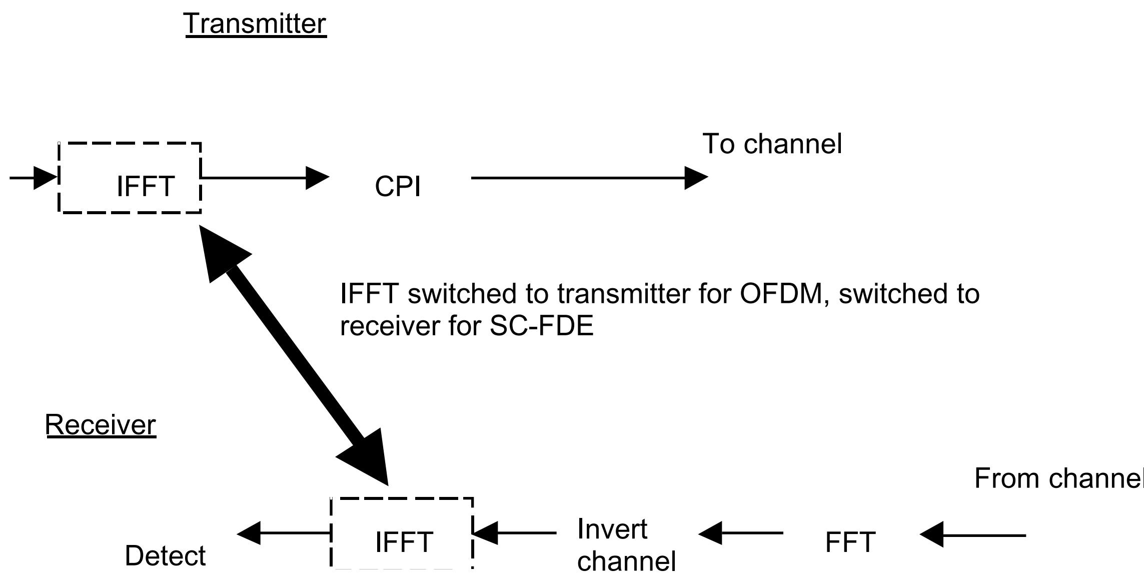
Both systems can be enhanced by coding (which is in fact required for OFDM systems), adaptive modulation and space diversity. In addition, OFDM can incorporate peak-to-average reduction signal processing to partially (but not completely) alleviate its high sensitivity to power amplifier nonlinearities.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



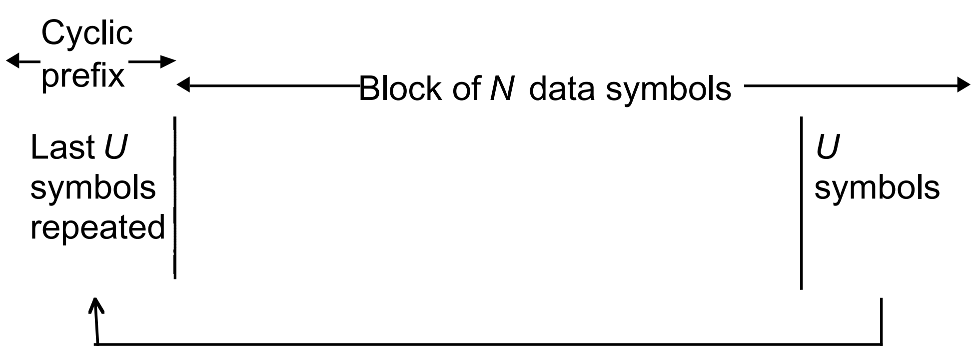
36 Figure 3.29- OFDM and Single Carrier-Frequency Domain Equalization (SC-FDE).

3.3.2 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 3.30).



37 **Figure 3.30 - OFDM and SC-FDE Convertible Modem Approach.**



38 **Figure 3.31 - Block Processing in Frequency Domain Equalization**

As shown in Figure 3.31, 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 U and last U symbols are identical unique word sequences of training symbols, the overhead fraction is $2U / (N+U)$.

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

3.4 The frequency range and the channel bandwidth

The frequency range and the downstream and upstream channel bandwidth of the PHY system are given in Table 3.6.

39 Table 3.6: Frequency Bands and Channel Bandwidth

Frequency Bands	Channel Bandwidth Options	Reference
a) 2.15- 2.162 GHz, 2.50- 2.690 GHz	2 to 6 MHz downstream, 200 kHz to 6 MHz upstream	FCC 47 CFR 21.901 (MDS) FCC 47 CFR 74.902 (ITFS, MMDS) Industry Canada SRSP-302.5 (Fixed Services operating in the 2500 to 2686 MHz band)
b) 3.5 GHz	1.75- 7 MHz downstream, 250 KHz to 7 MHz upstream	EN 301 021, CEPT/ERC Rec. 14-03 E, CEPT/ERC Rec. 12-08 E, Others (TBD)
c) 5.25 – 5.35 GHz, 5.725 – 5.835 GHz	Total 100 MHz downstream	FCC UNII
d) 10.5 GHz	3.5, 5 and 7 MHz	EN 301 021, CEPT/ERC Rec. 12-05 E

3.5 Duplex Schemes

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

3.5.1 TDD:

In **Time division duplex** (TDD) systems, the radio frame is divided into a downlink and an uplink section, offering flexible and 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.

3.5.2 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.

3.6 Downstream Channel

3.6.1 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 or 64 QAM) to provide a digital bit stream (e.g., 10 to 30 Mbps). Within each RF channel a frame structure is used to organize and schedule the transmission of voice, video and data traffic.

3.6.2 Modulation Schemes

The applicable modulation schemes for the downstream are QPSK, 16 QAM or 64 QAM (256 QAM downstream modulation can be an option.)

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. Complete description of the MAC / PHY support of adaptive modulation and coding is provided in Section 3.2.

3.6.3 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.

3.6.3.1 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.

3.6.3.2 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 _ Convolutional code with constraint length K=7. The Convolutional code is able to be configured to code rates 2/3, _, 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.

3.6.3.3 Symbol Mapping

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

3.6.3.4 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):

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

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

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

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

Where:

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

3.7 Upstream Channel

3.7.1 Upstream Multiple Access

The upstream multiple access method shall be TDMA.

The upstream channel bandwidth for MMDS channel allocation would be 6MHz. In TDD mode this 6MHz bandwidth can be dynamically allocated. This TDD flexibility permits efficient allocation of the available bandwidth and hence is capable of efficiently allocating the available traffic transport capacity for applications whose upstream to downstream traffic transport demand ratio can vary with time. FDD can be used by applications that require fixed asymmetric allocation between their upstream and downstream traffic transport demand. In FDD mode upstream traffic would typically be allocated 3MHz. This is half of the 6MHz bandwidth assuming symmetrical traffic requirements.

3.7.2 Upstream Modulation Format

The upstream modulation types can be the same as those available for downstream transmission; e.g., QPSK, 16QAM, or 64QAM. Modulation type, error correction, interleaving, etc, can be assigned to the upstream traffic for a particular Subscriber Station (SS) such that these parameters can be the same as in the downstream burst received by the SS. The SS accesses the 'quality' of the received signal from the downstream header, and the SS MAC decides on the best modulation and error correction to use for the channel conditions. This information is passed back up to the Access Point (AP) in the corresponding upstream burst, and the AP MAC uses this information to assign the modulation type and error correction to the next burst of data to be transmitted.

3.7.3 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.

3.7.3.1 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).

3.7.3.2 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.

1.1.1.3 Interleaving for the upstream channel

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

3.7.3.4 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):

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

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

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

Where:

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

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

3.8 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 MHz 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 SC-FDE System Capacity and Modulation Efficiency

4.1 System Capacity:

Table 4.1 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 _ = %0.15 or %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 4.1 presents an example of achievable system capacity where all coding and FDE overhead budget is being included.

40 Table 4.1: An Example of System Capacity Objectives.

Channel Spacing	Downstream Transmission Rate (Mb/s)		Upstream Transmission 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

4.2 SC-FDE System Throughput

For single-carrier systems, system throughput will vary with the operating modes. With frame structure given in Subsection 3.2, the SC-FDE system throughput is given as:

$$T = R\frac{N-U}{N} r \log_2 M .$$

If the design with U/2 = R • d, rounded up to the nearest power of 2, the throughput for SC-FDE system will then equal to:

$$T = R\frac{N-\lceil R \bullet d \rceil_2}{N} r \log_2 M ,$$

where $\lceil \bullet \rceil_2$ denotes rounding up to the nearest power of 2.

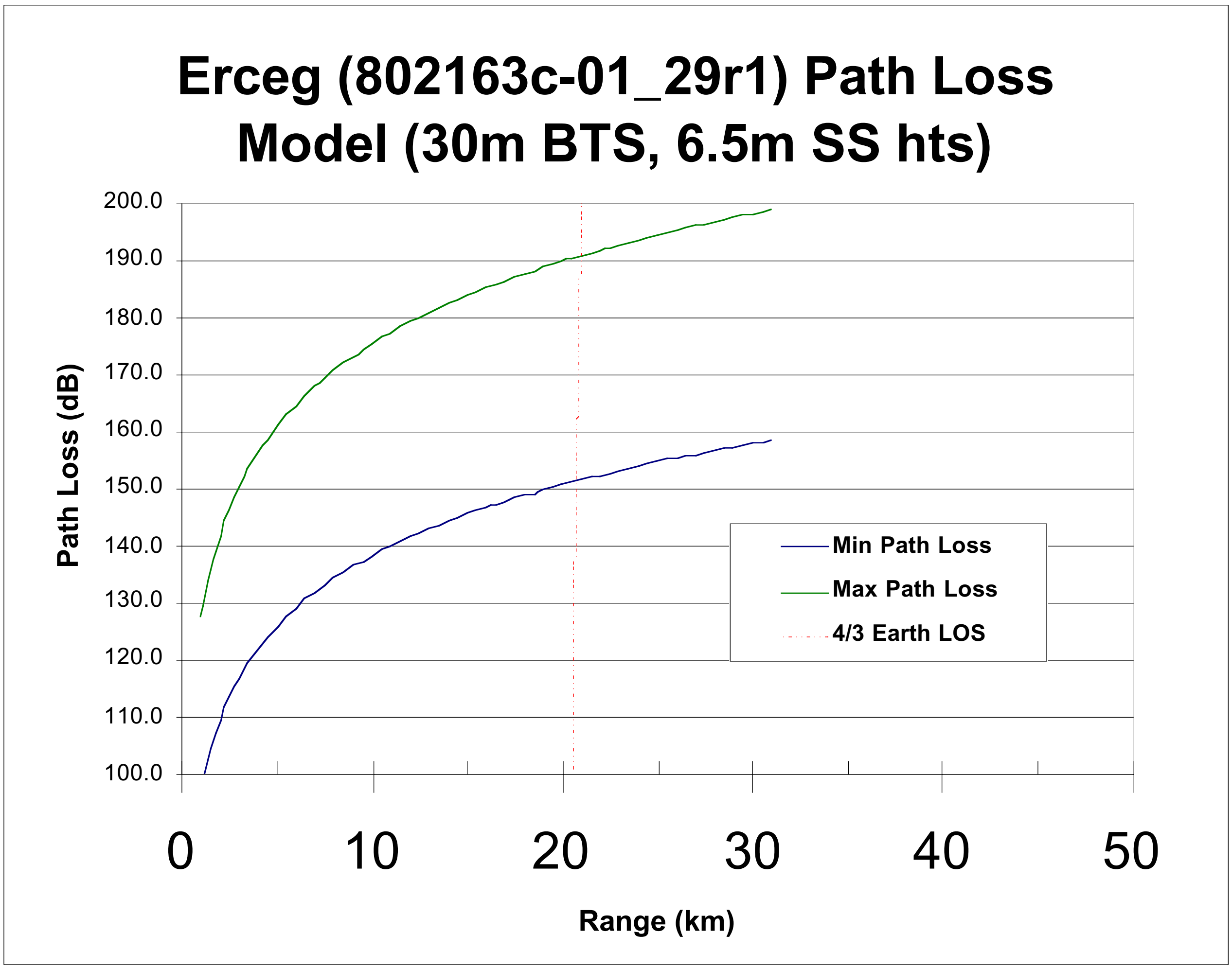
Next, Table 4.2 presents typical channel throughput for SC-DFE system with a 1.75 MHz channel Bandwidth. Similar typical results for higher channel bandwidths will be proportionally larger.

Table 4.2: Throughput for Various Models in 1.75 MHz Channel

System-Dependent Parameters		Link-Dependent Parameters		Traffic-Dependent Parameter			
Symbol [Sample] Rate (MS/sec)	Design Max Delay Spread (microsec)	Number of QAM States	Convolutional Code Rate	FFT Size 256	512	1024	2048
			1/2	1.453	1.477	1.488	1.494
		4	2/3	1.938	1.969	1.984	1.992
			3/4	2.180	2.215	2.232	2.241
			7/8	2.543	2.584	2.604	2.615
			1/2	2.906	2.953	2.977	2.988
	4	16	2/3	3.875	3.938	3.969	3.984
			3/4	4.359	4.430	4.465	4.482
			7/8	5.086	5.168	5.209	5.229
			1/2	4.359	4.430	4.465	4.482
		64	2/3	5.813	5.906	5.953	5.977
			3/4	6.539	6.645	6.697	6.724
			7/8	7.629	7.752	7.813	7.844
			1/2	1.395	1.447	1.474	1.487
		4	2/3	1.859	1.930	1.965	1.982
			3/4	2.092	2.171	2.210	2.230
			7/8	2.440	2.533	2.579	2.602
			1/2	2.789	2.895	2.947	2.974
1.5	10	16	2/3	3.719	3.859	3.930	3.965
			3/4	4.184	4.342	4.421	4.460
			7/8	4.881	5.065	5.158	5.204
			1/2	4.184	4.342	4.421	4.460
		64	2/3	5.578	5.789	5.895	5.947
			3/4	6.275	6.513	6.631	6.691
			7/8	7.321	7.598	7.737	7.806
			1/2	1.313	1.406	1.453	1.477
		4	2/3	1.750	1.875	1.938	1.969
			3/4	1.969	2.109	2.180	2.215
			7/8	2.297	2.461	2.543	2.584
			1/2	2.625	2.813	2.906	2.953
	20	16	2/3	3.500	3.750	3.875	3.938
			3/4	3.938	4.219	4.359	4.430
			7/8	4.594	4.922	5.086	5.168
			1/2	3.938	4.219	4.359	4.430
		64	2/3	5.250	5.625	5.813	5.906
			3/4	5.906	6.328	6.539	6.645
			7/8	6.891	7.383	7.629	7.752

5 LINK Budget Analysis

We have made a complete Link budget analysis for the various combinations of modulation format and channel bandwidth that were specified by Erceg s latest version of channel model for this proposal. The path loss given below was calculated using the median value for Condition C of the model in Erceg s latest version of the path model (802.16.3c-29r1). For each Downstream (D/S) and Upstream (U/S) pair we have calculated the maximum path length that could be supported given the 43 dBm EIRP at the BTS and a 40 dBm EIRP at the SS with typical values for SNR at the receiver for each modulation format. Some typical results are presented in Table 5.1.



41 **Figure 5.1: Path Loss Model (Condition C of the Erceg s 802.16.3c-29r1 Contribution.**

Table 5.1 present s channel model as per Erceg s contribution 802.16.3c-29r1. The selected channel is a typical MMDC channel at 2.5 GHz band.

Table 5.1: Channel Model Section as per Erceg s Contribution 802.16.3c-29r1

	Category		
	C	B	A
Parameter	**Flat, few trees**	**Inter mediate**	**Hilly, heavy trees**
a	3.6	4	4.6
b	0.005	0.0065	0.0075
c	20	17.1	12.6
Channel frequency	2.5	GHz	
Wavelength	0.12	m	
receive antenna height h=	6.5	m	
(hb is the height of the base station in m) hb=	80	m	
_ =(a —b hb +c /hb) _ =	4.116667	4.375	4.795
A =20 log10 (4 „ d0 /_)(_ being the wavelength in m)	80.40057		
s=	9.4		
PL =A + 10 _ log10 (d/d0) + DPl + DPh_+ s for d >d0,			
4/3 Earth Line of Sight =	32.5	km	

Based on the parameter selection in Table 5.1, we have generated link budget for various scenarios. Some typical results are for QPSK and 64 QAM that are presented in the following Tables 5.2 and 5.3, respectively. These results assume very similar scenarios for SC-FDE and OFDM systems.

Table 5.2: Typical Link Budget results for Single Carrier and OFDM for QPSK (1.5 and 6 MHz width)

	Single Carrier		512 Carriers		Single Carrier		512 Carriers	
Bandwidth	1.5 MHz		1.5 MHz		6.0 MHz		6 MHz	
Modulation type / Target SNR	QPSK	10 dB	OFDM	10 dB	QPSK	10 dB	OFDM	10 dB
Downstream								
EIRP (BTS)	43.0 dBm	20 w	43.0 dBm	20 w	43.0 dBm	20 w	43.0 dBm	20 w
Antenna Gain	3.0 dB		3.0 dB		3.0 dB		3.0 dB	
Back off	12.0 dB		14.0 dB		11.0 dB		14.0 dB	
Nominal 1 dB compression point	52.0 dBm	**158 w**	54.0 dBm	**251 w**	51.0 dBm	**126 w**	54.0 dBm	**251 w**
Normalized Price	**1.0**		**1.3**		**1.0**		**1.3**	
Path distance for targeted SNR	**8.0 km**		**8.0 km**		**8.0 km**		**8.0 km**	
Associated Path Loss (from 802.16.3c-29r1)	-153.8 dB		-153.8 dB		-153.8 dB		-153.8 dB	
Receive Antenna gain	14.0 dB		14.0 dB		14.0 dB		14.0 dB	
Power at Input to Receiver	-96.8 dBm		-96.8 dBm		-96.8 dBm		-96.8 dBm	
Receiver Noise Figure	5.0 dB		5.0 dB		5.0 dB		5.0 dB	
Equivalent Noise Power in channel BW	-107.2 dBm		-107.2 dBm		-101.2 dBm		-101.2 dBm	
SNR, Calculated	**10.4 dB**		**10.4 dB**		**4.4 dB**		**4.4 dB**	
Upstream								
EIRP (SS)	34.0 dBm	3 w	34.0 dBm	3 w	40.0 dBm	10 w	40.0 dBm	10 w
Antenna Gain	14.0 dB		14.0 dB		14.0 dB		14.0 dB	
Back off	6.0 dB		14.0 dB		11.0 dB		14.0 dB	
Nominal 1 dB compression point	26.0 dBm	**0.40 w**	34.0 dBm	**3 w**	37.0 dBm	**5 w**	40.0 dBm	**10 w**
Normalized Price	**1.0**		**4.0**		**1.0**		**4.0**	
Path distance for targeted SNR	**3.0 km**		**3.0 km**		**3.0 km**		**3.0 km**	
Associated Path Loss (from 802.16.3c-29)	-136.3 dB		-136.3 dB		-136.3 dB		-136.3 dB	
Receive Antenna gain	6.0 dB		6.0 dB		6.0 dB		6.0 dB	
Power at Input to Receiver	-96.3 dBm		-96.3 dBm		-90.3 dBm		-90.3 dBm	
Receiver Noise Figure	4.0 dB		4.0 dB		4.0 dB		4.0 dB	
Equivalent Noise Power in channel BW	-108.2 dBm		-108.2 dBm		-102.2 dBm		-102.2 dBm	
SNR, Calculated	**12.0 dB**		**12.0 dB**		**12.0 dB**		**12.0 dB**	

Table 5.3: Typical Link Budget results for Single Carrier and OFDM for 64QAM (1.5 and 6 MHz width)

	Single Carrier		512 Carriers		Single Carrier		512 Carriers	
Bandwidth	**1.5** MHz		**1.5** MHz		**6.0** MHz		**6** MHz	
Modulation type / Target SNR	**64 QAM**	25 dB	**OFDM**	25 dB	**64 QAM**	25 dB	**OFDM**	25 dB
Downstream								
EIRP (BTS)	43.0 dBm	20 w	43.0 dBm	20 w	43.0 dBm	20 w	43.0 dBm	20 w
Antenna Gain	3.0 dB		3.0 dB		3.0 dB		3.0 dB	
Back off	12.0 dB		14.0 dB		12.0 dB		14.0 dB	
Nominal 1 dB compression point	52.0 dBm	**158 w**	54.0 dBm	**251 w**	52.0 dBm	**158 w**	54.0 dBm	**251 w**
Normalized Price	**1.0**		**1.3**		**1.0**		**1.3**	
Path distance for targeted SNR	**6.5 km**		**6.5 km**		**4.5 km**		**4.5 km**	
Associated Path Loss (from 802.16.3c-29r1)	-150.1 dB		-150.1 dB		-143.5 dB		-143.5 dB	
Receive Antenna gain	14.0 dB		14.0 dB		14.0 dB		14.0 dB	
Power at Input to Receiver	-93.1 dBm		-93.1 dBm		-86.5 dBm		-86.5 dBm	
Receiver Noise Figure	5.0 dB		5.0 dB		5.0 dB		5.0 dB	
Equivalent Noise Power in channel BW	-107.2 dBm		-107.2 dBm		-101.2 dBm		-101.2 dBm	
SNR, Calculated	**14.2 dB**		**14.2 dB**		**14.7 dB**		**14.7 dB**	
Upstream								
EIRP (SS)	34.0 dBm	3 w	34.0 dBm	3 w	40.0 dBm	10 w	40.0 dBm	10 w
Antenna Gain	14.0 dB		14.0 dB		14.0 dB		14.0 dB	
Back off	6.0 dB		14.0 dB		6.0 dB		14.0 dB	
Nominal 1 dB compression point	26.0 dBm	**0.40 w**	34.0 dBm	**3 w**	32.0 dBm	**2 w**	40.0 dBm	**10 w**
Normalized Price	**1.0**		**4.0**		**1.0**		**4.0**	
Path distance for targeted SNR	**2.5 km**		**2.5 km**		**2.5 km**		**2.5 km**	
Associated Path Loss (from 802.16.3c-29)	-133.0 dB		-133.0 dB		-133.0 dB		-133.0 dB	
Receive Antenna gain	6.0 dB		6.0 dB		6.0 dB		6.0 dB	
Power at Input to Receiver	-93.0 dBm		-93.0 dBm		-87.0 dBm		-87.0 dBm	
Receiver Noise Figure	4.0 dB		4.0 dB		4.0 dB		4.0 dB	
Equivalent Noise Power in channel BW	-108.2 dBm		-108.2 dBm		-102.2 dBm		-102.2 dBm	
SNR, Calculated	**15.2 dB**		**15.2 dB**		**15.2 dB**		**15.2 dB**	

6 Minimum (Multipath) Performance

The introduction of Table 6.1 is to include minimum system parameters into the draft document. Entries in the following table are SNR s in dB to achieve a BER of 10^{-6}. **They are subject to change.** The numbers in this table are from static channel simulations, using SUI-2 and SUI-5 channels that were presented by Ariyavisitakul and Falconer in Reference [20]. There should be an implementation margin added, of perhaps several dB.

Note that the SNR numbers to achieve 10^{-6} average BER are tentative at this point - the simulations have a fair amount of variability at that level.

	QPSK				**16QAM**				**64QAM**				
	Rate 1/2	Rate 2/3	Rate 3/4	uncoded	Rate 1/2	Rate 2/3	Rate 3/4	uncoded	Rate 1/2	Rate 2/3	Rate 3/4	Rate 7/8	uncoded
Relative bit rate	5	6.67	7.5	10	10	13.3	15	20	15	20	22.5	26.25	30
SUI-2													
1 TX, 1 RX antenna	23	27	30	35	29			40			48		
SUI-5													
1 TX. 1 RX antenna	21	24	27	28	29	30	30	36	30	36	37	40	43

42 Table 6.1: Minimum Performance Specification of PHY Systems

7 Main Features and Benefits of the PHY Standard

This PHY standard for the IEEE802.16a air interface 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 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 efficiency.
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.

10) Important note that the expansion of the system bandwidth from 6 MHz to say 20 MHz, while encountering the same type of multipath channels, will result in multipath delay spreads that span 3 to 4 times as many data symbols. Frequency domain equalization (SC-FDE) is still viable, provided that the FFT block lengths, cyclic prefix lengths, and unique word lengths are increased by the corresponding factor of 3 to 4. The same considerations apply for OFDM systems. Thus for example the maximum FFT block length for a 20 MHz system might be on the order of 8,192.

8 Similarity to other standards:

The SC-FDE PHY is similar to some extent 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).

9 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.

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11 APPENDIX A: Channel Model For BWA PHY Systems

11.1 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 terrain 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.

11.2 RF Channel Models

11.2.1 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 λ 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.

11.2.2 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.

11.2.3 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:

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

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)

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

ΔPL_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**.

12 Appendix B: Block Turbo and Reed-Solomon Coding

12.1 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.

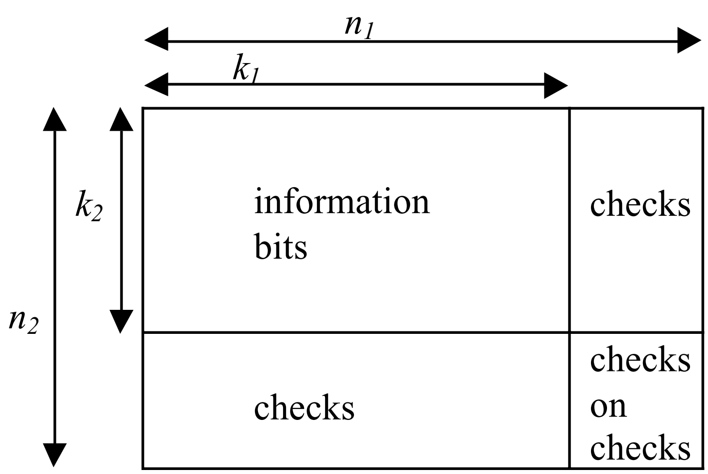


Figure C1 - Two-dimensional product code matrix

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	$x^6 + x + 1$

43 **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 $k_1 \times (n_2-k_2)$ 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}.

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}

44 **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}.

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}

45 **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 $(8,4)\times(8,4)$ 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:

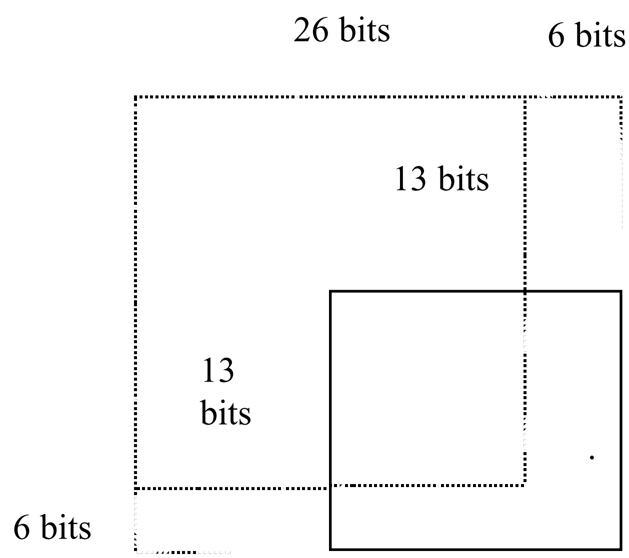
- the codes defined for the rows (x-axis) are binary (n_x, k_x) block codes
- the codes defined for the columns (y-axis) are binary (n_y, k_y) block codes
- data bits are noted $D_{y,x}$ and parity bits are noted $E_{y,x}$.

Shortened BTCs

To match packet sizes, a product code can be shortened by removing symbols from the array. In the two-dimensional case rows, 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 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.



46 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 constellation 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 stuffed 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 shortening 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 interleavers ~~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. QPSK/16QAM/64QAM modulation and interleave type 1 (no interleaver) was assumed.

47 **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
Channel Efficiency, QPSK/16QAM/64QAM	1.35/2.69/4.04 bit/symb/Hz	1.42/2.84/4.27 bit/symb/Hz	1.58/3.16/4.74 bit/symb/Hz
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

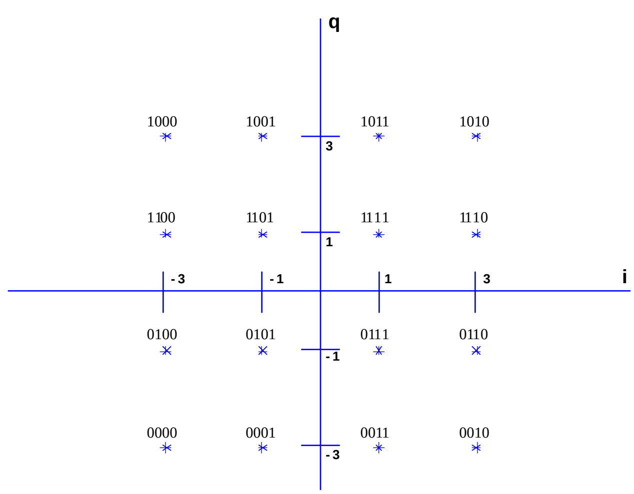
48 **TABLE C3: Typical performance for BTC with small blocks (upstream channel)**

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

Combining Turbo Product Codes With M-QAM

The downstream channel supports both continuous and burst mode operation and the FEC incorporates turbo product codes, for each of these applications. In order to provide the desired flexibility and the required QoS, TPCs are used in conjunction with adaptive modulation scheme where different modulation formats and TPC s are specified on a frame-by-frame basis. Mapping of the encoded bits into the desired modulation is arranged in Gray code format, as shown in Figure C5.

Fig.C5. Signal Mapping for 16QAM



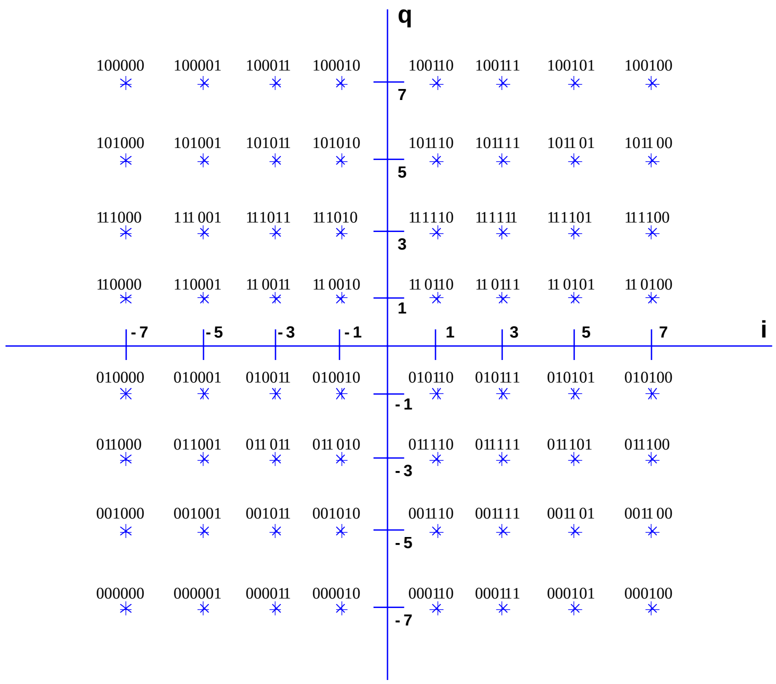
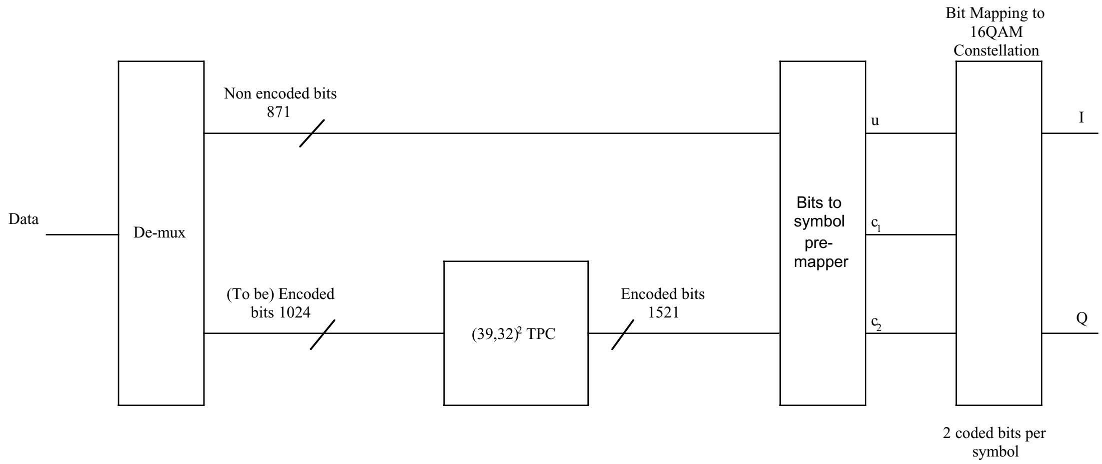


Fig.C5b. Signal Constellation for 64QAM

An additional optional mode of mapping, known as pragmatic turbo coded modulation may be implemented [13] at the subscriber station. This mode does not require any new turbo coding schemes, however, by modifying the implementation of the mapping table, an additional coding gain between 0.5dB to 0.75 dB for a range of coding rates and modulation techniques can be achieved. The following example illustrates the procedure.

Consider first that we have a $(63,56)^2$ TPC code described in Table C2. This has rate 0.79 and when combined with 16QAM in Gray code mapping fashion provides $0.79 \times \log_2 16 = 3.16$ bits/symbol/Hz of channel efficiency.

Now consider the $(39,32)^2$ 2-dimensional TPC from the same table. This code has shorter block length, its performance is not as good as the performance of the larger code, however it can be implemented with less complexity and reduced absolute latency (Note: that the latency of TPC codes is similar to the conventional RS codes, i.e. encoding latency is equal to one block size while the decoding latency for TPC is equal to 2 block lengths). When combined with 16QAM in a Gray code manner it provides 2.69 bits/symb/Hz. Figure C5c describes the basic scheme for the TPC in a TCM scheme.

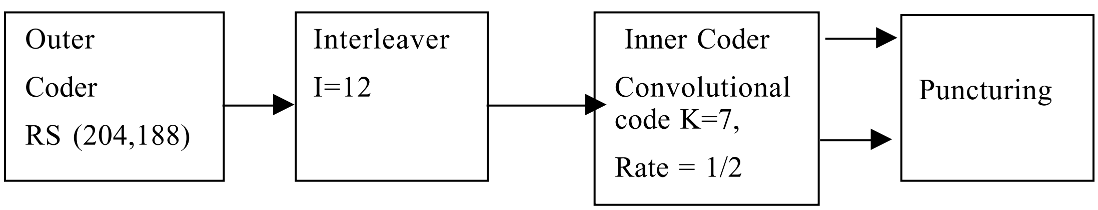


49 Figure C5c $(39,32)^2$ TPC TCM Encoder

Looking at the throughput of this system we have 1024+871=1895 bits entering the system and $(1521+871)/\log_2 16=600$ symbols emanating from the modulator. Thus the channel efficiency of this system will be 1895/600=3.16 bits/symb/Hz (similar to the channel efficiency of the coding scheme with larger block length), while both the encoder and decoder latency remains as for shorter $(39,32)^2$ codes. It is apparent that this scheme is applicable to all TPC codes described in the Section and the variation of number of uncoded bits will allow additional flexibility in system development and deployments.

12.2 Reed – Solomon Coding

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



50 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).

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

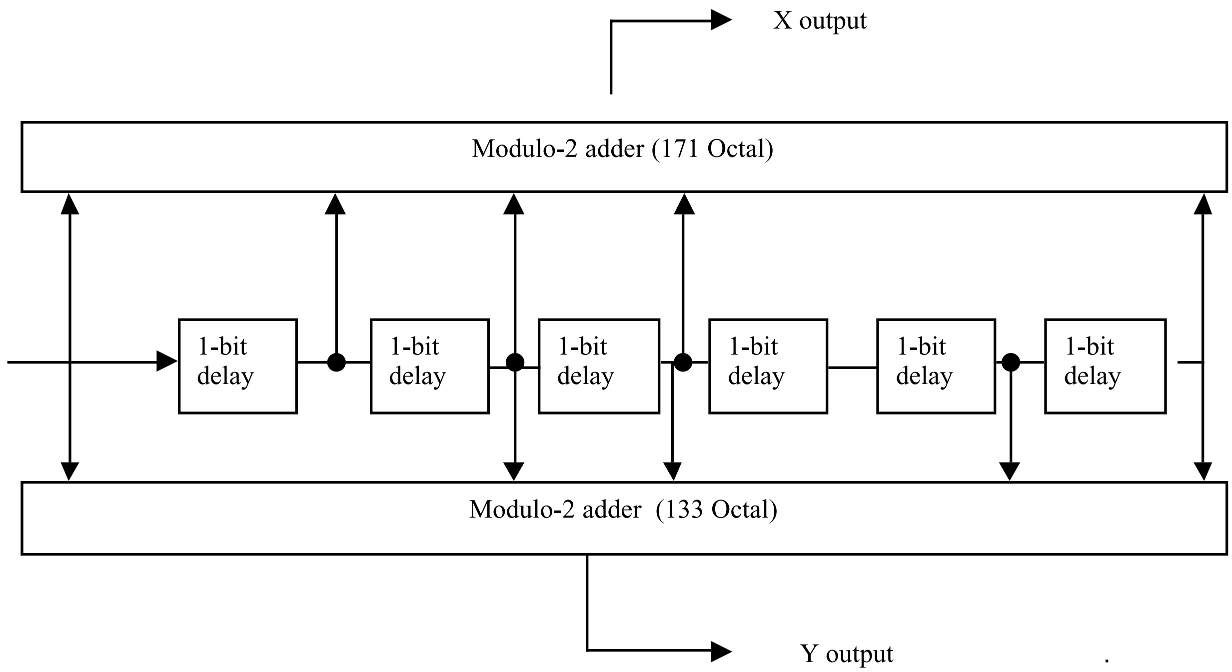
P(x) is the Field generator polynomial, and the code generator polynomial is:

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

Where: λ 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.



51 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.

52 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	171	133	X: 1 Y: 1 I=X_1 Q=Y_1	10	X: 10 Y: 11 I= $X_1Y_2Y_3$ Q= $Y_1X_3Y_4$	6	X:101 Y:110 I= X_1Y_2 Q= Y_1X_3	5	X°:10101 Y°:11010 I= $X_1Y_2Y_4$ Q= $Y_1X_3X_5$	4	X:1000101 Y:1111010 I= $X_1Y_2Y_4Y_6$ 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}).

53 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 _, 7/8, 16QAM, respectively.				