Project	**IEEE 802.16 Broadband Wireless Access Working Group <http://ieee802.org/16>**
Title	**Contributions and Comments for the 802.16.4 PHY Draft**
Date Submitted	**2001-02-24**
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Re:	Contributions to the 802.16.4 PHY draft as agreed during the interim meeting (session 11.5)
Abstract	The document contains the following proposed sections for the draft: • Convolutional encoder and puncturing • Interleaving • Dynamic adaptive modulation and coding • Transmit center frequency and symbol clock tolerance • Base station frame synchronization and coordination mechanism • Subcarrier based parallel polling • OFDM Preambles It also contains comments on: • the use of Reed-Solomon coding vs. ARQ with adaptive modulation • the different aspects of using channelization of 5, 10 or 20 MHz
Purpose	For review.
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Contributions and Comments for the 802.16.4 PHY Draft

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Contributions to the 802.16.4 PHY Draft

Note:

I used *italics* to mark comments, remarks and TBD issues that must not appear in the final document. I suggest keeping them until we finalize the document.

Convolutional encoder and puncturing

The serial data (*to add later where it comes from: scrambler or outer interleaving, to specify the padding*) shall be coded with a convolutional encoder of coding rate R=1/2, 2/3 or 3/4 corresponding to the desired data rate (*to decide later the data rates i.e. valid combinations of modulation and coding rates*). The convolutional encoder shall use the industry standard generator polynomials $g_0=133_8$ and $g_1=171_8$ of rate R=1/2 as shown in **Figure 1**. Bit denoted as "A" shall output from the encoder before the bit denoted as "B". Rates of 2/3 and 3/4 shall be derived by employing "puncturing", i.e. by omitting some of the encoded bits in the transmitter based on known patterns. The puncturing patterns are illustrated in **Figure 2** for R=3/4 and **Figure 3** for R=2/3. The rate R=1/2 is illustrated in **Figure 4**. Decoding by the Viterbi algorithm is recommended.

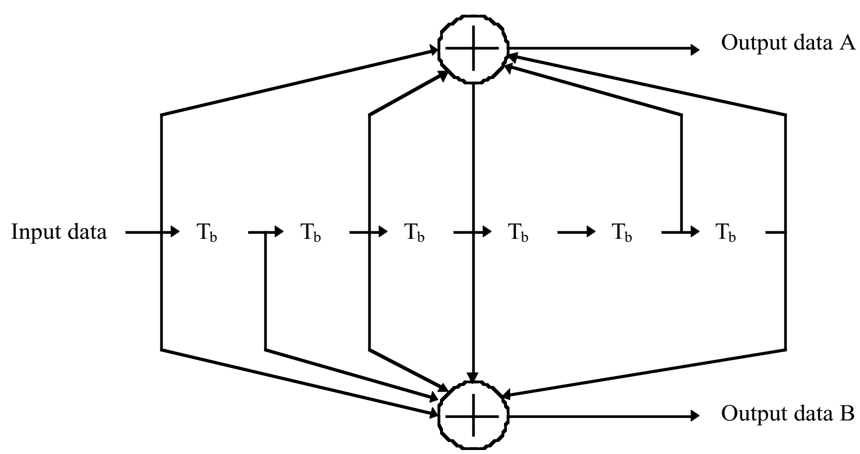

Figure 1: Convolutional encoder

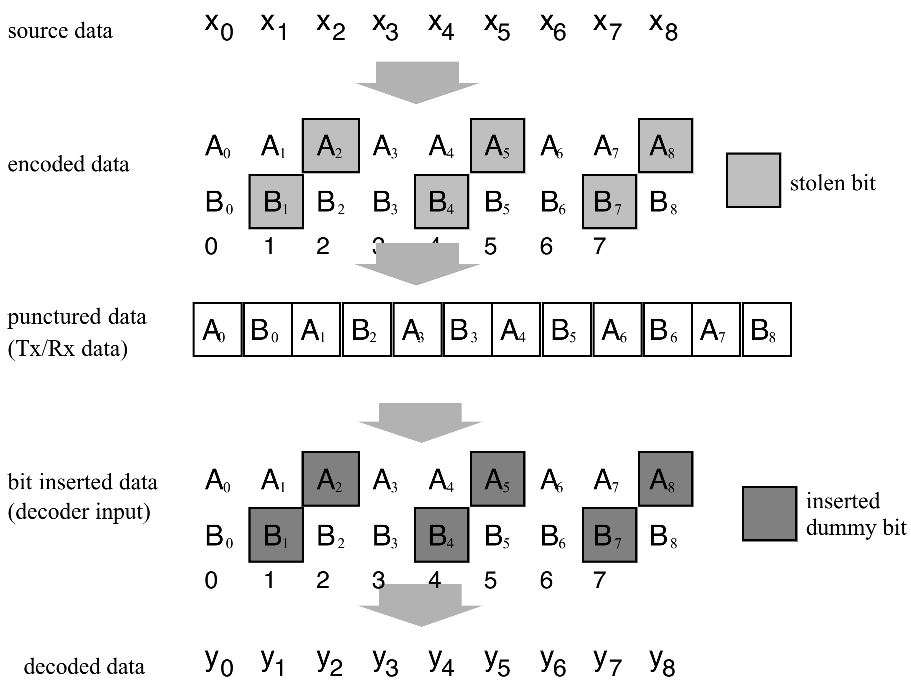

Figure 2: Puncturing for R=3/4, transmission and reception

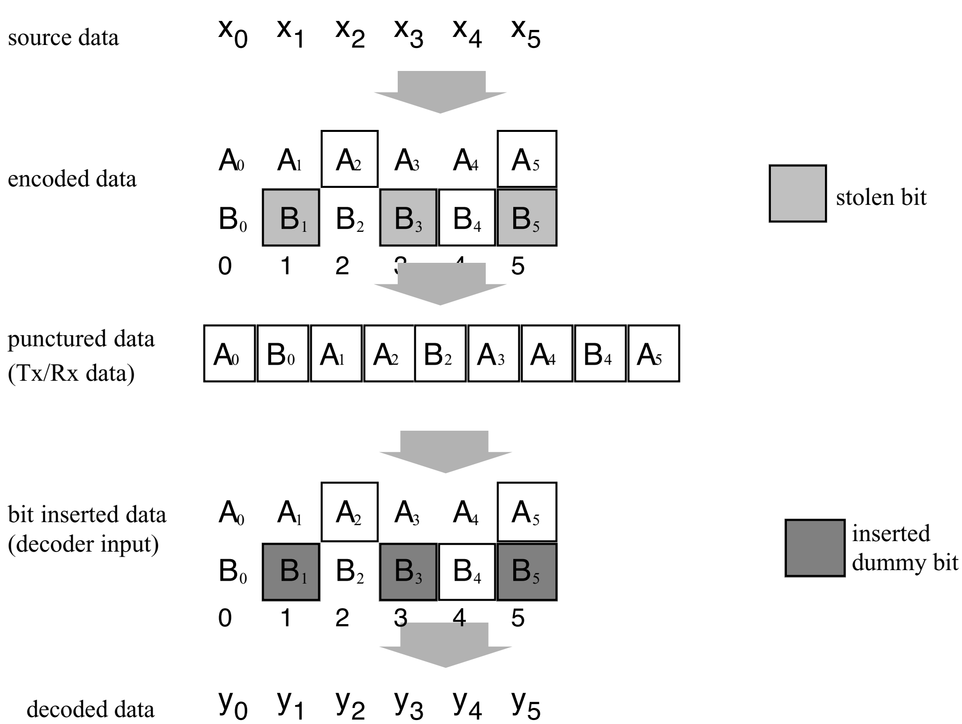

Figure 3: Puncturing for R=2/3, transmission and reception

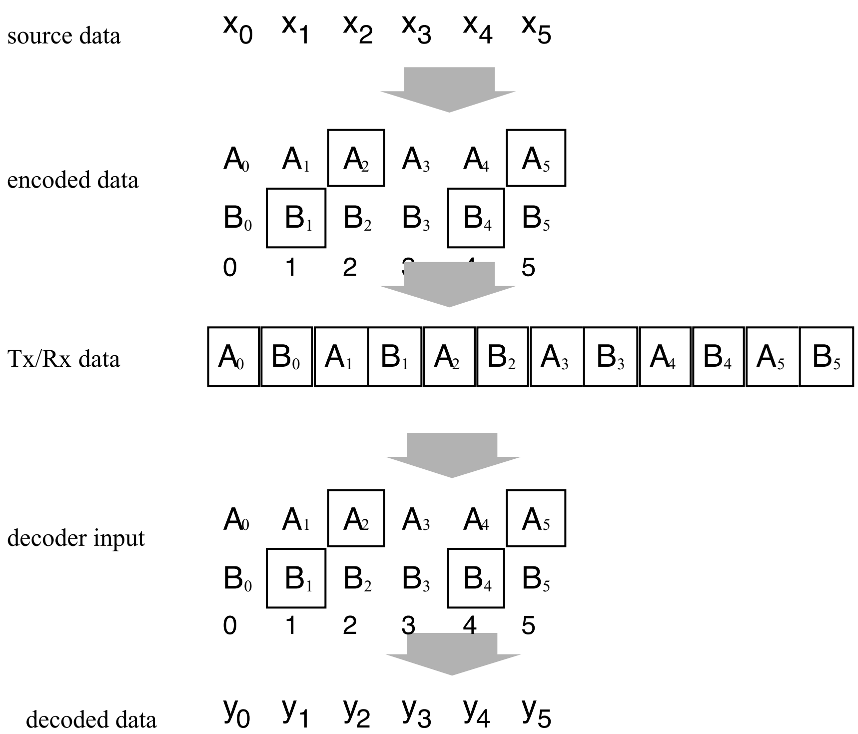

Figure 4: Bit ordering for R=1/2, transmission and reception

Interleaving

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

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

The first permutation is defined by the rule:

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

The second permutation is defined by the rule:

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

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

The first permutation is defined by the rule:

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

The second permutation is defined by the rule:

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

The first permutation in the deinterleaver is the inverse of the second permutation in the interleaver, and conversely. *Both permutations can be applied as they are for any number of subcarriers provided that* N_{CBPS} *is a multiple of 16. They can be easily changed so that this requirement is changed to* N_{CBPS} *be a multiple of 8 or 4.*

Dynamic adaptive modulation and coding

The ARQ mechanism can be combined with dynamic adaptive modulation and coding to meet the required BER/PER while maximizing the system through-output. Assume PER is P_e for a given packet size, modulation and coding. If up to *n* retransmissions (with same modulation and coding) are allowed, the overall PER theoretically improves to P_e^{n+1}. *However, to keep delay and overhead small, n is usually small, often limited to 1, which in turn limits the improvement brought by ARQ.* On most wireless channels the path loss and interference are variable but in unlicensed bands these may be extremely variable. In such a case, retransmissions with the same modulation and coding show a PER improvement much lower than the theoretical limit. We expect situations in which retransmissions require lower data rates to be successful. The solution is to combine the ARQ mechanisms with adaptive modulation and coding. *For example, two data rates (one rate refers here to a certain combination of modulation and coding) can be used for each connection. The first (high) data rate will be used for the first transmission while the second (low) data rate will be used for all retransmissions. Many other similar solutions can also be considered.*

Anyway, to enable ARQ with adaptive modulation and coding, the 802.16.4 MAC and PHY (*must*) have provisions to:

- change modulation and coding from CPE to CPE to accommodate channel diversity
- for same CPE, change modulation and coding from CID to CID to provide different QOS
- for same CID change modulation and coding on a packet by packet basis to implement efficient ARQ mechanisms

Transmit center frequency and symbol clock frequency tolerance

The transmitted center frequency and the symbol clock frequency shall be derived from the same reference oscillator. At the BS the reference frequency tolerance shall be +/- 5ppm or +/- 20ppm maximum (*here Zion shall add his arguments*). At the SS, both the transmitted center frequency and the symbol clock frequency shall be synchronized to the BS with a tolerance of +/- ?ppm (*TBD*) maximum. During the synchronization period as described in the 802.16 MAC, the SS shall acquire frequency synchronization with the specified tolerance before attempting any uplink transmission. During normal operation, the SS shall track the frequency changes and shall defer any transmission if synchronization is lost.

Base station frame synchronization and coordination mechanism

In sectorized environments, where several BS's are collocated in the same hub, the co-channel and adjacent channel interference can be significantly reduced if BS's are frame synchronized, i.e. they switch between transmit and receive and reverse at the same time. To implement this, the hub controller (or network controller) has (*must have*) access to the statistics of the uplink and downlink requests in all BS's, from which it can decide on an optimum uplink/downlink partition for the entire hub. The hub controller is (*must be*) also allowed to set the frame size and the uplink/downlink partition for the BS's. Therefore, the MAC provides (*must provide*) a standardized interface (*TBD*) to the hub controller. The MAC is (*shall be*) capable to:

- report upon request, the sizes (e.g. Kbytes) of total downlink and total uplink pending requests
- set its frame size upon request
- set its uplink/downlink partition upon request

For synchronization, the BS's need a common time reference, for example a GPS-like signal providing one pulse per second (*TBD*). Therefore, the BS can synchronize the start of a new MAC frame with the pulse received on its GPS-like input. The same pulse triggers the changes in the frame size and uplink/downlink partition in all BS's at the same time. To preserve synchronization over a long period of time, the BS's need also a common reference frequency. Consequently, the BS can synchronize (lock) its reference oscillator to a GPS-like external reference of 10 MHz (*TBD*).

Subcarrier based parallel polling

The 802.16 MAC provides several mechanisms that can be used by an SS to request from BS additional bandwidth: piggyback, unicast polling and multicast polling. In the later case, all SS's matching the group address can place their bandwidth request in the BW Request Contention part of the uplink. The (*proposed*) subcarrier based polling is an alternate way to answer the multicast polling without contention.

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

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

OFDM Preambles

Preambles are special training sequences that are prepended to the OFDM symbols carrying data. The receiver uses them for signal detection, AGC convergence, carrier recovery, symbol timing recovery and equalization. The 802.16.4 PHY uses a fully featured preamble and shortened versions of it depending on the burst position in the frame. The complete preamble is depicted in **Figure 5** and consists in 10 (TBD) short training symbols and 1 long training symbol.

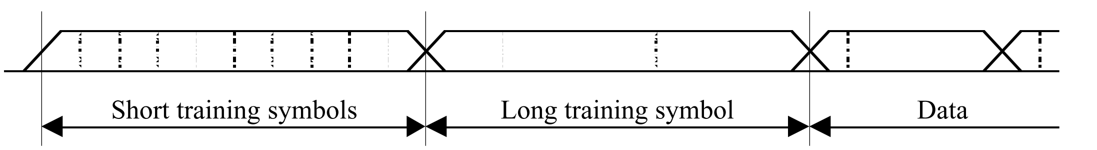

Figure 5: Complete preamble structure

A short training symbol has no cyclic prefix and consists of 12 carriers with an FFT size of 16 modulated by the elements of the sequence (*to discuss the factor*):

$$S^{16}_{-7,7} = \{0, 1+j, -1-j, 1+j, -1-j, -1-j, 1+j, \mathbf{0}, -1-j, -1-j, 1+j, 1+j, 1+j, 1+j, 0\}$$

Equivalently, the same 12 carriers are spaced at 4 times the inter-carrier spacing with an FFT size of 64:

$$S^{16}_{-31,31} = \{0, 0, 0, 0, 0, 0, 0, 1+j, 0, 0, 0, -1-j, 0, 0, 0, 1+j, 0, 0, 0, -1-j, 0, 0, 0, -1-j, 0, 0, 0, 1+j, 0, 0, 0, \mathbf{0}, 0, 0, 0, -1-j, 0, 0, 0, -1-j, 0, 0, 0, 1+j, 0, 0, 0, 1+j, 0, 0, 0, 1+j, 0, 0, 0, 1+j, 0, 0, 0, 0, 0, 0, 0\}$$

and at 16 times the inter-carrier spacing with an FFT size of 256:

$$S^{16}_{-127,127} = \{0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 1+j, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, -1-j, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 1+j, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, -1-j, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, -1-j, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 1+j, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, \mathbf{0}, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, -1-j, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, -1-j, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 1+j, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 1+j, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 1+j, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 1+j, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0, 0\}$$

Having a period of only 16 samples, short training symbols are suitable for signal detection and for fast AGC. Having enlarged inter-carrier spacing, short training symbols can be used for coarse carrier recovery for an offset up to half of the carrier spacing. **Table 1** summarizes the duration of the short training symbols and their offset recovery capabilities at 5.8GHz for different channel bandwidths (clock frequencies).

Table 1: Short training symbols vs. channel bandwidth

Channel BW	Short symbol duration	Total duration for short symbols	Allowed offset
5 MHz	3.2 µs	32 µs	26 ppm
10 MHz	1.6 µs	16 µs	53 ppm
20 MHz	0.8 µs	8 µs	107 ppm

The long training symbol is an OFDM symbol generated with the same FFT size as the data symbols but with a cyclic prefix (guard interval or GI) extended so that the overall length is 2 times the nominal length of a data symbol (TBD). (*Description changed from 802.11a for clarity and flexibility of the definition*.) It is BPSK modulated with a known/fixed pattern. It is used for fine carrier offset recovery, symbol timing recovery and equalization. The first two functions require the extended cyclic prefix; the equalization requires the same cyclic prefix as normal (data) OFDM symbols. For an FFT size of 64, the long training symbol is modulated by the sequence (*TBD*):

$S_{-31,31}$ = {0, 0, 0, 0, 0, 1, 1, -1, -1, 1, 1, -1, 1, -1, 1, 1, 1, 1, 1, 1, -1, -1, 1, 1, -1, 1, -1, 1, 1, 1, 1, **0**,
1, -1, -1, 1, 1, -1, 1, -1, 1, -1, -1, -1, -1, -1, 1, 1, -1, -1, 1, -1, 1, -1, 1, 1, 1, 1, 0, 0, 0, 0, 0}

For an FFT size of 256 … (*TBD*). **Table 2** summarizes the duration of the long training symbol and the nominal OFDM symbol for GI equal to 1/4 of the FFT period.

Table 2: Nominal symbol and long training symbol duration for GI=1/4

FFT size	Channel BW	Nominal symbol duration	Long training symbol duration
64	5 MHz	4 µs	8 µs
	10 MHz	8 µs	16 µs
	20 MHz	16 µs	32 µs
256	5 MHz	16 µs	32 µs
	10 MHz	32 µs	64 µs
	20 MHz	64 µs	128 µs

The preamble usage depends on the burst position within the frame. Complete preambles shall be used at the beginning of the downlink as well as for the bursts in the in Ranging Request Contention window. All other messages on the uplink except the BW Request in subcarrier based polling shall be sent with a shortened preamble that contains only the long training symbol and is depicted in **Figure 6**.

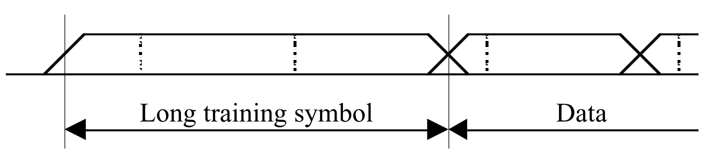

Figure 6: Shortened preamble: long training symbol with large GI

The BW Request in subcarrier based polling shall be prepended only with the minimal preamble needed for equalization, i.e. long training sequence with nominal cyclic prefix (same as data symbols). Such a preamble is illustrated in **Figure 7**.

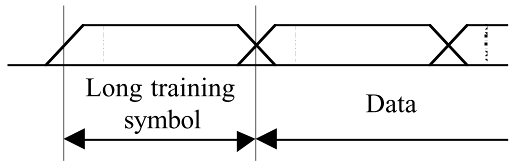

Figure 7: Minimal preamble: long training symbol with nominal GI

The minimal preamble can also be inserted during long bursts to refresh equalization for robustness, before changing the FFT size or before changing modulation. **Table 3** summarizes the use of preambles.

Table 3: Preamble usage vs. burst position

Burst position	Preamble	Description
Downlink	long	short and long training symbols
Ranging Req. Contention	long	short and long training symbols
BE Req. subcarrier based polling	minimal	long training symbol w/ nominal GI
Other uplink bursts	short	long training symbol w/ extended GI
Equalization refresh	minimal	long training symbol w/ nominal GI

Comments for the 802.16.4 PHY Draft

Reed-Solomon vs. ARQ with adaptive modulation

Convolutional coding combined with interleaving is used in most OFDM systems like DAB, DVB, Hyperlan and 802.11a. It can resolve deep channel fades and narrow band interference as long as the overall received signal to noise and interference ratio is still good. The convolutional code used in802.11a seems a very good compromise between implementation complexity and performance. Using higher order polynomials would make Viterbi decoder prohibitively expensive while adding very little in performance. Both the encoder and the puncturing patterns are industry standard. The interleaver is also well designed for data traffic: it does not introduce delay over one symbol and provides bit permutation across carriers and modulation levels. We strongly suggest keeping the 802.11a convolutional code, puncturing patterns and interleaving at least as the inner FEC layers.

When convolutional decoder fails, the reconstructed bit stream typically contains a long train of errors that are not correctable in this form by any other FEC techniques. To efficiently add another layer of coding, called hereby outer coding (as in DVB), one has to employ another level of interleaving, called hereby outer interleaving, that will split the long error burst into several small bursts. The outer interleaving has to operate on a larger time scale than the inner interleaving.

The advantage of adding an outer FEC layer is that it can further reduce the BER/PER. In a unidirectional broadcasting (i.e. with no feedback) application like DVB this is the only way to improve BER/PER. However, in data applications like 802.11a or 802.16, the ARQ mechanisms can replace and eventually outperform the outer coding. First, the message-response delay in such applications is important and thus having only one level of interleaving may by a clear advantage. Second, ARQ is based on feedback, it is adaptive and can achieve much lower BER/PER that any FEC. If adaptive modulation and coding are also used, the ARQ mechanisms can ensure constant BER/PER for a variable channel by changing the data rate. In turn, any fixed FEC technique will have the BER/PER dependent on channel variations. These issues become very important in unlicensed bands where the interference can vary dramatically from packet to packet.

Reed-Solomon codes are very efficient in rejecting short error bursts. They are particularly useful in channels where noise bursts are rare, short and random, caused by natural phenomena. They have little value in unlicensed bands where noise bursts can be caused by similar data transmission devices and can have a length comparable with the data packet. In terms of overhead, the Reed-Solomon code requires 8 bytes per data block while a CRC (needed for ARQ) uses only 4 bytes. Also, the CRC can capture errors with much higher probability than the Reed-Solomon. Therefore, the Reed-Solomon code cannot replace the CRC and ARQ mechanisms. The CRC are easy to implement, does not require an outer interleaving and does not introduce significant delay.

Channelization of 5, 10 or 20 MHz

Using a channelization of 10 MHz or 5 MHz has the advantage of doubling or quadrupling the number of channels in the UNII band. However, there are several advantages in using 20 MHz channels. The 20 MHz channel makes optimal use of the FCC part 15E. In UNII bands, FCC limits both the power (24 dBm for 5.25-5.35 GHz and 30 dBm for 5.725-5.825 GHz) and the power density (11 dBm/MHz for 5.25-5.35 GHz and 17 dBm/MHz for 5.725-5.825 GHz). The difference between the power and the power density is 13dB (20MHz) in both bands. In other words, a wider channel would have to use reduced power density while a narrower channel would have to use reduced power.

Therefore, users of smaller transmitted bandwidths will have to proportionally scale down the transmitter power to keep the power density within the FCC limits. In the best case this means proportionally scaling down the data rate. However, the narrower the channel the more sensitive is to fading and narrow-band interference. With a fade spanning over 4 MHz, the Viterbi decoder can easily recover data for a 20 MHz channel, can hardly recover data for a 10 MHz channel and cannot recover data for a 5MHz channel. For comparison the 6 MHz TV channel represents 0.66% of the maximum center frequency in UHF band while the 20 MHz channel represents 0.34% of the maximum center frequency in UNII band. Narrowing the channel will also increase the system sensitivity to phase noise, propagation-induced phase jitter. Going to narrower bands will practically decrease the deliverable data rate across the 100MHz bandwidth, while requiring more equipment.