

A Design of Next Generation Gigabit MIMO Wireless LAN System

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Abstract—This paper deals with a design of next generation Gigabit WLAN system based on IEEE802.11TGac's functional requirements. It achieves 1.1 Gbps throughput for 45 meter propagation distance with 80 MHz bandwidth on 5 GHz band. The proposed preamble ensures backward compatibility with the IEEE802.11a/n systems and has lower PAPR and comparable preamble efficiency with IEEE802.11n's preamble. It uses MIMO architecture to maintain both the high throughput and the high performance. Two options of channel encoder, binary convolutional code and low density parity check code are available to enhance the performance. Run test under in-door channel model demonstrates the excellent performance of the proposed system.

Index Terms—Gigabit, MIMO, WLAN, backward compatibility, IEEE802.11a/n

I. INTRODUCTION

The next generation wireless LAN (WLAN) systems should have ability to support numerous applications of wireless communication. This leads to a combination of the orthogonal frequency division multiplexing (OFDM) and multi input multi output (MIMO) techniques to provide a WLAN system with Gigabit throughput.

The IEEE802.11TGac have been discussing about the standard for very high throughput (VHT) WLAN system which has 1 Gbps throughput at medium access (MAC) layer [1]. This system is assigned to work in the 5 GHz band frequency, same as IEEE802.11a/n WLAN systems [2], [3]; therefore, ensuring the backward compatibility with those systems is included in the functional requirements. For evaluation, channel model B of IEEE802.11TGn [4] can be resampled and interpolated to model the in-door channel for IEEE802.11TGac system [5].

This paper deals with a design of next generation Gigabit WLAN system based on IEEE802.11TGac's functional requirements. It achieves 1.1 Gbps throughput for 45 meter propagation distance by utilizing 80 MHz bandwidth on 5 GHz band. The proposed preamble ensures backward compatibility with IEEE802.11a/n systems. This means that when the proposed preamble is transmitted, the IEEE802.11a/n WLAN devices can recognize it and consider that the 5 GHz band is busy. It is kind of protection mechanism to avoid collision within 5 GHz band. Further, we propose a novel

phase rotation before the OFDM modulation to lower the peak-to-average power ratio (PAPR) and provide the data length two times more than that of IEEE802.11n system to increase the preamble efficiency. MIMO architecture with four transmission streams and five receive antennas contribute 2nd-order diversity gain maintain both the high throughput and the high performance. As a forward error correction (FEC) scheme, besides the binary convolutional code (BCC) with soft Viterbi decoder, the performance enhancement using low density parity check code (LDPC) with min-sum algorithm decoder is also conducted. Run test under in-door channel model demonstrates the excellent performance of the proposed system.

The rests of this paper are organized as follows. The proposed next generation Gigabit WLAN system with backward compatibility to IEEE802.11a/n systems is briefly explained in section II. Section III describes the contributions of the proposed preamble. Section IV shows the results of system performance examination and link budget analysis under in-door environment. Finally, we draw some conclusions and future works in section V.

II. A NEXT GENERATION GIGABIT WIRELESS LAN SYSTEM WITH BACKWARD COMPATIBILITY TO IEEE802.11a/n

Block diagram of transmitter of the proposed system is shown in Fig. 1 where the partial transmit sequence (PTS) block is inserted in each stream to reduce the PAPR. The proposed receiver is displayed in Fig. 2 where diversity gain is expected by using five receive antennas. Three samples of modulation coding scheme (MCS) which defines the parameters to calculate the data rate are listed in Table I. By using a MAC protocol with 80% efficiency the 1 Gbps throughput at MAC layer can be achieved by using the MCS-31 with 400 ns of guard interval (GI) length.

A. Mixed Format Preamble

Since the aim is ensuring backward compatibility with IEEE802.11a/n systems, we propose a mixed format (MF) preamble which consists of fields that can be recognized by IEEE802.11a system named Legacy fields (L), i.e. L-short training field (L-STF), L-long training field (L-LTF),

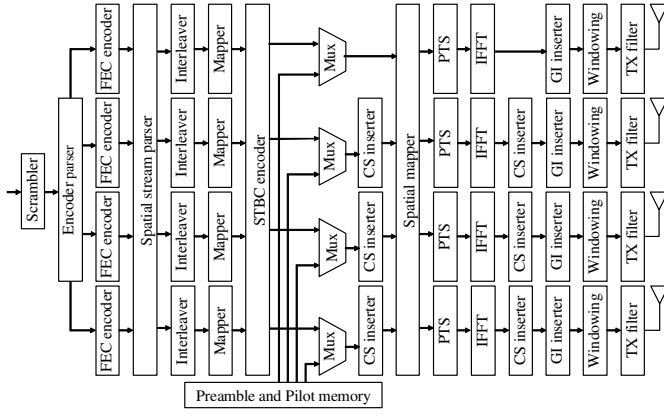


Fig. 1. Proposed Transmitter

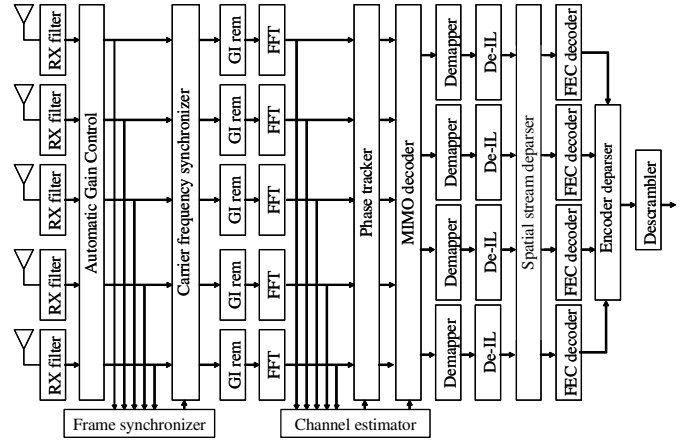


Fig. 2. Proposed Receiver

TABLE I
SAMPLES OF MODULATION CODING SCHEME FOR PROPOSED GIGABIT WLAN SYSTEM

MCS	Modulation	R	N_{BPCS} (i_{ss})	N_{SD}	N_{SP}	N_{CBPS}	N_{DBPS}	N_{SS}	Data rate [Mbps]	
									800ns GI	400ns GI
29	64-QAM	2/3	6	228	8	5472	3648	4	912	1013
30	64-QAM	3/4	6	228	8	5472	4104	4	1026	1140
31	64-QAM	5/6	6	228	8	5472	4560	4	1140	1266

and L-signal field (L-SIG), a field that can be recognized by IEEE802.11n system, called High Throughput (HT) field, i.e. HT-SIG, and fields for IEEE802.11ac system, named Very High Throughput (VHT) fields, i.e. VHT-SIG, VHT-STF, and VHT-LTF, as illustrated in Fig.3. Compared to IEEE802.11n system which also considers the backward compatibility with IEEE802.11a system, the proposed MF preamble is 8 μ s longer than IEEE802.11n's MF preamble. However we provide the data length which is two times longer than that of IEEE802.11n to mitigate this overhead problem. The transmitted signal on each transmit chain $i_{TX}, i = 1, 2, 3, 4$ can be written as:

$$\begin{aligned}
 s_{PPDU}^{(i_{TX})}(t) &= s_{LSTF}^{(i_{TX})}(t) + s_{LLTF}^{(i_{TX})}(t - t_{LLTF}) + s_{LSIG}^{(i_{TX})}(t - t_{LSIG}) \\
 &+ s_{HSIG}^{(i_{TX})}(t - t_{HSIG}) + s_{VSIG}^{(i_{TX})}(t - t_{VSIG}) + s_{VSTF}^{(i_{TX})}(t - t_{VSTF}) \\
 &+ \sum_{i_{LTF}=1}^{N_{LTF}} s_{VLTF}^{(i_{LTF}, i_{TX})}(t - t_{VLTF} - (i_{LTF} - 1)T_{VLTF}) \\
 &+ s_{Data}^{(i_{TX})}(t - t_{Data})
 \end{aligned} \quad (1)$$

where $t_{LLTF} = T_{LSTF}$; $t_{LSIG} = t_{LLTF} + T_{LLTF}$; $t_{HSIG} = t_{HLTF} + T_{HLTF}$; $t_{VSIG} = t_{HSIG} + T_{HSIG}$; $t_{VSTF} = t_{VSIG} + T_{VSIG}$; $t_{VLTF} = t_{VSTF} + T_{VSTF}$; $t_{Data} = t_{VLTF} + (N_{LTF})T_{VLTF}$.

The constants to calculate the timing of transmitted signal are listed in Table II. Next, each field of Eq. 1 is explained briefly as follows:

1) *L-STF*: is used for start-of-packet detection, AGC setting, initial frequency offset estimation, and initial time synchronization purpose. It is constructed by four times duplicating of IEEE802.11a STF followed by frequency shifting and phase rotating. The time domain waveform of L-STF on

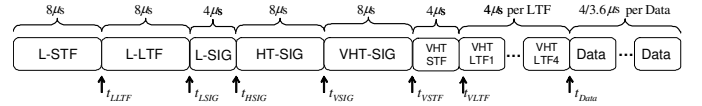


Fig. 3. The proposed mixed format (MF) preamble ensures backward compatibility with IEEE802.11a/n systems, and its time boundaries.

TABLE II
THE CONSTANTS FOR CALCULATION THE TIMING

Parameter	Value	Parameter	Value
Δ_F	312.5kHz (80MHz/256)	T_{LLTF}	8 (μ s)
T_{DFT}	3.2 μ s ($1/\Delta_F$)	T_{LSIG}	4 (μ s)
T_{GI}	0.8 ; 0.4 (μ s)	T_{HSIG}	8 (μ s)
T_{SYM}	4 ; 3.6 (μ s)	T_{VSIG}	8 (μ s)
T_{TR}	0.1 (μ s)	T_{VSTF}	4 (μ s)
T_{LSTF}	8 (μ s)	T_{VLTF}	4 (μ s)

transmit chain i_{TX} is:

$$s_{LSTF}^{(i_{TX})}(t) = \frac{1}{\sqrt{48 \cdot N_{TX}}} w_{T_{LSTF}}(t) \sum_{k=-122}^{122} r_k s_k e^{j2\pi k \Delta_F (t - T_{CS}^{i_{TX}})} \quad (2)$$

where $N_{TX} = 1, \dots, 4$ is number of transmit antenna, $\frac{1}{\sqrt{48 \cdot N_{TX}}}$ is the scale factor to maintain that the total power of the time domain signal as summed over all transmit chains is either one or lower than one. The constant 48 describes the number of used subcarriers by L-STF as well as by VHT-STF in one OFDM symbol. $w_{T_{LSTF}}(t)$ is the time windowing function which is defined as a rectangular pulse $w_T(t)$ of duration T as Eq. 3. Table II lists the duration of each field (T_{field}).

$$w_T(t) = \begin{cases} \sin^2(\frac{\pi}{2}(0.5 + \frac{t}{T_{TR}})) & \text{for } (-\frac{T_{TR}}{2} < t < \frac{T_{TR}}{2}) \\ 1 & \text{for } (\frac{T_{TR}}{2} \leq t < \frac{T-T_{TR}}{2}) \\ \sin^2(\frac{\pi}{2}(0.5 - \frac{(t-T)}{T_{TR}})) & \text{for } (\frac{T-T_{TR}}{2} \leq t < \frac{T+T_{TR}}{2}) \end{cases} \quad (3)$$

TABLE III
PROPOSED CYCLIC SHIFT VALUES FOR LEGACY PORTION ($T_{CS}^{i_{TX}}$)

T_{CS}^1	T_{CS}^2	T_{CS}^3	T_{CS}^4
0 ns	- 50 ns	- 100 ns	- 150 ns

where T_{TR} is the transition time between two consecutive fields. γ_k represents the proposed rotation of the tones in 80MHz channel to get low PAPR signal while considering the backward compatibility with IEEE802.11a/n system, as:

$$\gamma_k = \begin{cases} 1 & \text{for } k \leq -64, 0 < k \leq 64, \\ j & \text{for } -64 < k \leq 0 \\ -j & \text{for } k > 64 \end{cases} \quad (4)$$

where k is the subcarrier index in the spectral line $S_{-128,127}$. S_k are the four times duplication of IEEE802.11a STF symbols placed at indices that are a multiple of four. This generates waveforms which has a period of 0.8 μs , and the L-STF includes ten such periods, with a total duration of 8 μs . $T_{CS}^{i_{TX}}$ represents the cyclic shift (CS) for transmit chain i_{TX} to prevent unintentionally beamforming which considers compatibility with the legacy IEEE802.11a. The values of CS for legacy portion are specified in Table III.

2) *L-LTF*: is used for fine frequency offset estimation, time synchronization, and estimate the MIMO channel's characteristics for decoding the SIGNAL fields. The time domain representation of the L-LTF on transmit chain i_{TX} is:

$$s_{LLTF}^{(i_{TX})}(t) = \frac{1}{\sqrt{208 \cdot N_{TX}}} w_{T_{LLTF}}(t) \sum_{k=-122}^{122} \gamma_k L_k e^{j2\pi k \Delta_F (t - 2T_{GI} - T_{CS}^{i_{TX}})} \quad (5)$$

where L_k are the four times duplication of IEEE802.11a LTF symbols. The constant 208 describes the number of used subcarriers by L-LTF as well as by all SIG fields in one OFDM symbol. Equation 5 occurs in 3.2 μs time domain sequence. The L-LTF has 8 μs duration which obtained by repeating this sequence twice and prepending with 1.6 μs cyclic prefix.

3) *L-SIG field*: is inserted to maintain the backward compatibility with IEEE802.11a system. It contains IEEE802.11a SIGNAL format which is composed of 24 bits information, all are convolutional encoded at rate=1/2, interleaved and BPSK mapped. The stream of 48 complex numbers generated by these steps is defined as: $D_k, 0 \leq k \leq 47$. The time domain waveform of the L-SIG field on transmit chain i_{TX} is:

$$s_{LSIG}^{(i_{TX})}(t) = \frac{1}{\sqrt{208 \cdot N_{TX}}} w_{T_{LSIG}}(t) \sum_{k=-26}^{26} (D_k + P_k) (e^{j2\pi(k-96)\Delta_F(t-T_{GI}-T_{CS}^{i_{TX}})} + j e^{j2\pi(k-32)\Delta_F(t-T_{GI}-T_{CS}^{i_{TX}})} + e^{j2\pi(k+32)\Delta_F(t-T_{GI}-T_{CS}^{i_{TX}})} - j e^{j2\pi(k+96)\Delta_F(t-T_{GI}-T_{CS}^{i_{TX}})}) \quad (6)$$

where D_k and P_k are the data of L-SIG field and pilot, respectively. The data is null at the pilot's subcarriers.

4) *HT-SIG field*: is inserted to keep the backward compatibility with IEEE802.11n system. It needs to be decoded but the contained information does not need to be interpreted. It

consists of IEEE802.11n HT-SIGNAL format which is composed of HT-SIG1 and HT-SIG2 each containing 24 bits. All are convolutional encoded at rate=1/2, interleaved and BPSK mapped. The stream of 96 complex numbers generated by these steps is divided into two groups of 48 complex number: $D_{k,n}, 0 \leq k \leq 47, n = 0, 1$. The time domain representation of the HT-SIG in transmit chain i_{TX} is:

$$s_{HSIG}^{(i_{TX})}(t) = \frac{1}{\sqrt{208 \cdot N_{TX}}} \sum_{n=0}^1 w_{T_{HSIG}}(t) \sum_{k=-26}^{26} (j D_{k,n} + p_{n+1} P_k) (e^{j2\pi(k-96)\Delta_F(t-nT_{SYM}-T_{GI}-T_{CS}^{i_{TX}})} + j e^{j2\pi(k-32)\Delta_F(t-nT_{SYM}-T_{GI}-T_{CS}^{i_{TX}})} + e^{j2\pi(k+32)\Delta_F(t-nT_{SYM}-T_{GI}-T_{CS}^{i_{TX}})} - j e^{j2\pi(k+96)\Delta_F(t-nT_{SYM}-T_{GI}-T_{CS}^{i_{TX}})}) \quad (7)$$

where $D_{k,n}$ is the data of HT-SIG which allocated on the k -th subcarrier of n -th OFDM symbol. p_n is pilot polarity controller sequence which is generated by IEEE802.11a's scrambler with the "all ones" initial state and by replacing all "1's" with -1 and all "0's" with 1. Equation 7 results in a BPSK modulation in which the constellation of the data tones is rotated 90 [deg] relative to L-SIG.

5) *VHT-SIG field*: contains information about the transmitted frame and has special format as shown in Fig. 4. It composed of VHT-SIG1 and VHT-SIG2 each containing 24 bits. All are convolutional encoded at rate=1/2, interleaved and BPSK mapped. The stream of 96 complex numbers generated by these steps is divided into two groups of 48 complex number: $D_{k,n}, 0 \leq k \leq 47, n = 0, 1$. The time domain form of the VHT-SIG in transmit chain i_{TX} is:

$$s_{VHSIG}^{(i_{TX})}(t) = \frac{1}{\sqrt{208 \cdot N_{TX}}} \sum_{n=0}^1 w_{T_{VHSIG}}(t) \sum_{k=-26}^{26} (j D_{k,n} + p_{n+2} P_k) (e^{j2\pi(k-96)\Delta_F(t-nT_{SYM}-T_{GI}-T_{CS}^{i_{TX}})} + j e^{j2\pi(k-32)\Delta_F(t-nT_{SYM}-T_{GI}-T_{CS}^{i_{TX}})} + e^{j2\pi(k+32)\Delta_F(t-nT_{SYM}-T_{GI}-T_{CS}^{i_{TX}})} - j e^{j2\pi(k+96)\Delta_F(t-nT_{SYM}-T_{GI}-T_{CS}^{i_{TX}})}) \quad (8)$$

As shown in the upper side of Fig. 4, VHT-SIG1 field provides data length up to 2^{17} octets which is two times more than that in IEEE802.11n system. This makes the proposed MF preamble has comparable efficiency with IEEE802.11n MF preamble which will be verified in section III.

6) *VHT-STF*: allows more accurate AGC setting and reduce the dynamic range requirement on the ADC of the proposed system. It has the same construction as L-STF but with longer CS to get comparable power with the data field. The values of CS for each STS are specified in Table IV. The time domain representation the VHT-STF in transmit chain i_{TX} is:

$$s_{VSTF}^{(i_{TX})}(t) = \frac{1}{\sqrt{48 \cdot N_{TX}}} w_{T_{VSTF}}(t) \sum_{k=-122}^{122} \sum_{i_{STS}=1}^4 [\mathbf{Q}_k]_{i_{TX}, i_{STS}} \gamma_k S_k e^{j2\pi k \Delta_F (t - T_{CS}^{i_{STS}})} \quad (9)$$

where S_k are the four times duplication of IEEE802.11a STF symbols. $\mathbf{Q}_k$ is a spatial mapping matrix which maps each

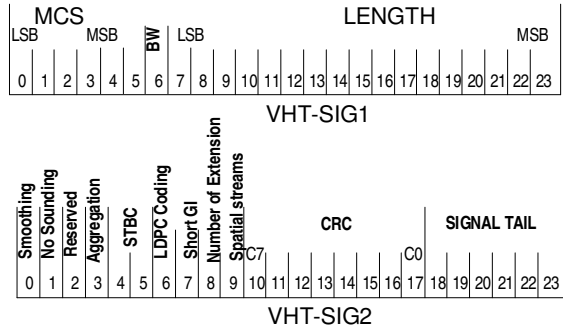


Fig. 4. Format of the proposed VHT-SIG. VHT-SIG1 provides the data length two times longer than that in IEEE802.11n system to increase preamble efficiency.

TABLE IV
PROPOSED CYCLIC SHIFT VALUES FOR NON-LEGACY PORTION (T_{CS}^{iSTS})

T_{CS}^1	T_{CS}^2	T_{CS}^3	T_{CS}^4
0 ns	- 400 ns	- 200 ns	- 600 ns

STS symbols onto transmit chain symbols $s_k^{(i_{TX})}$. For line of sight (LOS) environment $\mathbf{Q}_k$ is an identity matrix. For NLOS environment, expansion mapping of $\mathbf{Q}_k$ is applied.

7) *VHT-LTF*: estimates the MIMO channel's characteristics for decoding the Data field. It is constructed by two times duplication of the IEEE802.11n HT-LTF followed by frequency shift and phase rotation. Each VHT-LTF has 4 μ s duration. The time domain representation of the VHT-LTF on transmit chain i_{TX} is:

$$s_{VHT-LTF}^{(n, i_{TX})}(t) = \frac{1}{\sqrt{228 \cdot N_{TX}}} w_{VHT-LTF}(t) \sum_{k=-122}^{122} \sum_{i_{STS}=1}^4 [\mathbf{Q}_k]_{i_{TX}, i_{STS}} [\mathbf{P}]_{i_{STS}, n} \gamma_k H_k e^{j2\pi k \Delta_F (t - T_{GI} - T_{CS}^{i_{STS}})} \quad (10)$$

where n is the number of VHT-LTF in each transmit chain. H_k are the two times duplication of IEEE802.11n HT-LTF symbols. $\mathbf{P}$ is 4×4 orthogonal mapping matrix defined as:

$$\mathbf{P} = \begin{bmatrix} 1 & -1 & 1 & 1 \\ 1 & 1 & -1 & 1 \\ 1 & 1 & 1 & -1 \\ -1 & 1 & 1 & 1 \end{bmatrix} \quad (11)$$

The constant 228 describes the number of used subcarriers by VHT-LTF as well as by the data field in one OFDM symbol.

B. The Data Field

The Data field consists of the 16-bit SERVICE field, the physical (PHY) sublayer service data unit (PSDU), 24 TAIL bits for four encoding streams, and PAD bits, when BCC encoding is used. The Data field consists of the 16-bit SERVICE field and the PSDU, when LDPC encoding is used. All bits in the Data field are scrambled.

The SERVICE field is used for scrambler initialization. It is composed of 16 bits, all set to zero before scrambling. The TAIL bits are six bits of zero for each stream which are required to return the convolutional encoder to the "zero state".

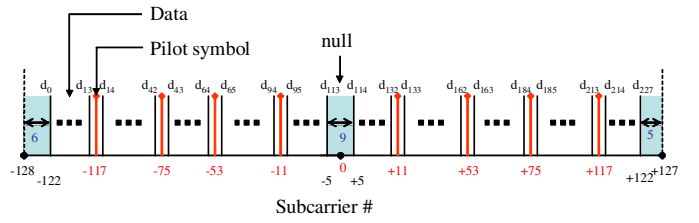


Fig. 5. Proposed data, pilot and null subcarriers allocation on OFDM symbol.

These TAIL bits are produced by replacing six scrambled "zero" bits following the message end with six nonscrambled "zero" bits. The PAD bits is appended so that the number of bits in the DATA field is a multiple of N_{CBPS} .

To reduce the probability of long sequences of zeros or ones, the Data field is scrambled by using frame synchronous scrambler which has generator polynomial $G(x) = x^7 + x^4 + 1$.

The scrambled data is encoded using forward error correction (FEC) encoder to enhance the performance against channel noise. BCC and LDPC are available for FEC. If BCC encoding is to be used, the scrambled data bits are divided by sending alternating bits to four different BCC encoders. The BCC encoder has generator polynomials $G_0 = 133_8$ and $G_1 = 171_8$ of rate $R = 1/2$. After encoding, the encoded data is punctured to achieve the rate selected by the MCS. If LDPC encoding is to be used the scrambled data bits are divided by sending alternating bits to two different LDPC encoders. The LDPC encoder is systematic, i.e., it encodes an information block, $\mathbf{c} = (i_0, i_1, \dots, i_{(k-1)})$, of size k , into a codeword, $\mathbf{c}$, of size n , $\mathbf{c} = (i_0, i_1, \dots, i_{(k-1)}, p_0, p_1, \dots, p_{(n-k-1)})$, by adding $n-k$ parity bits obtained so that $\mathbf{H} \cdot \mathbf{c}^T = 0$, where $\mathbf{H}$ is an $(n-k) \times n$ parity-check matrix.

Spatial stream parser divides the encoded data into blocks of $N_{CBPS}(i_{SS})$, $i_{SS} = 1, 2, 3, 4$ bits. If BCC encoding was used, each block is interleaved by a three steps permutation interleaver in frequency domain then mapped to PSK or QAM symbols. If LDPC encoding was used, since the structure of encoded data is already interleaved, no more frequency interleaving is performed, hence the parsed streams are immediately mapped to PSK or QAM symbols.

Two options of Spatial mapper $[\mathbf{Q}]$ are available, direct and extension mapping. The later mapping promises robust communication in NLOS environment.

The 256 inverse fast Fourier transform (IFFT) point is used to get the time domain OFDM signals. The indices 1 to 122 are mapped to the same numbered IFFT inputs, while the indices -122 to -1 are copied into IFFT inputs 134 to 255. The rest of the inputs: 123 to 133 and the 0 (dc) input are set to zero. After performing an IFFT, the output is cyclically extended as a GI. Two options of GI length are available to get the desired data rate.

Eight pilot symbols are inserted in the sub-carriers $k = -117, -75, -53, -11, 11, 53, 75$ and 117. Each STS has a different determined pilot pattern denoted as $P_{i_{STS}, n}^k$, $n = 1, 2, \dots, 8$. The proposed subcarriers allocation for data, pilot and null in one OFDM symbol is illustrated in Fig. 5.

The time domain waveform of the VHT-Data on transmit

chain i_{TX} can be written as:

$$s_{Data}^{(i_{TX})}(t) = \frac{1}{\sqrt{228 \cdot N_{TX}}} \sum_{n=0}^{N_{SYM}-1} w_{T_{SYM}}(t - nT_{SYM}) \sum_{k=-122}^{122} \sum_{i_{STS}=1}^4 [Q_k]_{i_{TX}, i_{STS}} (D_k + p_{n+2} P_{i_{STS}, n}^k) \gamma_k \cdot e^{j2\pi k \Delta_F (t - nT_{SYM} - T_{GI} - T_{CS}^{i_{STS}})} \quad (12)$$

C. The Receiver Side

In this part we introduce very briefly the receiver side. After all frame are synchronized and the GIs are removed, each stream is demodulated using the fast Fourier transform (FFT). Five streams of received training sequences are exploited to estimate the MIMO channel characteristics including phase errors. The minimum mean square error (MMSE) of MIMO decoder which is used to cancel the interference signals contributes 2^{nd} order diversity gain. This comes from MIMO linear decoder diversity which is stated as $N_T - N_R + 1$ [7]. Output of the MIMO decoder are four streams of the estimated symbols. Demapper remaps the estimated symbols onto sequence of bits. If BCC encoding was used, the deinterleaver returns the sequence of data bits into original sequence per block, before errors are corrected by soft decision Viterbi decoder. If LDPC encoding was used, the sequence of data bits is decoded by min-sum algorithm-based LDPC decoder. Finally, descrambler returns the decoded data bits into its original order.

III. CONTRIBUTION OF THE PROPOSED PREAMBLE

In this section we compare the proposed MF preamble with the MF preamble of IEEE802.11n system. A conventional 80 MHz channel with 256 subcarriers for Gigabit system is illustrated in Fig. 6. It consists of two 40 MHz channels which are compatible with IEEE802.11n 40 MHz system, i.e. the subcarriers in the upper 20 MHz is 90 [deg] phase rotated relative to the subcarriers in the lower one. This phase rotation, as a special form of PTS technique, reduces the PAPR of 40 MHz signal. However applying this construction in 80 MHz channel as it is yields in high PAPR signal.

Here we propose a new phase rotation to reduce the PAPR of 80 MHz signal with simple procedure. The subcarriers in the 20 MHz upper of the 40 MHz upper is phase rotated 90 [deg] c.c.w. relative to the subcarriers in the lower one as shown in Fig. 7, which is also a description of Eq. 4. This configuration is applied in all fields, so that the channel information of the 40 MHz upper which is taken from the LTFs preamble is valid to be used for decoding the SIG fields and the Data field.

The PAPR of time domain OFDM signal which has N samples can be calculated by:

$$PAPR \text{ (dB)} = 10 \log_{10} \frac{\max(|s_n|^2)}{E\{|s_n|^2\}}, \quad n = 0, \dots, N-1 \quad (13)$$

Table V lists the obtained PAPR of the proposed MF preamble of the Gigabit system using conventional and proposed phase rotation compared to IEEE802.11n MF preamble after four times over sampled. By applying the proposed phase rotation we get significant PAPR reduction from the conventional one.

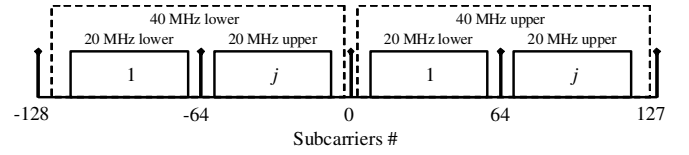


Fig. 6. Construction of the conventional 80 MHz channel.

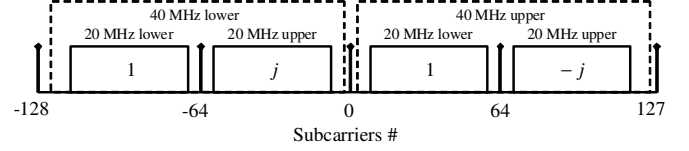


Fig. 7. Construction of the proposed 80 MHz channel for PAPR reduction.

This makes the proposed MF preamble has lower PAPR than IEEE 802.11n MF preamble. The PAPR value for the STF and LTFs are constant, while for SIGs and Data field are vary depend on the contained information.

Next, the preamble efficiency can be approached by:

$$\eta = \left(\frac{T_{SYM} \cdot N_{SYM}}{T_{PREAMBLE} + T_{SYM} \cdot N_{SYM}} \right) 100\% \quad (14)$$

where $N_{SYM} = \lceil \frac{8 \cdot LENGTH + 16 + 6 \cdot N_{ES}}{N_{DBPS}} \rceil$ is number of OFDM symbols in the Data field if BCC encoding was used, and N_{ES} is number of BCC encoders. The preamble efficiency of both systems for four spatial streams with maximum $LENGTH$ and $T_{SYM} = 4\mu s$ are listed in Table VI.

IV. PERFORMANCE EXAMINATION

We observe three scenarios by combining antenna configuration and FEC encoder in MCS-3 with 400 ns of GI duration. Channel model B of IEEE802.11Tgn is resampled to model the in-door environment for examining the proposed system. Table VII lists the simulation parameters.

- 1) Scenario-A: 4×4 MIMO with BCC as the FEC.
- 2) Scenario-B: 4×5 MIMO with BCC as the FEC.
- 3) Scenario-C: 4×5 MIMO with LDPC as the FEC.

Fig. 8 shows the performance comparison of the above scenarios. For BER 10^{-6} , Scenario-A, B and C need 54 dB, 40 dB and 35 dB of SNR, respectively. As expected, in Scenario-B and C which employ 4×5 MIMO, we get 2^{nd} -order diversity gain about 14 dB better than Scenario-A that uses square antennas configuration. Scenario C has the lowest SNR with the cost of highest complexity due to the use of LDPC.

The link budget analysis to calculate the propagation distance can be approached by:

$$d = \frac{\lambda}{4\pi 10^{\frac{L}{20}}} \quad (15)$$

where λ is transmitted wave length, $L = L_{d \leq 5} = 10 \log_{10}(P_x B \alpha) + G_{TX} - G_{RX} - (SNR + 10 \log(kTB) + NF + IM)$ is the path loss for $d \leq 5$ m, and $L = L_{d \leq 5} + 35 \log_{10}(d/5)$ for $d > 5$ m. These parameters and their values are included in Table VII. $\alpha \leq 1$ is the efficiency factor and constant 5 is the LOS break-point distance. The propagation distance of three

TABLE V

PAPR VALUE OF THE CONVENTIONAL AND THE PROPOSED MF PREAMBLE OF THE GIGABIT SYSTEM COMPARED WITH IEEE802.11n MF PREAMBLE AFTER FOUR TIMES OVERSAMPLED.

Field	Conventional [dB]	Proposed [dB]	IEEE802.11n [dB]
L-STF	5.82	4.47	5.25
L-LTF	6.90	5.43	5.79
HT-STF	5.82	4.47	5.25
HT-LTF	7.09	5.57	6.36

TABLE VI

PREAMBLE EFFICIENCY (η) COMPARISON FOR FOUR SPATIAL STREAMS WITH MAXIMUM LENGTH AND $T_{SYM} = 4\mu s$

Preamble Mode	LENGTH	N_{SYM}	$T_{PREAMBLE}$	η [%]
Proposed MF	131071 octet	230	56 μs	94.26
IEEE802.11n MF	65535 octet	243	48 μs	95.29

scenarios is shown in Fig. 9. In the LOS case, for throughput over 1.1 Gbps, Scenario-B and C reach 45 meter propagation distance, while Scenario-A reaches about 23 meter. In the NLOS case the formers can propagate up to 20 meter, while the later reaches about 13 meter.

V. CONCLUSION

We have been developing a next generation Gigabit MIMO WLAN system which achieves throughput 1.1 Gbps with 45 meter propagation distance. The proposed MF preamble ensures backward compatibility with IEEE802.11a/n systems. It has comparable efficiency with IEEE 802.11n MF preamble, and lower PAPR than IEEE 802.11n MF preamble. Run test under in-door environment shows an excellent performance. MIMO architecture with 4×5 antennas configuration contributes 14 dB better performance than MIMO square 4×4 , and the LDPC contributes 6 dB more. We will implement and evaluate this system on FPGA chips as a future work [8].

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TABLE VII
SIMULATION PARAMETER

Parameter	Value
MCS Index	31
Frequency Carrier (f_c)	5.2 [GHz]
Antenna Configuration	Square 4×4 , RX diversity 4×5
Bandwidth	80 [MHz]
Format Packet	Mixed Format Mode
FEC Encoder	BCC with soft Viterbi decoder LDPC with min-sum algorithm decoder
Spatial Mapping	Direct
MIMO Decoder	Linear MMSE
Guard Interval Length	400 [ns]
Antenna Gain (G_{TX}, G_{RX})	0 [dB]
Noise Figure (NF)	10 [dB]
Implementation Margin (IM) *	5 [dB]
Transmit Power per BW (P_x)	2.5 [mW/MHz]
Boltzmann Constant (k)	1.381×10^{-23} [J/K]
Temperature (T)	290 [K]
Light speed (c)	3.01×10^8 [m/s]
Channel	in-door channel model

* : possible performance degradation in the hardware implementation.

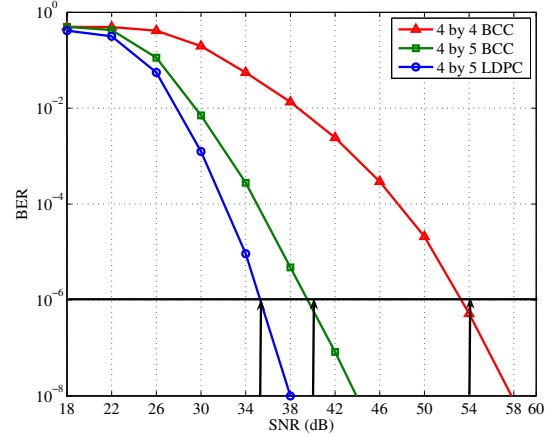


Fig. 8. Performance of the proposed Gigabit system under in-door channel model.

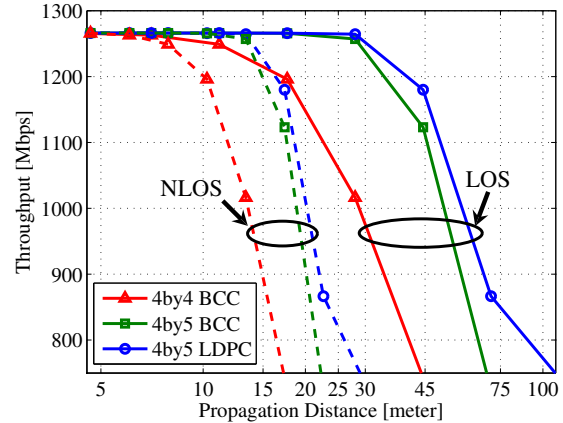


Fig. 9. Throughput v.s. propagation distance under in-door channel model.