Channel Sounding and PAPR Reduction in OFDM for WiMAX Using Software Defined Radio

B. Siva Kumar Reddy, B. Lakshmi

Abstract**—**This paper addresses the reduction of peak to average power ratio (PAPR) for the OFDM in Mobile-WiMAX physical layer (PHY) standard. In the process, the best achievable PAPR of 0 dB is found for the OFDM spectrum using phase modulation technique which avoids the nonlinear distortion. The performance of the WiMAX PHY standard is handled by the software defined radio (SDR) prototype in which GNU Radio and USRP N210 employed as software and hardware platforms respectively. It is also found that BER performance is shown for different coding and different modulation schemes. To empathize wireless propagation in specific environments, a sliding correlator wireless channel sounding system is designed by using SDR testbed.

Keywords**—**BER, Channel sounding, GNU Radio, OFDM/OFDMA, USRP N210.

I. Introduction

THERE is growing demand for broadband services world over. To achieve higher data rates to support broadband applications in a multipath environment, more effective modulation schemes in the form of multicarrier schemes are adopted. They present challenges of their own. For this issue, new technologies with high transmission abilities have been designed. The broadband wireless access has become the best way to meet escalating business demand for rapid internet connection and integrated "triple play" services. In addition to not only topographic but also technological limitations, wireless solution alternatives have been found. That is the very base of the WiMAX concept: a wireless transmission infrastructure that allows a fast deployment as well as low maintenance costs. Based on the IEEE 802.16-2004 standard, WiMAX allows for an efficient use of bandwidth in a wide frequency range, and can be used as a last mile solution for broadband internet access. This paper considers the high crest factor/PAPR of OFDM systems [1], [2]. The effects of a high PAPR are: the low power efficiency of the High Power Amplifier (HPA) and signal distortion caused by operation in the nonlinear region of the HPA.

Low power efficiency implies that the HPA must have a large dynamic range thus increasing cost of the system. Signal distortion induces a degradation of the BER. PAPR reduction techniques are used. The OFDM systems have some disadvantages, for example, the large Peak to Average Power Ratio (PAPR) [3] as well as high sensitivity to the synchronization errors [4]. An OFDM signal is the superposition of multiplexing of modulated sub-channel signals may produce an extreme peak signal with reference to the average signal level, which in turns to high PAPR. To deal with this PAPR problem, several techniques [5] have already been developed such as for instance coding and tone reservation [6], clipping/ filtering [7], [8], peak windowing [9], partial transmit sequence [10] and receiver correction algorithms [11] such as iterative decoding. As discussed in [5], each one of these distinctive methods has various quantities of effectiveness, and provides different sets of tradeoffs and also includes increased complexity, reduced spectral efficiency, and performance degradation.

- ➢ An alternative way of reducing the PAPR problem is focused around signal transformation i.e. Phase modulation [12]. This technique needs a signal transformation before amplification, then an inverse transformation at the receiver before demodulation. In this approach, the transformed signal has the best achievable PAPR, 0 dB and non-linear distortions are avoided. Hence, phase modulation is more preferable for PAPR reduction in OFDM based systems (Section 3). Here's how my idea compares to other people's approach.

Radio propagation measurement is necessary for creating propagation models and critical for conveying and preserving operational frameworks. There are various delay spread and path loss models depending on estimations made in a variety of environments, employing different carrier frequencies and antenna heights. In this paper, authors portray their first exertion in outlining a rapid wireless channel sounding system [13].

II. Software Defined Radio Experimental Setup

In this paper, we present the design of a Mobile-WiMAX PHY in an experimental setup shown in Fig. 1. It is built upon GNU Radio [14], an open source framework for software defined radio [15] and the universal software radio peripheral (USRP) [16], an open source transceiver platform. Compared to signal analyzers and oscilloscopes (a 3-GHz vector network analyzer (VNA) can cost US $20,000), our system is low cost. The cost of the proposed system is US$975, which enables large scale.

Advantages of such systems include their flexibility and reconfigurability since the devices can be dynamically programmed using software to control the hardware configurations. However, SDR operates without the knowledge of its environment and regardless of the changes in

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its surrounding radios and other contexts. Since the access to the radio spectrum is currently based on a fixed resource allocation regime, the radio spectrum scarcity becomes a major problem especially in the highly loaded allocated spectrum bands. This also obstructs the future development of new wireless applications. As a result, radios with the capability of dynamic spectrum access based on the environmental awareness and the ability of intelligent decision making are desirable.

Computers employed for the experimentation are built with: Intel Core i5 with core processors running with 2.45 GHz. Communication between computer and USRP N210 is provided by way of a Gigabit Ethernet cable that provides wideband signal transmission. USRP N210 board works as the interface between the analogue RF front end and the host PC where all the signal processing is performed in software. All high speed general purpose operations such as for instance decimation and interpolation, digital up and down conversion are performed on the FPGA (built in USRP). GNU Radio is most commonly used in academic and commercial environments to support wireless communications research and to carry out real-world radio systems. All signal processing blocks are written in Python and those are connected using C++ and both languages are communicated by SWIG (simplified wrapper and interface generator) interface compiler on GNU Radio. The combination of flexible hardware and open source software allow it to be a perfect platform for software radio development. In this paper all GNU schematics (signal flow graphs) are drawn as per mobile WiMAX specifications.

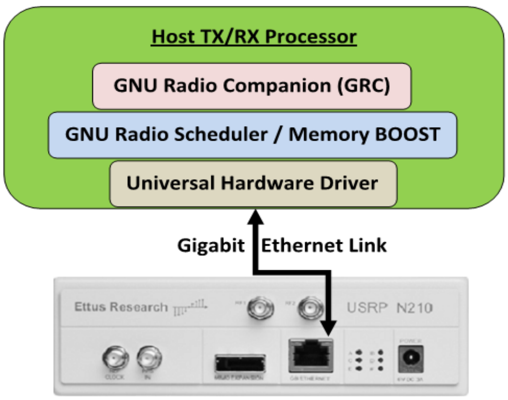

Fig. 1 SDR experimental setup with software and hardware

III. WiMAX Physical Layer Transmitter

WiMAX is an application of Ethernet and hence the whole standard is focused around the open systems interconnections (OSI) reference model. The physical layer is the lowest layer, in the connection of OSI model. It determines the frequency band, error correction techniques, the modulation/ multiplexing techniques, data rate, the synchronization between transmitter and receiver. The PHY layer is used to encode the binary digits that symbolize MAC frames into signals and to send and acquire these signals all over the communication media.

The WiMAX PHY layer transmitter section (shown in Fig. 2) incorporates various functional stages: (i) Forward Error Correction (FEC): comprises of scrambling, concatenated encoding and interleaving (ii) OFDM modulation and (iii) Phase modulation.

The data flow processing through physical layer is reported clearly in this section. A signal with 6 GHz frequency is captured from the environment by using USRP N210 with the assistance of CBX daughterboard. The USRP is communicated with GNU installed PC through a Gigabit Ethernet interface. The captured 6 GHz signal from file source is passed to the packet encoder ("Number of Bits per symbol" for M^{th} order QAM is $log_2(M)$). This block appends a bit sequence at the transmitter (tx) side, and finds the tx sequence on the receiver side to recognise the first bit of a packet.

Encoded data are sent through scrambler and it scrambles an input stream employing an LFSR (linear feedback shift register). This block impacts on the LSB only of the input data stream, i.e., on an "unpacked binary" stream, and generates the same format on its output. In coding hypothesis, concatenated codes structure a type of error-correcting codes which can be gained by linking inner and outer codes. During the paper, concatenated coding is structured like convolutional coding as inner code and Block coding as outer code. In this paper, Reed Muller (RM) and Reed Muller Golay (RMG) coders are considered as Block codes for concatenating. Scrambled data are passed to Reed-Muller encoder, in which only the first bit is used for in and output. Reed-Muller codes are listed as $RM(d, r)$, where d is the order of the code, r sets the size of code, and these are related as $n = 2^r$. RM codes are related to binary functions on field GF (2^r) over the elements 0, 1. $RM(1,r)$ codes are parity check codes of length $n=2^r$ rate and $R=(r+1)/N$ and minimum distance $d_{min}= n/2$. In RMG coder, assuming that "Number of bits to precode" is situated to zero, stand out Coset is likely to be applied. Any second request coset of a q-ary generalization of the first request Reed-Muller code might be zoned into Golay corresponding sets whose size registers just on a solitary parameter that is effectively computed from a chart connected with the coset. The RM/RMG encoded data are passed through CCSDS encoder block that executes convolutional encoding, applying the CCSDS standard polynomial ("Voyager"). Since the code rate is $1/2$, there will be 16 output symbols for every input byte. Next, data interleaving is utilized to expand e effectiveness of the FEC by dispersing burst errors inserted by the transmission channel over a long time. The interleaving is determined by way of a two-step permutation. First checks that adjoining coded bits are mapped onto non-adjacent subcarriers. The second permutation checks that adjoining coded bits are mapped alternately onto less or even more significant components of the constellation, thus eliminating long runs of low reliable bits. OFDM block produces OFDM symbols depending on the parameters FFT length, occupied tones, cyclic prefix (CP) length and modulation scheme specified on the block. The transmitted signal voltage to the antenna as a function of time throughout any OFDM symbol is defined as

$$S(t) = \left[e^{j2\pi f_c t} \sum_{k=N_{used}/2}^{N_{used}/2} a_k e^{j2\pi k \Delta f (t - T_g)} \right] \quad (1)$$

where t is the time elapsed of the OFDM symbol with $0 < t < T_s$, a_k is a complex number; the data to be transmitted on the carrier whose frequency o set index is k, during the OFDM symbol. It assigns a point in a QAM constellation, T_g is guard time, T_s is OFDM symbol duration, including the guard time, f is the carrier frequency spacing. Subsequently, carriers are cited by a carrier index; however, frequency offset index is required to construct the OFDMA signal.

The frequency offset index (k_{foi}) is defined in terms of its carrier index by following equation

$$k_{foi} = \begin{cases} k_{ci} - N_{used}/2, & \mathrm{k_{ci}} < N_{used}/2 \\ k_{ci} - N_{used}/2 + 1, & \mathrm{k_{ci}} \geq N_{used}/2 \end{cases} \quad (2)$$

where K_{ci} is carrier index and N_{used} (in this research work, $N=1024$, $N_{used}=896$) is the number of used carriers. These many number of individual subcarriers causes to high PAPR of generated OFDM signal. In this paper, the high PAPR problem is resolved using Phase modulation technique.

A. Constant Envelop-OFDM

A system with large PAPR makes the implementation of Digital-to-Analog Converter (DAC) and Analog-to- Digital Converter (ADC) to be extremely hard and also increase the complexity of RF amplifier design. To avoid high PAPR problem in OFDM based systems, phase modulation is performed on OFDM signals to get constant envelop-OFDM signals. This can be considered as a simple transformation of OFDM.

Let a constant-envelope phase varying signal be of the form

$$\sqrt{\frac{2E}{T}} \cos(2\pi f_c t + \phi(t)) \quad (3)$$

where T is the length of a basic signaling interval and E is the energy expended during this interval, as can be verified by integrating the square of (3) over a T-length interval. In each interval, an M-ary data symbol appears and the phase (t) follows some pattern in response to these symbols. f_c is the carrier frequency. The phase signal during the n[th] signaling interval $nT \leq t < (n+1)T$, can be written as

$$\phi(t) = \theta_i + 2\pi h C_N \sum_{k=1}^{N} d_{ik} q_k (t - iT_s), \mathrm{iT_s} \leq t \leq (i+1)T_s \quad (4)$$

The real data symbols d_{ik} modulate the orthogonal OFDM subcarriers $q_k(t)$ defined below. The normalizing constant $\mathrm{CN} = \sqrt{2/\sigma_i^2 N}$ ensures that the phase variance is independent of the number of OFDM subcarriers N, and σ_i^2 is the data variance with $\sigma_i^2 = 1$ for binary data. The key parameter, modulation index h that operates the spectral properties and the signal space properties of the CE-OFDM [12]. To assure a continuous phase at the symbol boundaries, the phase memory θi is employed in conjunction with a phase unwrapper to get better spectral efficiency at the receiver. Without loss of generality, full sine and cosine subcarriers divided by *1/Ts*, are utilized as a part of this paper to satisfy the subcarrier orthogonality necessity [12]:

$$q_k(t) = \begin{cases} \cos\left(\frac{2\pi k t}{T_s}\right), & \mathrm{k = 1,2,....N/2} \\ \sin\left(\frac{2\pi (k - N/2) t}{T_s}\right), & \mathrm{k = N/2+1,....N} \end{cases} \quad (5)$$

The subcarrier definition in (5) results from taking the IFFT of a Hermitian symmetric data vector.

Chunks to symbols block maps a stream of symbol indexes (unpacked bytes or shorts) to stream of oat or complex constellation points. Input is stream of short and output is a stream of float.

$$out[nD + k] = symbol\ \mathrm{table[in(n)D + k]}, \quad \mathrm{k = 0,1,....D - 1} \quad (6)$$

The combination of *gr_packed_to_unpacked_ XX* followed by *gr_chunks_to_symbols_XY* deals the general case of mapping from a stream of bytes or shorts into arbitrary oat or complex symbols.

Poly phase resampler block acknowledges a single complex stream in and yields a single complex stream out. Accordingly, it requires no additional glue to deal the input/output streams. This block is supplied to be predictable with the connection to the other PFB (poly phase filter banks) block. PFBs are a very powerful set of filtering tools that can efficiently perform numerous multi rate signal processing tasks. GNU Radio has a set of PFBs to be employed in all sorts of applications. This block accepts a signal stream and performs arbitrary resampling. The resampling rate can be any real number r. The resampling is acted by constructing N filters where N is the interpolation rate. Then D can be defined as, $D = floor(N/r)$. Using N and D, rational resampling is performed. Where N/D is a rational number close to the input rate r where $i + 1 = (i + D)\%N$. For the purpose of arbitrary rate, the interpolation between two points is required. For each value out, an output from the current filter, i, and the next filter $i + 1$ are considered and then linearly interpolate between the two based on the real resampling rate.

Multiply constant block is used to improve the amplitude level of the signal and it multiplies the signal with the constant value set by the user. Then, the signal is passed to USRP N210 and it is upconverted by RF hardware and transmitted into the air. The required parameters of USRP, like sampling frequency, clock rate, and number of channels, centre frequency and antenna gain are reconfigured by the user accordingly.

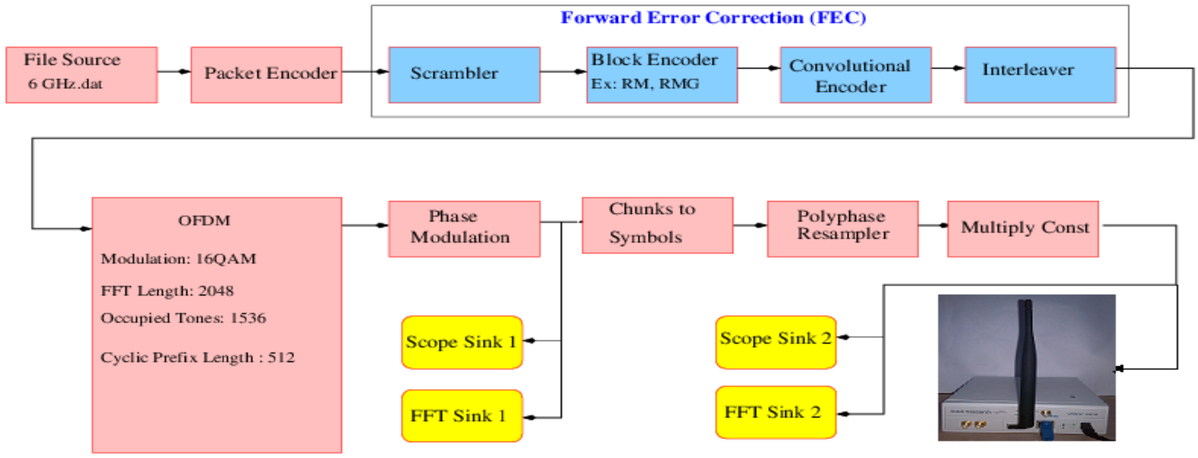

Fig. 2 WiMAX physical layer transmitter with USRP N210 as Sink

IV. Channel Sounding

The principle of the sliding correlator approach is to transmit a pseudo-noise (PN) sequence with a 60 nanosecond (ns) pulse duration and also to the receiver gets the wideband path loss and multipath delay profile. The transmitter and receiver paths of the sliding correlator system are shown as upper and lower parts of Fig. 3 respectively.

The transmitter sends *x*, a Galois linear feedback shift register (GLFSR) maximal length PN sequence of degree 10. The *x* can be analytically expressed as follows:

$$x(n) = \sum_{r} c[n - rN] \tag{7}$$

where, $N = 1023$ and c is a chip sequence of length 1023, c = [c_0,....c_{1022}]. Here $c_i \,\forall\, i \in [0,1022]$ takes the value of either +1 or -1.

Defining R_{cx} as the correlation output of c and x and using the properties of PN sequence,

$$R_{cx}[n] = \begin{cases} 1, & n = 0, N, \text{-}N, 2N, \text{-}2N, \ldots \\ -\dfrac{1}{N}, & \text{otherwise} \end{cases} \tag{8}$$

The input signal *x* from the GLFSR source is sent to a Root Raised Cosine (RRC) filter and then transmitted through USRP N210 sink in to air. The transmitted baseband signal can be expressed as:

$$x(t) = \sum_{n} p[t - nT_s] \tag{9}$$

where T_s is the period of the RRC generated pulse. The multipath channel's impulse response is defined as:

$$h(t) = \sum_{k=0}^{k-1} a_k \delta(t - \tau_k) \tag{10}$$

Here, *K* is the number of multipath in the channel, *k* is the complex gain of each multipath and *k* is the associated delay.

Let's assume $\tau_0 = 0$ since we concentrate on relative delay. In time domain, the received baseband signal is found by

$$y(t) = (x * h)(t) = \sum_{k=0}^{k-1} a_k x(t - \tau_k) \tag{11}$$

y(t) passed through receiver section as shown in Fig. 4. The time synchronizer block is used to get the proper phase of the RRC pulses. Proper timing synchronization results to the discrete received signal *y* as following,

$$\mathrm{y(j) = y[jT_s]} = \sum_{k=0}^{K-1} a_k \sum_{n} x[n] p[(j - n) - \tau_k] \tag{12}$$

The properties of the RRC filter suggest that $p[nT_s]=0$ if $n \neq 0$. Therefore

$$y(j) = \sum_{k=0}^{K-1} x(j - c_k) \tag{13}$$

Now, the correlator block generates

$$R_{cy}[n] = corr(c, y) = \sum_{k-0}^{k=1} a_k R_{cx}(n - c_k) \tag{14}$$

where, $R_{cx}=corr(c,x)$. Our approach probes the channel only at the integer multiples of the pulse duration, T_s. However, the multipath rays can arrive at other times, as well. Therefore, our observation at time T_s, $2T_s$,.....is actually an effect of multipath at nearby times.

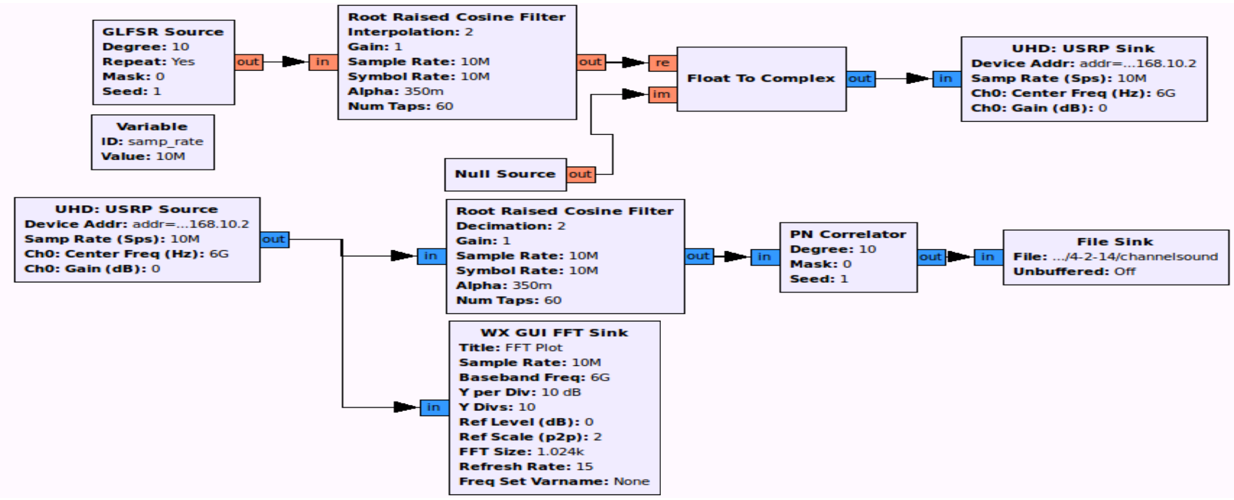

Fig. 3 GNU Schematic constructed for sliding correlator Channel Sounding system

V. Experimental Results

The experimental setup used for data transmitting and receiving is demonstrated in Fig. 1. The complete WiMAX physical layer transmitter (shown in Fig. 2) is implemented in

GNU Radio with hardware support of USRP N210 as per Mobile WiMAX specifications are noted in Table I. In presented Fig. 2, the scope sinks and FFT sinks are used to observe the output in time domain and frequency domain respectively. In general, the OFDM modulated signal has high peaks corresponds to average of the signal as depicted in Fig. 4 and transmitted OFDM spectrum is shown in Fig. 5. These high peaks are appeared in OFDM signal due to superimposition of high amplitude individual subcarriers, which leads to the power inefficiency in the RF section of the transmitter. To overcome this problem several techniques are proposed in literature but they also have their own advantages and disadvantages (The authors given complete information in [5]).

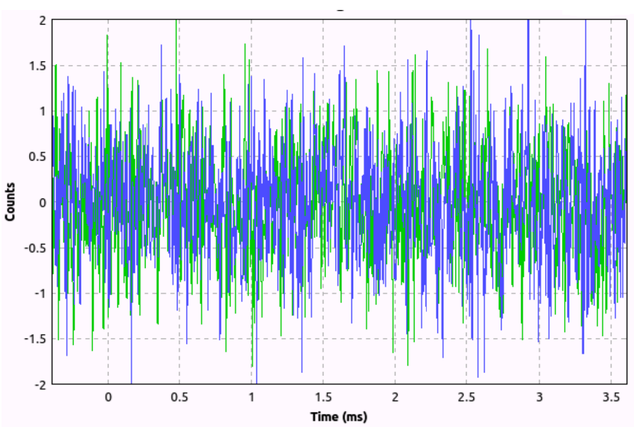

Fig. 4 The generated OFDM signal (having high PAPR)

In this paper, an alternate solution presented to overcome high PAPR problem is to clip OFDM signal before amplification to give lowest PAPR value. As shown in Fig. 2, a proposed PAPR reduction technique i.e. Phase modulation is performed on generated OFDM signal. The output phase modulated signal is depicted in Fig. 6 and it can be concluded that phase modulated signal has almost zero PAPR. Hence, PAPR problem in OFDM is resolved by the proposed phase modulation technique.

TABLE I
Experimental Parameters Defined

Parameters	Values
Number of Subchannels	1024
Number of pilots (P=N/8)	128
Number of data subcarriers (N-P)	896
Pilot position interval	8
Sample rate	11.111M
Carrier frequency	6 GHz
Digital Modulations	BPSK, QPSK, 16QAM, 64QAM and 256QAM

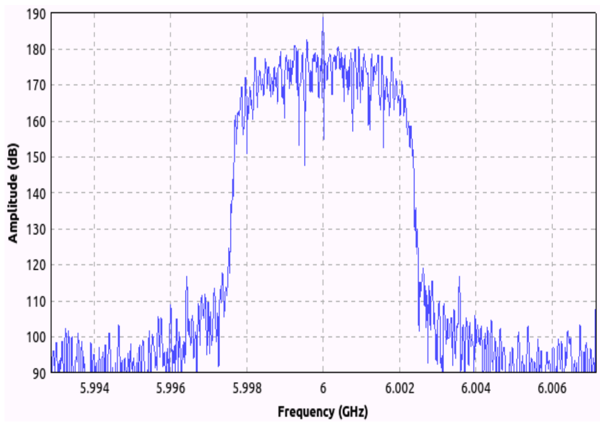

Fig. 5 The transmitted OFDM spectrum by USRP N210

The transmitted spectrum is passed through a channel model block (used as an AWGN channel in GNU Radio) and analyzed by varying channel noise levels. Received spectrum with zero channel noise is shown in Fig. 7 and with 100m channel noise is presented in Fig. 8.

Received spectrum by USRP N210 (worked as source) in lab environment is demonstrated in Fig. 10. From Figs. 7-9, it can be concluded that increment in noise results to high out of band radiation. Hence, to get specified OFDM spectrum, channel noise should be less. The received constellation diagrams with channel model block and lab environment are shown in Figs. 10 and 11 respectively.

The estimated SNR value of received signal through channel model block is 11.37148 units and with USRP N210 is 0.808060 (shown in Fig. 12). SNR values can be improved by using multiply constant block in GNU radio.

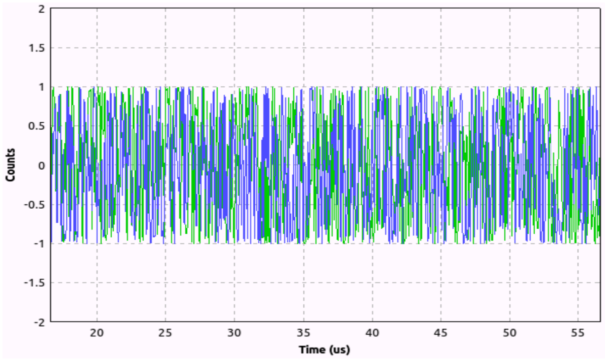

Fig. 6 Phase modulated OFDM signal having Constant Envelop

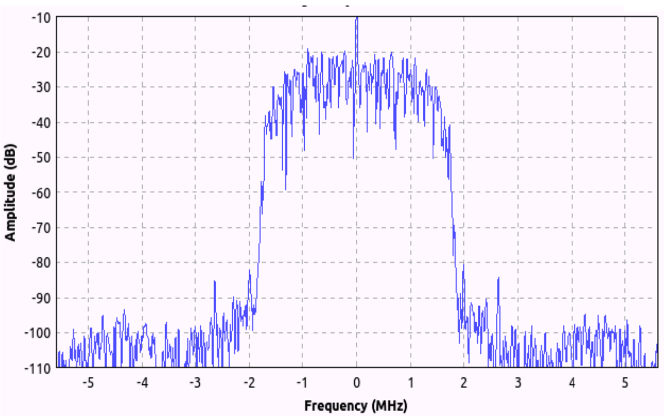

Fig. 7 Received OFDM spectrum when the channel noise is zero

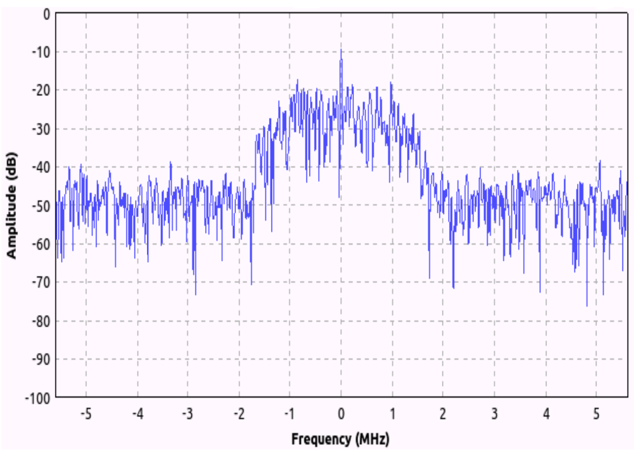

Fig. 8 Received OFDM spectrum when the channel noise is 100m dB

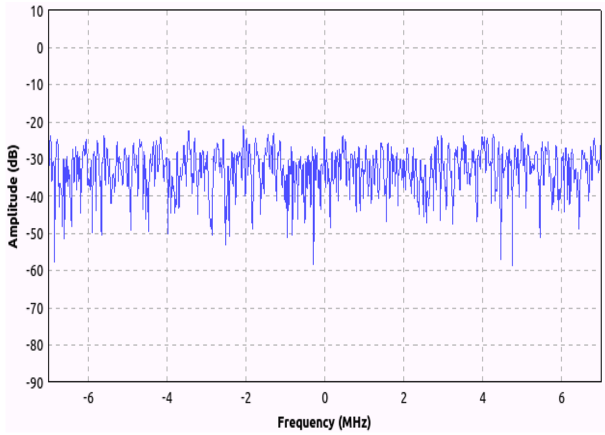

Fig. 9 Received OFDM spectrum with USRP N210 as Source

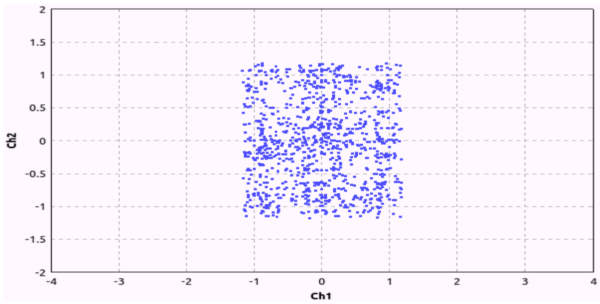

Fig. 10 Received Constellation mapping over channel model block

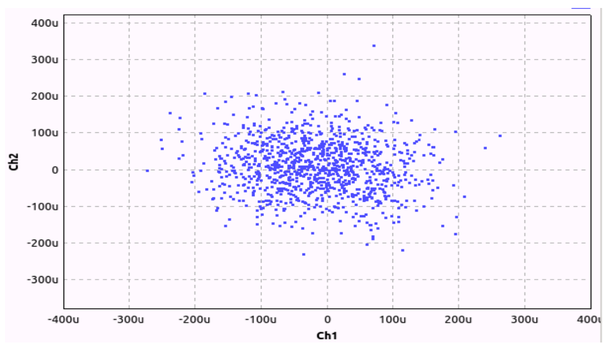

Fig. 11 Received Constellation mapping by USRP N210 over air

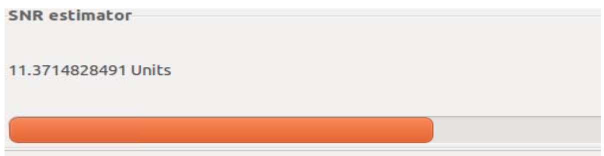

Fig. 12 The estimated SNR value on Number sink

Forward error correction (FEC) is involved in WiMAX PHY for transmitting redundant data to enable receiver to recover from errors and retransmission is not required. As indicated in Fig. 2, the incoming signal from the file source is channel coded by the scrambler, RMG encoder and interleaving separately and passed through the OFDM modulator to produce OFDM symbols. Afterwards OFDM symbols passed through the multiply constant block and channel model. The channel model is reconfigured by changing frequency offset and noise parameters. The generated OFDM symbols are demodulated, de-interleaved, decoded and descrambled respectively. The RMG coder in Fig. 13 is replaced with convolutional coder, RM coder and results are noted in Table II by varying modulation schemes and bits per byte in BER block. From the Table II, it can be concluded that BER performance has been improved as number of bits per byte increased. The error difference between with and without convolutional coder is more than RMG coder and RM coders. Hence, it can be concluded that convolutional coder is better than RM and RMG coders in channel coded systems. The coded and uncoded BER values of different coders are noted in Table III. The coded version shows better BER performance than uncoded version.

The indoor channel estimations were gotten utilizing the sliding correlator framework. Three transmitters were set up in distinctive parts of a wing and the recipient travelled through the wing. The multipath delay profile and the wideband path loss of each transmitter were stored for each location. It is evident from Figs. 14 and 15 that as the receiver moves away from the transmitter, the path loss becomes higher. Transmitter 2 will require less power to cover the whole wing compared to transmitter 1 as it is placed in the central location of the wing, resulting in less mean path loss from transmitter 2.

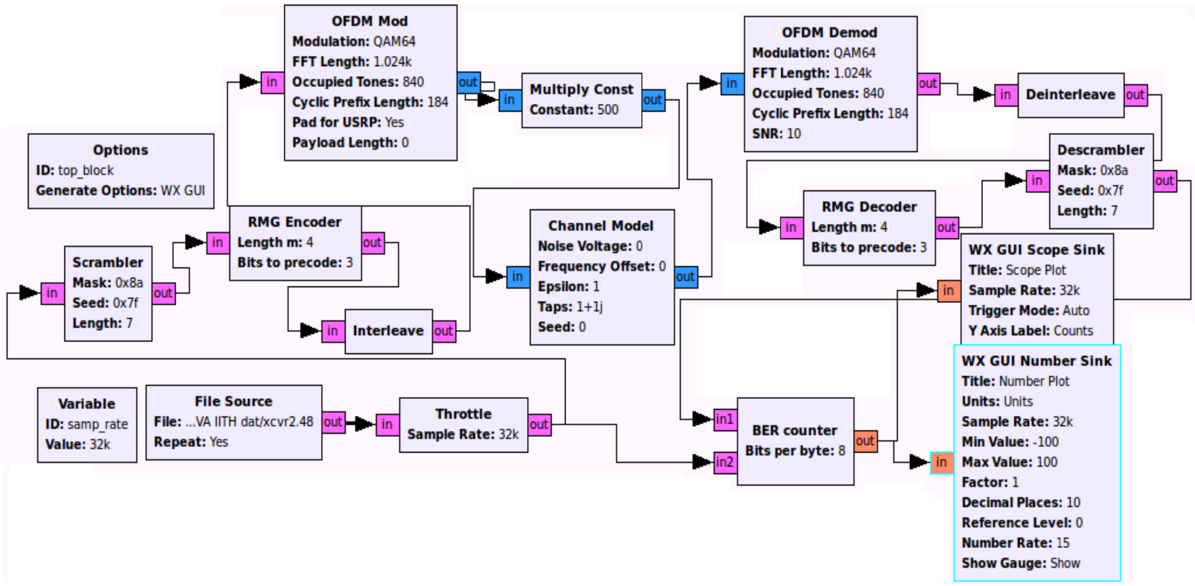

Fig. 13 GNU Schematic for BER of 64QAM-OFDM with RMG coder channel coding

TABLE II
RESULTS OBTAINED FOR BER PERFORMANCE WITH CHANNEL CODING OVER AWGN CHANNEL

Coder	Mod Scheme	Bits per byte	BER value
RMG Coder	BPSK	1	0.5001434684
RMG Coder	BPSK	4	0.4068774879
RMG Coder	BPSK	8	0.3892548680
		Average	0.43209194
RMG Coder	QAM64	1	0.5004814267
RMG Coder	QAM64	4	0.4189103246
RMG Coder	QAM64	8	0.4092760682
		Average	0.44288927
CCSDS Coder	BPSK	1	0.4585958719
CCSDS Coder	BPSK	4	0.5740551949
CCSDS Coder	BPSK	8	0.5869382620
		Average	0.53986311
CCSDS Coder	QAM64	1	0.4601847827
CCSDS Coder	QAM64	4	0.5721712112
CCSDS Coder	QAM64	8	0.5870822072
		Average	0.53981274
RM Coder	BPSK	1	0.5000710487
RM Coder	BPSK	4	0.4210021496
RM Coder	BPSK	8	0.3852755427
		Average	0.43544958
RM Coder	QAM64	1	0.5001475811
RM Coder	QAM64	4	0.42152449419
RM Coder	QAM64	8	0.4092907906
		Average	0.44365444

TABLE III
BER PERFORMANCE FOR UNCODED AND CODED SIGNALS

Coder	Bits per Byte	BER for Coded	BER for Uncoded
RMG Coder	1	0.3480649889	0.0099786464
RMG Coder	4	0.2802865505	0.0024868969
RMG Coder	8	0.0010421536	0.0002434485
CCSDS Coder	1	0.5534156561	0.0099835679
CCSDS Coder	4	0.5438777804	0.0024312227
CCSDS Coder	8	0.5416055322	0.0012434485
RM Coder	1	0.3504898846	0.0105615864
RM Coder	4	0.3020897806	0.0105615864
RM Coder	8	0.0202364717	0.0013201983

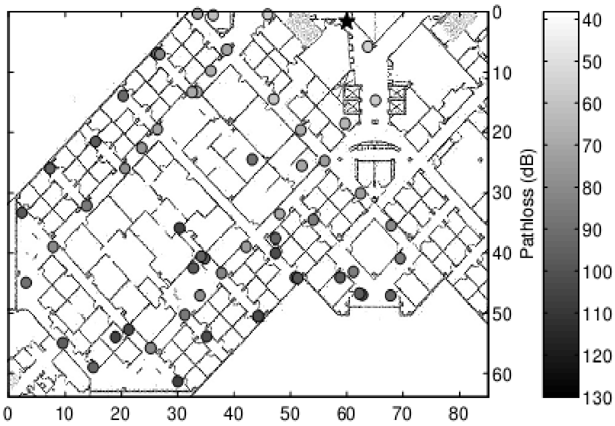

Fig. 14 Path loss data for indoor transmitter 1

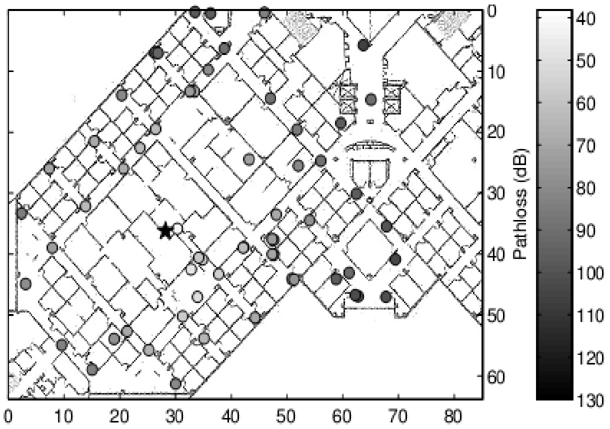

Fig. 15 Path loss data for indoor transmitter 2

VI. CONCLUSION

In this paper a transformation technique that eliminates the PAPR problem associated OFDM is developed for Mobile-WiMAX PHY standard. The phase modulation transform results in 0 dB PAPR constant envelope signals ideally suited for nonlinear, efficient amplification. Experimental results concluded that BER performance is improved as the number of bits per byte increases, and BER performance is better for convolutional coder compared to RM/RMG coders. Further, the outcomes of WiMAX PHY Transmitter are verified experimentally on SDR prototype with the support of GNU radio and USRP N210.

Future work should address MIMO-OFDM with all proposed features and implementation on a single SDR test bed. Different channel estimation methods will be applied and analysed for Channel Estate Information, which is a key factor for the design of Adaptive Modulation & Coding (AMC) techniques.

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