A Novel QAM Technique for High Order QAM Signaling

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Abstract. *The paper proposes a novel spread quadrature amplitude modulation (S-QAM) technique with high SNR improvement for high-order QAM channels. Simulated and experimental bit error rate (BER) performance analyses of the proposed technique in blind and non-blind equalizers are obtained by using single carrier (SC) WiMAX (IEEE 802.16-2004) radio. Instead of using any one particular type of channel profile, this study concentrates on true frequency selective Rayleigh fading channels in the real-time WiMAX radio environment around 3.5 GHz. The Constant Modulus Algorithm (CMA) blind equalizer has been compared with the popular non-blind equalizers, Recursive Least Squares (RLS) and Least Mean Squares (LMS) algorithm, as benchmarks. It has been proven in experimental and simulated channels that CMA blind equalizer, using the proposed technique, can be considered as a low complexity, spectrum efficient and high performance time domain equalizations to be embedded in a transceiver for the next generation communications. Furthermore the proposed technique has also reduced approximately till 5 dB and 7.5 dB performance differences between non-blind and blind equalizers for 16-QAM and 64-QAM, respectively. The simulation results have demonstrated that the simulated and experimental studies of the proposed technique are compatible with each other and extremely satisfying.*

Keywords

Experimental BER, adaptive blind training, WiMAX (IEEE 802.16-2004), constant modulus algorithm, S-QAM.

1. Introduction

IEEE 802.16 working group was set up in 1999 to develop a new standard for broadband wireless access (BWA) and published the first IEEE 802.16 standard in October 2001. In October 2004, the new standard 802.16-2004 was published, which is actually an amalgamation of 802.16 and 802.16a. In the first phase of the standard, Single-Carrier (SC) for 11-66 GHz and Multi-Carrier (MC) transmissions for sub-11 GHz frequency regions were considered for fixed wireless access. By the publications of IEEE 802.16-2004 [1], its applications have been extended to single carrier transmission for sub-11 GHz systems. Recently, the 802.16e standard was also ratified in December 2005 by allowing the upgrade from fixed BWA systems to mobile service provision up to vehicular speeds for sub-11 GHz systems [2].

Applications such as video and audio streaming, online gaming, video conferencing, Voice over IP (VoIP) and File Transfer Protocol (FTP) demand a wide range of QoS requirements like bandwidth and delay. Existing wireless technologies that can satisfy the requirements of heterogeneous traffic are very costly to deploy in rural areas and “last mile” access. The WiMAX radio provides an affordable alternative for wireless broadband access supporting a multiplicity of applications.

However, like in any wireless system, signal distortion due to channel fading, noise, inter symbol interference (ISI), carrier frequency offset (CFO) and Doppler can limit the overall transmission data rate and coverage. To minimize the degradation in system performance caused by the channel, channel estimation and equalization techniques must be performed to remove the effects of the channel. One of the best ways to mitigate these effects is to use blind or non-blind equalization techniques.

Blind equalization techniques have several advantages over training sequence based equalizers, i.e. avoiding of the training sequence simplifies the receiver architecture and saves the time duration of training period. During the experimental studies it has been observed that blind techniques significantly reduce the ISI without interfering the content of the incoming data. This feature can easily be exploited in a cooperative communication and repeater design.

Because of their low error performance, blind receivers have not been used in commercial applications where short packet duration and low latencies are essential and significant received signal-to-noise ratio (SNR) is available. When the orthogonal frequency division multiplexing (OFDM) was introduced to communication industry at late 90s, frequency domain equalization found quite simple and more robust to symbol synchronization by having a certain amount of cyclic prefix duration. Thus, the blind channel estimation and data recovery can also well be issued in Coded-OFDM using the pilot tones and cyclic prefix and therefore Digital Video Broadcasting-Terrestrial (DVB-T) standard does not promote the use of training preamble [3].

On the other hand, due to longer symbol duration, the OFDM systems have a limited mobility support, and also require a higher level of SNR and very expensive linear power amplifiers when it is compared with single carrier transmission. During the experimental studies it is observed that single carrier transmission even with high level of modulation depths, i.e. 16-QAM and 64-QAM, the data recovery is possible in a SNR level of as low as 8-10 dB using either non-blind or blind equalization. Therefore, it is observed that without involving with any deduction and implementation limitations on considered wireless standards, a BER performance evaluation of a blind equalization for the SC receivers is possible. The required data polarity correction and alignment can be made by using a very short code, i.e. Constant Amplitude Zero Autocorrelation Code (CAZAC) sequence, during the experiments.

However, without data aided correction blind techniques cannot recover true polarity of incoming signal [4], [5] unless someone use differential modulation techniques. A differential modulation cancels the polarity ambiguity price paid for 3 dB performance degradation when comparing with those using a coherent modulation. In experiments a CAZAC sequence is used to obtain data polarity, because it is required for BER calculations of blind techniques [6]. There are several techniques recovering data polarity as it is explained in [7].

One of the first experimental studies has been done by Wang in an underwater acoustic communication channel considering a decision feedback equalizer (DFE) to cancel ISI and their performance measures are limited by mean square error (MSE) comparisons and signal constellations demo [8]. Another work by Tanada demonstrates a blind equalization scheme based on parallel maximum likelihood sequence estimation (MLSE) method, where the work presents BER performances when the data is coded by Reed-Solomon technique using pi/4 shifted DQPSK with 6.25 kHz channel spacing in 400 MHz band [9]. The most inspiring work on the blind training –as far as author's concern- is the publication by Labat et al. [10], titled "… Can You Skip the Training Period?", which also triggered my work in the area of blind training [11] when the paper was published in 1998.

So far, blind equalizations were not considered for commercial and high performance applications. However, this study proves that for high level of modulations, i.e. 16-QAM and 64-QAM, the performance of blind equalizer can challenge the performances of those using non-blind equalizations. For the comparisons this study evaluates the BER performance analyses of blind and non-blind equalizations in a real time SC WiMAX radio experimental system and as well as in simulations of the system for the frequency selective Rayleigh fading channels.

The rest of the paper is organized as follows: Section 2 introduces the proposed spread quadrature amplitude modulation (S-QAM) technique. Section 3 explains the CMA based adaptive blind equalizer trainings which are as far as author's knowledge the best blind training techniques. Section 4 presents the experimental system and measurement conditions. Section 5 evaluates the obtained BER performances to verify the feasibility and robustness of the proposed technique and finally, the paper is concluded in section 6.

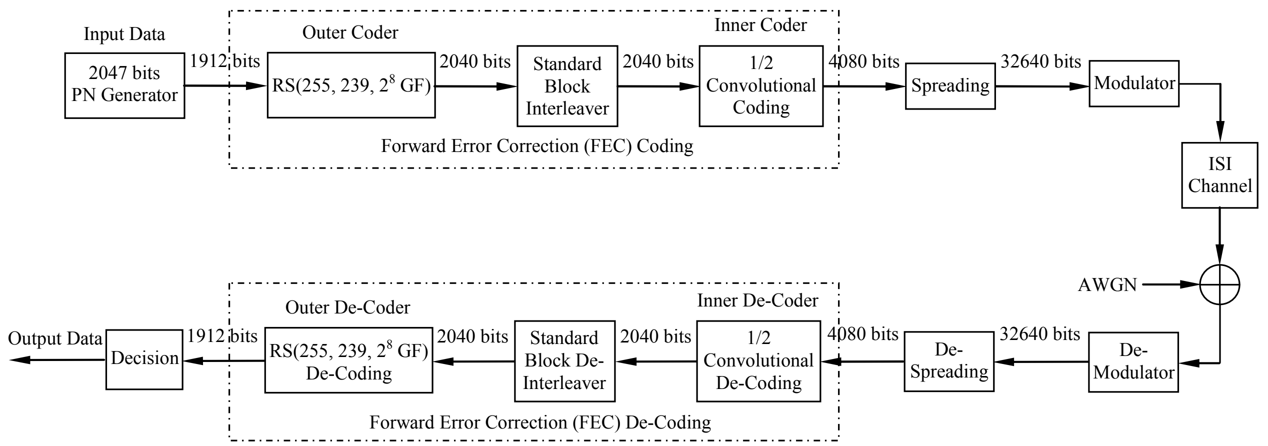


Fig. 1. The WiMAX model in single carrier wireless communication systems.

2. The Proposed Spread QAM Technique for High-Order QAM Modulation

Fig. 1 depicts the physical layer components of a WiMAX receiver and transmitter for single carrier communications [1], where Reed Solomon and convolutional coding for forward error correction (FEC) coding, and soft output Viterbi decoding for the FEC decoding are used. A 1912 (=8x239) bits of a 2047 bits PN sequence is used as the payload sequence, and coded by the (255, 239, GF 2^8) Reed-Solomon coding for the outer code (page 357, [1]), block interleaved (page 258, [1]) and then coded by the binary convolution code (CC) with the rate of 1/2 as an inner code (pages 258-259, [1]). The bit randomizer is also applied to the raw data. The output of the FEC encoder is spread and then modulated with one of the desired modulation types. The resulting modulated signal is transmitted over a multipath channel and corrupted by white Gaussian noise.

The receiver converts the received signal in the baseband and demodulates. The demodulated signal is despread and fed to FEC decoding block. In order to decode despread signals, soft output Viterbi algorithm in FEC decoding is used. Finally, output data is obtained at the output of decision block and then the desired performance comparisons are also performed.

The input to a spreading unit is a serial stream of bits derived from the FEC encoder output. Fig. 2 shows the generation of spreading data with a spreading factor of F_s [1]. Each input bit should be held for F_s symbol clocks as it is XORed with F_s consecutive outputs of a PN sequence generator operating at the symbol rate. The spreading factors from the set $F_s = 2^n$, $0 \le n \le n_{max}$, where $n_{max} = 3$ (for downlink), 4 (for uplink) can be used [1]. n_{max} parameter is equal to 3 in this study. The XOR output, spread bits, can be mapped to any desired modulation type.

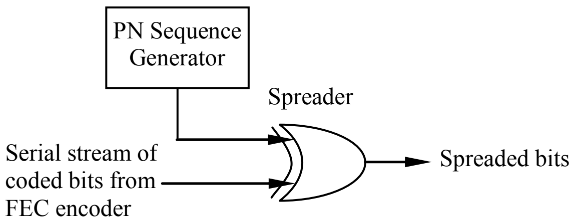


Fig. 2. The spreading process.

The spreading PN sequence generator should be constructed from the linear feedback shift register (LFSR) illustrated in Fig. 3. The characteristic polynomial for this LFSR is $1+x^{21}+x^{22}$. The PN sequence generator can be preset at the beginning of a spreading unit allocation with one of the seeds listed in Tab. 184 (page 378, [1]). The burst profile setting for spreading is used to select the seed to be used.

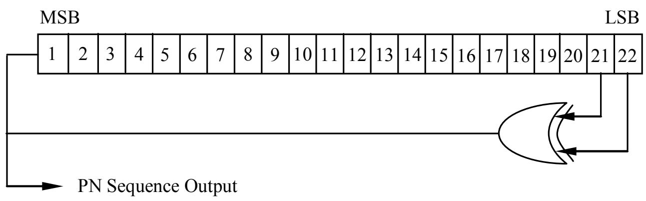


Fig. 3. Spreading PN sequence generator.

Seed 0 is the default setting. Selection of any seed other than Seed 0 requires use of the burst profile encoding for spreading parameters.

The received signal at the output of the spreader, as can be seen from Fig 2, Tx_{spread} is given by

$$Tx_{spread}(i*F_s+j)=\sum_{j=0}^{F_s} PN_{spread}(i*F_s+j)\oplus D_{coded}(i) \quad (1)$$

where, PN_{spread} is the output of the spreading PN sequence generator, D_{coded} is the output of the FEC encoder, F_s is the spreading factor, $\oplus$ is the logical XOR operator and i is the time index. The received signal at the output of the despreader, $Rx_{despread}$, is given by

$$Rx_{despread}(i)=\sum_{j=0}^{F_s} PN_{spread}(i*F_s+j)\hat{x}(i*F_s+j) \quad (2)$$

where $\hat{x}$ is the output of the equalizer (blind or non-blind).

When the output of the spreading block is mapped to BPSK symbols, Spread-BPSK (S-BPSK) modulation is obtained as in [1]. Fig. 4 compares the BER performances of two adaptive algorithms (LMS and RLS) for BPSK and S-BPSK modulations. It can be seen from Fig. 4 the BER performance obtained using the S-BPSK technique outperforms the conventional BPSK and provides high SNR improvement of approximately 10 dB for a BER value of 1E-3.

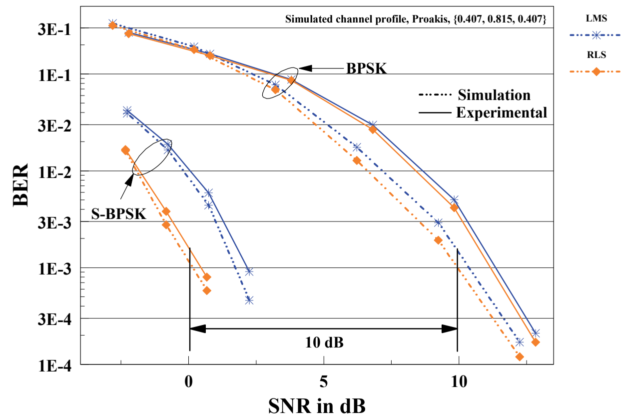


Fig. 4. BER performances of non-blind equalizers for experimental and simulated channels using BPSK and S-BPSK modulations.

The greatest novelty of this study, inspired by S-BPSK, is that spread bits are mapped to 16-QAM and 64-QAM symbols. Thus, the proposed technique is obtained called as Spread-16-QAM (S-16-QAM) and Spread-64-QAM (S-64-QAM). Although there exist some works on S-BPSK technique [12, 13], there are no any published work on spread QAM (S-QAM) and its higher degrees (S-16-QAM, S-64-QAM) for blind and non-blind equalizations, as far as author's concern. Thus, this contribution, inspired by S-BPSK, investigates the proposed technique in the context of high-order QAM blind and non-blind equalizations.

3. Blind Equalization

The baseband model of a digital communication channel can be characterized by a symbol-spaced Finite Impulse Response (FIR) filter and additive white Gaussian noise (AWGN) source. The received signal at the output of a wideband channel, $v(k)$, is given by

$$v(k) = \sum_{i=0}^{M} h(i)x(k-i) + \eta(k) \quad (3)$$

where, $x(k)$ is the transmit data sequence, assumed to be independent identically distributed (iid), $h(i)$ is the ith tap coefficients of the tapped-delay-line filter model of a channel, M+1 is the tap number of the channel, $\eta(k)$ is the iid AWGN component with zero-mean and variance $\sigma^2_{\eta(k)}$ and k is the time index. It should be mentioned here that in this study no offset frequency is considered and the samples are in symbol spaced, so the main equations hold for the case of strict carrier and symbol synchronizations. A linear transversal equalizer (LTE) or a soft decision data directed decision feedback equalizer (DFE) is used for blind equalizations. Because, firstly the irreducible phase error occurs [4] and secondly incorrect hard decision drives a DFE training to an instable region (called error propagation) which occurs more often in a blind training. Therefore, a LTE is used in this study and its output, $\hat{x}(k)$, is calculated by

$$\hat{x}(k) = \sum_{i=0}^{N} w(i)v(k-i) \quad (4)$$

where N+1 is the tap number of LTE, $w(i)$'s are the LTE coefficients. For an ordinary training case, the error function of an equalizer is calculated by $\varepsilon(k) = x(k - L_{\text{offset}}) - x(k)$, where a training sequence, known by both end of transmission, transmitter and receiver, is available, where the number indicated by L_{offset} is attained for the adjustment of the centre tap of equalizer filter. However, if a training sequence is not issued in the transmission, one of the blind algorithms has to be applied to recover the transmitted data. For the adaptive blind training, the CMA algorithm is one of the best training techniques that use the cost function

$$\bar{J}_{\text{CMA}}(W) = E\{(|\hat{x}(k)|^2 - \Delta_2)^2\} \quad (5)$$

where W is the equalizer coefficient vector as $W = [w(0), w(1),\ldots,w(N)]^T$ ($[.]^T$ indicates the transpose of the matrix [.]), $\hat{x}(k)$ is the kth estimate of the equalizer filter given by (4), $E\{\}$ is the expectation operator and Δ_2 is a real positive constant calculated by $\Delta_2 = E\{|x(k)|^4\} / E\{|x(k)|^2\}$ using the transmit data.

It should be mentioned here that if W_{opt} is obtained verifying the cost function (5), $\bar{J}_{\text{CMA}}(W_{\text{opt}})$, it produces to same results as $W_x = \exp(j\phi)W_{\text{opt}}$, $0 \le \phi \le 2\pi$ so, the algorithm always produces a phase error which cannot be corrected by the CMA criterion (see [4], [5]), therefore the phase of estimated symbol is not processed by a hard detector directly. The defined problem has to be solved by further operations using schemes either a differential modulation or a phase compensating coding techniques.

The error function to verify CMA criterion is

$$\hat{\varepsilon}(k) = \hat{x}(k)(\Delta_2 - |\hat{x}(k)|^2) \quad (6)$$

and similar to stochastic gradient algorithm the adaptation of W according to [14], [15] is given by

$$w(i+1) = w(i) + \mu\hat{\varepsilon}(k)v^*(k-i), \quad i = 0,1,\ldots N \quad (7)$$

where μ is the step size parameter of CMA, $\hat{\varepsilon}(k)$ is the kth estimate of error function using CMA criterion and $v^*(k-i)$ is the complex conjugate of $v(k-i)$.

4. The Experimental System

Fig. 5 shows a block diagram of the transmitter for the experimental WiMAX radio, where the baseband signal preparation is done in a PC and uploaded to a vector signal generator, E4438C by Agilent (0-6 GHz). In transmission a Raised-Cosine filter with a cut off rate of 0.35 is employed for the baseband filtering. A linear power amplifier with 22 dBm and 35 dBm of the IP1 and IP3 powers respectively, HMC409LP4 by Hittite Microwave, is used before sending signal to the antenna. Two types of antennas, a biquad directional antenna with approximately 9 dBi of gain and 60 degree of aperture angle, and an Omni directional antenna with the gain of less than 1 dBi (as it is measured during experiments) are used during the experimental tests.

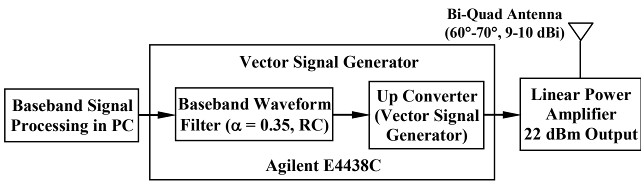
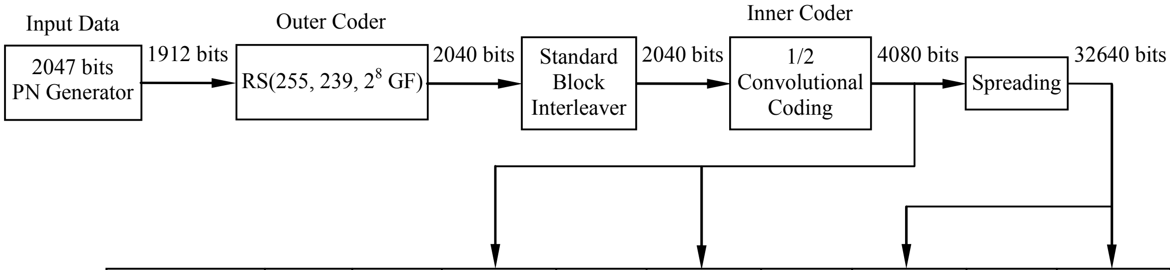


Fig. 5. Transmitter of Experimental Radio Set.

A program written in C code in a PC at the transmitter side prepares a long experimental data sequence as it is given in Fig. 6. The prepared sequence contains one 255 symbols of QPSK modulated PN sequence following by 4 sub-sequences each representing to a burst set formats, 16-QAM and 64-QAM, of the standard [1] and the proposed S-16-QAM and S-64-QAM. A CAZAC sequence with the length of 64 symbols, which is described in [1], page 379 is used as unique word and repeated three times to create a burst set preamble at the beginning of each sub-sequences. The standard does not support 1/2 CC for 64 QAM modulation (Table 175, [1]), however in order to compose the results comparable to each other 1/2 CC is employed as an inner coder for 64 QAM modulation. The same coded data sequence is spread and modulated by every modulation type and placed into the attained burst set. The burst made of as in Figu. 6 is stored to signal generator and transmitted repeatedly with the symbol rate of 20.48 MSample/s. The use of combined data packet, as in Fig. 6, provides a comparative analysis for the modulation types.

The data packet has a complex PN sequence to detect the beginning of the data packet, since there is no feedback link in the experimental test bench to get the starting time of transmission. Three 64 symbols of CAZAC sequences



Modulation	Comp. PN	UWs	16-QAM	UWs	64-QAM	UWs	S-16-QAM	UWs	S-64-QAM
Number of Samples	255	3X64	1020	3X64	680	3X64	8160	3X64	5440

Fig. 6. Transmit data packet structure.

are used for non-blind trainings and for the phase corrections of blind trainings. The transmitter repeats the transmission of the same packets for each measurement points and however the payload data has changed for every new packet in simulations.

An essential part of data packet preparations is that a fixed power is available at the transmitter. So, data subsequences are normalized to unit value for all modulation types. Instead of using SNR per bit (as it is used to compare the error performances of modulation types in Ziemer's [16] and Proakis's books [17]) the performances of modulation types are compared using the available SNR at the receiver. Therefore, comparisons of obtained throughput levels are more rational and considered to be comprehensive under the simulation and as well as experimental conditions.

The receiver of the experimental radio is given in Fig. 7, where a vector signal analyzer (WCA380 from Tektronix, 0-8 GHz.) is used for RF radio receiver. The baseband of the received signal is sampled by the sampling frequency of 20.48 Msamples/s and stored by the analyzer with length of 100 experimental data sequences given in Fig. 6. This sampled long sequence is downloaded to a PC for the baseband signal processing and BER calculations. A receiver algorithm, involving with synchronization, equalization and decoding, is implemented in a program written in C programming language.

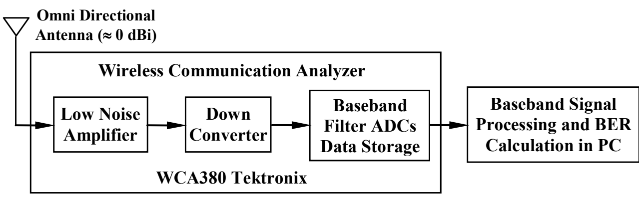


Fig. 7. Receiver of experimental radio set.

The receiver algorithm executes a synchronization using the PN sequence at the beginning of the long sampled sequence, and calculates the signal power for estimating the signal-to-noise ratio (SNR). In order to calculate the noise power of the channel, the RF signal turned off at the transmitter then the receiver samples the noise sequence which is the output of the demodulator at the receiver while the receiver is kept tuned to 3.45 GHz RF frequency. The sampled noise sequence is identical to noise on the received data which is sampled when the modulated signal is transmitted. By omitting the noise and dividing the power of the received signal by the power of the sampled noise sequence gives an estimate of the SNR.

During the experimental studies reference clocks of two device, transmitter and receiver, are synchronized to each other by a coaxial cable. Therefore, there is no offset frequency effect at the reception. Very small channel variations were occurred in the channel when people were passing between antennas, and there was no limitation on the environmental conditions in the test area.

In the receiver, firstly data normalization is made by setting the received signal plus noise power to a unit value by a separate process for every sub-sequence. The normalization is one of the essential step when implementing a channel estimation and equalization algorithms, since the transmit data also needs to be considered as having unit energy.

Two conventional adaptive non-blind training algorithms, LMS and RLS, and one adaptive blind training algorithm, conventional CMA, are used for non-blind and blind equalizations. The error count is made after equalization for raw BER calculation of the system without coding gain. The equalized data is decoded by inner decoder, deinterleaved and decoded again by the outer decoder. The final error count is obtained over decoded data in order to obtain the overall BER performance of the system. Here, the error rates of all modulation types are calculated at one measurement point which is single trail of the channel. Therefore, in order to obtain an average value as it is done in Monte-Carlo simulation programs an averaging process is required which is done as explained in the following section.

BER performance results are obtained on the grid shown in Fig. 8 and the effects of the modulation types are compared using the same channel. Three essential criteria are found to be useful for this kind of trails:

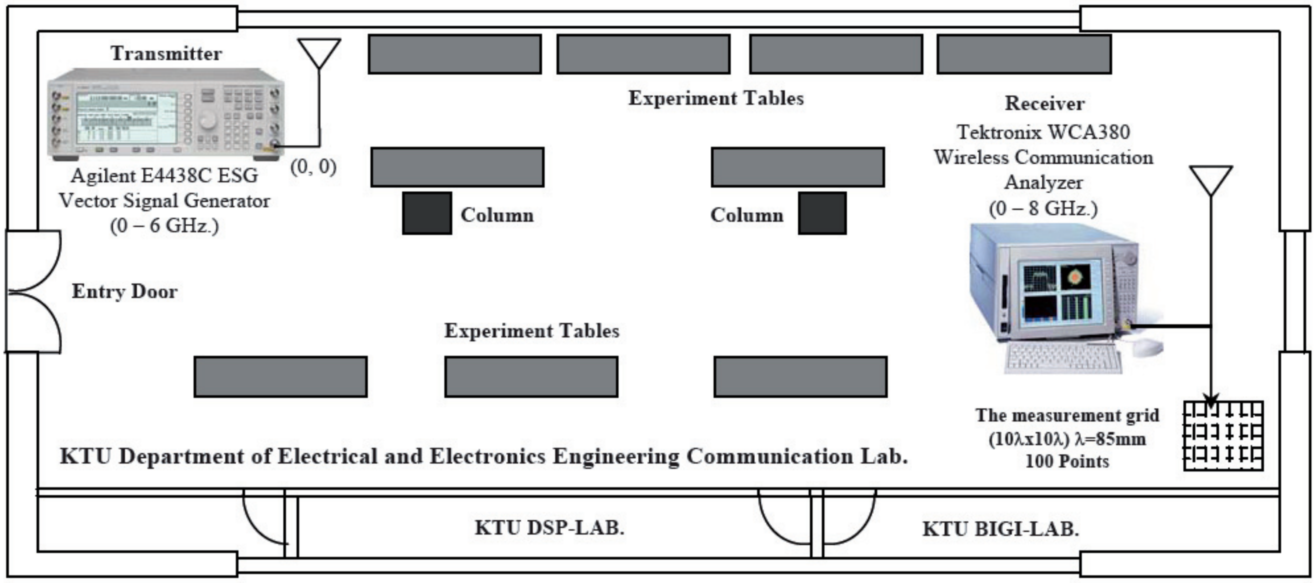


Fig. 8. The test area and measurement grid with 100 points.

1. The grid should be far enough from the transmitter that there would be no SNR differences between measurement points,
2. There should be enough number of measurement points that an individual channel could not make any big difference over averaged performance,
3. The measurement points should be apart from each other that the channel profiles would not correlate and in order to keep SNR margin between measurement points smaller the number of measurement points should also be limited.

Therefore a measurement grid of 100 points and minimum distance between measurements points 8.5 cm, approximated wavelength of the carrier, is chosen. The physical placement of the grid is shown in Fig. 8. Thus, the obtained BER performance is the averaged value over 100 separate channels with a similar expected SNR value. The measured highest SNR value, which is measured as 25 dB for 16-QAM and 64-QAM, is the starting point for BER performance analysis. The rest of the analyses were carried out by adding the sampled noise sequence to received signal in order to obtain the BER values for lower SNR levels. For different levels of SNR the noise sequence was multiplied by a constant, which is adjusted due to desired SNR level, before adding the noise sequence to the received data sequence.

5. Experimental and Simulation Results

In this study, conventional CMA for blind training and two conventional adaptive training algorithms for non-blind trainings, LMS and RLS, are employed for equalizations of experimental received data and in simulations. Since the obtained experimental BER performance is the averaged value over 100 separate channels as shown in Fig. 8, the simulations are performed over 100 channels by a Monte Carlo simulation. An 11 taps linear transversal equalizer (LTE) filter is used in both blind and non-blind training methods. The centre tap of LTE is set to unit value in blind trainings and otherwise the values of all taps are initialized to zero before starting training. The step size parameter of LMS was equal to 0.0025 for 16-QAM and S-16-QAM, and 0.00015 for 64-QAM and S-64-QAM, and the forgetting factor of RLS was 0.999 for all modulation types. Additionally, the step size parameter of CMA was equal to 0.00115 for 16-QAM and S-16-QAM, and 0.00005 for 64-QAM and S-64-QAM. The given training parameters are used in both experimental and simulation data. The non-blind trainings, LMS and RLS, are carried out using all three CAZAC sequences at the beginning of each assigned sub-sequences. Thus, 1152 (=6x192) and 1920 (=10x192) steps of non-blind training are executed before starting the recovery of incoming data for attained 16-QAM, 64-QAM, S-16-QAM and S-64-QAM modulation types, respectively.

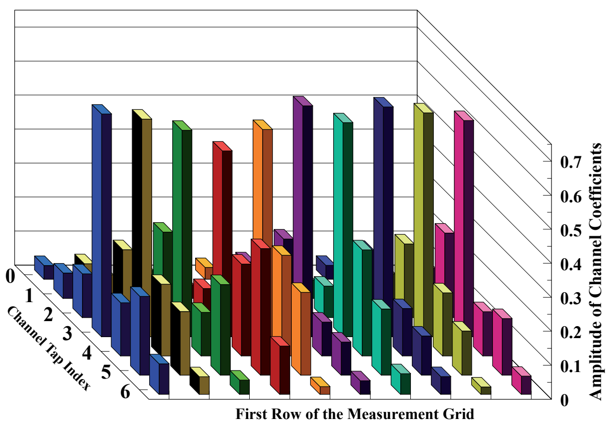


Fig. 9. A sampled channel profile obtained from experimental radio.

The blind equalizations were completed in four steps: First, the beginning of each sub-sequence is defined in the received data sequence. Second, the blind training is carried out over the entire length of each sub-sequence 8160 (=8x1020) steps for 16-QAM and S-16-QAM, 13600 (=20x680) steps for 64-QAM and 10880 (=2x5440) steps for S-64-QAM). Third, the ISI is cancelled by running the blindly-trained equalizer filter over the CAZAC sequences and received data for each sub-sequence. Forth and finally, the phase correction coefficient is obtained using the CAZAC sequences at the beginning of sub-sequence as in [6] and the phase correction of the related sub-sequence was done before detection and decoding of the sub-sequence.

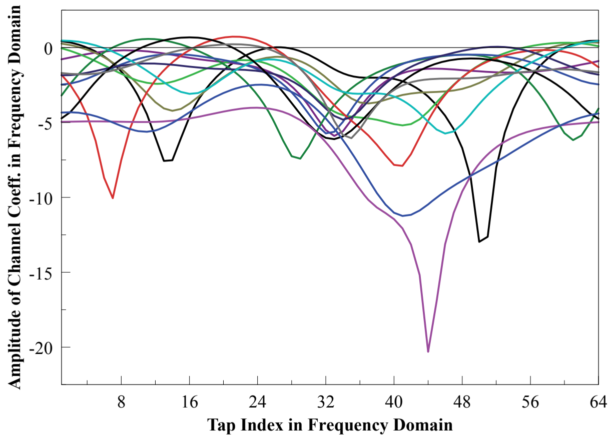


Fig. 10. Frequency responses of the channels given in Fig. 9.

Fig. 9 shows a sampled channel profiles with 7 taps observed above the noise floor of the receiver used in the experiments, at the first row of the measurement grid shown in Fig. 8. The frequency responses of the same channels are shown in Fig. 10.

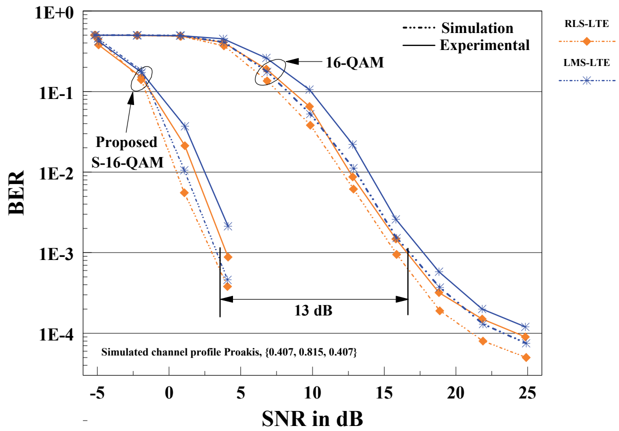


Fig. 11. BER performances of non-blind equalizers for experimental and simulated channels using 16-QAM and S-16-QAM modulations.

The obtained BER performances of non-blind trainings, LMS and RLS, and blind training, conventional CMA for simulated and experimental channel equalizations are given by Fig. 11 and 12 for 16-QAM and S-16-QAM, where 239/256 Reed Solomon and 1/2 Convolutional Coding are employed in cascade as it is explained in [1]. The dashed lines belong to simulation performances of equalizations using the channel profile given by (0.407, 0.815, 0.407), which is defined in [17]. The experimental data produces around 1.5 to 2 dB worse performances than the simulated channels' data since the channel delay spreads have got quite bigger in some measurement points as it is shown in Fig. 9.

Fig. 11 compares the BER performances of two non-blind equalizers (LMS and RLS) for 16-QAM and S-16-QAM. It can be seen from Fig. 11 the BER performance obtained using the proposed S-16-QAM technique outperforms the conventional 16-QAM and provides high SNR improvement of approximately 13 dB for a BER value of 1E-3. This can be explained by the spreading unit, preventing sequential error bursts during the blind and non-blind channel equalization process. The performance improvement by the proposed technique is very significant that, with little increase on the complexity, the conventional 16-QAM and 64-QAM have become a high performance modulation technique.

Fig. 12 also compares the BER performances of blind equalizer for 16-QAM and S-16-QAM. The performances of conventional CMA produce error floor both in experimental and simulated channels for conventional 16-QAM. On the other hand, CMA blind equalizer using the proposed S-16-QAM, improved in this study, provides satisfactory performance and outperforms the conventional 16-QAM performances and cancel error floor.

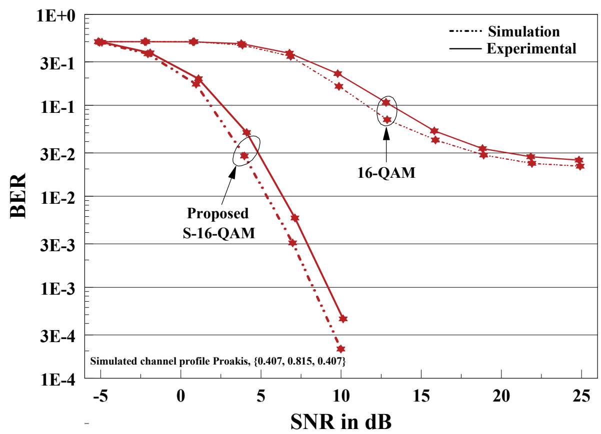


Fig. 12. BER performances of CMA blind equalizer for experimental and simulated channels using 16-QAM and S-16-QAM modulations.

The obtained comparative BER performances of non-blind and blind trainings for simulated and experimental channel equalizations are given by Fig. 13 for the proposed S-16-QAM.

This is one of the first real-time experimental studies of blind equalizations that the performances of non-blind equalizations are 5 dB better than the considered blind equalization technique. When the performances belong to S-16-QAM modulation are considered, the RLS training is around 1 dB better than the LMS however both algorithms are able to clear the error region below 5 dB of SNR. Of course, the coding helps to the LMS trainings much more than those using the RLS. Then, the simplicity of the LMS algorithm can easily put forward the LMS as the best candidate for a real time application. However, the blind training is not too far back, especially CMA blind equalizer, using the proposed method, also cancels the error floor around 10 dB of SNR and its performance is quite solid for S-16-QAM. The obtained performances of blind equalizer for S-16-QAM are quite important that the blind technique is as good as their non-blind counterparts in noise limited region (SNR < 20 dB), by opening new research areas in spread spectrum communications.

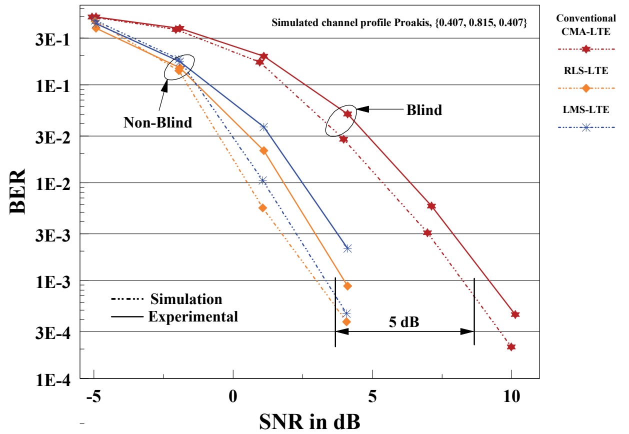


Fig. 13. Comparative BER performances of non-blind and blind equalizers for experimental and simulated channels in S-16-QAM modulation.

The obtained BER performances of non-blind and blind trainings for simulated and experimental channel equalizations are given by Fig. 14 and 15, respectively for 64-QAM and S-64-QAM.

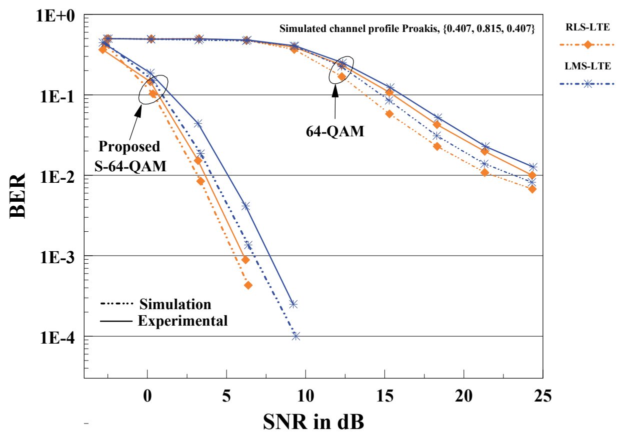


Fig. 14. BER performances of non-blind equalizers for experimental and simulated channels in 64-QAM and S-64-QAM modulations.

Fig. 14 evaluates the BER performances of two non-blind equalizers for 64-QAM and S-64-QAM. As can be seen from Fig. 14 the BER performance obtained using the proposed S-64-QAM technique outperforms the conventional 64-QAM.

For conventional 64-QAM modulation, the obtained performances of blind equalization start to be varying since CMA produces error floor both in experimental and simulated channels. However, CMA blind equalizer using the proposed S-64-QAM technique outperforms the conventional 64-QAM performances and cancels the error floor as can be seen from Fig. 15.

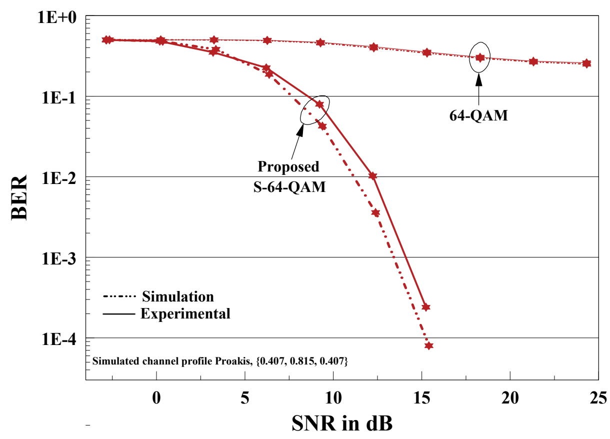


Fig. 15. BER performances of CMA blind equalizer for experimental and simulated channels in 64-QAM and S-64-QAM modulations.

The obtained comparative BER performances of non-blind and blind trainings for simulated and experimental channel equalizations are given by Fig. 16 for the proposed S-64-QAM.

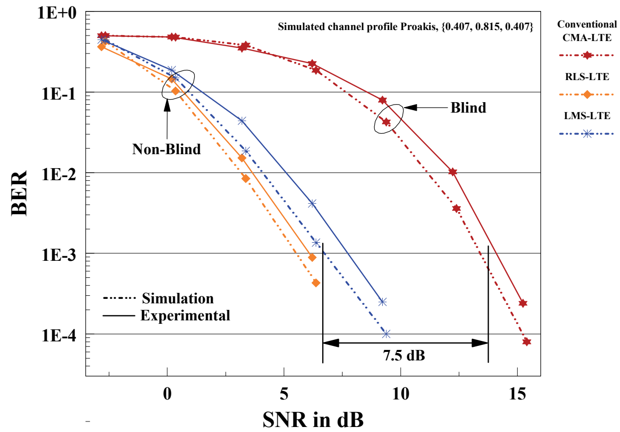


Fig. 16. Comparative BER performances of non-blind and blind equalizers for experimental and simulated channels in S-64-QAM modulation.

As can be seen from Fig. 16 the BER performances of non-blind equalizations are 7.5 dB better than the considered blind equalization technique. When the performances belong to S-64-QAM modulation are considered, the RLS training is around 2 dB better than the LMS however both algorithms are able to clear the error region below 10 dB of SNR. The blind training is not too far back, especially CMA blind equalizer, using the proposed technique, also cancels the error floor around 15.5 dB of SNR, and its performance is quite solid for S-64-QAM modulation.

6. Conclusion

This research has been carried out to study the BER performances of adaptive equalizations using blind and non-blind training techniques in order to obtain a comprehensive performance profile in both experimental and also simulated wireless communication channels for high-order QAM signaling. The non-blind equalizer, RLS and LMS, have been compared with the popular CMA blind equalizer as a benchmark. Since it prevents sequential error bursts, the proposed S-16-QAM and S-64-QAM, with little increase on the complexity, offer practical alternatives to blind equalization of higher-order QAM channels and provide significant equalization improvement over the conventional 16-QAM and 64-QAM. The obtained BER performances approve that blind equalizations can be used in future wireless communications. In particular, when the spectrum efficiency, low complexity and low level of received signal powers are considered in an embedded transceiver design, the blind techniques can easily be employed with bearable performance degradation. The simulation results have also demonstrated that the simulated and experimental studies of the proposed technique are supporting each other with a great compatibility.

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