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IMPULSIVE NOISE CANCELLATION IN SYSTEMS WITH OFDM MODULATION

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Orthogonal Frequency Division Multiplex (OFDM) is recently widely used in many wireless and wire line communication systems. The parallelism between RS code and OFDM modulation with consecutive suppressed carriers is known. In this paper the capability of OFDM modulation to act as an error correcting code is analyzed if used for transmission over channel with impulsive noise. New decoding algorithms of OFDM-RS code are presented with emphasis on the low implementation complexity. The decoding and error correcting performance of these new algorithms has been evaluated in series of simulations. As the channel model for simulations, Power Line Communication channel has been selected. Results of these simulations are presented as well.

Keywords: orthogonal frequency-division multiplexing (OFDM), Reed-Solomon (RS) codes, Impulsive noise cancellation, multicarrier systems

1 INDRODUCTION

Orthogonal frequency-division multiplexing [1] (OFDM) system represents very popular and broadly used modulation technique. Although it is based on principles which are quite old, thanks to the progress in development of fast signal-processing components and other supporting technologies in the near past, it was incorporated into many communication systems of recent days. It can be found in European digital audio and video broadcasting (HD-Radio, DVB-T, DVB-H), in wireless communication standards (IEEE 802.11a/g, WiMAX), mobile communication systems (IEEE 802.20, Flash-OFDM) and also in xDSL systems. It is a candidate for modulation technique for Power Line Communication systems as well.

All communication channels mentioned above suffer from the same very disrupting phenomenon — impulsive noise. It can be described as an additive disturbance that is caused by switching electric equipment in the transmission environment. Depending on the environment its amplitude can reach several dB over background noise and from spectral perspective the pulse has a pole and infinite energy [2]. The consequence is the occurrence of isolated or more often bursty errors even if OFDM modulation is used.

To be able to decrease the impact of this kind of noise on transmitted signal a very robust and powerful encoding scheme is needed. The drawback of such a approach is high redundancy added to the useful information. It results in the lower efficiency of transmission. To overcome this shadow side of encoding there would be an option to use already existing redundancy. One such a source of the redundant information could be OFDM modulation itself.

One of such a useful forms of redundancy in OFDM can be sequence of consecutive null symbols appended to the block of information symbols due to over sampling, spectral restrictions or spectral adaptability. As it was already shown in the [3] OFDM is a Reed-Solomon code and it can be used as a specific impulsive noise canceller [4]. Because of the nature of impulsive noise (one impulse = one symbol error) it is more suitable for its cancellation than the classical FEC decoder. However the classical FEC code is also necessary, since the impulsive noise is not the one and only negative phenomenon in the common transmission channels.

Several approaches based on OFDM modulation has been already proposed concerning mainly usage of the BoseChaudhuriHocquengem (BCH) code in the complex and real field in order to minimize the effect of the minor errors caused by the impulsive noise. Most of them are based on algorithms known for codes defined over Galois fields. For example Wolf used BCH codes in [5] and Abdelkefi in [4], Marvasti in [6] and Kumaresan in [8] used RS codes in cooperation with DFT. Wu [9] has even defined a class of the real-number block codes using the discrete cosine transform. For decoding he also used a modified Berlekamp-Massey algorithm and Forney algorithm. All these authors used similar decoding algorithms based on the already well known techniques for decoding RS or BCH codes over finite fields. They have modified them by introducing different thresholds based often on empirical estimations from simulations. The thresholds are used for taking decisions about further progress in decoding algorithms in contrast to exact results from calculations over finite fields. All of them are very sensitive to the errors caused by AWGN noise and have problems

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with stability and quite high complexity because of the certain amount of indeterminateness in their calculations.

The paper is organized as follows. In Section 2 short description of the OFDM modulation with focus on the redundancy and way to use this redundancy for impulsive noise estimation and cancellation is given. Also parallelism to the RS code is mentioned and based on it the real number codes are defined. In Section 3 new decoding techniques of the so called "OFDM-RS" code based on different applications of the "intelligent brute force" and dictionary based methods are present. The presentation starts with very simple method, based on pure code word dictionary search and later it continues with more complex one using iterative calculations and syndrome analyses. In contrast to decoding algorithms derived from the already well known algorithms designed for codes defined over finite fields, the new ones are much simpler and their application in the real communication systems is pretty straightforward. In Section IV some results of their simulation analysis are presented. As a model for impulsive noise one described for PLC communication channel in [10, 11] was used. In Section Conclusions the results of this work are summarized.

2 THEORETICAL PART

2.1 Description of OFDM modulation as RS spectral code

Modulation of the information signal in OFDM is based on the Inverse Discrete Fourier Transform (IDFT), where N input samples are computed using IDFT (1) and transformed from the frequency domain representation into the N samples in time domain. Input samples could be real or complex numbers according to the type of modulation used per each subchannel. Also some of them could be equal to zero, because of the used spectral schema (because of regulation, unfriendly transmission environment, or oversampling).

Let $X[k]$, $k = 0, 1, \dots, N-1$, be an input sample sequence over field of complex numbers, then the transformed signal (in time domain) $x[n]$, $n = 0, 1, \dots, N-1$ can be expressed as

$$x[n] = \frac{1}{N} \sum_{k=0}^{N-1} X[k] \phi^{-kn}, \tag{1}$$

where $\phi = e^{j\frac{2\pi}{N}}$ and represents Nth root of unity. In matrix expression (1) can be defined as

$$\boldsymbol{x} = \mathbf{X}\mathbf{Q}^H, \tag{2}$$

where $\mathbf{Q}^H$ denotes a Hermitian transpose (the combination of complex conjugation and ordinary matrix transposition) of orthonormal or unitary DFT matrix $\mathbf{Q}$ which can be described by following matrix:

$$\mathbf{Q}^H = \frac{1}{\sqrt{N}} \begin{bmatrix} 1 & 1 & 1 & \dots & 1 \\ 1 & \phi & \phi^2 & \dots & \phi^{N-1} \\ 1 & \phi^2 & \phi^4 & \dots & \phi^{2(N-1)} \\ 1 & \phi^3 & \phi^6 & \dots & \phi^{3(N-1)} \\ \vdots & \vdots & \vdots & \dots & \vdots \\ 1 & \phi^{(N-1)} & \phi^{2(N-1)} & \dots & \phi^{(N-1)^2} \end{bmatrix} . \tag{3}$$

In other words any vector $\boldsymbol{x}$ can be expressed as a linear combination of rows of $\mathbf{Q}^H$ matrix from (3). In case that only specific number of rows K is selected and $K < N$, the set of all linear combinations of these rows defines a K-dimensional subspace of the N-dimensional space. This subspace is a code. As shown in [12] this code can be seen as Reed-Solomon code over complex numbers.

Reed-Solomon codes over finite fields can be constructed in both time and frequency domains [13]. In case of frequency domain, the spectrum of a RS codeword lives in the same field as information symbols. To form a RS code, a block of $2t$ consecutive spectral components is set to zero [8] and then transformed back to the time domain. These $2t$ frequencies are parity frequencies and such a RS code is capable to correct t corrupted symbols. One can easily identify here the equivalence between $2t$ parity frequencies and $N-K$ not selected rows from the IDFT transformation matrix (3).

For RS codes as also for all the others block codes the encoding process using the codes generator matrix $\mathbf{G}$ can be described as

$$\boldsymbol{c} = \boldsymbol{i}\mathbf{G} \tag{4}$$

where the information vector $\boldsymbol{i} = (i_0, i_1, \dots, i_{K-1})$ is encoded into the code vector $\boldsymbol{c} = (c_0, c_1, \dots, c_{N-1})$. If we now use the analogy between RS codes definition in frequency domain and OFDM modulation (2, 3), we will get for the generator matrix $\mathbf{G}$ of the "OFDM-RS code" that can be expressed as:

$$\mathbf{G} = \mathbf{M}_G \mathbf{Q}^H \tag{5}$$

where matrix $\mathbf{M}_G$ consists of matrix identity with dimensions $K \times N$ and then zero matrix with dimensions $(N-K) \times N$.

$$\mathbf{M}_G = \begin{bmatrix} \mathbf{I} \\ \mathbf{0} \end{bmatrix} . \tag{6}$$

Because of the last $(N-K)$ rows of matrix $\mathbf{G}$ are zero vectors, they can be dropped so the final size of $\mathbf{G}$ will be $K \times N$. It needs to be said here that the form of $\mathbf{M}_G$ in (5) is only one example. The position of zero rows can be different and depending on it one can get codes with some specific characteristics. Concretely in case of real input data one can through $\mathbf{M}_G$ do easy manipulation of rows of $\mathbf{Q}^H$ matrix to achieve the real output signal after encoding. These manipulations are based on the scaling rows of $\mathbf{Q}^H$ matrix and achieving conjugate symmetries of rows about ± 1. For complex input data more complex manipulations needs to be done and there are also some restriction for input data [6].

The $(N-K)\times N$ parity check matrix $\mathbf{H}$ of rank $N-K$ can be defined as follows:

$$\mathbf{G}\mathbf{H}^H = \mathbf{0}\,. \tag{7}$$

The parity check matrix can be constructed quite easily from the DFT vectors associated with the roots defining the code. Practically it means that we will get $\mathbf{H}$ if we combine all not used rows of transform matrix $\mathbf{Q}^H$, which can be defined as:

$$\mathbf{H} = \mathbf{M}_H\mathbf{Q}^H \tag{8}$$

where matrix $\mathbf{M}_H$ can be expressed as:

$$\mathbf{M}_H = \begin{bmatrix}\mathbf{0}\\ \mathbf{I}\end{bmatrix}. \tag{9}$$

The zero matrix here has the size of $K\times N$ and the matrix of identity $\mathbf{I}$ is $(N-K)\times N$ matrix. As already mentioned in connection to the generator matrix $\mathbf{G}$, if the code has some specific characteristics (*ie* real code words) then the modified $\mathbf{M}_H$ matrix needs to be used, which will achieve conjugate symmetries of rows about ± 1.

2.2 Decoding of the OFDM sequence

If the transmission of the signal modulated with OFDM modulation through a memoryless channel is assumed, at the receiver side there will be the following signal r_n containing the original transmitted signal c_n and then contribution of additive white Gaussian noise (AWGN) g_n plus contribution of impulsive noise i_n:

$$r_n = c_n + g_n + i_n\,, \quad n\in\{0,\dots,N-1\}\,. \tag{10}$$

Intersymbol interference was not considered, however if there is such a channel with ISI then it is trivial to remove it with cyclic prefix, as it is usually done in OFDM systems. For demodulation purposes that sequence r_n needs to be transformed from time to the frequency domain using DFT what will result in sequence

$$R_n = C_n + G_n + I_n\,, \quad n\in\{0,\dots,N-1\}\,. \tag{11}$$

If the OFDM-RS code was applied in the transmitter then the component C_n contains certain number of consecutive zeros. But these zeros are now not visible in the received R_n sequence anymore because of contribution of the Fourier transform of the Gaussian background noise and Fourier transform of noise impulses. The transformed Gaussian background noise is still Gaussian, the transformed impulsive noise is a sum of complex sinusoids which frequencies corresponds to the positions of errors.

Decoding (error correction) of such a signal can be done in more ways. The mostly used one is decoding through syndromes. If we now assume that in the original sequence C_n there were zeros at positions a_i, $i = 0,\dots,N-K-1$ then the syndromes can be expressed as:

$$S_i = R_{a_i} - C_{a_i} = G_{a_i} + I_{a_i} = \sum_{n-0}^{N-1} g_n\phi^{na_i} + \sum_{n-0}^{N-1} i_n\phi^{na_i}. \tag{12}$$

If the number of impulsive noises is M and sequence $\{\sigma_m\}$ where $m\in\{0,\dots,M-1\}$ represents the location of the impulsive noise in the sequence r_n. Then the expression can be modified as

$$S_i = \sum_{n-0}^{N-1} g_n\phi^{na_i} + \sum_{n-0}^{N-1} i_{\sigma_m}\phi^{\sigma_m a_i}\,. \tag{13}$$

The decoding problem is the estimation of number of the sinusoids representing the impulsive noise together with their frequencies and amplitudes, polluted by background noise. The number of sinusoids and their frequencies takes the integer values in contrast to the amplitudes which are real numbers.

In most of the current works [5–9] about decoding of the real number codes especially the ones based on DFT transformation there are proposed algorithms derived from the decoding in the finite field domain of BCH and RS codes. Their common sign is quite a significant computational complexity and problems with stability.

Decoding algorithms presented in this paper are different. The approach used in this paper was to use very simple algorithms at the beginning and then modify them in order to improve their decoding capability as well as some of their other features, which are relevant for practical deployment.

One of the simplest approaches is to use minimum distance decoding. In case of DFT codes it means that for a received sequence $\boldsymbol{r}$ the algorithm picks up a code word $\boldsymbol{c}_{est}$ ($\boldsymbol{c}_{est}$ is any codeword from code word set) to minimize the Euclidean distance $d_{Euc}(\boldsymbol{r},\boldsymbol{c}_{est})$. Practically this can be realized through creation of the dictionary (look up table) containing all possible code words and their related information sequences. The decoder goes through the dictionary (code words part), fetching code words one by one and calculating Euclidean distance per each code word. The information sequence related to the most similar code word in sense of Euclidean distance is then selected as decoded one and transmitted information is derived. The dictionary itself is calculated by choosing all possible information sequences and transforming them through DFT into the time domain. The limitation of this decoding method is that the number of operations grows exponentialy with the length of the information sequence.

To decrease the size of the dictionary the decoding algorithm based on syndromes can be used. As shown in (14) the syndromes are completely independent from the information sequence and their values reflect the influence of the impulsive noise on the transmitted signal. The syndromes are practically computed using the parity check matrix $\mathbf{H}$ as follows:

$$\mathbf{S}_{ideal} = \boldsymbol{r}\mathbf{H}^H = \boldsymbol{c}\mathbf{H}^H + \boldsymbol{g}\mathbf{H}^H + \boldsymbol{i}\mathbf{H}^H = \boldsymbol{g}\mathbf{H}^H + \boldsymbol{i}\mathbf{H}^H + \boldsymbol{n}\mathbf{H}^H \tag{14}$$

where $\boldsymbol{n}$ stands for a channel noise containing impulsive and AWGN noise. Since the values of the received vector $\boldsymbol{r}$ are from the set of real numbers there is inevitably also the round-off noise $\boldsymbol{\omega}$ introduced during processing of that received sequence. That is the reason why the syndrome expression (14) can be adjusted as follows:

$$\mathbf{S} = \boldsymbol{n}\mathbf{H}^H + \boldsymbol{\omega}\,. \tag{15}$$

The new proposed decoding algorithm for OFDM-RS codes using syndromes is based on the code word estimation, which is achieved through estimation of the channel noise. Generally said, the decoder is checking the whole set of possible channel noise sequences ϑ and through (16) doing estimation of transmitted sequence $\hat{\hat{\boldsymbol{c}}}_j$ in time domain, $j \in \{1,\dots,M\}$, where $M = \mathrm{card}(\vartheta)$.

$$\hat{\hat{\boldsymbol{c}}}_j = \boldsymbol{r}_j - \hat{\boldsymbol{n}}_j\,. \tag{16}$$

The precision of estimation is then evaluated through calculation of syndromes (17).

$$\mathbf{S}_j = \hat{\hat{\boldsymbol{c}}}_j\mathbf{H}^H + \boldsymbol{\omega}_j\,, \quad j \in \{1,\dots,M\}\,. \tag{17}$$

The sequence matching the following precondition

$$\hat{\hat{\boldsymbol{c}}} = \hat{\hat{\boldsymbol{c}}}_j\,, \text{ where } d_{Eucl}(\mathbf{S}_j,\mathbf{0}) = \min\big\{d_{Eucl}(\mathbf{S}_j,\mathbf{0})\,, \; j \in \{1,\dots,M\}\big\} \tag{18}$$

or in other words the sequence with lowest syndrome value (the closest one to the zero vector) is picked up as the most probably transmitted one and decoded through the DFT producing the estimate of the input data $\hat{\mathbf{I}}$ (19).

$$\hat{\mathbf{I}} = \hat{\hat{\mathbf{I}}} + \boldsymbol{\omega} = \hat{\hat{\boldsymbol{c}}}\mathbf{G}^{-1} + \boldsymbol{\omega}\,. \tag{19}$$

Since the input data are generally restricted to a large but finite alphabet, the round-off noise in processing, encoding and decoding does not cause intrinsic errors, even before other effects are included. The upper bound of the numerical values associated with the alphabet is affected by the dynamic range of the processing system whereas the precision determines round-off noise tolerance levels.

The third new introduced algorithm for decoding of OFDM-RS codes is the modification of the previous one. The difference is that instead of syndromes the decision criteria about the candidate for transmitted codeword is based on the Euclidean distance of the estimated transmitted sequence $\hat{\hat{\mathbf{C}}}$ from estimated codeword $\hat{\mathbf{C}}$ specified as:

$$\hat{\hat{\mathbf{C}}} = \hat{\hat{\mathbf{C}}}_j\,, \text{ where } d_{Eucl}\big(\hat{\mathbf{C}}_j,\hat{\hat{\mathbf{C}}}_j\big) = \min\big\{d_{Eucl}\big(\hat{\mathbf{C}}_j,\hat{\hat{\mathbf{C}}}_j\big)\,, \; j = 1,\dots,M\big\} \tag{20}$$

where $\hat{\hat{\mathbf{C}}}_j$ represents $\hat{\hat{\boldsymbol{c}}}_j$ transformed into the frequency domain through DFT and $\hat{\mathbf{C}}_j$ stands for a codeword. The set of all code words θ is finite and let assume that $L = \mathrm{card}(\theta)$ then $\hat{\mathbf{C}}_j \in \theta$ can be derived from $\hat{\hat{\mathbf{C}}}_j$ as

$$\hat{\mathbf{C}}_j = \hat{\hat{\mathbf{C}}}_j + \boldsymbol{\omega}_j\,, \text{ where} \tag{21}$$

$$d_{Eucl}\big(\hat{\mathbf{C}}_j,\hat{\hat{\mathbf{C}}}_j\big) = \min\big\{d_{Eucl}\big(\hat{\mathbf{C}}_l,\hat{\hat{\mathbf{C}}}_j\big)\,, \; l = 1,\dots,L\,,\; j = 1,\dots,M\big\}\,. \tag{22}$$

To find solution for (22) it is not necessary to check all combinations of $\hat{\mathbf{C}}_l$ and $\hat{\hat{\mathbf{C}}}_j$, because $\hat{\mathbf{C}}_l$ are usually the complex numbers containing integer values. So simple rounding of $\hat{\hat{\mathbf{C}}}_j$ (in some cases also rescaling) will lead us to the closest $\hat{\mathbf{C}}_l$ vector for particular $\hat{\hat{\mathbf{C}}}_j$. The final codeword estimation $\hat{\mathbf{C}}$ could be done from $\hat{\hat{\mathbf{C}}}$ (20) as

$$\hat{\mathbf{C}} = \hat{\hat{\mathbf{C}}} + \boldsymbol{\omega}\,. \tag{23}$$

Since in this case the decision is done based on the approximate confidence measure one can easily find this algorithm as soft-decision based. It is in contrast to the previous one where the hard-decision approach is used.

Both algorithms give us the possibility to carefully check if the decoding algorithm was working properly. The check procedure is based on the final Euclidean distance between syndrome and zero vector (for hard-decision approach) or final Euclidean distance of the transmitted sequence $\hat{\hat{\mathbf{C}}}$ from estimated codeword $\hat{\mathbf{C}}$ (for soft-decision approach). The lower Euclidean distance is, the higher probability of the proper decoding of the transmitted sequence can be expected. On the other side if the Euclidean distance is higher than some specific threshold, one can assume malfunctions of decoder, particularly problem with convergence of the decoding algorithm due to wrong parameters (precision, dynamic range), or higher level of background noise, or possible overflow of the error correcting capacity.

In the previous two paragraphs the general algorithm of decoding OFDM-RS code was described. Since in both approaches (soft and hard-decision) one need to check the whole set of channel noise sequences the algorithm is from this point of view just hypothetical case with infinite number of operations. To make it practical the infinite set of all possible channel noise sequences needs to be restricted to the finite set. This could be achieved in different ways.

Based on the experimental results the new decoding algorithm has been defined. It decreases significantly the number of operations needed for decoding and is easily to implement. The algorithm consists of three basic steps:

- Estimation of the impulsive noise position and raw estimation of its amplitude
- Precise estimation of the impulsive noise amplitude
- Fine tuning of the impulsive noise vector estimation

The main target of the first step is to create a candidate for noise vector which has corrupted the transmitted data sequence. The candidate is supposed to have correct positions of impulsive noise impulses but their amplitudes will be just roughly estimated. The input for the first step is the dynamic range of the possible noise impulses and precision of the raw impulsive noise amplitude estimation. Based on these values the set Ψ of the initial amplitudes is created. This set could be defined in more ways. Good choice according to the experimental results seems to be set of amplitudes defined as

$$\Psi = \{A^{m_{\min}}, \dots, A^{m_{\max}}\} = \{10^{m_{\min}}, 10^{m_{\min}}j, \dots, 10^0, 10^0 j, \dots, 10^{m_{\max}}, 10^{m_{\max}}j\}, \quad m \in \{m_{\min}, \dots, 1, 0, 1, \dots, m_{\max}\} \qquad (24)$$

where values $m_{\min}$ and $m_{\max}$ are the input parameters. Placing values of amplitudes from Ψ one by one on all positions of the noise vector and calculating $d_{Eucl}(\mathbf{S}_j, \mathbf{0})$ using (17), (18) for hard decision or calculating (22) for soft decision approach algorithm one will get the Euclidean distance matrix $\mathbf{D}_{Eucl}$. In other words each element of the Euclidean distance matrix is calculated from different noise vector in form of:

$$\boldsymbol{n}_{j,n} = [\,0 \quad \dots \quad A^m \quad \dots \quad 0\,] \qquad (25)$$

where index n defines the number of iteration.

Afterwards the candidate noise vector $\boldsymbol{n}_{cand,n}$ is picked up as one related to the element fulfilling the expression (18) respectively (20).

In step two the amplitude of the noise impulse is estimated. The input parameter is precision of this estimation. The estimation is done based on the bisection method [14], where always the Euclidean distances for border and middle values are calculated and those with lowest value are chosen for the next iteration. The bisection method is stopped when the requested precision is reached.

Then the algorithm goes back to step one, do the new calculation of Euclidean matrix but now with noise vectors where one position has been assigned with already estimated amplitude. Then again candidate noise vector $\boldsymbol{n}_{cand,n+1}$ is selected and step two is applied. The sequence of positions of estimated amplitudes $\mathbf{P}_A$ is recorded to be used in the next step. The loop is finished after N steps or when all positions in candidate noise vector has assigned estimated amplitudes.

The fine tuning of amplitudes is done in the last step. The input parameter is now the precision of amplitude value estimation. It is the same parameter as in step two but now with lower value, since the higher precision is requested. The amplitude adjusting is based again on bisection method [14]. Similar to step two one amplitude value is selected and using bisection method looped till the requested precision is reached. Then the algorithm continues with amplitude on the next position. The order of the selected amplitudes is specified in vector $\mathbf{P}_A$ created in step two. When all N amplitudes are calculated the step three is over and the output is the impulsive noise candidate vector $\boldsymbol{n}_{cand}$.

3 EXPERIMENTAL PART

Both syndrome based decoding algorithms mentioned in the previous chapter (hard decision and soft decision approach) have been evaluated in the simulations. Since impulsive noise is significant in the PLC channel, the transmission through model of such a channel has been simulated. The transmitted sequence has been encoded with OFDM-RS code then transferred through the channel, received and decoded using decoding algorithms described in the text above. Simulation of the impulsive noise was part of the channel model. Although in the PLC channel different type of impulsive noises are present [10, 11], one can identify there two basic categories; background and impulsive noise. Based on this fact there was simulated only the final impact of the impulsive noise on the transmitted sequence as the result of all this particular noises. Thanks to this generalization two scenarios could be distinguished for channel with impulsive noise. In first one the impulsive noise with short duration of noise impulses was assumed. Short duration means that only one symbol of transmitted sequence has been corrupted. In second case there was ingested the impulsive noise with long impulses where more than one (two or three) symbol has been corrupted. Besides impulsive noise also impact of AWGN noise was simulated. The simulations have been done in software tool *Matlab v7*.

In presented simulations there were $N = 32$ carriers used for transmission in OFDM modulation. Since the oversampling was applied the IDFT/DFT was calculated from 64 samples. Also both decoding algorithms were working over 64 samples. The length of the guard interval was 16 samples. Among these 32 carriers 8 or 4 (depending on type of code) consecutive ones were equaled to zero. The remaining $K = 24$ or $K = 28$ were used for transmitting data information with 8-QAM modulation used within each carrier.

With settings described above several simulations has been done to prove the impulsive noise removing capability of OFDM-RS codes and our proposed decoding algorithms. Simulations are organized in the similar way and as the result the symbol error rate (SER) is plotted as a function of SNR (dB) for three different decoding algorithms; no error-correcting minimum distance decoding, syndrome based decoding algorithm with hard decision approach and syndrome based decoding algorithm with soft decision approach.

The first figure depicts the situation when very low level of AWGN noise and impulsive noise of short duration is present in the PLC channel.

It is clearly visible the difference between all three decoding techniques, best results are achieved with soft decision approach of the syndrome based decoding against the hard decision approach (coding gain about 6 dB) and the worst with minimum distance decoding.

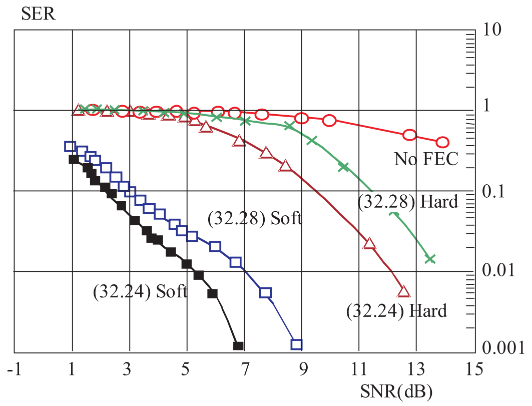

Fig. 1. OFDM-RS codes in channel with impulsive noise of short duration.

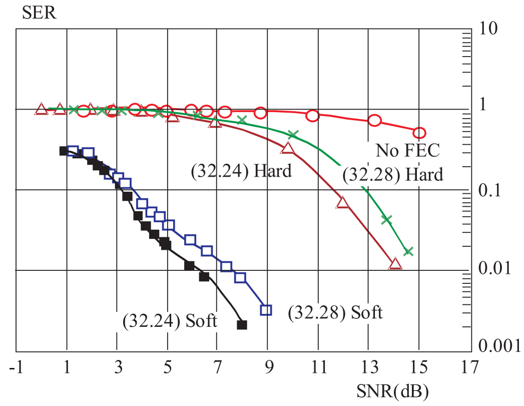

Fig. 2. OFDM-RS codes in channel with impulsive noise of long duration.

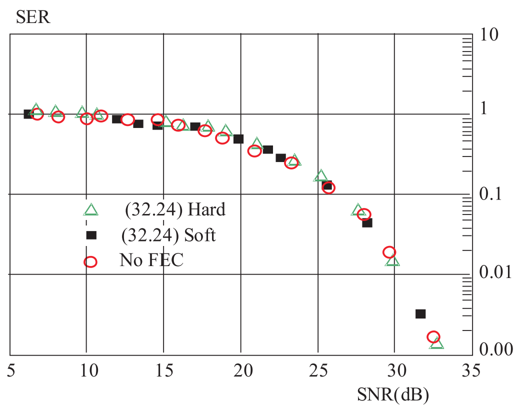

Fig. 3. OFDM-RS code N=32, K=24 in channel with AWGN noise.

Also comparison of the codes reveals us visible difference in correcting capabilities of the OFDM-RS (32, 24) and OFDM-RS (32, 28) code. The difference is about 2 dB and as expected in favor of OFDM-RS (32, 24).

Figure 2 shows the situation when long impulsive noise and low level of the AWGN noise is presented in the transmission channel. The results are similar to the previous case, where the best capability of correcting errors is achieved with soft decision approach (coding gain is 7 dB in comparison with hard decision), then hard decision approach follows and the last is minimum distance decoding. As expected also for this type of noise, OFDM-RS (32, 24) has performed slightly better than OFDM-RS (32, 28) code, in concrete values the gain for soft decision approach was 1.5 dB and for hard decision approach about 1 dB in favour of OFDM-RS (32, 24).

Both figures reveals also the fact that the long lasting impulsive noise has worse impact on the transmitted information as short one and all analyzed decoding techniques can less successfully cope with it. The difference is about 1–1.5 dB.

The last series of simulations analyze the ability of the OFDM-RS code to cope with the AWGN noise. In comparison to the previous simulations, now the impulsive noise was completely suppressed and only AWGN noise was presented but with more than 20 times higher variance value than before.

Figure 3 depicts us the results from these simulations. To make the picture clearer and more easily readable there are showed results only from OFDM-RS (32, 24) code since previous simulations has already proved the better results of the code with these parameters. Again there are compared all three decoding algorithms. As one can see the best results has been achieved with syndrome based algorithm with soft decision approach, but the differences are minimal. This reveals a very poor ability of the OFDM-RS codes to protect the useful signal against the impact of the Gaussian background noise.

4 CONCLUSIONS

In this paper there is proposed the new way of the impulsive noise correction based on the "OFDM-RS" coding techniques. The new decoding algorithms for these codes based on various applications of the "intelligent brute force" and dictionary based methods are proposed and analyzed. In contrast to already well known decoding algorithms, are these ones much simpler and closer to the implementation in the real communication systems. Besides the algorithms also controlling procedure for estimation quality of decoding and malfunction detection of the decoding algorithm is described.

The performance verification of the proposed codes and their decoding algorithms has been performed in the Broadband Power Line transmission environment. The algorithms were tested in three different noise scenarios; in environment with long lasting, short lasting and finally with AWGN noise.

The simulations proved good impulsive noise errors correcting capability, especially when the syndrome based decoding algorithm with soft decision approach has been used. Worse results have been achieved in channel with Gaussian background noise, but this is well known fact that DTF based codes have limited capability to cope with this type of noise. However the proposed algorithms represent just the initial approach and authors see a lot of space for optimization to improve their performance in case of usage in other than impulsive noise environments.

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