PERFORMANCE OF CODED OFDM IN IMPULSIVE NOISE ENVIRONMENT

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Abstract-Noise in the power line channel cannot be described by Additive White Gaussian Noise (AWGN). This channel suffers from impulsive noise and other narrow-band interferences. Impulsive noise, in particular, can degrade the performance of OFDM-based Power Line Communication (PLC) systems significantly. Therefore, channel coding is important to combat the impairments in this hostile medium, where the encoding and decoding of convolution codes over generated symbols for the transmission over impulsive noise channels. The encoder multiplies a vector of information symbols resulting from a modulation scheme choosing the inverse Fourier transforms as; the encoding procedure is similar to orthogonal frequency-division multiplex (OFDM) modulation. Simulation results will show the excellent performance of the Viterbi decoder. The performance of the OFDM is to study and find the probability of error performance verses signal to noise ratio with encoding and without encoding. In this paper, we study and simulate the performance of PLC channels in the presence of impulsive noise.

Keywords - Impulsive noise, OFDM, Viterbicoding and QPSK.

1. INTRODUCTION

Power Line Communication (PLC) technology is an emerging technology with potential results and recent developments as a multipurpose medium providing services such as energy distribution, voice, data and digital telecommunication services. Intense research activities focus the PLC’s due to all these advantages. Interest in Power Line Communication (PLC) technology as a broadband multimedia connectivity solution to and within the home continues to grow in a rapid pace. The driving advantage of this technology is that it exploits the already existing and ubiquitous power line distribution infrastructure to provide broadband multimedia services to customers. Because power lines were originally designed for AC power distribution at 50 Hz and 60 Hz, the characteristics of this channel present some technical challenges for data transmission at higher frequencies. Conventional communication systems differ from PLC systems not only in structure but also in physical properties. The performance of any communication system is limited by noise, attenuation and multipath propagation which are most significant channel properties [1].

Noise in PLC’s is much different from the classical additive white Gaussian noise. It can be classified into five types[2].They are colored background noise, narrow band noise, periodic impulsive...
noise asynchronous to the mains frequency, periodic impulsive noise synchronous to the mains frequency and asynchronous impulsive noise. The first three types of noise generally stay over long periods of time. The next two types are time varying and can be summarized as impulsive noise. Impulsive noise is mainly caused by power supplies or by switching transients in the network. It has a random occurrence and its duration varies from a few microseconds to milliseconds. Practical experiments in power lines [2] show that the power spectral density (PSD) of impulsive noise exceeds the PSD of background noise by minimum of 10-15 dB and may sometimes reach 50 dB. Manmade noise is typically impulsive. A relatively simple model that incorporates background noise and impulsive noise is suggested in [3] and is known as Middleton’s class-A noise. Middleton’s class-A noise model is the most widely used noise model in the performance analysis of PLC systems.

It is well known that the error performance of a communication system can be improved by using error control coding schemes such as BCH, Reed-Solomon (RS), Low Density Parity Check (LDPC) and convolution codes. Channel coding is a good way to combat with noise and improve the bit error rate of a system. Orthogonal Frequency Division Multiplexing (OFDM) is a mature multicarrier transmission technique that has been adopted in several wideband digital communication systems. Studies of OFDM system design are thus of great practical importance. The basic principle behind OFDM is to use a properly chosen linear transform at the transmitter and its inverse transform at the receiver. Most recently, the use of OFDM technique to combat with impulsive noise received a strong interest. OFDM performs better than single carrier in the presence of impulsive noise. This is due to OFDM spreads the effect of impulsive noise over multiple symbols due to Discrete Fourier Transform (DFT) operation. The effect of cyclic prefix in OFDM symbols can reduce the effect of multipath. Its basic premise is to divide the transmitted bit stream into many different sub streams and send these over many different sub channels. Typically, the sub channels are orthogonal to each other. The data rate on each of the sub channels is much less than the total data rate, and the corresponding sub channel bandwidth is much less than the total system bandwidth. The number of sub streams is chosen to insure that each sub channel has a bandwidth less than the coherence bandwidth of the channel, so the sub channels experience relatively flat fading. Thus, the Inter Symbol Interference (ISI) on each sub channel is small. The sub channel need not be contiguous, so it saves a lot of bandwidth while simultaneously achieving high data rates.

II. ORTHOGONAL FREQUENCY DIVISION MULTIPLEXING

A. Introduction

Orthogonal frequency division multiplexing (OFDM) is the modulation technique for European standards such as the Digital Audio Broadcasting (DAB) and the Digital Video Broadcasting (DVB) systems. As such it has received much attention and has been proposed for many other applications, including local area networks and personal communication systems. OFDM is a type of multichannel modulation that divides a given channel into many parallel sub-channels or subcarriers, so that multiple symbols are sent in parallel. OFDM is often motivated by two of its many attractive features: it is considered to be spectrally efficient and it offers an elegant way to deal with equalization of dispersive slowly fading channels.

B. Evolution of OFDM

The evolution of OFDM [4] can be divided into three parts. There are consists of Frequency Division Multiplexing (FDM), Multicarrier Communication (MC) and Orthogonal Frequency Division Multiplexing.
Frequency Division Multiplexing (FDM) has been used for a long time to carry more than one signal over a telephone line. FDM is the concept of using different frequency channels to carry the information of different users. Each channel is identified by the center frequency of transmission. To ensure that the signal of one channel did not overlap with the signal from an adjacent one, some gap or guard band was left between different channels. The concept of multicarrier (MC) communications uses a form of FDM Technologies but only between a single data source and a single data receiver. As multicarrier communications was introduced, it enabled an increase in the overall capacity of communications, thereby increasing the overall throughput. Referring to MC as FDM, however, is somewhat misleading since the concept of multiplexing refers to the ability to add signals together. MC is actually the concept of splitting a signal into a number of signals, modulating each of these new signals over its own frequency channel; multiplexing these different frequency channels together in an FDM manner; feeding the received signal via a receiving antenna into a de-multiplexer that feeds the different frequency channels to different receivers and combining the data output of the receivers to form the received signal.

OFDM is the concept of MC where the different carriers are orthogonal to each other. Orthogonal in this respect means that the signals are totally independent. It is achieved by ensuring that the carriers are placed exactly at the nulls in the modulation spectra of each other. Source for OFDM spectral efficiency is the fact that the drop off of the signal at the band is primarily due to a single carrier which is carrying a low data rate. OFDM allows for sharp rectangular shape of the spectral power density of the signal.

C. Principles of OFDM

The basic principle of OFDM is to split a high rate data stream into a number of lower rate streams that are transmitted simultaneously over a number of subcarriers. Because the symbol duration increases for lower rate parallel subcarriers, the relative amount of dispersion in time caused by multipath delay spread is decreased. Inter-symbol interference is eliminated almost completely by introducing a guard time in every OFDM symbol. In the guard time, the symbol is cyclically extended to avoid inter-carrier interference.

In OFDM design, a number of parameters are up for consideration, such as the number of subcarriers, guard time, symbol duration, subcarrier spacing, and modulation type per subcarrier. The choice of parameters is influenced by system requirements such as available bandwidth, required bit rate, tolerable delay spread, and Doppler values. Some requirements are conflicting. For instance, to get a good delay spread tolerance, a large number of subcarriers with small subcarrier spacing is desirable, but the opposite is true for a good tolerance against Doppler spread and phase noise.

Data Transmission using multiple carriers:

An OFDM signal consists of a sum of subcarriers that are modulated by using Phase Shift Keying (PSK) or Quadrature Amplitude Modulation (QAM). If \(d_i\) are the complex QAM symbol, \(N_s\) is the number of subcarriers, \(T\) is the symbol duration, and \(f_i = f_0 + i/T\) the carrier frequency, then one OFDM symbol starting at \(t = t_s\) can be written as:

\[
s(t) = \text{Re}\{\sum_{i=0}^{N_s-1} d_i \exp(j2\pi f_i(t - t_s))\}, \quad t_s \leq t \leq t_s + T
= 0, \quad t < t_s \text{ or } t > t_s + T
\]  

(2.1)

Often the equivalent complex notation is used, which is given by equation (2.2). In this representation, the real and imaginary parts correspond to the in-phase and quadrature parts of the OFDM signal, which have to be multiplied by a cosine and sine of the desired carrier frequency to produce the final OFDM signal.
As an example, figure 1 shows four subcarriers from one OFDM signal. In this example, all subcarriers have the phase and amplitude, but in practice the amplitudes and phases may be modulated differently for each subcarrier. Note that each subcarrier has exactly an integer number of cycles in the interval $T$, and the number of cycles between adjacent subcarriers differs by exactly one. This properly accounts for the orthogonality between subcarriers.

For instance, if the $j$th subcarrier from equation (2.2) is demodulated by down converting the signal with a frequency of $f_j = f_0 + j/T$ and then integrating the signal over $T$ seconds, the result is as written in equation (2.3). By looking at the intermediate result, it can be seen that a complex carrier is integrated over $T$ seconds. For the demodulated subcarrier $j$, this integration gives the desired output $d_j$ (multiplied by a constant factor $T$), which is the QAM value for that particular subcarrier. For all other subcarriers, this integration is zero, because the frequency difference $(i-j)/T$ produce an integer number of cycles within the integration interval $T$, such that the integration result is always zero.

$$dt = \sum_{i=0}^{N_s-1} d_i \exp \left( j 2\pi f_i (t - t_s) \right) dt = \sum_{i=0}^{N_s-1} d_i \int_{t_s}^{t_s+T} \exp \left( j 2\pi f_i (t - t_s) \right) dt = d_j T \quad (2.3)$$

The orthogonality of different OFDM subcarriers can also be demonstrated in another way. According to equation (2.1), each OFDM symbol contains subcarriers that are nonzero over a $T$ seconds interval. Hence, the spectrum of a single symbol is a convolution of group of Dirac pulses located at the subcarrier frequencies with the spectrum of a square pulse that is one for a $T$ second period and zero otherwise. The amplitude spectrum of the square pulse is equal to $\text{sinc}(\pi f T)$, which has zeros for all frequencies $f$ that are an integer multiple of $1/T$. This effect is shown in figure which shows the overlapping sinc spectra of individual subcarriers. At the maximum of each subcarrier spectrum, all other subcarrier spectra are zero. Because an OFDM receiver calculates the spectrum values at those points that correspond to the maxima of individual subcarrier, it can demodulate each subcarrier free from any interference from the other subcarriers. Basically, Figure 2 shows that the OFDM spectrum fulfills Nyquist’s criterion for an inter symbol interference free pulse shape. Notice that the pulse shape is present in frequency domain and note in the time domain, for which the Nyquist criterion usually is applied. Therefore, instead of inter-symbol interference (ISI), it is inter-carrier interference (ICI) that avoided by having the maximum of one subcarrier spectrum correspond to zero crossing of all the others.
Generation of sub-carriers using IFFT:

The complex baseband OFDM signal as defined by equation (2.2), is in fact nothing more than the inverse Fourier transform of \( N_s \) QAM input symbol. The time discrete equivalent is the inverse discrete Fourier transform (IDFT), which is given by

\[
s(n) = \sum_{i=0}^{N_s-1} d_i \exp \left( j2\pi \frac{in}{N} \right)
\]  

(2.4)

where the time \( t \) is replaced by a sample number \( n \). In practice, this transform can be implemented very efficiently by the inverse Fast Fourier transform (IFFT).

OFDM Transmitter Configuration:

Figure 3 shows the configuration of an OFDM transmitter. In the transmitter, the transmitted high speed data is first converted into parallel data of \( N \) sub channels. Then, the transmitted data of each parallel sub channel is modulated by PSK based modulation. Consider a quadrature modulated data sequence of the \( N \) channels \((d_0, d_1, d_2... d_{N_s-1})\) and \( d_{in} \) and \( d_{Qn} \) are \( \{1,-1\} \) in QPSK and \( \{\pm1, \pm3\} \) in 16-QAM. These modulated data are fed into an inverse fast Fourier transform (IFFT) circuit and an OFDM signal is generated.

One key principle of OFDM is that since low rate modulation scheme, where the symbols are relatively long compared to the channel time characteristics suffer less from inter-symbol interference caused by multipath. It is the advantageous to transmit a number of low rate streams in parallel instead of a single high rate stream. Since the duration of each symbol is long, it can be affordable to insert a guard interval between the OFDM symbols and thus the inter-symbol interference can be eliminated. The transmitter sends a cyclic prefix during the guard interval. The guard interval also reduces the sensitivity to time synchronization problems. The orthogonality of sub channels in OFDM can be maintained and individual sub channels can be completely separated by using an FFT circuit at the receiver when there are no ISI and inter-carrier interference (ICI) introduced by transmission channel distortion.

The spectra of OFDM signal are not strictly band limited, the distortion due to multipath fading causes each sub channel to spread the power into the adjacent channel. Moreover, the delayed wave with the delay time larger than 11 symbol time contaminates the next symbol. In order to reduce this distortion, a simple solution is to increase the symbol duration or the number of carriers. However, this method may be difficult to implement in terms of carrier stability against Doppler frequency and FFT size. Another way to eliminate ISI is to create a cyclically extended guard interval, where each OFDM symbol is preceded by a periodic extension of the signal itself.

The total symbol duration \( T_{total} = T_g + T_n \), where \( T_g = \text{guard time interval} \). Each symbol is made of two parts. The whole signal is contained in the active symbol, the last part of which is also repeated at the start of the symbol and is called a guard interval. When the guard interval is longer than the channel
impulse response or the multipath delay, the effect of ISI can be eliminated. However, the ICI or in band fading still exists. The ratio of the guard interval to the useful symbol duration is application dependent. The insertion of guard interval will reduce the data throughput; \( T_g \) is usually smaller than \( T_s/4 \).

**OFDM Receiver Configuration:**

![Figure 4. OFDM Receiver.](image)

Figure 4 shows the configuration of an OFDM receiver. At the receiver, received signal \( r(t) \) is filtered by a band pass filter, which is assumed to have sufficiently wide pass band to introduce only negligible distortion in the signal. An orthogonal detector is then applied to the signal where the signal is down converted to IF band. Then, an FFT circuit is applied to the signal to obtain Fourier coefficients of the signal in observation periods \([iT_{\text{Total}}, iT_{\text{Total}} + Ts]\). The BER depends on the level of the receiver’s noise. In OFDM transmission, the orthogonality is preserved and the BER performance depends on the modulation scheme in each sub channel.

### III. SYSTEM MODEL

#### A. Transmitter Model

The transmitter side of the BI – COFDM is shown in figure 5. [6]. The information bits \( \{u_k\} \) are first encoded by a convolutional encoder to produce a coded sequence \( \{c_k\} \). The coded sequence \( \{c_k\} \) is then mapped by the \( M \)-ary modulator into a symbol sequence. This symbol sequence is then passed through a serial to parallel converter, whose output is grouped into sets of \( N \) symbols \( \{S_0, S_1, \ldots, S_{N-1}\} \), where \( S_k \) belongs to \( M \)-ary constellation \( \Psi \). \( N \) is the number of subcarriers employed in OFDM and the symbol \( S_k \) is transmitted over \( k^{th} \) subcarrier. In order to generate the transmitted signal, an inverse discrete fourier transform (IDFT) is performed on the \( N \) symbols.

![Figure 5. Simplified block diagram of a Bit Interleaved Coded OFDM system.](image)
Typically $N$ is chosen to be the power of 2 and thus IDFT can implemented using Inverse Fast Fourier Transform (IFFT) algorithm. The IFFT yields the OFDM symbol consisting of sequence $\{s_0, s_1, ..., s_{N-1}\}$ of length $N$, where

$$s_k = \frac{1}{\sqrt{N}} \sum_{i=0}^{N-1} s_i e^{i 2\pi k i / N}, \quad 0 \leq k \leq N-1. \quad (3.1)$$

Assuming perfect synchronization and timing, the received symbols at the receiver are given by

$$r_k = s_k + i_k + g_k, \quad 0 \leq k \leq N-1 \quad (3.2)$$

### B. Impulsive Noise Channel Model

Note that the impulsive noise in equation (3.2) is separated in to two components: $i_k$ accounts for impulsive noise and $g_k$ represents Gaussian noise. In particular, $g_k$’s are independent and identically distributed (i.i.d) circularly symmetric complex Gaussian random variables with variance $\sigma_g^2$, whose probability density function is given by [2],

$$p_{g_k}(x) = \frac{1}{2\pi\sigma_g^2} \exp\left(-\frac{1}{2\sigma_g^2} x^2\right) \quad (3.3)$$

The impulsive noise variables $i_k$’s are also i.i.d with variance $\sigma_i^2$ and the p.d.f is given by Middleton’s class – A model as

$$p_{i_k}(x) = e^{-A}\delta(x) + \sum_{m=1}^{\infty} \frac{e^{-A m^2/2\sigma_i^2}}{m!} \exp\left(-\frac{1}{2\sigma_i^2} x^2\right) \quad (3.4)$$

with $\sigma_m^2 = \sigma_i^2 m/A$ and $\delta(\cdot)$ represents the Dirac delta function [3]. The parameter $A$ is called the impulsive index. For small values of $A$, only $1 - e^{-A} \approx 9.5\%$ samples are hit by impulses. So the noise is highly structured. A parameter $\Gamma = \sigma_g^2 / \sigma_i^2$ is defined as the Gaussian to impulsive noise power ratio for convenience.

### C. Iterative Receiver Model

![Figure 6. Iterative Receiver.](image)

**Operation of Receiver:**

The code word domain estimator neglects the dependencies among the code symbols $s_k$, while treating the i.i.d impulsive noise symbols $i_k$ correctly. Conversely, the information domain estimator neglects the dependence among the impulsive noise samples $I_k$, while treating the information symbols $S_k$ correctly. The idea is that by alternately using both estimators, at least to some extent, the whole statistical information provided by the received vector about the transmitted code word is exploited in the decoding process. This idea is similar to the idea of turbo decoding: here, instead of jointly decoding a concatenated code, two component-code decoders are used alternately. Each decoder exploits only the parity-check equations of one component code and neglects the dependencies introduced by the other. By alternately using both decoders, at least to some extent, the whole statistical information about the transmitted code word is exploited. The argument that the dependencies of one component code can be neglected is justified by using an interleaver. In our algorithm, the unitary transform $G$ plays the role of the interleaver. To convert the information between two domains, a transform matrix $G$ and its inverse matrix $G^{-1}$ are used as
shown in figure 6. [7]. In OFDM, the transform matrix $G$ is the inverse fourier transform matrix and is given by

$$G = \frac{1}{\sqrt{N}} \begin{bmatrix} e^{j\frac{2\pi}{N}0\times0} & e^{j\frac{2\pi}{N}0\times1} & \cdots & e^{j\frac{2\pi}{N}0\times(N-1)} \\ e^{j\frac{2\pi}{N}1\times0} & e^{j\frac{2\pi}{N}1\times1} & \cdots & e^{j\frac{2\pi}{N}1\times(N-1)} \\ \vdots & \vdots & \ddots & \vdots \\ e^{j\frac{2\pi}{N}(N-1)\times0} & e^{j\frac{2\pi}{N}(N-1)\times1} & \cdots & e^{j\frac{2\pi}{N}(N-1)\times(N-1)} \end{bmatrix}$$ (3.5)

The inputs to the code word domain estimator are the received symbol sequence $r$ and the estimated symbol sequence $\hat{s}$. With these inputs the code word domain estimator estimates the impulsive noise $\hat{i}$. The estimated impulsive noise $\hat{i}$ and the received code word are then transformed to the information domain as $\bar{I}$ and $R$ by multiplying them with the inverse transform matrix $G^{-1}$.

**Code word Domain Estimator:**

The inputs of the code word domain estimator are

$$r = s + i + g$$ (3.6)

and

$$\hat{s}^{(l-1)} = \beta_s^{(l-1)}s + \beta_i^{(l-1)}i + \beta_g^{(l-1)}g + d^{(l-1)}$$ (3.7)

where $r$ is the received vector, $\hat{s}^{(l-1)}$ is the estimated vector provided by the information domain estimator in the preceding iteration. $\beta_s^{(l-1)}$, $\beta_i^{(l-1)}$, $\beta_g^{(l-1)}$ are the scalar coefficients provided by the least square regression estimation. $d^{(l-1)}$ is the error term of the information domain estimator.

The high structure of the impulsive noise $i_k$ is the only non-Gaussian component in the code word domain and can be well distinguished from $s_k$, $g_k$ and $d_k$. Since the vectors $r$ and $\hat{s}$ are expected as i.i.d random variables, the MMSE estimation can be carried out for every $i_k$ and it is a function of $r_k$ and $\hat{s}_k$. The estimation of $i_k$ is as follows [6]:

$$\hat{i}_k = \frac{\sum_{m=0}^{\infty} \sum_{l=0}^{\infty} A_{mk} B_{ml} \text{Re}(r_{k+l}) \text{Im}(\hat{s}_{l}) \text{Re}(r_{k+l}) \text{Im}(\hat{s}_{l})}{\sum_{m=0}^{\infty} \sum_{l=0}^{\infty} A_{mk} B_{ml} \text{Re}(r_{k+l}) \text{Im}(\hat{s}_{l})}$$ (3.8)

where

$$\sigma_i^2 = \sigma_n^2 + \sigma_m^2 + \sigma_g^2$$ (3.9)

$$\sigma_s^2 = \beta_s^2 \sigma_s^2 + \beta_i^2 \sigma_m^2 + \beta_g^2 \sigma_g^2 + \sigma_D^2$$ (3.10)

$$\text{cov}(r, \hat{s}) = \beta_s \sigma_s^2 + \beta_i \sigma_m^2 + \beta_g \sigma_g^2$$ (3.11)

and the functions $b(r, \hat{s})$ and $\text{p}^\sim(r, \hat{s})$ are

$$b(r, \hat{s}) = 2 \sigma_m^2 [(\sigma_s^2 - \beta_i \text{cov}(r, \hat{s})) r + (\beta_i \sigma_s^2 - \text{cov}(r, \hat{s})) \hat{s}]$$ (3.12)

and

$$\text{p}^\sim(r, \hat{s}) = \frac{\exp\left(\frac{-\sigma_s^2 \sigma_m^2 - \text{cov}(r, \hat{s})}{\sigma_m^2 \sigma_i^2 + \text{cov}(r, \hat{s})^2}\right)}{\sigma_s^2 \sigma_m^2 + \text{cov}(r, \hat{s})^2}$$ (3.13)

The abbreviations used in the equation (3.8) are

$$\text{p}^{\sim}_{\text{Re}}(r, \hat{s}) = \text{p}^\sim(\text{Re}(r), \text{Re}(\hat{s}))$$ (3.14)

$$\text{p}^{\sim}_{\text{Im}}(r, \hat{s}) = \text{p}^\sim(\text{Im}(r), \text{Im}(\hat{s}))$$ (3.15)

where the operators $\text{Re}\{\cdot\}$ and $\text{Im}\{\cdot\}$ evaluate the real and imaginary parts of a complex number. Similarly the abbreviations for $b_{\text{Re}}(r, \hat{s})$ and $b_{\text{Im}}(r, \hat{s})$ are also used. In order to estimate the impulsive noise equation (3.8) is to be computed for every component $\hat{i}_k$.

**Information Domain Estimator:**

The information domain MMSE estimator takes $R$, $\hat{I}$ as inputs and estimates the code word.
\[ R = G^{-1}r \]  
(3.16)

and

\[ \hat{I}^{(l)} = \alpha_s^{(l)}S + \alpha_i^{(l)}I + \alpha_g^{(l)}Z + E^{(l)} \]  
(3.17)

where \( R \) is the received vector transformed into information domain, \( \hat{I} \) is the transformed vector of the estimated impulsive noise by the code word domain estimator. \( Z \) represents the transformed form of the background Gaussian noise. \( \alpha_s^{(l)} \), \( \alpha_i^{(l)} \), \( \alpha_g^{(l)} \) are the scalar coefficients given by the linear regression estimation.

The MMSE estimation of each \( S_k \) is given by [6]:

\[ \hat{S}_k = E(S_k | R_k, I_k) \]  
(3.18)

\[ = \frac{\Sigma_{k \in P_k} p(R_k, I_k, S_k) P(S_k)}{\Sigma_{k \in P_k} p(R_k, I_k) P(S_k)} \]  
(3.19)

Where \( p(R_k, I_k, S_k) \) is given as

\[ p(R_k, I_k | S_k) = \frac{\exp \left( -\frac{1}{2\sigma_s^2} (a - \text{cov}(R_k, I_k, S_k)) (a^* - \text{cov}(R_k, I_k, S_k)) \right)}{\sqrt{\pi \sigma_s \sigma_s^*}} \]  
(3.20)

The parameters \( a = R_k - S_k \), \( b = I_k - \alpha_s S_k \), and

\[ \sigma_s^2, \sigma_i^2, \sigma_g^2 \]  
(3.21)

\[ \text{cov}(R_k, I_k, S_k) = \alpha_i \sigma_i^2 + \alpha_g \sigma_g^2 \]  
(3.22)

\[ \text{cov}(R_k, I_k, S_k) = \alpha_i \sigma_i^2 + \alpha_g \sigma_g^2 \]  
(3.23)

\( P(S_k) \) are assumed to be equiprobable and they are cancelled in (3.18).

**Figure 7.** Pilot based Bit Interleaved Convolutionally Coded OFDM in impulsive noise.

The input data \( \{u_k\} \) is first encoded by a convolutional encoder to produce a coded sequence \( \{c_k\} \). The coded sequence \( \{c_k\} \) is then passed through a serial to parallel converter, whose output is grouped into sets of \( N \) symbols \( \{S_0, S_1, ..., S_{N-1}\} \), where \( S_k \) belongs to \( M \)-ary constellation \( \Psi \). \( N \) is the number of subcarriers employed in OFDM and the symbol \( S_k \) is transmitted over \( k \)-th subcarrier. In order to generate the transmitted signal, an inverse discrete fourier transform (IDFT) is performed on the \( N \) symbols.

After the signal is passed through the impulsive noise channel, the vector \( r \) is received at the receiver. First the impulsive noise is to be estimated by using the code word domain estimator as discussed in 3.3.2. Using the pilot symbols inserted at the transmitter the gaussian noise estimator nullifies the gaussian noise and FFT is implanted on that data. The output of FFT is demapped by an QPSK de-mapper followed by a convolutional decoder to get output data.
IV. RESULTS

Figure 8. Comparison of original impulsive noise and impulsive noise

Figure 9. Performance comparison of rate 1 by 3, ratehalf, without coding in gaussian noise environment.

Figure 10. Performance comparison of rate 1 by 3, ratehalf, without coding in impulsive noise environment.

Figure 11. Performance comparison of rate 1 by 3, ratehalf, without coding in impulsive noise and gaussian noise environment.
CONCLUSIONS AND FUTURE SCOPE

A novel combination of Coding techniques and OFDM is an attractive technique to combat with impulsive noise. The obtained simulation results show that, in the presence of impulsive noise convolutional coding improves the performance of OFDM based PLC systems significantly. In addition this combination minimizes the effect of impulsive noise.

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