Adaptive Permutation Coded Differential OFDM System for Power Line Communications

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***Abstract*—With a view to improving the capacity of the differential OFDM modulation scheme, specified in the narrowband power line communications (PLC) standards, in combating PLC channel associated noise, we hereby propose a permutation coded modulation scheme, which employs the hybridization of two kinds of DPSK (differential phase shift keying) modulations in an adaptive manner. The algorithm for deducing the encoded information from the hybridized modulations is described as well. This scheme is simulated and its performance is compared with a recently reported differential quinary PSK-OFDM system in the literature, whose behaviour has been shown to be better than the conventional permutation coded D8PSK-OFDM scheme at lower signal power to noise spectral density. Due to the simplicity in its encoding and decoding algorithms, this scheme is a good candidate for a number of low speed telemetry signaling in smart grids.**

Keywords*—Channel coding; G3-PLC; Differential phase shift keying; OFDM Modulation; Permutation Codes; Permutation coded DPSK-OFDM; Power Line Communications; PRIME

I. Introduction

Despite the advantages that power line communications (PLC) offers, it is however, plagued with severe channel noise, such as impulsive noise (IN), narrowband interference (NBI), frequency selective fading and additive white Gaussian noise (AWGN), which if not properly handled, can defeat the main purpose of using the existing power lines as a channel of communications. Attempts have been made to mitigate against some of these notorious noise, one of which is the reawakening of permutation coding (PC) by Vinck, when he proposed its use for PLC purposes [1]. PC is capable of combating NBI and IN in the narrowband PLC (NBPLC) channels [1]–[3]. The two NBPLC standards (i.e., PRIME and G3-PLC) have specified the use of an OFDM system, whose carrier modulation is done by the conventional differential phase shift keying (DPSK) [4], [5]. However, the effectiveness of this OFDM system is assisted with powerful channel coding schemes, such as Reed Solomon (RS) code and convolutional code (CC) proposed in the NBPLC standards [4], [5].

A special PC coded DPSK modulator, called the differential quinary PSK (DQuiPSK) modulator, as a component of OFDM system, was recently reported in [6]. This DQuiPSK modulator functions by constraining the number of its constellation points to macth the exact length of each PC word, which is 5 in this case. The scheme was devised, due to the interest of adapting PC into DPSK modulator of an order higher than 4. With a PC of codeword length 5, the most suitable DPSK modulator for mapping the PC symbols onto constellations, is the conventional D8PSK modulator. However, only 5 out of the 8 available constellation points are needed in the modulation, therefore 3 constellation points are left redundant, which in turn, may feature as *foreign symbol errors* (discussed later in Section II-B). That is why the DQuiPSK scheme was proposed. This scheme, however is unable to perform better than the conventional D8PSK scheme at extremely high E_b/N_o (signal power to noise spectral density), due to some possible *repetitive symbol errors* (discussed later in Section II-B). Therefore, by hybridizing these two OFDM components (i.e., DQuiPSK and D8PSK) in an adaptive manner, the entire system can have better performance, both at high and low E_b/N_o. A number of work involving adaptive modulation schemes has been reported in literature [7]–[9]. However, as far as we know, no one has reported an adaptive scheme consisting of DQuiPSK and D8PSK. Although the work is a conceptual design, its implementation is possible through the use of software defined radio hardware like the universal software radio peripheral (USRP), as done in [10] and [6].

Section II contains the foundational knowledge needed to understand the new concept presented in this study. Here, we first describe some of the notorious noise types associated with PLC, with brief descriptions of the noise models used in our simulation work. We then provide a brief overview of PC, DPSK and DQuiPSK, together with a detailed analysis of stregths of DQuiPSK and conventional DPSK at high and low E_b/N_o. After this, the concept of the adaptive modulation scheme is provided in Section III. The simulation work done is detailed in Section IV. Here, four different coded DPSK-OFDM schemes, including the proposed scheme, are simulated under AWGN, IN and NBI channel conditions. The outcomes of the simulations are presented in Section V. These are used to justify the usefulness of the proposed adaptive OFDM scheme. Section VI concludes the paper.

II. General Background

A. Power line channel associated noise

The narrowband power line channel is characterized with three different kinds of noise, namely background noise, which is usually referred to as AWGN, IN and frequency disturbance, which is regarded as NBI [11]–[13]. AWGN exists over the entire frequency region in the spectrum. Its effect decreases, as the transmission frequency increases and vice versa. This type of noise comes as a result of random processes like the flow of charges or thermal vibrations, which are normal phenomenon for any material at a temperature above absolute zero. As stated in [14] and [15], AWGN can be modeled using a Gaussian probability density function, $P(x)$ with power spectral density (PSD), σ^2_{AWGN} and mean 0, stated as:

$$P(x) = \frac{1}{\sqrt{2\pi\sigma^2_{\text{AWGN}}}} \mathrm{e}^{-y^2/2\sigma^2_{\text{AWGN}}} \tag{1}$$

NBI results from the interference caused by the connected equipments operating in the same frequency range as the PLC system. This noise type occurs in a narrow segment of the spectrum of operation, with time dependent amplitudes, which are usually above the floor level of AWGN [11]. A simplified model of NBI was presented in [16], where a parameter, P was used to define the probability of having NBI in an OFDM system. Some common sources of this noise, as stated in [15], [16] and [17], include the interference from amateur radios, AM transmissions and TV scanning frequencies.

IN is the noise type that has a flat broadband PSD, which can result into a multiple of large envelopes [16], [18], [19]. It can affect more than one frequency component in the transmitted data. Popular sources of this noise include household appliances like light dimmers, computers and hair dryers. As described in [15], a Markov model can be used to define IN. In this model, the strength of IN is defined by a parameter, T which is the ratio between the IN PSD, σ^2_{IN} and that of AWGN, σ^2_{AWGN}, given by:

$$T = \sigma^2_{\text{IN}}/\sigma^2_{\text{AWGN}} \tag{2}$$

B. Permutation coding and differential phase shift keying

PC is the process of representing codewords, each containing M non-repetitive symbols in a sequence. An example, where 5 bits are mapped onto 5 PC symbols is:

$$\left\{ \begin{array}{ccc} 00000 & 10000 & 01000 \\ \downarrow & \downarrow & \downarrow \\ 12340 & 13402 & 14023 \end{array} \right\} \tag{3}$$

As presented in [17], [18] and [20], a PC mapping can either be a distance increasing mapping (DIM), distance reducing mapping (DRM) or distance conserving mapping (DCM), using the general notation $Q(M, n, \delta)$. Here, M is the PC word length, n the number of bits to be mapped onto M PC symbols and δ is a small integer which defines the mapping type. If $\delta > 0$, the mapping is DIM, but it is a DRM if $\delta < 0$ and a DCM if $\delta = 0$.

In order to decode PC coded symbols at the receiving end, each received codeword is compared with all the possible codeword combinations in the PC sequence, and the word with the lowest Hamming distance (LHD) to the received codeword is selected as the decoded codeword.

Before passing through the channels, the PC coded symbols need to be modulated. In DPSK modulation, every successive baseband symbol is derived from the previously mapped symbol, hence the term *differential*. In other words, the symbols to be modulated are first differentially encoded, before mapping them onto constellations. For D8PSK, the constellation points $M_{\text{DP}} = 8$, while for DQuiPSK they are constrained to 5 [6]. In order to achieve good constellation mapping and demapping, the 5 constellation points needed for mapping the differentially encoded symbols are evenly distributed on the constellation graph, as depicted in Fig. 1 [6].

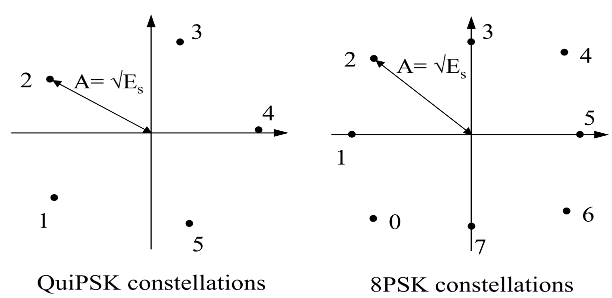

Fig. 1. Comparing QuiPSK constellations with 8PSK constellations

After demodulating the transmitted PC symbols, the wrongly demodulated symbols may result into what we term a *foreign symbol error* (FSE) and/or a *repetitive symbol error* (RSE), depending on the modulation type. FSE is the occurrence of a strange symbol, which ideally, should not be seen in a received codeword, while RSE is a situation where a certain symbol appears more than once in a received codeword. If a PC scheme is to be used in conjunction with a DPSK modulator, the PC codewords must have their symbols selected from a universal set U, whose elements are non-binary symbols between 0 and $M_{\text{DP}} - 1$. With this background information, we can mathematically explain the two terms FSE and RSE, using:

$$
\begin{aligned}
&e_{\text{pc}} \subset U, \\
&e_{\text{fse}} \subset U, \quad e_{\text{fse}} \cap e_{\text{pc}} = 0, \\
&e_{\text{rse}} \subset e_{\text{pc}}, \quad e_{\text{rse}} \cap e_{\text{fse}} = 0,
\end{aligned} \tag{4}
$$

where e_{pc} denotes the set of symbols to be permuted, so as to obtain all the codewords contained in the PC, while e_{fse} and e_{rse} are the elemental components of FSE and RSE respectively.

For our case study, the carrier modulator used in the OFDM system is D8PSK. Hence, M_{DP} = 8. The following thus gives the sample e_{pc} elements that may be used in the PC words, with U having symbols between 0 and 7:

$$
\begin{aligned}
&U = \{0,\ 1,\ 2,\ 3,\ 4,\ 5,\ 6,\ 7\}, \\
&e_{\text{pc}} = \{1,\ 2,\ 3,\ 4,\ 5\}
\end{aligned} \tag{5}
$$

Using the e_{pc} given in (5), the permuted codewords may be given as

$$
\left\{
\begin{array}{c}
12534, 21435, 13254, 24153, 21354, 12345, 23514 \\
23145, 15243, 51423, 25134, 53241, 41325, 21543 \\
31524, 35142, 14235, 12453, 34251, 54132, 42513 \\
32415, 34512, 43152, 54321, 52431, 45231, 35421 \\
52314, 45312, 43521, 53412
\end{array}
\right\} \tag{6}
$$

According to (4), possible e_{fse} elements are {0, 6, 7}, while e_{rse} are {1, 2, 3, 4, 5}. Hence, assumming the expected codeword at the receiver is {5 1 2 3 4}, due to FSE and RSE, we can demonstrate their effects on the received codewords using:

$$
\begin{aligned}
&\{5\ 7\ 2\ 3\ 0\}, \quad \text{due to FSE}, \\
&\{5\ 3\ 2\ 3\ 4\}, \quad \text{due to RSE}.
\end{aligned} \tag{7}
$$

According to DQuiPSK algorithm [6], its constellations are constrained to elements e_{pc}, as shown in (5). Hence, a system using DQuiPSK is only prone to RSE, which in effect, can make more than one codeword have the same LHD to the received codeword, and this may result in a decoding error. D8PSK is prone to both RSE and FSE, since all its constellation is the U elements. RSE is more likely to cause decoding errors than FSE. At very high E_b/N_o, a D8PSK system has a slimmer chance of having decoding errors than a DQuiPSK system. This is because the small amount of possible errors will be distributed between RSE and FSE error types, while for a DQuiPSK system the errors will be only RSE. Hence, a D8PSK system has a better behaviour at very high SNRs. This is why we have proposed the hybridization of the two systems in this research, in order to have better behaviours at both low and high SNRs.

C. Distance optimality for permutation coding

As done in [21], we use two matrices $\mathbf{E}$ and $\mathbf{E}^{(k)}$ to determine the distance optimality of a PC. $\mathbf{E}$ is a matrix containing elements $e_{i,j}$, representing the Hamming distance (HD) between permutation sequences, X_i, where $i = 1, 2, \ldots, |C|$, with $|C|$ being the cardinality of the PC. For example, according to (6), $X_1 = \{1\ 2\ 5\ 3\ 4\}$, and $X_2 = \{2\ 1\ 4\ 3\ 5\}$. Hence, by computation, the distance between X_1 and X_2 gives the $e_{i,j}$ element as $e_{1,2}$ = 4. By doing this for all the sequences in (6), $\mathbf{E}$ can be generated as a $|C| \times |C|$ matrix. Likewise, $\mathbf{E}^{(k)}$ is the distance matrix (with elements $e_{i,j}^{(k)}$) that is built by the symbols in position k, $1 \leq k \leq M$. So when the distances between sequences are computed, if there is a different symbol in position k for the sequences being considered, then that would contribute a 1 to $\mathbf{E}^{(k)}$. As such, M number of $\mathbf{E}^{(k)}$ matrices, whose dimensions are $|C| \times |C|$, are generated for a PC system. The magnitudes of these matrices, denoted by $|\mathbf{E}|$ and $|\mathbf{E}^{(k)}|$, can be represented by [21]:

$$
|\mathbf{E}| = \sum_{i=1}^{|C|} \sum_{j=1}^{|C|} e_{i,j} \quad \text{and} \quad |\mathbf{E}^{(k)}| = \sum_{i=1}^{|C|} \sum_{j=1}^{|C|} e_{i,j}^{(k)}. \tag{8}
$$

A PC is said to be distance optimal, if $|\mathbf{E}|$ is maximized, but to achieve that, all $|\mathbf{E}^{(k)}|$ need to be maximized. The upper bound on the distance that any PC mapping can attain is given by [20]:

$$
|\mathbf{E}_{\text{max}}| = M\left[2^{2n} - \left(2\alpha\beta + \beta + \alpha^2 M\right)\right], \tag{9}
$$

where $\alpha = \lfloor 2^n/M \rfloor$ and $\beta = 2^n \pmod{M}$.

According to (6), $|\mathbf{E}| = 4090$, and $|\mathbf{E}^{(1)}| = |\mathbf{E}^{(2)}| = |\mathbf{E}^{(3)}| = |\mathbf{E}^{(4)}| = |\mathbf{E}^{(5)}| = 818$, which is exactly the maxima obtainable. Hence, the PC is said to be optimal.

III. Adaptive modulation scheme

Fig. 2 depicts the proposed adaptive scheme. The data to be transmitted is first coded with an outer RS code and then with an inner non-binary PC, whose codeword length is 5, as presented in (6). A codeword length of 5 is proposed, because choosing a PC of word length equal to the modulation order, which is 8 in this regard, will require more complexities at the decoding end. The reason behind this complexity is that a codebook of codeword length 8 requires larger number of codewords than a codebook of codeword length 5, thereby resulting in an increased number of decision operations to be made by the receiver.

The mode of operation of our adaptive modulation scheme entails the use of threshold decision to adaptively decide which modulation/demodulation schemes are to be used out of the DQuiPSK and unconstrained D8PSK in the OFDM system.

By default, the threshold is set in favour of the DQuiPSK-OFDM scheme. In the adaptive modulator subsystem, the PC coded information is modulated, using a DQuiPSK-OFDM modulator, whose constellation points are constrained to match the codeword length.

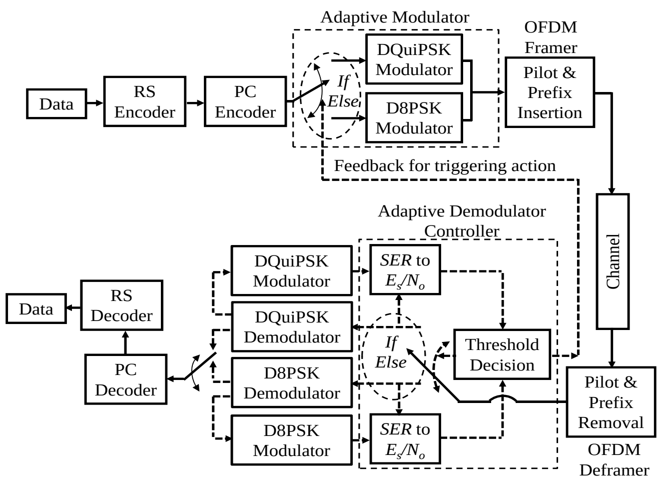

Fig. 2. Adaptive coded modulation scheme

At the receiving end, after the OFDM deframer operations, where the pilot and the cyclic prefix are removed, the received symbols are demodulated in order to obtain the transmitted PC symbols. However, in order to make use of the threshold detector for choosing the right choice of modulation scheme to be used for the subsequent train of data, our algorithm remodulates a copy of the demodulated symbols and compares the resulting modulated symbols with the received symbols (at the input of the demodulator). Here, due to channel effects, some errors are bound to occur in the comparison. The SER to E_s/N_o block converts the symbol error rate (SER) in the received symbols to the corresponding E_s/N_o values, using a relationship similar to (10) [22], [23].

$$SER \approx 2Q\left(\sqrt{2E_s/N_o}\sin\left(\pi/M\right)\right) \tag{10}$$

where E_s/N_o is the signal to noise ratio (SNR) of the MDPSK modulator and M the modulation order. Also, $E_s = 2E_b$. With this, E_b/N_o can be computed, provided the SER is known, and vice versa. Another approach to the SER estimation, is the use of the OFDM preamble, which is usually used for synchronization. With this, the preamble in the received signal is correlated with the copy of the preamble stored in the receiver's memory.

If the computed E_b/N_o is lower than a threshold Tr, the threshold detector chooses to use the DQuiPSK-OFDM scheme (see the *if else* ring on Fig. 2). Otherwise, the D8PSK-OFDM scheme shall be chosen. With this, an adaptive selection between the two schemes is attained, hence the term *adaptive-modulation*. This type of approach was employed in [7]. The resulting demodulated PC data then passes through the remaining decoding processes, until the original data is recovered. One good thing about the two schemes is their equal coding rates, since in each case, 5 RS bits are mapped onto 5 PC symbols. Also, since the PC symbols are directly mapped onto constellations, the two modulators have the same bit rates. Hence, switching between the two modulation schemes, by the aid of the threshold decision, does not incur any form of reduction in the bit rate, as opposed to the conventional adaptive modulation schemes, whose bit rates differ [7].

IV. Simulation Approach

The model represented in Fig. 2 was simulated under three common PLC channel associated impairments – IN, NBI and AWGN. For the NBI channel impairment, three values of NBI probability, $P = 1/48$, $P = 1/32$ and $1/16$ are used, employing the NBI model mentioned in Section II-A. These probabilities are extremely severe NBI cases. These have been used, in order to showcase the strength of the proposed scheme. More so, for IN, three values of the parameter, T, (described in Section II-A), have been employed. These T values include 0.2, 0.1 and 0.04.

In order to observe the strength of the proposed scheme, the conventional RS-CC-D8PSK-OFDM scheme (i.e., Scheme A from henceforth) proposed in the G3-PLC standard [5] and another two different PC coded modulation schemes, which are DQuiPSK-OFDM (i.e., Scheme B from henceforth) and D8PSK-OFDM (i.e., Scheme C from henceforth) were simulated alongside the proposed adaptive modulation scheme (i.e., Scheme D from henceforth), and the results are compared. The codewords used for the proposed scheme and the other two PC coded schemes, are the same as those presented in (6). These codewords have been adopted from [21], due to their optimized distance capability, with a minimum HD, $d_{\min}$ of 3. These codewords have better error correcting capability than those used in [6] (represented in (11) for the sake of emphasis), because they are not optimal and the $d_{\min}$ therein is 2.

$$\left\{\begin{array}{c} 51234, 51243, 51324, 51342, 51423, 51432, 52134 \\ 52314, 52143, 52341, 53214, 53241, 53421, 53412 \\ 53124, 53142, 41325, 41352, 41523, 41532, 41235 \\ 41253, 42135, 42153, 42315, 42351, 43125, 43215 \\ 43152, 43521, 43512, 43251 \end{array}\right\} \tag{11}$$

Table I contains the important specifications of all the four schemes considered in our simulations. By computation, the effective ratio of the coding rates of the simulated schemes is $R_A : R_B : R_C : R_D = 1.5 : 1 : 1 : 1$. Since Scheme A has a higher rate, its value has been compensated in the E_b/N_o computations, so as to ascertain fair comparisons. The input data is a random bit sequence generated in MATLAB, using the "*randi*" function.

V. Results

TABLE I. SPECIFICATIONS FOR SCHEMES A, B, C AND D

Scheme	Specifications
A	Codes: $(32, 24)$ RS & 1/2 rate CC with $\nu = 7$ and (117 155) Code gen. Modulator: D8PSK with 8 constellation points & OFDM with 64 FFT points and 16 CP length
B	Codes: $(32, 24)$ RS & PC of Q $(5, 5, 1)$ DIM & $d_{\text{min}} = 3$ Codewords: see (11) Modulator: D8PSK with 8 constellation points & OFDM with 64 FFT points and 16 CP length
C	Codes: $(32, 24)$ RS & PC of Q $(5, 5, 1)$ DIM & $d_{\text{min}} = 3$ Codewords: see (11) Modulator: DQuiPSK with 5 evenly distributed constellation points [6] & OFDM with 64 FFT points and 16 CP length
D	Codes: $(32, 24)$ RS & PC of Q $(5, 5, 1)$ DIM & $d_{\text{min}} = 3$ Codewords: see (11) Modulator: hybrid of DQuiPSK with D8PSK & OFDM with 64 FFT points and 16 CP length

The essence of the proposed Scheme D is revealed in the confusion rate curves generated for both Schemes B and C in Fig. 3, when using a combined AWGN+IN+NBI channel condition. As established in Section II-B, a DQuiPSK scheme, as in Scheme B, is prone to more confusions at very high E_b/N_o than a D8PSK scheme, as in Scheme C. Hence, this is why the adaptive demodulator controller needs to switch between the two Schemes B and C, in such situation. This preliminary result was used to determine the threshold Tr for switching between the two Schemes B and C. According to these results (Fig. 3), the performance of Scheme B crosses that of Scheme C at E_b/N_o value of 28 dB. Hence, the threshold value for the proposed algorithm is set at this value.

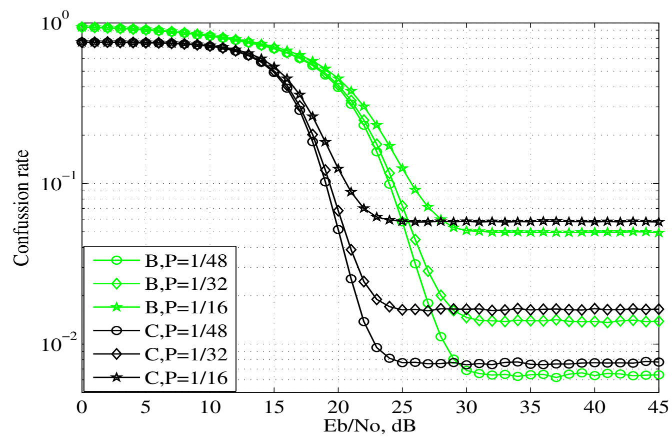

Fig. 3. Confusion rate curve for Schemes B and C, in the presence of AWGN+IN+NBI. P represents the NBI parameter. The value of T is fixed at 0.1

The simulation results obtained, when only an AWGN channel status is considered, is displayed in Fig. 4. According to these results displayed, there is a clear distinction between Scheme C and the two Schemes A and B. Its performance, both at high and low E_b/N_o, is better than Schemes A and B. At a BER of 5.5×10^{-4}, Scheme C has about 1.5 dB gain over Scheme A, and a 4 dB gain over Scheme B. This is due to the optimized codewords used, coupled with the fact that the DQuiPSK algorithm constrains the output constellation to 5, which in turn reduces the chances of having FSEs. A similar thing happens under a combined AWGN+IN condition, except that the introduced IN makes all the schemes perform relatively poor at extremely low E_b/N_o values, as Fig. 5 depicts. Under these two conditions (i.e., AWGN and AWGN+IN), the threshold Tr has a value less than the set value of 28 dB. Hence, the proposed Scheme D has the same behaviour as that of B, since the adaptive demodulator controller uses Tr to choose the better demodulator out of B and C. As such, the curve for Schemes D and B are seen to be overlapping.

The simulation results, when all the schemes are subjected to a combined AWGN+IN+NBI channel condition, are shown in Fig. 6. These outputs are similar to what was observed in [6], except with Scheme D, which is included in this study, coupled with the use of the optimal PC codewords.

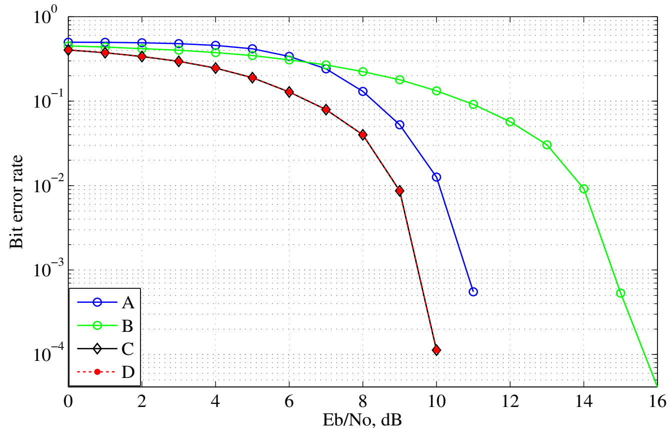

Fig. 4. Bit error rate curve for Schemes A, B, C and D, in the presence of AWGN

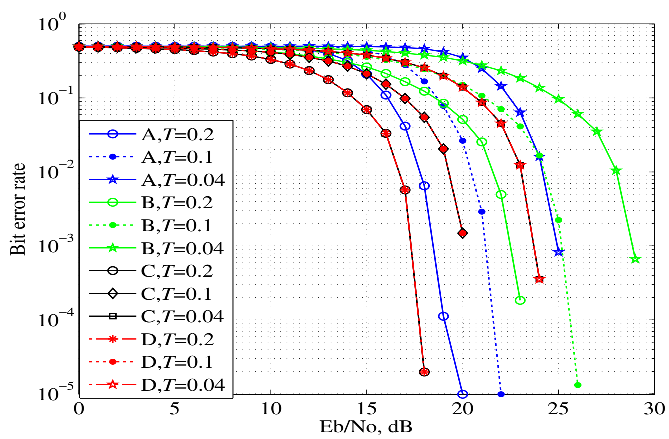

Fig. 5. Bit error rate curve for Schemes A, B, C and D, in the presence of AWGN+IN. T represents the IN parameter

At low E_b/N_o, Scheme B outperforms Scheme C, while the reverse is the case at high E_b/N_o. Scheme D thus switches from Scheme B to Scheme C at high E_b/N_o, by using the threshold descision in the adaptive demodulator controller. The switching points for Scheme D is so glaring at $E_b/N_o \geq 28$dB in all the NBI probabilities considered. Hence, the purpose of inventing Scheme D, in order to make the PC system have good behaviour at both high and low E_b/N_o, can said to be achieved.

VI. Conclusion

We have reported a special form of permutation coded modulation scheme, which is an adaptive hybridization of two differential modulation schemes, in an OFDM system. Its essence is to attain good performance, both at high and low E_b/N_o, when compared with the conventional schemes proposed in the NBPLC standards, under severe IN and NBI associated with PLC channels. Although adaptive modulation schemes have the disadvantage of introducing latency in a communication system, due to the decision computations like the threshold decision employed in this study, our scheme

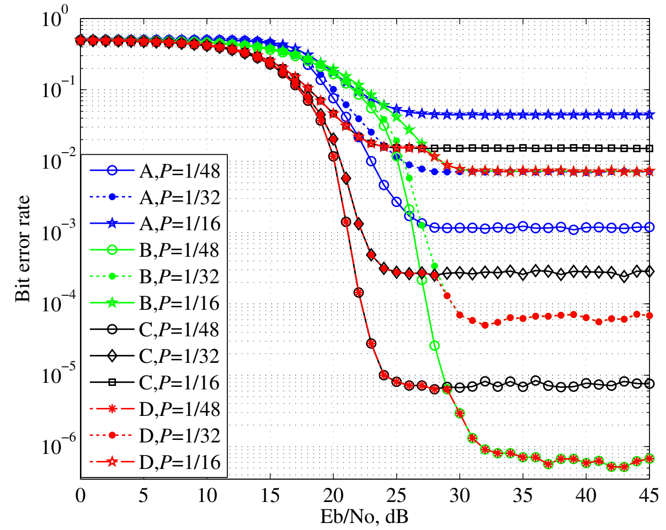

Fig. 6. Bit error rate curve for Schemes A, B, C and D, in the presence of AWGN+IN+NBI. P represents the NBI parameter. The value of T is fixed at 0.1

constitutes a good candidate for a number of applications in low speed communications used in smart grids, where emphasis are on control and command, for achieving good security and reliability. The next phase of this work is the implementation using the USRP.

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