Space-Time Block Codes with Rate ³ 1 for Power Line Channels

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Abstract-- **In this paper, the utilisation of space-time block codes (STBC) for high-speed communications over power line channels (PLC) affected by quasi-static flat Rayleigh fading and additive man-made asynchronous impulsive noise is presented. Although the diversity order is severely reduced in a 3-phase PLC environment with respect to wireless MIMO channels, our simulation results for a rate-1 real orthogonal STBC scheme demonstrate significant gains over an uncoded system operating with the same diversity advantage. In addition, taking advantage of the spatial orthogonality provided by the power line channel, a rate-3/2 real non-orthogonal space-time block code with reduced diversity order is constructed. It is shown that using the proposed high-rate code in a power line communication system yields good performance with very high spectral efficiency.**

I. INTRODUCTION

It is well known that the information capacity of a wireless communication system operating in a flat Rayleigh fading environment increases dramatically by employing multiple transmitting and receiving antennas [1], [2]. Often, multiple antennas are used at the receiver with some kind of scheme for combining the received signal, e.g. maximum ratio combining. However, in a wireless mobile scenario, it is difficult to efficiently use receiving antenna diversity at the mobile units, which need to be small, simple and cheap. Consequently, space diversity at the receiver has been almost exclusively used at the base station.

Recently, different transmit diversity techniques have been introduced to benefit from spatial diversity also in the downlink while putting the diversity burden on the base station. Space-time trellis coding (STTC) has been proposed [3], which combines signal processing at the receiver with coding techniques appropriate to multiple transmitting antennas. However, when the number of transmitting antennas is fixed, the decoding complexity of space-time trellis codes (measured by the number of trellis states in the decoder) increases exponentially with the transmission rate. In addressing the issue of decoding complexity, Alamouti discovered a remarkable scheme for transmission using two transmitting antennas [4]. For the same number of antennas this scheme is much less complex than space-time trellis coding, but there is a loss in performance compared to STTC. Despite this performance penalty, Alamouti's scheme is still appealing in terms of simplicity and performance. Space-time block coding (STBC), introduced in [5], generalises the transmission scheme discovered by Alamouti to an arbitrary number of transmitting antennas and is able to achieve the full diversity promised by the number of transmitting and receiving antennas.

In this paper we analyse the utilisation of space-time block codes for the transmission of data over the power line channel (PLC). The emphasis of the paper will centre upon short distance data transmission over 3-phase 415V power lines in the presence of asynchronous impulsive noise. In section II we present the impulsive noise model for such a channel. In section III we describe how to adapt the STBC technique to the power line environment, and present simulation results using a real orthogonal design with transmission rate 1. Taking advantage of the spatial orthogonality naturally provided by the PLC, a new rate-3/2 real non-orthogonal space-time block code is presented in section IV, which yields good performance with very high spectral efficiency. Section V presents our conclusions and final comments.

II. IMPULSIVE NOISE MODEL

Besides signal distortion, due to cable losses and (possible) multipath propagation, noise is the most crucial factor influencing digital communications over power line networks. Opposite to many other communication channels, the power line channel (PLC) does not represent an additive white Gaussian noise (AWGN) environment; in the frequency range from some hundred kilohertz up to 20 MHz it is mostly dominated by narrow-band interference and impulsive noise. In this work we are particularly interested in the asynchronous impulsive noise caused by switching transients in the network. The impulses have durations of some microseconds up to a few milliseconds with random arrival times. The psd of this type of noise can reach values of more than 50 dB above the background noise. The occurrence of such impulses may cause bit or burst errors in data transmission.

One suitable model for this type of noise is the complex memoryless additive white Class A noise (AWCN) channel model [6]. For the complex channel the class A pdf is given by

$$p_{n_k}(x) = \sum_{m=0}^{\infty} \frac{\alpha_m}{2\pi\sigma_m^2} \exp\left(-\frac{|x|^2}{2\sigma_m^2} \right) \qquad (1)$$

with $$\alpha_m = e^{-A} \frac{A^m}{m!} \qquad (2)$$

and complex valued argument x. The variance σ_m^2 is defined as

$$\sigma_m^2 = \sigma^2 \frac{(m/A)+T}{1+T} \qquad (3)$$

where σ^2 is the variance of the Class A noise. The Class A noise model combines an additive white Gaussian noise component g_k with variance σ_g^2 and an additive impulsive noise component i_k with variance σ_i^2 [7]. Therefore, the parameter $T = \sigma_g^2/\sigma_i^2$ is used in the Class A noise model. The second model parameter A is called the impulsive index. For small A, say A = 0.1, we get highly structured (impulsive) noise whereas for $A \to \infty$ the pdf becomes Gaussian. The variance σ_m^2 of x is determined by the channel state m = 0, 1, 2, 3, … using equation (3). Since the Class A noise is memoryless, the states are taken independently for every noise sample with probability $P(m)$ = α_m, which can be interpreted as a worst-case scenario to model impulsive noise on a power-line [7]. The channel state is unknown to the observer of the process and therefore its pdf is given by the expectation over all states.

III. SPACE-TIME BLOCK CODING FOR POWER LINE CHANNELS

In order to design an efficient STBC transmission scheme for the power line channel, we must take into account some important differences with respect to the wireless channel: 1) in an ST wireless system with multiple transmitting and receiving antennas, the signal received at antenna j (j = 1,…,m), at any time t, is the result of the combination of the signals emitted by the n transmitting antennas, which is not the case in the power line channel where each phase ideally provides an isolated path to the transmitted signal; 2) this isolation in turn provides a natural decoupling between any two signals corresponding to different symbols; 3) the fading effects are quite different from those that usually prevail in a wireless environment; and 4) the power line channel does not represent an AWGN environment as previously mentioned. With this in mind, and assuming we are using the three phases to obtain spatial diversity, the first difference means that the maximum spatial diversity order we can achieve is three. The second difference implies that assuming perfect isolation between the phases of the power line, we do not need to construct our transmission matrix using an orthogonal design [5]. Finally, differences 3 and 4 obviously imply that we have to employ a different channel model to evaluate the performance of the STBC scheme. We next consider a communication system with 3 emitting points at the transmitter and 3 receiving points at the receiver (Figure 1). Here, we use the terms *emitting points* and *receiving points* instead of the terms *transmitting antennas* and *receiving antennas* used in wireless communications.

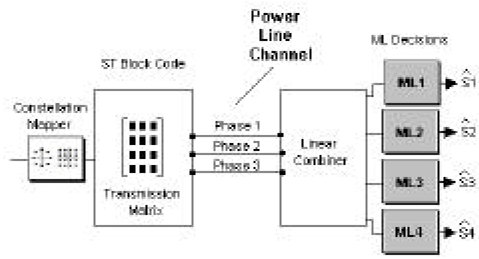

Fig. 1. Block diagram of a power line communication system that uses space-time block coding at the transmitter and linear combining/ML detection at the receiver.

As an example of our STBC construction, we have chosen a 4-PAM system. Using this modulation scheme the space-time block code can achieve unity transmission rate. An M-PAM scheme is also a useful example to investigate the effect of impulsive noise in the channel due to the unequal energies of the signals in the set.

A. Encoding

A space-time block code (STBC) is defined by a p x n transmission matrix G. The entries of the matrix G are linear combinations of the real variables $x_1, x_2, \ldots, x_k$ for the case of *real orthogonal designs*, or of the complex variables $x_1, x_2, \ldots, x_k$ and their conjugates for the case of *complex orthogonal designs* [5]. We can use a real orthogonal design when transmission at the baseband employs a real signal constellation such as M-PAM. However, for complex constellations like M-QAM or M-PSK, we must use a complex orthogonal design. For example, $G_c^{(2)}$ represents a code that utilises two emitting points and is based on a complex orthogonal design. This is Alamouti's scheme [4] and is defined by

$$G_c^{(2)}(z_1, z_2) = \begin{bmatrix} z_1 & z_2 \\ -z_2^* & z_1^* \end{bmatrix} \qquad (4)$$

Similarly, $G_r^{(3)}$ represents a space-time block code based on a real orthogonal design that utilises three emitting points.

$$G_r^{(3)}(x_1, x_2, x_3, x_4) = \begin{bmatrix} x_1 & x_2 & x_3 \\ -x_2 & x_1 & -x_4 \\ -x_3 & x_4 & x_1 \\ -x_4 & -x_3 & x_2 \end{bmatrix} \quad (5)$$

Since M-PAM modulation has a real signal constellation, we can use real-valued variables for the ST transmission matrix. Therefore, using the transmission matrix of (5), and considering a real constellation A of size 2^b (b = 2 for 4-PAM), the encoding algorithm proceeds as follows: 1) at time slot 1, kb bits arrive at the encoder and select k = M = 4 real symbols s_1, s_2, s_3, s_4; 2) the encoder populates the transmission matrix $G_r^{(3)}$ by setting x_i = s_i, and 3) at time slots t = 1, 2, 3, 4, the signals G_{t1}, G_{t2}, G_{t3} are transmitted simultaneously from emitting points 1, 2, 3. For example, at t = 1, $G_{11} = s_1$, $G_{12} = s_2$, $G_{13} = s_3$. At t = 2, $G_{21} = -s_2$, $G_{22} = s_1$, $G_{23} = -s_4$, and so on.

The rate R of this coding scheme is defined to be kb/pb, which is equal to k/p. In this case, k = 4 and p = 4, and the rate R = 1.

Note that for complex constellations, Alamouti's scheme [4] is the only case where we can achieve rate one. For more than two emitting points, complex orthogonal designs allow for rates less than one [5].

B. The decoding algorithm

At time t the signal r_t^i, received at the receiving point i, is given by

$$r_t^i = \boldsymbol{a}_{i,i} s_t^i + \boldsymbol{h}_t^i \quad (6)$$

where the path gains $\boldsymbol{a}_{i,i}$ for the power line channel are modelled as samples of independent zero-mean complex Gaussian random variables with variance 0.5 per real dimension (flat Rayleigh fading). The communication channel is assumed to be quasi-static so that the path gains are constant over a frame of length p and vary from one frame to another. The samples $\boldsymbol{h}_t^i$ of additive white Class A noise (AWCN) are independently identically distributed (i.i.d.) complex random variables according to Middleton's Class A noise model [6].

Maximum-likelihood detection at the receiver amounts to minimising the following decision metric, which takes into account only *three* signal paths

$$\Big(|r_1^1 - \boldsymbol{a}_{1,1}s_1|^2 + |r_2^1 + \boldsymbol{a}_{1,1}s_2|^2 + |r_3^1 + \boldsymbol{a}_{1,1}s_3|^2 + |r_4^1 + \boldsymbol{a}_{1,1}s_4|^2 +$$
$$|r_1^2 - \boldsymbol{a}_{2,2}s_2|^2 + |r_2^2 - \boldsymbol{a}_{2,2}s_1|^2 + |r_3^2 - \boldsymbol{a}_{2,2}s_4|^2 + |r_4^2 + \boldsymbol{a}_{2,2}s_3|^2 +$$
$$|r_1^3 - \boldsymbol{a}_{3,3}s_3|^2 + |r_2^3 + \boldsymbol{a}_{3,3}s_4|^2 + |r_3^3 - \boldsymbol{a}_{3,3}s_1|^2 + |r_4^3 - \boldsymbol{a}_{3,3}s_2|^2 \Big) \quad (7)$$

Expanding the above metric and deleting the terms that are independent of the code words, the maximum-likelihood detection rule leads to the decision variables

$$R_1 = \left(r_1^1 \boldsymbol{a}_{1,1}^* + r_2^2 \boldsymbol{a}_{2,2}^* + r_3^3 \boldsymbol{a}_{3,3}^* \right) \quad (8)$$
$$R_2 = \left(-r_2^1 \boldsymbol{a}_{1,1}^* + r_1^2 \boldsymbol{a}_{2,2}^* + r_4^3 \boldsymbol{a}_{3,3}^* \right) \quad (9)$$
$$R_3 = \left(-r_3^1 \boldsymbol{a}_{1,1}^* - r_4^2 \boldsymbol{a}_{2,2}^* + r_1^3 \boldsymbol{a}_{3,3}^* \right) \quad (10)$$
$$R_4 = \left(-r_4^1 \boldsymbol{a}_{1,1}^* + r_3^2 \boldsymbol{a}_{2,2}^* - r_2^3 \boldsymbol{a}_{3,3}^* \right) \quad (11)$$

and a decision in favour of s_i among all the symbols s of the signal constellation A is made if

$$s_i = \arg\min_{s \in S} |R_i - s|^2 + \left(-1 + |\boldsymbol{a}_{1,1}|^2 + |\boldsymbol{a}_{2,2}|^2 + |\boldsymbol{a}_{3,3}|^2 \right) s^2 \quad (12)$$

From (8) – (11) it can be seen that the diversity order for this scheme is 3.

C. Simulation results

In our computer simulation, the average energy of the 4-PAM signal constellation was scaled so that the average energy of the constellation points is one. Therefore, the variance $\boldsymbol{s}^2$ of the Class A noise at the input of each receiving point is 1/2SNR. The inter-arrival times of impulse noise were generated using a homogeneous Poisson process, with the state occurrence probabilities given by $\boldsymbol{a}_m$. The variance $\boldsymbol{s}_m^2$ for each possible channel state m is calculated using equation (3). We used a truncated version of (1) with states m = 0, 1, 2, and 3, which provides a very good approximation, and the class A noise parameters A = 0.1 and T = 10^{-3}. The performance results of the space-time code $G_r^{(3)}$ were compared with those obtained using an uncoded system operating with the same diversity advantage.

Figure 2 shows the Bit Error Rate versus SNR (dB) for both systems. It is seen that at the bit error rate of 10^{-3} the STBC scheme with the space-time code $G_r^{(3)}$ gives a gain of about 13 dB over the use of an uncoded system.

Figure 3 shows the BER performance of both schemes in flat Rayleigh fading and additive white Gaussian noise (AWGN). When these results are compared with those obtained in Figure 2, the performance degradation resulting from the presence of impulsive noise in the power line channel is evident.

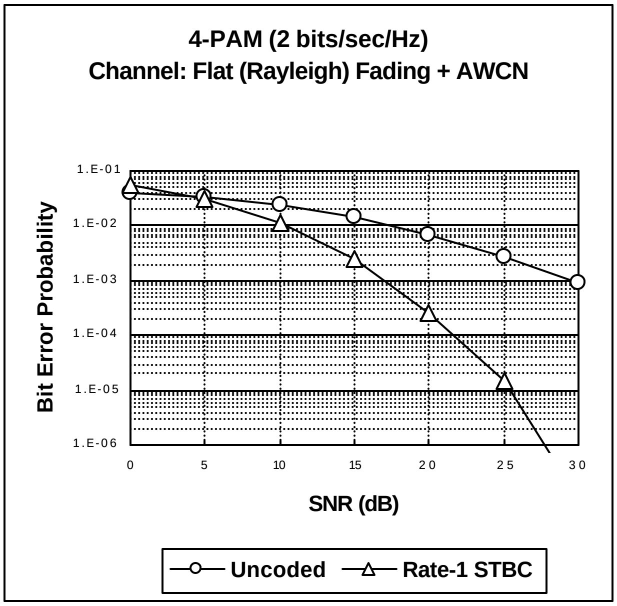

Fig. 2. Bit error probability (BER) versus SNR in the presence of flat Rayleigh fading and additive white Class A noise (AWCN) for the uncoded and rate-1 STBC schemes.

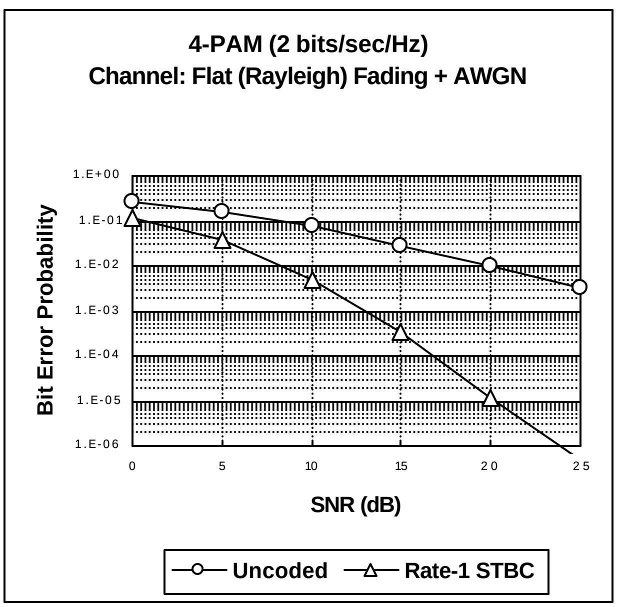

Fig. 3. Bit error probability (BER) versus SNR in the presence of flat Rayleigh fading and additive white Gaussian noise (AWGN) for the uncoded and rate-1 STBC schemes.

IV. RATE-3/2 NON-ORTHOGONAL STB CODE

As was pointed out in section II, the 3-phase power line channel provides isolation between the transmission paths. Therefore, the inherent spatial orthogonality of this channel allows us to remove the restrictions imposed in the transmission matrix by the theory of orthogonal designs [5]. In particular, it is now possible to trade off bandwidth for signal-to-noise-ratio. In other words, we are able to increase the symbol rate by reducing the achievable diversity order. In order to do this, however, the space-time block code has to be based on a non-orthogonal design. Consequently, in this section we present a new real non-orthogonal space-time block code with rate 3/2 and diversity order of 2. For a 4x3 transmission matrix like the one in (5), this is the maximum rate that can be achieved without completely losing the diversity gain provided by an STBC scheme.
Denoting the new code by $N_r^{(3)}$, its transmission matrix is expressed by

$$N_r^{(3)}(x_1, x_2, x_3, x_4, x_5, x_6) = \begin{bmatrix} x_1 & x_6 & x_2 \\ x_2 & x_1 & x_3 \\ x_4 & x_3 & x_5 \\ x_5 & x_4 & x_6 \end{bmatrix} \qquad (13)$$

A. Encoding

For the code (13) and 4-PAM modulation, the block of bits arriving at the encoder at time slot 1 is kb = 6x2 = 12 bits. The 4-PAM modulator maps the 12 bits onto 6 symbols that populate the transmission matrix (13) by setting $x_i = s_i$. Each row of the matrix is then transmitted at a specified time epoch from the three emitting points (phases), or what is the same, each column is transmitted from a specified emitting point at four time epochs.

B. Decoding

Assuming the same channel conditions of (6), maximum-likelihood detection at the receiver amounts to minimising the following decision metric

$$\Big(|r_1^1 - a_{1,1}s_1|^2 + |r_2^1 - a_{1,1}s_2|^2 + |r_3^1 - a_{1,1}s_4|^2 + |r_4^1 - a_{1,1}s_5|^2 +$$
$$|r_1^2 - a_{2,2}s_6|^2 + |r_2^2 - a_{2,2}s_1|^2 + |r_3^2 - a_{2,2}s_3|^2 + |r_4^2 - a_{2,2}s_4|^2 +$$
$$|r_1^3 - a_{3,3}s_2|^2 + |r_2^3 - a_{3,3}s_3|^2 + |r_3^3 - a_{3,3}s_5|^2 + |r_4^3 - a_{3,3}s_6|^2 \Big) \qquad (14)$$

Expanding the above metric and deleting the terms that are independent of the code words, the maximum-likelihood detection rule leads to the decision variables

$$R_1 = \left(r_1^1 \alpha_{1,1}^* + r_2^2 \alpha_{2,2}^*\right) \quad (15)$$
$$R_2 = \left(r_2^1 \alpha_{1,1}^* + r_1^3 \alpha_{3,3}^*\right) \quad (16)$$
$$R_3 = \left(r_3^2 \alpha_{2,2}^* + r_2^3 \alpha_{3,3}^*\right) \quad (17)$$
$$R_4 = \left(r_3^1 \alpha_{1,1}^* + r_4^2 \alpha_{2,2}^*\right) \quad (18)$$
$$R_5 = \left(r_4^1 \alpha_{1,1}^* + r_3^3 \alpha_{3,3}^*\right) \quad (19)$$
$$R_6 = \left(r_1^2 \alpha_{2,2}^* + r_4^3 \alpha_{3,3}^*\right) \quad (20)$$

Thus the minimisation of the original decision metric is equivalent to minimising the following decision metrics

$$\left|R_1 - s_1\right|^2 + \left(-1 + \left|\alpha_{1,1}\right|^2 + \left|\alpha_{2,2}\right|^2\right)\left|s_1\right|^2 \quad (21)$$

for detecting s_1, the decision metric

$$\left|R_2 - s_2\right|^2 + \left(-1 + \left|\alpha_{1,1}\right|^2 + \left|\alpha_{3,3}\right|^2\right)\left|s_2\right|^2 \quad (22)$$

for detecting s_2, the decision metric

$$\left|R_3 - s_3\right|^2 + \left(-1 + \left|\alpha_{2,2}\right|^2 + \left|\alpha_{3,3}\right|^2\right)\left|s_3\right|^2 \quad (23)$$

for detecting s_3, the decision metric

$$\left|R_4 - s_4\right|^2 + \left(-1 + \left|\alpha_{1,1}\right|^2 + \left|\alpha_{2,2}\right|^2\right)\left|s_4\right|^2 \quad (24)$$

for detecting s_4, the decision metric

$$\left|R_5 - s_5\right|^2 + \left(-1 + \left|\alpha_{1,1}\right|^2 + \left|\alpha_{3,3}\right|^2\right)\left|s_5\right|^2 \quad (25)$$

for detecting s_5, and the decision metric

$$\left|R_6 - s_6\right|^2 + \left(-1 + \left|\alpha_{2,2}\right|^2 + \left|\alpha_{3,3}\right|^2\right)\left|s_6\right|^2 \quad (26)$$

for detecting s_6.

From (15) – (20), it can be seen that the diversity order for this scheme is 2. When compared with the rate-1scheme of section III, it is obvious that the diversity loss for the new code is 2/3, which exactly compensates for the increase in the symbol rate from 1 to 3/2.

C. Simulation results

The performance results of the non-orthogonal space-time code $N_r^{(3)}$ were compared with those obtained using an uncoded system operating with the same diversity advantage. Figure 4 shows the BER performance of both systems in flat Rayleigh fading and additive white Class A noise (AWCN).

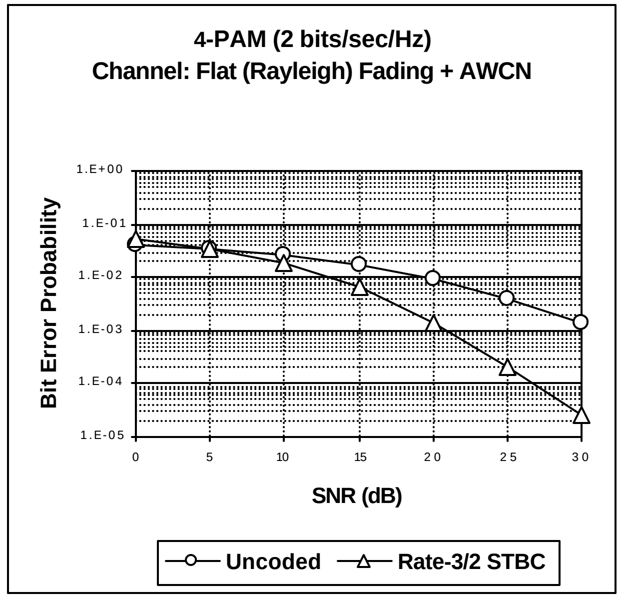

Fig. 4. Bit error probability (BER) versus SNR in the presence of flat Rayleigh fading and additive white Class A noise (AWCN) for the uncoded and rate-3/2 STBC schemes.

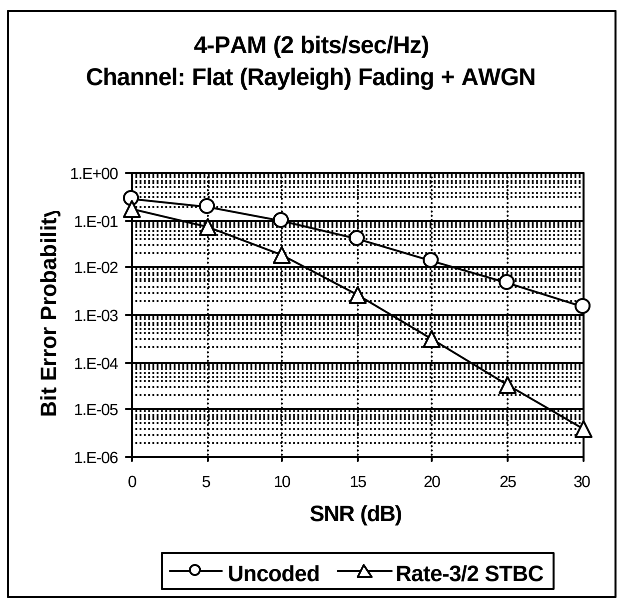

Fig. 5. Bit error probability (BER) versus SNR in the presence of flat Rayleigh fading and additive white Gaussian noise (AWGN) for the uncoded and rate-3/2 STBC schemes.

By observing the performance results depicted in Figures 2 and 4 for the codes $G_r^{(3)}$ and $N_r^{(3)}$, respectively, we note that the effect of reducing the diversity order is to shift the $N_r^{(3)}$ curve 3 dB to the right. In spite of this 3-dB penalty, $N_r^{(3)}$ still provides a gain of 10 dB over the use of an uncoded system for a BER value of 10^{-3}, and the additional advantage of a greater symbol rate.
Figure 5 shows the BER performance of both schemes in flat Rayleigh fading and additive white Gaussian noise (AWGN), which represents a less rigorous environment. As before, we observe a 3-dB penalty in the error performance of $N_r^{(3)}$ when compared with $G_r^{(3)}$.

V. CONCLUSIONS

The utilisation of space-time block coding techniques for high-speed data communications over 3-phase power line networks has been proposed in this paper. It is shown that the noise scenario in power line channels is not of the AWGN-type, and that the asynchronous impulsive noise caused by switching transients in the network can considerably affect the system performance.
We have described the design of real orthogonal codes with rate 1 and diversity order 3 for this type of channel. It is shown that these space-time block-coding schemes outperform an uncoded system operating with the same diversity advantage by about 13 dB. It is also shown that assuming perfect isolation between the phases of the power line the space-time transmission matrix needs no orthogonal design. Consequently, we have constructed a new real non-orthogonal space-time block code with rate 3/2 and diversity order 2.
There is a 3-dB penalty for this reduction in the diversity gain. In spite of this, the new STBC scheme still provides a gain of 10 dB over the use of an uncoded system for a BER value of 10^{-3}, and the additional advantage of a greater symbol rate. This scheme will be particularly useful for transmitting digital information over power line channels with both bandwidth and power restrictions [8].

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