Bit Error Rate Simulation Studies for PSSS with Multi-User Detection for Industrial Multipath-Fading Environments

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Abstract—Wireless connectivity offers industrial automation the flexibility, mobility and reconfigurability it requires to cope with the new challenges of industry 4.0. Communication in industrial applications like factory automation (e. g. closed-loop control) imposes ambitious demands on wireless communication systems as these applications require very low latencies at ultra-high reliabilities. Additionally, the industrial environment encourages harsh channel conditions due to numerous metallic reflectors in typical factory halls. Since state of the art wireless technologies are not able to completely fulfill these requirements especially in industrial wireless channels, novel approaches are being developed. In that regard, Parallel Sequence Spread Spectrum (PSSS) transmission is seen as a promising technology as it allows transmission with very short latencies. Nevertheless, its bit error rate performance has not yet been analysed for typical industrial wireless channels. This paper presents a PSSS system concept and two associated receiver structures which might be suitable for the targeted applications. A detailed mathematical description of the transmission of the PSSS signal in an Additive White Gaussian Noise (AWGN) channel. The performance of the system approach is analysed in industrial multipath-fading channels. Simulations show that the bit error rate curves of both receiver structures under consideration vary enormously for the considered channels. It can be shown that a multi-user detection receiver achieves better BER results than a single-user matched filter, but further ways of improvement need to be analysed.

I. INTRODUCTION

Nowadays, modern wireline fieldbuses are used for industrial communication. On the one hand, they provide very low latencies and high reliabilities in combination with a deterministic behaviour. On the other hand, they have limitations in flexibility, support of mobility and reconfigurability. Future industrial automation systems require a maximum degree of flexibility, mobility and reconfigurability as well as they need to be highly interconnected. This can only be enabled with wireless communication.

Generally, any wireless technology replacing a wireline fieldbus needs at least to provide a comparable performance in data transmission. Especially for applications in factory automation where requirements can be extraordinarily high with cycle times below 1ms at a Packet Error Rate (PER) down to $10^{-9}$, performance guarantees are essential to avoid unwanted down-times of machines. To make it more challenging, wireless communication systems face really harsh channel conditions in an industrial environment due to the numerous metallic reflectors in a factory hall, fast movement of machine parts and interferences by other manufacturing processes. Currently existing wireless technologies like e. g. WirelessHART [1] or WISA [2] are not able to provide the demanded robustness in the required cycle times.

In this paper we present and analyse a Code Division Multiple Access (CDMA) - based system approach for utilisation in highly demanding closed-loop control applications. It uses Parallel Sequence Spread Spectrum (PSSS) [3], [4], [5] because of its flexible transmission scheme and great potential to offer short latencies and high reliability. PSSS was standardized in IEEE 802.15.4.2011 for Low Rate-Wireless Personal Area Networks (LR-WPANs). We thoroughly describe the transmission in a time-dispersive channel and consider two types of receiver structures - the Single-User Matched Filter (SUMF) and the decorrelator receiver which is a Multi-User Detection (MUD) receiver.

The remainder of the paper is organized as follows: the following section describes the PSSS system concept and gives an equation of the received signal in an Additive White Gaussian Noise (AWGN) channel. In Sec. III the transmission via a time-dispersive channel is mathematically described and simplifications for PSSS are shown. Section IV presents the considered receiver structures. Numerical results together with a description of the assumed channel model are given in section V. A conclusion is drawn in section VI.

II. THE PSSS SYSTEM CONCEPT

When CDMA is applied, each transmit symbol is spread over time by multiplication with a signature sequence. This allows several users to transmit simultaneously in one frequency band and enables user separation on the receiver side by correlation. We consider a CDMA system with PSSS [3],
[4], which applies special pseudorandom binary sequences as signature sequences.

A. Signature Sequences

For PSSS the signature sequences \( g_k[i] \) for the users indexed by \( k = 0, 1, \ldots, K-1 \) are given as cyclically time shifted versions \( g[i-k] \) of one single \( m \)-sequence [6] \( g_0[i] \equiv g[i], i = 0, 1, \ldots, L-1 \) of length \( L \). This is in contrast to the conventional Direct Sequence Spread Spectrum (DSSS) approach where the users are characterised by different \( m \)-sequences. Obviously, PSSS requires synchronous transmission to separate the users at the receiver side.

As a consequence of the fact that one shifted \( m \)-sequence is used as spreading sequences, the crosscorrelations are described by the autocorrelation function [6] and they are given by

\[
\langle g_k | g_l \rangle = \begin{cases} 
1 & : k = l \\
-\frac{1}{L} & : k \neq l 
\end{cases}
\]  

(1)

In contrast to the application of orthogonal sequences like Walsh-Hadamard, the crosscorrelations are not 0 for \( k \neq l \). We used the notation

\[
\langle x | y \rangle = \sum_{i=0}^{L-1} x_i^* y_i
\]

for the scalar product of two complex vectors \( x \) and \( y \), and the signature vectors are normalised to the length \( |g_k| = \sqrt{\langle g_k | g_k \rangle} = 1 \).

B. Transmission in an AWGN Channel

To describe the PSSS receive signal under the influence of an AWGN channel we first arrange the signature sequences in column vectors of a \( L \times K \) matrix \( G \) with elements \((G)_{ik} = g_k[i]\). The discrete AWGN channel receive signal vector of one time slot can then be written as

\[
r = Gs + n,
\]

(2)

where \( n \) is a complex AWGN vector with variance \( N_0/2 \) in each real dimension, and the complex (e.g. M-PSK or M-QAM) modulation symbols \( s_k \) with average symbol energy \( E_{s} = E_s \) are arranged to a column vector \( s = (s_0, s_1, \ldots, s_{L-1})^T \). For \( M \) bits per symbol, the bit energy \( E_b \) is related to the symbol energy by \( E_s = \log_2(M)E_b \).

Although they are not perfectly orthogonal, the good autocorrelation properties of the \( m \)-sequences keep the multi-access interference (MAI) between the users rather small. This can be seen by writing the correlator output at the receiver for user number 0 as the scalar product

\[
\langle g_0 | r \rangle = s_0 + I_0 + n
\]

(3)

with an AWGN term \( n = \langle g_0 | n \rangle \) and a \( K \)-user MAI term

\[
I_0 = -\frac{1}{L} \sum_{k=1}^{K-1} s_k.
\]

(4)

For sufficiently large values of \( K \), according to the Central Limit Theorem, the sum of independent identically distributed (i.i.d.) random variables (RVs) can be modelled as a Gaussian RV. For 2-PSK with \( s_k = \pm \sqrt{E_b} \), the variance of this real-valued RV is given by

\[
\sigma^2_{MAI} = \frac{K-1}{L^2} E_b.
\]

(5)

For a 4-PSK, the same variance has to be added for the imaginary part. This equation shows that even for a fully loaded system with \( K = L \), the MAI effect becomes very small for increasing signature sequence length. For an AWGN channel this is one of the benefits of the PSSS spreading strategy.

C. The Cyclic Prefix

A time-dispersive channel leads to the superposition of delayed versions of the transmit signal. To avoid intersymbol interference between adjacent time slots, we apply a technique that is familiar to OFDM systems [7] and add a cyclic prefix (also called guard interval) to the transmit signal of each time slot. Assuming that the cyclic prefix is sufficiently long to absorb all relevant echoes that are caused by the channel, one can reduce the analysis to a single time slot and replace all convolutions by cyclic convolutions. In the following, we neglect the guard interval in the equations due to simplicity. However, one must keep in mind that the cyclic prefix slightly reduces both the spectral and the power efficiency.

III. TRANSMISSION IN A TIME-DISPERSIVE CHANNEL

A. The Continuous Channel Model

Before establishing a discrete model for time-dispersive channels, we have first to go back to the time-continuous channel. The continuous signature signals are defined by

\[
g_k(t) = \sum_{i=0}^{L-1} g_k[i] p(t - iT),
\]

(6)

where \( T \) is the chip pulse period and \( p(t) \) is the chip pulse which is assumed to be a Nyquist pulse with the property

\[
[p^*(t) * p(t)]_t=\pm T = \delta[i].
\]

(7)

Typically, sqrt-raised-cosine (RRC) pulses with roll-off factor \( \alpha (0 \leq \alpha \leq 1) \) are used which leads to an RF signal bandwidth

\[
B = \frac{1 + \alpha}{T}.
\]

(8)

The continuous transmit signal is given by

\[
s(t) = \sum_{k=0}^{K-1} s_k g_k(t).
\]

(9)

A static time dispersive channel is described by a linear time-invariant (LTI) system that is given by its impulse response \( c(t) \). For simplicity, we consider only a channel where
\(c(t)\) is the same for all users. The received signal for this channel is given by
\[
    r(t) = c(t) * s(t) + n(t) \tag{10}
\]
\[
    = \sum_{k=0}^{K-1} s_k h_k(t) + n(t), \tag{11}
\]
where \(n(t)\) is continuous AWGN with two-sided noise density \(N_0/2\), and the channel-signature signals
\[
h_k(t) = c(t) * g_k(t) \tag{12}
\]
are the channel outputs of the signature signals \(g_k(t)\). We note that for the non-dispersive channel with \(h_k(t) = g_k(t)\) Equation (11) leads back to the discrete model of Equation (2) when we identify the pulse matched filter (PMF) outputs \([p(-t)^* r(t)]_{t=T}\) with the components of the vector \(r\).

B. The Fractionally Spaced Discrete Channel Model

A set of sufficient statistics is given by the matched filter outputs \([h_k^c(-t)^* r(t)]_{t=0}\) which is just the set of scalar products of the receive signal \(r(t)\) with the base signals \(h_k(t)\) that span the space of all the possible transmit signals \(c(t) * s(t)\), see e.g. [8], [7]. For a general impulse response \(c(t)\), this space is not spanned by the chip pulse base \(p(t - iT)\) and, thus, the (PMF) outputs \([p(-t)^* r(t)]_{t=T}\) do not form a set of sufficient statistics because the channel matched filter (CMF) \(c^c(-t)\) is not included. As it is practically not possible to implement \(c^c(-t)\) or \(h_k^c(-t)\) by analogue filters, we apply fractionally spaced sampling at \(t = iT/2\) for a digital implementation with sampling frequency \(f_s = 2/T\) at twice the chip rate. This is the same principle known from equalisers where fractionally spaced \((T/2-)\) equalisers (see, e.g. [9], [10]) are applied to cope with the same problem. Since the base signals \(h_k(t)\) are limited to the bandwidth \(B\) given by Equation (8), they are completely characterised by their samples
\[
h_k[i] = \sqrt{T/2} h_k(iT/2). \tag{13}
\]
While proceeding this way, we have implicitly assumed that the signature signals have been periodically continued due to the introduction of a cyclic prefix. Therefore, the discrete channel-signature signals \(h_k[i]\) are \(2L\)-periodic in the time index \(i\). We arrange them in a \(2L \times K\)-matrix \(H\) given by its elements \([H]_{ik} = h_k[i]\) and write \(h_k\) for its column vectors. Due to the sampling theorem, the channel of Equation (11) can thus equivalently be described by its discrete version
\[
r = Hs + n = \sum_{k=0}^{K-1} s_k h_k + n, \tag{14}
\]
where \(n\) is a discrete AWGN vector with variance \(N_0/2\) in each real dimension and the elements \(r[i]\) of the received column vector \(r\) are related to the samples of the receive signal \(r(t)\) (after ideal low-pass filtering) by
\[
r[i] = \sqrt{T/2} r(iT/2). \tag{16}
\]
From Equation (14), a set of sufficient statistics can be obtained by the scalar products with the channel-signature base vectors \(h_k\) which can be expressed as sum
\[
(h_k[r]) = \sum_{i=0}^{2L-1} h_k^c[i] r[i] \tag{17}
\]
or, due to Parseval’s Equation, as an integral
\[
(h_k[r]) = \int_0^{2LT} h_k^c(t) r(t) dt. \tag{18}
\]
It is worth mentioning that, in analogy to Equation (12) for the continuous signals, the channel-signature base vectors \(h_k\) can be expressed as cyclic convolutions (denoted by \(\oplus\))
\[
h_k = c \oplus \tilde{g}_k, \tag{19}
\]
where the elements
\[
c[i] = \frac{1}{2} T c(iT/2) \tag{20}
\]
of the vector \(c\) are the (scaled) \(T/2\)-spaced channel samples and the elements
\[
\tilde{g}_k[i] = \sqrt{T/2} g_k(iT/2) \tag{21}
\]
of the vector \(\tilde{g}_k\) are the (scaled) \(T/2\)-spaced samples of the signature signals. Again, we have implicitly assumed that the signature signals have been periodically continued due to the introduction of a cyclic prefix that is long enough to absorb the delays of the channel. To apply the sampling in Equation (20), we assume that the channel impulse response \(c(t)\) already includes an ideal low-pass filter of bandwidth \(f_s/2 = 1/T\) with impulse response
\[
w(f_s) = \text{sinc}(tf_s). \tag{22}
\]
We note that much smoother than rectangular filter flanks are possible if the roll-off factor is smaller than \(\alpha = 1\). For a \(N\)-path channel with delay times \(\tau_m\), amplitudes \(a_m\), and phases \(\phi_m\), the channel coefficients \(c[i]\) given by Equation (20) can be written as
\[
c[i] = \sum_{m=1}^{N} a_m \text{e}^{j\phi_m} w(i - \tau_m f_s). \tag{22}
\]
We note that due to the band-limiting, the peaks \(\delta(t - \tau_m)\) in the infinite-bandwidth model must be replaced by \(f_s w(f_s(t - \tau_m))\). In practice, the coefficients \(c[i]\) according to some (stochastic) model for the parameter triples \((\tau_m, a_m, \phi_m)_{m=1}^{N}\) can be generated and stored prior to the actual system simulations. We shall discuss an example in Section V.

C. Simplifications for PSSS

As a consequence of the definition of PSSS, the channel-signature sequences \(h_k[i]\) for the different users are cyclically time shifted versions \(h[i - 2k]\) of one single sequence \([6]\)
\[
h_0[i] = h[i] \text{ of length } L. \tag{17}
\]
This has the nice consequence that
for a fully loaded system \((K = L)\), Equation (15) can be written as a cyclic convolution

\[
\mathbf{r} = \mathbf{h} \otimes \tilde{s} + \mathbf{n} \tag{23}
\]

\[
= \mathbf{c} \otimes \mathbf{g} \otimes \tilde{s} + \mathbf{n}, \tag{24}
\]

where \(\tilde{s}\) is the up-sampled symbol vector of length \(2L\) given by

\[
\tilde{s} = (s_0 \ 0 \ s_1 \ 0 \ s_2 \ 0 \ \ldots \ 0 \ s_{2L-1} \ 0)^T.
\]

IV. RECEIVER STRUCTURES

The first step in a CDMA receiver is to correlate the received signal with the reference transmit signals according to Equation (17). For single user detection, the signals of all other users are considered as noise while multi-user detection eliminates the interferences from all other users by some kind of equalisation. In this paper we consider a Single-User Matched Filter (SUMF) and a decorrelator receiver, which is a Multi-User Detection (MUD) receiver.

A. The Single-User Matched Filter Receiver

The SUMF receiver calculates the MF output \(\langle \mathbf{h}_k | \mathbf{r} \rangle\) of Equation (17) corresponding to the channel-signature vector for the user under consideration. The channel-signature vector contains the channel influence and the signature sequence. The remote station user in the downlink has to perform this only for its own signature vector corresponding to one branch in the upper part (a) of Figure 1, thereby treating the contributions of the other users as interference. As an example, for user number \(K - 1\), this results in

\[
\langle \mathbf{h}_{K-1} | \mathbf{r} \rangle = |\mathbf{h}_{K-1}|^2 s_{K-1} + \tilde{I}_{K-1} + n
\]

with an AWGN term \(n = \langle \mathbf{h}_{K-1} | \mathbf{n} \rangle\) and a \(K\)-user MAI term

\[
\tilde{I}_{K-1} = \sum_{k=1}^{K-2} \langle \mathbf{h}_{K-1} | \mathbf{h}_k \rangle s_k. \tag{26}
\]

In contrast to the AWGN channel MAI term \(I_0\) given by Equation (4), this MAI term \(\tilde{I}_{K-1}\) may become very critical. Consider, e.g., a 2-path channel with \(\tau_1 = 0\) and \(\tau_2 \approx T\). Then, depending on the amplitudes and phases of the second path, the interference term \(\langle \mathbf{h}_{K-1} | \mathbf{h}_{K-2} \rangle\) due to user number \(k = K - 2\) may reach a similar magnitude as \(|\mathbf{h}_{K-1}|^2\) and may thus severely corrupt the reception for user number \(K - 1\). We present an example for this in Section V.

B. The Decorrelator Receiver

Multi-User Detection receivers [11] utilise the information of the received signals for all users to obtain an estimate of the transmit signal for one single user. This means that they take into account the SUMF outputs \(\langle \mathbf{h}_k | \mathbf{r} \rangle\) of all branches in the upper part of Figure 1. We arrange them to the column vector \(\mathbf{H}^T \mathbf{r}\), where \(\mathbf{H}^T\) is the hermitian conjugate (i.e. complex conjugate and transposed) of the matrix \(\mathbf{H}\) defined by its elements \((\mathbf{H}^T)_{ki} = (\mathbf{H})_{ik}^*\). The upper part (a) can now be characterised more compactly as shown in the lower part (b) of the same figure.

\[
\text{Figure 1. Many parallel single-user matched filter (SUMF) receivers. The upper part (a) shows the individual branches. The lower part (b) shows the compact representation by a vector-matrix product.}
\]

The decorrelator is a linear receiver that can be interpreted geometrically by skew projections [12] of the receive vector \(\mathbf{r}\) of Equation (15) on the base \(\{\mathbf{h}_k\}_{k=0}^{K-1}\) that spans the vector space of transmitted signals. In the noise-free channel, the decorrelator outputs \(x_k\) are just the coordinates \(s_k\) of the transmitted signal. One can show [11], [7] that these projections are obtained by

\[
x = \mathbf{H}^+ \mathbf{r}, \tag{27}
\]

where

\[
\mathbf{H}^+ = (\mathbf{H}^\dagger \mathbf{H})^{-1} \mathbf{H}^\dagger \tag{28}
\]

is the Moore-Penrose pseudo-inverse matrix. We observe that the decorrelator simply multiplies the SUMF receiver output of Figure 1 (b) by the matrix \((\mathbf{H}^T \mathbf{H})^{-1}\), see the upper part (a) of Figure 2. From Equation (14) we conclude that the decorrelator vector is given by

\[
x = \mathbf{s} + \mathbf{\tilde{n}}, \tag{29}
\]

where

\[
\mathbf{\tilde{n}} = \mathbf{H}^+ \mathbf{n} \tag{30}
\]

is a correlated noise vector with autocorrelation matrix

\[
\mathbb{E}\{\mathbf{\tilde{n}} \mathbf{\tilde{n}}^\dagger\} = N_0 (\mathbf{H}^\dagger \mathbf{H})^{-1}. \tag{31}
\]
This structure is depicted in the lower part (b) of Figure 2. The decorrelator is followed by a (PSK or QAM) demodulator that makes decisions on its outputs $x_k$ to yield the estimated bit stream.

For a fully loaded PSSS scheme, the decorrelator becomes even more simple due to the cyclic structure of the signature sequences. To see this, we write $h^{-1}$ for the inverse (in the sense of cyclic convolutions) of $h$ and define

$$\hat{x} = h^{-1} \ast r.$$ (32)

From Equation (23) we conclude that

$$\hat{x} = \hat{s} + h^{-1} \ast n.$$ (33)

Down-sampling by taking every second sample of $\hat{x}$ yields Equation (29). We have thus seen that for a fully loaded PSSS system, the setup of Figure 2 is equivalent to the setup of Figure 3, and the decorrelator receiver is equivalent to a deconvolution (i.e. an inverse convolution) by the $h^{-1}$ which inverts the channel and the base signature sequence.

V. NUMERICAL RESULTS

We evaluate the bit error rate (BER) performance of a PSSS system with a chip period of $T = 50 \text{ ns}$ and a roll-off factor $\alpha = 0.5$ by simulation. According to Equation (8), this leads to a bandwidth of $B = 30 \text{ MHz}$. We consider a PSSS system with sequence length $L = 255$ and number of used sequences $K = 240$ which equals a load of 94%.

Based on primal experiences with channel measurements in industrial environments performed by Bosch, we assume a static 2-path Rayleigh fading channel as an extreme case for our simulations. This means that we set $N = 2$ in Equation (22), and the amplitudes $a_1, a_2$ are i.i.d. RVs that follow the Rayleigh distribution. The total average receive power is normalised to $E\{a_1^2\} + E\{a_2^2\} = 1$, where $E\{\cdot\}$ denotes the statistical expectation value. The phases $\phi_1, \phi_2$ are uniform i.i.d. RVs. We consider a fixed delay $\tau_2 - \tau_1 = 40\text{ ns}$ between the two paths which correspond to a run-length difference of 12 m. The physical reasoning for such a model is an obstructed (e.g. by knife edge bending) LOS path that is superimposed by a strong reflection from (e.g.) a metallic wall of the factory hall. We assume that the reflection is 3 dB weaker than the first component, i.e. $E\{a_2^2\} = \frac{1}{4}E\{a_1^2\}$. We generate an ensemble of 50 such independent static Rayleigh channels and study their BER performance curves for a BPSK modulation. The red plot in Figure 4 shows the average BER curve over all Rayleigh channels for the SUMF receiver together with the theoretical curve for BPSK in the AWGN channel as a reference. The average BER curve is drawn as a thick red curve with circle markers. Additionally, it is important to also analyse the performance for the individual Rayleigh channels as industrial applications need their required reliability at all times to avoid machines going into an error-state. Therefore, the average BER performance is compared with the BER curves for the individual Rayleigh channels. The maximum deviation from the average BER curve is shown in terms of the red-coloured error bars. We observe very strong statistical fluctuations between the BER curves for the different individual randomly generated Rayleigh channels. This is due to the fact that the channel is very flat: Because the runtime delay is in the order of the chip period, there is not much statistical fluctuation inside the bandwidth for

![Figure 3](image-url) Multiuser reception with the decorrelator implemented by the inverse cyclic convolution.

![Figure 4](image-url) Average BER curves for the SUMF and the MUD receiver in a 2-path Rayleigh channel. Maximum deviation of BER curves for the individual fading channels to the average curve is indicated by error bars.

![Figure 5](image-url) Average BER curves for the SUMF and the MUD receiver in a 2-path Rice channel. Maximum deviation of BER curves for the individual fading channels to the average curve is indicated by error bars.
each individual channel, but very high fluctuations between them. The average performance curve for the same 50 channels with a decorrelator MUD receiver is shown in Figure 4 by the blue plot. Again the maximum deviation of the individual BER curves from the average is indicated by error bars (blue-coloured). We observe as an effect of this receiver that the BER curves become much steeper. However, due to the flatness of the channel, some channels with a very poor BER performance occur.

For the next simulations, we consider a Rice channel with an undisturbed LOS component. It represents the most general channel and was also observed in our measurements. The LOS component has a fixed amplitude $a_1$ followed by a weak Rayleigh path with $E(a_2^2) = \frac{1}{4} a_1^2$ which corresponds to the Rice factor $K_{\text{Rice}} = 4$. The delay is again $\tau_2 - \tau_1 = 40$ ns. The red plot in Figure 5 shows the average BER curve for the SUMF receiver. Again, the red-coloured error bars indicate the maximum deviation of the BER performance for the individual Rayleigh channels from the average performance. Some improvements can be observed, but the performance is far from being satisfactory even for this weak delay path. The average BER curve with maximum deviations for the decorrelator MUD receiver is shown in Figure 5 by the blue plot. We observe an significant performance improvement. Recalling the discussion in Section IV and the fact that the delay time $\tau_2 - \tau_1 = 40$ ns is close to the chip period $T = 50$ ns, we note that the SUMF reception is severely disturbed by the second path whereas the decorrelator MUD receiver constructively combines both paths.

We have performed several other simulations with increasing Rice factor that show even better BER performance, especially for the MUD receiver. However, one should keep in mind that one can not rely on an unobstructed strong LOS path. Future evaluations of channel measurements may show the conditions the system has to cope with. We point out that the above simulations show a system without channel coding which is known to be very important especially for fading channels. Furthermore, one may think of antenna diversity to overcome the problem of flat fading channels.

VI. DISCUSSION AND CONCLUSION

When wireless fieldbus systems shall be replaced by wireless connectivity in factory automation, the requirements on latency and reliability are extremely high. In this paper, we presented and analysed a CDMA approach based on PSSS which looks promising in this context. The transmission of a PSSS signal over an AWGN channel and a time-dispersive channel was mathematically described. Two kinds of receiver structures, a SUMF and a decorrelator receiver, were analysed. The PSSS performance was numerically evaluated for both receiver types under the assumption of a two-path channel model with Rayleigh or Rice distribution. The channel parameters were chosen based on our first experiences with channel measurements in industrial environments. In each simulation, we generated 50 independent static Rayleigh and Rice channels. The simulation results show that there is a large variety in system performance for both the SUMF and the decorrelator receiver. Due to the flatness of the channel there are large statistical fluctuations between the BER curves. The performance of a SUMF receiver is not good enough to achieve a BER below $10^{-5}$ in the presence of channel echoes no matter a Rayleigh or a Rice channel is considered. The decorrelator receiver is better as the BER curves get steeper especially for the channels where the performance is bad. In a Rayleigh channel the required $E_b/N_0$ can be much higher than 20dB. In a Rice channel an $E_b/N_0$ below 20dB is needed to achieve a BER of $10^{-5}$.

We conclude that one has to think about ways of improvements also for a decorrelator receiver, to meet the performance requirements of systems in factory automation. This is not only the case for a PSSS system as the performance of other systems, like e. g. OFDM, will also suffer from a flat transmission channel. Further experience with industrial wireless channels is needed. In our simulations we did not consider channel coding, which can partly reduce the influence of a flat channel. Nevertheless, to fully unlock the potential of channel codes and achieve large improvements in BER performance, a combination with interleaving is necessary. This will negatively affect the latency. A second option to overcome the problem of flat fading channels is antenna diversity which requires further investigations.

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