Coded N-ary PPM UWB impulse radio with chaotic time hopping and polarity randomisation

Y.-P. Hong, S.-Y. Jin and H.-Y. Song

A coded N-ary pulse position modulated (PPM) ultra-wide bandwidth (UWB) impulse radio which exploits the chaotic time hopping and the polarity randomisation is considered. It is shown that the proposed system has a noise-like spectrum and its multi-user system is discussed. The bit error rate performance is confirmed by simulation based on the UWB indoor channel model.

Introduction: There has been much interest in chaotic communications, the main premise of which is that broadband signals generated by simple deterministic systems with chaotic dynamics can potentially replace pseudorandom carrier signals widely used in modern spread-spectrum communications [1]. A chaotic pulse position modulation (CPPM) was proposed in [2], which encodes the information in a pulse train by alteration of the time positions of pulses. This system avoids the difficulties, e.g. sensitivities to distortions and noise, of most other chaos-based communication schemes by using chaotically, that is aperiodically, timed pulse sequences rather than continuous chaotic waveforms. It belongs to the general class of time hopping (TH) ultra-wide bandwidth (UWB) impulse radio (IR) communications. A pseudo chaotic time hopping (PCTH) UWB-IR was proposed in [3], which combines the chaotically timed pulse sequences such as CPPM with the framed-time structure. Its structure is an N-ary PPM UWB-IR combined with a PCTH code based on the symbolic dynamics of a chaotic map. This radio is intended to coexist with conventional narrow bandwidth systems by both working in lower signal-to-noise ratio (SNR) at the same data rate and bit error rate (BER) due to the code and having the enhanced spread-spectrum characteristics, i.e. a reduced number of line spectrums, rather than conventional TH UWB-IRs due to a chaotic time hopping sequence removing periodic structures from the signal. As an alternative to PCTH, an interleaved convolutional time hopping UWB-IR was discussed in [4]. This radio replaces the PCTH code by an optimum convolutional code (from the viewpoint of free distances) to improve BER performance preserving the line spectrum properties. Unfortunately, this radio does not have a coding gain compared to the uncoded system because of using not the soft but the hard Viterbi decoder. In this Letter, we consider a coded N-ary PPM UWB-IR, which exploits the chaotically timed pulse sequences in the framed-time structure such as PCTH and polarity randomisation for the coexistence.

System structure: We consider the coded N-ary PPM UWB-IR shown in Fig. 1. This radio uses a convolutional code, the output linear equations of which are linearly independent. We can improve BER performance by using a better convolutional code rather than the PCTH code. Note that $N = 2^M$ and $T_f = N \times T_s$, where T_f is frame time and T_s is slot time. The symbol, $d_k' \in \{0, 1, \ldots, N-1\}$, can be determined by reading each successive M output bits, b_n, in decimal. We assume that the user data are independent identically distributed (i.i.d.) discrete uniform random variables (RVs). Then, d_k' are not i.i.d. because of a correlation between successive terms but uniformly improve security by the linear independence of output bits. The time slot index, d_k, can be i.i.d. discrete uniform RVs by interleaving d_k'. The transmitted signal is given by

$$s_{tr}(t) = \sum_{i=-\infty}^{\infty} p_i \times w(t - iT_f - d_i T_s) \tag{1}$$

where $p_i \in \{-1, 1\}$ is a polarity randomisation sequence and $w(t)$ is a transmit pulse. In this Letter, we use the second derivative of the Gaussian function for $w(t)$. The output of the correlator in the receiver for a uth time slot in a kth frame time is given by

$$m_{k,u} = \int_{kT_f+(u-1/2)T_s}^{kT_f+(u+1/2)T_s} r(t) \times w(t - kT_f - uT_s)dt \tag{2}$$

A vector $\bar{m}_k = (m_{k,0}, m_{k,1}, \ldots, m_{k,N-1})$ is deinterleaved and $m_{k,u}$ is used as the soft branch metric of a branch, the output of which is u in decimal in the following soft Viterbi decoder. We note that not the hard but soft Viterbi decoding has a coding gain in the coded N-ary PPM UWB-IR.

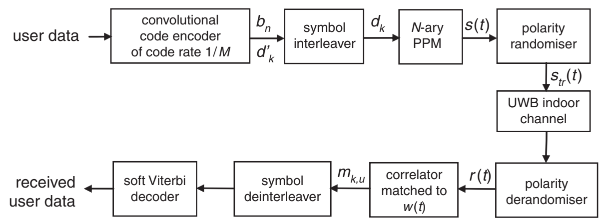

Fig. 1 *Coded N-ary PPM UWB-IR*

Line spectrum properties: Without the polarity randomiser, the line spectrums of the transmitted signal, $s(t)$, would exist at every $1/T_s$ Hz. These are the same line spectrum properties as those of the PCTH UWB-IR and these sparse lines can be set to fall outside the useful bandwidth by design [3]. Signals that cause line spectrums are highly undesirable since they have a high peak power spectral density (PSD) function. They hardly guarantee the low probability of detection (LPD) and have to back off the average power until the peak of the spectrum complies with the spectral mask established by the Federal Communication Commission (FCC). After polarity randomisation, the transmitted signal, $s_{tr}(t)$, is similar to that of joint PPM/PAM but the sign of the transmit pulse does not bear any information. The PSD function of an N-ary modulated signal with an i.i.d. input sequence is given by

$$\begin{aligned}\Phi(f) = \frac{1}{T^2}\sum_{n=-\infty}^{\infty}\left\{\left|\sum_{i=0}^{N-1} P_i S_i\left(\frac{n}{T}\right)\right|^2 \delta\left(f - \frac{n}{T}\right)\right\} \\ + \frac{1}{T}\left\{\sum_{i=0}^{N-1} P_i |S_i(f)|^2 - \left|\sum_{i=0}^{N-1} P_i S_i(f)\right|^2\right\}\end{aligned} \tag{3}$$

where T is a symbol period, $S_i(f)$ is the Fourier transform of an ith symbol, $s_i(t)$, of the constellation, and P_i is the marginal probability of the ith symbol. We assume that the polarity randomisation sequence, which is actually a long pseudorandom sequence, is i.i.d. uniform RVs. The transmitted signal, $s_{tr}(t)$, can then be treated as a pulse train with equiprobable $2N$ antipodal symbols. From (3), the PSD function of $s_{tr}(t)$ is as follows:

$$\Phi_{s_{tr}}(f) = \frac{1}{T}\sum_{i=0}^{2N-1} P_i |S_i(f)|^2 = \frac{1}{T_f}|W(f)|^2 \tag{4}$$

where $T = T_f$, $s_i(t) = w(t - iT_s)$, $s_{i+N}(t) = -s_i(t)$, and $W(f)$ is the Fourier transform of $w(t)$. Thus, the proposed system has no line spectrums and the shape of its PSD function is determined by the transmit pulse, $w(t)$. When T_s is fixed, we can get a lower PSD function by increasing N, i.e. increasing T_f. This results in compliance with the FCC mask, an improved LPD, and an improved BER performance in the case of orthogonal PPM. Fig. 2 shows the actual PSD of the system by Matlab.

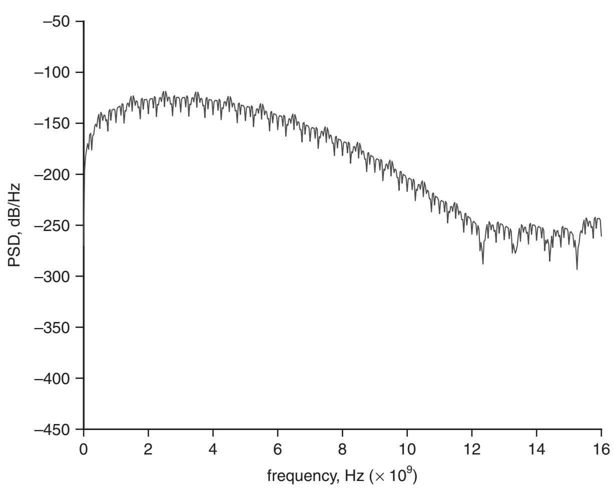

Fig. 2 *Actual PSD functions for N = 32, T_f = 32 ns, T_s = 1 ns, E_s = 1 W/Hz*

Multi-user system: In multi-user communications, an lth user uses his own sequence of pulses for the transmit pulse, given by

$$w^{(l)}(t) = \sum_{i=0}^{N_c-1} a_i^{(l)} \times w(t - iT_c) \tag{5}$$

where $a_i^{(l)} \in \{1, -1\}, i = 0, 1, \ldots, N_c - 1$, is the user signature sequence of the lth user and T_c is the chip time. A receiver structure for the lth user is the same as that of the single user system except that the correlator is matched to $w^{(l)}(t)$ given in (5) not $w(t)$. The PSD function of the transmitted signal can be calculated similar to the case of the single user system as follows:

$$\begin{aligned}\Phi_{s_{tr}}^{(l)}(f) &= \frac{1}{T_f}|W^{(l)}(f)|^2 \\ &= \frac{1}{T_f}|W(f)|^2 \left| \sum_{i=0}^{N_c-1} a_i^{(l)} \exp[-j2\pi fiT_c] \right|^2\end{aligned} \tag{6}$$

The multi-user system also has no line spectrums and the shape of its PSD function is dependent on both the basic pulse and the user signature sequence. Considering the multi-user interferences of the other users, the BER performance of the system mainly depends on the cross-correlation properties of user signature sequences. When the slot time is synchronised (not necessarily frame time) the BER performance of the system can be the same as that of the single user system by using the rows of a Hadamard matrix with optimum cross-correlation properties. It is noted that when the slot time is not synchronised but the chip time is synchronised, the rows of the Hadamard matrix do not guarantee the same good BER performance. In this case, the user signature sequences with optimal aperiodic cross-correlation properties are necessary.

Simulation results: The channels of personal wireless communications which are important application areas for UWB-IRs are dense multipath channels. To confirm the BER performance, we used the UWB-IR indoor channel model given in [5] by statistical analysis of the UWB-IR channel data obtained from an extensive measurement in a typical modern office environment. Fig. 3 shows the BER performance of the proposed single user system in the UWB-IR indoor channel when the distance between transmitter and receiver is 7 m. The proposed system uses the optimum convolutional code of the same code rate $1/5$ and constraint length 5 as those of the PCTH code of the 32-ary PCTH UWB-IR, respectively. An uncoded system transmits a pulse five times for each time slot index and uses the majority vote decoding for the same data rate, R_d, as that of a coded system. A rake receiver with 50 bins is used and perfect channel estimation is assumed. A guard interval, T_g, was inserted between time frames to obtain meaningful BER curves by decreasing serious intersymbol interference of the channel. Though inserting T_g, the PSD function given in (4) remains, except for the increase of T_f by T_g. The proposed system has a lower error floor below BER $= 10^{-5}$ rather than the PCTH UWB-IR. When d varies, the BER curves are correspondingly shifted and the error floors are preserved.

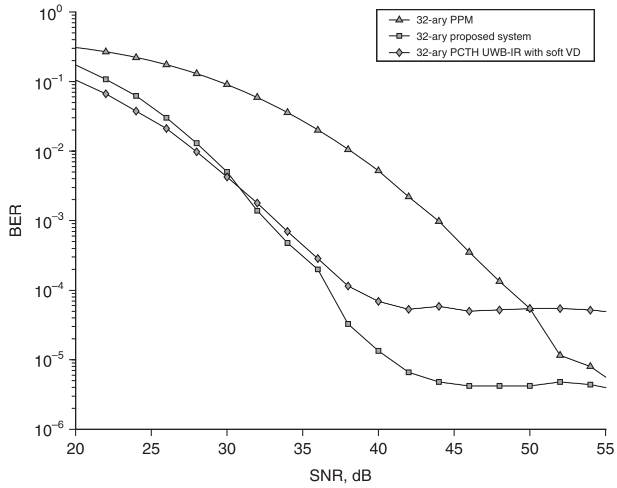

Fig. 3 *BER in UWB-IR indoor channel*
($N = 32$, $T_f = 32$ ns, $T_s = 1$ ns, $R_d = 7.58$ Mbit/s, $T_g = 100$ ns, $d = 7$ m)

Conclusion: We believe that the proposed system is a possible solution for the coexistence problem of the UWB-IR because the system works in lower SNR and has the line spectrum free PSD function. We note that the PSD functions given in (4) and (6) are also applicable to any other linear code, the output equations of which are linearly independent.

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