A NEW TRANSCEIVER DESIGN FOR MULTI-BAND UWB SYSTEM

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ABSTRACT

Pulse position modulation (PPM) and rake receive are usually used in single-band ultra-wideband (SB-UWB) systems. For multi-band ultra-wideband (MB-UWB) systems,we propose a new transceiver design based on spectral keying (SK) modulation. We develop a blockwise multipath channel model represented by a finite impulse response (FIR) filter bank for MB-UWB systems. At the transmitter end, an SK-modulated information symbol is sent through a sequence of P sub-symbol pulses that have distinct frequencies. At the receiver end, a minimum mean square error (MMSE) receiver bank followed by a simple mapping-based decision block is used to detect the received signals. Our scheme can increase the data rate by reducing the guard time between sub-symbol pulses aggressively and improve the system performance by blocking the coefficients of the channel impulse response (CIR) into vectors, while keeping the frequency diversity gain resulting from MB-UWB. Computer simulations illustrates the transceiver design and its effectiveness.

1. INTRODUCTION

Recently Ultra-wideband (UWB) technology has gained a growing attention from the industry, the standardization body and the academic community as a leading technology for the short-range wireless personal area network (WPAN). It will “unwire” people and provide us with seamless wireless integration of multiple digital devices for transmission of multimedia and other high-bandwidth data within the immediate area up to 10 meters.

Spectral keying (SK) [1] is a novel modulation scheme for UWB technology. In our design, a SK-modulated information symbol is represented by a sequence of P sub-symbol pulses which have different frequencies. The multipath channel is assumed to be known throughout the paper. Let T_b, T_s and T be the bin interval, the sub-symbol duration and the information symbol duration respectively. We then assume that $T = PT_s$, $T_s = MT_b$ and $P > 1$, $M > 1$ are integers. We know that it is a multirate discrete system since $T_s \neq T_b$. This motivates us to block the coefficients of the channel impulse response (CIR) and the received signals sampled at $\frac{1}{T_b}$Hz into vectors of length M.

It is shown in the next section that the blockwise multipath channel model has a better receiving property in the sense of less likely having frequency nulls. At the receiver end, a bank of P minimum mean square error (MMSE) receivers are designed and each of the P receivers follows a demodulator which is similar to the M-ary frequency shift keying (MFSK) envelope detector. For simplicity, we only consider to detect the sub-symbols of each information symbol first with a soft decision. Then a hard decision is made by mapping the sub-symbols and their associated soft decisions to the information symbol matrix. This will be made clear in the section of receiver design.

2. CHANNEL MODEL

In practice, the wireless channel cannot be time-invariant due to the time-varying nature of the channel medium and the movement of the transmitter or the receiver. For the indoor short-range UWB channel with very high data rate, however, we can assume it is a linear time-invariant (LTI) channel. The channel is described by

$$h(t) = \sum_{l=0}^{L_h-1} h_l \delta(t - lT_b - \tau_l), \tag{1}$$

where h_l's are the CIR coefficients and τ_l's are the relative time delays within a bin interval for each path,i.e., $0 \leq \tau_l < T_b$. The maximum delay spread is $(L_h - 1)T_b + \tau_{L_h-1}$. Assume that $L_h = (n+1)M$ and $n > 0$ is an integer. Let $l = \nu M + \mu, \nu = 0, 1, \ldots, n, \mu = 0, 1, \ldots, M-1$. (1) is rewritten as

$$h(t) = \sum_{\nu=0}^{n} \sum_{\mu=0}^{M-1} h_{\nu M+\mu} \delta(t - \nu MT_b - \mu T_b - \tau_{\nu M+\mu}). \tag{2}$$

Define a blocking operator $B_M(\bullet)$ via

$$\bar{h}_\nu = B_M(h_l) = \begin{bmatrix} h_{\nu M} \\ \vdots \\ h_{\nu M+i} \\ \vdots \\ h_{\nu M+M-1} \end{bmatrix}, \quad 0 \leq i \leq M-1, i \in \mathcal{Z}.$$

Let $H(z)$, $H_i(z)$ and $\bar{H}(z)$ be the Z-transform of $\{h_l\}_{l=0}^{L_h-1}$, $\{h_{\nu M+i}\}_{\nu=0}^{n}$ and $\{\bar{h}_\nu\}_{\nu=0}^{n}$ respectively. It holds that

$$H(z) = \sum_{i=0}^{M-1} z^{-i} H_i(z^M). \tag{3}$$

Then in the frequency domain, we can find that

$$\begin{array}{rcl} \|\bar{H}(e^{j2\pi f})\|^2 & = & \sum_{i=0}^{M-1} |H_i(e^{j2\pi f})|^2 \\ |H(e^{j2\pi f})| & = & |\sum_{i=0}^{M-1} e^{-j2\pi i f} H_i(e^{j2\pi M f})| \end{array}. \tag{4}$$

Remark 1 *$\bar{H}(z)$ has a better receiving property in the sense that it is less likely to have a frequency null. It can be readily shown that if we assume $\bar{H}(z)$ has a zero at $z = z_0$, then $H_i(z) = (z - z_0)\tilde{H}_i(z)$. From (3), $H(z) = (z^M - z_0)\sum_{i=0}^{M-1} z^{-i}\tilde{H}_i(z^M)$. For a large integer M, it is unlikely that $H(z)$ has M zeros uniformly distributed on a circle with a radius of $\sqrt[M]{z_0}$. It implies that there will be no performance degradation caused by the frequency nulls in $\bar{H}(z)$.*

3. SK MODULATION AND DEMODULATION

3.1. SK Modulation

SK modulation is a multi-band approach to UWB technology. We use a sequence of P sub-symbol pulses with distinct frequencies to represent an information symbol. It is illustrated in the following example and Figure 1. Let $P = 3$, the whole UWB spectrum is split into 3 disjoint sub-bands. Denote $S(f_x, f_y, f_z)$ as an information symbol sent through 3 sub-symbol pulses with frequency f_x, f_y and f_z sequentially, and assume $1 \leq x, y, z \leq 3$, $x \neq y \neq z$ and $x, y, z \in \mathcal{Z}$. There could be $3! = 6$ information symbols. We just randomly choose $N = 2^{\lceil \log_2^{P!} \rceil}$, i.e., $N = 4$ in this example to make a 2-bits mapping. For instance, $00 \rightarrow S(f_1, f_2, f_3)$; $01 \rightarrow S(f_1, f_3, f_2)$; $10 \rightarrow S(f_2, f_3, f_1)$; $11 \rightarrow S(f_2, f_1, f_3)$.

Let $p(t)$ be a rectangular pulse of height 1 with its support $[0, T_s]$. Then a sub-symbol pulse is obtained by modulating $p(t)$ with a sinusoidal signal $cos(2\pi f_{sub} t)$ of frequency f_{sub}. Denote E_s to be sub-symbol energy. Then a sub-symbol pulse can be described as $p(t)\sqrt{\frac{2E_s}{T_s}}cos(2\pi f_{sub} t)$. And the transmitted signal is

$$s(t) = \sum_{k=0}^{\infty} p(t - kT_s)\sqrt{\frac{2E_s}{T_s}}cos(2\pi f_{s_k} t), \tag{5}$$

where s_k represents the k-th sub-symbol pulse and $f_{s_k} \in \{f_1, f_2, \ldots, f_P\}$. Notice that starting from s_0, every P sub-symbol pulses represents an information symbol. That indicates those frequencies corresponding to the P sub-symbol impulses must be different. At the other end, the received signal $r(t)$ is

$$\begin{array}{rl} r(t) & = s(t) \bigotimes h(t) + v(t) \\ & = \sum_{k=0}^{\infty}\sum_{\nu=0}^{n}\sum_{\mu=0}^{M-1} h_{\nu M+\mu} p(\tilde{t} - kMT_b) \\ & \qquad \times \sqrt{\frac{2E_s}{T_s}} cos(2\pi f_{s_k}\tilde{t}) + v(t) \end{array}, \tag{6}$$

where $\tilde{t} = t - \nu M T_b - \mu T_b - \tau_{\nu M+\mu}$, $\bigotimes$ is the convolution operator and $v(t)$ is the zero-mean additive white Gaussian noise (AWGN) with variance σ_v^2.

3.2. SK Demodulation

Based on the binary frequency shift keying (BFSK) demodulation in [2] and the SK demodulator diagram in Figure 3, we down-sample the output of the i-th SK demodulator by M and then we have

$$\begin{array}{l} y_{kM+\mu+1}^{(i)} \\ = \sum_{\nu=0}^{n} h_{\nu M+\mu} e^{j\phi_{\nu M+\mu}^{(i)}} \xi^{(i)}(s_{k-\nu}) + v_{kM+\mu+1}^{(i)}, \end{array} \tag{7}$$

where the superscript (i) denotes the i-th demodulator. In the expression above, $\phi_{\nu M+\mu}^{(i)} = 2\pi f_i(\nu M T_b + \mu T_b + \tau_{\nu M+\mu})$ and $f_i \in \{f_1, f_2, \ldots, f_P\}$ is the frequency in the i-th demodulator; $v_{kM+\mu+1}^{(i)}$ is the sampled AWGN; $\xi^{(i)}[\bullet]$ is a binary mapping function defined as

$$\xi^{(i)}(s_k) = \left\{ \begin{array}{ll} 1 & \text{if the } k\text{-th sub-symbol pulse freq. is } f_i \\ 0 & \text{if the } k\text{-th sub-symbol pulse freq. isn't } f_i. \end{array} \right. \tag{8}$$

Define

$$\bar{y}_k^{(i)} = \begin{bmatrix} y_{kM+1}^{(i)} \\ y_{kM+2}^{(i)} \\ \vdots \\ y_{kM+M}^{(i)} \end{bmatrix}; \qquad \bar{v}_k^{(i)} = \begin{bmatrix} v_{kM+1}^{(i)} \\ v_{kM+2}^{(i)} \\ \vdots \\ v_{kM+M}^{(i)} \end{bmatrix};$$
$$\bar{h}_\nu^{(i)} = \begin{bmatrix} h_{\nu M} e^{j\phi_{\nu M}^{(i)}} \\ h_{\nu M+1} e^{j\phi_{\nu M+1}^{(i)}} \\ \vdots \\ h_{\nu M+M-1} e^{j\phi_{\nu M+M-1}^{(i)}} \end{bmatrix}; \qquad . \tag{9}$$

From (7) and (9), we have

$$\bar{y}_k^{(i)} = \sum_{\nu=0}^{n} \bar{h}_\nu^{(i)} \xi^{(i)}(s_{k-\nu}) + \bar{v}_k^{(i)} \tag{10}$$

Let $\bar{Y}^{(i)}(z)$, $\bar{H}^{(i)}(z)$, $\Xi^{(i)}(z)$ and $\bar{V}^{(i)}(z)$ be the Z-transform of $\bar{y}_k^{(i)}$, $\bar{h}_\nu^{(i)}$, $\xi^{(i)}(s_k)$ and $\bar{v}_k^{(i)}$ respectively, i.e.,

$$\begin{array}{ll} \bar{Y}^{(i)}(z) = \sum_{k=0}^{\infty} \bar{y}_k^{(i)} z^{-k}; & \bar{H}^{(i)}(z) = \sum_{\nu=0}^{n} \bar{h}_\nu^{(i)} z^{-nu}; \\ \Xi^{(i)}(z) = \sum_{k=0}^{\infty} \xi^{(i)}(s_k) z^{-k}; & \bar{V}^{(i)}(z) = \sum_{k=0}^{\infty} \bar{v}_k^{(i)} z^{-k}. \end{array} \tag{11}$$

Then we have the following equation in frequency domain

$$\bar{Y}^{(i)}(z) = \bar{H}^{(i)}(z)\Xi^{(i)}(z) + \bar{V}^{(i)}(z). \qquad (12)$$

4. RECEIVER DESIGN

4.1. Linear Equalizer

Following the i-th SK demodulator, a MMSE linear equalizer (LE) is designed. Hence, there are a filter bank of P MMSE-LE at the receiver end. For simplicity, we neglect the superscript (i) henceforth. From Figure 2, we have

$$\hat{\xi}(s_k) = \sum_{m=0}^{L_g-1} g_m y(kM - m), \qquad (13)$$

where $\{g_m\}_{m=0}^{L_g-1}$ are the filter coefficients. Assume that $L_g = QM, Q > 1$ is an integer. Let

$$\bar{g}_q = [g_{qM}, g_{qM-1}, \ldots, g_{qM-(M-1)}]^T,$$

where $(\bullet)^T$ is the transpose operator. Then (13) can be rewritten as

$$\hat{\xi}(s_k) = \sum_{q=0}^{Q-1} \bar{g}_q^T \bar{y}_{k-q}. \qquad (14)$$

Define

$$\begin{aligned}
\mathbf{H} &= \begin{bmatrix} \bar{h}_0 & \bar{h}_1 & \ldots & \bar{h}_n & & & \\ & \bar{h}_0 & \bar{h}_1 & \ldots & \bar{h}_n & & \\ & & \ddots & \ddots & \ddots & \ddots & \\ & & & \bar{h}_0 & \bar{h}_1 & \ldots & \bar{h}_n \end{bmatrix}_{Q\times(n+Q)} \\
\mathbf{Y} &= [\bar{y}_k^T, \bar{y}_{k-1}^T, \ldots, \bar{y}_{k-(Q-1)}^T]^T \\
\mathbf{\Xi} &= [\xi(s_k), \xi(s_{k-1}), \ldots, \xi(s_{k-n-Q+1})]^T \\
\mathbf{G} &= [\bar{g}_0^T, \bar{g}_1^T, \ldots, \bar{g}_{Q-1}^T]^T \\
\mathbf{V} &= [\bar{v}_k, \bar{v}_{k-1}, \ldots, \bar{v}_{k-(Q-1)}]^T
\end{aligned}$$

From (10), we know that

$$\mathbf{Y} = \mathbf{H}\mathbf{\Xi} + \mathbf{V}. \qquad (15)$$

Then (13) can be rewritten as

$$\hat{\xi}(s_k) = \mathbf{G}^T\mathbf{Y} = \mathbf{G}^T\mathbf{H}\mathbf{\Xi} + \mathbf{G}^T\mathbf{V}. \qquad (16)$$

Define the error signal as

$$\epsilon_k = \xi(s_k) - \hat{\xi}(s_k) = e_\delta^T\mathbf{\Xi} - \mathbf{G}^T\mathbf{H}\mathbf{\Xi} - \mathbf{G}^T\mathbf{V}, \qquad (17)$$

where $e_\delta = [0, \ldots, 0, 1, 0, \ldots, 0]^T$ is the delay vector whose $(\delta+1)$-th element is one and all the others are zero, and δ denotes the delay steps. Assume that both the signal and the noise are white and independent. The the standard MMSE solution [3] of $\mathbf{G}$ is

$$\mathbf{G} = (\mathbf{H}\mathbf{H}^H + \sigma_v^2 I)^{-1}(\mathbf{H}e_\delta), \qquad (18)$$

where $(\bullet)^H$ is the Hermitian operator.

4.2. Decision Rule

With the MMSE solution $\mathbf{G}$ obtained at the i-th receiver, we can make a soft decision about the estimate $\hat{\xi}(s_k)$ at time instant kT_s.It could be either one or zero. Since all the P receivers are parallel, we have all the P estimates $\{\hat{\xi}^{(1)}(s_k), \hat{\xi}^{(2)}(s_k), \ldots, \hat{\xi}^{(P)}(s_k)\}$ so that we can determine an information symbol transmitted at $[kT_s, (k+P)T_s]$. Recall the example in section 3.1, if the estimates we get from the first, the second and the third receiver are "zero, one and zero" at time instant $(3K+1)T_s$, $K \in \mathcal{Z}^+$. Similarly we get estimates "one, zero and zero" at time instant $(3K+2)T_s$ and "zero, zero and one" at $3(K+1)T_s$. Then we determine that 2-bits "11" are sent by use of the following information symbol matrix

$$11 \rightarrow \begin{bmatrix} 0 & 1 & 0 \\ 1 & 0 & 0 \\ 0 & 0 & 1 \end{bmatrix}.$$

5. SIMULATION STUDY

In this section, we simulate a real realization of the UWB channel model measured by Intel's researchers [4]. A modified double-Poisson lognormal channel model is adopted here with the parameters: cluster decay factor Γ=16ns; ray decay factor γ=1.6ns; cluster arrival rate $\Lambda = \frac{1}{60ns}$; ray arrival rate $\lambda = \frac{1}{0.5ns}$; standard deviation of lognormal fading term σ_{Intel}=4.8dB. A line-of-sight (LOS) UWB channel realization simulated from Intel's model is shown in Figure 4. We can find the clustering phenomenon in the figure that fits well with Intel's measurement.

Since UWB is an overlay radio system, the instantaneous power spectrum density limit of UWB is set by FCC to be -41.25dBm/MHz. Hence we only consider a low SNR scenario, i.e., SNR<15dB here. To compare with the conventional equalizer design of UWB, we simulate both the MMSE-LE based on the proposed blockwise channel model and a linear equalizer based on a non-blockwise channel model. In the simulation, we set L_h=180, n=17, M=10,L_g=48 and δ=23 for the MMSE-LE and L_h=185, L_g=215 and δ=190 for the conventional LE. We can see in the Figure 5 that the conventional LE does not work with a low SNR, but the MMSE-LE has a SNR gain of 5dB when the sub-symbol error rate is about -20dB. It can be explained that the blockwise channel model has a better receiving property as we state in the Remark 1.

6. CONCLUSION

In this paper, we present a new transceiver design of MB-UWB based on SK modulation. The proposed blockwise channel model is shown to have a better receiving property in the sense of less frequency nulls. Parallel MMSE-

LE receivers reduce the computation time. The system performance can be further improved if we consider to detect an information symbol directly from the estimates of P receivers, not to make a decision first at each receiver and then to map the information symbol matrix.

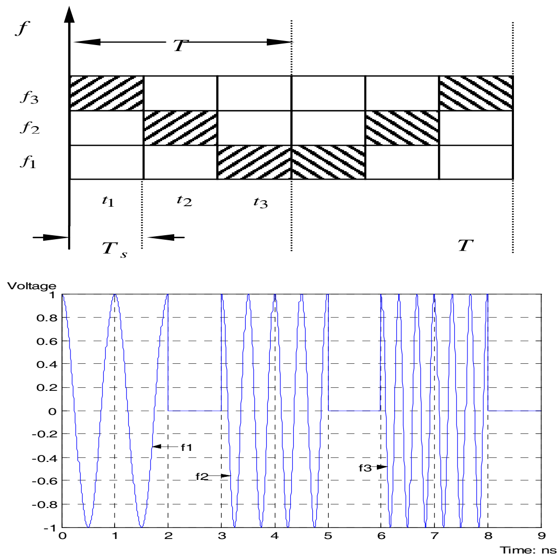

Fig. 1. Spectral Keying Modulation

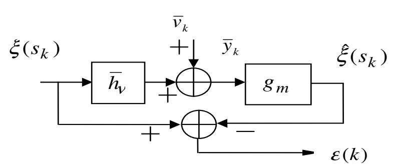

Fig. 2. MMSE Linear Equalizer

7. REFERENCES

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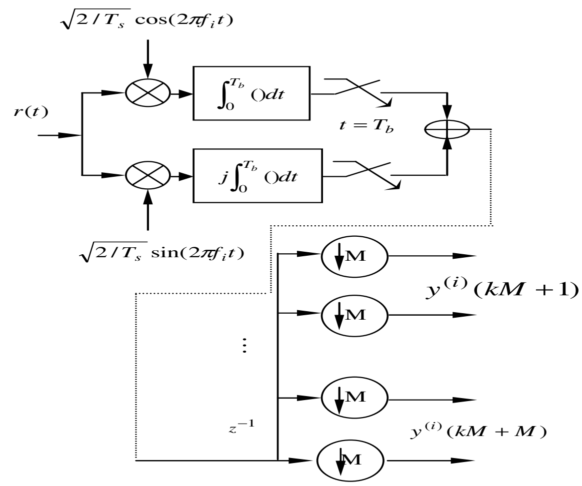

Fig. 3. Spectral Keying Demodulator

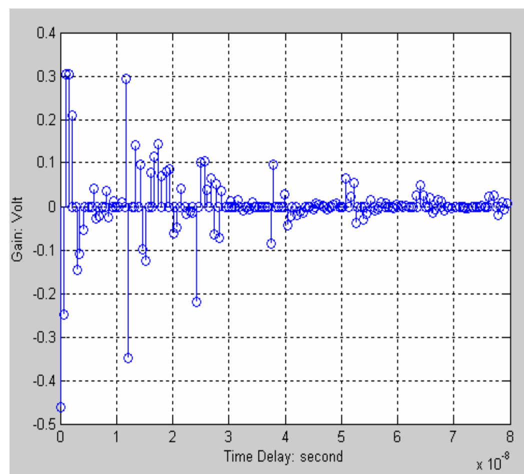

Fig. 4. A LOS Channel Realization of Intel UWB Model

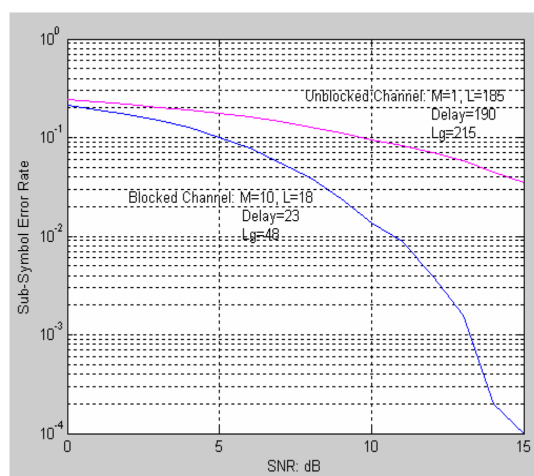

Fig. 5. MMSE-LE Receiver Performance