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Low-complexity ultrawideband transceiver with compatibility to multiband-OFDM

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***Abstract*— We present a low-complexity transceiver for ultrawideband communications with moderate (1-15Mbit/s) data rate. This transceiver is based on time-frequency-interleaved frequency-shift-keying (TF-FSK), and shows a high degree of compatibility with multiband-OFDM, the currently envisioned standard for high-data-rate ($>$100Mbit/s) UWB communications. We show that for dual-mode devices, the major part of the multiband-OFDM transceiver can be reused for the TF-FSK transceiver. We also study the performance of this transceiver in (standardized) UWB channels, and find that (depending on the data rate), coverage ranges of up to 30m (with LOS connection) are possible.**

I. INTRODUCTION

Ultrawideband (UWB) systems are defined as systems that have a relative bandwidth of more than 20%, or an absolute bandwidth of more than 500MHz. Such systems show many desirable properties, like immunity to multipath propagation, easier penetration of walls and floors, precise geolocation capabilities, and inherent security [1], [2], [3], [4], [5], [6]. Most importantly, they exhibit a very low power spectral density, so that the radiation they emit does not seriously disturb existing services. This allows them to operate simultaneously, and in the same frequency bands, as current systems.

Due to these beneficial properties, there has been intense interest in such systems, especially since the frequency regulator in the USA, the FCC (Federal Communications Commission), allowed the unlicensed operation of UWB emitters subject to restrictions in the spectral emission properties and applications [7]. Following that ruling, the IEEE has established a standardization body, IEEE 802.15.3a, for defining a physical-layer standard based on UWB transmission. The goal of this standard is achieving a data rate of 110Mbit/s at 10m distance, and higher data rates at shorter distances. In March 2003, more than 20 proposals were submitted to the IEEE. While the standard is not finalized yet, the only baseline proposal currently (March. 2004) still under consideration is the multiband-OFDM proposal described in Ref. [8].[1]

This proposal (which will also be briefly described in Sec. II), uses a combination of time-frequency (TF) interleaving and OFDM: the time-frequency plane is divided into units of 312.5 ns duration and 528 MHz width. A time-frequency code prescribes which of those TF-"chunks" are active for one specific user; this gives the multiple-access capabilities; within each TF-chunk, one OFDM symbol is transmitted, see Fig. 1. The use of OFDM allows good energy collection also in channels with high delay spreads. This system shows excellent performance, even at high data rates, but it does require advanced, and thus costly, technology. Analogue to digital conversion, as well as an FFT, must operate at a sampling rate of more than 528 MHz. This is necessary for the high data rates envisioned in the original call-for-proposals of the IEEE, which is needed, e.g., for transmission of (uncompressed) digital video.

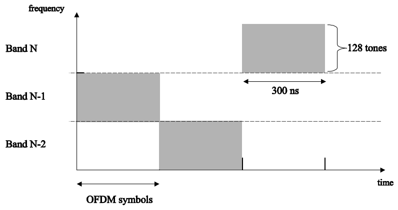

Fig. 1. Principle of time-frequency interleaved OFDM.

There are, however, also many applications where a much lower data rate is required - between 1 and 20 Mbit/s. Such applications encompass the current Bluetooth applications, as well as consumer electronic applications described, e.g., in [9], [10]. It is thus desirable to design a physical-layer mode that fulfills the following requirements:

- it can operate at those low data rates,
- it must have much lower complexity than the "normal" multiband-OFDM,
- it has to retain compatibility to multiband-OFDM, and allow an implementation of dual-mode transmitters (receivers) that create less cost and complexity than the sum of a high-mode and a low-mode transmitter (receiver).

This paper describes a low-complexity modem that fulfills all of the above requirements. It is based on time-frequency interleaved FSK. Using incoherent demodulation, it allows for extremely simple transceivers; by exploiting some ba-

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[1] This standard draft was selected as baseline approach by the IEEE. However, it has not yet passed the confirmation vote, and is thus not a definite choice.

sic similarities between OFDM and FSK, it also retains a high degree of compatibility with the full-complexity mode. However, we stress here that the modem is *not* theoretically optimum:

- due to its underlying structure of dividing each 500 MHz band into two subbands and transmitting different powers on them, it cannot achieve the same performance as OFDM even when a full-complexity receiver is used.
- when an incoherent receiver is used, it shows a performance loss on the order of 10dB in most UWB channels.
- when designing a stand-alone incoherent modem, a single-band pulse-based system would be preferable.

The main quality of our proposed scheme is the compatibility with the multiband-OFDM scheme, and the fact that dual-mode transmitters and receivers are exceedingly simple to build !!! There are applications where the transmitter is a full-complexity (multiband-OFDM) device, because it should be used for high data rate applications as well as for the low-complexity mode; however, the receiver is a low-complexity device. Such a situation might occur, e.g., when the transmitter is on a computer, while the receiver is on an MP3 player. In that case, the transmitter will usually be a dual-mode device. Similarly, there are applications where the transmitter is low-complexity only, but the receiver is required to be able to also handle full complexity.

The remainder of the paper is organized the following way: Section II reviews the multiband-OFDM physical layer, as it forms the basis for the compatibility analysis. Section III introduces the time-frequency interleaved FSK: after introducing the signal structure, we suggest specific implementations both for the transmitter and the receiver that make maximal use of the similarity to multiband-OFDM. Finally, simulations show the performance of this system in typical UWB channels. A summary and conclusions wraps up the paper.

II. MULTIBAND OFDM

As a starting point, we give a brief description of the multiband OFDM standard, as compatibility with this approach is a major goal for our novel transceiver. The source data (in packets of 1 kByte) are convolutionally encoded, interleaved, and mapped onto QPSK symbols. Each group of 100 such symbols is OFDM-modulated, i.e., it is serial-to-parallel converted; pilot tones and null tones are added, and the resulting 128 tones are subjected to an IFFT (inverse fast Fourier transform). A cyclic prefix is prepended, or a null padding is appended; in either case, the duration of prefix or postfix is 70ns, which is approximately equal to the delay spread of the channels this system is developed for. Each OFDM symbol, which lasts a total of 312.5 ns, is then upconverted, with a carrier frequency that changes from symbol to symbol. This carrier frequency is determined by a time-frequency code that is specific for each user. The employment of different TF codes gives the multiple-access capabilities. At the beginning of each packet, a training sequence is transmitted for channel estimation. A more detailed block diagram, which includes the coding, bit interleaving, and symbol mapping, is shown in Fig. 2. More details can be found in [8].

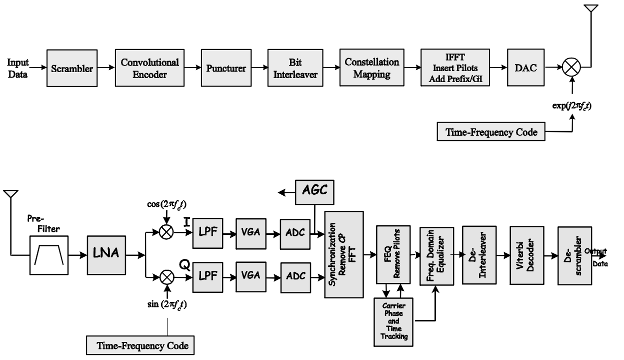

Fig. 2. Blockdiagram of an OFDM transceiver according to the IEEE proposal.

The system was designed and evaluated for the UWB channel models standardized by IEEE 802.15.3a [11]. These models are essentially Saleh-Valenzuela models, with the multipath components undergoing lognormal (instead of the usual Rayleigh) fading. Four different radio environments are defined, with different delay spread for each of them. CM1 corresponds to a LOS (line-of-sight) situation with a distance between transmitter and receiver less than 4 m. CM2 is a non-LOS situation, with a $0 - 4$ m distance between transmitter and receiver. CM3 is a NLOS situation between 4 and 10 m, and CM4 corresponds to heavy multipath, with 25 ns delay spread. The pathloss is modeled as free-space pathloss, i.e., to go with d^{-2} for all channel models. While this is not realistic for NLOS situations, it is the model that has been standardized by IEEE and has to be used for performance evaluations in that context.

III. TRANSCEIVER STRUCTURE FOR TF-FSK

A. Signal structure

The high-data-rate system has to combat considerable delay spread, namely up to 25 ns rms delay spread, and 200 ns maximum excess delay. With delay spreads extending over many symbol durations (5 ns or less), collection of energy is most easily achieved by OFDM. However, for the case of lower data rates, a simpler, pulse-based scheme can be used. Each symbol is represented by transmitting energy in either the lower or the upper half of a 528 MHz band - in other words, we use very wideband FSK on top of the time-frequency interleaving. In order to keep the signal structure as similar to OFDM as possible, multiple contiguous symbols are transmitted within one 528 MHz band, as outlined in Fig. 3. The transmit signal can thus be written as

$$\begin{aligned} s(t) \quad &= \quad E\sum_{i=0}^{\infty}\sum_{k=0}^{K-1} p[t-(k+iK)T_b]\exp(j2\pi b_{k+iK} f_{\text{offset}} t) \\ &\quad\quad \exp(j2\pi f_i t) rect[t,(i-1)KT_b, iKT_b] \qquad (1) \end{aligned}$$

where T_b is the bit duration, i.e., the inverse of the (coded) data rate; the b_n are the (coded) data bits 1 or 0; f_{offset} is 264 MHz; $rect(x,a,b)$ is a function that is unity when $a < x < b$

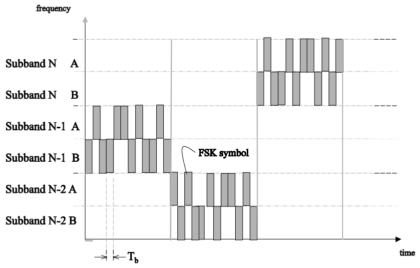

Fig. 3. Symbol structure of time-frequency-band FSK.

and zero otherwise; and the f_i the instantaneous carrier frequencies that are determined by a (periodic) time-frequency-interleaving code (in the example of Fig.3), $f_0 = f_c + \Delta f$, $f_1 = f_c$, $f_2 = f_c + 2\Delta f$, and $f_i = f_{i \bmod 3}$ for $i > 2$). K describes the number of bits that are transmitted on one carrier frequency (i.e., within a 312ns interval), and $p(t)$ is the basis pulse normalized to unit energy. Note that the bandwidth of the pulse is fixed in our system to 264MHz, but the duration of the pulse T_p can take on any value $1/264$ MHz$< T_p \leq T_b$.[2] Thus, different time-bandwidth products BT product can be used. The received signal $r(t)$ is

$$r(t) = h(t) * s(t) + n(t) \tag{2}$$

where $h(t)$ is the channel impulse response, and $n(t)$ is additive white Gaussian noise.

The use of FSK as modulation format allows either coherent or noncoherent reception. With incoherent reception, we first perform a downconversion of the signal to baseband, by multiplying with a signal $\exp(j2\pi f_i t) rect[t, (i-1)KT_b, iKT_b]$ (note that in the absense of a phase reference, this might lead to additional losses). Next, the receiver obtains signals $\widetilde{r}_l(t)$ and $\widetilde{r}_u(t)$ by filtering with bandpass filters whose passbands cover the range $0 - 264$ MHz, and $264 - 528$MHz, respectively. The decision variables x_u and x_l are then

$$x_l = \int_0^{T_r} |\widetilde{r}_l(t)|^2 dt \tag{3}$$

$$x_u = \int_0^{T_r} |\widetilde{r}_u(t)|^2 dt \tag{4}$$

The integration time T_r can be chosen according to the channel conditions, as discussed in Sec. IV; it is, however, upper-limited by T_b. For an uncoded system, it is decided at a $+1$ was sent if $x_l < x_u$. For coded systems, the(continuous) values of x_l and x_u are used as input of the Viterbi decoder.

Incoherent detection can entail a noticeable penalty, depending on the frequency selectivity of the channel. For a rough approximation, we can divide the available frequency band into several entities with a bandwidth B_c, the coherence bandwidth of the channel. We then can use the frequency analog of the common "block-fading" approximation, namely that all frequency components within such entity fade completely coherently, while different entities exhibit completely independent fading. If there are L such blocks within a 264 MHz band, the achievable SNR with coherent detection is a factor of L larger than for the case of incoherent detection.

In the first report and order of the FCC [7], it is required that the instantaneous bandwidth (defined as the 10 dB bandwidth) is larger than 528 MHz in order for a system to qualify as UWB. This would not be fulfilled if we just switch between the two frequency subbands $0 - 264$ and $264 - 528$. This problem can be avoided if we transmit both bands simultaneously; but with a $10dB$ power difference. The data then just determines which of the two subbands is attenuated. This leads to a certain amount of self-interferences, namely a 10 dB SIR. As the forward error correction code is designed to work at SNRs of 4 dB, the performance loss from such a scheme is negligible, as will be confirmed by simulations in later sections.

The distribution of the output of the incoherent detector, assuming that a $+1$ was sent, is then (see also Ref. [12])

$$p_0(x) = (\frac{x}{E_{na}})^{(M-1)/2} \exp(-\frac{x + E_{na}}{N_0}) I_{M-1}(\frac{\sqrt{x E_{na}}}{N_0/2}) \tag{5}$$

$$p_1(x) = (\frac{x}{E_a})^{(M-1)/2} \exp(-\frac{x + E_a}{N_0}) I_{M-1}(\frac{\sqrt{x E_a}}{N_0/2}) \tag{6}$$

where $2M = 2BT + 1$ is the bandwidth expansion factor, E_a and E_{na} are the symbol energies in the active and inactive band, respectively, and N_0 is the noise spectral density.

Intersymbol interference can become a limiting factor for the admissible data rate. This is especially true for incoherent detection. We suggest two ways to mitigate the effect of ISI:

- the duration of the $transmit$ symbol T_p is taken shorter than the bit duration T_b. This reduces the amount of intersymbol interference, while retaining the total transmit power (note that the FCC rulings allow a peak-to-average ratio of up to 20dB). Note that the integration time T_r at the receiver might be chosen larger than the pulse duration, but is upper-limited by the bit duration, $T_p \leq T_r \leq T_b$.
- the intersymbol interference can also be reduced by changing the frequency of the local oscillator after every symbol, instead of every 312ns, so that the transmit signal reads

$$\begin{aligned} s(t) &= E \sum_{i=0}^{\infty} p[t - iT_b] \exp(j2\pi b_i f_{\text{offset}} t) \\ &\quad \exp(j2\pi f_i t) rect[t, (i-1)T_b, iT_b] \end{aligned} \tag{7}$$

. In that case, only the energy that extends over N_p symbol durations (where N_p is the periodicity of the TF code) acts as interfering ISI. This value is very small at all data rates and channel models considered. The drawbacks of this scheme are (i) compatibility with the multiband-OFDM scheme is lost, and (ii) the local oscillator must be able to change its frequency much more frequently than in the scheme above (after each duration T_b). Furthermore, no signal can be received during the frequency-changing time, which is around $5 - 10$ ns for typical low-cost devices. This can lead to an appreciable loss of collectable energy. In the following, we will denote this scheme as "fast hopping", and the scheme of Eq. (1) as "slow hopping".

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[2] Bandwidth efficiency is not a major concern for the UWB applications this scheme is intended for.

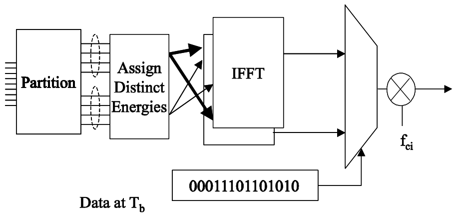

Fig. 4. Block diagram of a FSK transmitter using OFDM components.

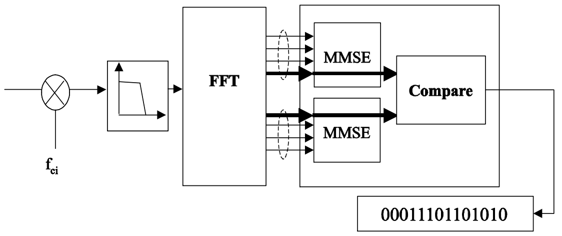

Fig. 5. FSK receiver structure using OFDM components.

B. Dual-mode transceiver structure

For a low-complexity-only device, the transmitter structure is straightforward, consisting of a local oscillator (whose frequency can be adapted according to the time-frequency code), plus a "standard" wideband FSK transmitter. One of the main goal of the reduced-complexity mode is compatibility with the current multiband-OFDM standard. Thus, a dual-mode transmitter (which is able to put out multiband-OFDM signals as well as FSK signals), should essentially be a multiband-OFDM transmitter with as little additional components as possible. In order to achieve that, we use the IFFT available in an OFDM transmitter to generate the FSK signal.[3] Figure 4 shows our new implementation. In it, the wideband signal generator is realized from components that are all available in an OFDM transmitter, plus a switch. The starting point is a set of data symbols that can be thought of as belonging to different frequencies. Those frequencies are partitioned in two groups, one representing the $0 - 264$ MHz range, one the $264 - 528$ MHz range. We are then performing an IFFT on each of the groups.

As the spacing between the tones is the inverse of the symbol duration, and the FSK symbols are shorter than the OFDM symbols, the required size of the IFFT is smaller than in the true OFDM case. Thus, even though we need two IFFTs, the hardware effort for those is smaller than in the OFDM case. In most cases, an IFFT is realized as a multi-stage Butterfly structure. In that case, we can just group existing elements of the OFDM butterfly in a different way, and obtain two IFFTs of smaller size, without a requirement for any additional components (apart from connectors and switches). At the output, we finally just need a switch controlled by the user data to decide which of the two IFFT outputs should be transmitted.

For the receiver, we again have to distinguish two different situations. The first is where it is built into a "low-complexity-mode-only" device. In that case, a standard incoherent FSK receiver can be used - which essentially just requires two bandpass filters and energy detectors, as outlined in Sec. III.A. For the case that the receiver should be able to process both multiband-OFDM signals and FSK signals, a different approach is preferable. As FFT components are available in the receiver, these can be exploited for equalization and coherent detection, which improves the performance of the system.

The principle is the same as for "normal" OFDM: the received signal can be represented by a number of tones, which are obtained by performing an FFT on a block of received signal samples (the sampling speed is determined by the requirements of the multiband-OFDM signal). The frequency-domain signals can then be equalized, using the channel knowledge obtained during the training sequence. As the FSK signals do not contain any cyclic prefix, the equalization is either imperfect, or more complicated (using, .e.g., the algorithm of [13]) than the "regular" one-tap OFDM equalizer. The tones belonging to each of the subbands are then combined with maximum-ratio or minimum mean-square-error (MMSE) combining. This allows to exploit the frequency diversity in the signal.

Note that the spacing of the tones is again determined by the duration of the FSK symbols, so that a $T_b f_s -$point FFT is used, where f_s is the sampling frequency of the receiver. The spacing between the processed tones is thus larger than in the regular multiband-OFDM. In strongly frequency-selective channels, like CM3 and CM4 of the IEEE 802.15.3a channels, the different frequency components of each OFDM tone thus do not add up completely coherently. This leads to an additional loss of performance.

IV. PERFORMANCE

In the following, we analyze (by simulations) the performance of the FSK scheme with incoherent detection. The simulations use the following assumptions: (i) one packet consists of 1024 bytes, as prescribed in the 802.15.3a standard. (ii) data are coded with a rate $1/2$ convolutional coder; decoding is done with a Viterbi decoder with traceback length 96, (iii) the given data rates are the rates of the source data (before the encoder), (iv) the packet error rates are shown as a function of distance, where it is assumed that the received power is inversely proportional to the square of the distance (free-space pathloss), see Sec. II, (v) for the multipath channels, the average over the 90% best realizations of the multipath channels is used for the computation of the PER, following the procedure often used in the IEEE 802.15.3a downselection process[4] (vi) for the slow hopping scheme, we use an integration time $T_r = T_p$. The pulse durations T_p are optimized for the different channels and data rates The OFDM curves that are shown for comparison have a source data rate of 110Mbit/s, and use optimum weighting the input of the Viterbi decoder.

Figure 6 shows the performance in AWGN. Note that (in the case of very small switching times of the local oscillator),

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[3] Note that this technique would not be efficient for a stand-alone FSK transmitter. It is advantageous only specifically in the context of a dual-mode (FSK+OFDM) transmitter.

[4] Note that the intersection of the PER curves with the 8% line gives the distance at which the 8% are achieved *on average*; it is not the 90% outage distance.

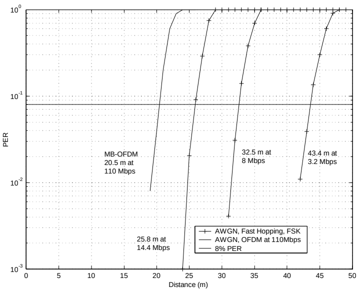

Fig. 6. Performance of the FSK scheme in AWGN as a function of the distance.

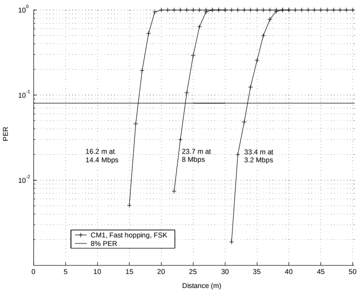

Fig. 7. Performance of fast hopping FSK in CM1.

there is no difference between fast hopping and slow hopping in that channel. We find that 8% packet error rates can be achieved over distances ranging from 25m (for the 14Mbit/s mode) to 43m (for 3Mbit/s). This compares very favorably with the target range of 10m (defined for the $same$ channel model). For comparison, we also show the performance of the 110Mbit/s mode (from [8]). We note that it can achieve distances that are slightly lower than the 14 Mbit/s mode, but not significantly so.

Figure 7 shows the performance of the fast hopping mode in the CM1 channel model. We find a noticable performance degradation at all data rates, which is due to the fading, as well as the fact that incoherent reception has a performance penalty in frequency-selective channels, as discussed in Sec. III.A. However, we see that 16m coverage is possible even with 14Mbit/s and 30m can be achieved with 3.2Mbit/s. In CM3, 12.7 m coverage can be achieved for 14Mbit/s. Figure 8 shows the impact of slow hopping on the performance. We see that for the 8Mbit/s mode, the coverage distance decreases only about 1.4m (from 23.7 to 22.3m). For the 3.2Mbit/s mode, the performance loss is even smaller. Similar results are also obtained in the other channel models (not shown here for space reasons). We stress, however, that this is only true if T_p and T_r are chosen correctly. A further important question is the impact of the transmission of energy in the "inactive" frequency band (so that the FCC requirement of 500MHz instantaneous bandwidth is fulfilled). Fig. 8 shows the performance when this out-of-band emission is taken into account for the slow-hopping simulations. Again, the performance decrease is very small (coverage distance decreases from 22.3 to 20.5 m); this holds also true for other channel models. This confirms our conjecture from Sec. III.A.

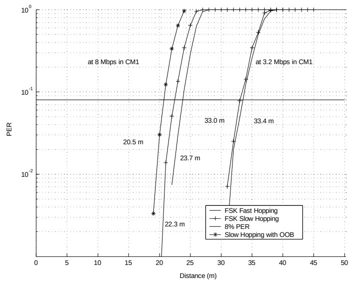

Fig. 8. PER as a function of distance in CM1 for 3.2 and 8 Mbit/s. Both slow and fast hopping are shown, as is "slow-hopping dual mode", which takes into account the -10dB out-of-band emissions required to obtain 500 MHz instantaneous bandwidth.

V. SUMMARY AND CONCLUSION

We have suggested a new signalling scheme for UWB transmission, based on a combination of wideband FSK with time-frequency interleaving. It allows extremely simple transceiver structures, and is thus well-suited for low-cost transceivers. Using the same basic signaling structure as multiband-OFDM, it is compatible with this current IEEE 802.15.3a standards proposal. TF-FSK signals can be received by a multiband-OFDM receiver, and multiband-OFDM transmitters can easily generate FSK signals, without requiring expensive additional components. There is a noticeable performance loss in frequency-selective channels when noncoherent detection is used. Still, for lower data rates, coverage distances of some 30m can be achieved for 3Mbit/s data rates, based on evaluations with the IEEE channel model.

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