PHYSICAL LAYER OF A BASE-BAND OFDM MODEM:
ALGORITHMS AND PERFORMANCE

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The base-band section of a modem employing Orthogonal Frequency Division Multiplexing (OFDM) is described in this paper. It utilizes the necessary algorithms to combat distortion due to the channel conditions and imperfect synchronization. A model describing these distortions is used to derive the algorithms. The system is realized in a prototype platform for the HIPERLAN/2 standard, but modifications for compliance to other broadband digital broadcasting and wireless networking OFDM systems are proposed. The performance of the prototype base-band modem is described.

Keywords: OFDM; Base-band receiver design; wireless networking.

1. Introduction

OFDM is nowadays chosen for many broadband digital broadcasting applications and wireless networks. The European Wireless Local Area Network (WLAN) standard, HIPERLAN/2\(^1,2\) employs OFDM, and so does the IEEE equivalent, 802.11a. Also OFDM has been proposed in the past for DAB and DVB-T, and more recently as one of the alternatives of the IEEE 802.16 standard for broadband outdoor communications.

The simplicity of the OFDM transceivers, built around Inverse Fast Fourier Transform (IFFT) and FFT pairs leads to low-cost implementations for frequency-selective channels.\(^3\) On the other hand, the major disadvantages of OFDM are its vulnerability to timing and frequency synchronization errors for the base-band
implementation, and its requirements for high linearity and low phase noise for the Radio Frequency (RF) implementation. The analysis in Refs. 4 and 5 for OFDM reception in frequency-selective channels led to a model for the received data symbols. Expressions for the Inter-Carrier Interference (ICI) and the Inter-Symbol Interference (ISI) terms were derived and used to determine the synchronizer performance that minimizes them.

This paper continues the analysis after assuming that the synchronization, although not ideal, is good enough to render the ICI and ISI terms negligibly. The remaining significant impairment terms of the model of the received data symbols are manipulated to point out the need for three algorithms to counter the impairments: one that spans the subcarriers only, another that spans the symbols only and a third that spans both the subcarriers and the symbols. The third algorithm is novel, as it is the manipulation of the synchronization metric. The other two algorithms are well-known and the effect of putting them together in a system is studied here. These algorithms are implemented in a hardware prototype that although is based on HIPERLAN/2, where appropriate, the necessary changes for compliance to other broadband digital broadcasting standards are described.

The paper is organized as follows: In Sec. 2, first the OFDM reception in frequency-selective channels is considered, based on Refs. 4 and 5, to identify the impairments introduced by nonideal synchronization. Then, the algorithms that cancel the introduced impairments are presented, together with the necessary synchronizer, and their implementation is discussed. In Sec. 3, a brief description of the prototype platform of the OFDM modem is given, followed by the detailed presentation of its performance. Finally, in Sec. 4, the conclusions are drawn.

2. Base-Band OFDM Modem

The base-band OFDM modem is shown in Fig. 1. It comprises a transmitter (Tx) and a receiver (Rx). The IFFT and FFT pair in the Tx and the Rx respectively partitions the modem in frequency- and time-domain sections.

The implementation of the Tx according to the HIPERLAN/2 standard or any other standard is straightforward. The frequency-domain Tx section first comprises bit manipulations: scrambling, Forward Error Correction (FEC), puncturing and interleaving. IEEE 802.16a also employs Reed–Solomon coding. Then constellation encoding and subcarrier (data and pilot) mapping for the IFFT follow. The time-domain Tx section after the IFFT comprises the cyclic prefix and the preamble addition.

On the other hand, the implementation of the Rx is not straightforward and is the goal of this section. The sources of the various impairments associated with an OFDM reception in a frequency-selective channel are identified and modeled based on 4.5. Having done so, the necessary algorithms to minimize the effect of the impairments are derived. The overall algorithmic system design is based on the
Fig. 1. Implementation of the base-band OFDM modem. The frequency-domain transmitter (Tx) section is followed by the time-domain Tx section. The OFDM signal is corrupted by the frequency selective channel and Additive White Gaussian Noise (AWGN), before being received by the time-domain section of the receiver (Rx). The modem concludes with the frequency-domain section of the Rx.
choice of the minimum computational complexity algorithms that yield an acceptable overall performance under the worst-case implementation scenarios (indoor channel models, analog impairments), as they are specified in the HIPERLAN/2 standard. These algorithms are implemented at the receiver either before the FFT (in the time-domain) or after (in the frequency-domain).

2.1. OFDM reception in frequency-selective channel
Assume that OFDM transmission with $K$ subcarriers is used. The $k$th subcarrier of the $l$th OFDM symbol is modulated by the data symbol $d_{l,k}$.

Then the transmitted signal is given by:

$$s(t) = \sum_{l=0}^{K/2-1} \sum_{k=-K/2}^{K/2} d_{l,k} \exp \left( j \frac{2\pi k}{T} (t - T_{CP} - lT_S) u(t - lT_S) \right),$$

where $T_S = T_{CP} + T$ is the OFDM symbol duration (which comprises of the cyclic prefix (CP) of duration $T_{CP}$ and the useful portion of duration $T$) and

$$u(t) = \begin{cases} 1 & 0 \leq t < T_S, \\ 0 & \text{elsewhere}. \end{cases}$$

This transmitter is implemented using an IFFT of length $N \geq K$.

Let's for the moment ignore the noise added by the channel. Then the transmitted signal suffers distortion from the frequency-selective channel, whose maximum dispersion is assumed to be less than $T_{CP}$ to avoid ISI. The impulse response of the channel $h_i$ is assumed not to change significantly during a single OFDM burst. This is justified by the very low Doppler frequency $f_D$ implied in Ref. 2, which leads to $f_D \cdot T < 2 \cdot 10^{-4}$ for the 5–6 GHz band. Hence from Ref. 5, the inter-carrier interference introduced by the temporal variance is negligible. For the same reason, the per-symbol deviation of the actual channel response from the estimated one is ignored in the analysis. Then the received signal, sampled multiples of $t_s = T/N$ is

$$r_n = \sum_i h_i(nT_s)s(nt_s - \tau_i),$$

where $\tau_i$ is the delay associated with tap $i$. The received data symbols are obtained using the FFT on $r_n$. They are

$$d'_{l,k} = d_{l,k}H_k,$$  

where, if ISI is avoided, $H_k$ describes the effect of the channel on the data symbols and depends only on the subcarrier index, i.e., is constant on a per-symbol basis.

Equations (3) and (4) hold under ideal synchronization of the transmitter and the receiver and describe the degradation of the signal in the time-domain and of
the data symbols in the frequency-domain, respectively. In practice, the synchronizer operates in the time-domain, and its effectiveness depends on the channel conditions, i.e., the Signal-to-Noise Ratio (SNR) and \((h_i, \tau_i)\). Hence, after the FFT, the receiver is only approximately synchronized to the transmitter: Some residual frequency offset \(\Delta f'\) exists, as the frequency offset between the Tx and the Rx RF carriers is estimated with finite precision. Also, the sampling time \(t'_s\) at the Rx is different from the Tx. Finally, there can be a symbol timing \(n'\) that is offset from \(n\). Taking into account these impairments, the sampled received signal at the time-domain becomes \(^4^{6}\)

\[
r_n = \exp(j 2\pi \cdot \Delta f' \cdot nt_s) \sum_i h_i(nt'_s) s((n - n')t'_s - \tau_i).
\]  

At the frequency-domain, the received data symbols are \(^4^{5}\)

\[
d'_{l,k} = ISI + ICI + d_{l,k} H_k [\exp(j 2\pi \cdot \Delta f' \cdot lT_S) \sin c(\pi \cdot \Delta f' \cdot T)]
\cdot \left[ \exp\left( j 2\pi \frac{k(t'_s - t_s)}{t_s T} \right) \sin c\left( \frac{\pi k(t'_s - t_s)}{t_s T} T \right) \right]
\cdot \left[ \exp\left( j 2\pi \frac{k}{N} \frac{N - n'}{N} \right) \right],
\]  

where \(ISI\) and \(ICI\) are inter-symbol and the inter-carrier interference terms and the three bracketed expressions are associated with the carrier frequency offset, the sampling time offset and the symbol timing offset, respectively. Note from Eq. (7) that the effect of the sampling time offset is that of a subcarrier varying frequency offset

\[
\Delta f_{eq}(k) = \frac{k(t'_s - t_s)}{t_s T}.
\]  

The inter-symbol interference term is due to the symbol timing offset and can be zero due to the use of the cyclic prefix. Hence some symbol time offset can be tolerated, as long as timing estimation is such as the FFT interval does not extend over a symbol boundary or the roll-off region of the previous symbol. \(^3^{5}\) Inter-carrier interference is caused by the loss of subcarrier orthogonality due to the frequency and sampling time offsets. In Refs. 3, 5 and 7, it is deduced that for normal ranges of \(\Delta f'\) and \(t'_s\), ICI can be ignored. This can be verified by simulation; the SNR loss due to various values of \(\Delta f'\) is shown in Fig. 2.

Hence, provided successful synchronization in the time domain, the frequency domain received data symbols can be obtained by ignoring the \(ISI\) and \(ICI\) terms in Eq. (7) and re-arranging the remaining terms

\[
d'_{l,k} = d_{l,k} \cdot \left[ H_k \exp\left( j 2\pi \frac{k}{N} \frac{N - n'}{N} \sin c\left( \frac{\pi k(t'_s - t_s)}{t_s T} \right) \right) \right]
\cdot \left[ \exp\left( j 2\pi \frac{k(t'_s - t_s)}{t_s T} IT_S \right) \right] \cdot \left[ \exp\left( j 2\pi \frac{k(t'_s - t_s)}{t_s T} IT_S \right) \right],
\]  

where the three bracketed terms represent the distortion on the transmitted data that the Rx has to cancel out after the FFT.
Fig. 2. SNR degradation as a function of frequency offset estimation error. This estimation error causes ICI. The actual frequency offset is 40 ppm, thus the loss shown is the worst-case.

The first bracketed term in Eq. (9)

$$D_1(k) \equiv H_k \exp\left( j2\pi k \frac{n'}{N} \right) \sin c\left( \pi \left( \frac{k(t'_s - t_s)}{t_s T} \right) \frac{N - n'}{N} \right) \sin c(\pi \cdot \Delta f' \cdot T)$$  (10)

depends only on the subcarrier $k$, i.e., is time-invariant. This term can be cancelled out using an estimate of $D_1(k)$ which depends on the subcarrier $k$, but not on the symbol number $l$. The second bracketed term

$$D_2(l) \equiv \exp(j2\pi \cdot \Delta f' \cdot lT_s)$$  (11)

depends only on the symbol $l$. It can be cancelled out by estimating $D_2(l)$ on a per-symbol basis, independent of the subcarriers. Finally, the third bracketed term

$$D_3(l, k) \equiv \exp\left( j2\pi \frac{k(t'_s - t_s)}{t_s T} lT_s \right)$$  (12)

depends on both the subcarrier $k$ and the symbol $l$. It can be cancelled out by estimating $D_3(l, k)$ both on a per-subcarrier and on a per-symbol basis.

### 2.2. Receiver algorithms

In the time-domain Rx section, the synchronization (symbol timing and frequency offset) is first estimated using the preamble, and then the estimated frequency offset is corrected. Finally, the cyclic prefix is removed, preparing the input to the FFT.

At the Rx frequency-domain section, de-mapping the FFT output and the various algorithms to cancel the impairment terms $D_1(k)$, $D_2(l)$ and $D_3(k, l)$ are...
followed by the constellation decoder and the inverse bit manipulations carried out
in the Tx, i.e., de-interleaving, de-puncturing, Viterbi decoding and unscrambling.
Note that a pair of 3-bit soft constellation decoder and 3-bit soft Viterbi decoder
is used. This greatly enhances the performance of the modem quantified using the
Packet Error Rate (PER), as shown in Fig. 3. Consequently, the de-interleaver in
between the two blocks has to handle 3-bit groups instead of single bits. Increasing
the soft bits beyond 3 still enhances the performance somewhat, but the advan-
tage diminishes as also shown in Fig. 3. The Viterbi complexity increases by 7.5%
(approximately 30 k gates more) if the soft bits are increased from 3 to 4. In the present
implementation, this increase is considered too much for the resulting performance
enhancement.

As the Tx implementation is detailed in the standard,\(^1\) the Rx bit manipula-
tions are not considered any further in this paper. In the following subsections,
the synchronizer from the time-domain Rx and the various blocks that cancel the
impairment terms \(D_1(k), D_2(l)\) and \(D_3(l,k)\) are discussed in detail.

The adverse effects of a channel, \(D_1(k)\), are well-known and not specific to
OFDM modems. The terms \(D_2(l)\) and \(D_3(l,k)\) are OFDM-specific, and their effect
grows with time. In Fig. 4, the effect of not compensating for these two impairment
terms on the SNR is compared as a function of the number of symbols in the
burst. Obviously for very short burst, their effect can be tolerated, but not for any
realistic data bursts. Also, the effect of \(D_3(l,k)\) is more than doubly as severe as
that of \(D_2(l)\).
2.3. Synchronization

Synchronization is carried out in the time-domain of the Rx. Its goal is the symbol timing and frequency offset estimation. If it is successful, only a small residual frequency offset $\Delta f'$ will remain and the symbol timing will be either on the actual OFDM symbol beginning, or in the later parts of the cyclic prefix where the channel influence is negligible. This is true as there is not a single ideal symbol timing index, but rather a window, whose length is determined by the length of the cyclic prefix and the roll-off of the channel. These synchronization conditions ensure that ISI and ICI can be ignored, and that the only degradation of the received data at the frequency-domain comprises the impairment terms $D_1(k), D_2(l)$ and $D_3(l,k)$ and any noise introduced by the channel.

Several types of synchronization algorithms have been evaluated using either correlation with the guard interval, the Schmidl and Cox correlation method, modified correlation methods to combat ISI in multipath fading channels, or joined time and frequency domain synchronization. The first approach typically results in delayed frame start position estimates while introducing overall system latency due to the need of using several OFDM symbols before an acceptable symbol timing estimation can be derived. The modified Schmidl and Cox synchronization method was found to marginally improve the timing estimation for the indoor fading channels under consideration, while the joined time-frequency estimator requires FFT and IFFT computations and although giving fine symbol timing estimation, it increases the overall hardware requirements. Since more

Fig. 4. Comparison of the effects of not compensating the terms $D_2(l)$ and $D_3(l,k)$ on the SNR. The degradation becomes larger as the number of symbols in the burst increase. The SNR given the proposed compensation for both terms is included for comparison.
advanced channels are not in consideration for the intended deployment scenarios, the Schmidl and Cox method\cite{10} has been adopted that offers both symbol timing and frequency offset estimations with acceptable accuracy and complexity.

The coherent detection of the transmitted PDUs is achieved by exploiting the repetitive nature of the preamble.\cite{1} The maximization of the normalized (by the received signal power) correlation between the two identical parts of the preamble provides both the symbol timing and the frequency offset estimations. The correlation calculation is based on the iterative formula\cite{10}

\[ r_{n+1} = r_n + x_n^* x_{n+2L} - x_n^* x_{n+L}, \]

where \( x \) are the incoming complex In-phase/Quadrature (I/Q) samples and \( r \) is the autocorrelation. Similarly the power \( p \) of the incoming complex I/Q samples is given by the iterative formula

\[ p_{n+1} = p_n + |x_{n+2L}|^2 - |x_{n+L}|^2. \]

Hence the Schmidl–Cox metric is given by

\[ M_n = \frac{|r_n|^2}{p_n^2}. \]

For HIPERLAN/2, \( L = 32 \) and this metric ideally (no channel or noise) comprises two large peaks. Each indicates that the two 32-samples long sections of the \( A \) and \( B \) parts of the preamble are detected. The position in time of the two large metric peaks provides the symbol timing estimation, and the value of \( r \) at the peaks provides the frequency offset estimation \( \Delta f_{\text{est}} \)

\[ \Delta f_{\text{est}} = \frac{\text{arg}(r_{n_{\text{peak}}})}{2\pi L t_s}. \]

The estimated frequency offset is corrected in the time-domain by multiplying the incoming complex I/Q samples \( r_n^{\Delta f} \) by the necessary complex exponential

\[ r_n = r_n^{\Delta f} \cdot \exp(-j2\pi \cdot \Delta f_{\text{est}} t_n). \]

The corrected I/Q samples \( r_n \) have the cyclic prefix removed and are processed by the FFT.

A peak detector implemented in the synchronizer detects the two large peaks. It also ignores the smaller peak that follows each of the two large metric peaks. These smaller peaks are due to the final 16 samples in the \( A \) and \( B \) parts of the preamble. The small peak between the two larger ones can be troublesome under nonideal channel conditions. As shown in Fig. 5(a), this middle peak can be comparable to the two ideally larger ones. Hence it is important to ignore it.

Due to the different preamble structures of both IEEE 802.11a and 802.16a, their metrics ideally have a plateau at unity. As shown in Figs. 5(b) and 5(c), due to adverse channel conditions, this plateau can be very jagged, making the
Fig. 5. Ideal and actual metrics for (a) HIPERLAN/2, (b) IEEE 802.11a and (c) IEEE 802.16a. The first forms peaks, while the latter two a plateau, which at 5 dB SNR is very jagged. After differentiation and smoothing with a comb filter, the two peaks appear again. Any intermediate peak can be avoided as in HIPERLAN/2.

peak detection described above useless. For these standards to work with the same synchronizer, a differentiator and a smoothing comb filter follow the calculation of Eq. (15). After dropping the sign, there are again two large peaks to be detected, using the same peak detector described above.
2.4. Sampling offset compensation

The objective of the sampling offset compensation is to cancel the effect of the
impairment term \( D_3(l, k) \), which is caused by the offset between the Tx and Rx
sampling clocks. In order to do so, the sampling offset \( t'_s - t_s \) in Eq. (12) is estimated
as follows.

Since the base-band clock and the RF carrier are generated from the same
reference,\(^1\) their offsets are proportional. So a carrier frequency offset \( \Delta f \) will gen-
erate a base-band sampling time offset, such that the resulting sampling time \( t'_s \) is
given by

\[
(\text{RF} + \Delta f)t'_s = \text{RF} \cdot t_s .
\]  

(18)

This yields

\[
t'_s = \frac{\text{RF}}{\text{RF} + \Delta f} t_s .
\]  

(19)

Hence using the frequency offset estimation \( \Delta f_{\text{est}} \) from the synchronizer (Eq. (16))
and the given RF, the sampling offset \( t'_s - t_s \) at the Rx can be estimated. Then,
based on Eq. (8), the subcarrier varying equivalent frequency offset can be esti-
mated as

\[
\Delta f_{\text{eq}}(k) = -k \frac{\Delta f_{\text{est}}}{T(\text{RF} + \Delta f_{\text{est}})} .
\]  

(20)

Hence \( D_3(l, k) \) can be corrected at the frequency-domain by multiplying each
of the \( K \) subcarriers of the FFT output by the complex exponential \( \exp(-j2\pi \cdot \Delta f_{\text{eq}}(k) \cdot lT_S) \) for every symbol \( l \).
The error in the estimation of the equivalent frequency offset due to the frequency offset estimation error $\Delta f'$ is

$$\left| \Delta f_{eq}(k) + k \frac{\Delta f}{T(RF + \Delta f)} \right| \approx \frac{k}{T \cdot RF} \Delta f',$$  

(21)

which is a very small fraction of $\Delta f'$ and increases linearly by the subcarrier index $k$.

2.5. Channel estimation and equalization

The objective of channel estimation and equalization is to estimate and cancel out the impairment term $D_1(k)$. Since $D_1(k)$ does not depend on the symbol number $l$, the second part of the preamble is used to estimate the $K$ values of $D_1(k)$ only once. The known FFT of the noiseless preamble symbol is divided by the FFT of the received preamble symbol, providing the inverse of the estimate of $D_1(k)$. This is the simple $y/x$ estimator. Equalization of every data symbol is achieved by multiplying the active subcarriers by this inverse estimate.

Note that for HIPERLAN/2, there are actually two identical preamble symbols to be used for channel estimation. The two received $C$ symbols are averaged to reduce the effect of noise to the $D_1(k)$ estimate. These are corrected regarding $D_3(l,k)$, but still suffer from the $D_2(l)$ impairment term and the additive $N(l,k)$ noise term. Hence, the received $C$ data for the two symbols are

$$c'_{l,k} = c_k D_1(k) D_2(l) + N(l,k), \quad l = 0, 1,$$

(22)

where $c_k = \pm 1$. The averaged received $C$ data symbol is

$$c_k' = c_k D_1(k) \frac{D_2(0) + D_2(1)}{2} + \frac{N(0,k) + N(1,k)}{2}.$$  

(23)

Hence the estimated value $D_{1,est}(k) \equiv c'_k/c_k$ varies from the actual as

$$D_{1,est}(k) = D_1(k) \frac{D_2(0) + D_2(1)}{2} + \frac{N(0,k) + N(1,k)}{2c_k}.$$  

(24)

As $\Delta f' \ll 1/T_S$, Eq. (24) yields

$$D_{1,est}(k) = D_1(k) \frac{1 + e^{j2\pi \cdot \Delta f' \cdot T_S}}{2} + \frac{N(0,k) + N(1,k)}{2c_k} \approx D_1(k) + \frac{N(0,k) + N(1,k)}{2c_k}.$$  

(25)

Hence the influence of $D_2(l)$ in the estimate of $D_1(k)$ is insignificant and the influence of the noise is halved by averaging.

The above analysis holds also for IEEE 802.11a, where there are also two $L$ preamble symbols for channel estimation. On the contrary, IEEE 802.16a has a single preamble symbol for channel estimation, loosing the advantage of noise averaging. Moreover, this symbol contains information every second subcarrier, hence channel estimates are obtained for every second subcarrier and interpolation between these channel estimates is needed for the intermediate subcarriers.
The $y/x$ channel estimator, although simple, is far from optimum, especially under deep fades that are typical of outdoor channels. In this case, equalization amplifies noise rather than the faded subcarriers. Two solutions have been studied: Either the use of more optimum and complex channel estimators like the Wiener channel estimator or the Least Means Square (LMS) decision feedback tracking equalizers (DFE), or adaptive modulation of the subcarriers that suffer from fading can be applied. Simulation yields superior results (1.5 dB gain for the Wiener approach and 0.5–1.0 dB gain for LMS DFE), but both alternatives are not implemented in the current prototype system due to hardware constraints. The adaptive modulation is not implemented either as it is not supported by the existing WLAN standards. The decision in favor to the $y/x$ channel estimator is justified by the performance of the resulting modem for indoor channels.

### 2.6. Fine phase tracking and correction

The objective of the fine phase tracking and correction blocks is to estimate and cancel out the effect of the impairment term $D_2(l)$. This term exists due to the inaccurate estimation of the carrier frequency offset that allows for the small residual frequency offset $\Delta f'$. The term depends only on the symbol number $l$, hence the pilot signals in each symbol are used for its compensation. After sampling offset compensation, channel equalization and normalization by the transmitted pilots, the magnitude of the received pilots is ideally unity. Their phase $\phi(l,k)$ is related to the phase of $D_2(l)$ as

$$\phi(l,k) = \arg(D_2(l)) + N(l,k), \quad k = -21, -7, 7, 21,$$

where $N(l,k)$ is the influence of the noise on the phase of the pilots. Hence averaging over the four pilots provides an estimate $\tilde{\phi}(l)$ of the phase of $D_2(l)$ with reduced noise influence

$$\tilde{\phi}(l) = \arg(D_2(l)) + \frac{1}{4} \sum_{k=\pm21,\pm7} N(l,k).$$

Fine phase correction is achieved by multiplying with $\exp(-j\tilde{\phi}(l))$.

Note that under the used simple $y/x$ channel estimator, some of the pilots could suffer from a spectral null. The effect on the overall performance of such a null on a pilot subcarrier is greater than that on a data subcarrier. Simulation shows that performance can be maintained to acceptable levels when a severely attenuated pilot is discarded from averaging or a more complex Maximum Likelihood (ML) phase estimator is used. Since such schemes are more complex and do not provide significant performance improvement for indoor deployment, they are not implemented in the current prototype.

### 3. Implementation and Performance of the Base-Band Modem

A prototype implementation of the OFDM modem for HIPERLAN/2 is realized on a Field Programmable Gate Array (FPGA) prototyping platform.
The platform hosts two ARM microprocessors and two XILINX Virtex E 2000, 0.18 µm, six metal layers, with 500 K usable gates and 832 Kb of additional RAM (BlockRAM) and built-in clock management circuitry (eight Digital Locked Loops, DLL) each. In the first FPGA, the frequency domain Rx is mapped. The total utilization of the first FPGA is 85%. The second FPGA includes the Tx, the time domain Rx, the interface to the Medium Access Control (MAC) layer and a slave interface to a bus of type Advanced Microprocessor Bus Architecture (AMBA). The total utilization of the second FPGA is 89%. Note that in the second FPGA, where the time domain Tx and Rx are hosted, the digital modulation/demodulation and interpolation/decimation filters that interface to the data converters are also included.

The prototype platform is used to determine the performance of the modem for the targeted HIPERLAN/2 system. Seven physical modes employing different modulation schemes (BPSK, QPSK, 16-QAM and 64-QAM) and coding rates are described by ETSI BRAN. Also, five channel models are described, namely channels A, B, C, D and E. These models are designed to simulate different transmission scenarios. Models A, B, C and D are for indoor and model E is for outdoor transmission. All these models are used for the performance evaluation.

In the following subsection, the performance of the synchronizer is first evaluated. Then the performance of the complete base-band OFDM modem under various conditions is obtained. Note that mostly the PER and not the Bit Error Rate (BER) is used as a performance meter, as the system goal specified in the standard is 10% PER for 54-byte long packets. Another measure of performance is the Error Vector Magnitude (EVM) that describes the discrepancy of the received constellation points from their ideal positions and is given by

\[
EVM = \frac{\sum l \sum_{k=-K/2}^{K/2} r_{l,k}^2 - d_{l,k}^2}{\sum l \sum_{k=-K/2}^{K/2} d_{l,k}^2},
\]

where \(r_{l,k}\) are the received data after the correction for the \(D_1(k), D_2(l)\) and \(D_3(l, k)\) impairment terms. Since the constellation power is normalized to unity, the denominator of Eq. (28) is unity.

### 3.1. Synchronizer performance

The goals of the synchronizer are to provide symbol timing and frequency offset estimations. The frequency offset estimator is zero-mean and its variance depends on the SNR and on the correlation length. Short correlation lengths decrease the estimation accuracy but increase the tracking range of the estimator, i.e., the maximum detectable frequency offset. The performance also depends on the channel, as shown in Table 1, for low SNR (5dB).

The performance of the estimator is very good at high SNR, where most of the physical modes are supposed to operate. This is demonstrated in Fig. 6(a).
Table 1. Synchronizer performance for 5 dB SNR and 40 ppm frequency offset.

<table>
<thead>
<tr>
<th>Channel</th>
<th>RMS delay spread</th>
<th>RMS frequency offset (kHz)</th>
<th>Probabilities (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>50.0</td>
<td>9.26</td>
<td>0</td>
</tr>
<tr>
<td>B</td>
<td>99.0</td>
<td>10.2</td>
<td>0.1</td>
</tr>
<tr>
<td>C</td>
<td>149</td>
<td>9.44</td>
<td>0.8</td>
</tr>
<tr>
<td>D</td>
<td>139</td>
<td>9.60</td>
<td>0.2</td>
</tr>
<tr>
<td>E</td>
<td>248</td>
<td>14.8</td>
<td>0.8</td>
</tr>
</tbody>
</table>

For channels A, B, C, D and E, there exists a frequency offset estimation Root-Mean-Square (RMS) error floor. It is due to the adverse channels and not to the fixed-point implementation of the synchronizer; simulating the synchronizer for 16 and 32-bit arithmetic yields similar results for any channel. For Additive White Gaussian Noise (AWGN), there is no such error floor.

The effect of the frequency offset estimation error floor of the frequency-selective channels can be evaluated using Fig. 2. For the 4.2 kHz RMS error of channel E, the SNR loss due to only the imperfect frequency offset estimation is 7.3 dB. This degradation is due to the channel. Under ideal channel, the frequency offset estimation error is much less; for any offset, it remains below 300 Hz. Simulations with an ideal synchronizer yield an SNR loss that ranges from 0.2 dB at 40 ppm offset, where the relative frequency offset estimation error is 0.145%, up to 6.3 dB at 0.4 ppm offset, where the relative frequency offset estimation error is 11.6%. Of these SNR losses, only 0.1 dB (40 ppm) up to 0.7 dB (0.4 ppm) are due to the 16-bit fixed-point implementation compared to 32-bit arithmetic.

It is possible after the initial frequency offset estimation and correction to perform a second correction using the second part of the preamble, where the correlation length is double. By doing so, the variance of the estimator is greatly reduced for low SNR (42% improvement at 5 dB), but is almost unaffected at high SNR (2.5% improvement at 23 dB). So at high SNR, where the demanding physical modes are expected to operate, the second correction scheme has little to offer. For example, at 64-QAM, the SNR gain for 10% PER is only 0.074 dB. At low SNR, where the estimation is improved considerably, the physical modes are more insensitive to frequency offsets, and the finer estimation does not improve the PER performance noticeably. At BPSK code rate $1/2$, the SNR gain is just 0.12 dB. Hence no fine frequency offset estimation and correction is employed in the modem. The frequency offset is corrected adequately since no significant improvement is achieved by using the exact frequency offset figure instead of the estimation. For 64-QAM with 40 ppm frequency offset and under channel A, the SNR for which 10% PER is achieved does not change. For BPSK code rate $1/2$, ideal estimation yields an improvement of the SNR by 0.3 dB.

The symbol timing estimator standard deviation is also depicted in Fig. 6(b). Its performance is qualified by the probabilities of miss and false alarm. The probability of miss is defined as the ratio of runs for which the timing estimator yields a delay
greater than the RMS delay spread of the channel, to the total runs. Likewise, for the probability of false alarm, the timing estimations that are less than a window below the RMS delay spread of the channel are used. This window is the cyclic prefix duration minus the channel RMS delay spread. Both probabilities are also given.
Table 2. Receiver performance for various timing offsets around the estimated timing sample.

<table>
<thead>
<tr>
<th>Timing offset (samples)</th>
<th>BER</th>
<th>PER (%)</th>
<th>EVM (dB)</th>
<th>Notes</th>
</tr>
</thead>
<tbody>
<tr>
<td>−16</td>
<td>3.53e−2</td>
<td>32.4</td>
<td>−21.1</td>
<td>ISI from previous symbol</td>
</tr>
<tr>
<td>−15</td>
<td>1.69e−3</td>
<td>3.20</td>
<td>−24.9</td>
<td>ISI from spreading of previous symbol</td>
</tr>
<tr>
<td>−14</td>
<td>0</td>
<td>0</td>
<td>−30.7</td>
<td>symbol due to channel</td>
</tr>
<tr>
<td>−12</td>
<td>0</td>
<td>0</td>
<td>−36.5</td>
<td></td>
</tr>
<tr>
<td>−10</td>
<td>0</td>
<td>0</td>
<td>−37.3</td>
<td></td>
</tr>
<tr>
<td>−8</td>
<td>0</td>
<td>0</td>
<td>−37.7</td>
<td></td>
</tr>
<tr>
<td>−6</td>
<td>0</td>
<td>0</td>
<td>−38.3</td>
<td>Optimum timing</td>
</tr>
<tr>
<td>−4</td>
<td>0</td>
<td>0</td>
<td>−38.1</td>
<td></td>
</tr>
<tr>
<td>−2</td>
<td>0</td>
<td>0</td>
<td>−37.9</td>
<td></td>
</tr>
<tr>
<td>−1</td>
<td>0</td>
<td>0</td>
<td>−37.3</td>
<td></td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>0</td>
<td>−36.7</td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>1.49e−1</td>
<td>78.4</td>
<td>−17.7</td>
<td>ISI from next symbol</td>
</tr>
<tr>
<td>2</td>
<td>3.57e−1</td>
<td>98.8</td>
<td>−13.2</td>
<td></td>
</tr>
</tbody>
</table>

in Table 1; they remain low even under very adverse conditions. Note that under adverse channels, post-FFT timing synchronization algorithms can improve performance. Such algorithms are not needed in indoor scenarios, as demonstrated by the low probabilities of miss and false alarm of the proposed pre-FFT synchronizer.

The robustness of the timing estimator is demonstrated by evaluating the effect of some offset around the timing estimation. To do so, a full 64-QAM frame is passed through a noiseless instance of channel $A$. The effect of a timing offset on BER, PER and EVM is listed in Table 2.

The symbol timing offset is normalized to the symbol timing estimation of the synchronizer. If the synchronizer estimate were accurate, then timing synchronization after that point would be disastrous, as ISI from the next symbol occurs. This is exactly the performance in Table 2, indicating the accuracy of the estimation.

Timing synchronization before timing estimation is not a problem, up to the point where the roll-off from the previous symbol due to the channel affects the cyclic prefix (e.g., channel $A$ has some influence up to half the cyclic prefix, or 390 ns). This is the reason why the EVM degrades only when early symbol timing synchronization by eight samples occurs, indicating a symbol timing window of eight samples.

It is obvious from the results that symbol timing synchronization is ensured as long as some samples back-off from the synchronizer estimation is used. For the five ETSI BRAN channels, the optimum value is two samples back-off.

### 3.2. Modem performance under AWGN

In order to estimate the performance of the base-band modem under AWGN, 100 full frames are received per SNR value. The preamble is the 4-symbol broadcast that allows symbol timing, frequency offset and channel estimation. It is followed
by 496 data symbols. For 64-QAM, this corresponds to 25,000 packets, or more than 1.08e7 data bits. The large number of frames guarantees a good estimation of the synchronizer performance, while the large number of packets (or bits) reduces the uncertainty of the PER estimation. Also the use of the whole frame in a single burst puts the maximum strain to the fine phase correction algorithm that operates on a per-symbol basis.

![PER versus SNR for the two most demanding physical modes of HIPERLAN/2. AWGN with or without frequency offset and the five ETSI BRAN channels are considered.](image-url)
Table 3. Required SNR for 10% PER under AWGN and ETSI BRAN channels. For 64-QAM, code rate 3/4, the SNR for ideal channel estimation are also included.

| Physical mode | | | ETSI BRAN channel and AWGN |
|---------------|---------------|---------------|----------|----------|----------|----------|----------|
|               | AWGN and carrier offset | ETSI BRAN channel | A | B | C | D | E |
| BPSK 1/2      | 3.07          | 5.13          | 7.00 | 7.11 | 5.87 | 6.30 |
| BPSK 3/4      | 4.53          | 10.6          | 12.1 | 12.3 | 9.22 | 12.2   |
| QPSK 1/2      | 5.70          | 8.88          | 10.5 | 10.9 | 6.95 | 10.9   |
| QPSK 3/4      | 8.05          | 13.9          | 15.3 | 16.0 | 13.2 | 16.6   |
| 16-QAM 9/16   | 11.5          | 15.4          | 16.4 | 17.4 | 13.9 | 18.3   |
| 16-QAM 3/4    | 14.3          | 20.0          | 21.5 | 22.4 | 17.8 | 25.8   |
| 64-QAM 3/4    | 19.9          | 24.8          | 27.0 | 28.1 | 23.1 | —      |

For ideal y/x channel estimation:

|               | — | — | — | — | — |

The PER as a function of the SNR for the three more demanding physical modes is depicted in Fig. 7. Similar curves are also given for 40 ppm frequency offset. The required SNR for each physical mode is presented in Table 3.

The necessary frequency offset correction of Eq. (17) is implemented by multiplying the I/Q samples with the sine and cosine of the correction term. This complex multiplication cannot be completed per sample if it is done with a clock of 40 MHz. Such a clock was originally used in the frequency offset correction block, so the correction term was updated every second sample. This leaded to an increased required SNR for 10% PER that reached 1.5 dB for the most demanding physical modes, compared to the final implementation with a clock of 80 MHz and the correct update of the term at every sample.

3.3. Modem performance under ETSI BRAN channels

Each ETSI BRAN channel\(^2\) comprises a 18-tap FIR filter with random coefficients: their magnitude follows the Rayleigh statistic, or the Rician for the Line-Of-Sight (LOS) component, around the given average, while their phase is uniformly distributed in \([-\pi, \pi]\). The delays between the filter taps are multiples of 10 ns and differ between the channel models, resulting to the different RMS delay spreads shown in Table 1. Due to the number of variable (random) parameters in the channel models, to accurately estimate the PER performance of the modem, 1000 different instances of the channels are created. The results are also depicted in Fig. 7 and Table 3.

For channel A, the results are in considerable agreement with those from Ericsson.\(^{14}\) This justifies the selection of the suboptimal y/x channel estimator. It is evident from the results that channel D is the easiest, with A close by. Also B and C impose similar impairments and channel E is the worst.

For 64-QAM, under channel E the modem fails as 10% PER is never reached, no matter how much is the SNR increased. There are two reasons for this failure.
Firstly, the delay spread of the channel $E$ instances is increased, causing ISI that can be reduced only if the cyclic prefix is increased. Indeed, making the cyclic prefix equal to the useful symbol duration halves the PER. Secondly, the fades are very deep. The deep fades at moderate SNR values effectively reduces the SNR of the faded subcarriers below the necessary for demodulation. Also, even at high SNR, 10% PER is still not achieved. To represent the whole dynamic range of the inverse of the channel magnitude, the wordlength of the fixed-point channel estimation arithmetic operations has to be drastically increased, leading to ideal $y/x$ channel estimation.

The effect of ideal $y/x$ channel estimation on the PER performance of 64-QAM is also shown in Table 3. A maximum of 1.5dB of SNR gain can be obtained by making the wordlength of the $y/x$ channel estimator and equalizer infinite. This SNR gain does not depend on the channel. Also note that the targeted 10% PER is not reached for channel $E$, even with ideal $y/x$ channel estimation, due to the increased delay spread.

Note that the difficulties of the proposed modem with channel $E$ are not considered a problem, since this channel is a model for outdoor deployments. There is no intention for such a deployment of the current system.

4. Conclusions

In this paper, a base-band OFDM modem is proposed. Based on an analysis of the effects of frequency-selective channels, the necessary receiver algorithms are derived and are implemented on a prototype platform for a HIPERLAN/2 system. The performance of the prototype modem is measured under AWGN and the various ETSI BRAN channels. The overall performance of the system is found to be satisfactory under indoor channel conditions. For outdoor deployment, a more optimum channel estimator is needed.

It is shown that the proposed receiver can be made compliant to other broadband digital broadcasting and wireless networking standards, by pointing out the necessary implementation changes. Following this, the realization of the proposed receiver on a reconfigurable SoC platform that will switch from standard to standard by changing the reconfigurable part of the SoC, is currently under investigation.

References

1. ETSI, Broadband Radio Access Networks (BRAN); HIPERLAN Type 2; Physical (PHY) Layer, TS 101 475 (1999).


