Physical Layer Performance of a Novel Fast Frequency Hopping-OFDM Concept

Tobias Scholand, Thomas Faber, Juho Lee, Joonyoung Cho, Yunok Cho and Peter Jung

***Abstract*— Orthogonal frequency division multiplexing (OFDM), facilitating inexpensive receiver structures and high data throughput at the same time, is emerging as a key radio transmission technology for wireless communications in various applications. When applied to time-varying multi-path environments as they occur in mobile radio scenarios, OFDM suffers from absent frequency diversity potential and, consequently, shows an insufficient system performance when frequency diversity techniques are not employed. Here, the authors illustrate an OFDM based system concept, using fast frequency hopping (FFH) within an OFDM symbol, termed FFH/OFDM concept. The FFH/OFDM concept exploits frequency diversity and, by choosing cyclic FFH patterns, facilitates a low-cost implementation of transmitters and receivers. Also, the deployment of advanced forward error correction (FEC) coding, in particular Turbo-Coding, is taken into account.**

***Index Terms*—Fast Frequency Hopping (FFH), Linear Pre-Coding, Mobile Communications, Orthogonal Frequency Division Multiplexing (OFDM), Turbo-Codes, Universal Mobile Telecommunications System (UMTS)**

I. Introduction

In the recent past, multicarrier (MC) techniques have attracted much interest as a potential air interface technology for evolved 3G mobile radio systems like UMTS (Universal Mobile Telecommunications System) [1] and mobile communications "beyond 3G" [2],[3]. In particular, orthogonal frequency division multiplexing (OFDM) [4], which allows both inexpensive transceivers and high data throughput, respectively, has been studied widely [2],[3],[5],[6]. However, when applied to time-varying mobile multi-path environments, the OFDM performance suffers from the absence of frequency diversity. To improve the diversity potential of OFDM based concepts the introduction of frequency domain spread spectrum techniques [7],[8], sometimes called MC-CDMA, combined with appropriate forward error correction (FEC) coding [1],[6] is advisable.

As a possible contribution to the UMTS downlink, the authors illustrate a physical layer concept, which considers several issues raised in the OFDM Study Item of 3GPP RAN WG1, see Sect. II. The novel system concept uses fast frequency hopping (FFH) within an MC symbol, thus combining FFH with OFDM transmission; it is therefore termed FFH/OFDM concept. Data symbols in the frequency domain are mapped to time domain samples by a unitary transformation setting out from a shuffled version of the well-known inverse discrete Fourier transform (IDFT) [9], cf. Sect. II. This feature can be viewed as the application of a specially selected version of linear pre-coding to regular OFDM [9]. The nature of the FFH/OFDM concept is to spread each data symbol over several, ideally all, frequency subcarriers to exploit frequency diversity [9]. Thus, the FFH/OFDM concept is closely related to spread spectrum techniques, e.g. OFDM-CDMA, however, taking a different, yet simpler approach to spread spectrum.

Another, however, less flexible approach towards exploiting the frequency diversity potential in mobile radio systems is taken by the so-called "DFT spreading" concept [10]. Although the "DFT spreading" concept has a more elaborate transmitter and a less complex receiver than the FFH/OFDM concept, the overall terminal implementation complexity and, therefore, the overall system implementation complexity are similar.

The FFH/OFDM concept can accommodate e.g. UMTS Turbo-Coding [11], to further increase the exploitation of diversity and therefore to reduce the block error ratio (BLER) at a given average signal-to-noise ratio (SNR). The application of the UMTS Turbo-Code to FFH/OFDM is discussed in Sect. II.A. Then, Sect. II.B gives a concise explanation of the FFH/OFDM symbol generation, summarizing the detailed discussion of [9]. Furthermore, requirements on the signal transmission needed to facilitate a quasi coherent data detection at the receiver are discussed in Sect. II.C. In Sect. II.D, the discrete-time received vector associated with the FFH/OFDM symbol transmission is introduced.

In Sect. III, the authors first illustrate that the

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orthonormality between subcarriers can be easily restored at the FFH/OFDM receivers, which facilitates simple receivers. Like in the case of regular OFDM, the mentioned FFH/OFDM receivers deploy block linear equalizers (BLEs), cf. Sect. III. Besides a straight-forward zero-forcing block linear equalizer (ZF-BLE), related to the one presented in [12], an MMSE (minimum mean squared error)-BLE shall be briefly addressed. The MMSE-BLE of Sect. III, which is related to the equalizer presented in [12], is more effective than the ZF-BLE with respect to the achievable BLER at a given SNR. Therefore, said MMSE-BLE has been applied in the system simulations.

The BLER performance of the FFH/OFDM concept is compared to the BLER performance of regular OFDM, both applying the UMTS Turbo-Code in several mobile environments, see Sect. IV. It is illustrated that the FFH/OFDM concept outperforms regular OFDM.

Throughout the manuscript we consider the discrete-time version of the communication system in the equivalent low-pass domain. All signals are assumed to be regularly sampled with a sampling rate equal to the user bandwidth B_{u}. Complex quantities are underlined in the text and in the equations. Matrix-vector calculus is employed for convenience. Matrices and vectors are in boldface italics. Matrices and vectors are represented by upper-case and by lower-case characters, respectively. Complex conjugation, matrix inversion, transposition and the Hermitian operation are denoted by $(\cdot)^{*}$, $(\cdot)^{-1}$, $(\cdot)^{\mathrm{T}}$ and $(\cdot)^{\mathrm{H}}$, respectively.

II. FFH/OFDM CONCEPT

A. FEC coding and data modulation

In general, each FFH/OFDM symbol contains M_{i} binary information bits, u_{μ}, $\mu = 1 \cdots M_{\mathrm{i}}$, which have the average information bit energy E_{b} measured at the transmitter and which are fed into the FEC encoder. As mentioned above, the UMTS Turbo-Code [11] is considered. To generate the FFH/OFDM symbol, M_{u} information bits are fed into the UMTS Turbo-Code encoder which generates $(3M_{\mathrm{u}}+12)$ encoded bits, including the twelve termination bits defined in [11]. Of course, M_{u} may differ from M_{i}. In order to match the number $(3M_{\mathrm{u}}+12)$ of encoded bits to consecutive FFH/OFDM symbols, puncturing and rate matching are required, cf. the general procedure of [11]. After appropriate puncturing and rate matching, the number of resulting encoded bits is given by

$$M_{\mathrm{c}} \leq (3M_{\mathrm{u}}+12). \tag{1}$$

Let us denote these encoded bits by c_{ν}, $\nu = 1 \cdots M_{\mathrm{c}}$. Now, assume that the symbol alphabet, from which the complex-valued data symbols $\underline{d}_{m}$, $m = 1 \cdots M$, are taken, has the cardinality $|\underline{\mathrm{V}}|$. Hence, each data symbol carries $\mathrm{ld}(|\underline{\mathrm{V}}|)$ encoded bits. A viable approach is the use of $2^{\mathrm{ld}(|\underline{\mathrm{V}}|)}$-QAM as the data modulation. It is particularly advisable to choose $\mathrm{ld}(|\underline{\mathrm{V}}|)$ as an ***even*** integer because Gray encoding is easily accessible in this case. Using $2^{\mathrm{ld}(|\underline{\mathrm{V}}|)}$-QAM data modulation, each FFH/OFDM symbol carries $\mathrm{ld}(|\underline{\mathrm{V}}|)\cdot M$ encoded bits. In order to obtain the best possible match between M_{c} and $\mathrm{ld}(|\underline{\mathrm{V}}|)\cdot M$, we demand the puncturing and rate matching to fulfil the requirement

$$N_{\mathrm{FFH/OFDM}} = \frac{M_{\mathrm{c}}}{\mathrm{ld}(|\underline{\mathrm{V}}|)\cdot M} \in \mathbb{N}. \tag{2}$$

$N_{\mathrm{FFH/OFDM}}$ is the number of FFH/OFDM symbols required to carry the information contained in the M_{u} information bits mentioned above. A special case is given, when $N_{\mathrm{FFH/OFDM}}$ is equal to one, i.e. when the number of encoded bits M_{c} perfectly fit on M $2^{\mathrm{ld}(|\underline{\mathrm{V}}|)}$-QAM data symbols.

Generally, the resulting code rate is

$$R = \frac{M_{\mathrm{u}}}{N_{\mathrm{FFH/OFDM}}\cdot\left(\mathrm{ld}(|\underline{\mathrm{V}}|)\cdot M\right)}. \tag{3}$$

With (3), the energy per encoded bit is given by

$$E_{\mathrm{c}} = \frac{M_{\mathrm{u}}}{N_{\mathrm{FFH/OFDM}}\cdot\left(\mathrm{ld}(|\underline{\mathrm{V}}|)\cdot M\right)}E_{\mathrm{b}}. \tag{4}$$

In the remainder of this manuscript, we will only consider quaternary phase shift keying (QPSK) data modulation and will interpret it as 4-QAM.

B. FFH/OFDM symbol generation

In this section, a concise version of Sect. II published in [9] will be given. To facilitate implementation efficient fast Fourier transform (FFT) and inverse fast Fourier transform (IFFT) algorithms at the receivers and the transmitters, we shall assume that the number of subcarriers M is an integer power of 2:

$$M = 2^{\omega}, \quad \omega \in \mathbb{N}. \tag{5}$$

The complex-valued QPSK symbols are arranged in the data vector

$$\underline{\boldsymbol{d}} = (\underline{d}_1, \underline{d}_2 \cdots \underline{d}_M)^{\mathrm{T}}. \tag{6}$$

Each data vector $\underline{\boldsymbol{d}}$ is mapped onto a unique FFH/OFDM symbol. In the FFH/OFDM concept, this mapping results in the FFH/OFDM symbol vector $\underline{\boldsymbol{b}}$, which represents the discrete-time version of the FFH/OFDM symbol. The FFH operation can be regarded as a linear pre-coding, described by an $M\times M$ linear pre-coding matrix $\underline{\boldsymbol{U}}_{(M)}$. With $\underline{\boldsymbol{U}}_{(M)}$, the linearly pre-coded data symbols form the pre-coded vector

$$\underline{\boldsymbol{\xi}} = \underline{\boldsymbol{U}}_{(M)}\underline{\boldsymbol{d}}. \tag{7}$$

With the $M\times M$ matrix $\underline{\boldsymbol{D}}_{(M)}$ of the M point inverse discrete Fourier transform (IDFT), cf. e.g. pp. 78ff. of [13], having the element

$$\left[\underline{\boldsymbol{D}}_{(Q)}\right]_{\nu,\mu}=\frac{1}{\sqrt{Q}}\left(\exp\left\{\mathrm{j}\frac{2\pi}{Q}(\nu-1)\cdot(\mu-1)\right\}\right),\nu,\mu=1\cdots Q\,, \quad (8)$$

in the ν th row and the μ th column, the OFDM symbol vector is given by

$$\underline{\boldsymbol{b}}_{\mathrm{OFDM}}=\underline{\boldsymbol{D}}_{(M)}\underline{\boldsymbol{d}}\,. \quad (9)$$

The FFH/OFDM symbol vector, however, is obtained by applying the IDFT to $\underline{\xi}$ introduced in (7), yielding

$$\underline{\boldsymbol{b}}=\underline{\boldsymbol{D}}_{(M)}\underline{\boldsymbol{\xi}}=\underline{\boldsymbol{D}}_{(M)}\underline{\boldsymbol{U}}_{(M)}\underline{\boldsymbol{d}}\,. \quad (10)$$

The product $\underline{\boldsymbol{D}}_{(M)}\underline{\boldsymbol{U}}_{(M)}$ of the IDFT matrix $\underline{\boldsymbol{D}}_{(M)}$ and the linear pre-coding matrix $\underline{\boldsymbol{U}}_{(M)}$ yields a unitary $M\times M$ FFH matrix

$$\underline{\boldsymbol{D}}_{(M)\mathrm{H}}=\underline{\boldsymbol{D}}_{(M)}\underline{\boldsymbol{U}}_{(M)},$$

$$\left[\underline{\boldsymbol{D}}_{(M)\mathrm{H}}\right]_{\nu,\mu}=\frac{1}{\sqrt{M}}\left(\exp\left\{\mathrm{j}2\pi(\nu-1)\cdot\frac{\left[\boldsymbol{\Phi}_{(M)}\right]_{\nu,\mu}}{M}\right\}\right), \quad (11)$$

$$\nu,\mu=1\cdots M,$$

which is a shuffled version of $\underline{\boldsymbol{D}}_{(M)}$. The element $\left[\boldsymbol{\Phi}_{(M)}\right]_{\nu,\mu}$, $\nu,\mu=1\cdots M$, of the $M\times M$ matrix $\boldsymbol{\Phi}_{(M)}$ refers to the quasi instantaneous frequency associated with the QPSK symbol $\underline{d}_{\mu}$ at sample time instant ν. Hence, $\boldsymbol{\Phi}_{(M)}$ contains the FFH pattern. To allow the data detection, this FFH pattern must be known to the receiver.

In the case of regular OFDM, we find

$$\left[\boldsymbol{\Phi}_{(M)}\right]_{\nu,\mu}=(\mu-1),\quad \nu,\mu=1\cdots M\,. \quad (12)$$

Now, let f_{ν}, $\nu=1\cdots M$, refer to the quasi instantaneous frequencies associated with the first data symbol $\underline{d}_{1}$, at the time instants $\nu=1\cdots M$. For convenience, these frequency values can be arranged in the frequency vector

$$\boldsymbol{f}_{(M)}=\left(f_1,f_2\cdots f_M\right)^{\mathrm{T}}. \quad (13)$$

To faciliate a best possible exploitation of the frequency diversity potential of the mobile radio channel for a given user bandwidth B_{u}, $\boldsymbol{f}_{(M)}$ is chosen as a shuffled version of the basic frequency vector $\left(0,1\cdots(M-1)\right)^{\mathrm{T}}$.

With (13), a cyclic FFH pattern is found by

$$\left[\boldsymbol{\Phi}_{(M)}\right]_{\nu,\mu}=\mathrm{mod}\left(f_{\nu}+\mu-1,M\right),\quad \nu,\mu=1\cdots M\,, \quad (14)$$

where $\mathrm{mod}(\cdot,\cdot)$ denotes the modulus. Each data symbol is distributed over all subcarriers during the transmission of an FFH/OFDM symbol. In the case of a cyclic FFH pattern represented by (14), $\underline{\boldsymbol{U}}$ becomes a right circulant matrix defined by

$$\underline{\boldsymbol{U}}_{(M)}=\underline{\boldsymbol{D}}_{(M)}^{\mathrm{H}}\underline{\boldsymbol{\Delta}}_{(M)}\underline{\boldsymbol{D}}_{(M)}\,. \quad (15)$$

In (15), the time-domain matrix $\underline{\boldsymbol{\Delta}}_{(M)}$, which is also the inverse Fourier transform of $\underline{\boldsymbol{U}}_{(M)}$, is a ***diagonal*** $M\times M$ matrix. Its diagonal elements form the column vector

$$\underline{\boldsymbol{\delta}}_{(M)}=\mathrm{diag}\left(\underline{\boldsymbol{\Delta}}_{(M)}\right). \quad (16)$$

With $\odot$ denoting the Hadamard (element-wise) product and with $\underline{\boldsymbol{\Delta}}_{(M)}$ and $\underline{\boldsymbol{\delta}}_{(M)}$ defined by (15) and (16), respectively, the newly generated FFH/OFDM symbol $\underline{\boldsymbol{b}}$ of (10) can now be represented as follows:

$$\underline{\boldsymbol{b}}=\underline{\boldsymbol{D}}_{(M)\mathrm{H}}\underline{\boldsymbol{d}}=\underline{\boldsymbol{\Delta}}_{(M)}\underline{\boldsymbol{D}}_{(M)}\underline{\boldsymbol{d}}=\underline{\boldsymbol{\delta}}_{(M)}\odot\left(\underline{\boldsymbol{D}}_{(M)}\underline{\boldsymbol{d}}\right). \quad (17)$$

Obviously, the FFH operation is identical to a proper weighting of the samples in a regularly generated OFDM symbol $\underline{\boldsymbol{b}}_{\mathrm{OFDM}}$ in the time domain.

Now, $\underline{\boldsymbol{b}}$ of (10) and (17) is transmitted over the mobile radio channel considered as a W path channel. To facilitate a single-tap equalization technique like in regular OFDM, the deployment of either cyclic prefixes or cyclic postfixes is required. However, the well-known techniques can easily be adapted to FFH/OFDM.

C. Requirements on the transmission from the receiver perspective

To facilitate a quasi coherent data detection, channel estimation has to be accomplished at the receivers. The channel estimation can either be blind or pilot signal based, requiring the transmission of pilot signals which are known to both the transmitters and the receivers, respectively. Since the pilot signal based channel estimation strategy is more robust than its blind counterpart, the authors will only consider pilot signal based channel estimation. In OFDM systems, usually pilot subcarriers are deployed, i.e. the pilot signals are allocated to specially selected subcarriers, which are simultaneously transmitted together with the data carrying subcarriers. However, to enable quasi coherent data detection, regular OFDM systems require the reception of the whole OFDM symbol before the channel estimation and the data detection can be accomplished. Since the OFDM symbols must be received completely, an inherent receiver latency cannot be avoided.

A different approach to channel estimation, which has been known from mobile radio systems such as e.g. UMTS, uses pilot signals which are transmitted either prior to, in between or after the data and are therefore termed preambles, midambles or postambles, respectively. In contrast to other schemes, preamble based channel estimation has the lowest latency and, therefore, has been selected by the authors. The preambles used by the authors are derived from chirp signals. With the number of pilot samples N_{pilot} taken at a sampling rate of B_{u}, the pilot signal vector is given by

$$\underline{\boldsymbol{g}}=\left(1\cdots\exp\left\{\mathrm{j}\pi\frac{\left(n_{\mathrm{pilot}}-1\right)^2}{N_{\mathrm{pilot}}}\right\}\cdots\exp\left\{\mathrm{j}\pi\frac{\left(N_{\mathrm{pilot}}-1\right)^2}{N_{\mathrm{pilot}}}\right\}\right)^{\mathrm{T}}. \quad (18)$$

In (18), the necessary energy normalization has been neglected. The duration T_{pilot} equal to $N_{\mathrm{pilot}}/B_{\mathrm{u}}$ of the chirp signals must be equal to or greater than the excess delay of the multi-path channel. To facilitate a simple implementation of a maximum-likelihood (ML) channel estimator, we choose N_{pilot} equal to the lowest possible integer power of 2.

The FFH/OFDM transmission is arranged in slots. Each slot begins with a cyclic pilot prefix of duration T_{prefix} followed by the pilot signal. Then, the cyclic prefix of the FFH/OFDM symbol with duration T_{prefix} is transmitted. Finally, the FFH/OFDM symbol itself fills the remainder of the slot, yielding a slot duration of

$$T_{\text{slot}} = T_{\text{pilot}} + 2T_{\text{prefix}} + T_{\text{symbol}}. \tag{19}$$

D. Discrete-time received vector of the FFH/OFDM symbol

When using an appropriate cyclic prefix at the transmitter and a corresponding received signal truncation, the mobile radio channel can be characterized by the $M \times M$ ***diagonal*** channel matrix $\underline{\boldsymbol{H}}_{\text{f}}$, each element of $\underline{\boldsymbol{H}}_{\text{f}}$ being a sample of the transfer function of the mobile radio channel taken at the center of each subcarrier [9]. At the receiver input, noise, modelled as additive white Gaussian noise (AWGN) with double-sided spectral noise density $N_0/2$, impairs the reception. With $\underline{\boldsymbol{n}}_{\text{f}}$ being the noise vector in the frequency domain containing M samples, the received vector can be represented as [9]

$$\underline{\boldsymbol{e}}_{\text{f}} = \underline{\boldsymbol{H}}_{\text{f}} \underline{\boldsymbol{D}}_{(M)}^{\text{H}} \underline{\boldsymbol{\Delta}}_{(M)} \underline{\boldsymbol{D}}_{(M)} \underline{\boldsymbol{d}} + \underline{\boldsymbol{n}}_{\text{f}}. \tag{20}$$

In what follows, we assume that N_0 is a-priori known to the receiver.

III. Receiver concept with block linear equalizer (BLE)

Setting out from (17), we see that the FFH operation is equivalent to the complex weighting of each row vector of $\underline{\boldsymbol{D}}_{(M)}$. Since $\underline{\boldsymbol{D}}_{(M)\text{H}}$ given in (11) is unitary for cyclic FFH patterns, we find [9]

$$\underline{\boldsymbol{\Delta}}_{(M)}^{-1} = \underline{\boldsymbol{\Delta}}_{(M)}^{\text{H}} = \underline{\boldsymbol{\Delta}}_{(M)}^{*}. \tag{21}$$

Hence, the $M \times M$ matrix $\underline{\boldsymbol{\Delta}}_{(M)}$ is also unitary. Equalization can be accomplished by first applying a single-tap frequency domain equalizer [9],[12]

$$\underline{\boldsymbol{M}}_{\text{f}} = \underline{\boldsymbol{H}}_{\text{f}}^{-1} \tag{22}$$

to $\underline{\boldsymbol{e}}_{\text{f}}$ of (20), followed by an M point IDFT operation $\underline{\boldsymbol{D}}_{(M)}$, the inverse FFH (IFFH) operation $\underline{\boldsymbol{\Delta}}_{(M)}^{*}$, and a further M point DFT operation $\underline{\boldsymbol{D}}_{(M)}^{\text{H}}$, yielding the estimate $\hat{\underline{\boldsymbol{d}}}$ of $\underline{\boldsymbol{d}}$, see (6) [9]:

$$\hat{\underline{\boldsymbol{d}}} = \underline{\boldsymbol{D}}_{(M)}^{\text{H}} \underline{\boldsymbol{\Delta}}_{(M)}^{*} \underline{\boldsymbol{D}}_{(M)} \underline{\boldsymbol{H}}_{\text{f}}^{-1} \underline{\boldsymbol{e}}_{\text{f}}. \tag{23}$$

The FFH/OFDM equalizer $\underline{\boldsymbol{D}}_{(M)}^{\text{H}} \underline{\boldsymbol{\Delta}}_{(M)}^{*} \underline{\boldsymbol{D}}_{(M)} \underline{\boldsymbol{H}}_{\text{f}}^{-1}$ carries out the zero-forcing equalization based on the received vector containing a data block. It is therefore also termed zero-forcing block linear equalizer (ZF-BLE) [9]. For a most efficient implementation, the ZF-BLE contains the channel equalization in the frequency domain and the IFFH operation in the time-domain [9].

Setting out from the MMSE criterion and with (4) we find

$$\hat{\underline{\boldsymbol{d}}} = \underline{\boldsymbol{D}}_{(M)}^{\text{H}} \underline{\boldsymbol{\Delta}}_{(M)}^{*} \underline{\boldsymbol{D}}_{(M)} \left(\underline{\boldsymbol{H}}_{\text{f}} + \frac{\underline{\boldsymbol{H}}_{\text{f}}^{\text{H}-1}}{\left(E_{\text{c}}/N_0\right)} \right)^{-1} \underline{\boldsymbol{e}}_{\text{f}}, \tag{24}$$

cf. also [9]. Although the MMSE-BLE represented by $\underline{\boldsymbol{D}}_{(M)}^{\text{H}} \underline{\boldsymbol{\Delta}}_{(M)}^{*} \underline{\boldsymbol{D}}_{(M)} \left(\underline{\boldsymbol{H}}_{\text{f}} + \underline{\boldsymbol{H}}_{\text{f}}^{\text{H}-1} / \left(E_{\text{c}}/N_0\right)\right)^{-1}$ does not restore the aforementioned orthonormality it allows a lower BLER at a given SNR than the ZF-BLE.

Approximated log-likelihood ratios (LLR's) to be fed into the channel decoder are based on the estimate $\hat{\underline{\boldsymbol{d}}}$ at the BLE output, resulting in continuous-valued estimates of the encoded bits. These approximated LLR values are then fed into a max-log-MAP (maximum a-posteriori) based Turbo-decoder, cf. e.g. [14].

Regular OFDM performs best when an MF based BLE is used. However, in the case of FFH/OFDM, deploying a simple MF based BLE results in considerable intersymbol interference which degrades the performance. Therefore, the MMSE based BLE was found to be the best solution in combination with Turbo-Coding.

IV. Simulation results

The BLER performance of the considered OFDM and FFH/OFDM concepts, both using the UMTS Turbo-Code, was studied in simulations as a function of the average E_{b}/N_0. The carrier frequency was f_{c} equal to $1.9\ \text{GHz}$ and the user bandwidth B_{u} equaled $5\ \text{MHz}$. The number of subcarriers M was 256. Each OFDM symbol and each FFH/OFDM symbol therefore carried M equal to 256 QPSK symbols. The number of information bits per Turbo-Code frame M_{u} was 167. At least, one thousand block errors were counted per selected average E_{b}/N_0 value. In the case of FFH/OFDM, cyclic FFH was used; the chosen frequency sequence $\boldsymbol{f}_{(M)}$ given in (13) allowed the frequency hopping of each data symbol over all M subcarriers.

The simulation results refer to the ETSI/ITU channel models Vehicular A (VA) and Indoor Office A (IOA), which are defined in [15]. The mobile speed was $100\ \text{km/h}$ in the case of VA and $3\ \text{km/h}$ in the case of IOA. The corresponding excess delay values were $6.6\ \mu\text{s}$ and $4.4\ \mu\text{s}$ in the case of VA and IOA, respectively. Therefore, the slot durations T_{slot} introduced in (19) were $77.2\ \mu\text{s}$ and $66.4\ \mu\text{s}$ in the case of VA and IOA, respectively. The channels are almost time invariant during a slot. The simulation results are shown in Fig. 1.

First, the receiver had perfect knowledge of $\underline{\boldsymbol{H}}_{\text{f}}$. In the case of regular OFDM, the SNR values, which are at least required for a BLER equal to 10^{-2}, are $\approx 11.1\ \text{dB}$ and $\approx 18\ \text{dB}$ in the case of VA and IOA. When FFH/OFDM is considered, the corresponding SNR values are $\approx 10.1\ \text{dB}$ and

≈ 15.5 dB for VA and IOA. The performance gains of the FFH/OFDM concept are ≈ 1 dB and ≈ 2.5 dB, respectively, at a BLER of 10^{-2} in the case of VA and IOA. As expected the FFH/OFDM concept outperforms the regular OFDM concept.

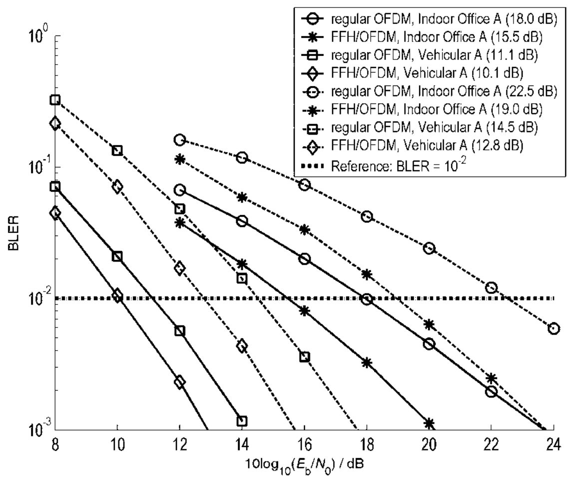

Fig. 1 Simulation results showing the block error ratio (BLER) versus the signal-to-noise ratio (SNR); with UMTS Turbo-Code;
solid lines (-): perfect knowledge of the channel at the receiver;
dashed lines (--): with ML channel estimation.

Since the IOA channel has a lower excess delay and, consequently, a lower delay spread, than the VA channel, its frequency diversity potential is lower within the user bandwidth B_u equal to 5 MHz [9]. Thus, the variance of the LLR values generated at the equalizer output in the IOA channel is larger than in the VA channel. For this reason, the IOA related performance curves require a larger SNR for a given BLER than the VA related curves. Notice, however, that the LLR values are perfect in the case of regular OFDM whereas an approximation is used in the case of FFH/OFDM. The approximated LLR values have less systematic errors in the case of channels which similarize single path channels, e.g. like the IOA channel at B_u equal to 5 MHz, than in the case of channels with significant multi-path components, like e.g. the VA channel B_u equal to 5 MHz. Therefore, compared to the case of the IOA channel, a lower performance gain of the FFH/OFDM concept is to be expected in the case of the VA channel. In addition, the LLR variance associated with regular OFDM is larger than the LLR variance associated with the FFH/OFDM concept in the case of the IOA channel because the FFH/OFDM concept benefits from the exploitation of the frequency diversity potential already at the BLE. Owing to the larger delay spread of the VA channel, the differences in variance of the LLR values are smaller when comparing regular OFDM and the FFH/OFDM concept. Therefore, the performance gain of the FFH/OFDM concept over regular OFDM is larger in the case of the IOA channel than in the case of the VA channel.

Now, ML channel estimation is used at the receiver. In the case of regular OFDM, the SNR values, which are at least required for a BLER equal to 10^{-2}, are ≈ 14.5 dB and ≈ 22.5 dB for VA and IOA. When FFH/OFDM is taken, the corresponding SNR values are ≈ 12.8 dB and ≈ 19.0 dB in the case of VA and IOA, respectively. Still, as expected, the FFH/OFDM concept outperforms the regular OFDM concept. The channel estimation errors cause SNR degradations of $\approx 2.7 \ldots 3.5$ dB in the case of the FFH/OFDM concept and $\approx 3.4 \ldots 4.5$ dB in the case of regular OFDM, clearly showing the greater robustness of the FFH/OFDM concept.

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