

Low Complexity Equalization with and without Decision Feedback and its Application to IFDMA

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Abstract—The Interleaved Frequency Division Multiple Access (IFDMA) scheme is a promising candidate for next generation mobile radio systems. IFDMA is based on compression, repetition and subsequent user dependent frequency shift of a modulated signal. As in OFDMA, multiple access is enabled by the assignment of overlapping but mutually orthogonal subcarriers to each user. It combines the advantages of single and multi carrier transmission such as low peak to average power ratio, orthogonality of the signals of different users even for transmission over a time dispersive channel and low complexity. In this paper, a linear low complexity frequency domain equalizer for IFDMA is presented and extended by subsequent decision feedback with initialization. Simulation results show the good performance for data transmission using IFDMA with frequency domain equalization over a mobile radio channel.

I. INTRODUCTION

Presently, research for beyond 3rd generation mobile radio systems is in progress world wide. A future mobile radio system has to meet challenging requirements. On the one hand it should enable different types of services from data rates of a few kbit/s up to several Mbit/s. Moreover, it should provide high flexibility and granularity as well as high performance. On the other hand, low cost and hence, low complexity implementation is requested, especially for mobile terminals.

High rate data transmission generally implies a small symbol duration. For transmission over time dispersive channels with a large maximum channel delay compared to the symbol duration many consecutive data symbols are affected by inter symbol interference. Thus, common time domain equalization methods based on, e.g., the Viterbi algorithm, become very complex. One method to overcome this problem is the use of multicarrier transmission. A well-established representative for multicarrier transmission is OFDMA. It provides low computational complexity and good performance at the same time. Orthogonality of the signals for different users is not affected by transmission over time dispersive channels. Furthermore, by appropriate choice of the guard interval for OFDMA only rough time synchronization is necessary. However, OFDMA signals are sensitive to frequency offsets and provide high envelope fluctuations.

Another solution is block transmission of single carrier signals. If subsequent data blocks are assumed to be separated by a cyclic prefix, low complexity equalization is possible by the use of linear frequency domain equalizers [9]. A well known representative for single carrier transmission is

DS-CDMA. It provides low envelope fluctuations and good robustness to frequency offsets. However, for transmission over time dispersive channels orthogonality of the signals of different users is lost and computationally complex algorithms for user separation are necessary.

A further promising candidate for a future mobile radio system is Interleaved Frequency Division Multiple Access (IFDMA) [1] [14]. It is based on compression, repetition and subsequent user dependent frequency shift of a modulated signal. As in OFDMA, multiple access is enabled by the assignment of overlapping but mutually orthogonal subcarriers to each user. IFDMA combines the advantages of spread-spectrum signals such as, e.g., DS-CDMA, and multi carrier solutions such as, e.g., OFDMA i.e., it provides a low peak to average power ratio as well as low complexity for user separation, because also for IFDMA the orthogonality of the signals is not affected by transmission over a time-dispersive channel [1]. However, IFDMA is closely related to OFDMA [12], [10], [11] and hence, it is also expected to be sensitive to frequency offsets [16], [12]. Also for IFDMA linear low complexity equalization in frequency domain is possible [15]. In this paper, a linear low complexity frequency domain equalizer for IFDMA is described. As linear equalization is suboptimum, further improvement of the frequency domain equalizer by subsequent non-linear decision feedback equalization in time domain is introduced.

The paper is organized as follows. In Section II a short description of IFDMA is given. It is recalled that IFDMA transmission can be described by a cyclic overall channel model that can also be applied to OFDMA or blockwise transmission of DS-CDMA signals. In Section III, based on the cyclic overall channel model a linear frequency domain equalizer is introduced for IFDMA. It is shown how linear frequency domain equalization for signals with cyclic prefix can be extended by decision feedback. Finally, in Section IV simulation results for IFDMA transmission over a mobile radio channel are given for different low complexity equalizers.

II. SYSTEM MODEL

In this section, an IFDMA system model consisting of IFDMA modulation, block transmission of the IFDMA signal over a time dispersive channel and IFDMA demodulation is described. Based on a block transmission model a cyclic overall channel model including IFDMA modulation and

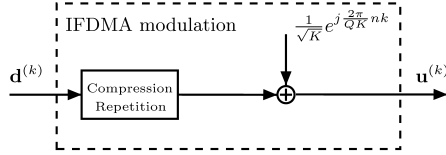


Fig. 1: IFDMA Modulation

demodulation is presented. The overall channel model is used in order to introduce frequency domain equalization (FDE) for IFDMA.

In the following, all signals are represented by their discrete time equivalents in the low-pass domain. We consider an IFDMA system with K users and user indices $k = 0, \dots, K - 1$. The data vector $\mathbf{d}^{(k)}$ of user k consists of Q linearly modulated data symbols at symbol rate $1/T_s$ and is given by

$$\mathbf{d}^{(k)} = (d_0^{(k)}, \dots, d_{Q-1}^{(k)})^T. \quad (1)$$

According to [1], IFDMA signal generation is accomplished by compression in time of vector $\mathbf{d}^{(k)}$ by factor K and subsequent K -fold repetition of the compressed signal. Thereafter, a user specific rotation corresponding to a shift in frequency domain is applied to the compressed and repeated signal. Hence, an IFDMA signal of length $N = KQ$ denoted as

$$\mathbf{u}^{(k)} = (u_0^{(k)}, \dots, u_{N-1}^{(k)})^T \quad (2)$$

consists of N chips $u_n^{(k)}$, $n = 0, \dots, N - 1$, at chip rate $1/T_c = K/T_s$ given by

$$u_n^{(k)} = \frac{1}{\sqrt{K}} \cdot d_{n \bmod Q}^{(k)} \cdot e^{j \frac{2\pi}{K} nk}; \quad n = 0, \dots, N - 1, \quad (3)$$

where mod designates modulo operation and $1/\sqrt{K}$ is a normalization factor, cf. Fig. 1. Compression of the data vector in time by factor K can be interpreted as a spreading in frequency domain by the same factor [1]. On the one hand, the K repetitions of the compressed data vector lead to the formation of Q equidistant local maxima distributed over the whole bandwidth $B = 1/T_c$ of the IFDMA signal. The distance between adjacent local maxima in the signal of one user is given by $\Delta f = B/Q = 1/(QT_c)$. On the other hand, inbetween adjacent maxima $K - 1$ equidistant zeros occur. The distance between adjacent zeros is given by $\frac{B}{KQ} = \frac{1}{T_c KQ}$ [1]. Hence, the signals of $K - 1$ additional users can be accommodated in terms of multiple access by a user specific frequency shift $k \cdot \Delta f$. Signals of different users remain orthogonal to each other as long as the maxima of the spectrum of one user exactly coincide with the zeros in the spectra of the other users, i.e., as long as perfect frequency synchronization can be assumed [1].

In order to avoid inter block interference, subsequent IFDMA blocks are assumed to be separated by a cyclic prefix (CP) that exceeds the maximum channel delay. For IFDMA, a cyclic prefix can be implemented easily by increasing the repetition factor. We consider a repetition factor of $K + K_\Delta$ where $K_\Delta \cdot Q$ represents the number of chips at the beginning of the IFDMA signal that can be interpreted as cyclic

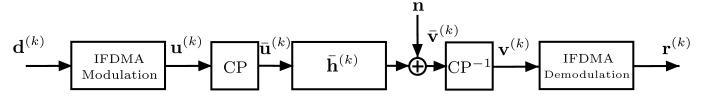


Fig. 2: IFDMA system

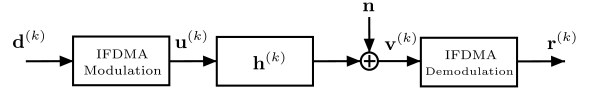


Fig. 3: IFDMA system with cyclic channel

prefix. Hence, $K_\Delta \cdot Q$ has to be an integer. An IFDMA signal according to

$$\bar{\mathbf{u}}^{(k)} = (\bar{u}_{-K_\Delta Q}^{(k)}, \dots, \bar{u}_{KQ-1}^{(k)})^T \quad (4)$$

with cyclic prefix consists of $\bar{N} = (K + K_\Delta)Q$ chips at chip rate $1/T_c = K/T_s$. The chips are given by

$$\bar{u}_n^{(k)} = \frac{1}{\sqrt{K}} \cdot d_{n \bmod Q}^{(k)} \cdot e^{j \frac{2\pi}{K} nk}; \quad n = -K_\Delta Q, \dots, KQ - 1. \quad (5)$$

We introduce transmission over a channel including transmit filter, physical time dispersive channel and receive filter. This channel is modeled by a finite impulse response (FIR) filter of length M . During transmission of one data block the channel is assumed to be time invariant. The channel is given by

$$\bar{\mathbf{h}}^{(k)} = (\bar{h}_0^{(k)}, \dots, \bar{h}_{M-1}^{(k)})^T. \quad (6)$$

Additionally, we define an additive white Gaussian noise (AWGN) vector $\bar{\mathbf{n}}$ of length \bar{N} according to

$$\bar{\mathbf{n}} = (\bar{n}_{-K_\Delta Q}, \dots, \bar{n}_{KQ+M-1})^T. \quad (7)$$

The received signal before removal of the cyclic prefix

$$\bar{\mathbf{v}}^{(k)} = (\bar{v}_{-K_\Delta Q}^{(k)}, \dots, \bar{v}_{KQ+M-1}^{(k)})^T \quad (8)$$

is given by

$$\bar{\mathbf{v}}^{(k)} = \bar{\mathbf{u}}^{(k)} * \bar{\mathbf{h}}^{(k)} + \bar{\mathbf{n}}, \quad (9)$$

cf. Fig. 2. In order to describe the effect of the cyclic prefix we define a channel

$$\mathbf{h}^{(k)} = (h_0^{(k)}, \dots, h_{M-1}^{(k)})^T \quad (10)$$

with $h_n^{(k)} = \bar{h}_n^{(k)}$ for $n = 0, \dots, M - 1$ that includes the channel given by Eq. (6) as well as insertion and removal of the cyclic prefix, cf. Fig. 2. Additionally, we define a noise vector after removal of the cyclic prefix according to

$$\mathbf{n} = (n_0, \dots, n_{N-1})^T. \quad (11)$$

According to [2], transmission of an arbitrary digital sequence with cyclic prefix over a channel as given by Eq. (10) can be described by a cyclic convolution with the channel impulse

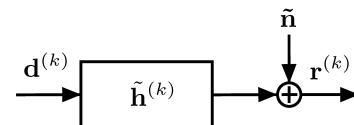


Fig. 4: IFDMA system with modified cyclic overall channel

response. Thus, the received IFDMA signal after removal of the cyclic prefix

$$\mathbf{v}^{(k)} = (v_0^{(k)}, \dots, v_{N-1}^{(k)})^T \quad (12)$$

is given by

$$\mathbf{v}^{(k)} = \mathbf{u}^{(k)} \circledast \mathbf{h}^{(k)} + \mathbf{n}, \quad (13)$$

where \circledast designates cyclic convolution, cf. Fig. 3.

In the following, a modified cyclic channel shall be described that contains IFDMA modulation, transmission over $\mathbf{h}^{(k)}$ and IFDMA demodulation, cf. Fig. 4. At the IFDMA demodulator, for each user the user specific rotation is reversed and the chips that belong to one data symbol $d_q^{(k)}$, $q = 0, \dots, Q-1$ are added up. According to [1] the elements of the demodulated IFDMA signal vector

$$\mathbf{r}^{(k)} = (r_0^{(k)}, \dots, r_{Q-1}^{(k)})^T \quad (14)$$

are given by

$$r_q^{(k)} = \frac{1}{\sqrt{K}} \sum_{l=0}^{K-1} v_{lQ+q}^{(k)} \cdot e^{-j \frac{2\pi}{QK} (lQ+q)k}, \quad q = 0, \dots, Q-1. \quad (15)$$

An IFDMA system according to Fig. 2 can be described by a modified cyclic overall channel

$$\tilde{\mathbf{h}}^{(k)} = (\tilde{h}_0^{(k)}, \dots, \tilde{h}_{Q-1}^{(k)})^T \quad (16)$$

at symbol rate $1/T_s = 1/(KT_c)$ in time domain [3]. Let $\lfloor x \rfloor$ designate the largest integer smaller than or equal to x . Using Eq. (2), (13) and (15), for $Q > M$ the elements of the modified channel can be described by

$$\tilde{h}_q^{(k)} = \begin{cases} h_q^{(k)} \cdot e^{j \frac{2\pi}{QK} qk} & , \quad q \leq M \\ 0 & , \quad M < q < Q \end{cases}, \quad (17)$$

whereas for $Q \leq M$ the elements of the modified channel are given by

$$\tilde{h}_q^{(k)} = \sum_{l=0}^{\lfloor \frac{M}{Q} \rfloor} h_{q+lQ}^{(k)} \cdot e^{j \frac{2\pi}{QK} (q+lQ)k}, \quad q < Q. \quad (18)$$

The elements of the equivalent noise vector

$$\tilde{\mathbf{n}} = (\tilde{n}_0, \dots, \tilde{n}_{Q-1})^T \quad (19)$$

after IFDMA demodulation are given by

$$\tilde{n}_q = \frac{1}{\sqrt{K}} \sum_{l=K_\Delta}^{K+K_\Delta-1} n_{lQ+q} \cdot e^{-j \frac{2\pi}{QK} lk}; \quad q = 0, \dots, Q-1. \quad (20)$$

Using the modified cyclic overall channel given by Eq. (16) and the noise given by Eq. (20), blockwise data transmission using IFDMA can be expressed by a simple vector/matrix equation at symbol rate $1/T_s$. Let $\tilde{\mathbf{H}}^{(k)}$ designate the circulant convolution matrix whose first row is given by the modified channel vector from Eq. (16). Thus, the vector of received data symbols is given by

$$\mathbf{r}^{(k)} = \tilde{\mathbf{H}}^{(k)} \cdot \mathbf{d}^{(k)} + \tilde{\mathbf{n}}, \quad (21)$$

cf. Fig. 4. The model given in Eq. 21 is not only applicable for IFDMA but also for, e.g., OFDMA and CDMA. It is known that also for block transmission of DS-CDMA signals with cyclic prefix a cyclic overall channel representation in time domain at symbol rate $1/T_s = 1/(KT_c)$ can be obtained if the channel including insertion and removal of the cyclic prefix, is extended by CDMA modulation and demodulation, i.e., spreading and despreading operation. OFDMA is known as a transmission scheme of data symbols $d_q^{(k)}$, $q = 0, \dots, Q-1$ in frequency domain [4]. Thus, for OFDMA with cyclic prefix a cyclic overall channel representation in time domain at symbol rate $1/T_s$ is given by combination of the channel including insertion and removal of the cyclic prefix, Inverse Discrete Fourier Transform (IDFT) and Discrete Fourier Transform (DFT).

III. EQUALIZATION

A. Frequency Domain Equalization

It is well known that frequency domain equalization initially investigated by [5] for cyclic channels provides low computational complexity. In this section, a frequency domain equalizer for IFDMA [15] based on the modified cyclic overall channel $\tilde{\mathbf{H}}^{(k)}$, cf. Eq. (21), is presented. As a similar model is also valid for OFDMA and CDMA, the equalization concept described in the following is also applicable for IFDMA and CDMA.

In the following, we assume perfect knowledge of the channel coefficients. Let \mathbf{F}_Q designate a $(Q \times Q)$ Discrete Fourier Transform (DFT) matrix and let \mathbf{F}_Q^{-1} designate the corresponding $(Q \times Q)$ Inverse Discrete Fourier Transform (IDFT) matrix. The received signal of user k in frequency domain is given by

$$\begin{aligned} \mathbf{F}_Q \mathbf{r}^{(k)} &= \mathbf{F}_Q \tilde{\mathbf{H}}^{(k)} \mathbf{d}^{(k)} + \mathbf{F}_Q \tilde{\mathbf{n}} \\ &= \mathbf{F}_Q \tilde{\mathbf{H}}^{(k)} \mathbf{F}_Q^{-1} \mathbf{F}_Q \mathbf{d}^{(k)} + \mathbf{F}_Q \tilde{\mathbf{n}}. \end{aligned} \quad (22)$$

As complex exponentials are eigenfunctions of $\tilde{\mathbf{H}}^{(k)}$, $\mathbf{F}_Q \tilde{\mathbf{H}}^{(k)} \mathbf{F}_Q^{-1}$ results in a diagonal matrix [2] according to

$$\begin{aligned} \mathbf{D}^{(k)} &= \text{diag}\{D_0^{(k)}, \dots, D_{Q-1}^{(k)}\} \\ &= \mathbf{F}_Q \tilde{\mathbf{H}}^{(k)} \mathbf{F}_Q^{-1}. \end{aligned} \quad (23)$$

A frequency domain equalizer for IFDMA can be described by a $(Q \times Q)$ matrix $\mathbf{E}^{(k)}$ according to

$$\mathbf{E}^{(k)} = \text{diag}\{E_0^{(k)}, \dots, E_{Q-1}^{(k)}\}. \quad (24)$$

Let \mathbf{I}_Q designate the $(Q \times Q)$ identity matrix. The zero-forcing condition is given by

$$\mathbf{D}^{(k)} \cdot \mathbf{E}_{ZF}^{(k)} = \mathbf{I}_Q. \quad (25)$$

If matrix $\mathbf{D}^{(k)}$ is non-singular Eq. (25) results in

$$\mathbf{E}_{ZF}^{(k)} = (\mathbf{D}^{(k)})^{-1}. \quad (26)$$

Thus, the diagonal elements of the zero-forcing equalizer matrix $\mathbf{E}_{ZF}^{(k)}$ are given by

$$E_{ZF,q}^{(k)} = \frac{1}{D_q^{(k)}}, \quad q = 0, \dots, Q-1. \quad (27)$$

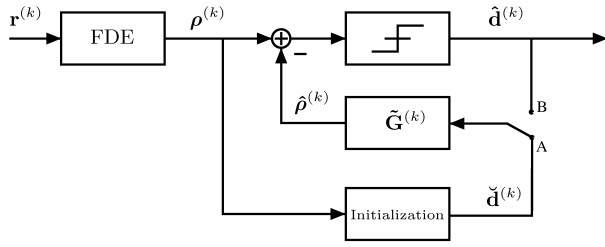


Fig. 5: FDE with subsequent decision feedback

A linear equalizer based on the MMSE criterion can be derived by

$$E\{|\mathbf{r}^{(k)} - \mathbf{d}^{(k)}|^2\} \xrightarrow{!} \min \quad (28)$$

and results in

$$E_{MMSE,q}^{(k)} = \frac{D_q^{*(k)}}{|D_q^{(k)}|^2 + \sigma_n^2/\sigma_d^2}, \quad q = 0, \dots, Q-1, \quad (29)$$

where x^* designates the conjugate of x and where σ_n^2 and σ_d^2 represent the variances of the noise $\tilde{\mathbf{n}}$ and of the input signal $\mathbf{d}^{(k)}$, respectively [5].

B. Decision Feedback Equalization

Linear equalization like the FDE-approach described in Section III-A or in [15] is suboptimum for equalization of time dispersive channels [7]. As optimum equalization is very complex, a common method to improve the receiver performance is subsequent non-linear time domain equalization based on the decision feedback principle [6]. For a non-cyclic channel the elements of the output vector of the decision feedback equalizer can be calculated iteratively. For a cyclic channel as it is the case for block transmission using cyclic extensions the first elements of output signal depend on the last ones. For that reason an iterative calculation is not straightforward. One method to overcome this problem is to use unique words (UW) instead of a cyclic prefix [6]. In the following, a solution is presented that is suitable for block transmission with cyclic prefix as it is the case for IFDMA, but also for CDMA and OFDMA. Before a decision for the elements of the output vector of the decision feedback equalizer is made, an initialization is run in order to estimate the elements that are required for decision feedback.

We consider a frequency domain equalizer with subsequent decision feedback according to Fig. 5. A linear frequency domain equalizer as described in Section III-A is used as feed forward filter. The feedback filter is chosen as a combination of the modified channel given by Eq. (16) and the equalizer given by Eq. (26). It can be described as

$$\tilde{\mathbf{G}}^{(k)} = \mathbf{F}_Q^{-1} \mathbf{E}^{(k)} \mathbf{D}^{(k)} \mathbf{F}_Q. \quad (30)$$

Let

$$\boldsymbol{\rho}^{(k)} = (\rho_0^{(k)}, \dots, \rho_{Q-1}^{(k)})^T \quad (31)$$

designate the output of the feed forward filter in time domain. $\boldsymbol{\rho}^{(k)}$ is given by

$$\boldsymbol{\rho}^{(k)} = \mathbf{F}_Q^{-1} \mathbf{E}^{(k)} \mathbf{F}_Q \mathbf{r}^{(k)}. \quad (32)$$

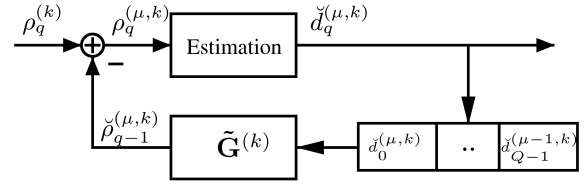


Fig. 6: Initialization algorithm

The noise at the output of the feed forward filter is given by

$$\begin{aligned} \tilde{\mathbf{n}} &= (\tilde{n}_0, \dots, \tilde{n}_{Q-1})^T \\ &= \mathbf{F}_Q^{-1} \mathbf{E}^{(k)} \mathbf{F}_Q \tilde{\mathbf{n}} \end{aligned} \quad (33)$$

The output of the equalizer after decision feedback can be described by

$$\hat{\mathbf{d}}^{(k)} = (\hat{d}_0^{(k)}, \dots, \hat{d}_{Q-1}^{(k)})^T. \quad (34)$$

In the following, an iterative initialization algorithm is used in order to estimate the elements of vector $\mathbf{d}^{(k)}$. For $M < Q$ only the elements $d_0^{(k)}, \dots, d_{M-1}^{(k)}$ and for $M \geq Q$ all elements of vector $\mathbf{d}^{(k)}$ have to be estimated. The algorithm is very close to an iterative equalization method described in [8]. The principle is depicted in Fig. 6. Let

$$\tilde{\mathbf{g}}^{(k)} = (\tilde{g}_0, \dots, \tilde{g}_{Q-1})^T. \quad (35)$$

designate the vector containing the Q channel coefficients in the first column of the circulant channel matrix given by Eq. (30). We consider one iteration step with index μ . For $q = 0, \dots, Q-1$ from $\rho_q^{(\mu,k)}$ estimates $\check{d}_q^{(\mu,k)}$ for the Q elements of the data vector $\mathbf{d}^{(k)}$ are calculated. The Q estimates are stored in memory. In order to reduce inter symbol interference, from each element $\rho_q^{(k)}$ of the output vector of the feedforward filter the $q-1$ -th element $\check{\rho}_{q-1}^{(\mu,k)}$ of a feedback signal is subtracted. The elements of the feedback signal are given by

$$\check{\rho}_{q-1}^{(\mu,k)} = \sum_{n=1}^q \tilde{g}_n^{(k)} \check{d}_{q-n}^{(\mu,k)} + \sum_{n=q+1}^{Q-1} \tilde{g}_n^{(k)} \check{d}_{q-n}^{(\mu-1,k)}. \quad (36)$$

For $\mu < 0$ it is assumed that $\check{d}_q^{(\mu,k)} = 0$. Hence, the iteration can be described by

$$\rho_q^{(\mu,k)} = \sum_{m=1}^{Q-1} \tilde{g}_m^{(k)} d_{q-m}^{(k)} + \tilde{n}_q - \check{\rho}_{q-1}^{(\mu,k)}. \quad (37)$$

In the following, the estimation process is described. From Eq. (37) we obtain

$$\frac{\rho_q^{(\mu,k)}}{\tilde{g}_0^{(k)}} = d_q^{(k)} + \frac{1}{\tilde{g}_0^{(k)}} \underbrace{\left(\sum_{m=1}^{Q-1} \tilde{g}_m^{(k)} d_{q-m}^{(k)} + \tilde{n}_q - \check{\rho}_{q-1}^{(\mu,k)} \right)}_{\tilde{\mathbf{n}}}. \quad (38)$$

According to [8] $\tilde{\mathbf{n}}$ can be modeled as a complex Gaussian stochastic process. For $d_q^{(k)}$ we assume QPSK symbols with $d_q^{(k)} \in \{\pm 1 \pm j\}$. Hence, the real part and the imaginary part of $d_q^{(k)}$ can be estimated independently and the estimation reduces to a binary estimation problem. The optimum estimate based

on the MMSE criterion of a binary estimation problem [13] is given by

$$\check{d}_q^{(\mu,k)} = E\{d_q^{(k)}|\rho_q^{(\mu,k)}\} \quad (39)$$

and can be solved using the Bayes theorem. Let \check{n}_{Re} and \check{n}_{Im} designate the real and the imaginary part of the complex stochastic process \check{n} , respectively. The estimate $\check{d}_q^{(\mu,k)}$ of the data symbol $d_q^{(k)}$ is given by

$$\check{d}_q^{(\mu,k)} = \tanh \frac{\Re\{\rho_q^{(\mu,k)}\} - E\{\check{n}_{Re}\}}{\sigma_{\check{n}_{Re}}^2 h_0^{(k)}} + j \cdot \tanh \frac{\Im\{\rho_q^{(\mu,k)}\} - E\{\check{n}_{Im}\}}{\sigma_{\check{n}_{Im}}^2 h_0^{(k)}}, \quad (40)$$

where $\sigma_{\check{n}_{Re}}^2$ and $\sigma_{\check{n}_{Im}}^2$ designate the variances of real and imaginary part of the complex stochastic process \check{n} , respectively. The values for $\sigma_{\check{n}_{Re}}^2$ and $\sigma_{\check{n}_{Im}}^2$ are given by

$$\sigma_{\check{n}_{Re}}^2 = \sigma_{\check{n}_{Im}}^2 = \frac{\sum_{m=1}^{Q-1} |\tilde{g}_m^{(k)}|^2}{2(\tilde{g}_0^{(k)})^2} + \frac{\sigma_{\check{n}}^2}{2(\tilde{g}_0^{(k)})^2}, \quad (41)$$

where $\sigma_{\check{n}}^2$ designates the variance of the noise at the output of the feed forward filter. As the data symbols $d_q^{(k)}$ as well as the noise at the output of the feed forward filter are assumed to be zero-mean, $E\{\check{n}_{Re}\}$ and $E\{\check{n}_{Im}\}$ are given by

$$E\{\check{n}_{Re}\} = -\frac{1}{\tilde{g}_0^{(k)}} \Re \left\{ \sum_{n=1}^{Q-1} \tilde{g}_n^{(k)} \check{d}_{q-n}^{(\mu,k)} \right\} \quad (42)$$

and

$$E\{\check{n}_{Im}\} = -\frac{1}{\tilde{g}_0^{(k)}} \Im \left\{ \sum_{n=1}^{Q-1} \tilde{g}_n^{(k)} \check{d}_{q-n}^{(\mu,k)} \right\}. \quad (43)$$

IV. SIMULATION RESULTS AND CONCLUSION

In Fig. 7 simulation results for uncoded IFDMA transmission over a mobile radio channel with different frequency domain equalizers are compared to each other. The simulation parameters are given in Table I. As expected, MMSE frequency domain equalization (FDE) without decision feedback equalizer (DFE) provides better performance than FDE without DFE based on the zero forcing condition. The performance of the MMSE based FDE can be further improved using subsequent DFE with a feedback filter according to Eq. (30). For MMSE based FDE with DFE and initialization a performance improvement of about 1 dB can be achieved compared to MMSE based FDE without DFE. As for one iteration step performance results are very similar compared to results for initialization with more than one iteration step, for practical implementation one iteration step will be sufficient. Further investigations may extend the given equalization concept to coded transmission.

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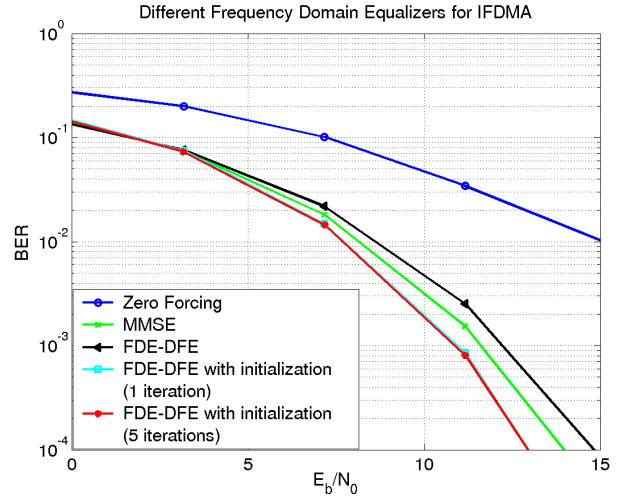


Fig. 7: Simulation Results

TABLE I: Simulation parameters

Bandwidth	20 MHz	Carrier frequency	5 GHz
Datarate per user	1.25 Mbit/s	Guard interval	7 μ s
Number of users	16	Channel model	COST 207
Block length	64 symbols		TU, 70 km/h

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