MULTI-CARRIER CDMA IN INDOOR WIRELESS RADIO NETWORKS

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Abstract

This paper examines a novel digital modulation/multiple access technique called Multi-Carrier Code Division Multiple Access (MC-CDMA) where each data symbol is transmitted at multiple narrowband subcarriers. Each subcarrier is encoded with a phase offset of 0 or π based on a spreading code. Analytical results are presented on the performance of this modulation scheme in an indoor wireless multipath radio channel.

Introduction

This paper examines the performance of a new spread spectrum transmission method called "MC-CDMA" in an indoor wireless environment. MC-CDMA may be a suitable modulation technique in the indoor environment where the dispersive character of indoor propagation [1] allows for the exploitation of this technique.

With MC-CDMA, each data symbol is transmitted over N narrowband subcarriers where each subcarrier is encoded with a 0 or π phase offset. If the number of and spacing between subcarriers is appropriately chosen, it is unlikely that all of the subcarriers will be located in a deep fade and consequently frequency diversity is achieved. As an MC-CDMA signal is composed of N narrowband subcarrier signals [2] each with a symbol duration, T_b, much larger than the delay spread, T_d, an MC-CDMA signal will not experience significant intersymbol interference (ISI). Multiple access is achieved with different users transmitting at the same set of subcarriers but with spreading codes that are orthogonal to the codes of other users.

Basic Principles of MC-CDMA

The generation of an MC-CDMA signal can be described as follows. As shown in Fig. 1, a single data symbol is replicated into N parallel copies. Each branch of the parallel stream is multiplied by one chip of a spreading code of length N and then binary phase-shift keying (BPSK) modulated to a subcarrier spaced apart from its neighboring subcarriers by F/T_b *Hz* where F is an integer number. The transmitted signal consists of the sum of the outputs of these branches.

For $F = 1$, this scheme is similar to performing Orthogonal Frequency Division Multiplexing (OFDM) [3] on a Direct-Sequence spread-spectrum signal [4]. Recently, there has been a growing interest on idea of combining OFDM and DS-CDMA [5] [6] [7] [8] [9]. Modern DSP methods make the implementation of MC-CDMA feasible and attractive. With $F = 1$, the transmit bandwidth is minimized. However, larger values of F may be desired to further increase the transmit bandwidth, i.e., to achieve a larger frequency diversity gain, without increasing the complexity in signal processing due to large spreading factors, N.

The transmitted signal corresponding to the *kth* data bit of the *mth* user ($a_m[k]$) is

$$s_m(t) = \sum_{i=0}^{N-1} c_m[i]\, a_m[k] \cos\left(2\pi f_c t + 2\pi i \frac{F}{T_b} t\right) p_{T_b}(t - kT_b) \qquad (1)$$

$$c_m[i] \in \{-1, 1\}$$

where $c_m[0], c_m[1], \ldots, c_m[N-1]$ represents the spreading code of the *mth* user and $p_{T_b}(t)$ is defined to be an unit amplitude pulse that is non-zero in the interval of $[0, T_b]$. The input data symbols, $a_m[k]$, are assumed to takes on values of -1 and 1 with equal probability.

Channel Model: Dispersive Rayleigh Fading

In this paper, we address a frequency-selective channel with $1/T_b \ll BW_c \ll F/T_b$ where BW_c is the coherence bandwidth. This model implies that each modulated subcarrier with transmission bandwidth of $1/T_b$ does not experience significant dispersion ($T_b \gg T_d$). As Doppler shifts are very small and typically in the range of 0.3-6.1 Hz [10] in the indoor environment, it is also assumed that the amplitude and phase remain constant over the symbol duration, T_b.

- **Uplink**

For uplink transmissions, i.e., from terminals to the base station, the base station receives each signal from different users through different channels depending on the location of the terminal. The transfer function of the continuous-time fading channel assumed for the *mth* user in the indoor environment can be represented as

$$H_m\left[f_c + i\frac{F}{T_b}\right] = \rho_{m,i}e^{j\theta_{m,i}} \tag{2}$$

where $\rho_{m,i}$ and $\theta_{i,m}$ are the random amplitude and phase of the channel of the *mth* user at frequency $f_c+i(F/T_b)$. Corresponding to the case in which the direct line-of-sight (LOS) path from the transmitter to receiver is obstructed, the random amplitudes, $\rho_{m,i}$, are assumed to be independent and identically distributed (IID) Rayleigh random variables (r.v.s) [11] for all users and subcarriers. The absence of a line-of-sight path corresponds to a worst-case propagation channel. The assumption of independent fading at the subcarriers is appropriate for channels where $F/T_b \gg BW_c$.

The random phases, $\theta_{m,i}$, are assumed to be IID uniform random variables on the interval $[0, 2\pi]$ for all users and subcarriers.

- **Downlink**

For downlink transmissions, i.e., from the base station to the terminals, a terminal receives interfering signals designated for other users ($m = 1, 2, \ldots, M-1$) through the same channel as the wanted signal ($m = 0$). Thus, the user index of the parameters characterizing the channel may be repressed as follows

$$\rho_{m,i} = \rho_{0,i} \qquad \theta_{m,i} = \theta_{0,i} \tag{3}$$

for $m = 0, 1, \ldots, M-1$.

Receiver Model

For M active transmitters, the received signal is

$$r(t) = \sum_{m=0}^{M-1}\sum_{i=0}^{N-1}\rho_{m,i}c_m[i]\,a_m[k]\cos\left(2\pi f_c t + 2\pi i\frac{F}{T_b}t + \theta_{m,i}\right) + n(t) \tag{4}$$

where n(t) is additive white Gaussian noise (AWGN). The local-mean power at the *ith* subcarrier of the *mth* user is defined to be $\overline{p_{m,i}} = E\rho_{m,i}^2/2$. Assuming the local-mean powers of the subcarriers are equal, the total local-mean power of the *mth* user is defined to be $\overline{p_m} = N\overline{p_{m,i}}$. To simplify the analysis, it is assumed that exact synchronization with the desired user ($m = 0$) is possible. In addition, it is assumed that the system operates synchronously with each user having the same clock. As shown in Fig. 2, demodulating each subcarrier includes applying a phase correction, $\hat{\theta}_{0,i}$, and multiplying the *ith* subcarrier signal by a gain correction, $d_{0,i}$. In the analysis, it is assumed that perfect phase correction can be obtained, i.e., $\hat{\theta}_{0,i} = \theta_{0,i}$. For the *kth* bit, the decision variable is

$$\upsilon_o = \sum_{m=0}^{M-1}\sum_{i=0}^{N-1}\rho_{m,i}c_m[i]\,d_{0,i}a_m[k]\frac{2}{T_b}\int_{kT_b}^{(k+1)T_b}\cos\left(2\pi f_c t + 2\pi i\frac{F}{T_b}t + \theta_{m,i}\right)\cos\left(2\pi f_c t + 2\pi i\frac{F}{T_b}t + \theta_{0,i}\right)dt + \eta \tag{5}$$

where η is the corresponding AWGN term.

In this paper, we will consider two standard diversity reception techniques: Equal Gain Combining (EGC) and Maximum Ratio Combining (MRC).

- Equal Gain Combining

With Equal Gain Combining, the gain factor at the *ith* subcarrier is given as

$$d_{o,i} = c_0[i]\,. \tag{6}$$

This method yields the following decision variable

$$\upsilon_o = a_0[k]\sum_{i=0}^{N-1}\rho_{0,i} + \beta_{int} + \eta \tag{7}$$

where the interference term, β_{int}, is given as

$$\beta_{int} = \sum_{m=1}^{M-1}\sum_{i=0}^{N-1} a_m[k]\,c_m[i]\,c_0[i]\,\rho_{m,i}\cos\tilde{\theta}_{m,i}\,. \tag{8}$$

with $\tilde{\theta}_{m,i} = \theta_{0,i} - \theta_{m,i}$. As the in-phase component, $\rho_{m,i}\cos\tilde{\theta}_{m,i}$, is a zero-mean Gaussian r.v., the interference term, β_{int}, has a zero-mean Gaussian distribution. The variance of β_{int} and the noise component are

$$\sigma^2_{\beta_{int}} = (M-1)\overline{p_m} \qquad \sigma^2_{\eta} = N\frac{N_0}{T_b} \tag{9}$$

respectively.

- Maximum Ratio Combining

For MRC, the gain factor at the *ith* subcarrier is

$$d_{o,i} = \rho_{0,i}c_0[i]\,. \tag{10}$$

The corresponding decision variable for MRC is almost the same as Eq.(7) with $\rho_{0,i}$ replaced by $\rho^2_{0,i}$ in the desired signal component and $\rho_{m,i}$ with $\rho_{m,i}\rho_{0,i}$ in the interference component.

For large *N*, the interference term, β_{int}, is approximately a zero-mean Gaussian r.v. The variance of β_{int} and the noise component are

$$\sigma^2_{\beta_{int}} = 2\frac{(M-1)}{N}\overline{p_m}\,\overline{p_0} \qquad \sigma^2_{\eta} = 2\frac{N_0}{T_b}\overline{p_0} \tag{11}$$

respectively.

Uplink Bit Error Rate (BER)

- EGC

For equal gain combining, the corresponding probability of error is

$$Pr\left(error \middle| \{\rho_{0,i}\}_{i=0}^{N-1}, (M-1)\overline{p_m}\right) = \frac{1}{2}erfc\left(\sqrt{\frac{\frac{1}{2}\left(\sum_{i=0}^{N-1}\rho_{0,i}\right)^2 T_b}{(M-1)\overline{p_m}T_b + NN_0}}\right). \tag{12}$$

In this paper, three approximations for the distribution of

$$\rho_0 = \sum_{i=0}^{N-1} \rho_{0,i} \tag{13}$$

will be considered.

1. In the limiting case of large N, $\Sigma\rho_{0,i}$ can be approximated by the Law of Large Numbers (LLN) to be $NE\rho_{0,i}$. Using the LLN yields the following average BER expression

$$BER = \frac{1}{2}erfc\left(\sqrt{\frac{\pi}{4}\frac{\overline{p_0}T_b}{\frac{(M-1)}{N}\overline{p_m}T_b + N_0}}\right). \tag{14}$$

2. For the small values, the distribution of ρ_0 may be approximated by

$$f_{\rho_0}(\rho_0) = \frac{\rho_0^{2N-1}e^{-\frac{\rho_0^2}{2b}}}{2^{N-1}b^N(N-1)!}. \tag{15}$$

where $b = \frac{\overline{p_0}}{N}[(2N-1)!!]^{1/N}$. The average BER can be obtained by averaging Eq.(12) over Eq.(15).

3. A third possible approximation can be obtained by applying the Central Limit Theorem (CLT) for the limiting case of large N to yield an average BER of

$$BER \cong \frac{1}{2}erfc\left(\sqrt{\frac{\pi}{4}\frac{\overline{p_0}T_b}{(2-\frac{\pi}{2})\frac{\overline{p_0}}{N}T_b + (M-1)\frac{\overline{p_m}}{N}T_b + N_0}}\right) \tag{16}$$

- MRC

For MRC, the average BER can be determined to be

$$BER = \int_0^{\infty} \frac{1}{\overline{p_{0,i}}(N-1)!}\left(\frac{r}{\overline{p_{0,i}}}\right)^{N-1} e^{-\frac{r}{\overline{p_{0,i}}}} \frac{1}{2}erfc\left(\sqrt{\frac{r^2T_b}{\frac{(M-1)}{N}\overline{p_m}\overline{p_0}T_b + \overline{p_0}N_0}}\right)dr. \tag{17}$$

An approximation for the BER, obtained by applying the LLN, is

$$BER \cong \frac{1}{2}erfc\left(\sqrt{\frac{\overline{p_0}T_b}{\frac{(M-1)}{N}\overline{p_m}T_b + N_0}}\right). \tag{18}$$

Downlink Bit Error Rate

Using Eq.(3), the BER in the downlink can be calculated in a similar manner.

- EGC

Applying the orthogonality of the codes to Eq.(8), the interference has a variance of $\sigma^2_{\beta_{int}} = 2(M-1)\left[1-\frac{\pi}{4}\right]\bar{p}_0$. The corresponding BER is

$$BER \cong \frac{1}{2}erfc\left(\sqrt{\frac{\pi}{4}\frac{\overline{p_0}T_b}{2\frac{(M-1)}{N}\left[1-\frac{\pi}{4}\right]\overline{p_0}T_b + N_0}}\right). \tag{19}$$

- MRC

For MRC in the downlink, the BER can be approximated by

$$BER \cong \frac{1}{2} erfc\left(\sqrt{\frac{\overline{p_0}T_b}{2\frac{(M-1)}{N}\overline{p_0}T_b + N_0}}\right) \tag{20}$$

where the variance of the interference is

$$\sigma^2_{\beta_{int}} = (M-1)\,N\,[E\rho^4_{0,i} - (E\rho^2_{0,i})^2]\,. \tag{21}$$

Numerical Results

- **Uplink**

Plots of the bit error rates versus the number of co-channel interferers are given in Fig. 3 for a signal-to-noise ratio ($SNR = \overline{p_0}T_b/N_0$) of 10dB. To calculate the BERs, it was assumed that each interfering signal has a local-mean power equal to the local-mean power of the wanted signal.

As it can be seen in Fig. 3, the approximations for EGC produce relatively close curves. According to all approximation, MRC outperforms EGC for any number of interferers. Comparing Eq.(14) and Eq.(18), the improvement in performance between MRC and EGC is $-1.05dB$.

- **Downlink**

Plots of the BER for downlink transmissions are given in Fig. 4. For both diversity methods, the bit error rates were approximated with the LLN. Examining the curves in Fig. 4, it can be seen that for a small number of users (i.e., in a noise limited channel, $M = 0$) MRC outperforms EGC. However, for a large number of users, EGC has a superior performance. This reflects the observation that MRC distorts the orthogonality between users.

This result suggests that the performance could be enhanced further by restoring the orthogonality of the interfering signals. In particular, an equalization scheme which makes the subcarrier amplitudes identical irrespective of any channel fading, i.e., $d_i = 1/\rho_i$, would entirely remove all interference if orthogonal codes are used. However, a fundamental problem is the fact that in a Rayleigh-fading channel $\mathrm{E}(1/\rho_i^2) \to \infty$.

Hence, a perfect equalization scheme would excessively amplify the noise power on some subcarriers. Our continued research indicates that a conditional equalization scheme, i.e., $d_i = (1/\rho_i)\,u(\rho_i - \rho_{TH})$, that disregards subcarriers below a threshold, ρ_{TH}, results in substantially lower BERs than the results in Fig. 4 for MRC and EGC. Moreover, simulations presented at PIMRC’94 [6] suggest that multi-signal detection schemes can further enhance the performance of MC-CDMA.

Conclusion

In this paper, a new spread spectrum technique was introduced and its bit error rate for a Rayleigh fading dispersive channel was analyzed. This method combines the advantages of bandspreading, code division and frequency diversity of Direct-Sequence CDMA with the advantages of Multi-Carrier Modulation. Depending of the choice of the *F*-parameter, the transmit spectrum can either be minimized with the strong attenuation of adjacent channel sidelobes which is typical in OFDM [3] or intentionally widened to achieve a larger frequency-diversity performance gain without excessively increasing the spreading factor. For wireless office communication systems operating in deregulated ISM bands, the latter is very desirable since these systems are often designed with minimum signal processing complexity in order to reduce terminal power consumption.

For the two diversity techniques considered in our analysis, MRC performed better than EGC in the uplink it but appeared less effective in combating interference in the downlink. MRC distorts the orthogonality of the codes and consequently performed worse for a large number of users in the downlink. Comparing the performance of EGC in the uplink to the downlink at a bit error rate of 10^{-3}, there was an increase in capacity from 8 users to 20 users in the downlink. This improvement was due to the greater degree of phase control in the downlink that allowed for some of the benefits from the orthogonality of the codes to be utilized.

This paper presented early results on this novel modulation, diversity and multiple access technique. It did not address implementation and complexity aspects. Parallel implementation of a multiple branch receiver (as in Fig. 2) may lead to an excessively complex receiver architecture. It should however be realized that MC-CDMA is very similar to a transmission technique where an FFT operation is performed on Direct-Sequence CDMA signals. Standard FFT circuits may simplify the implementation. Moreover, Multi-Carrier Modulation (MCM) or Orthogonal Frequency Division Multiplexing (OFDM), as applied in our concept, is known to be an efficient method for performing channel equalization. MC-CDMA modem design will be discussed in detail in a paper by Fettweis et al. [7].

A direct comparison between DS-CDMA and MC-CDMA requires careful choice of the channel model and its effects on the detection process. Presumably, the assumption of a highly selective channel with multiple resolvable paths will favor MC-CDMA. A fair comparison should further be made for receivers with comparable complexity. A DS-CDMA system utilizing a RAKE receiver should be compared with a MC-CDMA system using Weiner filtering on all of the subcarrier signals.

Equalization can be performed in the uplink if the transmitter predistorts the signals, i.e., if it anticipates the channel transfer function. This requires feedback information from the base station.

The performance of MC-CDMA may be affected by nonlinearities in the transmitter. The joint downlink signal is not necessarily a constant-envelope signal as in BPSK or QPSK modulation. As in OFDM transmissions, MC-CDMA may suffer from this nonlinearity if the number of subcarriers is too large.

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Biographies

Jean-Paul M.G. Linnartz

Jean-Paul M.G. Linnartz was born in Heerlen, The Netherlands, in 1961. He attended Gymnasium B at the Scholengemeenschap St. Michiel in Geleen. He received his Ir. (M.Sc. E.E.) degree in Electrical Engineering Cum Laude from Eindhoven University of Technology, The Netherlands, in 1986. During 1987-1988, he was with The Netherlands Organization for Applied Scientific Research, Physics and Electronics Laboratory F.E.L-T.N.O., The Hague, where he worked UHF propagation and frequency assignment techniques. From 1988-1991 he was Assistant Professor at Delft University of Technology, where he received his Ph.D. (Cum Laude) on multi-user mobile radio nets in December 1991. In June 1992, he received the Dutch Veder Prize for his research on traffic aspects in mobile radio networks. Since January 1992, he has been an assistant professor in the Department of EECS at the University of California at Berkeley. In February 1993 he published his first book, entitled “Narrowband Land-Mobile Radio Networks”.

Nathan Yee

Nathan Yee was born in Honolulu, Hawaii, in 1970. He received his undergraduate B.S. degree from the University of California at Santa Barbara in 1991. In 1993, he received his M.S. degree from the University of California at Berkeley.

Gerhard Fettweis

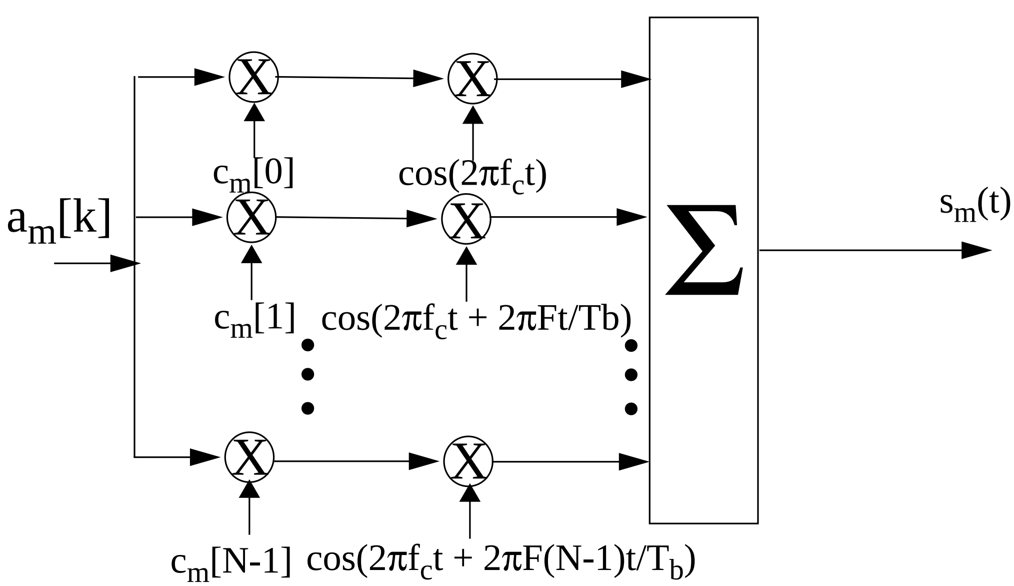

Fig. 1 Transmitter Model

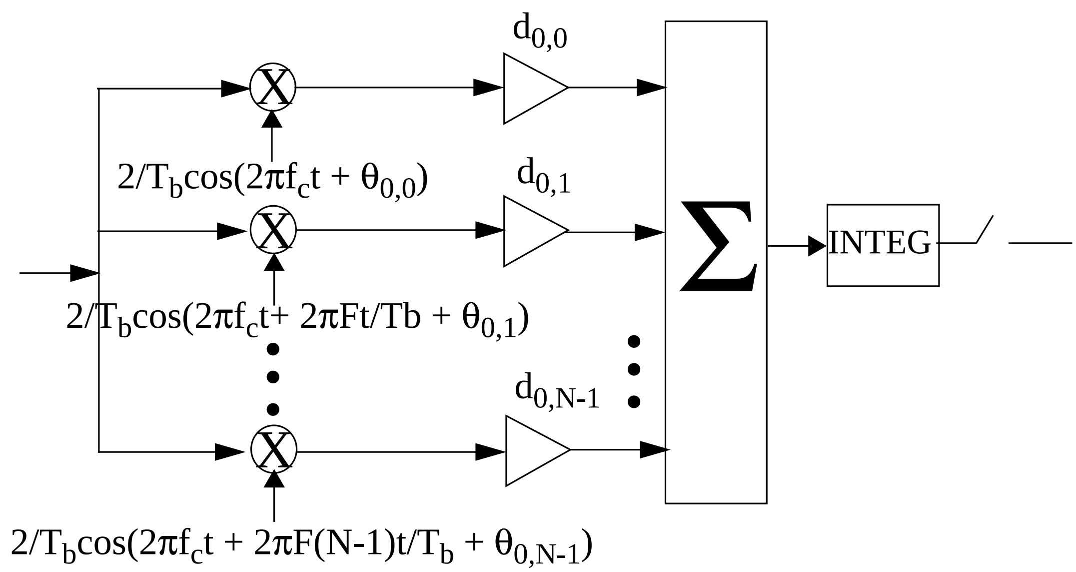

Fig. 2 Receiver Model

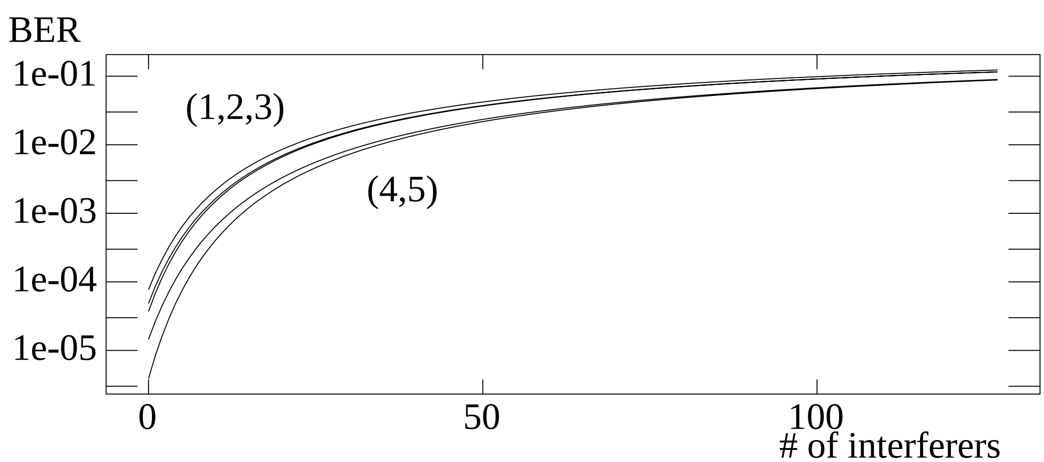

Fig. 3 BER for EGC using the small argument approx. (1), CLT (2), and LLN (3) and for MRC exact (4) and approx. using LLN (5) in the uplink versus the number of interferers m = 0, 1, ... , 127. The SNR is 10dB and N = 128.

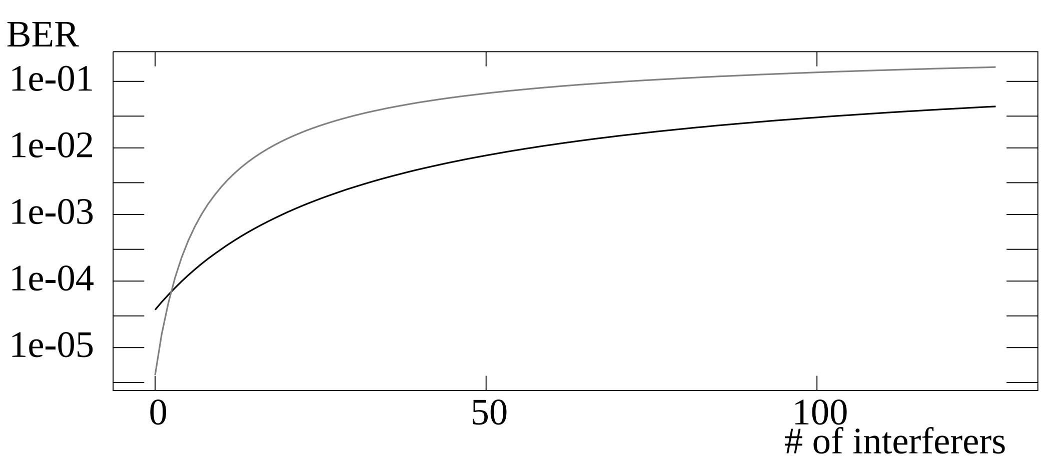

Fig. 4 BER for EGC(——) and MRC(----) in the downlink versus the number of interferers, m = 0, 1, ... , 127. The SNR is 10 dB and N = 128.