The Effect of Synchronisation Errors on MC-CDMA Performance

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ABSTRACT

In this contribution, we investigate the sensitivity of MC-CDMA systems to synchronisation errors. We show that the signal-to-noise ratio (SNR) of the MC-CDMA system at the input of the decision device in the presence of a carrier frequency offset or a clock frequency offset is a rapidly decreasing function of the number of carriers. For a maximal load, carrier phase jitter and timing jitter give rise to a degradation that is independent of the spectral content of the jitter; moreover, the degradation caused by carrier phase jitter is independent of the number of carriers. A constant carrier phase offset and a constant timing offset cause no degradation of the MC-CDMA system performance.

1. INTRODUCTION

Recently, different combinations of orthogonal frequency division multiplexing (OFDM) and code division multiple access (CDMA) have been investigated in the context of high data rate communication over dispersive channels [1-8]. One of these systems is multicarrier CDMA (MC-CDMA), which has been proposed for downlink communication in mobile radio. In MC-CDMA the data symbols are multiplied with a higher rate chip sequence and then modulated on orthogonal carriers.

The use of a large number of carriers makes a multicarrier system very sensitive to some types of synchronisation errors [9-13]. Synchronisation errors can be classified in two classes : carriers phase errors and timing errors. The influence of carrier phase errors between the carrier oscillators at transmitter and receiver, has been investigated in [9-10]. When using a free-running local oscillator, exhibiting a frequency offset and Wiener phase noise, simulations in [9] show that MC-CDMA performance rapidly degrades for an increasing number of carriers. In [10], a phase-locked local oscillator was used to get rid of a carrier frequency offset and of phase noise components that fall within the loop bandwidth of the PLL. For a maximal load, the MC-CDMA performance degradation caused by the resulting phase jitter was shown to be independent of the number of carriers and of the spectral content of the jitter.

The influence of sampling time errors on the performance of OFDM has been investigated in [11-13]. In [11-12], it was shown that for a large number of carriers, the OFDM system severely suffers from a clock frequency offset between the transmitter clock and the sampling clock of the receiver. In [12-13], it was proposed to use a phase-locked loop for the timing recovery, to eliminate the clock frequency offset. The timing jitter resulting from the PLL exhibits a performance degradation independent of the number of carriers.

In this contribution, we investigate the sensitivity of the MC-CDMA system to synchronisation errors by incorporating the carrier phase errors and timing errors in the end-to-end transfer function, obtaining a *time-varying* equivalent filter. This unified approach allows us to present a single analysis, valid for both carrier phase errors and timing errors. In section 2, we present the description of the MC-CDMA system in the presence of synchronisation errors. In section 3, we consider the sensitivity of MC-CDMA to a constant carrier phase offset, a carrier frequency offset and carrier phase jitter. In section 4, we focus on the influence of timing errors on the MC-CDMA performance, more specifically a constant timing offset, a clock frequency offset and timing jitter. Conclusions are drawn in section 5.

2. SYSTEM DESCRIPTION

The conceptual block diagram of a downlink MC-CDMA system is shown in Fig. 1. The data symbols $\{a_{i,m}\}$, transmitted to the user m during the ith symbol interval, are multiplied with the corresponding CDMA chip sequences $\{c_{n,m}|n=0,...,N-1\}$, $c_{n,m}$ denoting the nth chip of the sequence belonging to the user m, resulting in the samples $\{b_{i,n}|n=0,...,N-1\}$:

$$b_{i,n} = \sum_m \sqrt{E_{sm}}\, a_{i,m} c_{n,m} \quad (1)$$

where E_{sm} is the mth user's energy per symbol. Sequences belonging to different users are assumed to be orthogonal (e.g. Walsh-Hadamard codes). The sample $b_{i,n}$ is modulated on the nth carrier of a set of N orthogonal carriers, and all modulated carriers are summed to obtain the transmitted time-domain samples $s_{i,k}$:

$$s_{i,k} = \frac{1}{N}\sum_{n=0}^{N-1} b_{i,n} e^{j2\pi\frac{kn}{N}} \quad (2)$$

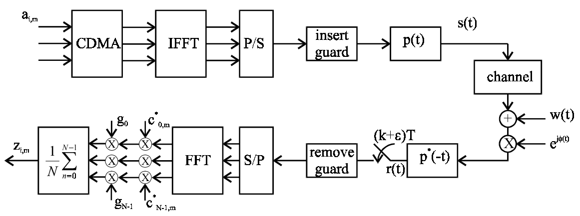

Figure 1 : MC-CDMA transceiver

As in conventional OFDM, the intersymbol interference can be avoided by cyclically extending the transmitted signal with a guard interval νT. The samples $\{s_{i,k}|k=-\nu,...,N-1\}$ are applied to a transmit filter with transfer function $P(f)$, yielding the transmitted signal $s(t)$. The transmit filter is a unit energy square root Nyquist filter.

The transmitted signal is applied to the dispersive channel with channel transfer function $H_{ch}(f)$ and is disturbed by additive white Gaussian noise $w(t)$, with uncorrelated real and imaginary parts, each having a power spectral density of $N_0/2$. At the receiver, the signal disturbed by a carrier phase error

$\phi(t)$ is applied to the receiver filter, which is matched to the transmit filter. The resulting signal $r(t)$ is sampled at the instants $t_{i,k}=i(N+\nu)T+kT+\varepsilon_{i,k}T$, $\varepsilon_{i,k}$ denoting the normalised timing error of the kth sample in the ith symbol interval ($|\varepsilon_{i,k}|<1/2$). When the carrier phase error is slowly varying as compared to T, the samples $r_{i,k}$ can be written as :

$$r_{i,k} = \sum_m h_{eq}\left(kT - mT; t_{i,k}\right) s_{i,m} \tag{3}$$

where $h_{eq}(t;t_{i,k})$ is the *time-varying* impulse response; its Fourier transform (with respect to the variable t) $H_{eq}(f;t_{i,k})$ is given by :

$$H_{eq}\left(f; t_{i,k}\right) = e^{j\phi_{i,k}} H(f) e^{j2\pi f \varepsilon_{i,k} T} \tag{4}$$

where $H(f)$ consists of the cascade of the transmit filter, the channel transfer function and the receiver filter, i.e. $H(f)=|P(f)|^2 H_{ch}(f)$, and $\phi_{i,k}$ is the carrier phase error at the instant $t_{i,k}$. Assuming the duration of $h_{eq}(t;t_{i,k})$ does not exceed the duration of the guard interval, the transients at the edges of a symbol will affect the signal only during the guard interval, such that the adjacent symbols will have no influence on the observed symbol. It can be verified that the samples $r_{i,k}$ *outside* the guard interval are given by :

$$r_{i,k} = \frac{1}{N} \sum_{n=0}^{N-1} b_{i,n} H_n\left(t_{i,k}\right) e^{j2\pi \frac{nk}{N}} + w_{i,k} \quad k = 0, \ldots, N-1 \tag{5}$$

where

$$H_n\left(t_{i,k}\right) = \frac{1}{T} \sum_{m=-\infty}^{+\infty} H_{eq}\left(\frac{n}{NT} + \frac{m}{T}; t_{i,k}\right) \tag{6}$$

The receiver selects the N samples $r_{i,k}$ outside the guard interval for further processing and disregards the other ν samples. The selected samples are demodulated using a fast Fourier transform (FFT); the nth output of the FFT is multiplied with the chip $c_{n,m}$ of the considered user m and applied to a one-tap equaliser with coefficient $g_{n,i}$, the filter coefficient belonging to the nth FFT output during the ith symbol interval. The resulting samples are summed yielding the samples at the input of the decision device

$$\begin{aligned} z_{i,m} &= \frac{1}{N} \sum_{n=0}^{N-1} g_{n,i} c_{n,m}^{*} \sum_{k=0}^{N-1} r_{i,k} e^{-j2\pi \frac{kn}{N}} \\ &= \sqrt{E_{sm}}\, a_{i,m} I_{i,m,m} + \sum_{\ell \neq m} \sqrt{E_{s\ell}}\, a_{i,\ell} I_{i,\ell,m} + W_{i,m} \end{aligned} \tag{7}$$

where $W_{i,m}$ is a zero-mean complex-valued Gaussian noise term and

$$I_{i,\ell,m} = \frac{1}{N^2} \sum_{k=0}^{N-1} \sum_{n,n'=0}^{N-1} g_{n',i} c_{n,\ell} c_{n',m}^{*} e^{j2\pi \frac{k(n-n')}{N}} H_n\left(t_{i,k}\right) \tag{8}$$

The quantity $I_{i,\ell,m}$ denotes the multi-user interference (MUI) at the correlator output of the mth user, originating from the ℓth user during the ith symbol interval. The equaliser attempts to compensate the attenuation and phase shift of the useful component, caused by the dispersive nature of the channel and the presence of synchronisation errors, by scaling and rotating the outputs of the FFT.

In the absence of synchronisation errors, the equivalent filter $H_{eq}(f;t_{i,k})$ reduces to

$$H_{eq}\left(f; t_{i,k}\right) = H(f) \tag{9}$$

and the quantities $I_{i,\ell,m}$ yield

$$I_{i,\ell,m} = \frac{1}{N} \sum_{n=0}^{N-1} c_{n,\ell} g_{n,i} H_n c_{n,m}^{*} \tag{10}$$

where H_n is related to $H(f)$ through (6). Assuming the receiver is able to accurately estimate H_n, a zero-forcing equaliser multiplies the outputs of the FFT with $g_{n,i}=1/H_n$; this reduces the quantities $I_{i,\ell,m}$ to $I_{i,\ell,m}=\delta_{\ell,m}$, signifying that the equaliser succeeds in compensating for the attenuation and the phase shift of the useful signal component as well as in eliminating the multi-user interference. The scaling of the FFT outputs however influences the additive noise and results in an increase of the noise power level, thus a degradation as compared to the case of an ideal channel.

In the case of an ideal channel ($H_{ch}(f)=1$) and in the absence of synchronisation errors, the equivalent filter yields

$$H_{eq}\left(f; t_{i,k}\right) = \left|P(f)\right|^2 \tag{11}$$

As the ideal channel yields $H_n=1$, no scaling or rotation of the FFT outputs by the equaliser are necessary (e.g. $g_{n,i}=1$); the channel introduces no attenuation of the useful component nor any multi-user interference.

When the carrier phase error or the timing error can be modelled as random processes, the quantities $I_{i,\ell,m}$ are random variables. The useful component can be decomposed as

$$I_{i,m,m} = E\left[I_{i,m,m}\right] + \left(I_{i,m,m} - E\left[I_{i,m,m}\right]\right) \tag{12}$$

where the first contribution denotes the average useful component and the second term the fluctuation of the useful component. Defining the signal-to-noise ratio (SNR) at the input of the decision device as the ratio of the power of the average useful component to the power of the remaining components, the SNR at the correlator output of the mth user is given by

$$\begin{aligned} SNR_m = E_{sm} \left| E\left[I_{i,m,m}\right]\right|^2 \Big\{ E\left[\left| W_{i,m}\right|^2\right] \\ + E_{sm} E\left[\left| I_{i,m,m} - E\left[I_{i,m,m}\right]\right|^2\right] + \sum_{\ell \neq m} E_{s\ell} E\left[\left| I_{i,\ell,m}\right|^2\right] \Big\}^{-1} \end{aligned} \tag{13}$$

In the case of an ideal channel and in the absence of synchronisation errors, (13) yields $SNR_m=E_{sm}/N_0$. Hence, the degradation caused by a non-ideal channel and/or in the presence of synchronisation errors is defined as $D_m=10\log((E_{sm}/N_0)/SNR_m)$ (dB).

3. Carrier Phase Errors

In this section, we investigate the sensitivity of the MC-CDMA system to carrier phase errors in the absence of timing errors. In downstream communication, all transmitted carriers exhibit the same carrier phase error, as they are up-converted by the same oscillator. In this case (6) yields

$$H_n\left(t_{i,k}\right) = e^{j\phi_{i,k}} H_n \tag{14}$$

In the following, we separately consider a constant carrier phase offset, a carrier frequency offset and carrier phase jitter.

3.1 Constant Carrier Phase Offset

In the case of a constant carrier phase offset $\phi(t)=\phi$, (14) reduces to

$$H_n\left(t_{i,k}\right)=e^{j\phi}H_n \quad (15)$$

It can be verified that the outputs of the FFT are rotated over an angle ϕ as compared to the case of a zero carrier phase offset. As a rotation of the FFT outputs has no influence on the noise power, a constant phase offset can be compensated without loss of performance, by simply rotating the outputs of the FFT over (an estimate of) $-\phi$, i.e. $g_{n,i}=e^{-j\phi}$.

3.2 Carrier Frequency Offset

A frequency offset ΔF between the carrier oscillators at the transmitter and the receiver yields a carrier phase error $\phi(t)=2\pi\Delta Ft$. Assuming a slowly varying phase error as compared to T, i.e. $\Delta FT<<1$, the quantities $I_{i,\ell,m}$ are given by

$$I_{i,\ell,m}=\frac{1}{N}\sum_{n,n'=0}^{N-1}c_{n,\ell}H_nG\left(\frac{n-n'}{N}+\Delta FT\right)e^{j2\pi\Delta FTi(N+\nu)}g_{n',i}c^*_{n',m} \quad (16)$$

where

$$G(x)=\frac{1}{N}\sum_{k=0}^{N-1}e^{j2\pi kx} \quad (17)$$

In the following, we consider the case of an ideal channel, so the degradation only originates from the carrier frequency offset. Assuming the receiver can estimate the carrier frequency offset ΔF, the equaliser attempts to eliminate the attenuation and rotation of the useful component by multiplying the FFT outputs with $g_{n,i}=exp(-j2\pi\Delta FTi(N+\nu))/G(\Delta FT)$. Assuming all users have the same energy per symbol E_s and the number of users is M, it follows from (13) that all users exhibit the same SNR and degradation

$$D_m=-10\log\left|\frac{\sin\pi N\Delta FT}{N\sin\pi\Delta FT}\right|^2 +10\log\left(1+\frac{E_s}{N_0}\frac{M-1}{N-1}\left(1-\left|\frac{\sin\pi N\Delta FT}{N\sin\pi\Delta FT}\right|^2\right)\right) \quad (18)$$

In Fig. 2, the degradation (18) is shown for the maximum load, i.e. $M=N$. The degradation shown in Fig. 2 yields an upper bound for the degradation for $M<N$ users. We observe a high sensitivity of the MC-CDMA system to the carrier frequency offset. In order to obtain small degradations, only small frequency offsets are allowed, i.e. $\Delta F<<1/NT$.

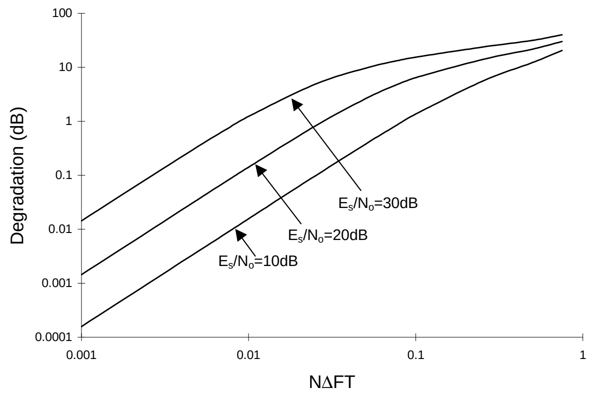

Figure 2 : Carrier frequency offset, M=N

3.3 Carrier Phase Jitter

In order to get rid of the frequency offset, a phase-locked local oscillator can be used for IF to baseband conversion. The phase-locked loop (PLL) also eliminates the phase noise components that fall within the bandwidth of the PLL. The residual phase jitter can be modelled as a zero-mean stationary process with jitter spectrum $S_\phi(f)$ and jitter variance σ_ϕ^2. Assuming slowly varying phase jitter, the bandwidth B of the jitter spectrum $S_\phi(f)$ needs to be $BT<<1$. The quantities $I_{i,\ell,m}$ are stationary processes and, assuming small jitter variances, i.e. $\sigma_\phi^2<<1$, they can be approximated by

$$I_{i,\ell,m}\cong\frac{1}{N}\sum_{n,n'=0}^{N-1}c_{n,\ell}H_n\left(\delta_{n,n'}+\frac{1}{N}\sum_{k=0}^{N-1}j\phi(kT)e^{j2\pi\frac{k(n-n')}{N}}\right)g_{n',i}c^*_{n',m} \quad (19)$$

For small jitter variances, the equaliser is essentially the same as in the absence of phase jitter, so that the FFT outputs are multiplied with $g_{n,i}=1/H_n$.

In the case of an ideal channel ($H_n=1$), it was shown in [10] that the fluctuation of the useful component and the multi-user component mainly consists of the low frequency components ($<1/NT$) and the high frequency components ($>1/NT$) of the jitter spectrum, respectively. For the highest load ($M=N$) and all users exhibiting the same energy per symbol E_s, the sum of the fluctuation of the useful component and the multi-user component is independent of the spectral content of the jitter spectrum and of the number of OFDM tones, and only depends on the jitter variance, given by

$$\sigma_\phi^2=\int_{-\infty}^{+\infty}S_\phi(f)df \quad (20)$$

For N active users, the degradation $D_m=10log(1+(E_{sm}/N_0)\sigma_\phi^2)$, shown in Fig. 3, yields an upper bound for $M<N$ users.

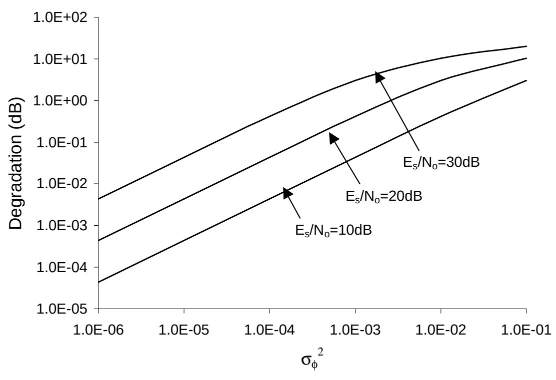

Figure 3 : Carrier phase jitter, M=N

4. TIMING ERRORS

In this section, we investigate the sensitivity of the MC-CDMA system to timing errors. The deviation from the correct sampling instant by an amount $\varepsilon_{i,k}T$ influences the quantities $H_n(t_{i,k})$ as follows

$$H_n\left(t_{i,k}\right)=\frac{1}{T}\sum_{m=-\infty}^{+\infty} H\left(\frac{n}{NT}+\frac{m}{T}\right)e^{j2\pi\varepsilon_{i,k}T\left(\frac{n}{N}+m\right)} \qquad (21)$$

In downstream communication, all carriers exhibit the same timing error as the signals are synchronised at the base station. In the following, we consider a constant timing offset, a clock frequency offset and timing jitter in the case of an ideal channel ($H_{ch}(f)=1$); the transmit and receiver filters are square root raised cosine filters with rolloff α.

4.1 Constant Timing Offset

In the case of a constant timing offset $\varepsilon_{i,k}=\varepsilon$, the quantities $I_{i,\ell,m}$ yield

$$I_{i,\ell,m}=\frac{1}{N}\sum_{n=0}^{N-1} c_{n,\ell} H_n\left(t_{i,k}\right) g_{n,i} c^*_{n,m} \qquad (22)$$

The equivalent filter is bandwidth limited : $H_{eq}(f;0)=0$, $|f|>(1+\alpha)/2T$, $0\leq\alpha\leq 1$. For frequencies n/T outside the rolloff area, the sum (21) reduces to one contribution

$$H_n\left(t_{i,k}\right)=H_n e^{j2\pi\varepsilon\frac{\mathrm{mod}(n;N)}{N}} \qquad \frac{n}{T}\notin\left(\frac{1-\alpha}{2T},\frac{1+\alpha}{2T}\right) \qquad (23)$$

where $mod(x;N)$ denotes the modulo-N reduction of x, yielding a result in ($-N/2,N/2$). For frequencies n/T inside the rolloff area, the sum (21) consists of two contributions for which it can be verified that

$$\left|H_n\left(t_{i,k}\right)\right|<\left|H_n\right| \qquad \frac{n}{T}\in\left(\frac{1-\alpha}{2T},\frac{1+\alpha}{2T}\right) \qquad (24)$$

In Fig. 4, we observe that for frequencies n/T outside the rolloff area, the nth output of the FFT exhibits a constant amplitude $|H_n|=1$ and is rotated over an angle $2\pi\varepsilon mod(n;N)/N$. For frequencies n/T inside the rolloff area, the outputs of the FFT are rotated over some angle and the amplitudes are reduced as compared to $|H_n|$.

The equaliser multiplies the outputs of the FFT with $g_{n,i}=1/H_n(t_{i,k})$, reducing the quantities $I_{i,\ell,m}$ to $I_{i,\ell,m}=\delta_{\ell,m}$. This equalisation signifies a rotation over an (estimate of the) angle $-2\pi\varepsilon mod(n;N)/N$ for FFT outputs outside the rolloff area and a scaling and rotation for FFT outputs inside the rolloff area. The equaliser compensates for the attenuation and rotation of the FFT outputs and avoids the MUI; the noise power is given by

$$E\left[\left|W_{i,k}\right|^2\right]=N_0\frac{1}{N}\sum_{n=0}^{N-1}\frac{1}{\left|H_n\left(t_{i,k}\right)\right|^2}\geq N_0 \qquad (25)$$

which yields an increase of the noise power level as compared to the case of a zero constant timing offset caused by the carriers inside the rolloff area. For $\alpha=0$, a constant timing offset introduces no degradation. For $\alpha\neq 0$, the sensitivity of the MC-CDMA system to constant timing offsets can be eliminated by not using the carriers in the rolloff area.

4.2 Clock Frequency Offset

When sampling is performed by means of a free-running clock with a relative clock frequency offset of $\Delta T/T$, the normalised timing error is given by $\varepsilon_{i,k}=(k+i(N+\nu))\Delta T/T$. Assuming $\Delta T/T<<1$ the quantities $I_{i,\ell,m}$ yield

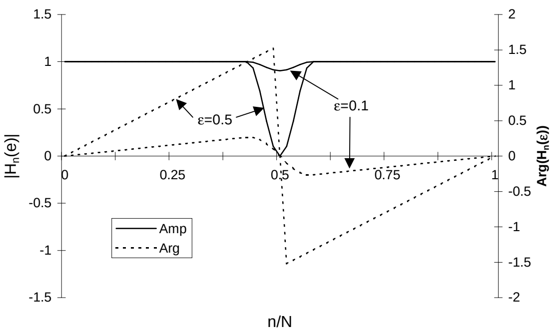

Figure 4 : Dependency of $|H_n(\varepsilon)|$ and $Arg(H_n(\varepsilon))$ on the carrier index n, $\alpha=0.15$

$$I_{i,\ell,m}=\frac{1}{N}\sum_{n,n'=0}^{N-1} c_{n,\ell}\, g_{n',i}\, c^*_{n',m}\,\frac{1}{T}\sum_{k=-\infty}^{+\infty} H\left(\frac{n}{NT}+\frac{k}{T}\right) G\left(\frac{n-n'}{N}+\left(\frac{n}{N}+k\right)\frac{\Delta T}{T}\right)e^{j2\pi\left(\frac{n}{N}+k\right)i(N+\nu)\frac{\Delta T}{T}} \qquad (26)$$

where $G(x)$ is defined in (17). The deterministic character of the clock frequency offset causes no fluctuation of the useful power but only yield an attenuation of the useful power and multi-user interference.

For $\alpha=0$, the quantities $I_{i,\ell,m}$ reduce to

$$I_{i,\ell,m}=\frac{1}{N}\sum_{n,n'=0}^{N-1} c_{n,\ell}\, G\left(\frac{n-n'}{N}+\frac{\mathrm{mod}(n;N)}{N}\frac{\Delta T}{T}\right) e^{j2\pi\frac{\mathrm{mod}(n;N)}{N}i(N+\nu)\frac{\Delta T}{T}} g_{n',i}\, c^*_{n',m} \qquad (27)$$

The equaliser multiplies the outputs of the FFT with $g_{n,i}=exp(-j2\pi(mod(n;N)/N)\ i(N+\nu)\Delta T/T\ /\ G((mod(n;N)/N)\ \Delta T/T))$. Assuming each user exhibits the same energy per symbol E_s, the degradation of the SNR at the input of the decision device is user independent and is shown in Fig. 5. It follows that the MC-CDMA system strongly depends on the clock frequency offset and in order to obtain small degradations, we need a small relative frequency offset $\Delta T/T<<1/N$. For $\alpha\neq 0$ and no carriers in the rolloff area, the results in Fig. 5 yield an upper bound for the degradation. We can avoid the degradation of the MC-CDMA system caused by a clock frequency offset by using a timing correction mechanism in front of the FFT.

4.3 Timing jitter

In order to get rid of a constant timing offset and a clock frequency offset, we can perform synchronised sampling, e.g. by means of a PLL. The normalised timing error $\varepsilon_{i,k}$ resulting from the PLL can be modelled as zero-mean stationary Gaussian noise with spectrum $S_\varepsilon(f)$ and jitter variance σ_ε^2. Assuming slowly varying timing jitter, the bandwidth B of the jitter spectrum is limited, i.e. $BT<<1$. For small variances $\sigma_\varepsilon^2<<1$, we obtain the quantities

$$I_{i,\ell,m}\cong\frac{1}{N}\sum_{n,n'=0}^{N-1} c_{n,\ell}\left(\delta_{nn'}+j2\pi\tilde{H}_n\frac{1}{N}\sum_{k=0}^{N-1}\varepsilon_{i,k}e^{j2\pi\frac{k(n-n')}{N}}\right)g_{n'}c^*_{n',m} \qquad (28)$$

where

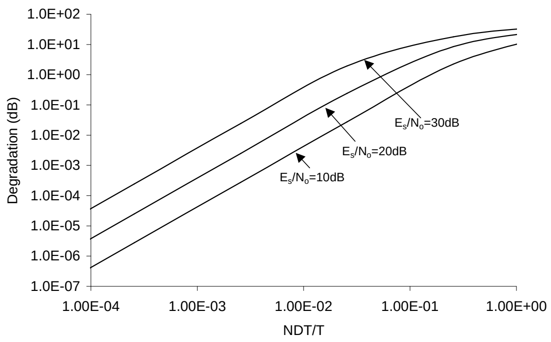

Figure 5 : Clock frequency offset, M=N

$$\tilde{H}_n = \frac{1}{T} \sum_{m=-\infty}^{+\infty} H\left(\frac{n}{NT} + \frac{m}{T} \right)\left(\frac{n}{N} + m \right) \quad (29)$$

from which it follows that for $\ell=m$ the useful component consists of an average useful component and a fluctuation about its average; for $\ell \neq m$, (28) indicates the presence of MUI. For small jitter variances, the equaliser is essentially the same as in the absence of timing jitter, which implies that the equaliser coefficients can be approximated by $g_{n,i}=1$. Assuming all users have the same jitter spectrum $S_\varepsilon(f)$ and energy per symbol E_s, it can be verified in the case of the maximum load $M=N$ that the sum of the useful power and the MUI power is independent of the spectral contents of the jitter but only depends on the jitter variance σ_ε^2

$$\sigma_\varepsilon^2 = \int_{-\infty}^{+\infty} S_\varepsilon(f) df \quad (30)$$

For $\alpha=0$, the maximal load ($M=N$) and for large N ($N \rightarrow \infty$), the degradation is given by

$$D_m = 10 \log\left(1 + \frac{E_s}{N_0} \frac{\pi^2}{3} \sigma_\varepsilon^2 \right) \quad (31)$$

which signifies that for large N the degradation becomes essentially independent of the number of carriers and origins from the MUI only. The degradation of the SNR (31) is shown in Fig. 6.

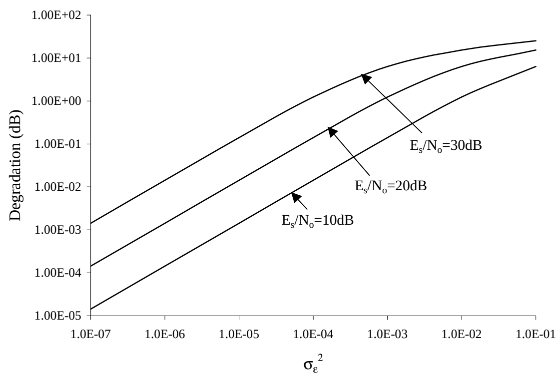

Figure 6 : Timing jitter, M=N

5. CONCLUSIONS

In this contribution, we have investigated the effect of synchronisation errors on the performance of a MC-CDMA system. A constant phase offset and a constant timing offset can be compensated without loss of performance while the MC-CDMA performance degrades for time-varying timing and carrier phase errors. In the case of a carrier frequency offset or a clock frequency offset, the MC-CDMA performance rapidly degrades and strongly depends on the number of carriers. The sensitivity of the MC-CDMA system to a carrier frequency offset or a clock frequency offset can be avoided by using phase-locked loops for carrier and timing recovery, exhibiting a residual carrier phase jitter and timing jitter. For the maximum load, the degradation caused by this carrier phase jitter and timing jitter is independent of the spectral content of the jitter, and is independent of (carrier phase jitter) or only slightly depends on (timing jitter) the number of carriers. Comparing these results with the results of an OFDM system [12], we find that MC-CDMA and OFDM have the same performance for carrier phase errors and essentially the same performance for timing errors.

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