Sensitivity of a MC-CDMA Beyond 3G System to RF Impairments

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ABSTRACT

The MC-CDMA transmission technique is seen as a candidate for beyond 3G systems. It benefits from the high bandwidth efficiency of multi-carrier systems and the flexibility and granularity offered by CDMA for resources allocation.
In the framework of the IST MATRICE European project, a MC-CDMA beyond 3G system has been defined and extensively studied. This paper describes the sensitivity of this system to receiver non-linearity, I&Q mismatch, carrier frequency offset and phase noise. Simulation results show that the higher efficiency of future B3G implies severe constraints on RF front-end.

I. INTRODUCTION

The third generation terrestrial mobile system (UMTS-UTRA) is currently being launched. It aims at offering a large variety of services (circuit and packet services, low to high bit rates), as well as greater capacity compared to second-generation systems (e.g. GSM). The evolution from 2G to 3G represents a change in many aspects: new technology, change of focus from voice to mobile multimedia, simultaneous support of several QoS classes in a single radio interface. Despite the high capacity offered by the 3G technology, the rapid growth of Internet services and increasing interest in portable computing devices are likely to create a strong demand for high-speed wireless data services, presumably with a maximum information bit rate of more than 2-20Mbps in a vehicular environment and possibly 50-100Mbps in indoor to pedestrian environments, using a 50-100MHz bandwidth [1].
It is clear that provision of such high bit rates requires the development of a new technology. One of the most promising candidate techniques for achieving high data rate transmission in a mobile environment is multi-carrier CDMA (MC-CDMA) which divides a wide signal bandwidth into several sub-channels, where several information bearing signals can coexist by using code separation. In the framework of the IST MATRICE European project [3], a MC-CDMA Beyond 3G system have been defined and extensively studied. This paper evaluates the sensitivity of this MC-CDMA system to RF receiver impairments thanks to the modelisation of RF at baseband level. It consists in measuring the degradation introduced by the receiver non-linearity, I&Q mismatch, carrier frequency offset and phase noise.
The paper is organized as follows : in section II, the system specifications are described, in section III, the RF impairments are modeled, and in section IV the simulation results are given and interpreted.

II. SYSTEM DESCRIPTION

The main features of the radio frame of the demonstrator are the followings [4]:

- The radio frame is derived from the UMTS-TDD one. We consider a 10ms frame of 15 slots, where the first slot of each frame is dedicated to DL transmission.
- Sampling frequency F_S = 57,6 MHz, carrier frequency F_0 = 5GHz,
- FFT size : N=1024 (736 useful subcarriers and 288 null subcarriers for spectral shaping), cyclic prefix of 216 samples, spreading factor : 32
- Supported constellation : QPSK and 16-QAM.
- Channel coding : convolutional code with 1/2 and 3/4 coding rate (RCPC convolutional code from IEEE 802.11a standard).

The slot structure is described in Figure 1. It contains a synchronization sequence (S), 4 full OFDM symbols for channel estimation (C), 24 ODFM symbols for data (D) and a guard interval (G).

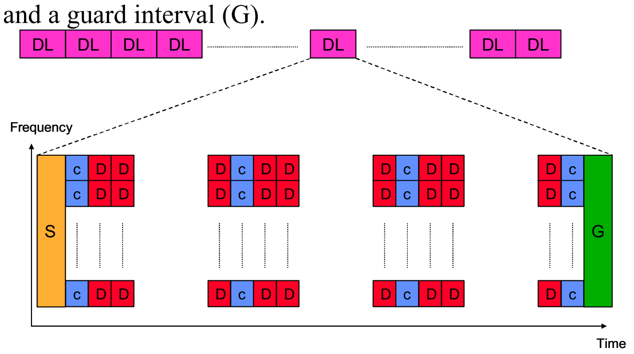

Figure 1 : Frame structure.

Figure 2 gives a functional view of the transmitter and Figure 3 presents the receiver structure [4]. The RF impairments are placed at the front of the receiver.

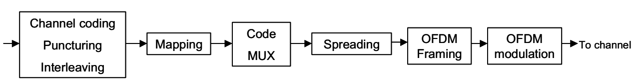

Figure 2: Transmitter.

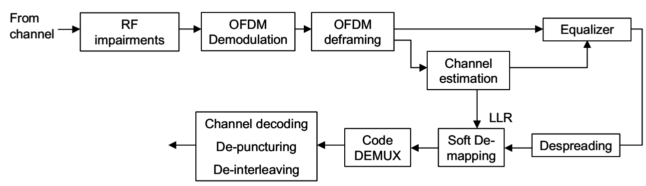

Figure 3: Receiver.

The channel estimation procedure exploits the a priori knowledge of the 4 full pilots transmitted in each slot. At the FFT output, the received signal of a full pilot symbol corresponding to the i^{th} position within the radio frame $i \in S=\{1,10, 19, 28\}$ is :

$$y_i[k] = h_i[k]s[k] + v[k]$$

where s[k] is the pilot symbol transmitted on the k^{th} sub-carrier and $h_i[k]$ is the channel coefficient for the i^{th} pilot position. v[k] is a zero-mean complex gaussian sample with variance σ^2.

If we assume that the coherence bandwidth of the channel spans much more than M+1 subcarriers, $h_i[k]$ is nearly constant over M adjacent subcarriers and a simple averaging will decrease the influence of the noise :

$$\tilde{y}_i[k] = y_i[k].s[k]$$

$$\hat{h}_i[k] = \frac{1}{M+1} \sum_{n=-M/2}^{M/2} \tilde{y}_i[k+n]$$

Then, we need to evaluate the channel coefficient for data ODFM symbols : $\hat{h}_i[k]$ for $i \notin S$.

To do so, channel estimates are interpolated between two consecutive pilot symbols :

$$\hat{h}_{i+j}[k] = \hat{h}_i[k] + \frac{j}{9}\left(\hat{h}_{i+9}[k] - \hat{h}_i[k]\right) \quad j = 1,...,8; i = 1,10,19,28$$

The equalizer is a normalized Minimum Mean Square Equalizer (MMSE) :

$$g_i[kS_F + l] = \frac{\hat{h}_i^*[kS_F + l]}{\left|\hat{h}_i[kS_F + l]\right|^2 + \lambda} \frac{1}{NORM_i[k]}; l = 0,...,S_F - 1; k = 0,...,23$$

$$NORM_i[k] = \sum_{l=1}^{S_F} \frac{\left|\hat{h}_i[kS_F + l]\right|^2}{\left|\hat{h}_i[kS_F + l]\right|^2 + \lambda}$$

where $\hat{h}_i[kS_F + l]$ is the channel estimate of the $(kS_F+l)^{th}$ subcarrier, and λ is the average SNR per subcarrier.

The Log Likelihood Ratios (LLR) used by the soft-demapping to improve the decoder performance are computed according to the following formula :

$$LLR_i[k] = \frac{NORM_i[k]}{S_F} \sum_{l=1}^{S_F} \left|\hat{h}_i[kS_F + l]\right|^2$$

This LLR does not correspond to the optimum LLR formulation, or even the simplified formula proposed by Kaiser [5]. On the other hand, it induces only a 0,5 dB loss for a great complexity improvement.

III. RF IMPAIRMENT MODELS

In this section we will note $V_e(t)=I(t)+jQ(t)$ the input and $V_s(t) = I_1(t)+jQ_1(t)$ the output signals of the RF front-end.

Receiver non-linearity

Front-End RF non linearities are modelled by a polynomial function which coefficients are computed thanks to inter-modulation (IM) product values:

$$V_s(t) = \sum_{k=1}^{M} a_k V_e(t)^k$$

For a 2-tone signal $V_e(t) = A\cos(w1t) + A\cos(w2t)$, IM products appear when developing the 3-order polynomial $a_1 x + a_3 x^3$:

$$V_s(t) = \left(a_1 + \frac{9}{4}a_3 A^2\right) A\cos(w1t) + \left(a_1 + \frac{9}{4}a_3 A^2\right) A\cos(w2t)$$

$$+ \frac{3}{4}a_3 A^3 \cos(2w1 - w2) + \frac{3}{4}a_3 A^3 \cos(2w2 - w1)$$

3-order interception point (IP3) amplitude is computed thanks to polynomial coefficients:

$$A_{IP3} = \sqrt{\frac{4}{3}\left|\frac{a_1}{a_3}\right|}$$

For $a_1 = 1$, we get : $a3 = \frac{4}{3{A_{IP3}}^2}$ and with further developments: $a5 = \frac{8}{5{A_{IP5}}^4}$.

I&Q mismatch

Figure 4 presents the effect of the IQ mismatch on the received signal [6]. Using our notations, the following relation holds :

$$V_s(t) = \alpha V_e(t) + \beta V_e(t)^*$$

with

$$\alpha = \cos(\Delta\Phi) + j\varepsilon \sin(\Delta\Phi)$$

$$\beta = \varepsilon \cos(\Delta\Phi) - j\sin(\Delta\Phi)$$

$\Delta\Phi$ and ε are respectively the phase and gain imbalances.

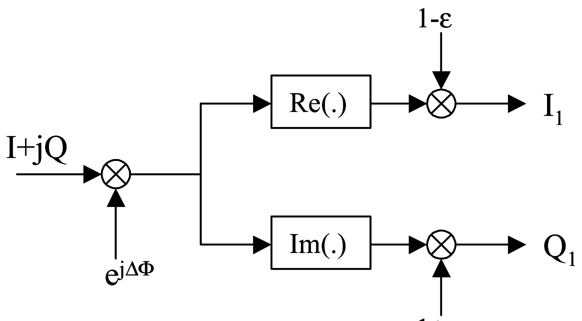

Figure 4 : I&Q mismatch.

After FFT, the signal on the n^{th} sub-carrier is:

$$y_n = \alpha H_n X_n + \beta H^*_{N_c-1-n} X^*_{N_c-1-n} + w_n$$

where X_n is the summation of all the spread data that have been transmitted on the n^{th} subcarrier.

The IQ mismatch creates interference from the mirror frequency. Moreover, this interference is independent from the useful signal. Since the spreading factor (32) is much smaller than the FFT size (1024), the spread data on the

$(N-n)^{th}$ subcarrier are different from ones transmitted on the n^{th} subcarrier. This interference can be taken as an additive noise. If the number of spreading codes is large enough, this interference can also be considered as Gaussian.

If the RF front-end is properly designed, the parameters ε and ΔΦ are small enough so that the small scale approximation is valid. Then parameters α and β can be approximated by:

$$\alpha = 1$$
$$\beta = \varepsilon - j\Delta\Phi$$

The variance of the interference term is thus scaled by $|\beta|^2 = \varepsilon^2 + (\Delta\Phi)^2$. This means that for a fixed $|\beta|^2$, various configurations of ε and ΔΦ should give the same performance.

Carrier frequency offset

The carrier frequency offset is created by different local oscillator frequencies at the transmitter and the receiver. In practice, it is modelled very easily by :

$$V_s(t) = V_e(t)e^{j2\pi\Delta Ft}$$

Phase noise

In multi-carrier systems, sensitivity to phase noise depends on the frequency of this noise. That's the reason why we have developed a phase noise model whose spectra is in full agreement with frequency synthesizer simulations results. Moreover, the parameters used in the phase noise have been chosen in such a way that they can be directly used by RF designers in their frequency synthesizer architecture.

The local oscillator introduced a random phase rotation $\Delta\Phi(t)$ to the received signal when it is down converted :

$$V_s(t) = V_e(t)e^{j\Delta\Phi(t)}$$

The phase noise power spectral density is commonly modelled as a Lorentzian spectrum [11]. This behaviour can be approximated in time domain by a Wiener random process, whose spectrum decreases as a function of the inverse of the square frequency. The variance of the Wiener phase noise process is a linear function of the time: $\sigma_{\Delta\Phi}{}^2(t) = 2Dt$ where D is the diffusion factor of the free running oscillator.

The Wiener phase noise is generated iteratively as follows:

$\Delta\Phi_1(n+1) = \Delta\Phi_1(n) + X(n)$ where X(n) ~ N(0,2D)

Then it is filtered by the Phase Locked Loop, whose filter transfer function is:

$$G(p) = \frac{p^2}{p^2 + 2p.\xi.w_n + w_n{}^2}$$

w_n and ξ are the PLL cut-off pulsation and resonance coefficients. The sensitivity of MC-CDMA to phase noise has already been studied in [7], but assuming a perfect channel estimation. Here, we evaluate the sensitivity of the complete system, including channel estimation, LLR computation, and channel decoding.

III. SIMULATION RESULTS

To perform our simulations, we used a modified BRAN E channel model. It is obtained by interpolation and resampling of the 20 MHz bandwidth original model, to adapt it to a 57,6 MHz channel bandwidth. We also assumed a perfect frame synchronization.

It is well known that RF impairments degrade more severely the performance of high order modulation transmission schemes. Thus we decided to limit our study to the most sensitive configuration which corresponds to the combination of 16-QAM modulation and a coding rate of 3/4.

Receiver non-linearity

The non-linearity of the receiver RF front-end can be characterized by its 3^{rd} and 5^{th} interception points. We have evaluated the sensitivity of the receiver to these two parameters. The IP3 and IP5 levels are given relatively to the received signal power. Figure 5 presents the obtained results. The IP3 and IP5 levels must be respectively 17.5 dB and 20 dB higher than the received signal power. For comparable cellular receivers such as in UMTS TDD mode, the IP3 is imposed by the adjacent channel level. In this particular case, IP3 has only to be 10 dB higher than the received signal power. This mean that the higher efficiency of future Beyond 3G systems imposes more severe specifications on receivers and probably higher power consumption to cope with higher linearity.

Previously, the receiver non-linearly was seldom studied at baseband level since this specification was mainly imposed by out of band signals (aliasing of adjacent channels due to intermodulation). These simulations results show that the deformation of the wanted signal itself is no longer negligible.

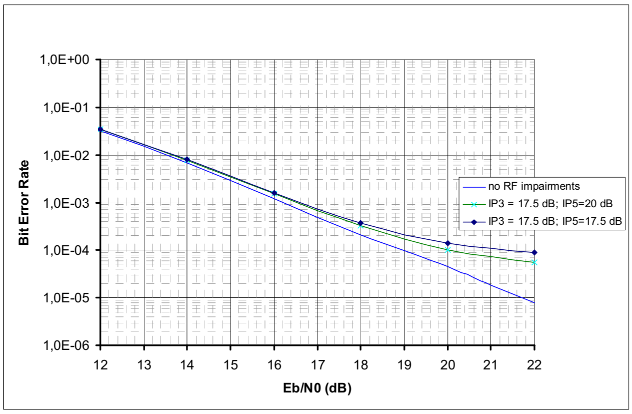

Figure 5 : Receiver non-linearity.

I&Q mismatch

In this section, we intend to measure the degradation introduced by I&Q mismatch, and also to validate the following assumption : for a fixed $|\beta|^2$, various configurations of ε and ΔΦ should give the same performance.

Figure 6 shows the results obtained for $|\beta| = 0.04$and the following configurations:

- Case (a) : ε = 0.04 and ΔΦ = 0,
- Case (b) : ε = 0 and ΔΦ = 0.04 (2.3°),
- Case (c) : ε = 0.0346 and ΔΦ = 0.02 (1.15°),
- Case (d) : ε = 0.02 and ΔΦ = 0.0346 (2°)

It is clear that the assumption is validated. It is important since it gives a degree of freedom to the RF designers.

Figure 7 presents the performance obtained with various $|\beta|$. If we want to keep the degradation below 1 dB, then $|\beta|$ should not be superior to 0.04. In practice, this means that the maximum phase offset is around 2° and the gain offset is 10log(1+ε) < 0.2dB. These figures are in line with the RF specifications of IEEE 802.11a commercial products. On the other hand, these specifications should be reached for a much broader bandwidth, which may cause some problems to RF designers. Development of low complexity compensation system for such impairments is highly recommended to relax RF constraints.

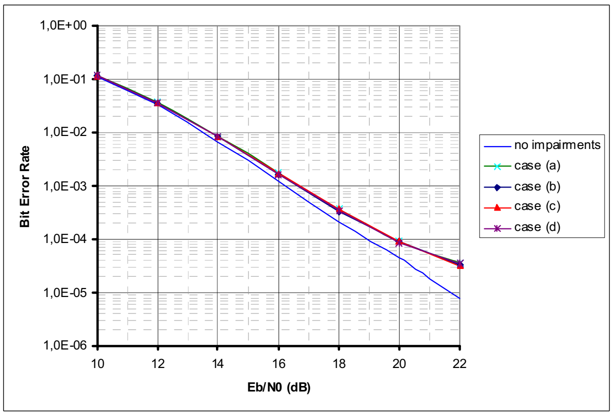

Figure 6 : Influence of ε and ΔΦ for a fixed β.

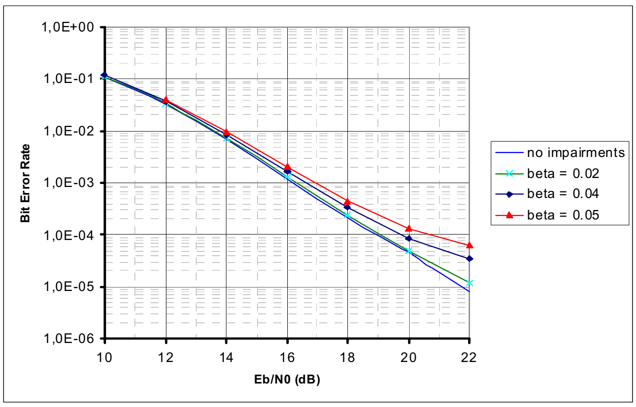

Figure 7 : influence of $|\beta|$.

Carrier frequency offset

All multi-carrier receivers are known to be very sensitive to carrier frequency offset [7]. It creates inter-carrier interference which is added to the multiple access interference in a MC-CDMA context. We have evaluated the sensitivity of the receiver according to the normalized carrier frequency offset: $\delta = N_{FFT}\Delta F T_s$. It should be noted that no compensation algorithm was implemented in the receiver. This means that the presented results give the residual carrier frequency offset that can be supported by the system.

The results are presented in Figure 8. As expected, the receiver is very sensitive to carrier frequency offset. It should be less than 1% of the sub-carrier spacing. Moreover, the estimation may be quite tricky since a small estimation error, δ=0.02 instead of 0.01, leads to an error floor.

The oscillator stability after compensation is thus 0.01*56.25 kHz/(5 GHz) = 0.01125 ppm.

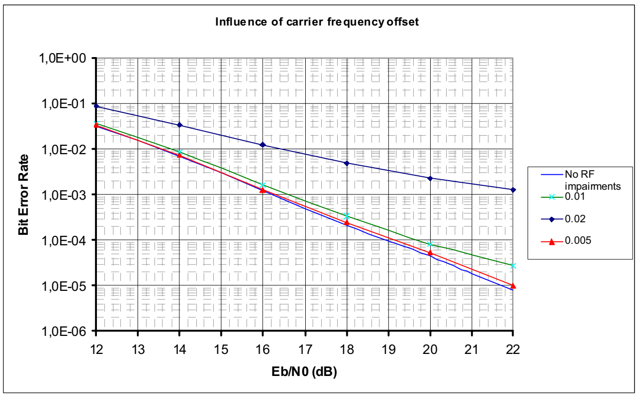

Figure 8 : sensitivity to carrier frequency offset.

Phase noise

At the receiver side, phase noise introduces a Common Phase Error (CPE) on every sub carrier and Inter Carrier Interference (ICI) [9]. In our study, we assume that the CPE is perfectly estimated and corrected, while ICI is not compensated. The resonance coefficient of the PLL filter is set to $\xi = 1/\sqrt{2}$ and cut-off frequency is either equal to 57.6 kHz (sub-carrier spacing), 100 kHz or 1MHz. Two different specifications of the oscillator have been selected:

- Good quality : -120 dBc at 1 MHz (D=19.7)
- Medium quality : -100 dBc at 1MHz (D=1973.9)

Figure 9 and Figure 10 present the performance obtained with the 2 specifications and the 3 cut-off frequencies. We observe that the a very accurate oscillator is required to support a 16-QAM modulation. We also remark that the performances degrade with a smaller cut-off frequency.

If we note S(w) the phase noise spectrum and H(w) the PLL filter transfer function, the SINR degradation for a fully loaded system is [7]: $D = 10\log\left(1 + \sigma_{\Delta\Phi}^2\right)$ where

$\sigma^2_{\Delta\Phi} = \int_{-\infty}^{+\infty} S(w)|H(w)|^2 dw \approx 2\int_{w_n}^{+\infty} S(w)|H(w)|^2 dw$ is the phase noise variance. It increases when the cut-off frequency decreases, which explains the observed simulation results.

As observed in non-linearities studies, the phase noise requirements of beyond 3G systems become more and more severe. In UMTS products, the –120dBc attenuation of phase is usually expected at 15 MHz, that is at least 3 channel from the wanted signal. To increase even more performances of future systems, a co-design between baseband and RF will be probably necessary.

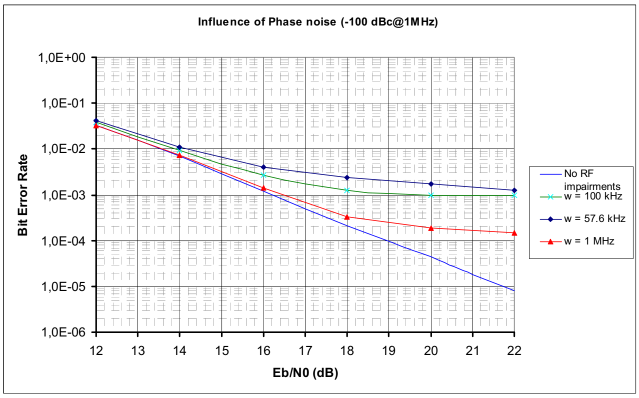

Figure 9 : sensitivity to phase noise (medium quality).

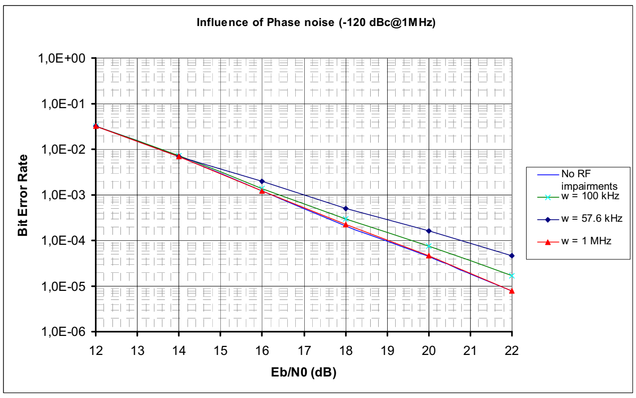

Figure 10 : sensitivity to phase noise (good quality).

VI. CONCLUSION

In this paper, we have described the specification of a MC-CDMA system that have been defined for a beyond 3G system and evaluated its sensitivity to RF impairments. They consist in receiver non linearity, I&Q mismatch, carrier frequency offset and phase noise.

The main result from this study is the very high sensitivity of the receiver to carrier frequency offset and phase noise. The residual carrier frequency offset remaining after compensation should be lower than 1% of the sub-carrier spacing. Moreover, a small compensation error leads to a dramatic error floor which destroys the performance of the system. Concerning phase noise, an oscillator with a good quality is required to support a 16-QAM modulation. A power spectral density of –120dBc at 1MHz is required.

It has been shown that the behaviour of RF front-end in the channel bandwidth is going to become the key point for future high data-rate system. Compensation of RF impairments at base-band level will probably be unavoidable. Moreover, a cross-layer optimisation should be done to reflect these new requirements at higher levels.

This work will serve as a basis for the specification of a MIMO RF front-end that will be developed in the IST 4MORE European project [10].

ACKNOWLEDGEMENTS

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