USING A DIRECT CONVERSION RECEIVER IN EDGE TERMINALS – A NEW DC OFFSET COMPENSATION ALGORITHM

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ABSTRACT – The basic concept of EDGE (Enhanced Data rates for GSM and TDMA/136 evolution) is to provide higher data rate transmission, over 384 kbits/s, than in the GSM and TDMA/136 systems of today. In order to be able to reach such a high data-rate within the GSM bandwidth a very good receiver design is required. Further, from a terminal perspective, the receiver design needs to be efficient from both a cost, current and size point of view. A direct conversion receiver fulfill these criterias. However, such an architecture gives rise to a DC offset that must be taken care of in order to do good demodulation. In this paper a DC offset compensation algorithm based on joint estimation of the radio channel and the DC offset is presented, giving competitive receiver performance for the EDGE system.

I. INTRODUCTION

The idea behind EDGE (Enhanced Data rates for GSM and TDMA/136 Evolution) is to provide higher data-rate transmission per radio time-slot than in the GSM system of today. This is achieved by introducing a more efficient modulation method, $3\pi/8$ offset 8-PSK, within the GSM burst- and frequency-structure. By introducing the new modulation together with the link adaptation and incremental redundancy concepts, EDGE allows for data rates which meet the IMT-2000 requirements, i.e. over 384 kBits/s. Different aspects of the EDGE concept have been presented and evaluated earlier in, e.g., [3] [4], [5], [6] and [9]. In [9], a preliminary modulation proposal for EDGE was assumed, whereas in [3], [4], [5], and [6] the standardized modulation, $3\pi/8$ offset 8-PSK, was used.

However, there is a price in terms of requirements on the receiver distortion in order to decode the received data at such a high data rates. Further, from a terminal perspective, the receiver design needs to be efficient from a cost, size and current consumption point of view in order to allow high volume production.

An example of such an efficient receiver architecture fulfilling these mobile terminal requirements is a direct conversion receiver see e.g. [7]. The idea behind the direct conversion receiver is that the incoming carrier is directly converted down to baseband without use of any intermediate frequencies. However, such a receiver architecture also suffers from some problems, for instance the DC-offset problem. The DC offset arises mainly from three different sources, see [7] and figure 1: (1) Local oscillator (LO) signal leaking to and reflecting off the antenna and self-downconverting to DC through the mixer, (2) a large near-channel interferer leaking into the LO and self-downconverting to DC, and (3) transistor mismatch in the signal path. By careful front-end receiver design, the leakage due to (2) and (3) can be reduced to some extent, but the DC offset can not be completely eliminated, and it therefore has to be taken care of in the baseband processing.

The first and most immediate idea for reducing the DC offset is to do long-term averaging of the baseband signal, and remove the DC by subtracting the DC estimate. However, in the EDGE system, as well as in all high performance cellular systems using frequency hopping, long time average is not possible, due to different DC offsets for different carrier frequencies. Further, doing short time averaging, i.e., over only one EDGE burst ($\approx$ 140-150 data samples), gives rise to a residual DC offset which degrades the receiver performance especially for channel coding schemes with high code-rate. Thus, other approaches have to be used.

In this paper a DC-offset compensation algorithm for EDGE, based on joint estimation of the DC offset and the radio channel, is presented. In chapter 2, the origin of the DC offset is analyzed, in chapter 3 the algorithm is described, and in chapter 4 simulation results are presented. Finally, in chapter 5 the conclusions are presented.

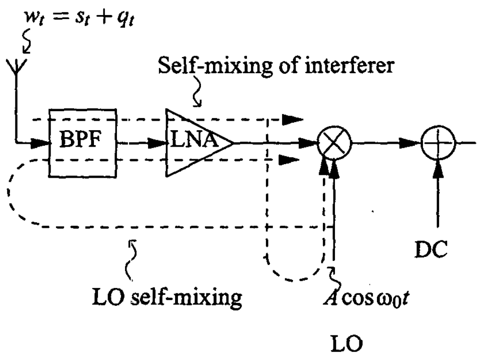

Figure 1: Three main sources of leakage giving rise to DC offset in a direct conversion receiver. 1) LO self-mixing, 2) self-mixing of interferer (q_t), and 3) DC due to transistor miss-match in the signal path.

II. ANALYSIS OF THE DC OFFSET

In order to make a good DC compensation, it is important to understand the origin of the DC offset problem. Therefore, we start with an analysis of the DC offset problem.

Assume that the received signal after the front-end band pass filter (BPF in figure 1), can be written as

$$\begin{aligned} w_t &= s_t + q_t \\ &= r_t \cos(\omega_0 t + \varphi_t) + p_t \cos(\omega_1 t + \theta_t) \end{aligned}$$

where s_t is the desired signal at carrier frequency ω_0, and q_t is a large near-channel interfering signal at carrier frequency ω_1, and where ω_0 and ω_1 are within the pass band of the band-pass filter. The local oscillator can be expressed

$$\begin{aligned} LO_I(t) &= Acos(\omega_0 t) \\ LO_Q(t) &= Asin(\omega_0 t) \end{aligned}$$

where LO_I and LO_Q are the local oscillators of the $I-$ and $Q-$ component respectively. The LO signal and the received signal are multiplied in the mixer, and due to LO leakage, interferer leakage, and transistor mismatch, see figure 1, the output from the I and Q component mixers can be written as

$$\begin{aligned} \tilde{I}_t &= \left(w_t + \alpha_I' LO_I(t)\right)(\beta_I w_t + LO_I) + I_{DC} \\ \tilde{Q}_t &= \left(w_t + \alpha_Q' LO_Q(t)\right)(\beta_Q w_t + LO_Q) + Q_{DC} \end{aligned}$$

Here α' and β', represent the leakage factors for the LO- and received signal- leakage respectively and I_{DC}, Q_{DC} are the DC offsets due to transistor mismatch.

The $\tilde{I}$ and $\tilde{Q}$ components are then analog to digital converted and low-pass filtered. After certain normalizations, and using the assumption that $|p_t|^2 \gg |r_t|^2$, i.e., the near-channel interferer is much larger than the desired signal, the digital I, Q baseband components, sampled at the symbol rate, can be written as

$$\begin{aligned} \tilde{I}_n &= I_n + a|\tilde{p}_n|^2 + I_{DC} \\ \tilde{Q}_n &= Q_n + b|\tilde{p}_n|^2 + Q_{DC}. \end{aligned}$$

where I_n, Q_n are the desired I and Q components, and I_{DC}, Q_{DC} are the DC components on the I and Q component respectively. Further, $a|\tilde{p}_n|^2$ and $b|\tilde{p}_n|^2$ are the low-pass filtered and sampled squared envelope of the interfering signal on the I and Q channels respectively.

In case of digital transmission over radio channels with inter-symbol interference, such as for instance in GSM or EDGE, the desired I and Q components can be written in a complex notation, according to (see e.g. [8])

$$\begin{aligned} S_n &= I_n + jQ_n = \sum_{k=0}^{L} h_k u_{n-k} + e_n \\ &= H^T U_n + e_n \end{aligned}$$

where $H = [h_0, \dots, h_L]$ is a vector of complex-valued channel filter taps, $U_n = [u_n, \dots, u_{n-L}]$ is a vector of complex-valued transmitted symbols, and e_t is some kind of complex valued noise with mean zero. Thus, the base band signal in our case can be written according to

$$\begin{aligned} \tilde{S}_n &= \tilde{I}_n + j\tilde{Q}_n && (2.1) \\ &= H^T U_n + (a + jb)|p_n|^2 + I_{DC} + jQ_{DC} + e_n. \end{aligned}$$

From (2.1) one can see that we have one pure DC component and one component that depends on the envelope of the interfering signal. If we assume that we can suppress the varying DC component by careful front end radio receiver design, the following baseband model for the received signal is appropriate

$$\tilde{S}_n^1 = \tilde{I}_n + j\tilde{Q}_n = \sum_{k=0}^{L} h_k u_{n-k} + m + e_n \qquad (2.2)$$

where m is a parameter describing the unknown DC offset. Now we have a model of the received signal and we can develop the DC compensation algorithm.

III. THE ALGORITHM

In order to do synchronization and channel estimation, one needs to do de-rotation of the received sequence in order to compensate for the inherent $3\pi/8$ offset between consecutive EDGE symbols. After the de-rotation the signal can be written according to

$$\begin{aligned}
\tilde{S}_n &= e^{-j\frac{3\pi}{8}n}\tilde{S}_n^1 & (3.3)\\
&= e^{-j\frac{3\pi}{8}n}\sum_{k=0}^{L} h_k u_{n-k} + me^{-j\frac{3\pi}{8}n} + \tilde{e}_n \\
&= e^{-j\frac{3\pi}{8}n}\sum_{k=0}^{L} h_k e^{j\frac{3\pi}{8}(n-k)} v_{n-k} + me^{-j\frac{3\pi}{8}n} + \tilde{e}_n \\
&= \sum_{k=0}^{L} \tilde{h}_k v_{n-k} + me^{-j\frac{3\pi}{8}n} + \tilde{e}_n \; n = 1,\ldots,N_B
\end{aligned}$$

where N_B is the number of symbols in the burst, v_t is an 8-PSK signal, i.e., $v_t = e^{j\frac{\pi}{4}n_t}$, $n_t = 0,\ldots,7$. A standard correlation algorithm can be used for synchronization. For the following we assume that the training sequence starts at $k = n_0$.

The next step is to estimate the channel and the DC offset. As can be seen from (3.3), the DC offset is rotating due to the de-rotation of the received burst. Thus, a reasonable signal model to be used in the channel estimator is

$$\begin{aligned}
\hat{S}_n &= \sum_{k=0}^{L} \tilde{h}_k v_{n-k} + me^{-j\frac{3\pi}{8}n} = \tilde{H}^T \tilde{U}_n, \\
& \quad n = L,\ldots,N_{TS}-1 & (3.4)
\end{aligned}$$

where N_{TS} is the number of training symbols, $\tilde{H} = [\tilde{h}_0,\ldots,\tilde{h}_L,m]^T$ and $\tilde{U}_n = [v_n,\ldots,v_{n-L},e^{-j\frac{3\pi}{8}n}]^T$. Assuming Gaussian white noise one can show that joint estimation of the channel taps and the DC offset using Least Squares (LS) estimation is equivalent to ML estimation, and is therefore an (asymptotically) optimal estimator. Thus, we use LS to jointly estimate the channel taps and DC offset, i.e., (assuming the training sequence starts at $n = n_0$)

$$\hat{\tilde{H}} = \left(\sum_{k=L}^{N_{TS}-1} \tilde{U}_k \tilde{U}_k^H\right)^{-1} \sum_{k=L}^{N_{TS}-1} \tilde{U}_k^* \tilde{S}_{n_0+k}. \quad (3.5)$$

The estimated DC offset is then subtracted from $\tilde{S}_n$, i.e,

$$\begin{aligned}
\hat{S}_n &= \tilde{S}_n - e^{-j\frac{3\pi}{8}(n-n_0)}\hat{m}, \\
&= \tilde{H}^T U_n + (m-\hat{m}) + \tilde{e}_n \\
&\approx \hat{\tilde{H}}^T U_n + \tilde{e}_n,\; t = 1,\ldots,N_B. & (3.6)
\end{aligned}$$

Now, we have removed the DC offset, and the signal $\hat{S}_n$ together with the estimated channel filter taps $\hat{H}$ is then used in the equalization process.

IV. SIMULATION RESULTS

The receiver used in the simulations is similar to what have been used in [9], and comprises of a receiver filter, a synchronization algorithm, a channel and a DC estimation unit according to what has been described above, an equalizer, a deinterleaver, and a Viterbi decoder. On the received signal, a random DC offset is added in order to model the impairments due to the direct conversion receiver. The DC offset has been assumed to be constant over one EDGE burst, but different for different bursts, implying long time averaging is not possible. Further, a five-tap channel model is used, i.e., L=4 in (3.4). The equalizer is sub-optimum with approximately four times the complexity per burst compared to the 16-state Viterbi algorithm for GMSK.

In order to evaluate the DC compensation algorithm, the receiver performance for the above mentioned receiver is compared to the receiver performance for 1) a reference receiver with the same receiver parameters as described above but without DC offset and DC compensation algorithm, and 2) a receiver including a DC component, but using only burst averaging to compensate for the DC offset. Two EDGE 8-PSK EGPRS (Enhanced General Packet Radio Service) coding schemes, MCS-5 (code rate 0.37, data rate 22.4 kbits/s/time slot), and MCS-9 (code rate 1.0, data rate 59.2 kbits/s/time slot) see [1], has been used in the simulations. Further, simulation results for two scenarios are presented:

- A static channel disturbed by additive white Gaussian noise, and
- A TU3 channel, see [2], with no frequency hopping, disturbed by one co-channel interferer.

The simulation result can be found in figures 2-5.

As can be seen in the figures, the receiver performance for the direct conversion receiver including the DC compensation algorithm is less than 1 dB worse than the receiver performance for the reference receiver at 10% Block Error Rate for both coding schemes and channel scenarios. Further, note that the simple averaging method gives reasonable performance in the MCS-5 case, i.e, with much coding, but it is useless in the case of no coding (MCS-9). This is because averaging over only one burst and subtracting the DC estimate gives rise to a residual DC offset of the order $1/\sqrt{N}$, where N is the number of data used in the averaging process. The origin offset, around -20 dB relative to the signal strength in the EDGE case, which seems to be small, is however large enough to degrade the receiver performance for coding schemes with high code rate. Thus, it shows that in order to be able to manage the highest data rates in EDGE it is very important to use a correct channel model which includes a possible DC offset in the demodulation process.

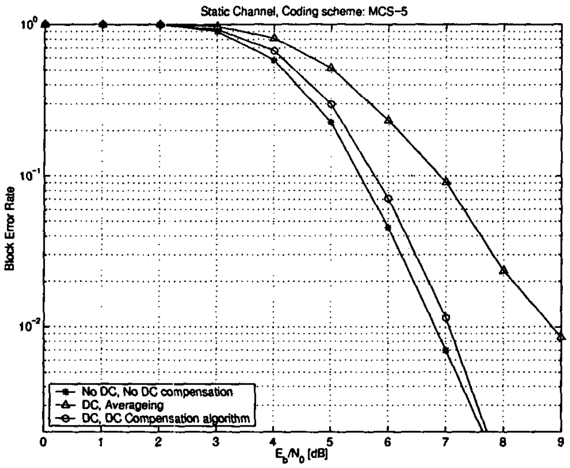

Figure 2: Block Error Rate, Static channel, Coding scheme: MCS-5.

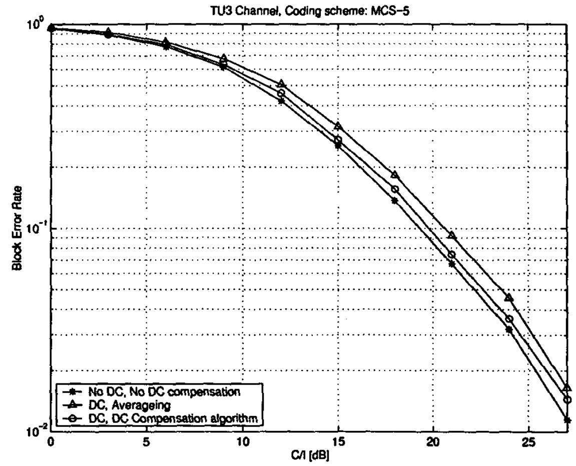

Figure 4: Block Error Rate, TU3 channel, Coding scheme: MCS-5.

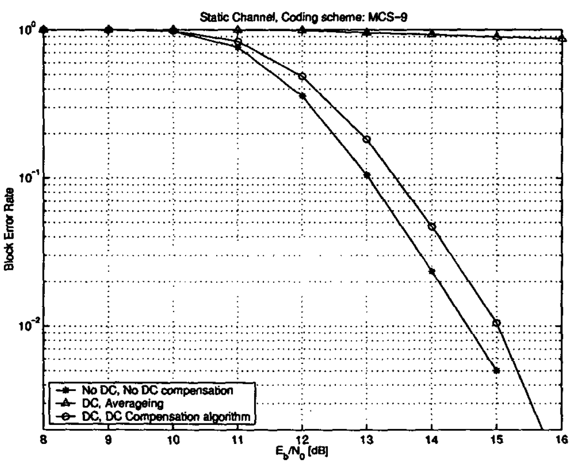

Figure 3: Block Error Rate, Static channel, Coding scheme: MCS-9.

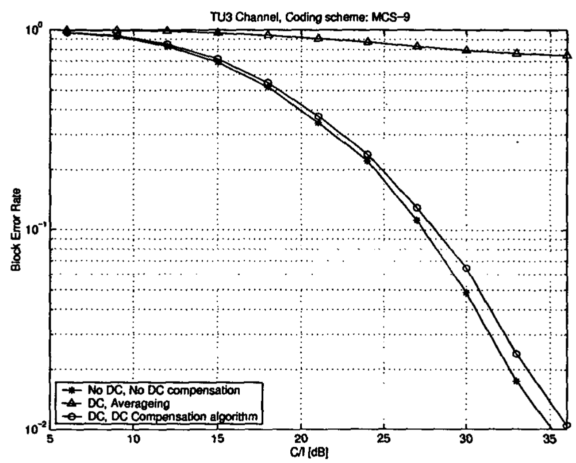

Figure 5: Block Error Rate, TU3 channel, Coding scheme: MCS-9.

V. CONCLUSIONS

In this paper a new DC compensation algorithm, taking care of DC offsets which occur when using a direct conversion receiver, is presented. The algorithm is investigated assuming a direct conversion receiver used in 8-PSK mode EDGE. It can also be used in the EDGE GMSK mode as well as in other TDMA systems. From the simulations one can draw two conclusions.

1. If using a direct conversion receiver for EDGE it is necessary to use an advanced DC offset compensation algorithm. The algorithm, based on simultaneous channel tap and DC offset estimation, presented in this paper, gives competitive receiver performance compared to other more advanced and expensive receiver architectures.

2. In order to be able to manage the highest data rates in EDGE it is very important to include internally generated receiver distortions, such as DC offsets, in the channel model.

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