$\Sigma\Delta$ ADC Design Considerations for an UMTS Receiver

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Abstract— **This paper gives an insight on how to derive the minimum required specifications both for the anti-aliasing filter (AAF) and the analog-to-digital converter (ADC) in the case of an UMTS receiver. Based on the receiver specifications, we come up with a design plan for both discrete time $\Sigma\Delta$ ADC as well as for a continuous time $\Sigma\Delta$ ADC.**

I. Introduction

The ability to process higher user bit rates is the major objective of today's wireless systems. To fulfill this demand, third generation (3G) wireless systems are introduced. For these systems, wideband code-division multiple access (WCDMA) technology seems to be the widely adopted air interface. Rather than being separated in frequency and time, users are now separated by orthogonal codes. Here, we will focus on the FDD (Frequency Division Duplex) mode of UMTS (universal mobile telecommunication system), the standard most relevant in Europe.

The paper starts by summarizing the minimum receiver specifications as they are specified in the standardization documents. Next, a commonly used receiver topology, the low-IF receiver is discussed together with some architectural issues. Finally, a design plan is given for a discrete time $\Sigma\Delta$ ADC as well as for a continuous time $\Sigma\Delta$ ADC.

II. Minimum UMTS Receiver Specifications

This section gives some general remarks concerning WCDMA and briefly discusses the receiver characteristics as they are specified by [1]. We focus on the specifications that are relevant for ADC designers.

A. General Info

Although UMTS/FDD is designed to operate in different paired bands, this paper concentrates on the first operating band (called region I). The specifications for the other frequency bands are similar and can be found in [1]. For region I, the uplink frequency band (Tx) is from 1920-1980 MHz and the mobile reception (downlink frequency band or Rx) is within the band of 2110-2170 MHz. The channel spacing is 5 MHz. After spreading and scrambling, the quadrature phase-shift keying (QPSK) modulation occupies a bandwidth (B) nearly equal to the chip rate R_c of 3.84 Mcps.

B. Processing Gain and Minimum Required SNR

The UMTS/FDD standard describes a number of test scenarios for which the user bit rate (R_u) is fixed at 12.2 kbps and the bit error rate (BER) must be below 0.1%. To obtain this BER, a minimum Signal-to-Noise Ratio (SNR), commonly known to and used by ADC designers, at the output of the receiver is required. However, in the field of communication, the ratio E_b/N_0 is used instead of SNR. E_b is the energy per information bit and N_0 is the noise power spectral density. The relationship between E_b/N_0 and SNR is given by:

$$\frac{E_b}{N_0} = \mathrm{SNR} + G_p \text{ [dB]} \tag{1}$$

Here G_p denotes the user data processing gain. This processing gain originates from two processes (G_s and G_c). The spreading gain G_s results from spreading and despreading in WCDMA. Despreading does not change the total noise and interferer power nor the total power of the wanted signal. However noise and interferers, which do not correlate with the despreading code, are spreaded into a wider bandwidth. In contrast, the desired signal energy is concentrated in a bandwidth that corresponds to the channel symbol rate (R_s = 30 kbps) during the despreading. As a result, the SNR in the bandwidth of the despreaded data is improved with the spreading gain:

$$G_s = 10\log\frac{R_c}{R_s} = 10\log\frac{3.84\text{Mcps}}{30\text{kbps}} = 21 \text{ dB} \tag{2}$$

An additional coding gain (G_c) of 4 dB is obtained from signal decoding. Spreading gain and coding gain result into the processing gain G_p:

$$G_p = G_s + G_c = 25 \text{ dB} \tag{3}$$

The minimum required E_b/N_0 for a BER of 0.1% is determined by simulations in [2] and should be:

$$\left.\frac{E_b}{N_0}\right|_{\text{min}} = 7 \text{ dB} \tag{4}$$

C. Crest factor

Since several user- and control- signals are multiplexed onto one frequency channel, a single radio channel behaves like band-limited noise and significant amplitude variations may

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occur. These amplitude variations are described by the crest factor ξ which is defined as:

$$\xi = P_{\text{PEAK}} - P_{\text{AV}} \tag{5}$$

where P_{PEAK} is the 99.9% limit of the instantaneous-power distribution in decibels and P_{AV} the mean value of the input power. In UMTS/FDD, the crest factor can vary between 4.5 dB and 11 dB depending on the number of code channels and the selected codes [3].

D. Reference Sensitivity and Maximum Input Level

The reference sensitivity level (P_{ref}) is an important measure for the receiver as it expresses the minimum mean power received at the antenna port of the receiver for which the BER may not exceed 0.1%. It is defined by the power of the signal that is carrying the actual user data:

$$P_{\text{ref}} = -117 \text{ dBm} \tag{6}$$

In this paper signal powers correspond to the total power in the channel bandwidth of 3.84 MHz.

Form eq. (3), (4) and (6), we can find the maximum allowable noise energy N_{max} in the channel bandwidth before despreading at the antenna port:

$$N_{\text{max}} = P_{\text{ref}} + G_p - \left.\frac{E_b}{N_0}\right|_{\text{min}} = -99 \text{ dBm} \tag{7}$$

The maximum input level on the other hand is defined as the maximum input power of the selected channel as received at the antenna port for which the BER should not exceed 0.1%.

$$P_{\text{max}} = -25 \text{ dBm} \tag{8}$$

E. Blocking Characteristics

The worst case interferer power spectrum is defined by the so-called blocking characteristics. A desired signal, 3 dB above P_{ref} at a frequency f_0, should be received correctly (BER $< 0.1\%$) in the presence of unwanted interferers at frequency f_{block}. Two different types of blocking signals can be distinguished:

1) Modulated WCDMA channels for interferers falling into the RX band or into the first 15 MHz below or above this band. This type of blocking is denoted in-band blocking.

2) Continuous waves (CW) for interferers falling more than 15 MHz outside the Rx band, which are denoted out-of-band blocking.

For a wanted signal situated at the edge of the Rx band, the worst case blocking power spectrum is shown in Fig. 1. Note that f is defined as the frequency difference from f_0 instead of its effective value.

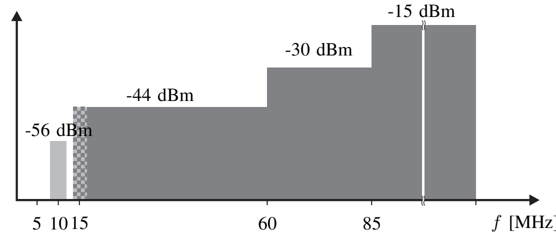

Fig. 1. Worst case blocking signals, light gray: in-band and dark gray: out-of-band blockers. The signal powers are specified over the channel bandwidth.

III. Receiver Topology

In many wireless communication systems, low-IF receivers (see Fig. 2) are chosen to retrieve the information signal from the outside world. Such a receiver consists of an antenna, an analog part, an ADC and a digital signal processor (DSP). The analog part is made up of a quadrature down converter mixer, an anti-aliasing filter (AAF) and an automatic gain control (AGC). In a low-IF receiver, the RF signal is mixed down to a non-zero low or moderate intermediate frequency. Low-IF receiver topologies have many of the desirable properties of zero-IF architectures, but avoid the DC offset and $1/f$-noise problems and also even-order distortion has less effect [4]. The main disadvantage is that the image rejection ratio requirement is larger. This is due to the fact that in a zero-IF architecture the image signal is originating from the wanted signal itself, while in the low-IF case it is originating from a blocking signal which can be much larger than the wanted signal (see e.g. Fig. 1).

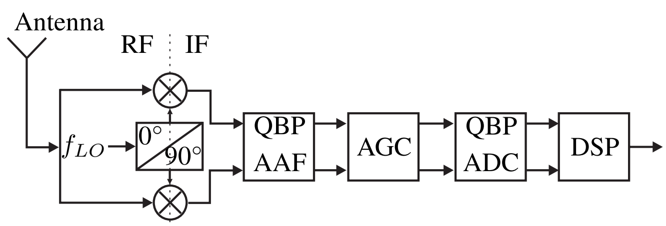

Fig. 2. Low-IF receiver architecture.

We will now discuss successively the following different parts of Fig. 2: the ADC, the AAF and the AGC.

A. Analog-to-Digital Converter

A quadrature bandpass (QBP) $\Sigma\Delta$ ADC [4] is used since it is well suited for the use in low-IF receivers, where both the I and Q signal need to be digitized. This ADC directly performs 'complex A/D conversion', instead of digitizing separately the two signals with two real bandpass (BP) $\Sigma\Delta$ ADCs. To achieve the same order of noise shaping, the solution with one QBP $\Sigma\Delta$ ADCs will need only half the amount of integrators compared to the solution with two BP $\Sigma\Delta$ ADC.

Among $\Sigma\Delta$ ADCs, two major types exist based upon the place where the sampling operation occurs. For Discrete Time (DT) $\Sigma\Delta$ ADCs, it occurs at the modulator input, for Continuous Time (CT) $\Sigma\Delta$ ADCs that is at the quantizer input. Therefore, in contrast to DT modulators, CT modulators provide an implicit anti-aliasing filtering (IAAF). Next, for reason of power efficiency, both the DT and the CT $\Sigma\Delta$ ADC need to be implemented with a feedforward structure [5]. With this structure, the signal transfer function (STF) of the DT ADC can be made identically one. Hence, the wanted signal and the nearby blocking signals have the same ADC gain. In CT $\Sigma\Delta$ ADCs, nearby blocking signals can have a much higher gain factor than the wanted signal. However, with the technique of [5], the STF is equal to one for all frequencies, except for those lying in the aliasing bands. There the STF equals $\frac{\text{NTF}(z)}{\text{NTF}_\infty}$, resulting in the implicit anti-aliasing filtering.

B. Anti Aliasing Filter

We are interested to integrate the whole receiver, without the need for external components, like SAW filters. To meet this objective, no channel selection is done in the analog domain. Only the minimum required filtering to suppress adjacent channels which would alias as co-channel interferers due to the ADC sampling action is implemented.

For the AAF we prefer a Butterworth filter since its frequency response is maximally flat in the passband. Other filters may have sharper cut-offs but for an AAF this property is not required. Moreover, a low Error-Vector-Magnitude (EVM) can be achieved by using a Butterworth filter, without significantly increasing the filter complexity, compared to filters with a sharper cut-off. EVM is the root-mean-square (rms) error between the ideal constellation points and the actual symbols at the optimal sampling instants. The EVM is an important measure since it can be linked to the BER (and E_b/N_0) of the receiver and therefore it needs to be limited [3].

For a Butterworth filter the AAF attenuation is given by:

$$|\text{AAF}(f)|^2 = \frac{1}{1+\varepsilon^2\left(\frac{f}{f_p}\right)^{2N}} \tag{9}$$

where N is the filter order, f_p the maximal passband frequency and ε the attenuation at the passband edge. Since f is defined as the frequency difference from f_0 instead of its effective value, eq. (9) defines the attenuation of a complex QBP AAF, with an asymmetrical transfer characteristic. For the same reason as for the ADC, one complex QBP AAF is used instead of two BP AAFs. Note that for QBP filtering, f_p is equal to $B/2$ in contrast to lowpass filters where it would be B = 3.84 MHz. Furthermore, simulations reveal that it is necessary to limit ε to 1 dB in order to be able to achieve the desired BER.

C. Automatic Gain Control (AGC)

To avoid saturation (instability) of the ADC, the maximum IF signal power must be below the full-scale value (maximum stable input) of the ADC. To ensure this, there is a constant and a variable gain between the antenna and the ADC input. The constant gain (K_c) is adjusted to ensure that the maximally expected input level equals the full-scale value. The variable gain (K_v) will supply extra gain if the input level is below this maximum input level. This way, the dynamic range (DR) specifications of the ADC itself can be limited. In this paper, we assume that the variable gain is not continuous but varies in discrete steps. This is a good compromise between complexity and performance.

IV. Design Specifications

A. General Remarks

To implement an UMTS receiver, two conditions are of major concern. The first condition is related to the AAF. Signals (f_i, $P_{\text{block},i}$) which may alias as co-channel interferers must be attenuated in such a way that they are negligible compared to $N_{\max}$, see eq. (7). We concentrate all the energy of the out-of-band interferer into one single tone situated at the edge of the aliasing band ($f_i = f_s - B/2$) with f_s the sampling frequency. This is the worst case situation since this tone experiences both the least AAF and IAAF attenuation. For DT $\Sigma\Delta$ ADCs it is required that:

$$|\text{AAF}(f_i)|^2 P_{\text{block},i} = P_{\min} \ll N_{\max} \tag{10}$$

$P_{\min}$ is chosen such that aliasing interferers are negligible compared to $N_{\max}$. A good idea to chose for a safety distance of 20 dB since the sum of every interferer, noise term and non-linearity needs to be below $N_{\max}$. Therefore we design separately every term 20 dB below $N_{\max}$.

$$P_{\min} = N_{\max} - 20 \text{ dB } \ll N_{\max} \tag{11}$$

For CT $\Sigma\Delta$ ADCs, the AAF requirement is less stringent due to the existence of the IAAF. Now, it is required that:

$$|\text{AAF}(f_i)|^2 |\text{IAAF}(f_i)|^2 P_{\text{block},i} = P_{\min} \ll N_{\max} \tag{12}$$

or, with the definition of IAAF (see higher):

$$|\text{AAF}(f_i)|^2 \frac{|\text{NTF}(f_i)|^2}{\text{NTF}_\infty^2} P_{\text{block},i} = P_{\min} \tag{13}$$

Both for DT and CT $\Sigma\Delta$ ADCs, we assume that the wanted signal is situated at the edge of the Rx band. For the AAF this is the worst case, since the high level blocking signals are closer than in the situation where the wanted signal is in the center of the Rx band. For this situation, $P_{\text{block},i}$ is shown in Fig. 1.

The second condition originates from the fact that the NTF must be designed in such a way that the equivalent input in-band noise (IBN) should be negligible compared to $N_{\max}$. This means that the IBN at the input of the ADC needs to fulfill the following condition:

$$\text{IBN} = \int_{-B/2}^{B/2} |\text{NTF}(f)|^2 Q^2 df = K_c K_v P_{\min} \ll K_c K_v N_{\max} \tag{14}$$

where Q^2 denotes the noise spectrum:

$$Q^2 = \frac{\Delta^2}{12}\frac{2}{f_s}\ ,\ \Delta = \frac{2A_{\text{fs}}}{2^b - 1} \tag{15}$$

with A_{fs} the full-scale amplitude of the ADC and b the number of bits in the internal quantizer.

In conclusion, both conditions, eq. (10) and (14) must be satisfied in order to obtain the required UMTS receiver specifications in case of a DT $\Sigma\Delta$ ADC. In case of a CT $\Sigma\Delta$ ADC, eq. (13) and (14) must be fulfilled. Note that in this situation, the NTF is a common design parameter. Consequently, the attenuation of the quantization noise and signals which may alias as co-channel interferers can not be designed separately which is in contrast to DT $\Sigma\Delta$ ADCs.

V. Design Plan for DT $\Sigma\Delta$ ADCs

The design plan for DT $\Sigma\Delta$ ADCs is organized as follows: first we will focus on the AAF requirements, next on the ADC. This is possible since eq. (10) and (14) can be designed separately.

A. Minimum Sampling Frequency

In this section, the minimum sampling frequency for a DT ADC, for which the aliasing components are sufficiently suppressed (eq. (10)) is calculated. Therefore we determine the maximum allowable aliasing signals for a Butterworth filter with a given order N. With eq. (10) we obtain:

$$P_{\text{block},i} = \frac{P_{\text{min}}}{|\text{AAF}(f_i)|^2} \tag{16}$$

In Fig. 3, the maximum allowable aliasing signals for the Butterworth-AAF with $N \in \{2,3,4\}$ are given. The step curve on this figure is equal to Fig. 1 and indicates the worst case out-of-band blocking signals ($P_{\text{block},i}$). The highest intersection of the fluent curve with the step curve gives us the minimum sampling frequency $f_{s,\text{min}}$. This graphical method is also illustrated in Fig. 3.

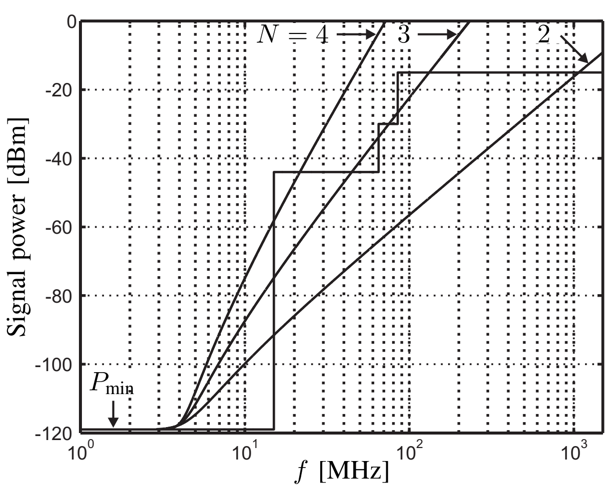

Fig. 3. Blocking signals and the maximum allowable aliasing signals

The minimum sampling frequency $f_{s,\text{min}}$ can also be determined analytically, given a filter order N, by substituting eq. (9) for $\text{AAF}(f_i)$ in eq. (10) and f_i by $f_s - B/2$ and by filling in the correct out-of-band blocking signal level. This value can be readily derived from the graph of Fig. 3.

$$f_{s,min} = \left(1 + \left(\frac{10^{\frac{P_{\text{block},i}-P_{\text{min}}}{10}} - 1}{\varepsilon^2}\right)^{\frac{1}{2N}}\right)\frac{B}{2} \tag{17}$$

In Table I, the minimum sampling frequency $f_{s,\text{min}}$ and minimum oversampling ratio OSR_{min} is given for $N \in \{2,3,4\}$. One can conclude that a third order Butterworth filter is a sensible choice for the AAF. Both the minimum required sampling frequency and the filter order are within the technological feasibility.

TABLE I
Minimum sampling frequency and minimum OSR

N	$f_{s,\text{min}}$	OSR_{min}
2	1073 MHz	279
3	132 MHz	34
4	22 MHz	6

B. Estimation of the Dynamic Range

Under the condition of the results of the previous section (third order AAF), we will make an estimation of the required DR to ensure that also condition (14) is fulfilled. At this moment, we have calculated the maximum allowable noise energy N_{max} = -99 dBm, introduced a safety distance (SD) of 20 dB which results in P_{min} = -119 dBm. Furthermore, the AAF also attenuates the in-band and out-of-band blockers resulting in residual blocking signals. The average power of the maximum residual blocking signal together with its crest factor is $P_{\text{block}}^{\text{res}}$ = -53 dBm. Finally, the AGC headroom (AGC_{HR}) has been set to 10 dB.

The estimation of the DR requirements for a DT $\Sigma\Delta$ ADC with a third order AAF is illustrated in Fig. 4. The axis in the middle contains the signal power levels at the antenna of the receiver. At the left and the right of it, the power levels are related to the full-scale of the ADC. In case no AGC is used, (the right of Fig. 4) the maximal input power is projected on the full-scale of the ADC, resulting in the value of the constant gain K_c. Now, that the constant gain is known, the maximal IBN at the input of the ADC is given by eq. (14), resulting in a DR requirement of 105 dB. In case an AGC is used (the left of Fig. 4), the maximal residual blocking signal $P_{\text{block}}^{\text{res}}$ needs to be projected to the full-scale of the ADC, taking into account the AGC headroom of 10 dB. In this case, the DR requirement becomes 66 dB.

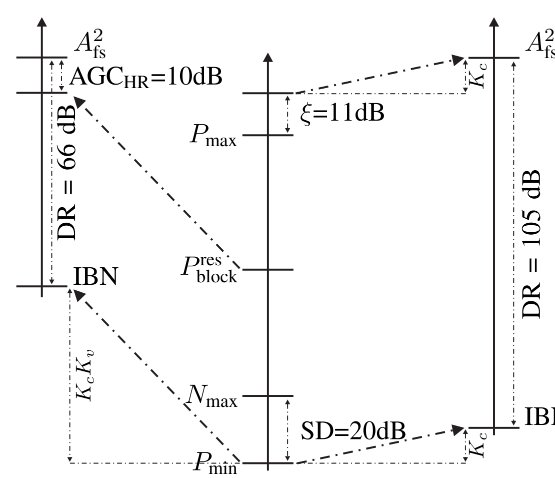

Fig. 4. DR requirement, with (left) and without (right) AGC.

The DR requirement for the presented receiver topology, with a third order AAF and with the AGC, can be met with a third order $\Sigma\Delta$ QBP modulator. The modulator has an internal 2-bit quantizer, a moderate NTF_∞ of 1.5 and is oversampled with OSR_{min} = 34 and has been designed and simulated with [6].

VI. Design plan for CT $\Sigma\Delta$ ADCs

For CT $\Sigma\Delta$ ADCs, the design plan will be different since the attenuation of the quantization noise and signals which may alias as co-channel interferers has a common origin. Therefore they can not be designed separately in contrast to DT $\Sigma\Delta$ ADCs. For CT $\Sigma\Delta$ ADCs, the following design plan is suggested:

1) Use a Butterworth AAF with order N. Since we try to design a receiver structure with an as low as possible filter order, only two situations with $N \in \{1, 2\}$ are discussed.

2) Given the filter order N, the residual blocking signals after AAF can be calculated and, for its part, the required DR can be estimated with the method described in section V-B.

3) Next, calculate the minimum required OSR to fulfill both the conditions, eq. (13) and (14) for different modulator orders k and number of bits in the quantizer b. The results are shown in Fig. 5 and Fig. 6 for N = 1 and N = 2.

4) Based upon these figures, a choice for N, k and b in function of the minimum required OSR can be made.

Comparing the receiver specifications in case of a DT with the ones in case of a CT $\Sigma\Delta$ modulator, we can conclude that in both situation a third order $\Sigma\Delta$ QBP ADC with an internal 2-bit quantizer, a moderate $NTF_\infty = 1.5$ and an OSR about equal to 34 is sufficient to fulfill the receiver specifications. However for a CT $\Sigma\Delta$ modulator the AAF specifications are less stringent due to existence of the IAAF. In this case a second order AAF is required which is in contrast with the necessity of a third order AAF in case of a DT $\Sigma\Delta$ modulator.

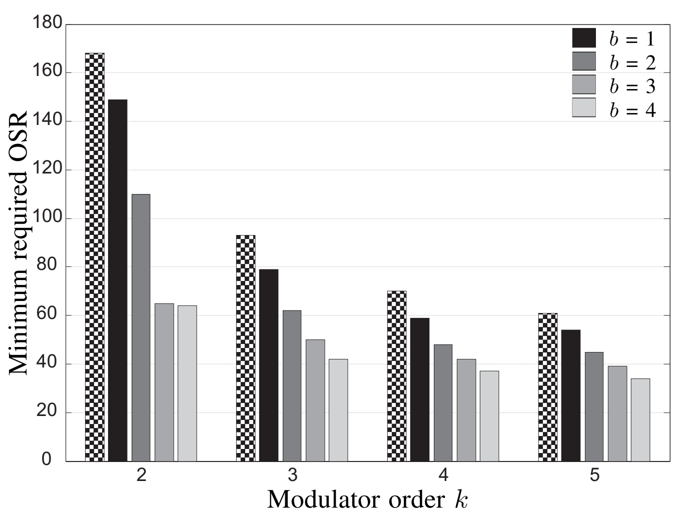

Fig. 5. Minimum required OSR to fulfill eq. (13) (draughtboard) and eq. (14) (others), in case of first order Butterworth AAF and a $NTF_\infty = 1.5$. The required DR is about 90 dB in this case.

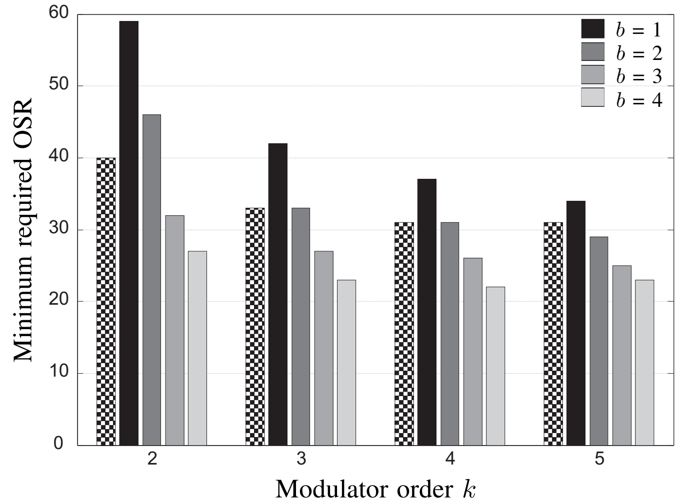

Fig. 6. Minimum required OSR to fulfill eq. (13) (draughtboard) and eq. (14) (others), in case of second order Butterworth AAF and a $NTF_\infty = 1.5$. The required DR is about 70 dB in this case.

VII. Conclusion

After having given a presentation of the minimum UMTS receiver specifications, the low-IF receiver has been discussed. It has been pointed out that the usage of a QBP $\Sigma\Delta$ modulator in such a receivers is a good choice for the ADC. For DT as well as for CT $\Sigma\Delta$ ADCs a design plan is given. Therefore, the minimum UMTS receiver specifications have been translated into design conditions recognizable by ADC designers. Based upon these conditions, the minimum OSR has been determined. For DT $\Sigma\Delta$ ADCs, the attenuation of the quantization noise and signals which may alias as co-channel interferers can be designed separately. For CT $\Sigma\Delta$ ADCs however, this is not the case. So, two different graphical methods, one for DT and one for CT, to determine the minimum OSR are given.

Acknowledgment

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