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DIGITAL COMPENSATION OF RECEIVER CLIPPING FOR DVB RECEPTION ON LOW-POWER MOBILE PHONES

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ABSTRACT

Battery life-time is a critical issue for digital television (DVB) viewing on mobile phones. The number of quantization steps used in the analog-to-digital converter (ADC) is an important factor in the total power consumption of a DVB receiver. The OFDM signals require a large resolution of the ADC. We propose a digital post-processing method that allows reduction of the peak-to-average margin by compensating for clipping artifacts when the ADC is saturated. Clipped OFDM peaks cause spurious signals on empty subcarriers, which can be used to recover the original signal and to eliminate clipping artifacts.

1. INTRODUCTION

At the moment of writing this paper, the analog PAL television transmitters are being switched off in many European countries and these are replaced by Digital Video Broadcasting (DVB). The DVB standard provides terrestrial transmission to the home (DVB-T) and to handhelds (DVB-H), based on Orthogonal Frequency Division Multiplexing (OFDM).

This provides reliable transmission over multipath channels, because it subdivides the data stream into many parallel subcarriers each of which sees a narrow band, i.e., non-selective channel. However, the nearly Gaussian distribution of the OFDM signal is seen as one of its main disadvantages. When signal peaks in the OFDM signal are clipped, all subcarriers are affected. Practical systems require a margin of about 8 to 9 dB headroom between the average signal power and the clipping level of the transmit and receive chain. Many papers have been written about the effect of this Peak to Average Power Ratio (PAPR) problem of OFDM, and on methods to mitigate it. In most papers thus far, the high PAPR is seen as a problem for the linearity and the power consumption of the power amplifier at the transmitter. However, in an OFDM broadcast system, the transmitter can be built according to professional standards with sufficient power available, while the receiver may need to be integrated in a cellular mobile phone or other handheld device with only limited battery capacity. This motivated us to focus on the handheld receiver, where the ADC increasingly becomes a main drain of battery power. The power consumption of typical analog–to–digital converters (ADC) is proportional to the number of quantisation steps L, so it is exponential in the ADC resolution b, expressed in bits ($L = 2^b$). To minimize power consumption, we propose to allow a modest degree of clipping of the ADC and to compensate it by digital post–processing.

The results are believed to be relevant for low–power design of OFDM receivers, for instance to prolong battery life during digital television reception on mobile phones.

The remainder of this paper is organized as follows. Section 2 discusses the concept. Section 3 gives a baseband model of an OFDM communication system and a statistical model of the OFDM signal at the receiver, including the receive ADC. Then, Section 4 describes our clipping correction algorithms and evaluates their performance for an idealized DVB-H system. Section 5 explores whether the algorithms can coexist with Doppler compensation algorithms.

2. INTUITIVE DESCRIPTION

In contrast to [1–3], we exploit the empty subcarriers. Digital TV standards such as DVB-T and DVB-H use empty, i.e. unmodulated, subcarriers at the outskirts of their spectrum. We pursue the idea that the clipping induces spurious components at these empty subcarriers and these can assist the receiver to reconstruct the original, i.e., not clipped signal. Figure 1 illustrates that a (time–domain) signal clipped on one input sample transforms into a complex exponential at the FFT output.

Figure 2 gives a block diagram of a simplified compensation algorithm which correlates the complex exponentials to estimate the amount of clipping that occurred.

Our simulations show that a significant reduction of the headroom of the ADC is possible. This algorithm delivers about 2 dB more headroom for the signal power at an SNR

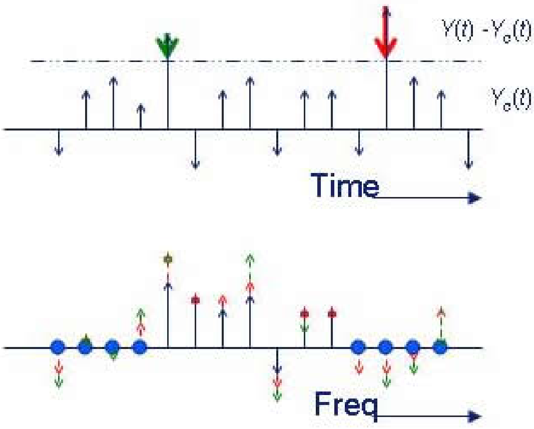

Figure 1: *Time and frequency–domain signal, at ADC output and at FFT output, respectively.*

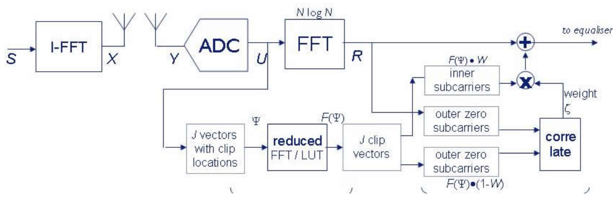

Figure 2: *Block Diagram of OFDM receiver with ADC clip restoration*

of 30 dB, but its effectiveness is reduced if many samples are clipped. We also propose a more elaborate algorithm that gives even more headroom.

The idea to exploit spurious signals received at empty subcarriers has been considered for various other applications. Wolf [4] showed that man-made burst noise can be mitigated. Tureli et al. [5] and Wu et al. [6] studied a method to refine the frequency synchronization by observing empty subcarriers. Our algorithm can be combined with theirs in a straightforward manner, resulting in a multi-dimensional optimisation problem.

3. MODEL OF THE COMMUNICATION SYSTEM

Digital Video Broadcasting uses OFDM transmission with N_s = 2048 or 8192 subcarriers. The subcarriers are indexed from $-N_\mathrm{s}/2$ to $N_\mathrm{s}/2-1$, such that the m-th subcarrier has frequency mf_s, where f_s is the subcarrier spacing. We represent the user data as N_s complex numbers $\underline{S} = (S_{-N_\mathrm{s}/2}, \ldots, S_{N_\mathrm{s}/2-1})$ where subcarrier m carries the data symbol S_m. In DVB and in most other OFDM standards, only a subset of the subcarriers are modulated, while the lowermost M_L and the uppermost M_R subcarriers are unmodulated, so the corresponding $S_m = 0$. For the 2K-mode of DVB-T, $N_\mathrm{s} = 2048$, $M_L = 172$ and $M_R = 171$. The set of empty subcarriers is denoted by $\mathcal{M}_0$, its size $|\mathcal{M}_0| = M_L + M_R$ is denoted by M. We consider an OFDM transmit symbol that lasts from $t = 0$ to $T_\text{symbol} = 1/f_\mathrm{s}$, to ensure orthogonality between the subcarriers. Yet, the transmitter adds a cyclic prefix of duration T_guard, so the transmitted signal can be written in complex baseband representation as

$$s(t) = \sum_{m=-N_\mathrm{s}/2}^{N_\mathrm{s}/2-1} S_m e^{2\pi i m f_\mathrm{s} t}, \quad t \in [-T_\text{guard}, T_\text{symbol}). \tag{1}$$

Additive noise from the receiver's low-noise amplifier and interference from adjacent channel users is denoted by η. The receiver ignores the signal received in the interval $[-T_\text{guard}, 0)$, since it suffers from inter-symbol interference, and samples the signal in the interval $[0, T_\text{symbol})$ at the ADC. This signal is $x(t) = \eta(t) + r(t)$ with

$$r(t) = \sum_{m=-N_\mathrm{s}/2}^{N_\mathrm{s}/2-1} R_m e^{2\pi i m f_\mathrm{s} t}, \quad t \in [0, T_\text{symbol}), \tag{2}$$

with $R_m = \sum_{m'} H_{mm'} S_{m'}$ and H the channel matrix in the frequency domain. Assuming sampling at the Nyquist frequency, $T_\text{sample} = (N_\mathrm{s} f_\mathrm{s})^{-1}$ and the n-th sample is $x_n = x(nT_\text{sample}) = r_n + \eta_n$, $n = 0, \ldots, N_\mathrm{s} - 1$. The set of all N_s samples is denoted by $\underline{x} = (x_0, \ldots, x_{N_\mathrm{s}-1})$. The in-phase and quadrature components of each sample x_n pass through separate (but identical) ADCs. Each ADC discretises its input w and outputs $q(w)$, the best approximation in a discrete set with L different quantisation levels, $D = \{d_1, \ldots, d_L\} \subset \mathbb{R}$ with $d_1 < d_2 < \ldots < d_L$, according to

$$q(w) = \arg\min_{d \in D} |d - w|, \tag{3}$$

so the function q is completely determined by D. If the ADC input w lies between d_k and d_{k+1}, the discretisation error $|q(w) - w|$ is at most equal to $(d_{k+1} - d_k)/2$, but if the input sample $w > d_L$, or $w < d_1$ the ADC is said to *clip* the sample. When clipping occurs, the discretisation error $|q(w) - w|$ can be large. We consider the case of a large L and finite d_1 and d_L, such that the error introduced by the quantizer is dominated by clipping, so that

$$q(w) = \begin{cases} -C & \text{if } w \le -C \\ w & \text{if } -C < w < C \\ C & \text{if } w \ge C. \end{cases} \tag{4}$$

The complex baseband representation of the sample after the ADC is

$$y_n = q(\mathrm{Re}(x_n)) + \mathrm{i}q(\mathrm{Im}(x_n)) =: Q(x_n), \tag{5}$$

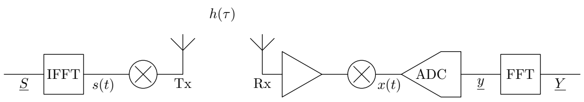

Figure 3: *A diagram showing the system model.*

or, in vector notation, as in Fig. 3,

$$\underline{y} = Q(\underline{x}), \tag{6}$$

where $\underline{y} = (y_0, \ldots, y_{N_s-1})$. The discretized samples are Fourier transformed, which yields the received signal on the subcarriers:

$$Y_m = \frac{1}{N_s} \sum_{n=0}^{N_s-1} y_n e^{-2\pi i m n / N_s}, \qquad m = -N_s/2, \ldots, N_s/2 - 1. \tag{7}$$

If the ADC would have no distortion, i.e., $q(w) = w$ for all w, and if there would be no noise, $\eta(t) \equiv 0$, then $Y_m = R_m$.

4. COMPENSATION ALGORITHMS

In this section we describe three methods to reconstruct a clipped signal.

The first method involves strict equation-solving. The second one is easy to implement, because it estimates the degree of clipping per sample individually, but it ignores correlation between the artifacts if more than one sample is clipped. Hence it results in an approximate solution if multiple samples are clipped simultaneously. The third method takes an MMSE approach and exploits the statistical properties of the signal and noise to reconstruct the expected received signal before clipping. The model is worked out for signals and noise with Gaussian distributions.

4.1. Compensation algorithm for noise-free channel

Initially we assume a noise-free channel. So we can write $r_n = y_n + c_n$ with

$$\begin{aligned} \mathrm{Re}(c_n) &= 0 \text{ if } |\mathrm{Re}(y_n)| < C, \\ \mathrm{Im}(c_n) &= 0 \text{ if } |\mathrm{Im}(y_n)| < C. \end{aligned} \tag{8}$$

Denoting the number of clippings by N_c, there are N_c non-zero parameters, one for each clipped real or imaginary part of a sample.

After the Fourier transform, we have that $R_m = Y_m + C_m$. Since the real and imaginary parts of R_m are zero if m corresponds to an empty subcarrier, we have that $C_m = -Y_m$ for $m \in \mathcal{M}_0$. These equations are equivalent to $2M$ equations for N_c unknown parameters, which can be written in matrix form as

$$v = Au, \tag{9}$$

where v is a known $2M$-vector with the real and imaginary parts of $-Y_m$ for the subcarriers where $R_m = 0$, u is an N_c-vector with the unknown values of $\mathrm{Re}(c_n)$ or $\mathrm{Im}(c_n)$ at a clipped sample and A is an $2M \times N_c$-matrix of which the elements are the appropriate real and imaginary parts of the Fourier-transform matrix. If $N_c \leq 2M$, u can be uniquely determined by solving any subset of N_c equations, or by

$$u = (A^T A)^{-1} A^T v. \tag{10}$$

If $N_c > 2M$, $A^T A$ has $N_c - 2M$ eigenvalues equal to zero, so u cannot be determined uniquely and this equation-solving clip correction algorithm breaks down.

Wolf [4] proposed this algorithm to repair the effects of impulsive noise.

4.2. Compensation algorithm for single clips

Each of the diagonal elements of $A^T A$ is equal to M/N_s^2. The off-diagonal elements are *partial* ($M < N$) correlations of an integer number of rotations of complex exponentials, so these are usually non-zero. However, off-diagonal components are smaller than the diagonal elements. Approximating $A^T A$ by its diagonal leads to the simpler approximation

$$\tilde{u} = \frac{(N_s)^2}{M} A^T v. \tag{11}$$

The corresponding clip-corrected complex-valued signal in the frequency domain is given by

$$\tilde{Y} = Y - \frac{N_s}{M} F P_{\text{clip},\underline{y}} F^{-1} P_{\text{empty}} Y, \tag{12}$$

where F is the Fourier transform matrix, F^{-1} its inverse, P_{empty} projects onto the M empty subcarriers and $P_{\text{clip},\underline{y}}$ projects the real and imaginary parts of its argument onto the positions where the corresponding real or imaginary parts of $\underline{y}$ are clipped.

This algorithm is equivalent to making the approximation that the out-of-band clip artifacts u for every clip location are orthogonal, so that clipping can be corrected by inverting the clip artifacts one by one. The effect of noise is not excessive because each of the $2M$ out of band samples is attenuated by $1/M$. Hence the signal–to–noise ratio on compensated, i.e., previously clipped (pre–FFT) samples deteriorates only marginally and the effect at the FFT output is negligible.

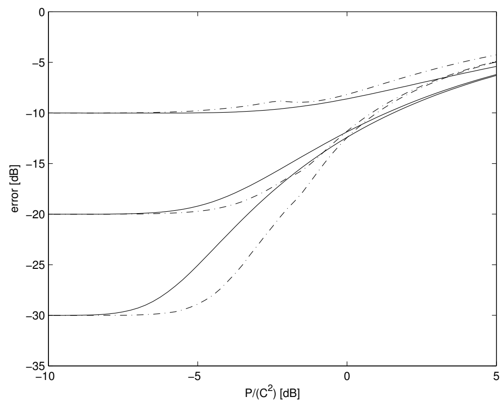

Figure 4: *Average squared error $\sum_m |\hat{R}_m - R_m|^2$ (in dB) versus received signal power P/C^2 (in dB) for SNR = $10\,dB$ (upper curves), $20\,dB$, and $30\,dB$ (lower curves). The average is taken over 100 realizations of signal and noise. Solid curves: $\hat{R}_m = Y_m$, dashed/dotted curves: approximate clip correction using (12).*

4.2.1. Simulation Results

Fig. 4 shows the performance simulated for DVB-H. We consider a system with $N_\text{s} = 2048$ subcarriers, out of which $M = 343$ subcarriers are empty. Then each of the $N_\text{s} - M$ non-empty subcarriers carries a signal $\sqrt{P}\,R_m$ where the R_m are i.i.d. circularly symmetric complex Gaussian distribution with mean zero and variance $1/(N_\text{s}-M)$, and where P is the received signal power. The noise, with power P/SNR, is evenly distributed among all N_s subcarriers, so each subcarrier also carries an i.i.d. complex Gaussian noise signal with variance $P/(N_\text{s}\text{SNR})$. For a given SNR, the effect of clipping depends on the clipping level C only through the ratio P/C^2.

The graphs show that clip correction using (12) gives a gain of about 2 dB when the SNR = 30 dB. Without clip correction, the error starts rising above the noise level due to clipping at $P/C^2 = -7\,\text{dB}$, with clip correction this happens at $P/C^2 = -5\,\text{dB}$, so the receiver's AGC requires 2 dB less back-off. For higher values of P/C^2, clip correction using (12) gives worse results than not using any correction. When SNR = 20 dB, the gain is reduced to about 1 dB, when SNR = 10 dB, there is no improvement at all.

For each clipped signal sample, the simple algorithm calculates or retrieves from a look-up-table N complex values. Real-time calculation would involve 1 complex rotation (multiplication) per value. Out of these, M complex values are used to perform a correlation (M complex multiplications and M complex additions). The outcome of the correlation is then used to scale the other $N - M$ values ($N - M$ complex multiplications) and added to the $N - M$ subcarriers that contain user data ($N - M$ complex additions). Hence the algorithm requires $2N$ complex multiplications per clipped sample.

4.3. MMSE algorithm for Gaussian signal and noise

Recall the sampled signal

$$\underline{y} = Q(\underline{r} + \underline{\eta}).$$

The minimum mean square error (MMSE) estimate of $\underline{r}$ is given by

$$\hat{\underline{r}}(\underline{y}) = \mathbb{E}_{\underline{r},\underline{\eta}}[\underline{r} | Q(\underline{r} + \underline{\eta}) = \underline{y}] \tag{13}$$

where the expectation is calculated with the *a priori* distributions of $\underline{r}$ and $\underline{\eta}$. The distribution of the time domain signals at any sample moment becomes Gaussian if the number of OFDM subcarriers, N_s, goes to infinity, even if each subcarrier signal is non-Gaussian. We therefore make the approximation that $\underline{r} = (r_0, \ldots, r_{N_\text{s}-1})^T$ has a circularly symmetric Gaussian distribution with mean zero, $\mathbb{E}[\underline{r}] = 0$, $\mathbb{E}[\underline{r}\underline{r}^T] = 0$ (the superscript T denotes the transpose), and correlation matrix $\mathbb{E}[\underline{r}\underline{r}^H] =: G_\text{r}$ (the superscript H denotes the hermitean conjugate). Our particular case of an OFDM signal where the signals transmitted on different subcarriers are uncorrelated implies $\underline{r} = F^{-1}\underline{R} = F^{-1}H\underline{S}$, so that $G_\text{r} = F^{-1}G_R(F^{-1})^H$, $G_R = HG_SH^H$, H is the $N_\text{s} \times N_\text{s}$ channel matrix in the frequency domain, and G_S is a diagonal $N_\text{s} \times N_\text{s}$ matrix with $\mathbb{E}[|S_m|^2]$ on the position corresponding to subcarrier m. The diagonal elements of G_S corresponding to empty subcarriers are equal to zero. We assume that the noise is Gaussian as well, with correlation matrix G_n. It can be shown (see [7]) that

$$\hat{\underline{r}}(\underline{y}) = G_\text{r}(G_\text{r} + G_\text{n})^{-1} \frac{\int_{Q^{-1}(\underline{y})} \underline{x}\, p_\text{x}(\underline{x})\, d\underline{x}}{\int_{Q^{-1}(\underline{y})} p_\text{x}(\underline{x})\, d\underline{x}}, \tag{14}$$

where $Q^{-1}(\underline{y})$ is the set of all signals $\underline{x}$ for which $Q(\underline{x}) = \underline{y}$, so $Q^{-1}(\underline{y})$ is semi-infinite space with dimension equal to the number of clippings, where clippings in the real and imaginary parts are counted separately. The distribution p_x is Gaussian with correlation matrix $G_\text{r} + G_\text{n}$. We approximate the mean of $\underline{x}$ over $Q^{-1}(\underline{y})$ by the point in $Q^{-1}(\underline{y})$ where p_x assumes the largest value, so

$$\hat{\underline{r}}(\underline{y}) \approx G_\text{r}(G_\text{r} + G_\text{n})^{-1} \arg \min_{\underline{x} \in Q^{-1}(\underline{y})} \underline{x}^H (G_\text{r} + G_\text{n})^{-1} \underline{x}. \tag{15}$$

Locating the minimum of $\underline{x}^H(G_\text{r} + G_\text{n})^{-1}\underline{x}$ over $Q^{-1}(\underline{y})$ amounts to solving a quadratic minimisation problem with linear constraints. There exists a vast literature on solving such problems, see e.g. [8].

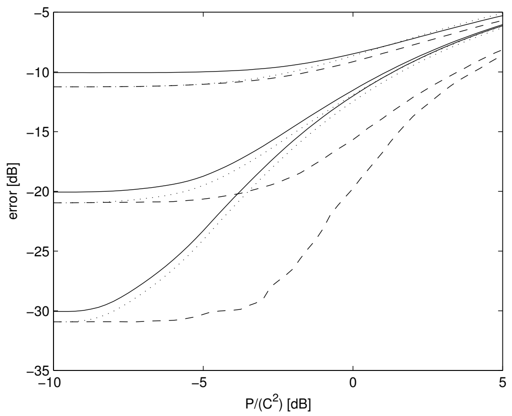

Figure 5: *Squared error $\sum_m |\hat{R}_m - R_m|^2$ (in dB) versus received signal power P/C^2 (in dB) for SNR $= 10\,dB$ (upper curves), $20\,dB$ (middle curves) and $30\,dB$ (lower curves). Solid curves: $\hat{R}_m = Y_m$, dotted curves: $\hat{R}_m = G_R(G_R + G_N)^{-1}Y_m$, dashed curves: full clip correction.*

4.3.1. Simulation Results

We used the same DVB-H system as in Section 4.2.1, but performed the clip correction according the more elaborate algorithm developed in Section 4.3. Since the correction algorithm has two steps, the solution of the minimisation problem and the MMSE correction with the matrix $G_\text{r}(G_\text{r} + G_\text{n})^{-1}$, we also showed the effect of the MMSE correction alone. The results are shown in Fig. 5, for SNR-values of 10 dB, 20 dB and 30 dB and a frequency-flat channel $H = I_{N_\text{s}}$. Not surprisingly, this algorithm performs better than our single clip correction algorithm: when the SNR $= 30\,$dB, the back-off gain is about $5\,$dB, at SNR $= 20\,$dB it is about $3\,$dB and at SNR $= 10\,$dB it is less than $1\,$dB. Note that the full algorithm always always give better results than not using any correction at all and that the MMSE correction alone gives minor gains.

5. DOPPLER CHANNELS

Mobile reception and frequency offsets cause crosstalk among subcarriers and spurious signals on empty subcarriers, see e.g., [9]. Following the model from [10] the channel matrix in the frequency domain can be written as

$$H = H_0 + (f_\text{Doppler}/f_s)\Xi H_1. \tag{16}$$

Here the Doppler spread $f_\text{Doppler} = v f_c/c$ with v the speed of the receiver, f_c the carrier frequency and c the speed of light; Ξ is the frequency domain crosstalk matrix, with off-diagonal elements

$$\begin{aligned}\Xi_{m,n} &= [N_s(1 - \exp(2\pi\text{i}(n-m)/N_s))]^{-1}\\ &\approx 1/(m-n),\end{aligned}$$

and H_0 and H_1 are diagonal matrices. In terms of a scattering model with L scatterers we can write

$$H_0(m) = \frac{1}{\sqrt{L}} \sum_{\ell=1}^{L} h_\ell \exp[-2\pi\text{i} m f_s \tau_\ell] \tag{17}$$

$$H_1(m) = \frac{1}{\sqrt{L}} \sum_{\ell=1}^{L} h_\ell \cos(\theta_\ell) \exp[-2\pi\text{i} m f_s \tau_\ell], \tag{18}$$

where the channel amplitudes h_ℓ are i.i.d. $CN(0,1)$, the delays τ_ℓ are i.i.d. exponentially distributed with mean T_rms and the angles of incidence θ_ℓ are i.i.d. uniformly distributed in $[0, 2\pi)$. In the rich scattering limit ($L \to \infty$) the $H_0(m)$ and $H_1(m)$ become Gaussian with the following correlations [10, 11]:

$$\mathbb{E}[H_0(m)H_0^*(m')] = \frac{1}{1 + 2\pi\text{i}(m-m')f_s T_\text{rms}}, \tag{19}$$

$$\mathbb{E}[H_1(m)H_1^*(m')] = \frac{1/2}{1 + 2\pi\text{i}(m-m')f_s T_\text{rms}}, \tag{20}$$

$$\mathbb{E}[H_0(m)H_1^*(m')] = 0. \tag{21}$$

Ideally, the receiver would estimate both H_0 and H_1 and perform an estimate of S using the complete channel matrix, including the off-diagonal elements, i.e., the MMSE algorithm with

$$\begin{aligned}G_R &= P_0 H (I_{N_\text{s}} - P_\text{empty}) H^H\\ G_N &= \sigma^2 I_{N_\text{s}},\end{aligned}$$

where P_0 is the signal power per non-empty subcarrier and σ^2 is the noise power per subcarrier.
Alternatively, the receiver may choose to simplify the estimation of S_m by treating the inter-carrier interference as additional Gaussian noise, i.e.,

$$\begin{aligned}G_R &= P_0 H_0 (I_{N_\text{s}} - P_\text{empty}) H_0^H\\ G_N &= \sigma^2 I_{N_\text{s}} + P_0 \left(\frac{f_\text{Doppler}}{f_s}\right)^2 \Xi H_1 (1_{N_\text{s}} - P_\text{empty}) H_1^H \Xi^H.\end{aligned}$$

A third option for the receiver is to not estimate H_1 but to average over H_1 in the noise correlation matrix, i.e.,

$$\begin{aligned}G_R &= P_0 H_0 (I_{N_\text{s}} - P_\text{empty}) H_0^H\\ G_N &= \sigma^2 I_{N_\text{s}} + \frac{P_0}{2} \left(\frac{f_\text{Doppler}}{f_s}\right)^2 \Xi (1_{N_\text{s}} - P_\text{empty}) \Xi^H.\end{aligned}$$

In this case the effective noise power on subcarrier m with $|m| \ll (N-M)/2$ is increased from σ^2 to

$$(G_N)_{mm} \approx \sigma^2 + P_0 \left(\frac{f_\text{Doppler}}{f_s}\right)^2 \frac{\pi^2}{6}.$$

This extra noise leads to a lower SNR and hence a worse performance of the clip correction algorithms.

6. CONCLUSION

We proposed an algorithm to compensation clipping. It allows for a significant reduction in power consumption of the A/D converter by reducing the margin needed between the average received power of an OFDM signal and the clip level of the ADC of an OFDM receiver.

The advantage of a clip compensation algorithm is that the design specifications of the ADC can be relaxed. We found that for high SNR the AGC back-off or clipping threshold can be reduced by 2 dB when using the algorithm for single clips, so that one may use only 79% of the number of quantization steps needed hitherto. This corresponds to a reduction of power consumption by the ADC by 21%. However our algorithm involves a number of digital signal processing operations. A more involved MMSE clip correction algorithm performs even better, at the expense of even more digital processing.

Typically, in many designs it appeared favorable to trade analog operations, such as long-tailed transistor pairs in ADC comparator circuits for digital operations in the baseband processor.

A practical implementation of an OFDM receiver requires analog and digital bandpass filters. Our algorithms can operate without modification if analog filters are applied before the ADC. In fact, strict removal of channel noise and interference on the empty subcarriers contributes to the reliable operation of the algorithm. However digital filtering between the ADC and the FFT can reduce the ability of the algorithm to compensate clipped samples. Preferably digital filtering is performed by over-dimensioning the FFT.

ICI caused by Doppler deteriorates the performance of the clip correction algorithms that treat it as noise. A clip correction algorithm that uses the full channel knowledge can still improve the signal quality. We leave it as a suggestion for future work to integrate the algorithms for clip correction, man made noise mitigation, frequency tracking, phase noise compensation and full channel estimation, including the Doppler effect.

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