Optimal Output Back-Off in OFDM Systems with Nonlinear Power Amplifiers

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Abstract—A well-known problem of OFDM signals is their high peak-to-average power ratio, which can cause the clipping of peaks in the transmitted signal due to the limited linear range of high power amplifiers. Operating the amplifier solely in the linear region, i.e., with a large output back-off, results in an undistorted transmitted signal, but at the expense of a low SNR at the receiver. Driving the amplifier into saturation, on the other hand, maximizes the received SNR, but causes severe nonlinear signal distortions. In this paper we study the resulting trade-off. We show both theoretically and by simulations that, despite the distortions, it is optimal to operate the amplifier close to saturation. Furthermore, we analytically derive a modification of the Bussgang noise cancellation algorithm which was recently shown in the literature to improve the EXIT characteristic of the algorithm.

I. INTRODUCTION

The OFDM technique has received a lot of interest in the literature because it allows a spectrally efficient transmission over frequency-selective broadband channels while still allowing for low-complexity receiver implementations. During the past decade, OFDM has become the basis of many standards, for example IEEE 802.11a, IEEE 802.16, and 3GPP LTE, the designated successor of UMTS.

However, OFDM also suffers from some drawbacks. One important problem is the high peak-to-average power ratio (PAPR) of the transmitted signal which originates from the superposition of many independent subcarriers. The resulting power peaks can drive the high power amplifier (HPA) into saturation which leads to nonlinear signal distortions. Many different techniques have been proposed that aim at reducing the PAPR on the transmitter side (c.f. [1] for a literature overview). An alternative approach, which is used in this paper, is to clip the signal deliberately, accepting the resulting impairments, and to rely on receiver-side techniques instead to mitigate the distortion [2], [3].

While it is possible to circumvent the PAPR problem altogether by operating the amplifier solely in the linear range, this comes at the expense of a low signal-to-noise ratio (SNR) at the receiver or an overdimensioned and thus inefficient HPA. Operating the amplifier close to saturation, on the other hand, leads to a higher SNR but also to severe signal distortions. The resulting trade-off is the topic of the present work.

The ratio between the maximum possible transmit power and the actual mean transmit power is called the output back-off (OBO) of the transmitter. In this paper we investigate the impact of the OBO on the system performance and determine its optimal value. We use two different criteria for performance; firstly, the mutual information between channel input and output, and secondly, the frame error rate that is obtained by a practical receiver algorithm, the so-called Bussgang Noise Cancellation (BNC) algorithm. We furthermore derive a modification of the BNC algorithm which has, in a slightly different form, recently been proposed in an ad-hoc manner in the literature, motivated by means of the EXIT chart technique.

The remainder of this paper is organized as follows. Section II introduces our model of the system in general and the nonlinear amplifier in particular. Section III contains an information-theoretic analysis of the problem, followed by a study of a practical receiver using the BNC algorithm in Section IV. Finally, Section V concludes the paper.

We use the following notation: Upper and lower case letters denote frequency and time domain quantities, respectively. Vectors are written in bold face. E[·] denotes statistical expectation. We do not explicitly distinguish between random variables and their realizations; the meaning will always be clear from the context.

II. SYSTEM MODEL

A. Transmitter and Channel Model

We consider a coded OFDM baseband system as depicted in Fig. 1. Information bits \( u \) are encoded with a channel code of rate \( r \), and the resulting code bits are interleaved by a pseudorandom channel interleaver. Groups of \( m \) code bits are mapped to symbols \( X_k \in \mathcal{X} \), where \( \mathcal{X} \subset \mathbb{C} \) is the \( 2^m \)-ary modulation alphabet. Then, symbol blocks \( X = [X_0, \ldots, X_{K-1}] \) are modulated via an \( LK \)-point IFFT operation, yielding the time-domain samples \( x = [x_0, \ldots, x_{LK-1}] \), and a cyclic prefix of \( LN_{cp} \) samples is inserted. Here, \( K \) is the number of subcarriers, and \( L \) denotes the oversampling factor.

The resulting signal is fed through a nonlinearity \( g(x) \), yielding the clipped signal \( x^c \). If deliberate clipping of the discrete baseband signal is employed in order to reduce the PAPR [2], \( g \) is applied to the samples \( x_n \). However, \( g \) could
also model the nonlinear characteristic of the amplifier, in which case the input to \( g \) is the signal \( x(t) \) after D/A-conversion. The nonlinearity will be discussed in detail in Section II-B.

The nonlinear distortion leads to in-band and out-of-band noise (unless the clipping is performed on the critically sampled signal, \( L = 1 \), which results in in-band distortion only). In order to remove the out-of-band signal components, the clipped signal \( x^c \) is filtered by a low-pass filter (LPF), and the output \( x^f \) is transmitted over a frequency-selective fading channel.

In this paper, we assume perfect time and frequency synchronization of the receiver, and a maximum channel delay spread that is shorter than the cyclic prefix duration. Under these assumptions, the received signal is free of intersymbol interference and, in the context of this paper, it suffices to study the transmission of a single OFDM symbol. Furthermore, we assume perfect channel knowledge at the receiver, and a quasi-static fading channel which can be considered as constant during one symbol duration. Then, the sampled received signal can be written as

\[
y_n = h_n \ast x_n^f + w_n,
\]

where \( \ast \) denotes convolution, \( w_n \) is AWGN with variance \( \sigma_w^2 \), and \( h_n \) is the impulse response of the fading channel, which has an average gain of \( \sigma_h^2 = E[\sum_n |h_n|^2] \). The processes \( x, h, \) and \( w \) are mutually independent.

The receiver removes the cyclic prefix and performs an FFT on the received blocks. The resulting frequency-domain signal model can be described in vector notation as

\[
Y = H \odot X^f + W,
\]

where the vector elements correspond to the data-carrying subchannels. In (2), \( \odot \) denotes the elementwise vector product.

B. Characterization of the Nonlinearity

Due to their large bandwidth, amplifiers like traveling-wave tubes and solid state power amplifiers can be modeled as memoryless nonlinearities \( g : \mathbb{C} \rightarrow \mathbb{C} \) acting on the complex baseband signal. The effect of \( g \) is a distortion of the signal amplitude, and potentially also a phase shift. Since both effects only depend on the amplitude and not on the phase of the input signal, \( g(x) \) can be expressed as

\[
x^c = g(x) = \frac{x}{|x|} G(|x|) e^{j\phi(|x|)},
\]

where \( G \) and \( \Phi \) are the AM/AM and AM/PM characteristics. We apply two normalizations to the considered nonlinearities:

- In order to abstract from the amplification itself, we require \( g(x) \) to have a small-signal gain of one:
  \[
  \lim_{x \to 0} \frac{g(x)}{x} = 1.
  \]
- We normalize the maximum output amplitude \( A \), and hence the maximum output power, to one:
  \[
  \sup_{x \in \mathbb{C}} |g(x)| = 1.
  \]

Two particular nonlinearities are examined in the following:

- A soft limiter (SL) with the characteristic
  \[
  G_{sl}(|x|) = \min\{|x|, 1\}, \quad \Phi_{sl}(|x|) = 0
  \]
  which is well suited to model digital clipping.
- A traveling-wave tube (TWT) according to the Saleh-model [4]. The parameters of this model are chosen to meet the normalizations (4), (5), and to match a measured TWT characteristic, resulting in the model
  \[
  G_{twt}(|x|) = \frac{|x|}{1 + |x|^2/4}, \quad \Phi_{twt}(|x|) = -\frac{\pi}{3} \frac{|x|^2}{|x|^2 + 5/2}.
  \]

C. The Bussgang Decomposition

The time-domain OFDM signal \( x \) consists of a superposition of many independent subcarriers. According to the central limit theorem, \( x \) can be approximated as a circularly symmetric zero-mean Gaussian process with variance \( \sigma_x^2 \). An important tool for the analysis of nonlinearities driven by Gaussian processes is the Bussgang decomposition [5]–[7], which we will briefly review in this section.

The Bussgang theorem states that the output of a memoryless nonlinearity \( g \), driven by a zero-mean Gaussian process \( x \), can be decomposed into

\[
x_n^c = g(x_n) = \alpha x_n + d_n
\]

with \( E[x_n d_n^*] = 0 \). The difference \( d_n \) between the output and the linearly scaled input will be called Bussgang noise in the following.

The scaling factor \( \alpha \) depends on the input variance and is given as

\[
\alpha = \frac{E[|x|^2]}{E[|x|^2]^2}
\]

\[
= \frac{2}{\sigma_x^2} \int_0^\infty \xi^2 \cdot g(\xi) e^{-\frac{\xi^2}{\sigma_x^2}} d\xi.
\]

The signal power is affected in two ways by the nonlinearity. Firstly, the clipping operation scales the signal power by a factor of

\[
\beta^c = \frac{E[|x|^2]}{E[|x|^2]}
\]

\[
= \frac{2}{\sigma_x^2} \int_0^\infty \xi^2 \cdot G^2(\xi) e^{-\frac{\xi^2}{\sigma_x^2}} d\xi.
\]
Secondly, the low-pass filter removes the out-of-band components of the Bussgang noise \(d\), yielding a further power loss of \(\beta^f = \frac{E[|x|^2]}{E[|x|^2]}\), which can be calculated numerically \([8]\). The total power loss is thus given as

\[
\beta = \beta^f \beta^l = \frac{E[|x|^2]}{E[|x|^2]}. \tag{13}
\]

For the soft limiter, the integrals (10) and (12) can be solved analytically:

\[
\alpha_{sl} = 1 - \exp \left(-\frac{1}{\sigma_x^2} + \frac{\sqrt{\pi}}{2\sigma_x} \text{erfc}\left(\frac{1}{\sigma_x}\right)\right) \tag{14}
\]

\[
\beta_{sl} = 1 - \exp \left(-\frac{1}{\sigma_x^2}\right). \tag{15}
\]

Denoting the filtered Bussgang noise as \(d^f\), the frequency domain signal model (2) becomes

\[
Y = H \odot (\alpha X + D^f) + W. \tag{16}
\]

The output back-off (OBO) of the transmitter is defined as the ratio between the maximum possible output power of the HPA and the mean transmit power,

\[
\text{OBO} = \frac{A^2}{E[|x|^2]} = \frac{1}{\beta \sigma_x^2}. \tag{17}
\]

With this definition, the received SNR becomes

\[
\text{SNR} = \frac{\sigma_h^2}{\sigma_w^2} = \frac{\sigma_h^2}{\sigma_w^2} \frac{1}{\text{OBO}}, \tag{18}
\]

and the ratio of bit energy to noise power spectral density is

\[
\frac{E_b}{N_0} = \frac{K + N_{cp}}{r \cdot m \cdot K} \frac{\sigma_h^2}{\sigma_w^2} \frac{1}{\text{OBO}}. \tag{19}
\]

For fixed transmission parameters, \((K + N_{cp})/(r \cdot m \cdot K)\) is constant. Thus for a given fraction \(\sigma_h^2/\sigma_w^2\), which is determined by the physical scenario, \(E_b/N_0\) is inversely proportional to the OBO. However, a small OBO leads to severe clipping distortions. The resulting trade-off is the motivation for this study.

### III. INFORMATION-THEORETIC ANALYSIS

In this section, we examine the impact of the OBO on the mutual information between channel input and output. We refer to this quantity as \textit{constrained capacity}, because we do not maximize the mutual information with respect to the input distribution, but rather assume it to be Gaussian.

To the best of our knowledge, a precise evaluation of the mutual information of the channel model (1) is computationally infeasible due to the memory introduced by the LPF

\[
\text{\textbf{Fig. 2. Simplified memoryless channel model}}
\]

and the frequency-selective channel. We therefore base this analysis on the simplified memoryless channel model

\[
y_n = g(x_n) + \tilde{w}_n \tag{20}
\]

as depicted in Fig. 2. Both the channel input \(x\) and the additive noise \(\tilde{w}\) are modeled as stationary zero-mean white Gaussian processes with variance \(\sigma_h^2\) and \(\sigma_w^2 = \sigma_h^2/\sigma_w^2\), respectively. (We have introduced the effective noise process \(\tilde{w}\) with scaled variance in order to keep the notation within this paper consistent.)

The main difference between the channel model (20) and the true time-continuous channel is that the model assumes independent input samples. As discussed in Section II-A, the time-continuous nonlinearity leads to spectral sidelobes which are subsequently suppressed by a low-pass filter, leading to a loss of information. Since this effect is not captured by the simplified model, the following analysis gives only an upper bound on the true constrained capacity. However, we expect this upper bound to be rather tight, because numerical evaluations show that the out-of-band power is small in comparison to the total signal power.

Using the circular symmetry of the channel output \(y\), it has been shown in \([9]\) that the constrained capacity of the channel (20) in [bit / channel use] can be computed as

\[
I(x; y) = -\int_0^\infty p_{|y|}(\eta) \log_2 \frac{p_{|y|}(\eta)}{2\pi\eta} d\eta - \log_2(\pi\sigma_w^2), \tag{21}
\]

where \(p_{|y|}(\eta)\), the pdf of \(|y|\), is given as

\[
p_{|y|}(\eta) = \frac{4\eta}{\sigma_h^2\sigma_w^2} \int_0^\infty \xi \exp \left(-\frac{\xi^2}{2\sigma_h^2} - \frac{\eta^2 + G(\xi)^2}{\sigma_w^2}\right) \times I_0 \left(\frac{2\eta G(\xi)}{\sigma_w^2}\right) d\xi. \tag{22}
\]

Here, \(I_0(z)\) is the modified Bessel function of the first kind.

Fig. 3 shows the constrained capacity as a function of the OBO, where the input power \(\sigma_x^2\) in the numerical evaluation
of (22) was chosen according to (17). Note that the traveling-wave tube according to model (7) cannot achieve an output back-off of less than about 1.1 dB under the constraint of Gaussian input. This is due to the fact that the output amplitude does not saturate for large \(|x|\), but rather reaches a maximum at \(|x| = 2\) and then decreases towards zero for \(|x| \to \infty\).

It can be observed that for both nonlinearities, the plots for \(\frac{\sigma^2_h}{\sigma^2_w} = 0\) dB and 5 dB reach their maxima at the minimal possible OBO. At \(\frac{\sigma^2_h}{\sigma^2_w}\) of 10 dB and 15 dB, the constrained capacity is maximized by larger output back-off factors, because the clipping distortion becomes the dominant source of signal corruption. But even in this SNR range, the OBO which maximizes the constrained capacity is only about 1 dB above the minimum possible OBO.

The analysis in this section has shown that, from an information-theoretic point of view, it is best to operate nonlinear power amplifiers in an OFDM system close to saturation. It is obvious that in a practical system, the receiver must be aware of the clipping distortion and employ suitable equalization algorithms. In the next section, we will study the performance of one particular receiver algorithm as a function of the OBO.

IV. PERFORMANCE OF THE SOFT BNC ALGORITHM

Recall the signal model (16)

\[
Y = H \odot (\alpha X + D^{(i)}) + W
\]

(23)

where the vector elements correspond to the data-carrying subchannels. Since \(D\) and \(D^{(i)}\) differ only in the out-of-band components, we drop the index \((\cdot)^{(i)}\) in the following for simplicity.

Assuming that the information sequences are equally probable a priori, the receiver attempts to find the ML data estimate

\[
\hat{u} = \arg \max_u p(Y|u).
\]

(24)

Due to the nonlinearity, however, finding the exact solution to (24) is computationally infeasible, and one has to resort to suboptimal algorithms. Following the principle of Turbo equalization [10], the state-of-the-art approach is to employ an equalizer that is able to exploit prior information about the symbols. The equalizer and the soft-in soft-out (SISO) channel decoder are scheduled iteratively for several times, exchanging soft information during the process.

Two algorithms for the mitigation of clipping distortion that have gained a lot of interest in the literature are Decision-Aided Reconstruction (DAR) [11] and Bussgang Noise Cancellation (BNC) [12]. A comparison in [13] has recently shown that the latter is the more promising choice. Therefore, we use the BNC algorithm in this study.

The principle of Bussgang noise cancellation is to estimate the term \(D^{(i)}\) in (23) and subtract it from the received signal. A block diagram of the whole baseband receiver is depicted in Fig. 4, where the superscript \((\cdot)^{(i)}\) denotes the iteration index.

![Block diagram of the iterative soft BNC algorithm](image)

These are processed by a SISO channel decoder, resulting in the posterior LLRs \(\lambda^{(i)}_{\text{post}}\). The extrinsic information \(\lambda^{(i)}_{\text{extr}} = \lambda^{(i)}_{\text{post}} - \lambda^{(i)}_{\text{pri}}\) is fed to a soft mapper, which computes the conditional mean of the data symbols (often called soft symbols in the literature)

\[
\hat{X}^{(i)} = E[X|\lambda^{(i)}_{\text{extr}}],
\]

(25)

serving as the input to the Bussgang noise estimator.

A. An Upper Bound on the BNC Performance

Since the BNC algorithm discards information by subtracting the distortion term \(D\) (which is a deterministic function of \(X\)), we anticipate that this approach is unable to reach capacity. Therefore, we investigate the possible performance of the BNC approach by giving an upper bound on the achievable data rates, before turning to the estimation algorithm itself.

Clearly, the best performance that the BNC algorithm can possibly achieve occurs when the Bussgang noise estimate converges to the true value, \(\hat{D}^{(\infty)} = D\). Then, the remaining signal is

\[
\hat{Y}^{(\infty)} = \alpha H \odot X + W.
\]

(26)

It is well known that the capacity for this channel is achieved by a complex Gaussian input distribution. In the AWGN case \((H_k = \sigma_h, \forall k)\), the mutual information \(I(X;\hat{Y}^{(\infty)})\) is simply given as

\[
I(X;\hat{Y}^{(\infty)}) = \log_2 \left( 1 + \frac{|\alpha|^2 \sigma^2_h \sigma^2_w}{\sigma^2_w} \right)
= \log_2 \left( 1 + \frac{|\alpha|^2 \sigma^2_h}{\beta \sigma^2_w} \frac{1}{\text{OBO}} \right),
\]

(27)

which is plotted in Fig. 5 for the same parameters as the constrained capacity in Fig. 3. It can be seen that the plots in both figures are essentially identical over a wide parameter range. While (26) is indeed only an upper bound on the achievable data rate (note, for example, that the peak power constraint that is introduced by the nonlinearity is not present anymore in (26)), this examination at least indicates that the BNC approach has the potential to operate close to the constrained channel capacity.

B. Estimation of the Bussgang Noise

The main part of the iterative BNC algorithm has not been discussed yet, namely the estimation of the Bussgang noise term \(D\). Since the true Bussgang noise in time domain is
denote as \( d = g(x) - \alpha x \), a natural approach to estimate \( d \) is\(^1\) \[12\], \[13\]

\[
\hat{d} = g(\hat{x}) - \hat{\alpha} \hat{x},
\]

where \( \hat{x} \) is the IFFT of the soft symbols \( \hat{X} \). In the following, we derive a modification of this estimator which improves the performance of the BNC algorithm.

We assume that the structure of the estimator is given as\(^2\)

\[
\hat{d} = g(\hat{x}) - \hat{\alpha} \hat{x},
\]

and we attempt to choose \( \hat{\alpha} \) such that the remaining Bussgang noise power is minimized, given the information \( \lambda_{\text{extr}} \) from the decoder. Denoting \( \hat{x}^c = g(\hat{x}) \), we find

\[
\hat{\alpha} = \arg \min_{\alpha} \mathbb{E} \left[ \| \hat{d} - \hat{d}^* \| \|^2 \left| \lambda_{\text{extr}} \right\|^\alpha \right]
\]

\[
= \arg \min_{\alpha} \mathbb{E} \left[ \| \hat{d}^H \hat{d} - \hat{d}^* \hat{d} - \hat{d}^H \hat{d} \| \lambda_{\text{extr}} \right]
\]

\[
= \arg \min_{\alpha} \hat{d}^H \hat{d} \mathbb{E} [ d \lambda_{\text{extr}} ] - \mathbb{E} [ \hat{d}^H \lambda_{\text{extr}} ] \hat{d}
\]

\[
= \arg \min_{\alpha} \| \hat{x}^c \|^2 + \| \alpha \|^2 \| x \|^2 - \alpha^* \hat{x}^c \hat{x} - \alpha \hat{x}^c \hat{x}
\]

\[
= : \arg \min_{\alpha} f(\alpha, \alpha^*).
\]

In (34) we have defined the objective function of our optimization problem as \( f(\alpha, \alpha^*) \) where, according to the Wirtinger calculus \[15\], the parameter \( \alpha \) and its complex conjugate are regarded as two independent variables. We can now find a stationary point of \( f \) by setting the partial derivative to zero:

\[
\frac{\partial f(\alpha, \alpha^*)}{\partial \alpha^*} = 0 \iff \alpha = \frac{\hat{x}^H \hat{x}^c - \mathbb{E} [ \hat{x}^H d \lambda_{\text{extr}} ]}{\| \hat{x} \|^2}.
\]

A precise evaluation of \( \mathbb{E} [ \hat{x}^H d \lambda_{\text{extr}} ] \) is difficult. However, since \( \mathbb{E} [ \hat{x}^H d ] = 0 \) according to Bussgang’s theorem, we conjecture that \( | \mathbb{E} [ \hat{x}^H d \lambda_{\text{extr}} ] | \ll | \hat{x}^H \hat{x}^c | \) and therefore

\[
\hat{\alpha} \approx \frac{\hat{x}^H \hat{x}^c}{\| \hat{x} \|^2},
\]

which finally results in the estimator

\[
\hat{d} = \hat{x}^c - \frac{\hat{x}^H \hat{x}^c}{\| \hat{x} \|^2} \hat{x}.
\]

A similar estimator was recently suggested in \[16\], where the authors replace the true \( \alpha \) in (28) by

\[
\alpha \bigg|_{\sigma_2^2 = \frac{1}{K} \| \hat{x} \|^2} ,
\]

i.e., the integral (10) evaluated using the mean power of \( \hat{x} \) instead of the variance \( \sigma_2^2 \) of the original signal. However, the proposal in \[16\] is rather ad-hoc and motivated by means of the EXIT chart technique \[17\]. In our simulations, we found the difference between estimator (38) and the one from \[16\] to be negligible.

### C. Numerical Results

In order to verify the statement from Section III that it is optimal to operate the amplifier close to saturation, we now study the impact of the OBO on the frame error rate (FER) of a practical system that uses the BNC algorithm.

In our simulations, we use an LDPC code of rate \( r = 1/2 \) and blocklength 2048. We use QPSK mapping \((m = 2)\) and an OFDM system with \( K = 1024 \) subcarriers. Thus, each codeword corresponds to exactly one OFDM symbol. The cyclic prefix length is \( N_{\text{cp}} = 72 \), causing a power loss of about 0.3 dB. Fig. 6 presents results that are obtained with a soft limiter in the transmitter. The results for the TWT are similar and therefore not presented here. The figure shows the frame error rates after the zeroth, first and second iteration of the BNC algorithm, where the zeroth iteration refers to the original data estimate, i.e., before the Bussgang noise estimator is invoked for the first time. We found the possible improvement after the second iteration to be negligible.

According to (19), we set \( E_0/N_0 = C/\text{OBO} \) where the constant \( C \) is chosen such that a target FER in the order of \( 10^{-3} \) is reached. In contrast to the information-theoretic study in Section III, the clipping is applied here to an oversampled signal with an oversampling factor of \( L = 4 \). The resulting out-of-band signal components are filtered out, causing a power loss, which explains why the plots start at an output back-off of around 0.5 dB instead of 0 dB.

In the case of an AWGN channel (Fig. 6a), the minimum FER is reached at an OBO of about 0.8 dB, which is only slightly higher than the minimum possible OBO. For the frequency-selective channel (Fig. 6b), the optimal output back-off is around 1.1 dB, which is still a very low value.
thus conclude that the simulations have verified the result of the information-theoretic analysis. The optimal output power back-off of an OFDM system with a nonlinear amplifier is close to zero, because the benefit of the increased SNR outweighs the drawback of the nonlinear signal distortion.

V. CONCLUSIONS

We have studied an OFDM transmission system with a nonlinear amplifier at the transmitter. While it is possible to operate the amplifier mainly in the linear range, i.e., with a large output power back-off, this comes at the expense of a low signal-to-noise ratio at the receiver. Driving the amplifier into saturation, on the other hand, maximizes the received SNR, but introduces severe clipping noise into the signal.

In this paper, we have shown that the mutual information between channel input and output is maximized when the transmitter operates at a very low output back-off. We have furthermore shown by simulations, where the Bussgang Noise Cancellation algorithm has been used in the receiver, that the frame error rate is indeed minimized by a low output back-off.

Finally, we have derived a modification to the BNC algorithm which has, in a similar form, recently been suggested in an ad-hoc manner in the literature.

It should be stressed, however, that we have only considered a single link in this investigation. In a wireless network, where interference from other users is of concern, we expect that the overall system performance will benefit from a higher output back-off. This is because in an interference-limited system, lowering the transmit power of all transmitters leads only to a moderate decrease of the signal-to-interference-plus-noise-ratio (SINR) at the receivers, but still to a significant reduction of the distortion noise power.

REFERENCES