

Optimizing Throughput for Limited Receiver Circuit Power

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Abstract—Maximizing the battery life time of mobile devices and sensor nodes increasingly becomes a challenge, and receiver power consumption tends to become more problematic than delivering adequate transmit power. We address the challenge of achieving the highest possible throughput per Watt of receiver circuit power. Our results show that optimum and adaptive tuning of the front-end parameters of the receiver can result in substantial power savings, compared to the common practice of a design for worst case conditions. We obtain a closed form solution for maximum throughput and the corresponding optimal overall system specifications. We confirm that handling the interference from nearby channels has a large influence, and our analysis concludes that adaptive control of the IP3 performance has an overarching impact. We further describe how the adaptive overall system settings can be translated into optimum gain and IP3 specifications of each of the individual stages that form the receiver cascade, considering both the accumulation of circuit noise and distortion products. The example of a WLAN system is elaborated to illustrate our method.

I. INTRODUCTION

In current mobile receivers, such as laptops and cell phones, the radio frequency (RF) signal processing is a major factor that influences battery life time. In the absence of disruptive new approaches, we expect this trend will continue for the foreseeable future. Currently, mobile users spend more and more time downloading data. Therefore, the receiver power consumption becomes a more prominent power consuming part of the battery limited side of the transceiver chain.

The aim of our analysis is to find an appropriate operation point for each of the analog stages of a power-constrained receiver, such that its user data throughput is maximized.

For our analysis we restrict ourselves to a commonly used receiver architecture. A broad band signal is processed at radio frequency (RF) by the analog front end, where the wanted signal only occupies a small portion of the front-end bandwidth. The analog front end amplifies, down-converts, and filters the desired signal from the received broadband signal and presents the signal to the ADC.

Therefore, the first stages of our receiver need to handle a broadband signal which usually contains strong adjacent channel interferers, with a priori not fully known statistical properties [1]. These signals need to be handled with adequate linearity to avoid excessive distortion spill-over into the band of the desired signal [2]. Typically, a higher linearity requirement leads to a higher power consumption of the analog

circuit. Commonly, the linearity is specified for the highest power of the interference at which the receiver should still operate. This results in an overly linear design at all lower values of the interference power, which is a waste of energy and reduces battery lifetime unnecessarily.

Commonly, an RF design is based on a set of system specifications, determined by standardization [2], which may include packet error rates, sensitivity and modulation. An RF designer strives to achieve the lowest power design possible, for a fixed overall specification target [3], [4]. However, we address the converse, given an available receiver circuit power budget we determine the maximum throughput. To achieve this optimum we do not define the system requirements a priori. Rather, we assume the key system settings are variable. This optimum might seem to depend on a large number of independent variables, but since we are designing at the minimum power consumption in a given IC process it is known from theory that several variables can be linked in this minimum [5]. This reduces the number of independent variables which determine the optimum of the throughput.

A main contribution of this work is that we reduce the optimization of the throughput to one variable from which all other variables can be derived and that this optimum is unique. Furthermore, the result shows that by far the most dominant factor in power consumption for low power designs is the linearity requirement. This is graphically shown by varying the interference power relative to the wanted signal power and plotting the subsequent throughput versus power consumption.

II. MODELING SYSTEM THROUGHPUT

To determine which variables are dominant in low power design we strive to optimize the throughput T for a given receiver circuit power budget P_r , according to

$$\hat{T} = \max T. \quad (1)$$

Here the maximum is taken over all possible settings of the RF stages, provided that the total consumed power does not exceed P_r . We further use the capacity expression as a measure of achievable throughput T for the modeled AWGN system.

$$T = \log_2 \left(1 + \frac{S}{N_{tot}} \right) \quad (2)$$

where S is the input signal power. The total noise plus distortion, relative to power levels at the input, is

$$N_{\text{tot}} = N_{th} + \frac{N_r}{G_{\text{tot}}} + \frac{N_d}{G_{\text{tot}}}, \quad (3)$$

where N_{th} is the noise in the channel, N_r the electronics noise added by the analog circuits of the receiver, and N_d is the distortion caused by an interferer. Other noise sources such as LO leakage, DC leakage and images are not considered, since they are related to the architecture, topology and layout, which are beyond the scope of this paper. The AWGN noise in the channel is given by

$$N_{th} = kTB, \quad (4)$$

where k is Boltzmann's constant, T is the temperature and B is the bandwidth of the desired signal. Further, G_{tot} is the total maximum power gain of the analog receiver given by

$$G_{\text{tot}} = \prod_{m=1}^M G_m, \quad (5)$$

where G_m is the gain of the m^{th} stage in the cascade (Figure 1). We now need a more detailed model of N_d and N_r .

A. Distortion and Noise

We obtain the distortion power N_d in (3) via the third order input referred intercept point $IP3$. The $IP3$ is a measure of the linearity of the analog circuits in the receiver [2]. Total $IP3$, $IP3_{\text{tot}}$, is defined as

$$IP3_{\text{tot}} = P_{\text{int}} \sqrt{\frac{P_{\text{out}}}{N_d}}, \quad (6)$$

where P_{out} is the output power in the channel of the interferer due to the received channel interference power P_{int} . $IP3_{\text{tot}}$ corresponds to the extrapolated input power at which P_{out} and N_d are equal. For an analog receiver the output power is $P_{\text{out}} = G_{\text{tot}} P_{\text{in}}$, with P_{in} the input power. Therefore we can substitute P_{out} in (6) by $G_{\text{tot}} P_{\text{int}}$. The distortion can now be expressed as

$$N_d = \frac{G_{\text{tot}} P_{\text{int}}^3}{IP3_{\text{tot}}^2}. \quad (7)$$

We simplify our model by assuming that after further channel selectivity filtering, Nyquist sampling and A/D conversion, the baseband processing engine of the receiver has no further knowledge of this distortion signal, and experiences it as AWGN. The total worst case $IP3$ of M stages can be calculated via [5]

$$IP3_{\text{tot}} = \left(\sum_{m=1}^M \frac{\prod_{j=1}^{m-1} G_j}{IP3_m} \right)^{-1}, \quad (8)$$

under the worst case assumption that all distortion components are in-phase. Here $IP3_m$ is the third order intercept point of the m^{th} stage.

Next to the distortion signals, the analog front end adds noise N_r to the desired signal. This addition of noise is

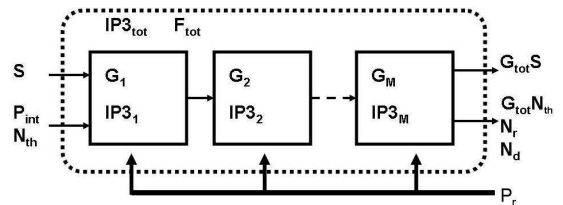


Fig. 1: Receiver cascade, consisting of M stages.

modeled via the noise figure (NF), and is defined as $NF = 10 \log_{10}(F_m)$. Here, the noise factor F_m is defined as:

$$F_m = \frac{SNR_m}{SNR_{m+1}}, \quad (9)$$

where SNR_m is the SNR at the input, and SNR_{m+1} is the SNR at the output of the m^{th} stage in a cascade. Note that the noise figures do not model the contribution by distortion. The total noise-factor of M stages in a cascade, F_{tot} , can be calculated via Friis formula

$$F_{\text{tot}} = 1 + \sum_{m=1}^M \frac{F_m - 1}{\prod_{j=1}^{m-1} G_j}, \quad (10)$$

where F_m the noise-factor of the m^{th} stage and G_j the gain of the j^{th} stage. Moreover, the total noise factor must satisfy

$$F_{\text{tot}} = 1 + \frac{N_r}{G_{\text{tot}} N_{th}} \quad (11)$$

where G_{tot} is the total gain of the analog circuit and N_r is the variance of the electronics noise added by all analog circuits weighed with the partial gains. Note that the distortion noise does not contribute to the noise figure. The electronics noise can now be expressed as

$$N_r = (F_{\text{tot}} - 1) N_{th} G_{\text{tot}} \quad (12)$$

By combining (3) (7) and (12), the total noise plus distortion, normalized to power levels present at the input, is now given by

$$N_{\text{tot}} = F_{\text{tot}} N_{th} + \frac{P_{\text{int}}^3}{IP3_{\text{tot}}^2}, \quad (13)$$

while achieving $IP3_{\text{tot}}$ using (8) and F_{tot} using (10). Figure 1 now depicts how the receiver is modeled, every individual stage has variable gain (G_1, \dots, G_M) and $IP3$ ($IP3_1, \dots, IP3_M$), thus allowing for variable $IP3_{\text{tot}}$ and variable F_{tot} of the receiver cascade.

B. Optimum Throughput

Our aim is to optimize the power budget P_r over the various stages such that throughput T is optimum under the constraint of a given technology, a given received power S , total gain G_{tot} needed to drive the ADC and channel parameters N_{th} and P_{int} . The optimization is done in a two staged approach by using F_{tot} and $IP3_{\text{tot}}$ as an intermediate design variable.

Section II-C.1 first addresses how for a chosen F_{tot} and $IP3_{\text{tot}}$ one can optimize the power budget P_r by optimally distributing the gains (G_1, \dots, G_M) and $IP3$'s

($IP3_1, \dots, IP3_M$) over the cascade. In fact, this section addresses a mathematical formalization of a commonly encountered design problem in RF design, of optimizing each circuit block to meet a given spec for the receiver. In [5] a double Lagrangian tool is proposed for this exercise, to solve the distribution of the gains and $IP3$ for the cascade. Paper [4] summarizes [5] for $M = 2$. This Minimum Power Cascade Optimization (MPCO) solves:

$$P_{\min} = \min \left(\sum_{m=1}^M P_m \right), \quad (14)$$

where P_m is the power dissipation of circuit block m , for a given technology, as will be covered by (18). Section II-C extends the MPCO by further optimizing F_{tot} and $IP3_{\text{tot}}$ to satisfy our end goal of maximizing the throughput per unit of consumed circuit energy, thus searching for

$$\hat{T} = \max_{F_{\text{tot}}, IP3_{\text{tot}}} (T), \quad (15)$$

where the available receiver circuit power P_r , satisfies $P_r = P_{\min}$ as in (14). By substituting (2) and (3) in (15), we can show that maximizing (15) corresponds to

$$\hat{N}_{\text{tot}} = \min_{F_{\text{tot}}, IP3_{\text{tot}}} (N_{\text{tot}}). \quad (16)$$

A main contribution of this paper is that we can reduce the minimization of (16) to one variable. This is achieved by expressing the power optimal $IP3_{\text{tot}}$, called $\widehat{IP3}_{\text{tot}}$, as a closed form function (22) of the figure of merits related to the used IC design process, the available receiver circuit power P_r , the power optimal total noise factor \widehat{F}_{tot} and total gain G_{tot} . Therefore, we can write N_{tot} as a function of \widehat{F}_{tot} and the available receiver circuit power P_r . Our claim is that in the end (16), for a given power budget P_r , is equal to

$$\hat{N}_{\text{tot}} = \min_{F_{\text{tot}}} (N_{\text{tot}}(\widehat{IP3}_{\text{tot}})). \quad (17)$$

We will call this an Maximum Throughput Cascade Optimization Method (MTCO).

C. Minimum Power Cascade Optimization Method

1) *Linearity Factor Model*: Traditionally, designers aim at achieving the system specifications for $IP3_{\text{tot}}$, F_{tot} , and G_{tot} at minimal power dissipation. This, justifies the use of an equivalent figure of merit (EFOM) [5], such as

$$P_m = \frac{f_m G_m IP3_m}{\kappa_m}, \quad (18)$$

where f_m is the power limiting bandwidth and κ_m is the power linearity factor of the m^{th} stage. The most appropriate parameter to chose for f_m highly depends on the circuit functionality. For LNAs with a dominant pole, the bandwidth is an appropriate choice. By using an EFOM, P_m theoretically does not depend on the noise figure F_m .

2) *Dual Lagrange Optimization Method*: So,

$$P_{\min} = \min_{G_1, \dots, G_M, IP3_1, \dots, IP3_M} \left(\sum_{m=1}^M \frac{f_m}{\kappa_m} G_m IP3_m \right), \quad (19)$$

while achieving the G_{tot} using (5), $IP3_{\text{tot}}$ using (8), and F_{tot} using (10). In this optimization process, the f_m , κ_m and F_m of a cascade are taken as constant. The individual F_m is kept constant because the F_m is fundamentally limited by the topology and used technology. A closed form expression can be derived [5] for the minimal analog signal conditioning (ASC) power dissipation as a function of the overall noise factor F_{tot} ,

$$P_{\min} = IP3_{\text{tot}} \left(\sqrt{\frac{f_M}{\kappa_M} G_{\text{tot}}} + \sqrt{\frac{F_w}{F_{\text{tot}} - F_1}} \right)^2, \quad (20)$$

where the "weighed excess noise factor" F_w is defined as

$$F_w = \left(\sum_{m=1}^{M-1} \sqrt[3]{\frac{f_m}{\kappa_m} (F_{m+1} - 1)} \right)^3. \quad (21)$$

D. Maximum Throughput Cascade Optimization Method

Whereas (20) gives the minimum power P_{\min} needed to satisfy a required $IP3_{\text{tot}}$, we can conversely claim that the best $\widehat{IP3}_{\text{tot}}$ that one can achieve for a given available P_r equals

$$\widehat{IP3}_{\text{tot}} = P_r \left(\sqrt{\frac{f_M}{\kappa_M} G_{\text{tot}}} + \sqrt{\frac{F_w}{F_{\text{tot}} - F_1}} \right)^{-2}. \quad (22)$$

Now, we can rewrite the total noise (3) as a function depending on F_{tot} , P_{int} and P_{\min}

$$N_{\text{tot}} = F_{\text{tot}} N_{\text{th}} + \frac{P_{\text{int}}^3}{P_r^2} \left(\sqrt{\frac{f_M}{\kappa_M} G_{\text{tot}}} + \sqrt{\frac{F_w}{F_{\text{tot}} - F_1}} \right)^4, \quad (23)$$

The noise factor is a real positive number which is $F_{\text{tot}} \geq F_1$.

1) *Maximizing Throughput*: By combining (1), (2), (3), and (23), we can maximize T by minimizing N_{tot} . We require that

$$\left. \frac{dN_{\text{tot}}(F_{\text{tot}})}{dF_{\text{tot}}} \right|_{F_{\text{tot}} = \widehat{F}_{\text{tot}}} = 0 \quad (24)$$

and obtain \widehat{F}_{tot} as $\widehat{F}_{\text{tot}} =$

$$F_1 + \frac{4F_w}{\left(-\sqrt{\frac{f_M}{\kappa_M} G_{\text{tot}}} + \sqrt{\frac{f_M}{\kappa_M} G_{\text{tot}}} + 2^{5/3} \left(\frac{F_w N_{\text{th}} P_r^2}{P_{\text{int}}^3} \right)^{1/3} \right)^2} \quad (25)$$

Applying this value of \widehat{F}_{tot} means that (12) turns into $N_r/G_{\text{tot}} =$

$$(F_1 - 1) N_{\text{th}} + \frac{4F_w N_{\text{th}}}{\left(-\sqrt{\frac{f_M}{\kappa_M} G_{\text{tot}}} + \sqrt{\frac{f_M}{\kappa_M} G_{\text{tot}}} + 2^{5/3} \left(\frac{F_w N_{\text{th}} P_r^2}{P_{\text{int}}^3} \right)^{1/3} \right)^2}, \quad (26)$$

TABLE I: Typical design choices for cascades in an ASC (LNA [6], Mixer [7], and Output buffer [8]). Where, numbers denoted in italics are considered as variable in this paper.

	LNA	Mixer	Buffer
NF [dB]	1.7	9.1	16
Gain [dB]	16	10.2	0
IP3 [dBm]	4	10.7	19
κ_m	$7.65 \cdot 10^9$	$17.8 \cdot 10^9$	$1.22 \cdot 10^9$
f_m [MHz]	100	2500	22

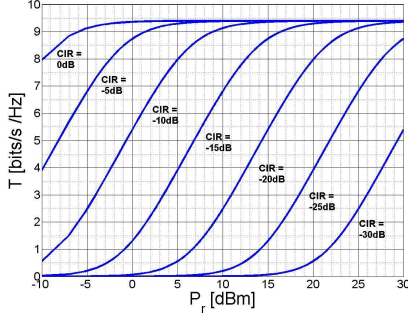


Fig. 2: Throughput for different values of CIR, versus receiver power P_r for a WLAN system.

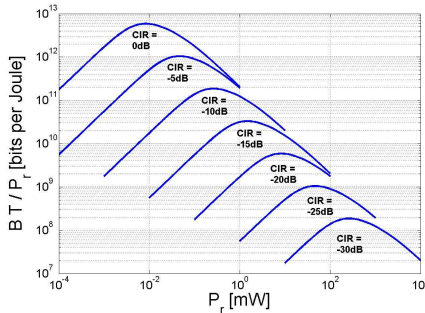


Fig. 3: Bits per Joule for different values of CIR, versus receiver power P_r for a WLAN system.

and (7) turns into $N_d/G_{tot} =$

$$\frac{P_{int}^3}{P_r^2} \frac{1}{16} \left(\sqrt{\frac{f_M}{\kappa_M} G_{tot}} + \sqrt{\frac{f_M}{\kappa_M} G_{tot} + 2^{5/3} \left(\frac{F_w N_{th} P_r^2}{P_{int}^3} \right)^{1/3}} \right)^4, \quad (27)$$

which we insert in (2) and (3). We now have found an analytically closed form solution which maximizes the throughput for a given circuit power budget P_r . We also found closed form solutions for \hat{F}_{tot} and $\hat{IP3}_{tot}$ that achieve the optimum throughput, respectively (25) and (22) with (25) inserted. The individual gain (G_1, \dots, G_M) and $IP3$ ($IP3_1, \dots, IP3_M$) per stage follow from [5].

III. NUMERICAL RESULTS

Figure 2 and 3 show the results for the system specifications, $S/N_{th} = 30$ dB, $k = 1.38 \times 10^{-23}$, $T = 295$ K, $B = 22$ MHz, and $G_{tot} = 65$ dB, when using the EFOM of the building blocks of Table I. The considered technology is 90nm CMOS. The characteristic frequencies f_m are chosen to satisfy the frequency requirements of an IEEE802.11b system in the 2.4 GHz band. Therefore, f_m is for the LNA $f_1 = 100$ MHz,

the mixer $f_2 = 2500$ MHz, and the output buffer $f_3 = 22$ MHz. Channel to interference ratio (CIR) is defined as, $CIR = S/P_{int}$. The closed form solution for optimum throughput as a function of available receiver power, using (2), (3), (26) and (27), is depicted in figure 2. For large available receiver power the throughput approaches the throughput for a signal 30 dB above thermal noise, and LNA with noise figure $F_1 = 1.7$ dB. At small available receiver power N_r and N_d become dominant. The figure shows that when the CIR is decreased, the power needed to achieve a certain throughput is increased. A relation between CIR and power consumption was reported earlier [1]–[5], but we now have formalized this relation.

The closed form solution for maximizing throughput can be extended to express bits per Joule as a function of available receiver power BT/P_r , using (1), (2), (3), (26) and (27). The result is depicted in figure 3. When the CIR is decreased, the efficiency in bits per joule for a given available receiver power is decreased as well.

IV. CONCLUSIONS

An analytically closed form solution has been presented which maximizes the throughput for a given available receiver circuit power. From the maximized throughput all other receiver system specifications such as $IP3$ and F can be derived. In turn, these derived system specifications allow us to derive the specification of the individual stages which form the receiver cascade. Furthermore, the closed form solution allows us to formalize the relation between interference power and achievable throughput for a given available receiver circuit power budget.

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