Circuit Aware Design of Power-Efficient Short Range Communication Systems

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***Abstract*— We consider the maximization of the overall power efficiency when communicating over a noisy *single-input single-output* channel while satisfying a certain throughput constraint. The total power dissipation includes the transmit power as well as the analog circuit power consumption. In the context of battery operated short range communication, where low power, low cost and small size are key requirements (e.g. standard IEEE 802.15.4), this circuit aware system optimization is crucial given the growing importance of "*Green Communication*". In fact, the transmit power in such applications reaches values in the order of or is even smaller than the power dissipation of certain analog components along the signal path. Using an appropriate information-theoretic framework we derive the optimal bit-resolution of the Analog-to-Digital Converter (ADC), the optimal choice of the noise figure for the low noise amplifier (LNA) and the optimal operating input back-off (IBO) of the Power Amplifier (PA) as a function of the path-loss (i.e. the communication distance). This work is an extension to the work [1], where only the ADC and transmit power have been considered.**

I. Introduction

In his famous work [2], Shannon showed that the maximal achievable rate of an AWGN channel with given transmit power P_T and bandwidth B is given by

$$R = B\log_2(1+\mathrm{SNR}) = B\log_2\left(1+\frac{G_c P_T}{N_0\cdot B}\right), \quad (1)$$

where G_c is the radio path-gain[1] and N_0 is the one-sided noise spectral level (in Joule). The classical measure for power efficiency in communications takes only the radiated energy per bit $E_b = P_T/R$ into account. Since the rate function is monotonically increasing with the bandwidth B, the minimum received signal energy per information bit is obtained by taking the bandwidth to infinity

$$\left[\frac{E_b}{N_0}\right]_{\min} = \lim_{B\to\infty}\frac{P_T}{N_0\cdot R} = \frac{\ln 2}{G_c}. \quad (2)$$

Obviously, since there is no penalty from taking the bandwidth to infinity, the maximum power-efficiency is obtained at infinite bandwidth. Besides it is assumed that the receiver has access to the channel data with infinite precision. This classical information theoretic approach is motivated by long range communication and thus neglects the conversion and processing power. However, when communicating over smaller distances using energy-constrained devices (e.g. sensor networks or on-chip communication), the transmit power can be comparable to the analog/digital processing power. Thus, the power consumption of certain analog and digital components can have a strong impact on the total power consumption. In this paper we aim to jointly minimize, the power consumption of certain analog transmitter and receiver components, in addition to the radiated power. This is motivated by the fact that the transceiver chip transmits and receives almost at the same data rate. The problem of jointly minimizing the transmission and the electronic processing power has been considered in [3], [4] among others. A common assumption is that the power consumptions of the analog components are constant parameters. The best strategy in that case is to employ bursty transmission with an optimized duty cycle. Nevertheless, the power consumptions of the analog components are generally mutually coupled with other system parameters (bandwidth, modulation scheme, noise power, bit resolution, input back-off...) which in turn directly affect the channel capacity. Therefore, we carefully model the system components in order to get well founded results. Fig. 1 shows a block diagram of a wireless transceiver. The encoding and decoding parts are not yet considered in the optimization, but could be subject of future work.

[1]Although $G_c \leq 1$, it is more convenient to refer to it as path-gain, i.e. the inverse path-loss.

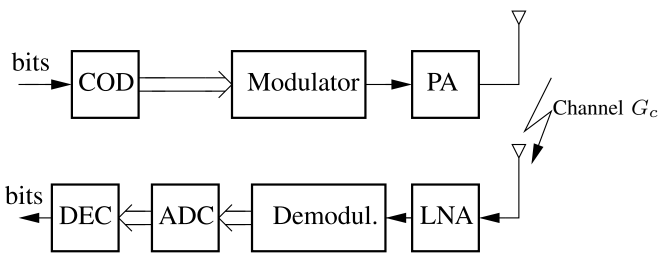

Fig. 1. Components of a SISO transceiver.

In the following, we briefly present the results of our literature search on communication circuits and their modeling [5], [6].

A. *Power consumption of the Power Amplifier (PA)*

The power consumption of the power amplifier is denoted by P_{PA} and already includes the radiated power P_T. They are in fact related by

$$P_T = \eta P_{\mathrm{PA}}, \tag{3}$$

where $\eta \leq 1$ is the drain power efficiency of the amplifier. The drain power efficiency η usually depends on the ratio of the instantaneous amplitude of the RF sinusoid a_x to the maximum output voltage A. Let consider the simplified push-pull amplifier in Fig. 2, having unit voltage amplification without loss of generality, and let describe it by the following simple model

$$x = a_x \sin(\omega t + \varphi_x) \tag{4}$$

$$u_t = \mathrm{sign}(\sin(\omega t + \varphi_x)) \cdot \min(|a_x \sin(\omega t + \varphi_x)|, A) \tag{5}$$

$$i_d = i_L \approx I_d \sin(\omega t + \varphi_x), \tag{6}$$

where ω is the carrier frequency. Based on this simplified push-pull amplifier model and depending on the operating mode, i.e., the case when the amplifier operates on its linear region or when it is overdriven into saturation, we get

$$\eta = \frac{\int_0^{\frac{2\pi}{\omega}} A \cdot i_d \mathrm{d}t}{\int_0^{\frac{2\pi}{\omega}} u_t \cdot i_L \mathrm{d}t} = \begin{cases} \frac{\pi}{4}\frac{a_x}{A} & \text{for } a_x < A \\ \frac{\sqrt{1-(\frac{A}{a_x})^2} + \frac{a_x}{U_0}\arcsin\frac{A}{a_x}}{2} & \text{for } a_x \geq A. \end{cases}$$

For a complex Gaussian input alphabet with variance σ_x^2 the amplitude a_x is Rayleigh distributed

$$f_{a_x}(a_x) = \frac{2a_x}{\sigma_x^2}\exp(-\frac{a_x^2}{\sigma_x^2}), \tag{7}$$

and we obtain the average PA efficiency

$$\bar{\eta} = \int_0^\infty \eta(a_x) f_{a_x}(a_x) da_x \approx \frac{1}{8}\pi^{3/2}\frac{1}{z}\mathrm{erf}(z) - \frac{(-4+\pi)(\exp(-z^2) + \mathrm{Ei}(-z^2))z^2}{4}, \tag{8}$$

where $z = \frac{A}{\sigma_x}$ represents the Input back-off (IBO) of the power amplifier, $\mathrm{Ei}(u) = \int_{-\infty}^{u} \mathrm{e}^t/t \cdot \mathrm{d}t$ denotes the exponential integral, and the approximation follows from the Taylor expansion $\eta \approx 1 - A^2/(6a_x^2)$ for $a_x \geq A$.

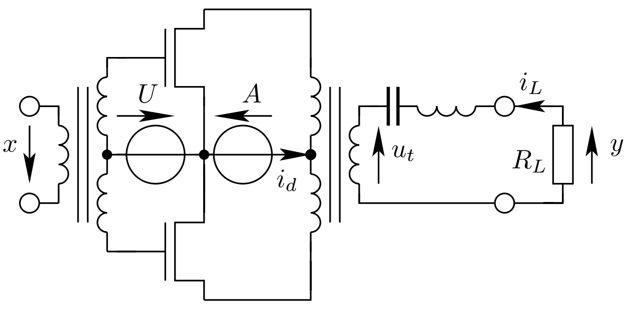

Fig. 2. Simplified push-pull transformer amplifier.

Obviously, there is a trade-off between power efficiency and the signal distortion due to clipping. To quantify the distortion effects we decompose the output of the amplifier y into a desired signal component and an uncorrelated distortion n_{PA} using the Bussgang theorem

$$y = \alpha \cdot x + n_{\mathrm{PA}}. \tag{9}$$

Assuming a soft-limiter type of nonlinearity ($y \approx \min(a_x, A)\sin(\omega t + \varphi_x)$) and an IBO z, α and the variance of n_{PA} take the values

$$\alpha = 1 - \mathrm{e}^{-z^2} + \frac{\sqrt{\pi}z}{2}\mathrm{erfc}(z), \tag{10}$$

and

$$\sigma_{n_{\mathrm{PA}}}^2 = (1 - \mathrm{e}^{-z^2} - \alpha^2)\sigma_x^2. \tag{11}$$

Finally, the PA power consumption can be written as

$$P_{\mathrm{PA}} = \frac{P_T}{\bar{\eta}} = \frac{\alpha^2\sigma_x^2 + \sigma_{n_{\mathrm{PA}}}^2}{\bar{\eta}} = \frac{1 - \mathrm{e}^{-z^2}}{\bar{\eta}}\sigma_x^2, \tag{12}$$

where $\bar{\eta}$ is given in (8).

B. Power consumption of the low noise amplifier (LNA)

Although, there are many performance factors in LNAs, we concentrate on four important parameters such as power gain, bandwidth, noise figure and power dissipation. These are included in a figure-of-merit expression $\mathrm{FOM}_{\mathrm{LNA}}$ [7]

$$\mathrm{FOM}_{\mathrm{LNA}} = \frac{G_{\mathrm{LNA}} \cdot B \cdot N_0}{(N_F - 1) \cdot P_{\mathrm{LNA}}}, \tag{13}$$

where G_{LNA} is the power gain, N_F is the noise figure, P_{LNA} is the power consumption and B is the operating bandwidth. Note that we multiplied the figure-of-merit definition in [7] by N_0, just to make it dimensionless. A noise figure of value N_F means that the LNA enhances the thermal noise level N_0 by the factor N_F. It turns out that for "good" designed LNAs the $\mathrm{FOM}_{\mathrm{LNA}}$ can be considered as an invariant quantity that only depends on the process technology and certain transistor parameters, and thus can not be influenced by our system optimization. Recent LNA designs found in the literature exhibit an $\mathrm{FOM}_{\mathrm{LNA}}$ in the range of 10^{-7} to 10^{-9}. The gain G_{LNA} is not subject of the optimization and is set to 10, so that the effect of noise from subsequent stages can be neglected.

C. Power consumption of the analog-to-digital converter (ADC)

The ADC complexity grows with the resolution b and the bandwidth B and it heavily affects the complexity of the following digital signal processing, e.g. the required memory size. In fact, it has been observed that new ADC architectures like pipelined ADCs are thermal noise limited and thus their minimum possible power is proportional to $N_0 \cdot 2^{2b} \cdot f_s$, where f_s is the sampling frequency [6]. In other words, under Nyquist rate sampling, the power needed for converting a complex signal with bandwidth B can be modeled as (see [6] for more motivation on this):

$$P_{\mathrm{ADC}} = 2 \cdot c_{\mathrm{ADC}} \cdot N_0 \cdot 2^{2b} \cdot B, \tag{14}$$

with some proportionality constant c_{ADC} depending on the ADC architecture. Although (14) is still not representative for the broad amount of actual ADC designs (especially at low resolution), it can be seen as the fundamental limit on conversion power due to thermal noise. Eq. (14) results in a trade-off between power consumption and performance loss due to quantization. It is therefore of interest to design the system parameters like bandwidth and ADC resolution in order to minimize the total power consumption including the conversion power.

D. Achievable Rate

Taking into account the distortion noise caused by the power amplifier and the noise figure of the LNA, we obtain the effective SNR as

$$\mathrm{SNR} = \frac{G_c \cdot \alpha^2 \cdot \sigma_x^2}{G_c \cdot \sigma_{n_{\mathrm{PA}}}^2 + N_F \cdot N_0 \cdot B}. \tag{15}$$

In [8], a lower bound on the achievable rate under output quantization by means of an MMSE approach has been derived, as follows

$$R \geq B \log_2 \left(\frac{1 + \mathrm{SNR}}{1 + \mathrm{SNR}/\lambda_q} \right), \tag{16}$$

where λ_q is the signal-to-distortion ratio related to the ADC resolution b. This bound is tight at low SNR and its derivation does not assume uncorrelated additive quantization noise. In order to maximize λ_q, an AGC-circuit is placed before the ADCs, which scales the noisy inphase and quadrature signals by a factor such that the ADCs are driven with an optimal input power. In this case, $\lambda_q \propto 2^{2b}$ and we can approximate the rate under finite resolution b as

$$R \approx B \log_2 \left(\frac{1 + \mathrm{SNR}}{1 + \mathrm{SNR} \cdot 2^{-2b}} \right), \tag{17}$$

which is consistent with the fact, that the achievable rate with infinite P_T is $2bB$. The ADC resolution is commonly chosen such that the ADC distortion noise is about 10dB below the overall noise level. However, such an approach is inappropriate for designing low power systems.

Using these energy consumption models, we aim to jointly minimize the transmission and the analog processing power consumption with respect to the different system parameters (resolution, bandwidth, noise figure, input back-off). The considered setting is free of a bandwidth constraint, which is quasi-true in UWB, but the optimization can be easily modified to include a bandwidth limitation. We also determine the fractions of power that should be optimally allocated to each of the components (PA, LNA and ADC).

II. Total Power Minimization

Let us consider the total power spent in the ADC and in the transmission,

$$P_{\mathrm{total}} = P_{\mathrm{PA}} + P_{\mathrm{ADC}} + P_{\mathrm{LNA}} = P_{\mathrm{ADC}}(1 + r_1 + r_2), \tag{18}$$

where we introduced the power ratios

$$r_1 = \frac{P_{\mathrm{PA}}}{P_{\mathrm{ADC}}} \text{ and } r_2 = \frac{P_{\mathrm{LNA}}}{P_{\mathrm{ADC}}}. \tag{19}$$

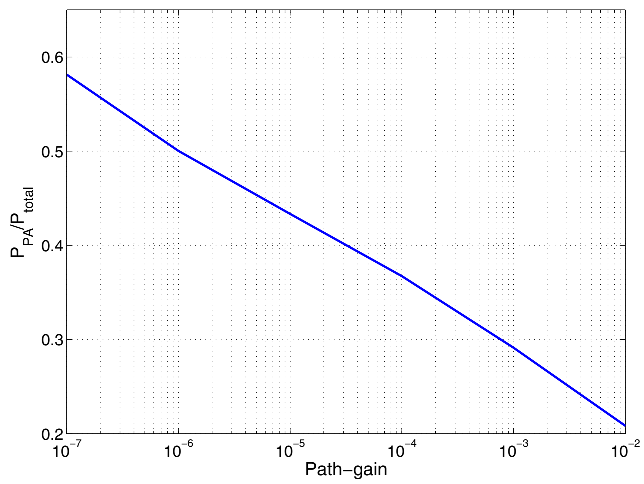

Fig. 3. Optimal PA power fraction (radiated power included) vs. the path-gain.

We aim to minimize the total power consumed for a given target rate, which might be interesting for low power applications. To this end, we modify (17) using (18) and (14)

$$R \approx \frac{P_{\mathrm{total}} \log_2 \left(\frac{1+\mathrm{SNR}(r_1, r_2, z, b)}{1+\mathrm{SNR}(r_1, r_2, z, b) \cdot 2^{-2b}} \right)}{(1 + r_1 + r_2) c_{\mathrm{ADC}} N_0 2^{2b+1}}. \tag{20}$$

It turns out, that the SNR in (15) is only a function of r_1, r_2, the resolution b and the IBO z. Evidently, we should minimize the total power for a given target rate to maximize the following objective function with respect to the power ratios r_1 and r_2, the ADC resolution b and the IBO z.

$$f(b, z, r_1, r_2) = \frac{\log_2 \left(\frac{1+\mathrm{SNR}(r_1, r_2, z, b)}{1+\mathrm{SNR}(r_1, r_2, z, b) \cdot 2^{-2b}} \right)}{(1 + r_1 + r_2) 2^{2b}}. \tag{21}$$

Fortunately, this optimization does not depend on the special target rate. However, we were not able to find a closed form solution. Using numerical optimization methods, the global optimizer of this function is easily found.

III. Numerical Results

As an example, let us take $c_{\mathrm{ADC}} = 96000$, which is a typical value of recent comparator-based ADCs [6], and $FOM_{\mathrm{PA}} = 10^{-7}$ [7]. Fig. 3 illustrates the behavior of the optimal fraction of PA power as a function of the path-gain G_c. Observe that, only for very small G_c (long range communication), the transmission power is dominant, in accordance to the classical approach.

Fig. 4 shows the behavior of the optimized conversion-to-transmission power ratio versus the path-gain G_c. As already mentioned, for large G_c, the ADC power becomes significant compared to the transmit power. On the other hand, the fraction of power that should be dedicated to the LNA shown in Fig. 5 becomes maximum at moderate path-loss. All in all, at very short communication distances, the conversion power

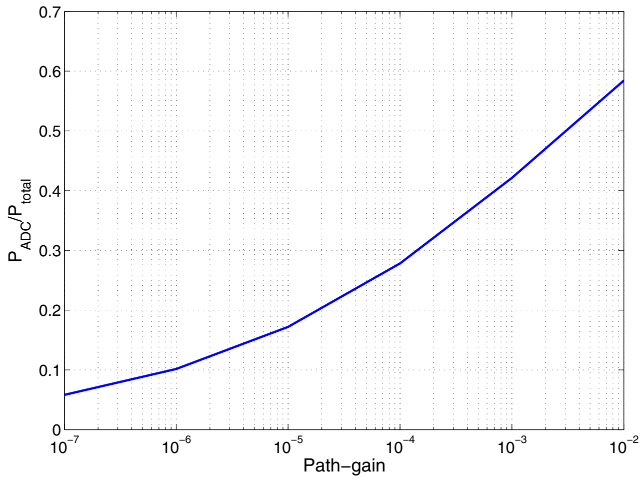

Fig. 4. Optimal ADC power fraction vs. the path-gain.

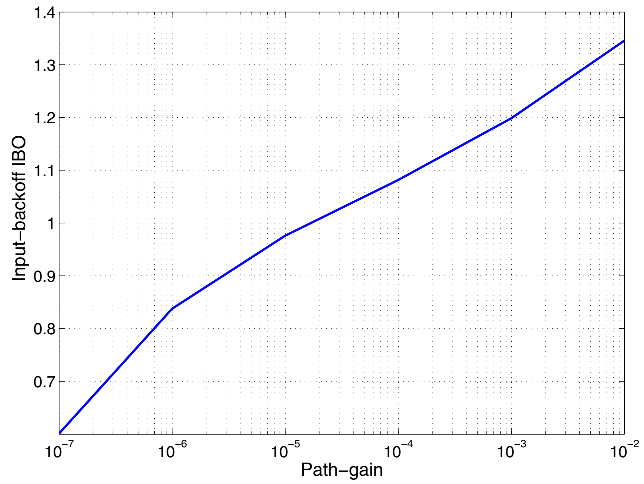

Fig. 6. Optimal operating IBO vs. the path-gain.

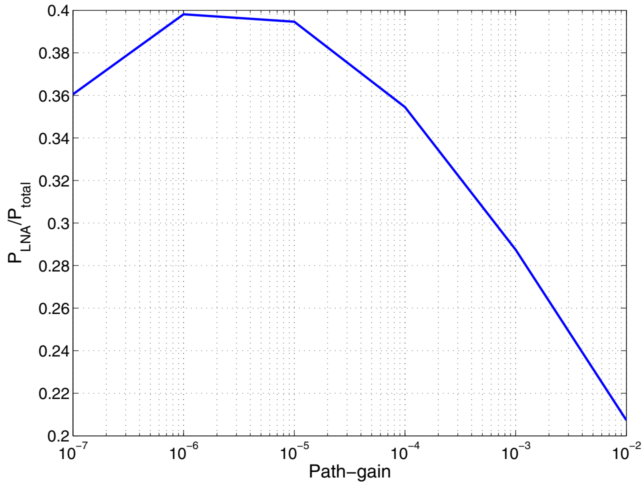

Fig. 5. Optimal LNA power fraction vs. the path-gain.

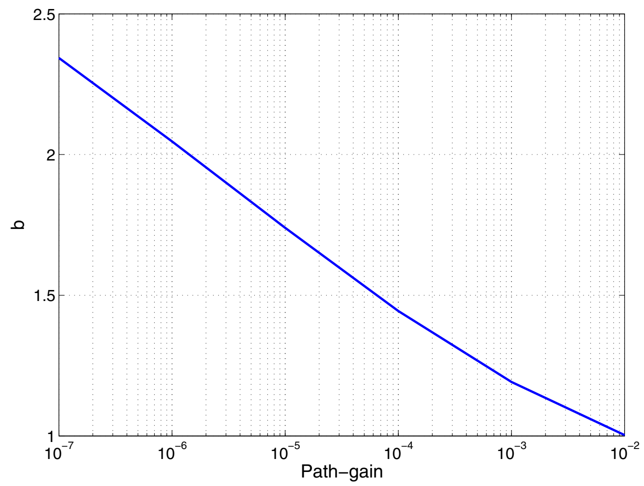

Fig. 7. Optimal ADC resolution vs. the path-gain.

is dominant.

Fig. 6 shows the optimal operating IBO of the power amplifier. At short distances, it is advantageous to operate the PA at high IBO values, i.e., at low drain efficiency, in order to reduce the distortion effects, due to the negligible PA power dissipation. On the other hand, Fig. 7 shows, that the optimal resolution for short range communication is indeed quite low and converges to the extreme case of 1 bit.

This suggests, that 1-bit ADCs may be a good choice for low power short range communication. Note, that if single bit hard-decision is used, the implementation of the digital receiver is considerably simplified [9], [10], [11]. In particular, automatic gain control (AGC), linearity requirements of RF components and multipliers for signal correlation are no more necessary. The optimal choice of the noise figure for the LNA is shown in Fig. 8. Clearly, there is no advantage from putting a lot of effort to reduce the noise figure at a high channel gain, i.e., at short communication distances, since this would require much higher LNA power relatively to the transmission power. Summarizing, the normalized combined energy per bit required to communicate across the noisy channel is depicted in Fig. 9. The term $\frac{P_{\text{total}}}{N_0 R_{\max}}$ corresponds to the minimal energy per bit. Remarkably, even at a very small path-loss, the required total energy per bit is quite large due to the ADC and LNA power consumption.

IV. Conclusion

We presented a new channel capacity optimization framework that takes into account the characteristics and the power consumptions of the ADC, LNA and PA. In the context of low power short range communication, we showed that, low-resolution (or maybe 1-bit) sampling exhibits a good performance, while reducing the total power consumption. It is also beneficial to design LNAs with poor noise figure in terms of overall power consumption. Besides, even if the system is

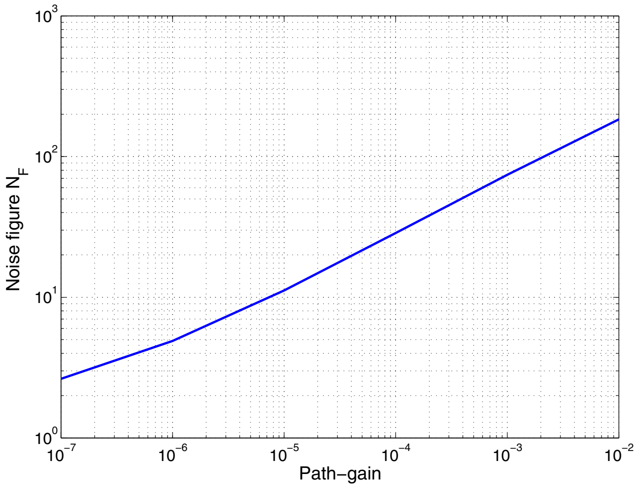

Fig. 8. Optimal noise figure vs. the path-gain.

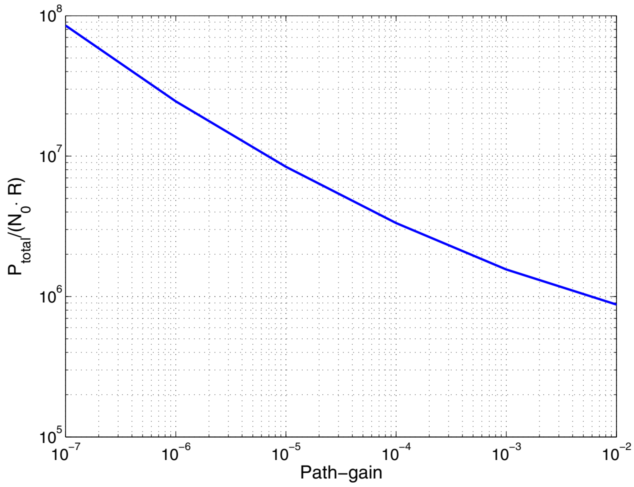

Fig. 9. Normalized combined energy per bit vs. the path-gain.

free of a bandwidth constraint, there is no advantage from taking the bandwidth to infinity contrary to the result stated by the Shannon theory. We believe that these results also hold for more general channel settings, e.g. MIMO channels. Additionally, the consideration of the decoding energy would be an interesting extension of this work.

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