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Adaptive nonlinear interference suppressor for cognitive radio applications

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Abstract—To utilize the radio frequency spectrum efficiently a Cognitive Radio (CR) can operate as a secondary user in a frequency band which is licensed to a primary user. To this end, the CR must sense the spectrum continuously to find empty frequency channels for its transmission. The transmitted signal by the local transmitter of the CR, however, induces a strong local interference in the local receiver of the CR. Hence a half-duplex transceiver is used where the transmit and sense operations are done in separate time slots. The time-slotted operation though, reduces the throughput of the CR. This paper proposes application of an adaptive Nonlinear Interference Suppressor (NIS) to suppress this strong local interference to enable simultaneous transmit and sense. We present experimental results of a transceiver testbed that uses an implementation of the NIS, fabricated in 140 nm CMOS technology. These experiments show that the NIS can substantially suppress the local interference with low complexity and power consumption. This suggests the application of the NIS for spectrum sensing in CR.

Index Terms—Cognitive radio, Multimode transceivers, Interference suppression, Adaptive circuits, Nonlinear systems, adaptive systems.

I. INTRODUCTION

Cognitive Radio (CR) is recognized as a solution to efficiently utilize the radio frequency spectrum. A frequency band, which is comprised of several frequency channels, is officially allocated to a Primary User (PU). The CR can temporarily occupy an empty channel in this band as a Secondary User (SU). To this end the CR must continuously sense the frequency spectrum over a wide frequency range to find empty channels for its operation [1].

If a duplex transceiver is used then the sense and transmit operations could be performed, simultaneously. The transmitted signal by the Local Transmitter (LTX) of the CR, however, is received by the Local Receiver (LRX) of the CR, inducing a local interference which can be several orders of magnitude larger than what would be received from the PU by the LRX. This large local interference desensitizes the LRX, and prevents simultaneous transmit and sense. Hence

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currently a half-duplex transceiver is used with separate time slots for transmit and sense. This half-duplex operation can cause interference to the PU and also reduces the throughput of the CR [1].

Traditionally duplexer passive filters are used to isolate the LTX and LRX in Frequency Division Duplex (FDD) transceivers. These duplexers however, have fixed pass-bands and stop-bands and hence are not suitable for CR application. The antenna cancellation scheme as proposed in [2], [3], requires an additional transmit antenna. The analogue cancellation method [2], [4], [5] cannot achieve an adequate interference suppression, without high analogue complexity.

An attractive approach is to suppress the local interference by passing the received signal through an adaptive memoryless nonlinearity [6]. This Nonlinear Interference Suppressor (NIS) is placed before active components of the LRX to prevent desensitization of the LRX. The NIS can significantly suppress a strong interference and pass the weak desired signal with a small amount of nonlinear distortion, provided that it is adapted as a function of the envelope of the received interference at the NIS input. Assuming that the coupling path of the interference from the LTX to the LRX is linear, the NIS adaptation signal can be digitally estimated using an adaptive model of this path and the baseband interference as the input of this model. Both the interference coupling path and the NIS are subject to environmental changes. Hence the path model is adapted during the LRX operation to track these changes such that the average power of the interference at the NIS output is minimized.

In this paper, we present experimental results of a transceiver testbed which uses the NIS chip, fabricated in 140 nm CMOS technology [7]. Performance of the NIS in terms of interference suppression and reception quality of the weak desired signal is presented. Finally, the pros and cons of using the NIS are discussed. The NIS approach is originally developed for multimode transceivers, where a local transmitter of one communication standard induces a large interference on a local receiver of another one. Although the experimental results are based on the multimode operation scenarios, the similarity of these two applications suggests the NIS as a potential solution for simultaneous transmit and sense in CR.

II. SYSTEM MODEL

A. Transceiver model

The CR transceiver, including the NIS, is shown in Fig. 1. The Local RX (LRX) is meant to receive a desired signal, transmitted by a Remote TX (RTX), while the Local TX (LTX) is active. The RTX here can be the PU transmitter (spectrum sensing) or a transmitter of another standard (multimode operation). The LTX FE up-converts the baseband interference $i(t)$ to a frequency f_i and transmits it as $i_t(t)$. At the LRX, the desired signal transmitted by the RTX is received in the presence of a part of $i_t(t)$ coupled from the LTX. The combination of these two signals is passed through a Band Pass Filter (BPF), which select the band of operation. After the BPF, the NIS input $x(t)$ includes both the desired signal and interference as:

$$x(t) = A_d(t) \cos(2\pi f_d t + \varphi_d(t)) + A_i(t) \cos(2\pi f_i t + \varphi_i(t)), \quad (1)$$

where A_d , φ_d , f_d , A_i , φ_i , and f_i are envelope, phase and center frequencies of the desired signal and interference at the NIS input, respectively. The power of the desired signal and interference at the NIS input are denoted by P_d and P_i .

B. Nonlinear Interference Suppressor, principle of operation

The NIS can be built by combining a linear amplifier (with gain of $-c$) and a limiter with an adaptable limiting amplitude $l(t)$ as shown in Fig. 2. The limiter gain for the weak signal is smaller than the gain for the strong signal, because of its compressive behavior. On the other hand the amplifier has the same gain for both weak and strong signals. By adapting $l(t)$ proportional to the envelope of the received interference, the gains of the limiter and amplifier for the strong interference can be made equal but of opposite sign. Thus, there is an optimal adaptation signal that leads to cancellation while the weak desired signal is amplified. The described principle for interference suppression is not restricted to the hard-limiter NIS and can also be applied to a soft-limiter, which is implemented in practice.

The high frequency components at the NIS output around $3f_i$ and higher harmonics can be simply filtered out. By neglecting these components $y(t)$ can be written as [6]:

$$\begin{aligned} y(t) &\simeq A_{d,y}(t) \cos(2\pi f_d t + \varphi_d(t)) \\ &+ A_{i,y}(t) \cos(2\pi f_i t + \varphi_i(t)) \\ &+ A_{IM}(t) \cos(2\pi(2f_i - f_d)t + 2\varphi_i(t) - \varphi_d(t)), \end{aligned} \quad (2)$$

where $A_{i,y}$, $A_{d,y}$, and A_{IM} are envelopes of the interference, desired signal and the main intermodulation at the NIS output. The NIS power gains for the desired signal and interference are defined as:

$$g_d = \frac{A_{d,y}^2}{A_d^2}, \quad g_i = \frac{A_{i,y}^2}{A_i^2} \quad (3)$$

There is an optimal adaptation signal $\tilde{l}(t)$, which minimizes the interference power $P_{i,y} = E(A_{i,y}^2)$ at the NIS output. For the hard-limiter NIS, assuming that $A_i \gg A_d$, by adapting the NIS according to $\tilde{l}_{HL}(t) = \frac{\pi}{4} c A_i(t)$, one can obtain: $g_i = 0$,

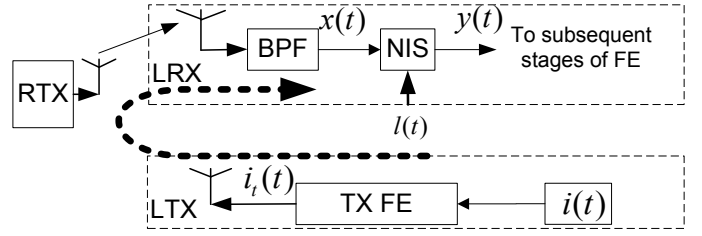


Fig. 1: CR transceiver with the NIS.

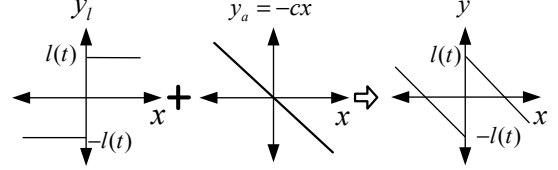


Fig. 2: NIS input-output characteristic with a hard-limiter.

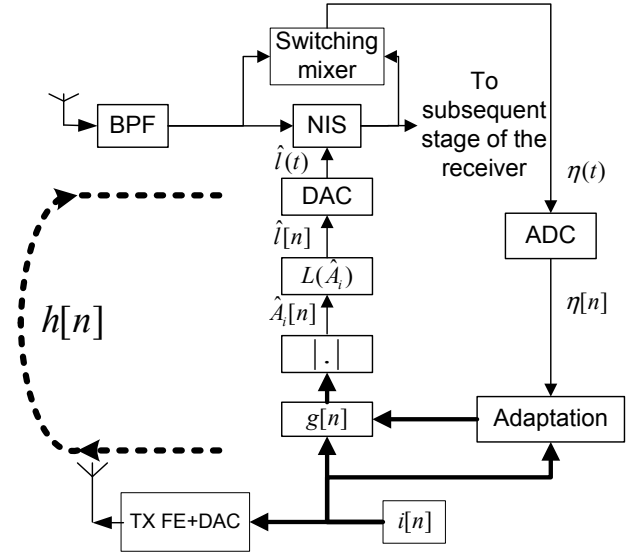


Fig. 3: Closed-loop adaptation of the NIS.

$g_d = \frac{c^2}{4}$, and $A_{IM} = A_d$ [6]. For the general case of a soft-limiter NIS, $\tilde{l}(t)$ is a function of $A_i(t)$, which will be denoted as $\tilde{l}(t) = L(A_i(t))$.

C. NIS adaptation

Fig. 3 shows closed-loop adaptation of the NIS [8]. Complex-valued signals are shown with bold lines. We assume that the coupling path of the interference is linear with a baseband impulse response $h[n]$. Hence $\tilde{l}[n] = L(|h * i[n]|)$, where $*$ denotes convolution. To generate an estimate $\hat{l}[n]$ of $\tilde{l}[n]$, an adaptive finite impulse response filter $g[n]$ is used with $i[n]$ as its input. The envelope of $g * i[n]$ is calculated and the function $L()$ is applied on the estimated envelope $\hat{A}_i[n] = |g * i[n]|$. Finally $\hat{l}[n] = L(\hat{A}_i[n])$ is converted to the analogue signal $\hat{l}(t)$. Since the coupling path and the NIS are subject to environmental changes, $g[n]$ must be adapted to track these changes. To this end an error signal $\eta(t)$ that is approximately proportional to $\hat{l}(t) - \tilde{l}(t)$ is extracted by downconverting $y(t)$ using $x(t)$ as the local oscillator of a

switching mixer, which is implemented in the same package as the NIS circuit. By processing $\eta[n]$ and $i[n]$ together, $g[n]$ is adapted such that the power of the residual interference at the NIS output is minimized.

For a narrow-band constant-envelope interference $A_i(t)$ changes only because of variations in the interference coupling path. Since these variations are slow, $\hat{l}(t)$ can be estimated directly by integrating $\eta[n]$ [6].

III. TRANSCEIVER TESTBED

A testbed, as shown in Fig. 4, is developed to characterize the NIS circuit and investigate the NIS performance in the receiver. The NIS and the switching mixer are fabricated as one chip in 140 nm CMOS technology. The chip is packaged and mounted on a PCB and the PCB is enclosed in a Faraday cage to fix the circuit's electromagnetic condition. Block diagram of the testbed is shown in Fig. 5. The baseband desired signal and interference are generated in a PC. They are combined digitally, uploaded to National Instruments (NI) Flex RIO module #1, converted to the analog IQ signals by NI5781 module #1, and up-converted by a Vector Signal Generator (VSG). The output of the VSG includes the interference ($f_i=1.85$ GHz) and desired signal ($f_d=1.87$ GHz) and is connected to the NIS input directly by a coaxial cable. Flex RIO#2 and NI5781 module#2 are used to generate $l(t)$ and digitize $\eta(t)$. The adaptation is done via the PC. The NIS output signal is down-converted by a commercial IQ mixer, digitized using the adapter module #1, and sent to the PC by FPGA module #1 for further processing.

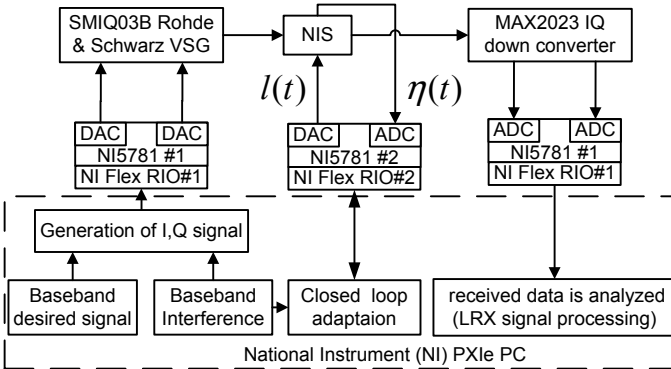


Fig. 5: Block diagram of the testbed, including LTX, RTX, and LRX with the NIS.

IV. MEASUREMENT RESULTS

A. Measurement results for Single Tone (ST) signals

As the first step we find the optimal adaptation signal $\tilde{l}(A_i)$ for a ST interference with envelope of A_i such that $A_{i,y}^2$ is minimized. The obtained function $\tilde{l}(A_i) = L(A_i)$ is stored in a look-up table and is used in the following experiments. To measure the NIS impact on the desired signal and the interference, we use a ST as the interference and another ST as the desired signal 60 dB smaller than the interference. The adaptation signal is set to $l = \tilde{l}(A_i)$ and g_d and g_i are measured and are shown in Fig. 6. The Interference Suppression (IS) is

defined as $IS = \frac{g_d}{g_i}$ and equals the amount of improvement in Signal to Interference Ratio (SIR) from the NIS input to output. For $0 \text{ dBm} < P_i < 11 \text{ dBm}$, although g_i is less than -30 dB, it is not zero as would be expected for a memoryless NIS. This nonzero g_i originates from a phase misalignment between the limiter and the amplifier. The misalignment depends on P_i . At $P_i \simeq -1 \text{ dBm}$ and 10 dBm , a perfect alignment is obtained and g_i becomes extremely small. For $0 \text{ dBm} < P_i < 11 \text{ dBm}$, g_d is approximately constant and drops when P_i goes below 0 dBm . When P_i becomes small, the soft limiter behaves linearly. Hence the interference is suppressed by subtracting the outputs of two linear blocks which have the same gain for the interference and the desired signal. Hence both signals are suppressed. For a varying envelope interference, variation of $A_i(t)$ leads to variation of $g(A_i(t))$. Hence the amplitude modulation of the interference is transferred to the modulation of the desired signal. This leads to Cross Modulation (CM) distortion of the weak desired signal. To extend the range that g_d is constant to smaller P_i , the slope of the soft-limiter can be increased. Alternatively the CM distortion can be compensated digitally. An IS of at least 35 dB is observed for $-2 \text{ dBm} < P_i < 11 \text{ dBm}$.

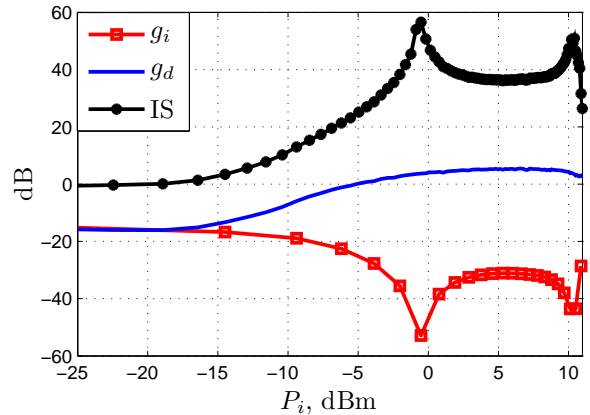


Fig. 6: g_i and g_d for a single tone interference and desired signal.

B. Closed-loop adaptation for constant envelope interference

In this section, for the interference a GMSK modulation and for the desired signal a 16QAM modulation is used, with 1 MHz and 6 MHz bandwidths, respectively. The SIR at the NIS input is -40 dB and $P_i = 10 \text{ dBm}$. The closed-loop adaptation method as described in [6] is used to adapt the NIS. Fig. 7 shows the NIS output spectrum. An interference suppression of about 43 dB is observed. Also the intermodulation component of the desired signal and the interference is seen at the image frequency of the desired signal with respect to the interference.

C. Closed-loop adaptation for varying-envelope interference

In this section, for both the interference and the desired signal a 16 QAM modulation is used, with 1.5 MHz and 6 MHz bandwidths, respectively. Root raised cosine pulse shaping

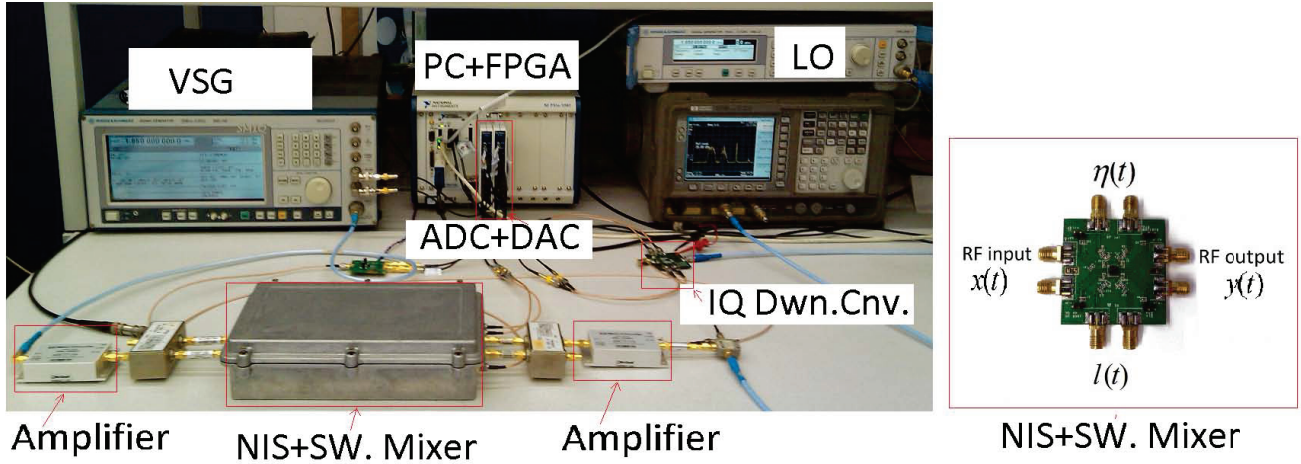


Fig. 4: Transceiver testbed.

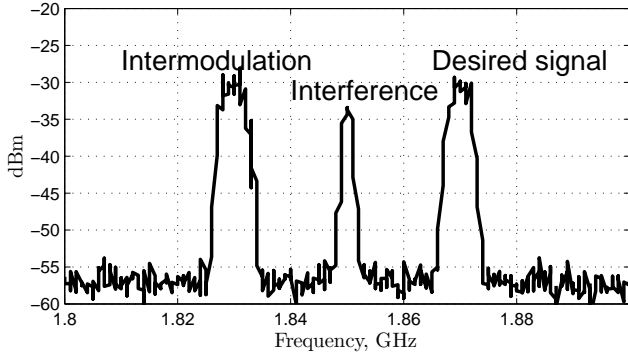


Fig. 7: NIS output spectrum, for GMSK interference.

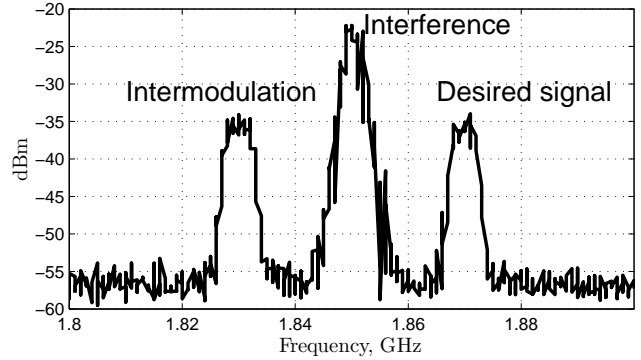


Fig. 8: NIS output spectrum, for 16QAM interference.

with roll-off-factor of 0.5 is used for both signals. The closed-loop adaptation method, shown in Fig. 3, is used to adapt the NIS. The SIR at the NIS input is -40 dB and $P_i = 4$ dBm. Fig. 8 shows the NIS output spectrum after convergence of the adaptation loop to the steady state condition. An interference suppression of about 33 dB is observed. Also the intermodulation component is seen at the image frequency of the desired signal with respect to the interference. Constellation diagram of the desired signal is shown in Fig. 9a. The constellation points of the transmitted signal and the received signal are shown in red and blue, respectively. The cross-modulation distortion of the received signal is observed in Fig. 9a. By multiplying the received signal by $\frac{1}{\sqrt{g(\hat{A}_i[n])}}$ the cross modulation is compensated. Fig. 9b shows the constellation diagram after the compensation. We see that the constellation is significantly improved. As the result of the compensation, modulation error ratio is improved from 18 dB to 25 dB.

V. POWER CONSUMPTION

The NIS approach includes analog and digital circuits. The analog circuits include the NIS circuit, DAC and ADC. The power consumption of the current realization of the NIS circuit is proportional to P_i and reaches to 35 mW for $P_i = 11$ dBm. For a varying envelope interference with about 5 MHz

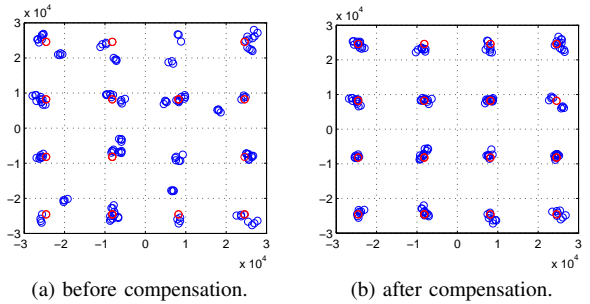


Fig. 9: Constellation diagram of the received desired signal.

bandwidth, the required bandwidth to accurately generate the adaptation signal is about 50 MHz. The power consumptions of a 14 bit DAC and 6 bit ADC with 100 MSPS are 16 mW and 5 mW, respectively [9], [10]. A detailed analysis of power consumption of the digital part of the NIS adaptation can be found in [11] which results in 10 mW for 100 MHz sample frequency. The total power consumption of the NIS method then would be about 66 mW. It must be noted that: firstly, these numbers are loose upper bounds, secondly, power consumption of the ADC, DAC and digital processing decreases as bandwidth of the interference decreases, thirdly, power consumption of the digital part can be reduced by

further optimization of the adaptation algorithms.

VI. ADVANTAGES AND DISADVANTAGES OF THE NIS APPROACH

Advantages of the NIS approach can be summarized as:

- 1-Significant interference suppression,
- 2-Low complexity and power consumption,
- 3-Possibility of a fully integrated solution.

Despite its advantages, using the NIS leads to the following undesired effects:

1-Spectral mirroring: All the components around $2f_i - f_d$ at the NIS input are mapped to f_d at the NIS output. This phenomenon which is referred to as spectral mirroring, leads to a 3 dB floor on the noise figure of the NIS. Also if an external interference is present at around $2f_i - f_d$, then it would be mapped to the frequency channel of the desired signal. This necessitates using a BPF before the NIS to suppress this component. For the CR applications, where a wide frequency agility is required, a tunable BPF may be used.

2-Harmonic generation: Owing to the strong nonlinear effects in the NIS, higher order harmonics of the interference are generated. These harmonics are far away from f_i and can be filtered out easily. They must be however, filtered immediately to prevent saturation of the NIS. To this end, the current implementation of the NIS circuit has a bandpass filter with a 3dB bandwidth from 1.75GHz to 1.95GHz at the NIS output. This filter limits the frequency range of the desired signal. The current implementation has a 3 dB bandwidth from 1.75GHz to 1.95GHz for the desired signal.

VII. CONCLUSION

A cognitive radio (CR) must sense the spectrum continuously to find empty frequency channels for its transmissions. The transmitted signal by local transmitter of the CR, however, induces a strong local interference in local receiver of the CR, forcing the transmit and sense to be done in separate time slots. To enable simultaneous transmit and sense an adaptive memoryless nonlinearity can be used to significantly suppress the local interference. In this paper, we presented experimental results of a transceiver testbed with this Nonlinear Interference Suppressor (NIS) implemented in 140nm CMOS technology. Measurement results show that the strong interference can be substantially suppressed with a small penalty to the quality of the desired signal. A closed-loop adaptations method is implemented which shows up to 43 dB improvement in the desired Signal to Interference power Ratio (SIR) for a GMSK interference. For a 16 QAM interference up to 33 dB of improvement in SIR is measured. The power consumption of the NIS approach is estimated to be below 66 mW. The current implementation has a 3 dB bandwidth from 1.75GHz to 1.95GHz for the desired signal.

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