Performance Evaluation of Decoupled Space–Time Delayed Decision–Feedback Sequence Estimation in Mobile–Radio Environments [(*)]

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Abstract–**This work aims at investigating the performance of a decoupled space–time processing structure based on a delayed decision–feedback sequence estimator (D–ST–DDFSE), in the context of the enhanced data rates for GSM evolution (EDGE) system. The main idea in using decoupled space–time processing is to separate co–channel interference (CCI) reduction from inter–symbol interference (ISI) suppression. Consequently, all degrees of freedom of a space–time front–end are dedicated to treat CCI, leaving ISI to be suppressed by a temporal equalizer. Due to the 8–PSK modulation and the large delay spread values compared to the symbol period, optimum detection becomes too complex in the EDGE system, which makes DDFSE a promising scheme for ISI suppression. The performance of D–ST–DDFSE is analyzed through link–level simulations under the context of COST 259 channel models for Typical Urban (TU) and Bad Urban (BU) propagation scenarios. Improved performance of this D–ST technique over a conventional space–time equalizer is observed.**

I. INTRODUCTION

The mobile–radio environment of incoming mobile communication systems are characterized by strong co–channel interference (CCI) and severe intersymbol interference (ISI), which are key factors that limit performance and capacity. Third generation systems as the enhanced data rates for GSM evolution (EDGE) are characterized by high data rates requiring the use of equalization at both ends of the link. Furthermore, tight reuse configurations lead to strong CCI that limits system capacity.

In the uplink, the use of antenna array diversity at the base station constitutes a classical solution for suppressing CCI and combating multipath fading [1]. The ISI due to multipath may be suppressed spatially by the adaptive antenna array. However, the rich multipath of practical propagation environments may limit the performance and it would be required too many antennas to overcome the effects of ISI and CCI. By introducing some kind of temporal equalization generally improves performance. Therefore, space–time (ST) processing techniques explore both the spatial and temporal structure of the received signals to obtain full (path) diversity. The classical space–time processing structure is the space–time linear equalizer (ST–LE) where a temporal equalizer follows each antenna element in order to exploit spatial and temporal diversity [2] simultaneously. In this case the minimum mean square error (MMSE) criterion is employed for CCI reduction and ISI suppression. Enhanced ISI suppression is obtained by the use of a maximum likelihood sequence estimator (MLSE) equalizer following an adaptive antenna array. However, in rich multipath scenarios the problem of insufficient degrees of freedom degrades MLSE performance due to residual CCI at the output of the array. Furthermore, in ISI–dominated scenarios with small values of angular separation, mainbeam user paths may severely reduce output signal–to–interference–plus–noise–ratio (SINR) and degrade bit–error–rate (BER) performance.

It is known that the optimum solution against ISI is the MLSE equalizer while an MMSE criterion is more robust against CCI. Thus, it is reasonable to state that it would be desirable to treat ISI with an MLSE equalizer, which is the optimum detector in the presence of ISI. Similarly, CCI is better combated with an MMSE equalizer. The idea of separating CCI and ISI suppression has been studied by several authors [3–7].

A decoupled space–time (D–ST) processing technique can make use of the individual advantages of an MMSE–based algorithm for CCI reduction and an MLSE–based algorithm for ISI suppression. This is done by separating CCI and ISI mitigation in two stages. A canceling filter is employed to generate a modified version of the training sequence that adapts the array, so that it cancels only CCI leaving all ISI structure to be suppressed within a temporal equalizer, whose parameters are obtained from the coefficients of the canceling filter. D–ST equalization was introduced in [8].

In this work we evaluate the performance of a D–ST processing structure based on a Delayed Decision–Feedback Sequence Estimator, namely D–ST–DDFSE, in the context of the Enhanced Data rates for GSM Evolution (EDGE) system. Due to the 8–PSK modulation and the

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large delay spread values compared to the symbol period, an optimum detection becomes too complex, which makes DDFSE a promising scheme for EDGE since it presents a good trade–off between performance and complexity [9]. The performance of D–ST–DDFSE is compared to that of ST–LE for Typical Urban (TU) and Bad Urban (BU) propagation scenarios of the context of the COST 259 channel model [10]. Link–level simulation results show that improved performance is achieved with this D–ST structure.

In the remainder of this paper, we organize the sections as follows: The D–ST–DDFSE structure is presented in section II. In section III, TU and BU scenarios of COST 259 channel model are characterized. Link–level simulations results are presented in section IV; and in section V we conclude the paper drawing some conclusions and perspectives.

II. D–ST–DDFSE STRUCTURE

As shown in Fig. 1, D–ST–DDFSE structure is composed by three basic elements: *i*) An antenna array and a temporal filter at the output of each antenna, which we will call a space–time (ST) linear front–end; *ii*) An FIR canceling filter; *iii*) A DDFSE equalizer, which is divided in two parts: A MLSE and a feedback filter.

A filtered version of the training sequence is employed as the desired signal to jointly adapt the array weights and the coefficients of the FIR canceling filter. After the training period, the coefficients of the channel estimator filter represent the desired channel impulse response modified only by the multiplication of the array weights whose values are such that they spatially cancel only the multipaths of the co–channel interferer. Therefore, the estimated channel impulse response is converted into the parameters of the temporal equalizer following the array that performs ISI suppression. The purpose of the ST front–end is to provide both spatial and temporal diversity for CCI reduction when fading is frequency selective. However, it may be reduced to an AA front–end when fading for all CCI sources is flat.

The channel estimator is used to synthesize an estimate of the overall CIR of the desired user. This is done by joint adapting the array weights and the FIR filter coefficients in order to minimize the mean square value of the error signal indicated in Fig. 1. The order of the channel estimator should be sufficiently large to capture the most delayed paths of the desired user. On the contrary, ISI due to the most delayed paths are suppressed by the ST front–end. At the end of the training period, ISI suppression is done by employing the estimated CIR within the DDFSE equalizer, while CCI reduction is performed at the ST front–end.

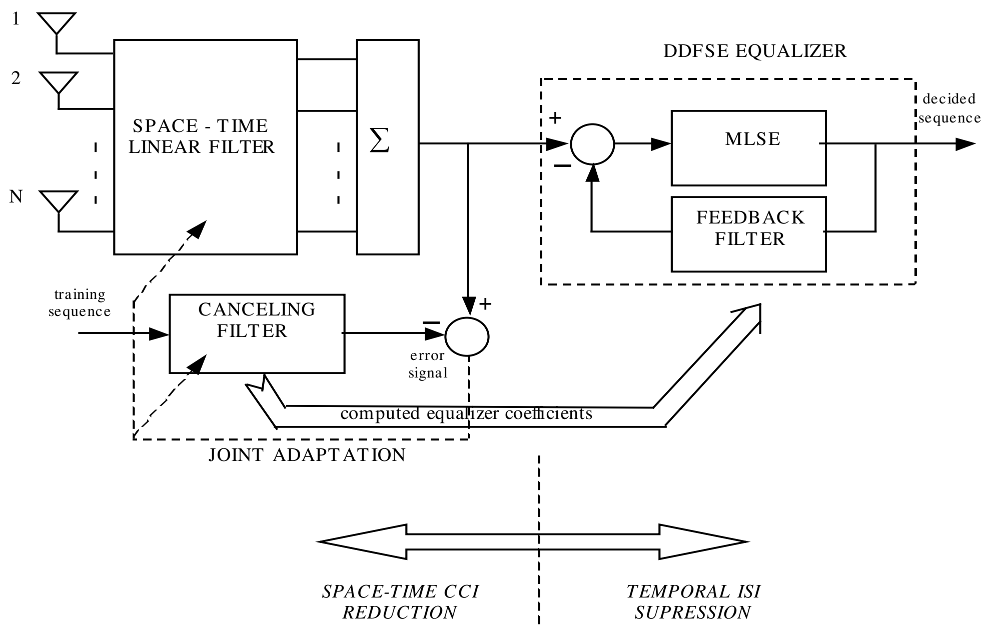

Fig. 1: The general structure of an adaptive D–ST–DDFSE.

The equivalent baseband signal received at the ST front– end can be denoted by:

$$\mathbf{x}(t)=\sum_{n=-\infty}^{+\infty} a_n \mathbf{h}(t-nT)+\sum_{l=1}^{L}\sum_{n=-\infty}^{+\infty} b_{n,l}\mathbf{h}_l(t-nT)+\mathbf{n}(t) \quad (1)$$

where

a_n symbol sequence of the desired user;

$b_{n,l}$ symbol sequence of the *l–th* interferer;

$\mathbf{h}(t)$ *M*x1 equivalent channel of the desired user;

$\mathbf{h}(t)$ *M*x1 equivalent channel of the interferer;

$\mathbf{n}(t)$ *M*x1 white noise vector;

L number of co–channel interferers.

Let us assume a ST linear front–end of *M* antennas with *P* taps per antenna and a canceling filter of *P* taps where,

$$\mathbf{W}=[\mathbf{w}_1^{\mathbf{T}}\ \mathbf{w}_2^{\mathbf{T}} \ldots \mathbf{w}_P^{\mathbf{T}}\ ]^{\mathbf{T}}, \quad \mathbf{w_i}=[w_{i1}\ w_{i2}\ \ldots\ w_{iM}]^{\mathbf{T}} \qquad (2)$$

is an (*M*x*P*) x 1 vector of the coefficients of the ST linear front–end;

$$\mathbf{c}=[c_1\ \ c_2 \ldots c_P]^{\mathbf{T}} \qquad (3)$$

represents the coefficients of the channel estimator filter;

$$\mathbf{X}=[\mathbf{x}_1^{\mathbf{T}}\ \mathbf{x}_2^{\mathbf{T}} \ldots \mathbf{x}_L^{\mathbf{T}}]^{\mathbf{T}}, \ \ \mathbf{x}_i=[x_{i1}\ x_{i2}\ \ldots\ x_{iN}]^{\mathbf{T}} \qquad (4)$$

is an (*M*x*P*) x 1 vector of received signal;

$$\mathbf{d}_{\mathrm{n}}=[d_{n-1}\ d_{n-2} \ldots d_{\mathbf{n-P}}]^{\mathbf{T}} \qquad (5)$$

is a *P*x1 vector of training symbols.

The optimum solution for both $\mathbf{W}$ and $\mathbf{c}$ maximizes the SINR at the output of the ST front–end as defined below:

$$(\mathbf{W}_o,\mathbf{c}_o)=\arg\max_{W,C}\frac{E\left\{\left\|\mathbf{W}^{\mathrm{H}}\mathbf{X}\right\|^2\right\}}{E\left\{\left\|\mathbf{W}^{\mathrm{H}}\mathbf{X}-\mathbf{c}^{\mathrm{H}}\mathbf{d}\right\|^2\right\}} \qquad (6)$$

where $E\{\bullet\}$ denotes the expectation operator. It can be shown that the maximization of (6) is equivalent to the minimization of the squared value of the error signal shown in Fig. 1. We must place a constraint on vector $\mathbf{c}$ to solve (6). In this work we work with the constraint $\mathbf{f}^{\mathbf{T}}\mathbf{c}=1$, $\mathbf{f}=[0\ldots f_j \ldots 0]$, where $f_j=1$. The value of *j* determines the training delay of the canceling filter [7] and will identify the number of causal and anti–causal taps of the estimated CIR. The arrow in Fig. 1 represents the acquisition of the equalizer parameters from the coefficients of the training filter. Therefore, adaptation is performed at the array and the canceling filter only.

III. COST 259 CHANNEL MODEL

The channel model suggested by COST 259 [10] is a wideband directional channel model capable of providing channel impulse responses in both spatial and temporal domains. It was validated using measurements in the 1GHz to 2GHz range, but it is expected to be applicable at least in the range 450MHz to 5GHz.

The channel responses of the Typical Urban (TU) and the Bad Urban (BU) scenarios are based on the macrocell radio environment of the COST 259. The desired user and the co–channel interferer are uniformly distributed within a 120º sector and cell radius was assumed to be 1 Km. Due to high bit rates of the EDGE system (T ≅ 3.7μs), where T is the symbol interval, channel time–variations during a time–slot are very small, and it are not considered here.

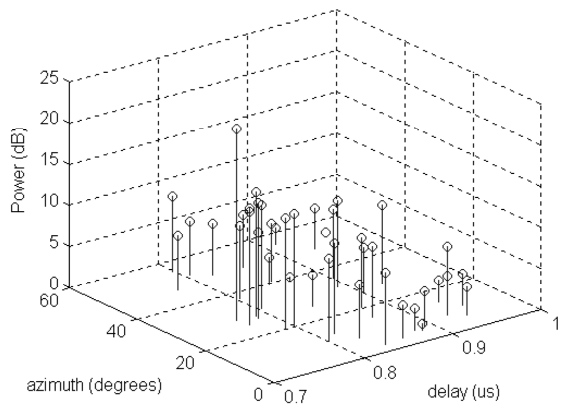

Fig. 2: TU channel realization. Small delay and angle spreads.

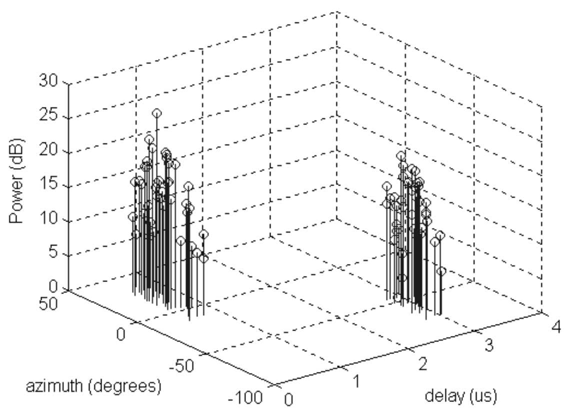

Fig. 3: BU channel realization. Clusters of scatterers and larger delay and angle spreads.

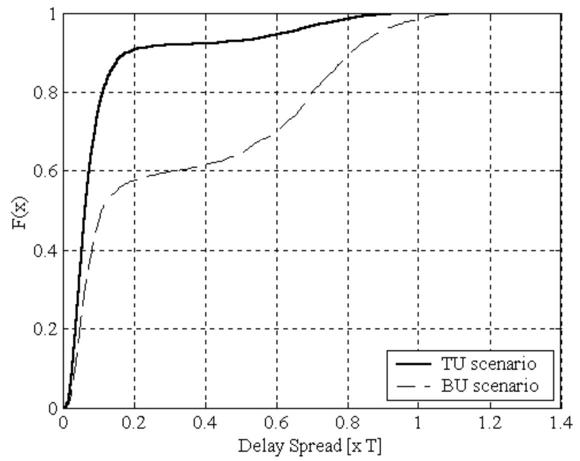

Fig. 4: CDFs of RMS delay spread for TU and BU scenarios.

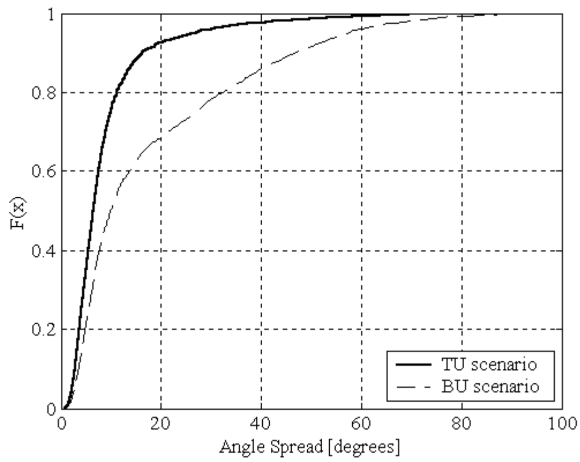

Fig. 5: CDFs of RMS angle spread for TU and BU scenarios.

In order to better characterize COST 259 channel models, a typical channel realization of TU and BU channels is illustrated in Fig. 2 and 3, respectively. In Fig. 4 and 5 we plot the CDFs of the RMS delay and angle spread, respectively, for both TU and BU scenarios. The CDF curves were obtained from a total 5000 snapshots of the COST 259 directional channel response. It can be seen that in 80% of the cases, delay spread is not superior to 10% of a symbol period in the TU case while in the BU case it reaches 70% of the symbol period. Similarly, angle spread does not exceed 10º in TU, while it reaches 30º in BU. From Fig. 2 to 5 we observe the TU scenario is characterized by the presence of a unique cluster of scatterers which is located local to base station in most of times, leading to small delay and angle spreads. In BU scenario, additional clusters of scatterers lead to larger delay and angle spreads.

IV. Simulation Results

The BER performance of the D–ST–DDFSE on TU and BU scenarios of the COST 259 channel model is evaluated here. The modulation scheme (8–PSK), slot format and symbol rate used in all simulations follow those of the EDGE system [11, 12], except the pulse shaping function, where we use a raised cosine with a roll–off factor of 35%. The recursive least squares (RLS) algorithm is used as the adaptation algorithm with a forgetting factor of 0.98. The first 26 symbols are used for training while the remaining 114 symbols are used for adaptation in the decision–directed mode.

A. Performance evaluation for a pure ISI scenario

The performance of D–ST–DDFSE is evaluated here and compared to that of the ST–LE on BU and TU scenarios, with no CCI. We use a 2–antenna–element array in all the structures. The channel estimator of the D–ST–DDFSE has 4 taps on the TU scenario and 6 taps on the BU scenario. The MLSE memory of the D–ST–DDFSE is set to 1. The DFE employs 3 feedback taps on TU scenario and 5 feedback taps on the BU scenario. The ST–LE has 3 taps per antenna and the D–ST–DDFSE uses an AA front–end in this pure ISI scenario. We consider two spatial diversity approaches. In the first one, the 2 antenna elements are separated by d=0.5λ while in the second one, such a separation is d=10λ, where λ is the wavelength. When only ISI is present, it is expected that the D–ST–DDFSE can exploit the spatial diversity at the antenna array and temporal diversity at the DDFSE equalizer to mitigate ISI.

In Fig. 6, it can be seen that the D–ST structure has the best performance on the TU scenario, especially for d=10λ, where the performance improvement is more pronounced. We observe also that the E_b/N_0 gain of D–ST–DDFSE increases with E_b/N_0. For d=0.5λ and a target uncoded BER of 10^{-3}, the E_b/N_0 gain of D–ST–DDFSE over ST–LE is approximately 3dB.

In Fig. 7 the BU scenario is considered. We observe that the improvement of D–ST–DDFSE over ST–LE is much greater in the BU where more spatial and temporal diversity is present. As well as on the TU scenario, the performance improvement of D–ST–DDFSE over ST–LE on BU scenario is more pronounced when full spatial

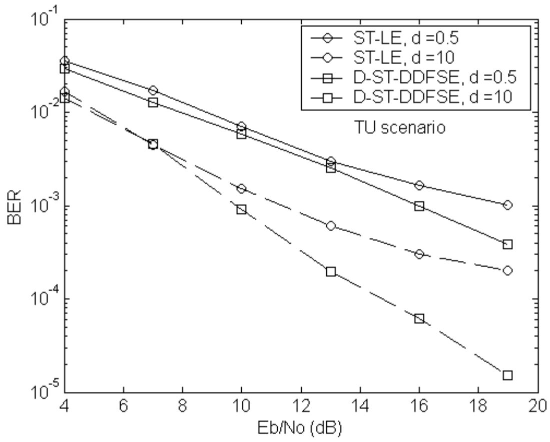

Fig. 6: Performance of D–ST–DDFSE on the TU scenario. The performance improvement of D–ST–DFSE increases as E_b/N_0 increases.

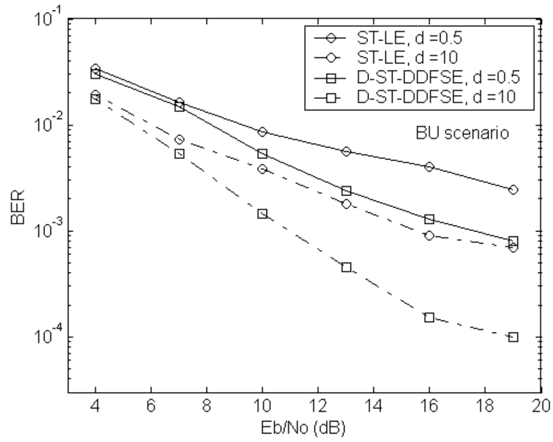

Fig. 7: Performance of D–ST–DDFSE on the BU scenario. The E_b/N_0 gain of D–ST–DDFSE over ST–LE is 5dB for a target uncoded BER of 10^{-3} and d=10λ .

diversity is provided by the antenna array, i.e. d=10λ. In this case, for target uncoded BER of 10^{-3} the E_b/N_0 gain of D–ST–DDFSE over ST–LE is remarkably 5dB. We choose here the ST–LE as a reference of comparison since it is the classical ST equalizer. However, in [13] we show that performance gains of D–ST–DDFSE over its conventional counterpart (ST-DDFSE) are also verified.

B. Performance evaluation in the presence of CCI

In the next simulations we include a single co–channel interferer. The SIR is 0dB and both the desired user and the co–channel interferer follow the same channel profile. We first consider the TU scenario. A ST front–end with 2 taps per antenna is employed in D–ST–DDFSE to enhance CCI reduction with ST processing. The ST–LE also has 2 taps per antenna. In Fig. 8 it can be seen that the performance of D–ST–DDFSE is superior to that of the ST–LE and an E_b/N_0 gain of approximately 6dB is observed for a target uncoded BER of 10^{-3}. We have observed a slight improvement on the performance of D–ST–DDFSE when 3 taps per antenna are used instead of 2. In Fig. 9, BU scenario is considered. Here, the ST front–end of both ST–LE and D–ST–DDFSE has 3 taps per antenna. The E_b/N_0 gain of D–ST–DDFSE over ST–LE is more than 7 dB for a target uncoded BER of $4x10^{-3}$. We verified that the performance gain of D–ST–DDFSE over ST–LE is more significant on BU scenario, where more temporal (delay) and spatial (angle) diversity are present.

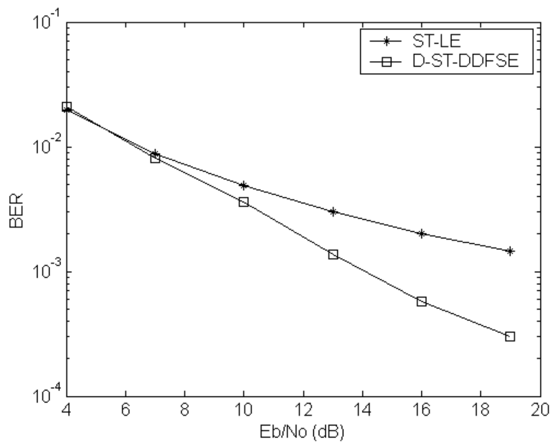

Fig. 8: Performance gain of D–ST–DDFSE over ST–LE on the TU scenario with CCI. An E_b/N_0 gain of approximately 6dB is observed at BER=10^{-3}.

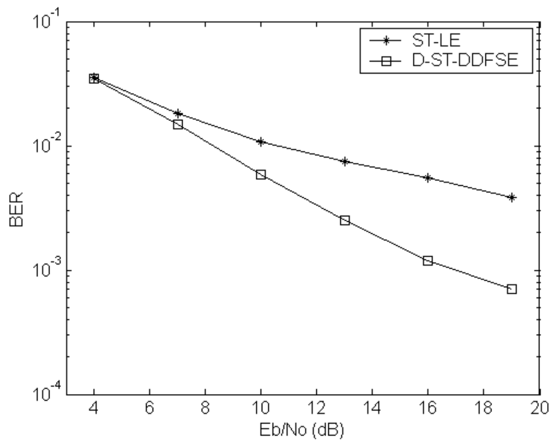

Fig. 9: Performance gain of D–ST–DDFSE over ST–LE on the BU scenario with CCI. We observe that the E_b/N_0 gain is more than 7dB at BER = $4x10^{-3}$

V. CONCLUSIONS AND PERSPECTIVES

The D–ST–DDFSE has presented good results on COST 259 BU and TU scenarios, indicating the advantage of using the idea of D–ST processing together with DDFSE equalization. In the TU case, performance gains of the proposed D–ST equalizer are pronounced for medium to high E_b/N_0 values. This was verified for pure ISI scenarios as well as in the presence of CCI. In the BU case the performance improvement of D–ST–DDFSE over ST–LE is great, indicating the robustness of the proposed D–ST structure in worst–case situations, where a high level of CCI and ISI is present. These simulation results reinforce that D–ST–DDFSE can perform quite well within the EDGE system.

The continuity of this work include a system–level evaluation of D–ST–DDFSE in order to clarify the potential user and data rate capacity gains when employing this D–ST structure in tight frequency reuse pattern scenarios.

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