# Decoupled Space–Time Equalization for CCI/ISI Suppression in Mobile Communication Systems *

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*Abstract* – **In this contribution, we present a family of decoupled space–time (D–ST) equalizers for improved reception in the presence of co–channel interference (CCI) and inter–symbol interference (ISI) in mobile communication systems. Basically, a D–ST equalizer separates CCI and ISI suppression in two processing stages. In the first one, an adaptive array (AA) spatially cancels only the co–channel interferers, while in the second one an equalizer performs temporal ISI suppression. The main idea of D–ST equalization is to use all the degrees of freedom of the array to improve CCI reduction and to preserve all ISI structure of the desired signal to be better exploited by a temporal equalizer following the array. In order to accomplish this task, a canceling filter is employed to generate a filtered version of the training sequence used to cancel the user signal in array coefficient adjustment. Simulation results are presented to compare the performance of some D–ST equalizers to that of conventional space–time structures in CCI/ISI–limited scenarios.**

## I. INTRODUCTION

INCOMING mobile communication systems are characterized by tight reuse configurations and a large delay spread, which makes co–channel interference (CCI) and inter–symbol interference (ISI) key factors that limits performance and capacity. The classical solution to combat CCI is the use of antenna diversity [1, 2]. However, ISI that is originated from multipath components of the desired signal are considered as additional CCI sources [3, 4] that should be suppressed by the antenna array also. In propagation environments characterized by a large variety of multipaths, the problem of insufficient degrees of freedom may lead to a poor CCI suppression, degrading signal reception. space–time (ST) processing techniques in general improve performance by including some kind of temporal processing in conjunction with a spatial only processing. Simultaneous CCI/ISI mitigation may be obtained with a space–time linear equalizer, where a temporal filter is used at the output of each antenna element. Another possible structure consists of a temporal equalizer at the output of the antenna array. The use of fractionally spaced taps may enhance CCI reduction, but a good performance will depend on the degree of correlation within the subchannels. The former ST structures in general perform better than the adaptive array (AA) in scenarios dominated by CCI and ISI. However they have two main drawbacks. First, the problem of insufficient degrees of freedom still degrades the performance of ST structures when a large number of multipath components is present. As a consequence, residual CCI at the output of the array degrades signal detection. Second, the presence of desired user multipaths falling at the mainbeam of the array severely affects output desired signal power and even the use of an equalizer cannot compensate for this loss of signal–to–interference–plus–noise ratio (SINR).

The former problems can be combated by treating CCI and ISI separately. Previous works have applied similar ideas. In [5, 6] an adaptive array is used with a maximum likelihood sequence estimator (MLSE) as the temporal equalizer. In [7], an MLSE is used with an ST front–end instead of the AA. In [8], a decision–feedback equalizer (DFE) following the array is used. The D–ST equalization technique was introduced in [9].

In this work, we evaluate the performance of a family of decoupled space–time (D–ST) equalizers that combat CCI and ISI separately, in mobile–radio environments. In D–ST equalization, a canceling filter is employed to generate a filtered version of the training sequence used to cancel the user signal in array coefficient adjustment, so that the array 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.

In the remainder of this paper, we organize the sections as follows. In section II, we describe our space–time signal model. In section III, conventional space–time equalization is briefly explained. D–ST equalization is presented in section IV. In section V, simulations results are presented in order to evaluate and compare the performances of the D–ST equalizers. At last, in section VI, we draw some conclusions.

## II. SPACE–TIME SIGNAL MODEL

We will assume throughout the paper the signal received at the $M$–element antenna array is given by:

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

where $T$ is the symbol period, $\mathbf{h}(t)$ is the $M$x1 vector representing the baseband equivalent channels, $h_k(t)$, $k$=1,…,$M$, of the signals received at each antenna element and $\{a_n\}$ is the desired symbol sequence. Accordingly, $\{b_{n,l}\}$ are the symbol sequence of the $l$–$th$ co–channel interferer, $l$=1,…,$L$. and $\mathbf{h_l}(t)$ corresponds to the $M$x1 vector representing the baseband equivalent channels of the interferers signals received at the array. The $M$x1 noise vector, represented by $\mathbf{n}(t)$, is assumed white and gaussian and it is uncorrelated (spatially and temporally), such that $\mathrm{E}\{\mathbf{n}(t).\mathbf{n}^{\mathrm{H}}(t-\tau)\} = \sigma^2\mathbf{I}\delta(\tau)$. A raised cosine pulse shape is used to characterize the impulse response of both the desired user and the co–channel interferers. Signal envelopes are Rayleigh distributed and time–variant during a processing interval.

## III. CONVENTIONAL SPACE–TIME EQUALIZATION

A general configuration of a conventional space–time equalizer is illustrated in Fig. 1. A space–time linear filter (STLF) with $M$ antennas and $L$ taps per antenna performs simultaneous CCI and ISI suppression. Following the STLF, a non–linear temporal equalizer (TE) mitigates residual ISI. We may employ a DFE, an MSLE equalizer or a delayed decision–feedback sequence estimator (DDFSE) equalizer [10]. The two former equalizers may be viewed as particular cases of the last one, when the MLSE memory is either zero (DFE) or equal to the total number of channel states (full state MLSE). Parameter adaptation of both the STLF and the TE is driven by a known training sequence. After the training period, hard decisions are employed as the desired signal to track small channel variations. Since the SLTF deals with both CCI and ISI, the problem of insufficient degrees of freedom may degrade performance. Furthermore, due to the imbalance in the energy of the desired and co–channel interferer signals, the STLF tend to combat ISI more. As a consequence, residual CCI at the output of the array degrades the performance of the TE. Another problem is attenuation on the desired signal due to the presence of multipaths from the desired user falling at the mainbeam of the array.

In this situation, it may be difficult for the TE to recover this loss of signal–to–interference–plus–noise ratio (SINR).

## IV. DECOUPLED SPACE–TIME EQUALIZATION

The Fig. 2 illustrates the general structure of a D–ST equalizer. A filtered version of the training sequence is employed as the desired signal to form en error signal that jointly adapts the array weights and the coefficients of a canceling filter. After the training period, the coefficients

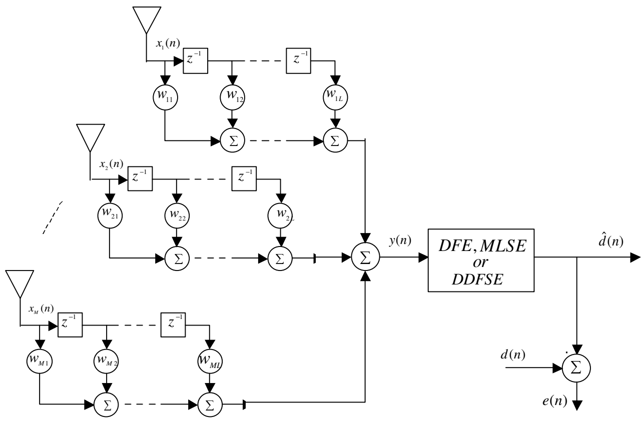


Fig.1: A conventional space–time equalizer.

of the canceling 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 interferers.

Therefore, the estimated channel impulse response is converted into the parameters of the TE following the array that performs ISI suppression.

In propagation environments with delay spread, the array benefits from the decoupling of CCI and ISI suppression since all the degrees of freedom will be available to cancel the multipaths of co–channel interferers. Furthermore, with D–ST equalization the desired signal does not suffer from undesirable attenuation when we have a small angular separation among the desired user multipaths. We first describe the adaptive array and the optimization criterion used in D–ST. Afterwards, some comments about parameter acquisition of the temporal equalizer from the canceling filter is made.

### *A. Optimization Criterion*

Let us assume an adaptive array of $M$ antennas and a FIR canceling filter of L taps. Let us call $\mathbf{w}=[w_1\ w_2 \ldots w_N]^{\mathrm{T}}$, the array weights and $\mathbf{c}=[c_1\ c_{2\ldots} c_L]^{\mathrm{T}}$ the coefficients of the canceling filter. The vector of received samples is denoted by $\mathbf{x}=[x_1\ x_2 \ldots x_N]^{\mathrm{T}}$ and $\mathbf{d} = [d_1\ d_{2\ldots} d_L]^{\mathrm{T}}$ is the vector of training symbols.

The optimum solution for both $\mathbf{w}$ and $\mathbf{c}$ maximizes the SINR at the output of the antenna array defined below:

$$\begin{aligned}(\mathbf{w}_{opt},\mathbf{c}_{opt}) &= \arg\max_{w,c}\mathrm{SINR} \\ &= \arg\max\frac{E\left\{\left\|\mathbf{w}^H\mathbf{x}\right\|^2\right\}}{E\left\{\left\|\mathbf{w}^H\mathbf{x}-\mathbf{c}^H\mathbf{d}\right\|^2\right\}}\end{aligned} \quad (2)$$

where $E\{\bullet\}$ denotes the expectation operator. It can be shown that the maximization of (2) is equivalent to the minimization of the following cost function:

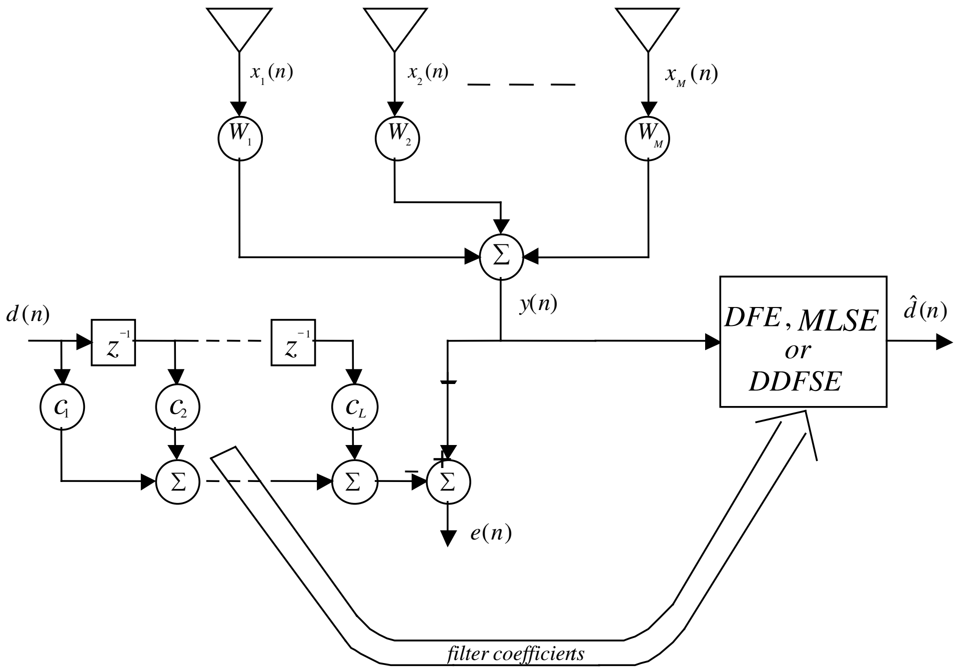


Fig. 2: The general structure of a D–ST equalizer.

$$J(\mathbf{w},\mathbf{c}) = E\left\{\left\|\mathbf{w}^H\mathbf{x} - \mathbf{c}^H\mathbf{d}\right\|^2\right\} \quad (3)$$

Indeed, (3) represents the mean square value of the error signal shown in Fig. 2. The optimization of this cost function is carried out by choosing a suitable constraint for the vector $\mathbf{c}$ in order to avoid the trivial solution. We describe two different constraints, where the first one is linear and the second one is quadratic. The two approaches for minimizing $J(\mathbf{w}, \mathbf{c})$ are described above:

*i.* min $J(\mathbf{w}, \mathbf{c})$ subject to $\mathbf{f}^T\mathbf{c}=1$, $\mathbf{f}=[0\ldots1\ldots 0]$. (4)

*ii.* min $J(\mathbf{w}, \mathbf{c})$ subject to $\|\mathbf{c}\|^2=g$. (5)

In (4), only the *j–th* element of vector $\mathbf{c}$ is different from zero and equal to 1. The value of $j$ determines the training delay of the canceling filter [8] whose performance is delay–dependent. In (5), the energy in the filter coefficients $\mathbf{c}$ is equal to some constant $g$ [11]. In this work we adopted (4) as the constraint of the canceling filter.

*B. Temporal Equalizer*

At the end of the training period, the output of the adaptive is corrupted only by ISI due to desired user multipaths that should be suppressed by the TE following the array.

The parameters of the TE are calculated from the estimated coefficients of the canceling filter that represents the overall channel impulse response. By identifying the causal and anti–causal taps in the estimated channel impulse response, we convert them into the coefficients of a feedfoward and feedback sections of a DFE [12]. The coefficients of the canceling filter can also be employed to provide channel state information to an MLSE equalizer or a DDFSE equalizer. The arrow in Fig. 2 represents the acquisition of the equalizer parameters from the coefficients of the canceling filter. Therefore, adaptation is performed in the array and in the canceling filter only. Assuming $M$=2 receiving antennas with weights $w_1$ and $w_2$ and no noise, it can be shown [11] that the mean square error is minimum if each coefficient c(k), $k$=0,…,$L$–$1$ of the canceling filter is given by:

$$\mathrm{c(k)} = \mathrm{w_1 h_1(k)} + \mathrm{w_2 h_2(k)} \quad (6)$$

where $h_1(k)$ e $h_2(k)$ are the *k–th* element of the channel impulse response at antenna 1 e 2, respectively. Therefore the overall channel impulse response seen by the TE is a linear combination of the channel impulse response perceived at each antenna.

Three D–ST equalization structures arise, namely, D–ST–DFE, D–ST–MLSE and D–ST–DDFSE. Similarly to the conventional ST equalization case, the two former D–ST equalizers are particular cases of the latter.

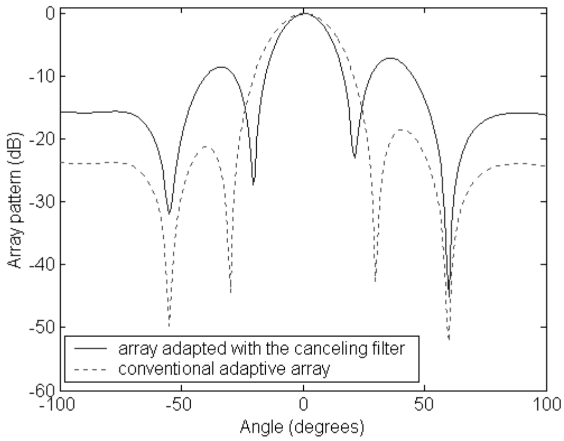

Fig. 3: Array pattern of the antenna array adapted with the canceling filter for scenario 1. All user paths are beamformed.

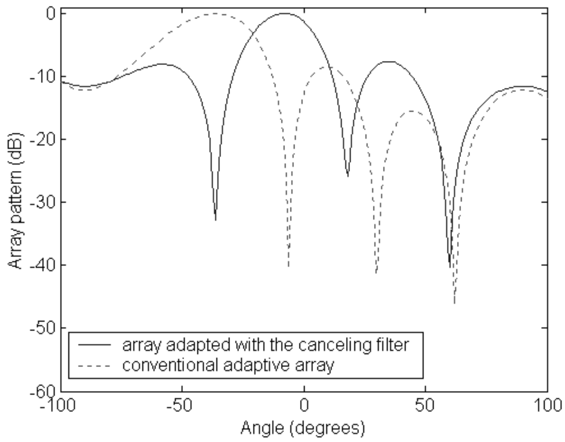

Fig. 4: Array pattern of the antenna array adapted with the canceling filter for scenario 2. The attenuation on the desired signal is avoided when we have a small angular separation.

## V. SIMULATION RESULTS

Computer simulation results are presented here to demonstrate the performance of D–ST equalization structures. We start with two illustrative scenarios. The major difference between them is that in scenario 2 there is one user multipath falling at the mainbeam of the array. The impact of using a canceling filter to adapt the array can be observed from the array pattern. This is showed in Fig. 3 and 4 for scenarios 1 and 2, respectively. We employ five antennas in scenario 1 and four antennas in scenario 2. The desired path in both scenarios is that with a direction of arrival (DOA) of 0º. The input signal–to–noise ratio (SNR) is 20dB. The canceling filter has 3 taps on both scenarios. In Fig. 3 we observe that, due to the presence of the canceling filter, the array beamforms towards all user paths, rather then suppressing them. In Fig. 4 the desired signal power is not affected by a small angular separation, since the array adapted with the canceling filter attempts to preserve all user paths.

The output SINR gain of the conventional adaptive array and the array adapted with the canceling filter is plotted in Fig. 5 and 6 as a function of the input SNR, for scenarios 1 and 2, respectively.

TABLE 1: Channel characterization for scenario 1.

| *SCENARIO 1* | DOA | Delay (T) | Gain |
|---|---|---|---|
| User paths | –30º, 0º, 30º | 0, 1, 2 | 1, 1, 0.5 |
| Interferer paths | 60º | 0 | 1 |

TABLE 2: Channel characterization for scenario 2.

| *SCENARIO 2* | DOA | Delay (T) | Gain |
|---|---|---|---|
| User paths | 0º, –6º, 30º | 0, 1, 2 | 1, 1, 0.5 |
| Interferer paths | 60º | 0 | 1 |

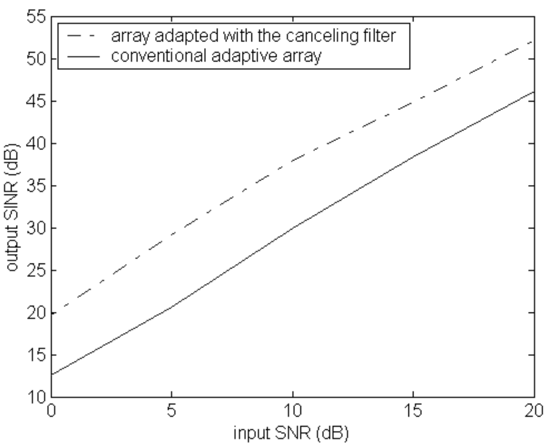

Fig. 5: SINR gain of the array adapted with the canceling filter over the conventional adaptive array, for scenario 1.

Comparing Fig. 5 and 6 it can be seen that the presence of a user multipath at the mainbeam does not affect significantly the performance of the array adapted with the canceling filter, while the conventional adaptive array tends to exhibit a floor in the output SINR for medium to high input SNR values.

Next, we evaluate the BER performance of D–ST equalizers on the Typical Urban (TU) and Hilly Terrain (HT) channel profiles [13] with CCI. A single co–channel interferer with a signal–to–interference (SIR) ratio of 0dB is considered. Both the desired user and the co–channel interferer follow the same channel profile. We assume that the mobile speed is equal to 50 Km/h. Each run represents a transmitted time–slot of 140 symbols from which 26 are for training. The pulse shaping function is a raised cosine with a roll–off factor of 35%. Besides, we use 8–PSK modulation since our system context is the Enhanced Data Rates for Global Evolution (EDGE) [14].

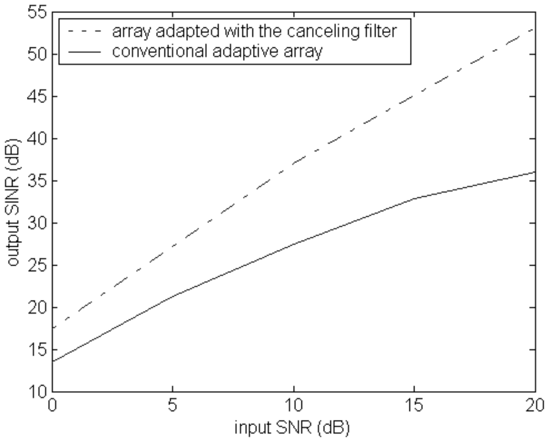


Fig. 6: SINR gain of the array adapted with the canceling filter over the conventional adaptive array, for scenario 2.

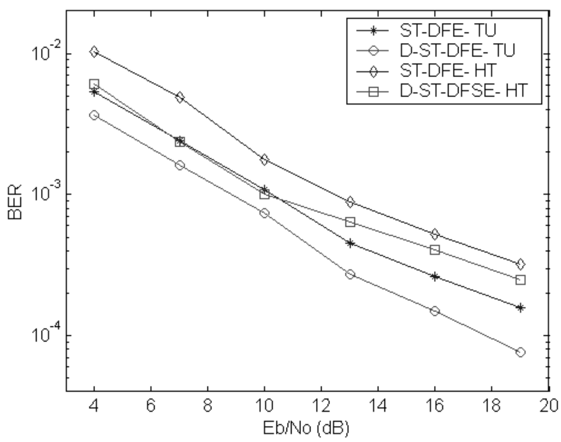


Fig. 7: BER performance of D–ST–DFE and ST–DFE on TU and HT channel profiles.

The recursive least squares (RLS) algorithm is used for adaptation. For each of the equalizer options, we compare the performance of the D–ST technique with that of its conventional counterpart with 2 taps per antenna. The canceling filter has 7 taps on both channel profiles. In Fig. 7, we observe that the performance of the D–ST–DFE is better than that of ST–DFE on TU and HT profiles. On HT, their performances approximate for high $Eb/N_0$ values. Concerning D–ST–MLSE and ST–MLSE, we work with a 1–state Viterbi equalizer in a reduce–complexity approach. In TU this is reasonable since delay spread is concentrated within 1 symbol period, however in HT we expect some performance degradation. Fig.8 shows that a considerable performance improvement of D–ST–MLSE over ST–MLSE is verified on both TU and HT profiles.

The D–ST–DDFSE and ST–DDFSE are shown in Fig. 9. Both strategies offer a good trade–off between performance and complexity [10]. The performance improvement of D–ST–DDFSE over its conventional counterpart is verified on both channel profiles, and it is more pronounced on the HT case For a target uncoded BER of $10^{-3}$ the $Eb/N_0$ gain of D–ST–DDFSE over ST–DDFSE is approximately 8dB.

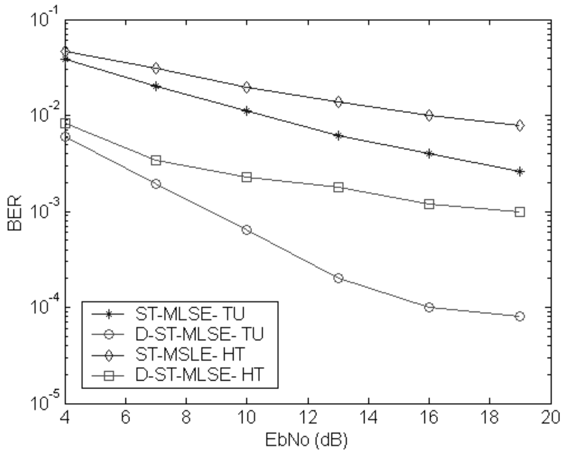


Fig. 8: BER performance of D–ST–MLSE and ST–MLSE on TU and HT channel profiles.

## VI. CONCLUSIONS

In this work we have evaluated the performance of D–ST structures based on DFE, MLSE and DDFSE. It was verified that the D–ST equalization technique is well suited for rich multipath scenarios limited by CCI and ISI. The benefits of D–ST equalizers arise from the separation of CCI and ISI suppression in two stages, allowing the antenna array to suppress only CCI, and leaving ISI to be suppressed by a temporal equalizer. Such a separation is done by adapting the array with a filtered version of the training sequence.

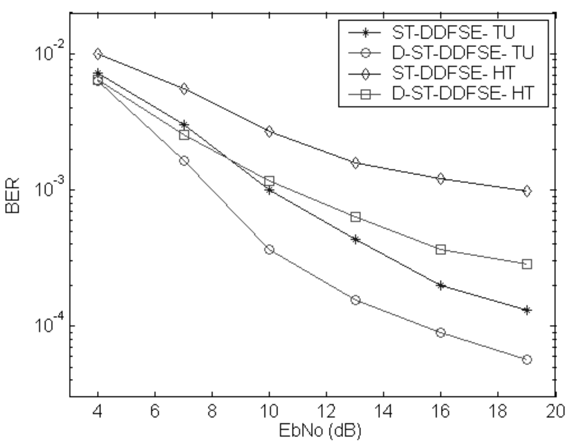


Fig. 9: BER performance of D–ST–DDFSE and ST–DDFSE on TU and HT channel profiles.

Concerning the array, we showed that this adaptation criterion provides a higher output SINR, while the temporal equalizer leads to more reliable decisions and a better BER performance in practical channels as in TU and HT.

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