DISTRIBUTED SPACE-TIME CODING STRATEGIES FOR WIDEBAND MULTIHOP NETWORKS: REGENERATIVE VS. NON-REGENERATIVE RELAYS

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ABSTRACT

Distributed space-time coding (DSTC) is a rather novel paradigm that merges ideas from space-time coding (STC) and multihop networks (MHN) to design a wireless network capable of improving the performance considerably with respect to single hop networks (SHN). The basic advantage of DSTC comes from allowing multiple nodes to share their antennas to create a *virtual* transmit array and then implement a distributed space-time coding technique over the virtual array. The major differences between DSTC and conventional STC are: i) detection errors at the relay nodes and ii) possible lack of synchronization between source and relay nodes. In this work, we study these problems and compare different DSTC techniques based on decode and forward and amplify and forward strategies. Finally, we show the trade-off curves between rate and diversity gain for DSTC systems.

1. INTRODUCTION

Cooperation among users in a wireless network can increase the capacity in the uplink multiuser channel [10]. Cooperation was proved to be very useful to combat shadowing effects, as shown in [3], and it can occur in several different ways, as suggested in many recent works, like e.g. [7], [1], [11], [5], [8], [4]. In [6], [7], different cooperation strategies were compared. The coding strategies proposed in [6], [7] were essentially based on time repetition coding. Clearly, this implies a loss in terms of rate. A different approach to obtain the maximum diversity gain, without suffering from the rate loss of repetition coding, consists in merging the idea of cooperation with space-time coding, giving rise to the so called *distributed* space-time coding, where source (S) and relay (R) share their antennas to create a *virtual* transmit array to transmit towards the destination (D). This idea was studied in detail by Anghel, Leus, and Kaveh, as in [1] for example. The overall channel between S and D, including the S-R channel, was supposed to be, equivalently, a Rayleigh fading channel. Differently from [1], in this work we incorporate explicitly the two major issues that make DSTC different from conventional STC: i) the decoding errors in the R node and ii) the different arrival times of the packets arriving at D from S or R. We consider both cases where the relay acts as a regenerative system or simply as a repeater. In the first case we talk about decode and forward (D&F), whereas in the second case we have an amplify and forward (A&F) scheme. The D&F scheme has, potentially, better performance than A&F, but A&F is much simpler to implement, because it consists only of the RF section (essentially antenna and amplifier). We will also compare the performance of the maximum likelihood decoder, for the D&F scheme, in case of BPSK transmission, with sub-optimal but simpler detectors. With respect to the flooding method underlying the Opportunistic Large Array idea recently proposed in [8], we consider here a coordinated strategy between S and R's, so as to get full spatial diversity gain.

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2. DISTRIBUTED SPACE-TIME BLOCK CODING

We consider a scenario based on the following assumptions: (a1) all channels are FIR of maximum order L_h and time-invariant over at least a pair of consecutive blocks; (a2) the channel coefficients are zero mean complex Gaussian random variables; (a3) the transmission scheme for all terminals is a block strategy, where each information block $\boldsymbol{s}(n)$ is composed of M symbols and it is linearly encoded, so as to generate the N-size vector $\boldsymbol{x}(n) := \boldsymbol{F}\boldsymbol{s}(n)$, where $\boldsymbol{F}$ is the $N \times M$ precoding matrix - a CP of length L is inserted at the beginning of each block, to facilitate elimination of inter-block interference and channel equalization at the receiver - L is chosen equal to $L_h + L_d$, where L_d is the delay between the time of arrivals of packets arriving at D from S and from R; (a4) the information symbols are i.i.d. BPSK and each symbol may assume the values A or $-A$ with equal probability[1]; (a5) the received data are degraded by additive white Gaussian noise (AWGN), denoted by the vector $\boldsymbol{w}$; (a6) the radio nodes are quasi-synchronous. We assume that each terminal is equipped with one antenna, but the extension to multiple receive antennas is straightforward [9]. In the present setup, since D has only one antenna, there cannot be any spatial multiplexing gain ad thus we concentrate on diversity gain. Throughout the paper, we use the following notation: $h_{sd}(i)$, $h_{sr}(i)$, and $h_{rd}(i)$, are the impulse responses between S and D, S and R, and R and D, respectively; $\boldsymbol{H}_{sd}$, $\boldsymbol{H}_{sr}$ and $\boldsymbol{H}_{rd}$ are $N \times N$ circulant Toeplitz matrices, with entries $H_{sd}(i,j) = h_{sd}((i-j) \mod N)$, $H_{sr}(i,j) = h_{sr}((i-j) \mod N)$ and $H_{rd}(i,j) = h_{rd}((i-j) \mod N)$, respectively.For the sake of clarity, the DSTC protocol that we propose here is a distributed version of the block Alamouti scheme [2], but other STC techniques may be chosen instead, depending on the desired trade-off between diversity gain and transmission rate. Furthermore, we adopt an OFDM strategy to simplify the analysis, but other linear precoding strategies may be used, as suggested in [9]. The transmission protocol is sketched in Fig. 1. We consider a pair of consecutive time slots (TS) where, in the first TS, S transmits and R (and possibly D) receives; in the second TS, S and R trans-

[1] This assumption is made only for simplifying our derivations, but there is no restriction to use higher order constellations.

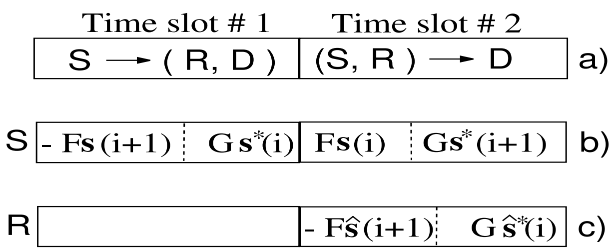

Fig. 1. Structure of transmit slots: a) time slots and block indices; b) information blocks sent by S; c) information blocks sent by R.

mit together and D receives. Within each time slot, we transmit two blocks of symbols according to the following strategy. During the first TS, S sends, consecutively, the blocks $-A_s\boldsymbol{F}\boldsymbol{s}(i+1)$ and $A_s\boldsymbol{G}\boldsymbol{s}^*(i)$[2] In the second TS, S transmits, consecutively, $A_s\boldsymbol{F}\boldsymbol{s}(i)$ and then $A_s\boldsymbol{G}\boldsymbol{s}^*(i+1)$. The node R may operate as an A&F or as a D&F node, as detailed next.

Amplify & Forward: In the A&F case, R simply amplify the received signal and retransmits the received blocks. This means that R, in the second time slot, transmits $A_r(-\boldsymbol{H}_{sr}\boldsymbol{F}\boldsymbol{s}(i+1)+\boldsymbol{v}_R(i))$ first and then $A_r(\boldsymbol{H}_{sr}\boldsymbol{G}\boldsymbol{s}^*(i)+\boldsymbol{v}_R(i+1))$.

Decode & Forward: In the D&F case, R decodes the received vectors and provides the estimated symbol vectors $\hat{\boldsymbol{s}}(i)$ and $\hat{\boldsymbol{s}}(i+1)$. Then, it transmits the blocks $-A_r\boldsymbol{F}\hat{\boldsymbol{s}}(i+1)$ first and then $A_r\boldsymbol{G}\hat{\boldsymbol{s}}^*(i)$.

The amplitude coefficients A_s and A_r are used to impose the power available at the S and R nodes, respectively. In the A&F case, differently form the D&F case, the coefficient A_r depend also on the S-R channel as well as on the noise at the R node. Clearly, A_r changes depending on which strategy is implemented in R. What is important to remark is that, whichever is the R strategy, in the second time slot, S and R transmit simultaneously.
For the sake of simplicity, we consider here an OFDM transmission strategy where $\boldsymbol{F}=\boldsymbol{W}$ and $\boldsymbol{G}=\boldsymbol{W}$, where $\boldsymbol{W}$ is the DFT matrix, with entries $W_{kl}=e^{j2\pi kl/N}/\sqrt{N}$. Thanks to the combination of Alamouti's coding and OFDM, the overall systems is equivalent to a series of N parallel channels. For each sub-carrier, we have a system of two equations in two unknowns. More specifically, let us denote with $\boldsymbol{r}_k(i):=[r_k(i),r_k(i+1)]^T$, the two-components vectors of received symbols pertaining to the k-th sub-carrier, in the i-th time slot. We introduce also the vectors $\boldsymbol{s}_k(i):=[s_k(i),s_k(i+1)]^T$, $\hat{\boldsymbol{s}}_k(i):=[\hat{s}_k(i),\hat{s}_k(i+1)]^T$ and $\boldsymbol{w}_k:=[w_k(i),w_k(i+1)]^T$, referring to the k-th sub-carrier, with $k=0,\ldots,N-1$. Hence, in the D&F case, we can write (we drop the block index i for simplicity of notation, because the same relationships hold true for all blocks)

$$\begin{aligned}\boldsymbol{r}_k = \; & \alpha_1\begin{bmatrix}|H_{sd}(k)|^2 & -H_{sd}^*(k)H_{rd}(k)\\ H_{sd}(k)H_{rd}^*(k) & |H_{sd}(k)|^2\end{bmatrix}\boldsymbol{s}_k\\ & + \alpha_2\begin{bmatrix}|H_{rd}(k)|^2 & H_{sd}^*(k)H_{rd}(k)\\ -H_{sd}(k)H_{rd}^*(k) & |H_{rd}(k)|^2\end{bmatrix}\hat{\boldsymbol{s}}_k+\boldsymbol{w}_k. \quad (1)\end{aligned}$$

where $H_{sr}(k)=\sum_{l=0}^{L_h-1}h_{sr}(l)\,\mathrm{e}^{-j2\pi\,lk/N}$, $H_{sd}(k)=\sum_{l=0}^{L_h-1}h_{sd}(l)\,\mathrm{e}^{-j2\pi\,lk/N}$ and $H_{rd}(k)=\sum_{l=0}^{L_h-1}h_{rd}(l)\,\mathrm{e}^{-j2\pi\,lk/N}$; α_1 and α_2 incorporate all the amplitude coefficients on the S-D and on the R-D paths. The set of equations (1) reduces to the well known Alamouti scheme, when there are no decision errors at the relay, i.e. $\hat{\boldsymbol{s}}_k=\boldsymbol{s}_k$ and $\alpha_1=\alpha_2$. In such an ideal case, in fact, setting $\alpha_1=\alpha_2=\alpha$, (1) reduces to

$$\boldsymbol{r}_k=\alpha\begin{bmatrix}|H_{sd}(k)|^2+|H_{rd}(k)|^2 & 0\\ 0 & |H_{sd}(k)|^2+|H_{sd}(k)|^2\end{bmatrix}\boldsymbol{s}_k+\boldsymbol{w}_k. \quad (2)$$

The noise vector $\boldsymbol{w}$ has independent components and its covariance matrix $\boldsymbol{C}_w$ has diagonal elements equal to $C_w(k)=\sigma_d^2\,(|H_{sd}(k)|^2+|H_{rd}(k)|^2)$. Clearly, (1) reduces to (2) when the SNR at the relay tends to infinity, i.e. when there are no decision errors at the relay.

In the A& F case, proceeding as before, we end up with the following relationships:

$$\boldsymbol{r}_k=\begin{bmatrix}g_k & 0\\ 0 & g_k\end{bmatrix}\boldsymbol{s}_k+\boldsymbol{\nu}_k, \quad (3)$$

where $g_k:=A_s^2|H_{sd}(k)|^2+A_r^2|H_{rd}(k)|^2|H_{sr}(k)|^2$ and $\boldsymbol{\nu}_k$ is a Gaussian vector with zero mean and diagonal covariance matrix $\boldsymbol{C}_\nu=\sigma_\nu^2\boldsymbol{I}$, with $\sigma_\nu^2=(A_r^2|H_{rd}(k)|^2\sigma_r^2+\sigma_d^2)\,(A_s^2|H_{sd}(k)|^2+A_r^2|H_{rd}(k)|^2|H_{sr}(k)|^2)$.

Since S and R are not co-located, the blocks transmitted from S and R arrive at D at different times. This is a specific difference of DSTC with respect to STC. However, if the difference in arrival times τ_d is incorporated in the CP used from both S and R, D is still able to get N samples from each received block, without inter-block interference (IBI). In such a case, the different arrival time does not cause any trouble to the final receiver. In fact, let us take as a reference starting time the time of arrival of the i-th block from R. If the block coming from S arrives with a delay of L_d samples, the only difference with respect to the case of perfect synchronization, thanks to OFDM, is that the transfer function $H_{sd}(k)$ in (1) will be substituted by the matrix $H_{sd}(k)e^{-j2\pi L_d k/N}$. From (1), it is clear that such a substitution will not affect at all the useful term, but it will only affect the interfering term. However, in the hypothesis of Rayleigh fading channel, $H_{sd}(k)$ is a circularly symmetric random variable, so that any rotation is not going to affect its statistical properties. Hence, the combination of Alamouti (more generally, orthogonal STC) and OFDM is robust with respect to lack of synchronization between the time of arrivals of packets from S and from R. The price paid for this robustness is the use of a CP of length L_d+L_h, instead of just L_h.

2.1. ML detector for D&F systems

We derive now the structure of the maximum likelihood (ML) detector for D&F systems adopting BPSK constellations. We denote with $\mathcal{S}$ the set of all possible transmitted vectors $\boldsymbol{s}_k$ and with $p_{e1}(k)$ and $p_{e2}(k)$ the error probabilities on $s_k(1)$ and $s_k(2)$, respectively, conditioned to a given channel realization, at the relay node. After detection, at node R, we have $\hat{s}_k(l)=s_k(l)$, with probability $(1-p_{el}(k))$, and $\hat{s}_k(l)=-s_k(l)$ with probability $p_{el}(k)$, $l=1,2$. To simplify the expression of the likelihood function, we normalize the received vector $\boldsymbol{r}_k$ so that the resulting noise vector $\boldsymbol{v}_k$ had independent components with unit variance. We introduce the vector $\boldsymbol{z}_k=\boldsymbol{r}_k/(\alpha\sigma_n\sqrt{|H_{sd}(k)|^2+|H_{rd}(k)|^2})$.

[2]The reason for using such a sequence will be clear when we will consider the A&F scheme.

Since the symbols are independent, the probability density function of the vector $\boldsymbol{z}_k$, conditioned to having transmitted $\boldsymbol{s}_k$, is (see [9] for the details):

$$
\begin{aligned}
f_{\boldsymbol{z}_k|\boldsymbol{s}_k}(\boldsymbol{z}|\boldsymbol{s}_k) = \tfrac{1}{\pi^2}& \\
\cdot \left[(1-p_{e1}(k))(1-p_{e2}(k))\right.& \exp\left\{-|\boldsymbol{z}-\boldsymbol{A}_k(1,1)\boldsymbol{s}_k|^2\right\} \\
+p_{e1}(k)p_{e2}(k)& \exp\left\{-|\boldsymbol{z}-\boldsymbol{A}_k(-1,-1)\boldsymbol{s}_k|^2\right\} \\
+(1-p_{e1}(k))p_{e2}(k)& \exp\left\{-|\boldsymbol{z}-\boldsymbol{A}_k(1,-1)\boldsymbol{s}_k|^2\right\} \\
+p_{e1}(k)(1-p_{e2}(k))& \left.\exp\left\{-|\boldsymbol{z}-\boldsymbol{A}_k(-1,1)\boldsymbol{s}_k|^2\right\}\right], \quad (4)
\end{aligned}
$$

where we have introduced the matrix $\boldsymbol{A}_k(p,q)$ defined as follows (we drop the dependence by the index k)

$$
\boldsymbol{A}_k(p,q) = \begin{bmatrix} \alpha_1|H_{sd}|^2+\alpha_2|H_{rd}|^2 p, & \alpha_2 H_{sd}^* H_{rd} q - \alpha_1 H_{sd}^* H_{rd} \\ \alpha_1 H_{sd} H_{rd}^* - \alpha_2 H_{sd} H_{rd}^* p, & \alpha_1|H_{sd}|^2+\alpha_2|H_{rd}|^2 q \end{bmatrix}
$$

The ML detector is then

$$
\hat{\boldsymbol{s}}_k = \arg\max_{\boldsymbol{s}_k \in \mathcal{S}} \left\{ f_{\boldsymbol{r}_k|\boldsymbol{s}_k}(\boldsymbol{r}_k|\boldsymbol{s}_k) \right\}, \quad (5)
$$

with $f_{\boldsymbol{z}_k|\boldsymbol{s}_k}(\boldsymbol{z}_k|\boldsymbol{s}_k)$ as given in (4). The ML detector (5) assumes the knowledge of the vector of error probabilities $p_{e_1}(k)$ and $p_{e_2}(k)$, $k=0,\ldots,N-1$, occurring at the relay node. This requires an exchange of information between R and D. This information has to be updated with a rate depending on the channel coherence time. An alternative, although sub-optimum, symbol-by-symbol detection scheme that does not require such a knowledge can be set up simply assuming that the decision errors at the relay are negligible and thus approximating (1) with (2). In such a case, it is not necessary to know the relay error probabilities at the D node. We will show later on the difference in performance between these two alternatives.

In the detection schemes seen so far, D processes only the vectors received in the time slot #2, and ignores the vectors received in the time slot #1. However, if D is in a listening mode also during the first time slot, we have shown in [9] how an appropriate combination of the vectors received over all the time slots improves the performance of the system.

3. PERFORMANCE AND CONCLUSION

We evaluated the performance of our proposed schemes in terms of average BER and information rates. To make a fair comparison between single-hop and multihop schemes, we assume that the overall transmit power is always the same.

Example 1. Comparison of different strategies: We show now some performance results, obtained using the following parameters: The block length is $N=16$; the channel are simulated as FIR filters of order $L_h=6$, whose taps are iid complex Gaussian random variables with zero mean and variance $1/d^2$, where d is the distance in the link. These distances are normalized with respect to the S-D distance, so that $d_{sd}=1$, $d_{sr}=0.5$, $d_{sd}=0.5$. In Fig. 2, we compare, in a slow fading environment, the average BER vs. the SNR_D at the destination node, obtained using the following strategies: a) Decode and forward (blue lines) using ML detector (solid line) or sub-optimal detector: theoretical value (dashed line) found in [9] and simulation results (circles); b) amplify and forward (red line); c) single hop (non-cooperative) case (black line). The SNR_R at the relay is equal to 20 dB. The sub-optimal detector, in the D&F case, is obtained by simply taking the sign of the real part of the vector $\boldsymbol{r}_k$ in (1). We can observe the floor on

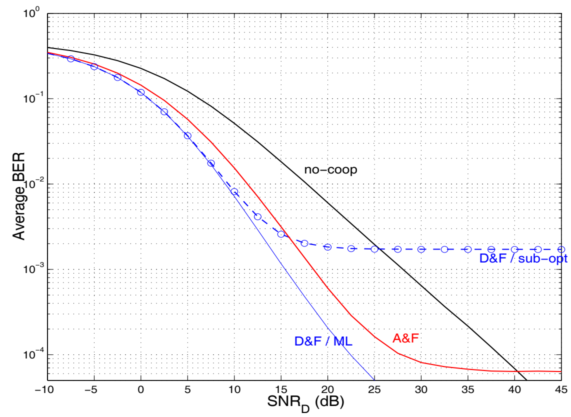

Fig. 2. Average BER vs. SNR_D (dB) achieved with different strategies: a) Decode and forward (blue lines) using ML detector (solid line) or sub-optimal detector: theoretical value (dashed line) and simulation results (circles); b) amplify and forward (red line); c) single hop (non-cooperative) case (black line); $SNR_R = 20$ dB.

the BER of the sub-optimum receiver for the D&F scheme. That floor is due to the errors at the relay node. It is also evident the diversity gain achieved with all the cooperative schemes with respect to the non-cooperative schemes. In fact, we notice that all cooperative curves exhibit a behavior of the average BER, at high SNR, proportional to $1/SNR^2$ (at least for SNR_D values below the values giving rise to the floor), whereas in the non-cooperative case we observe a behavior proportional to $1/SNR$. It is also interesting to notice that, in the D&F case, the sub-optimum detector exhibits performance very close to the optimal ML detector, for average BER values greater than the floor. This indicates that the sub-optimum detector is indeed a good choice because it is certainly less complex to implement than the ML detector and, most important, it does not require any exchange of information about the BER between R and D. The price paid for this simplicity is that the R node must have a sufficiently high SNR to guarantee that the final BER be above the floor. Finally, comparing the D&F and A&F schemes, we observe that the D&F method performs better than the A&F at low and intermediate SNR_D values, but for high values of SNR_D, the A&F performs better. This shows that A&F is indeed a valuable choice.

Example 2. Rate-diversity trade-off for the D&F scheme: We have showed that our system achieves maximum spatial diversity if the detection errors at the relay can be neglected [9]. The price paid for this cooperation diversity gain is the rate loss, induced by the insertion of the time slots necessary to let S to send its data to R. If all the links use a BPSK constellation, the loss factor is $1/2$. To reduce this loss factor, we can use higher order constellations on the S-R link, for a given constant symbol duration. In such a way, using a constellation $\mathcal{A}$ of cardinality $|\mathcal{A}|$, the loss factor is $1/(1+1/\log_2(|\mathcal{A}|)) := 1/(1+1/m)$, having set $m=\log_2(|\mathcal{A}|)$. On the other hand, cooperation increases the final SNR and then it induces a rate increase. To quantify the overall balance in terms of rate, we compare the maximum rates achievable with our double-hop system and with a single-hop system. We define as achievable rate the maximum number of bits per symbol (bps) that can be detected with an arbitrarily low error probability, provided that sufficient error correction coding is incorporated in the system, conditioned to the assumptions (a1)-(a6)[3]. Thanks to the combination of orthogonal space-time coding and OFDM, the block transmission over dispersive channels is equivalent to the transmission over parallel non-dispersive sub-channels (sub-carriers). The equivalent S-D link, over each sub-channel, in the presence as well as in the absence of the relay link, is a symmetric BPSK channel with crossover probability given by the bit error probability on that sub-carrier, conditioned to the channel realization. Thus, the maximum number $R(k|\boldsymbol{h})$ of bits/symbol that can be transmitted on the k-th sub-carrier reliably (using sufficiently long error correction codes) for a given channel realization $\boldsymbol{h}$, is[4] $R(k|\boldsymbol{h}) = \frac{1}{1+1/m}\,\mathrm{C_{BSC}}(\mathrm{P_{e|h}(k)})$ bps, where $P_{e|h}(k)$ denotes the binary error probability on the k-th sub-carrier, conditioned to the channel realizations, and $C_{BSC}(p) := 1 + p\log_2(p) + (1-p)\log_2(1-p) := 1 - H(p)$ is the capacity of the (equivalent) binary symmetric channel[5] with crossover probability p.

As an example, we report in Fig.3 the achievable rate vs. the SNR_D in D, for an SNR_R in R equal to 15 dB, for different choices of the constellation used in the S-R link, achieved with or without cooperation. We can see that, at high SNR_D, the non-cooperative case approaches the maximum value, equal to 1 bps, whereas the cooperative cases tend to an asymptote less than 1, depending on the adopted constellation. We observe from Fig.3 that, for $SNR_R = 15$ dB, for practical values of SNR_D, increasing the constellation order from BPSK to 16-QAM in the S-R link improves the achievable rate. However, a further increase of the order, from 16-QAM to 64-QAM, does not induce any appreciable gain because of the higher BER at the relay. Nevertheless, it is interesting to remark that, at lower/medium SNR_D at the final destination (within a range depending on SNR_R), the cooperative case can outperform the non-cooperative case also in terms of rate, because the diversity gain induces an SNR_D increase that more than compensates the rate loss due to the exchange of data between S and R.

In conclusion, in this work we have shown that a distributed space-time coding scheme using different nodes to build a virtual transmit array can be very effective to induce a useful diversity gain. Comparing A&F and D&F strategies, we have shown that A&F exhibits a very good balance in terms of simplicity and performance. Interestingly, even though our system was designed only to maximize the space diversity gain, we have shown that it can induce some improvement in terms of achievable rates, at low and moderate SNR. In a parallel work we have investigated DSTC techniques, valid for destination nodes having multiple antennas, where, instead of diversity, we exploit the transmit virtual array in order to get a spatial multiplexing gain. We are currently investigating the combination of relaying strategies with space-time coding schemes that are flexible enough to make available the best trade-off between spatial multiplexing gain and diversity gain.

[3] It is important to remark that the rate defined above is smaller than the capacity of the system, because the proposed scheme is designed to maximize the spatial diversity gain and not to maximize information rate.

[4] In the balance we have considered also the slots i-th and $(i+1)$-th dedicated to S-R communication.

[5] We can use this formula because the S-D is always BPSK, regardless of the constellation used in the S-R link.

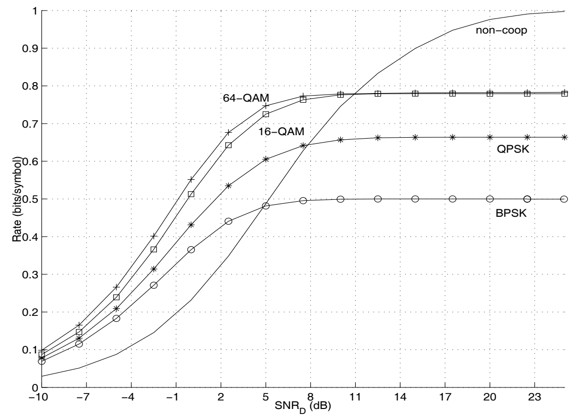

Fig. 3. Achievable rate (bps) vs. SNR (dB) - non cooperative case (solid line), cooperative case using: BPSK (circle marker), QPSK (star marker), 16-QAM (square marker), 64-QAM ('+'), in the S-R link; $SNR_R = 15$ dB.

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