Interference Subspace Rejection in Wideband CDMA: Modes for Mixed-Power Operation*

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Abstract— **Multiuser CDMA detectors suppress interference to provide improved noise immunity, increased capacity, higher data rates and reduced precision requirements for power control. A new multiuser detector structure is formulated which offers a number of implementation modes ranging in performance from that of interference cancellation (IC) detectors to that of linear receivers, yet provides more attractive performance/complexity tradeoffs. It exploits both space and time diversities as well as the array-processing capabilities enabled by multiple antennas and carries out simultaneous channel and timing estimation, signal combining and interference rejection. The improved performance enables increased utilization of wideband CDMA networks, especially at high data rates.**

I. Introduction

Third generation wireless systems will deploy wideband CDMA [1] access technology to achieve data transmission at variable rates with different mobility and quality of service (QoS) requirements. Standards [1] call for increasing the transmission rate from the 14.4 Kbps voice rate currently supported up to 384 Kbps for mobile users and 2 Mbps for portable terminals. Current industrial concerns are how to provide such multi-rate services in the broadband channels of 5 to 15 MHz likely to become available. A significant improvement in spectrum efficiency stands out as the key issue.

The call capacity of wireless CDMA systems is limited by the so-called near-far situations resulting from some highly interfering transmissions to/from other mobiles within and outside the cell. In these so called near-far problem cases, we can improve the transmission quality or reduce the transmitted power by reducing the interference. In turn, for the same transmission quality, the number of calls supported within the cell may be increased, resulting in improved spectrum utilization.

Power control is presently used to minimize the near-far problem, but with limited success. It requires frequent power control updates, typically 800 times per second, to reduce the power mismatch between the lower-rate and higher-rate users. Even tighter power control with twice the number of updates is expected in future CDMA systems, but the near-far problem will not be completely eliminated. Multiuser detectors [2]-[8] introduce interference suppression to provide potential benefits such as improvements in capacity and reduced precision requirements for power control. However, these detectors may not be cost-effective to build with a sufficient performance gain over present day systems [2]. Reaching a satisfactory performance/complexity tradeoff remains a prime concern.

In the formulation proposed here, we upgrade the spatio-temporal array-receiver (STAR) [9], a single-user receiver, by incorporating multiuser detection by interference subspace rejection (ISR) [10]-[12] at the signal combining step. The upgraded multiuser receiver STAR-ISR offers different modes that improve in performance between IC detectors and linear receivers, and require increasing complexity for implementation. At the low end, STAR-ISR reconstructs interferences from channel and data hard-decision estimates, then suppresses them like IC methods. ISR avoids the error-sensitive subtraction and implements instead a more near-far resistant linearly-constrained filtering derived with the aid of hard-decision feedback at a complexity comparable to IC. Compared to the linear receivers at the high end, STAR-ISR implements nulling along different interference subspace decompositions with much-reduced complexity. It fully exploits both space and time diversities as well as the array-processing capabilities of multiple antennas while carrying out simultaneous channel and timing estimation, signal combining and interference rejection.

II. Formulation and Background

We consider the uplink of an asynchronous cellular CDMA system where each base-station is equipped with a receiving antenna-array of M sensors. Ability to process asynchronous transmissions will be better demonstrated in [11] while application to the downlink will be addressed in [12]. For the sake of simplicity, we assume for now that all users transmit with the same modulation and at the same rate. We also assume that the base-station knows the spreading codes of all the terminals with which it communicates. The BPSK bit sequence for a mobile with index u is first differentially encoded at the rate $1/T$, where T is the bit duration. The resulting DBPSK sequence

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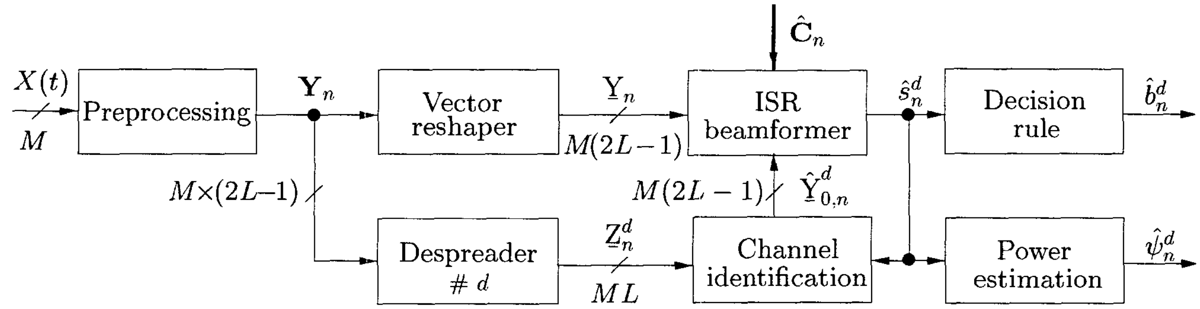

Fig. 1. Block diagram of one of the receiver modules (for the desired user) of upgraded multi-user STAR-ISR.

$b^u(t)$ is then spread by a personal PN code $c^u(t)$ at a rate $1/T_c$, where T_c is the chip pulse duration. The processing gain is given by $L = T/T_c$. We assume the use of long codes. Finally, we assume a multipath fading environment with P resolvable paths where the delay spread $\Delta\tau$ is small compared to the bit duration (*i.e.*, $\Delta\tau \ll T$).

At time t, the observation vector received by the antenna array of one particular cell, shown in Fig. 1, can be written as follows:

$$X(t) = \sum_{u=1}^{U} X^u(t) + N^{\mathrm{th}}(t) \ , \tag{1}$$

where U is the total number of mobiles received at the selected base-station from inside and outside the cell, $X^u(t)$ is the received signal vector from the mobile u, and $N^{\mathrm{th}}(t)$ is the thermal noise received at each antenna element. The preprocessing unit in Fig. 1 successively implements matched-pulse filtering, sampling at the chip rate and framing over $2L-1$ chip samples[1] at the bit rate, to yield the $M \times (2L-1)$ matched-filtering observation matrix:

$$\mathbf{Y}_n = \sum_{u=1}^{U} \mathbf{Y}_n^u + \mathbf{N}_n^{\mathrm{pth}} \ , \tag{2}$$

where each user u contributes its user-observation matrix $\mathbf{Y}_n^u$, and where the base-band preprocessed thermal noise (*i.e.*, after matched-pulse filtering) contributes $\mathbf{N}_n^{\mathrm{pth}}$.

Due to asynchronism and multipath propagation, each user-observation matrix carries information from the current as well as from the previous and future symbols of the corresponding user. We therefore have:

$$\mathbf{Y}_n^u = s_n^u \mathbf{Y}_{0,n}^u + s_{n-1}^u \mathbf{Y}_{-1,n}^u + s_{n+1}^u \mathbf{Y}_{+1,n}^u \ , \tag{3}$$

where $s_n^u = \psi_n^u b_n^u$ denotes the signal component of user u and where the canonic user-observation matrices $\mathbf{Y}_{k,n}^u$, respectively for $k = -1, 0, +1$, stand for the delayed, current and advanced versions of the normalized propagation channel $H^u(t)$ spread by $c^u(t)$. The total received power $\psi^u(t)^2$ that holds the normalization factor in s_n^u is affected by path-loss, Rayleigh fading and shadowing. We assume that both $H^u(t)$ and $\psi^u(t)^2$ vary slowly and are constant over the bit duration T.

We divide the mobiles received at a base-station into two subsets, one comprising those whose received signal powers are relatively high and a second whose powers are relatively low. Power mismatch (*i.e.*, near-far situations) arises on the uplink due to imperfect power control for path-loss and shadowing variations or when we intentionally increase the power of particular users (*e.g.*, "priority links", acquisition, higher-order modulations or higher data-rates in mixed-rate traffic). For now, we assume that all users transmit with the same modulation at the same rate (see simulations later). To receive the low-power users adequately, we attempt to eliminate the interference produced by the high-power users. For simplicity, we assume that the high-power users can be received adequately without interference suppression.

Let us assume the presence of NI strong interfering mobiles with indices $i = 1, \ldots, NI$. With respect to the desired user, assigned the index d, we can now rewrite the vector-reshaped (see Fig. 1) matched-filtering observation matrix before despreading as:

$$\underline{Y}_n = \underline{Y}_{0,n}^d s_n^d + \underline{I}_{\mathrm{ISI},n}^d + \underline{I}_n + \underline{N}_n \ , \tag{4}$$

where the first canonic matched-filtering observation vector $\underline{Y}_{0,n}^d$ appears as the "spread channel" vector of the desired user. The total interference vector before despreading:

$$\underline{I}_n = \sum_{i=1}^{NI} \underline{Y}_n^i = \sum_{i=1}^{NI} \left\{ s_n^i \underline{Y}_{0,n}^i + \underline{I}_{\mathrm{ISI},n}^i \right\} \ , \tag{5}$$

is the sum of the interfering signal vectors $\underline{Y}_n^i$ and:

$$\underline{I}_{\mathrm{ISI},n}^u = s_{n-1}^u \underline{Y}_{-1,n}^u + s_{n+1}^u \underline{Y}_{+1,n}^u \ , \tag{6}$$

[1]The number of chips is fixed here to $2L-1$ to yield despread observation matrices $\mathbf{Z}_n$ with dimension $M \times L$, as initially required for channel identification by STAR [9] (see Fig. 1). This dimension can be reduced, but we omit the discussion for simplicity.

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is the ISI vector from user u. Note that the ISI vectors correspond to the symbols from all users that lie at both edges of the processed frame. They give rise to an edge effect which can be overcome by the constraints introduced below [10]. In large-processing-gain situations, the self-ISI vector $\underline{I}^d_{\mathrm{ISI},n}$ can be combined for simplicity with the uncorrelated spatio-temporal noise vector $\underline{N}_n$, leading to the following data vector model before despreading:

$$\underline{Y}_n = \underline{Y}^d_{0,n} s^d_n + \underline{I}_n + \underline{N}_n \; . \tag{7}$$

In previous work [9] we assumed that the spatio-temporal noise vector $\underline{I}_n$ is spatially uncorrelated and merged it into $\underline{N}_n$. Hence, the signal component of the desired user could be extracted by simple spatio-temporal maximum ratio combining (MRC) or matched beamforming in a single-user receiver structure as follows[2]:

$$\hat{s}^d_n = \mathrm{Real}\left\{ \underline{W}^{d\,H}_n \underline{Y}_n \right\} = \mathrm{Real}\left\{ \frac{\hat{\underline{Y}}^{d\;H}_{0,n} \underline{Y}_n}{\|\hat{\underline{Y}}^d_{0,n}\|^2} \right\} \; . \tag{8}$$

However, due to the near-far situations explained earlier, power control on the uplink is no longer sufficient to equalize the received signal powers and the uncorrelated noise assumption becomes untenable. We next introduce a near-far resistant solution for the desired users that rejects these additional interference terms in $\underline{I}_n$.

III. Interference Subspace Rejection

So far, we used the single-user STAR to receive both the interferers and the desired users independently of each other. While the matched beamformer of (8) is optimal in uncorrelated white noise, it is suboptimal when receiving the desired users due to spatio-temporal correlation of the interference terms. To allow the accommodation of additional users in the presence of much stronger interfering mobiles in the target cell, we upgrade STAR to enhance its near-far resistance by constraining the beamformer of (8) to reject the total interference $\underline{I}_n$ in (7).

In the general case, the total interference $\underline{I}_n$ is an unknown random vector which lies at any moment in an interference subspace spanned by a matrix, say $\mathbf{C}_n$ (*i.e.*, $\underline{I}_n \in \mathrm{Vec}\{\mathbf{C}_n\}$) with dimension depending on the number of interference parameters (*i.e.*, power, data, multipath components and delays) estimated separately. As discussed next, a number of alternative techniques are available to construct the constraint matrix $\mathbf{C}_n$. To achieve near-far resistance, the beamformer must conform to the following theoretical constraints (see Fig. 1):

$$\begin{cases} \underline{W}^{d\,H}_n \underline{Y}^d_{0,n} = 1 \; , \\ \underline{W}^{d\,H}_n \mathbf{C}_n = 0 \; , \end{cases} \Rightarrow \begin{cases} \underline{W}^{d\,H}_n \underline{Y}^d_{0,n} = 1 \; , \\ \underline{W}^{d\,H}_n \underline{I}_n = 0 \; , \end{cases} \tag{9}$$

[2]Other operations that, 1) estimate the data symbols and the power of the desired user, and 2) identify the channel from the signal component $\hat{s}^d_n$ as shown in Fig. 1 are explained in detail in [9],[10].

Mode	$\hat{\mathbf{C}}_n$	N_c
ISR-TR	$\left[\sum_{i=1}^{NI} \hat{\psi}^i_n \sum_{k=-1}^{+1} \hat{b}^i_{n+k} \hat{\underline{Y}}^i_{k,n}\right]$	1
ISR-R	$\left[\ldots, \sum_{k=-1}^{+1} \hat{b}^i_{n+k} \hat{\underline{Y}}^i_{k,n}, \ldots\right]$	NI
ISR-RH	$\left[\ldots, \sum_{k=-1}^{0} \hat{b}^i_{n+k} \hat{\underline{Y}}^i_{k,n}, \hat{\underline{Y}}^i_{+1,n}, \ldots\right]$	$2NI$
ISR-H	$\left[\ldots, \hat{\underline{Y}}^i_{-1,n}, \hat{\underline{Y}}^i_{0,n}, \hat{\underline{Y}}^i_{+1,n}, \ldots\right]$	$3NI$

Tab. 1. Common constraint matrix $\hat{\mathbf{C}}_n$ (the generic columns shown above are actually normalized to 1) and the corresponding number of constraints or columns N_c for each ISR mode.

The first constraint guarantees a distortionless response to the desired user while the second rejects the interference subspace and thereby cancels the total interference. We shall refer to this modification of the beamforming step of STAR as interference subspace rejection (ISR).

In Tab. 1, we show how to form the constraint matrix $\hat{\mathbf{C}}_n$ for different modes, which decompose or regroup interference vectors from different interference subspace characterizations. The TR (total realization) mode nulls the total interference vector and hence requires accurate estimation of all the channel and data parameters of the NI interferers. The R (realizations) mode nulls the signal vector of each interferer and hence becomes robust to power estimation errors. The H (hypotheses) mode nulls the signal vector from each interfering bit of each interferer and hence introduces robustness to data estimation errors. The RH (reduced hypotheses) mode, a hybrid between the R and H modes, reduces the number of constraints from $3NI$, for the H mode, to $2NI$.

In [11] we propose an additional mode that introduces robustness to channel identification errors. We also develop a new option that increases the space dimension with larger observation frames to reduce the relative effect of noise enhancement from a large number of constraints [11]. Note that the first two of the three decision-feedback (DF) ISR modes (namely TR, R and RH) require a delay of a symbol duration to allow estimation of $\hat{b}^i_{n+1}$. Hence joint (*i.e.*, the desired user is among the interferers) and multi-stage ISR implementations are treated among many other options in [11] (analysis is provided there as well).

With an estimate of the constraint matrix $\hat{\mathbf{C}}_n$ made available for a selected mode as shown in Tab. 1, we obtain the ISR combiner (*i.e.*, the constrained spatio-temporal beamformer) $\underline{W}_n^d$ by:

$$\mathbf{\Pi}_n = \mathbf{I}_{M*(2L-1)} - \hat{\mathbf{C}}_n \left(\hat{\mathbf{C}}_n^H \hat{\mathbf{C}}_n\right)^{-1} \hat{\mathbf{C}}_n^H , \quad (10)$$

$$\underline{W}_n^d = \frac{\mathbf{\Pi}_n \hat{\underline{Y}}_{0,n}^d}{\hat{\underline{Y}}_{0,n}^{d\ H} \mathbf{\Pi}_n \hat{\underline{Y}}_{0,n}^d} , \quad (11)$$

where $\mathbf{I}_{M*(2L-1)}$ denotes a $M*(2L-1) \times M*(2L-1)$ identity matrix. First, we form the projector[3] $\mathbf{\Pi}_n$ orthogonal to the constraint matrix $\hat{\mathbf{C}}_n$. Second, we project the estimate of the desired response vector $\hat{\underline{Y}}_{0,n}^d$ and normalize it[4]. The estimate $\hat{\underline{Y}}_{0,n}^d$ as well as the estimates $\hat{\underline{Y}}_{k,n}^i$ in Tab. 1 are reconstructed by spreading and delaying or advancing the corresponding channel estimates $\hat{\mathbf{H}}_n^d$ and $\hat{\mathbf{H}}_n^i$ (or $\hat{\underline{H}}_n^d$ and $\hat{\underline{H}}_n^i$) each provided by the channel identification unit of STAR [9],[10] (see Fig. 1).

With the above linearly-constrained combining or beamforming step, STAR-ISR exploits both space and time diversities as well as the array-processing capabilities of multiple antennas and carries out simultaneous channel and timing estimation, signal combining and interference rejection. In contrast, post-combining techniques (*e.g.*, [13],[14]) do not exploit multi-user detection in channel identification while pre-combining versions (*e.g.*, [14],[15]) do not exploit diversity advantages in multi-user detection (see more detailed discussion in [10]).

ISR also offers a novel interference rejection paradigm. On one hand ISR-TR is close to IC (*e.g.*, [5]-[7]) or linear IC (*e.g.*, [16]) methods. However, it replaces sensitive subtraction by more robust nulling. More generally, ISR can be interpreted as *linearly-constrained* linear IC with much higher near-far resistance. On the other hand ISR-H is close to a decorrelator-type receiver [3]-[4],[8]. It is even more similar to the projection receiver (PR) [17] when $M = P = 1$. In the general case, however, ISR characterizes interference from a new data model that merges both space and time (see above). Additionally, it implements nulling along different interference subspace decompositions (*i.e.*, TR, R, H, etc... [11],[12]) with much reduced complexity.

Overall, ISR modes improve in performance between IC detectors (closer to ISR-TR at the low end) and linear receivers (closer to ISR-H at the high-end), and require increasing complexity for implementation (see number of constraints in Tab. 1). Additionally, they completely mitigate the edge effect and can reject self-ISI with nulling [10]. They efficiently apply to the downlink [12], allow additional performance improvement by partial ISR [18], and can increase near-far resistance of channel identification [19]. They can even be adapted to implement MMSE-type criteria or to process multi-code or multi-rate data.

We explain in the following the performance/complexity tradeoffs and implementation issues resulting from the above ISR modes.

IV. Performance Evaluation

A. Simulation Configuration

We consider the uplink of an asynchronous wideband CDMA system with chip rate of 4.096 Mcps operating at a carrier frequency of 1.9 GHz. We assume a selective Rayleigh fading propagation environment characterized by $P = 3$ multipaths having a relative power profile of 0, -6 and -10 dB, respectively[5]. The corresponding multipath delays drift linearly at the rate of 2 ppm due to clock imprecision and to user mobility. The delay spread is fixed to $\Delta\tau = 8$ chips. We also assume a carrier recovery error of 0.04 ppm corresponding to a frequency error of 75 Hz at 1.9 GHz. An additional Doppler frequency of 8.8 Hz further contributes to channel time-variations expected at a low-speed mobility at 5 Kmph. We implement a closed power control loop at 1600 Hz, corresponding to an update period of 2560 chips (*i.e.*, 0.625 ms). To make this power control loop even more realistic, we degrade the power control link by a transmission delay of 0.625 ms and a BER of 10%. The power control increments are fixed at ±0.5 dB.

In the following, the desired user denotes one of the mobiles employing the default DBPSK modulation while the interferers denote the set of *NI* in-cell mobiles using differential 8-PSK modulation. All users have the same symbol-rate of 128 KBaud, corresponding to an equal spreading factor $L = 32$. The corresponding data-rates are 128 and 384 Kbps, respectively, corresponding to 64 and 192 Kbps with simple 1/2 rate FEC channel coding and decoding. The additive white noise in the observation vector mostly represents the plurality of other weak-rate users inside or outside the desired cell. To target the same nominal BER after differential decoding for a given SNR - our main evaluation criterion in this work - the nominal received power for both modulations is fixed to ${\psi_{\text{opt}}^d}^2 = 1$ and ${\psi_{\text{opt}}^i}^2 = 1/\sin^2(\pi/8)$ (*i.e.*, about 8.3 dB) for $i = 1, \ldots, NI$, respectively.

B. Simulation Results

We show below the performance results for ISR in terms of BER versus input SNR after despreading for users of both data-rates using the TR, R and H modes. For comparison, we provide reference performance results for SIC

[3]This projector is computed once for all desired users.

[4]These operations are actually implemented in a much simpler way that exploits redundant or straightforward computations in the data projection and the normalization.

[5]Other tested power-profiles of (0,0,0) and (0,-3,-6) dB resulted in negligible changes to the simulation results reported.

	$M=1$ - $NI=8$		$M=1$ - $NI=16$		$M=2$ - $NI=16$	
	BPSK	8-PSK	BPSK	8-PSK	BPSK	8-PSK
SIC	6.8	—	—	—	5.3	—
PIC	<u>12.8</u>	11.1	—	—	<u>14.5</u>	10.8
ISR-TR	<u>9.3</u>	8.8	—	—	<u>8.0</u>	7.1
ISR-R	6.4	<u>6.7</u>	—	—	<u>4.6</u>	4.3
ISR-H	5.6	<u>5.8</u>	<u>11.4</u>	<u>11.4</u>	<u>3.6</u>	3.2
SUB	**3.7**	**3.6**	**3.7**	**3.6**	**1.0**	**0.6**

Tab. 2. Required SNR (in dB) with respect to the BPSK user (*i.e.*, 128 Kbps) at a BER of 5×10^{-2} in 4.096 Mcps. For each technique and each (M, NI) configuration, the limiting SNR value between BPSK and 8-PSK is the underlined one.

and PIC[6], which likely offer the most acceptable performance/complexity tradeoffs today. We also provide the single-user bound (SUB) of ISR for both data-rates.

In Fig. 2, we provide the BER curves in the presence of $NI = 8$ interfering users when the number of receiving antennas at the base-station is fixed at $M = 1$. For both BPSK (Fig. 2a) and 8-PSK (Fig. 2b) we show the expected performance ranking among the ISR modes, steadily improving from TR to R, to H. For both modulations, PIC performs worse than ISR-TR. Fig. 2a, however, suggests that SIC takes advantage of power ranking when ultimately removing the strong 8-PSK interferers from the weak BPSK signal. When it comes to cancellation among the strong 8-PSK users themselves in Fig. 2b, power control, though imperfect, appears to render parallel interference cancellation more efficient than the successive one. SIC outperforms ISR-TR and falls close behind ISR-R for BPSK users. Also, it performs worse than PIC for 8-PSK users. However, we shall see below that ISR modes outperform both IC methods when considering the joint performance of BPSK and 8-PSK modulations.

Figs. 2a and 2b show that for both modulations ISR-H has the best interference rejection capabilities (*i.e.*, near-far resistance) with a simple 2 dB loss displacement compared to the SUB due to noise enhancement. Noise enhancement is less significant in the two other ISR modes, much less in the TR mode than in the R mode. However, residual interference is the dominant factor in limiting their performance due to their weaker robustness to estimation errors, more so for the TR mode than for the R mode. As suggested by the previous discussion, interference cancellation methods are more sensitive to reconstruction errors than suppression methods despite the absence of any kind of noise enhancement.

To gain deeper insight into the performance advantage of ISR over IC methods, we report in Tab. 2 the SNR required for a BER of 5×10^{-2}, assuming that channel coding/decoding can translate this probability of error into a practical QoS for high data-rates. The SNR is always given relative to the BPSK user, meaning that the SNR values reported for 8-PSK are simply displaced by 8.3 dB. SNR values reported for the SUB just confirm the need for such a power mismatch between both modulations to practically achieve the same BER. For each ISR and IC method, the higher SNR between the two modulations is the limiting SNR that guarantees the required QoS for both data links. Its value is underlined in Tab. 2.

In the two first columns of Tab. 2, the limiting SNR values for the situation considered suggest row-wise that: 1) SIC cannot accommodate 8-PSK users and hence cannot operate with mixed modulations; 2) PIC operating in the vicinity of 12.8 dB performs about 7 dB worse than the best ISR mode, ISR-H, and about 3.5 dB worse than the simplest ISR mode, ISR-TR; 3) ISR-R, operating at around 6.7 dB, offers intermediate performance between ISR-TR at 9.3 dB (*i.e.*, ~2.6 dB gain) and ISR-H at 5.8 dB (*i.e.*, ~0.9 dB loss); 4) ISR significantly outperforms IC methods and offers a variety of extremely efficient modes with different performance/complexity tradeoffs (see discussion below); 5) This performance gain increases at lower BER.

In a second set of experiments, we reassess the tested techniques in a heavily loaded system by doubling the number of 8-PSK interferers to $NI = 16$. Compared to the previous results, Figs. 3a and 3b show the same performance ranking among ISR modes and between ISR and IC methods. However, as indicated in the second pair of columns in Tab. 2, only ISR-H can accommodate BPSK and 8-PSK users while other techniques quickly saturate from interference. ISR-H looses as much as 7.7 dB relative to the SUB from noise enhancement, but shows tremendous near-far resistance capabilities and is still able to operate at a potential 11.4 dB in a heavily loaded system[7].

In a third set of experiments, we keep the number of interferers at $NI = 16$ and increase the number of receiving antennas to $M = 2$. Qualitatively, Figs. 4a and 4b lead to

[6]The incorporation of PIC and SIC into STAR subtracts interference and combines data simultaneously from both the temporal and spatial diversity branches according to a new data-model.

[7]A system-level simulation that complements the link-level simulation given herein is planned in a future work to confirm the capacity achievable at a given SNR.

(a) (b)

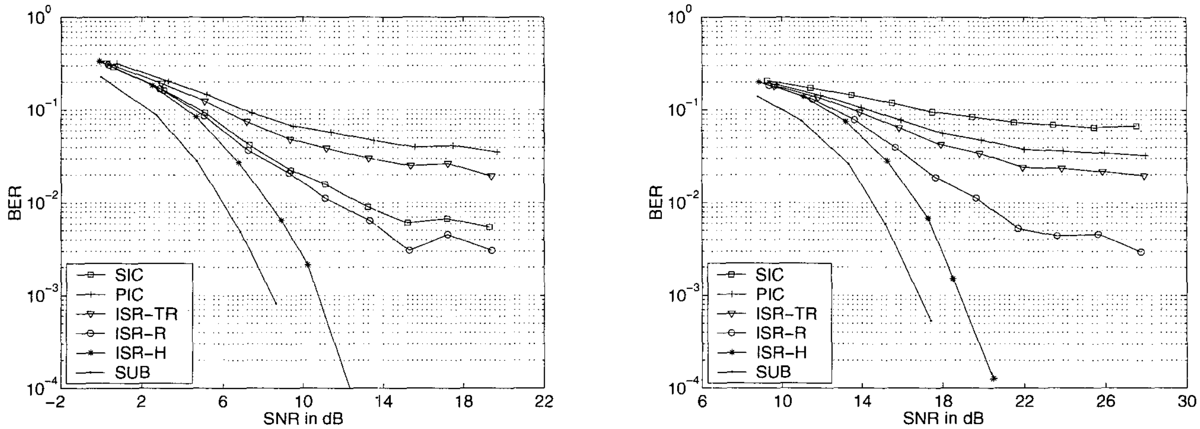

Fig. 2. BER versus SNR for $M = 1$ antenna and $NI = 8$ interferers in 4.096 Mcps. (a): BPSK @ 128 Kbps. (b): 8-PSK @ 384 Kbps.

(a) (b)

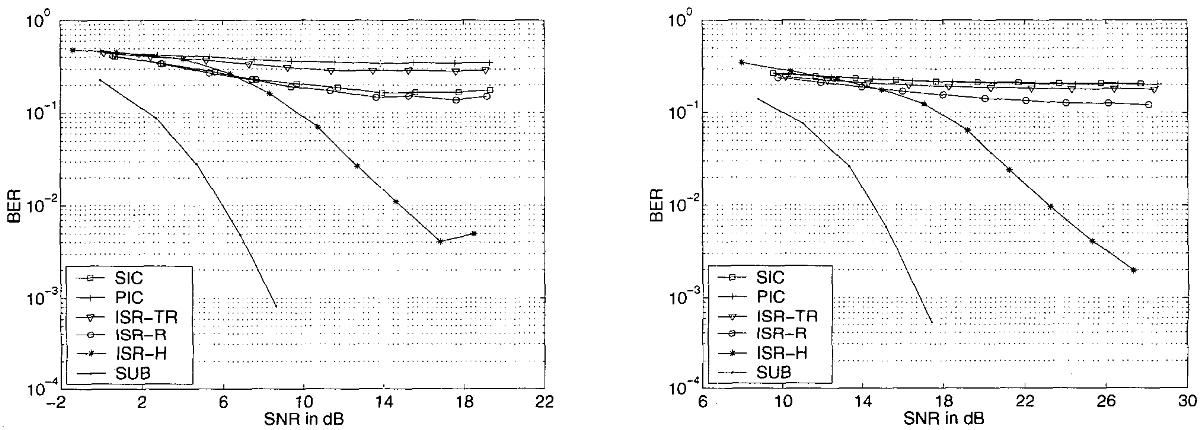

Fig. 3. BER versus SNR for $M = 1$ antenna and $NI = 16$ interferers in 4.096 Mcps. (a): BPSK @ 128 Kbps. (b): 8-PSK @ 384 Kbps.

(a) (b)

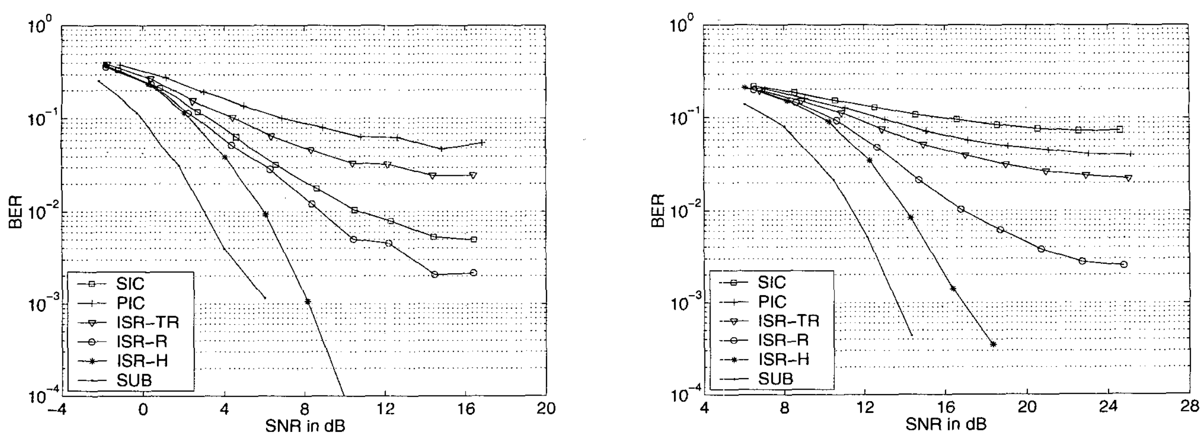

Fig. 4. BER versus SNR for $M = 2$ antennas and $NI = 16$ interferers in 4.096 Mcps. (a): BPSK @ 128 Kbps. (b): 8-PSK @ 384 Kbps.

the same conclusions as drawn from Fig. 2 with half the number of 8-PSK interferers and antennas. Compared to this initial situation, use of two antennas guarantees about 3 dB antenna-gain in additive noise reduction (see SUB curves). However, the resulting enhancement in interference rejection does not necessarily compensate for the increased amount of interference. A closer look at the two last columns of Tab. 2 indicates that SNR improves with BPSK and 8-PSK alike for all techniques but SIC with 8-PSK, and PIC with BPSK. It suggests that ISR benefits more from the enhanced interference rejection achievable with antenna-arrays than IC methods. In general, the results of the third experiment indicate that the more near-far resistant the technique is, the better it is able to exploit antenna-arrays for even more efficient interference rejection. Larger receiver arrays increase the performance advantage of ISR over IC methods.

V. Discussion and Conclusions

A new multi-user receiver structure, STAR-ISR, has been presented. Simulation results and complexity assessments indicate that the various ISR modes offer different performance/complexity tradeoffs. As performance improves from one mode to another, the complexity required increases while the resulting SNR advantage decreases making the last dB of gain even more expensive to obtain. In the short term, ISR-TR stands out as the best choice for implementation. It requires a low order of complexity that lies in the same range as IC methods; yet it outperforms them by significant SNR gains. As demand for service capacity increases in the long term, ISR-R and ISR-H (and other modes in [11],[12]) will ultimately become more attractive and affordable and so will become antenna-arrays.

As we move from one ISR mode to another, the factor that most affects complexity is the number of interferers to be suppressed, as well as the number of desired users to be processed simultaneously. In the situation depicted in Fig. 2, a conservative assessment indicates that ISR combining requires a computational complexity ranging from about 0.2 (ISR-TR) to 2.5 (ISR-H) Gops (10^9 operations per second) per user depending on the ISR mode implemented. STAR complexity (*i.e.*, both despreading and channel identification) adapts to channel time-variations. It can be reduced to 0.1 Gops for the slow Doppler case of Fig. 2 and results in a total complexity ranging from about 0.3 to 2.6 Gops for STAR-ISR.

Processors offer today computational power approaching 10 Gops and double speed almost every year and a half. To best exploit the available processing power, we suggest that ISR multi-user systems be designed initially to support a limited number of high-power users per cell or sector. With further advances in the capability to support greater complexity, the set of interferers suppressed may be increased to include lower-power lower-rate interferers as well.

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