Chapter in “WIRELESS COMMUNICATIONS IN THE 21ST CENTURY”, IEEE Press, edited by M. Shafi, J. Mizusawa, T. Hattori and S. Ogose.

Spatial-Temporal Signal Processing for Broadband Wireless Systems

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Abstract

This chapter reviews the basic techniques of adaptive spatial and temporal processing, which have been shown to yield substantial improvements in capacity and performance of wireless systems. After a review of appropriate transmission channel models, we introduce the elements of linear and nonlinear space-time processing, and outline their capabilities for combating interference and multipath. We then describe some applications and discuss adaptation issues. While many of the applications focus on receiver space-time processing, there is growing interest and promise in transmitter space-time processing as well.

1. Introduction - Motivation and Configurations for Space-Time Processing

Broadband digital multimedia communications is a fast-growing segment of the total traffic that will be carried by third generation cellular and other wireless communications systems. Rapid growth and increasing demands for bandwidth and near-ubiquitous coverage, combined with the performance usually associated with wired or fibered systems pose difficult challenges for wireless system designers. Each broadband data signal (usually defined as being equal or greater than 2 Mb/s) will occupy a relatively large portion of the overall allocated system bandwidth;

thus necessitating efficient frequency reuse among different users in the same small area. The use of “smart antennas” – generally consisting of arrays of antenna elements together with associated signal processing to control and combine the elements – can effectively meet many of these design challenges[1]. Adaptive antenna arrays[2,3,4,5], at either the transmitting end, the receiving end or both, can reduce or eliminate the effects of fading and multipath delay spread. As well, by increasing the effective antenna aperture, they can increase the antenna gain. With sufficiently large separation among elements, they can counteract the effects of large-scale signal variations (shadowing). Through their interference and fading mitigation properties, adaptive arrays can also provide a useful complement to equalization and coding techniques, for example reducing a code's minimum interleaving requirements in slowly fading channels[6].

A receiving antenna array comprises a set of individual antenna elements arranged in a 2- or 3-dimensional pattern, whose outputs are combined. Selection, equal gain or maximal ratio combining is well known approaches to providing diversity protection against fading and to minimize the effects of delay spread caused by multipath. In this article we are concerned with more general spatial processing, which can also reduce or eliminate interference from signals of other users of the same cellular communications system. Since the capacity of cellular systems is mainly determined by their ability to withstand cochannel and adjacent channel interference, smart antenna arrays can have a direct and positive impact on system capacity. Spatial processing can also be combined with temporal processing such as time domain filtering or equalization [7,8,9,10]. Advances in the technologies of digital signal processing, antenna fabrication and radio front ends, which are paving the way toward “software radios”, are also making relatively sophisticated spatial-temporal processing feasible and practical – at the transmitter, the receiver or both. Ultimately <u>space division</u> multiple access (SDMA) is possible – in which different users’ signals in the same vicinity, and using the same frequency band, can coexist without excessive interference.

Spatial processing (which can be thought of as spatial filtering) can shape an antenna pattern so as to emphasize desired signals and null out undesired signals (such as interferers or troublesome multipath components). It does this for example by appropriately weighting and combining the outputs of the individual antennas in the array. We shall see that spatial processing is complementary to temporal processing, which can eliminate or minimize intersymbol and/or cochannel interference from sampled channel responses. Spatial and temporal processing both separate and process signals based on differences among their "signatures" in space and time.

2. Channel Models for Multielement Arrays.

A model, showing radio channels between a single transmitting antenna, designated "i", and an array of *N* receiving antennas, is shown in Figure 1. The model assumes the use of a receiver and downconverter at the output of each receiving antenna element. The *N*-dimensional vector representing the complex baseband outputs of the receiving array is

$$\mathbf{r}_i(t) = \sum_{\ell} a_{i\ell}\mathbf{h}_i(t-\ell T) + \mathbf{v}(t) \tag{1}$$

where $\{a_{i\ell}\}$ are complex - valued data symbols transmitted at time ℓT from antenna i, $\mathbf{h}_i(t)$ is a complex vector impulse response from antenna i to the array, and $\mathbf{v}(t)$ is a vector representing additive noise.

The vector $\mathbf{h}_i(t)$ is also called the spatial-temporal signature [11] of the radio link between transmitter i and the array.

For *K* transmitters (i=0, 1, 2, …*K*-1),

$$\begin{aligned}\mathbf{r}(t) &= \sum_{i=0}^{K-1}\sum_{\ell} a_{il}\mathbf{h}_i(t-\ell T) + \mathbf{v}(t)\\ &= \sum_{\ell}\mathbf{h}(t-\ell T)\mathbf{a}_{\ell} + \mathbf{v}(t)\end{aligned} \tag{2}$$

where $\mathbf{h}(t) = [\mathbf{h}_0(t) \quad \mathbf{h}_1(t) \quad .. \quad \mathbf{h}_{K-1}(t)]$ is a N by K matrix channel impulse response, and

$$\mathbf{a}_{\ell} = \begin{bmatrix} a_{0\ell} \\ a_{1\ell} \\ . \\ a_{K-1,\ell} \end{bmatrix}. \tag{3}$$

The components of the vector impulse responses $\mathbf{h}_i(t)$ explicitly describe the properties of the radio channels between the ith transmitter and the array. Corresponding to the matrix impulse response $\mathbf{h}(t)$ is the matrix frequency response $\mathbf{H}(f)$. Note that the data symbols from different transmitters, $\{a_{i\ell}\}$ are assumed to be transmitted at the same symbol rate $1/T$, but are otherwise not necessarily synchronized. Small differences among symbol rates from different transmitters could be accounted for by corresponding slow time variations in the $\{\mathbf{h}_i(t)\}$, as long as the average symbol rates remain identical.

In the corresponding physical model, each radio channel in Figure 1 may represent the superposition of one or more physical radio paths with different delays and complex gains: a combination of direct (line of sight) paths, and/or reflections from scattering objects in the vicinity of the transmitter or receiver, and/or diffraction around obstacles. The signal from a given source may thus arrive at a given receiving antenna element from several different directions, with different delays and gains; the distribution of angles of arrival of a signal is called its angular spread. A large angular spread is generally found when the scattering/reflecting objects are close to or surrounding the receiving antenna. In this case, the outputs of receiving elements spaced by about $\lambda/2$ (where λ is the wavelength) will be nearly uncorrelated [12]. A small angular spread will require a larger spacing to achieve small correlation, and is associated with environments where the scattering/reflecting objects are far from the receiver, and subtend only a small angle from it. For a small angular spread, a correlation coefficient of less than 0.5 to 0.7 requires a minimum inter-element spacing of roughly [12] $\frac{1.9\,\lambda}{\pi.\text{angular spread (in radians)}}$.

The arrival of multiple versions of the same signal (multipath components) with different delays causes intersymbol interference if the delay differences are on the same order as a symbol interval or more.

Transmission over multiple uncorrelated channels provided by spatial or other forms of diversity is well known as an effective antidote to fading and multipath delay spread. Fading mitigation by spatial diversity is still effective even for correlation coefficients among diversity channels as high as 0.5 [12,13].

Recently there has been a renewed interest in spatial channel modeling as a result of the interest in applying spatial-temporal processing of the types discussed in here. A number of analytical and measurement-based models have been developed [14] for correlation, spatial signature variation [11] and other statistical properties of signals received at antenna array elements. These models are useful for analyzing the performance and for simulating systems which use antenna arrays. In general, an array's ability to combat multipath and/or fading is maximized if the correlation among elements is as low as possible. For effective separation of interfering signals, desired and interfering signals should have very different spatial-temporal signatures, either because they have quite different directions of arrival, or because their respective angular spreads, seen from the receiving array, are large and different.

3. Receiver Space-Time Processing

3a. Linear Space-Time Processing

Optimal array-based multiuser detection receivers for minimizing error probability in the presence of Gaussian noise, channel dispersion and interference, can be shown to include linear-processing front ends in the form of generalized spatial-temporal matched filter structures, followed by sampling, and nonlinear (generally exponentially-complex) decision-making [15]. Less general, but more practical, receiver structures are based on linear processing, together with sampling and decision making by simple quantization. Figure 2 shows a general linear space-time processing scheme for making a decision on the n th data symbol a_{0n}, from transmitter 0. The complex baseband output of each of N antenna array elements is fed to linear filters represented by a vector impulse response $\mathbf{w}_0(t)$, whose outputs are summed and sampled at the symbol rate (at times $\{0, T, 2T, ..nT..\}$. The array elements are usually omnidirectional; but directional elements may be used as well, providing coverage over partially overlapping directional beam patterns fanning out from the receiver. In such a case, the array may be considered to provide a kind of direction-of-arrival sampling. Reference [16] shows that time and space filtering and sampling operations provide a canonical space-time reception processing in terms of a fixed basis that is independent of the channel parameters.

In the general linear space-time receiver of Figure 2, the output on which the decision is based is

$$
\begin{aligned}
y_{0n} &= \int_{-\infty}^{\infty} \mathbf{w}_0(\boldsymbol{t})^H \, \mathbf{r}(nT-\boldsymbol{t})\, d\boldsymbol{t} \\
&= \sum_{\ell} \int_{-\infty}^{\infty} \mathbf{w}_0(\boldsymbol{t})^H \, \mathbf{h}(nT-\ell T-\boldsymbol{t})\mathbf{a}_{\ell}\, d\boldsymbol{t} + \int_{-\infty}^{\infty} \mathbf{w}_0(\boldsymbol{t})^H \, \mathbf{v}(nT-\boldsymbol{t})\, d\boldsymbol{t}
\end{aligned}
\qquad (4)
$$

3a.1 Zero-forcing criterion[17]

Under the zero-forcing criterion, y_{0n} should have no interference from other transmitters, and no intersymbol interference; i.e. $y_{0n} = a_{0,n-D} + u_{0n}$, where D is an integer delay chosen to make the filters { $w_{ik}(t)$ } causal, and

$$u_{0n} = \int_{-\infty}^{\infty} \mathbf{w}_0(t)^H \mathbf{v}(nT - t)dt \qquad (5)$$

Henceforth we set D=0 for notational simplicity (allowing non-causal filters). For additive white Gaussian noise with power spectral density $N_0/2$, the mean and variance of u_{0n} are respectively zero and

$$N_0 \int_{-\infty}^{\infty} \left|\mathbf{w}_0(t)\right|^2 dt = N_0 \int_{-\infty}^{\infty} \left|\mathbf{W}_0(f)\right|^2 df \; . \qquad (6)$$

The condition for zero-forcing the co-channel and intersymbol interference is that the K-dimensional overall vector response $\mathbf{q}(t)$ satisfies

$$\mathbf{q}(nT)^T \equiv [\int_{-\infty}^{\infty} \mathbf{w}_0(t)^H \mathbf{h}(nT-t)dt]^T = \begin{cases} [1,0,0,...0] \text{ for } n = 0 \\ [0,0,0,...0] \text{ for } n \neq 0 \end{cases} \qquad (7)$$

Taking Fourier transforms over the index n, we get the equivalent equation in the frequency domain

$$\sum_m \mathbf{W}_0(f + \frac{m}{T})^H \mathbf{H}(f + \frac{m}{T}) = [T,0,0,...0], \qquad (8)$$

where $\mathbf{W}_0(f)$ and $\mathbf{H}(f)$ are the Fourier transforms of $\mathbf{w}_0(t)$ and $\mathbf{h}(t)$, respectively. If the channels and/or frequency responses are strictly bandlimited to say $|f| \leq B/T$, there are at least int($2B$) nonzero terms in the summation for any frequency f (and no more than 1+int ($2B$)), where int (x) denotes the largest integer equal or less than x. Thus for any frequency f, the above matrix equation represents a system of K linear equations in int($2B$) unknown vectors $\mathbf{W}_0(f+m/T)$. Since each of these vectors has N components, there are a total of at least N int($2B$) unknowns in the K linear equations. A unique solution for any frequency f may be found, assuming the K columns of the matrix $\mathbf{H}(f)$ are linearly independent, if the number of interfering transmitters, K, satisfies

$$K = N \, \mathrm{int}(2B). \tag{9a}$$

Figure 3 shows spectra with excess bandwidths of 0, 100% and 200%, and their interference suppression capabilities, corresponding to B=1/2, 1 and 3/2, respectively. Conversely, there is no solution for a larger number of interfering transmitters, and there are in general an infinite number of possible solutions for a smaller number; i.e. for

$$K < N \, \mathrm{int}(2B). \tag{9b}$$

As a special case, we have the familiar result that for B=1/2 (the minimum Nyquist bandwidth), N=1 (one antenna), and K=1 (no interferers), a unique equalizer frequency response satisfying (8) is simply T times the inverse of the channel, $H_0(f)^{-1}$. In general, equation (9a) implies that the number of simultaneous users whose mutual interference can be mitigated by the system of Figure 2 is proportional to the number of antenna elements and also to the system's excess bandwidth. Interference-suppression capabilities for CDMA systems with finite numbers of filter taps have been obtained in [18].

The ability of N+1 antenna elements to separate up to N+1 users was pointed out by Winters et al[19,20,40]. See also [21,22]. The ability of a linear equalizer to suppress a number of interferers proportional to excess bandwidth was foretold in a paper by Shnidman[23]. See also[24,25]. That the number of separable (orthogonal) signals is proportional to the bandwidth/symbol rate ratio is a well known result of signal theory, and is the basis for spread spectrum CDMA multiple access schemes, as well as for FDMA and TDMA signal separation schemes for multiple access. In wireless systems however, orthogonality among different users' transmitted signals is often lost as a result of channel multipath or frequency or time slot reuse.

What is the best way to maximize the system capacity; i.e. the number of interfering users for which signal separation is possible: by increasing the number of antenna elements or by increasing the excess bandwidth? Increasing the number of antennas is clearly preferable, since bandwidth is a finite resource. However, there are often physical and cost constraints which limit the number of antennas at a site. Thus the ability of linear equalization, coupled with excess bandwidth, to augment the interference suppression capability of an antenna array is very valuable. For TDMA and FDMA systems, the excess bandwidth is usually moderate (B is typically in the range of 0.5 to 1 for these systems), and thus temporal linear processing is often more useful for equalization than for interference suppression. However, temporal processing is very powerful against interference in direct sequence CDMA systems, where B is significantly greater than 1; for example a binary CDMA system with a spreading gain of 32 would have B=16, and up to 32 interferers per antenna element would theoretically be suppressible. However these theoretical limits are only attainable if the set of equations (8) is non-singular for any f. Furthermore, even if a zero-forcing solution exists, its filtering operation may cause significant enhancement of additive thermal noise. In practice, noise enhancement and digital signal processing and adaptation considerations limit the number of interferers which can be effectively suppressed to something on the order of half to three quarters of the maximum theoretical number. Nevertheless the effectiveness of linear filtering in removing interference from received CDMA signals leads to significant gains in system capacity and to reduced sensitivity to near-far interference effects. Furthermore, as will be seen later, supplementing temporal processing with spatial processing in CDMA systems furthers these benefits.

Before leaving the issue of theoretical capabilities of linear combining systems to combat multipath ISI, cochannel interference, and noise, we consider a common special case of linear combining, in which the filters $\{w_{0k}(t)\}$ are replaced by memoryless complex weights $\{w_{0k}\}$; i.e. the only "memory" available to the receiver results from any relative time delays in signals reaching the K antenna elements. It may appear that there is no capability to combat intersymbol interference in this case. However that is not necessarily true, as shown by Clark[26,27]. The sampled array output is expressed, analogous to (7) as

$$\mathbf{q}(nT)^T = \mathbf{w}_0{}^H \mathbf{h}(nT) = \begin{cases} [1,0,0,..0] \text{ for } n = 0 \\ [0,0,0,..0] \text{ for } n \neq 0 \end{cases} \qquad (10)$$

If the $\mathbf{h}_i(t)$ are all time-limited to say $|t| \leq ST$, then there are $2S+1$ non-zero samples of each $\mathbf{h}_i(t)$, and (10) represents a set of $(2S+1)K$ equations in the N unknowns $w_{01,\ldots}w_{0N}$. Thus the ISI and interference can be eliminated by this "memoryless" array, provided that

$$N \geq (2S+1)K \ . \tag{11}$$

Thus the required number of antenna elements is proportional to the time duration of the multipath impulse responses and to the total number of users. Clark et al[27] showed numerical results which demonstrated the ability of "memoryless" arrays to effectively mitigate intersymbol interference.

3a.2 Minimum Mean Squared Error Criterion

The above zero-forcing results illustrate the ability of linear equalizers and/or linear spatial combiners to completely suppress ISI and cochannel interference up to limits imposed by bandwidth, number of antenna elements and number of interferers. However, noise enhancement and adaptation can be problematic for receivers based on zero-forcing. Minimization of the total mean squared error (MSE), consisting of thermal noise and residual ISI and cochannel interference at the equalizer or combiner output, is usually a more useful criterion for evaluation and adaptation purposes. The analytical problem can be formulated as either minimization of the MSE of one user, say user "0",

$$\mathrm{MSE} = \mathrm{E}\left|y_{0n} - a_{0n}\right|^2 .$$

The optimum set of linear combining filters turns out to be representable as a bank of filters matched to the components of the $\{\mathbf{h}_k(t)\}$ channel responses, whose outputs are sampled at T second intervals, and routed to and subsequently combined by sets of transversal filters[24]. A general expression for the minimum total MSE for a N-input, N-output linear system was derived by Salz[22]. A related zero-forcing problem, minimizing the output noise variance under the zero-forcing constraint (8) for a K-input, K-output linear system was solved by Van Etten[21]. In all of these results, no constraint was placed on the memory or complexity of the filter or filters $w_k^{(l)}(t)$.

Practical equalizers $\mathbf{w} = \begin{bmatrix} \mathbf{w}_{-M} \\ : \\ \mathbf{w}_M \end{bmatrix}$, based on transversal filters, have finite numbers of tap coefficients, and, following fixed anti-aliasing filters, operate at sampling rates that equal the symbol rate or a multiple of it. For a linear equalizing combiner, with say 2*M*+1 taps per antenna element, and with a sampling rate of $1/\Delta$, the output corresponding to equation (4) is

$$y(iT) = \mathbf{w}^H \mathbf{r}_i \,, \tag{12}$$

where

$$\mathbf{w} = \begin{bmatrix} \mathbf{w}_{-M} \\ : \\ \mathbf{w}_M \end{bmatrix}, \tag{13}$$

and

$$\mathbf{r}_i = \begin{bmatrix} \mathbf{r}(iT + M\Delta) \\ : \\ \mathbf{r}(iT - M\Delta) \end{bmatrix} \tag{14}$$

are $(2M+1)N$ - dimensional vectors representing the tap coefficients and the channel output samples respectively. Minimization of the MSE expression (19) for user "0" results in

$$\mathbf{w}_{opt} = \mathbf{R}^{-1}\mathbf{v} \tag{15}$$

where **R** is the channel output autocorrelation matrix,

$$\mathbf{R} = \mathrm{E}\left(\mathbf{r}_i \mathbf{r}_i^H\right) = \begin{bmatrix} \mathbf{R}_{-M,-M} & .. & \mathbf{R}_{-M,M} \\ : & & : \\ \mathbf{R}_{M,-M} & .. & \mathbf{R}_{M,M} \end{bmatrix} \tag{16a}$$

where $\mathbf{R}_{ij}$ are N by N square matrices

$$\mathbf{R}_{ij} = \sum_n \mathbf{h}(i\Delta - nT)\mathbf{h}(j\Delta - nT)^* + \frac{2N_0 B_r}{T}\mathbf{I}\delta_{ij}, \tag{16b}$$

$\mathbf{I}$ is an identity matrix, δ_{ij} is the Kronecker delta function and $\mathbf{v}$ is the desired channel propagation vector

$$\mathbf{v} = \mathrm{E}(\mathbf{r}_i a_{0i}^*) = \begin{bmatrix} \mathbf{h}_0(M\Delta) \\ \vdots \\ \mathbf{h}_0(-M\Delta) \end{bmatrix}. \tag{17}$$

assuming uncorrelated unit variance data symbols and white noise. The minimum MSE is then

$$\mathrm{MMSE} = 1 - \mathbf{v}^H \mathbf{R}^{-1}\mathbf{v}\,. \tag{18}$$

With the addition of decision feedback, $\mathbf{w}$ is augmented by F complex-valued feedback coefficients, and $\mathbf{r}_i$ is augmented by the previous F decisions $\{a_{0n}\}_{n=1}^{F}$.

The optimum set of tap coefficients and minimum MSE is still given by (15) and (18) respectively, but

$$\mathbf{v} = \begin{bmatrix} \mathbf{h}_0(M\Delta) \\ \vdots \\ \mathbf{h}_0(-M\Delta) \\ \cdots\cdots\cdots \\ 0 \\ \vdots \\ 0 \end{bmatrix} \tag{19}$$

and

$$\mathbf{R} = \begin{bmatrix} & \mathbf{R}_{11} & & \vdots & \mathbf{R}_{12} \\ & & & \vdots & \\ \cdots & \cdots & \cdots & \vdots & \cdots \\ & \mathbf{R}_{21} & & \vdots & \mathbf{R}_{22} \end{bmatrix} \tag{20}$$

where $\mathbf{R}_{11}$ has the same form as (16a), $\mathbf{R}_{22}$ is a F by F identity matrix, and

$$\mathbf{R}_{12} = \mathbf{R}_{21}{}^{H} = \begin{bmatrix} \mathbf{h}_0(M\Delta + T) & .. & \mathbf{h}_0(M\Delta + FT) \\ : & & : \\ \mathbf{h}_0(-M\Delta + T) & .. & \mathbf{h}_0(-M\Delta + FT) \end{bmatrix}. \tag{21}$$

3a.3 Max. SINR and Error Probability

The output signal-to-interference-plus-noise ratio (SINR) is another measure of the performance of a spatial and/or temporal processing receiver that suppresses interference and noise. It is defined as the ratio of the squared magnitude of the desired signal to the sum of the mean squared values of the total interference and noise, both measured at the sampling instant at the output of the linear combiner. It can be shown, by application of the Matrix Inversion Lemma, that the set of tap coefficients that minimizes MSE, given by (15), also maximizes SINR[40] . The maximum SINR is given by

$$\text{Max. SINR} = \frac{1 - MMSE}{MMSE} \tag{22a}$$

where MMSE is given by (18). It can be shown by the Matrix Inversion Lemma that another formula equivalent to (22a) is [40]

$$\max \text{SINR} = \mathbf{v}^{H} \mathbf{R}_{I+N}^{-1} \mathbf{v} \tag{22b}$$

where $\mathbf{R}_{I+N}$ is the covariance matrix of the interference plus noise.

A formulation for the probability distribution for SINR was obtained by Shah and Haimovich [28] for the case of $K>N$ equal power, Rayleigh-faded interferers, with no additive noise. Gao, Smith and Clark generalized to this result to include the effects of noise and unequal power interferers [29].

Symbol error probability is also an important performance measure. A simple and useful Chernoff-type upper bound on the symbol error probability is [20]

$$P_e \leq \exp(-\frac{1}{2MMSE}) \tag{23}$$

where MMSE, given by (18), applies for unit data symbol variance. Winters and Salz gave a relatively simple upper bound on bit error probability for spatial processing, accurate for low bit error rates [30]. They bound (23) in terms of the determinant of $\mathbf{R}_{I+N}$, and further simplify the bound for up to 7 antennas in terms of interferer powers.

3b Nonlinear Space-Time Processing

Maximum likelihood is an even more powerful detection approach for temporal and spatial anti-interference processing[31 32]. However for receivers with multi-element antenna arrays, the differences in performance among linear, decision feedback and maximum likelihood detectors are typically less than they would be for non-spatial-processing receivers operating in the presence of severe multipath [33]. This is because the spatial processing adds extra degrees of freedom. For example, suppose the jth frequency response component $H_{0j}(f)$ of $\mathbf{H}_0(f)$ has spectral nulls, which would be best handled by a decision feedback equalizer or maximum likelihood sequence detector if there is only one antenna element j ; but it is unlikely that the N responses to N antenna elements will have the same nulls, and thus linear spatial-temporal processing might yield almost equivalent performance to that of decision feedback processing, if previous decisions of the desired signal are fed back. An asymptotic result shows that an optimum linear N-branch space-time receiver requires only one additional diversity branch to achieve maximum likelihood receiver performance[34] against multipath .

However if previous decisions of the interfering signals are also fed back (some or all of these would generally be available at a cellular radio base station which is simultaneously receiving signals from many terminals), substantial gains in interference suppression performance can result. This is called centralized decision feedback[35,36,37], and is a form of interference cancellation[38]. Figure 4 shows a centralized decision feedback receiver for signals from interfering sources “0” and “1”. If say, source 0’s signal is substantially larger, source 1’s decision could be delayed so that essentially all of source 0’s interference is cancelled from it before a decision is made, as suggested in Figure 4.

An important design question for linear or nonlinear space-time processing is how many tap coefficients to provide for each of the N forward filters. A very useful rule of thumb result was recently obtained in [39]: the number of tap coefficients for which the MMSE performance approaches that of infinite-tap processors. Analysis in [39], supported by simulations, shows that for a decision feedback equalizer structure, in which the number of antennas exceeds the number of interferers, the forward filters should each span a number of symbol intervals roughly equal to $D[1+(2K-1)\varphi]$, where D is the span of the delay spread in symbol intervals, K is the total number of interfering signals (including the desired signal), and φ is the input SNR (in dB) divided by 10. Of this total, the causal and anticausal portions are respectively $C=D(K-1)\varphi$ and $A=D(1+K\varphi)$. The number of feedback coefficients should be $D+C$. These numbers were shown to also be valid for practical maximum likelihood sequence estimation reception and (approximately) for linear reception. Note that they are independent of the number of antenna elements.

In the next section, we consider some applications of these results.

3c Applications of Receiver Space-Time Processing

The capacity of cellular systems increases with the number of interfering signals that the receivers in the system can tolerate with adequate performance. Winters[40,19] showed that a receiver with an array of N antennas can theoretically tolerate (suppress) up to N-1 interferers. The capacity is thus proportional to N [20]. Furthermore, for independent flat Rayleigh fading on each of the N paths, $K+N$ antennas will suppress up to K-1 interferers, and also provide up to N–fold path diversity for each user [20]. Correlations of up to 0.5 among the N paths was shown[41] to have only a minor effect. The ability of adaptive arrays to increase capacity in TDMA cellular systems has been confirmed in laboratory and field experiments [20 42 43].

Direct sequence CDMA cellular systems are robust to interference, but their capacity can be significantly improved by appropriate time domain processing. Verdú [15] and others have derived optimum and suboptimum receivers for multiple user CDMA systems. Linear or decision feedback equalization is one form of time domain processing which has been found effective in suppressing interference (and thereby increasing system capacity) in CDMA systems with short spreading codes [44 45 46]. Short spreading codes are those whose length equals a bit interval or a small multiple of it, and which do not change from bit to bit. The interference suppression capability of linear filtering used with direct sequence CDMA follows as a consequence of equation (9) from its large excess bandwidth (approximately equal to the spreading gain). Short-code CDMA systems using adaptive equalization have been shown to approximately double or triple the capacity of systems using conventional (matched filter) reception, and also include the function of RAKE-type reception in the presence of multipath [44,45,46,47]. Furthermore, such systems generally have much less sensitivity to variations in interferers' power due to imperfect power control than do conventional direct-sequence CDMA systems.

Figure 5 shows a realization of linear filtering applied to a direct sequence CDMA receiver, for the case where multipath delay spread and interference span up to two bit intervals, and therefore the linear filtering extends over two bit intervals. Close examination of the processing depicted in this figure reveals that it is equivalent to transversal filtering with complex tap coefficients $\{w_{0j}^*\}$, spanning two bit intervals ($2T$). It is also equivalent to conventional RAKE receiver processing[13], with one principal difference: in the RAKE receiver, the $\{w_{0j}^*\}$ would be replaced by linear combinations of ± 1 spreading code chip values, with weights determined by estimated multipath component gains. This equivalence indicates that temporal linear processing to deal jointly with multipath and interference need not be significantly more complex than conventional RAKE receiver techniques based on correlation.

Combining time-domain processing of CDMA with antenna array processing as in Figure 2 yields further capacity benefits [32,48] - primarily because of the exploitation of the extra degrees of freedom mentioned earlier. In effect, spatial and temporal processing are complementary to each other. For example, interference or multipath components from similar directions may have similar spatial signatures, but can be eliminated by temporal processing, since their temporal signatures will likely be different. Interfering components with similar temporal signatures (for example interferers with highly correlated spreading code signatures) but different directions of arrival, may be eliminated by the spatial processing. It is worth noting that the benefits of adaptive array processing by itself, without adaptive temporal processing to combat interference, may be minimal in CDMA systems with large spreading gains. Instead, switched-beam antenna systems are used with conventional CDMA systems. The reason is that the number of interferers is usually much larger than the practical number of adaptive antenna elements which can be implemented[1]. Figure 5 illustrates spatial-temporal processing through the combination of other array outputs.

The dependence of MMSE on the number of antenna elements and on the number of adaptive tap coefficients at the output of each element is illustrated[48] in Figures 6 and 7. In these examples, the spreading gain is 8; short spreading codes are randomly chosen for each user; each interfering user's signal is received at each antenna element over 3 independent Rayleigh-distributed multipath components with equal average power, with overall delay spread ≤ 6 chips; interferers' directions of arrival are random; and E_b/N_0 of each received signal at each element is 15 dB. The sampling rate into each forward filter is the chip rate. Figure 6 shows MMSE versus number of forward tap coefficients per element, as the number of interfering user signals increases from 2 to 18, when there are 4 antenna elements. It indicates that an appropriate choice for the number of forward tap coefficients is approximately the number of chips per bit plus the maximum expected delay spread, in chip intervals. For a typical CDMA operating point, with a MMSE of about –10 dB, Figure 7 shows the rapid increase in the number of tolerable interfering signals as the number of antenna elements is increased; this number approaches the product of the spreading gain and the number of elements.

Figure 8 shows the significant reduction in MMSE that is possible from centralized decision feedback processing[37], as compared to linear and conventional decision feedback processing. The conditions are the same as in the above example, with the exception of E_b/N_0, which is 18 dB. Similar results for non-CDMA systems are shown in [36], which also derives a closed-form asymptotic expression for MMSE for infinite-length temporal processing. An extension to partial centralized decision feedback connectivity, where, for example, out-of-cell interferers are dealt with by forward temporal and spatial processing but not by feedback processing, is found in [49].

4. Recent Space-Time Wireless Communication Architectures

Recently, much interest has centered on the simultaneous use of multiple transmitting and receiving antennas as a means of increasing the capacity of restricted bandwidth links, almost without limit [50,51,52]. The application generally involves a single wireless link between a transmitting end and a receiving end. With K distinct parallel data streams transmitted from K antennas (all at the same physical location), and with $N \geq K$ receiving antenna elements, whose outputs are linearly combined into K receivers, it is possible to create essentially K "parallel data pipes", each carrying independent data[52]. The total end-to-end bit rate, and therefore the system capacity, is increased by a factor of K without any bandwidth expansion. However for this capacity expansion to be realized, the transmission channels between each pair of transmitting and receiving antennas should be independent. This will be possible if the antenna elements are spaced at least half a wavelength apart, and if dense scattering objects cause independent fading on all the paths. The system can use linear spatial, and if necessary temporal processing at the receiving array similar to what has been previously described, with a separate coefficient vector $\mathbf{w}_k$, k=1,2,…K, optimized according to either a zero-forcing or a MMSE criterion, for each of the K receivers. The receiver performance is further enhanced by the use of nonlinear cancellation: the K data symbols are detected sequentially in order of decreasing output SINR, and as each one is detected, its response is subtracted from the inputs, thus reducing the interference to later detected symbols. The idea has been successfully demonstrated in hardware [52] in an indoor office environment. The "bandwidth efficiency" (bits/s/Hz) possible with this approach is theoretically limited only by the number of parallel transmitters and antennas that can be deployed at each end of a link.

The approach has also been generalized to "space-time" coding, in which block or convolutional codes, and decoding is applied across the parallel streams of data, again using a combination of linear receiver spatial processing and an interference cancellation approach [50,51,53,54,55].

5. Adaptation issues

Adaptation of equalizer and antenna array tap coefficients can be based on using receiver decisions in a "decision-directed" mode as a reference desired output. Initially, with all tap coefficients set to arbitrary values, say zero, a known "training" sequence of data may be transmitted and supplied as a reference, until the receiver decisions are sufficiently reliable to use as a reference. Alternatively, directions of arrival, and hence spatial signatures, for individual received signal components can be estimated using a variety of methods such as MUSIC or ESPRIT [5], but these approaches lose effectiveness if the number of multipath components associated with each received signal is large.

A likely candidate for a training-based adaptation algorithm is the simple LMS (least mean squares) algorithm, which continually adjusts the adaptive coefficients in the direction of reduced mean squared error [56]. At time nT, the coefficient vector is updated as

$$\mathbf{w}_0(n+1) = \mathbf{w}_0(n) - m_n e(n)^* \mathbf{r}_n \tag{23}$$

where $e(n) = y(nT) - a_{0n}$ is the error at time nT, and m_n is an appropriately normalized step size scalar parameter (proportional to the inverse of the current estimated input signal variance).

Convergence of the LMS algorithm to the vicinity of the minimum mean squared error typically requires a number of training symbols (iterations) equal to at least 10 times the number of coefficients being adapted; for arrays with significant correlation among element outputs, much longer convergence times are required [56]. The LMS algorithm is best suited to applications which permit long training times, and to slowly varying channels.

An alternative to this LMS adaptation approach is to take a “least squares approach”; i.e. find coefficients which directly minimize the sum of squares of errors between desired output data symbols and spatial-temporal array outputs. In this approach we <u>estimate</u> the channel autocorrelation matrix **R** and the desired channel propagation vector **v**, from processing and time-averaging the appropriate products of channel outputs and receiver decisions [56],

i.e. $$\hat{\mathbf{R}} = \frac{1}{TB}\sum_{i=1}^{TB} \mathbf{r}_i \mathbf{r}_i^H \quad \text{and} \quad \hat{\mathbf{v}} = \frac{1}{TB}\sum_{i=1}^{TB} \mathbf{r}_i a_{0i}^* \tag{24}$$

where TB is the length of the training period, in symbols, and computing the tap coefficients from (15);

i.e.

$$\hat{\mathbf{w}} = \hat{\mathbf{R}}^{-1}\,\hat{\mathbf{v}}. \tag{25}$$

This is called DMI (direct matrix inversion), and is performed once (at the end of each training period) to compute the coefficients relevant to that training period. The length of the training period, measured in data symbols, needed for convergence to within 1 to 3 dB of the minimum mean squared error, is roughly twice the number of coefficients to be adapted [56]. At the end of the training period, the adaptation could be switched to decision-directed LMS, in order to track slow channel variations. The convergence of the DMI algorithm is much faster than that of LMS. However it is more complex. If the solution of (25) is implemented with Cholesky factorization, the total number of complex multiplications, including those for estimating **R** and **v** can be shown to be on the order of

$\frac{M^3}{6} + 6M^2$, where M is the total number of coefficients, assuming that the training period is $2M$ symbols. For the LMS algorithm, the total number of multiplies would be about $2M$ times the number of symbols in the training period. If the required training period for the LMS algorithm to converge is very long, the total complexity (as measured by number of multiplications) could be similar for the two algorithms, or even greater for LMS.

The Recursive Least Squares (RLS) algorithm is a recursive version of DMI, in which the coefficients are updated once per training symbol so as to minimize the sum of squares of errors up to that time [56]. If the coefficients are required to be estimated only once during the training period, the DMI approach uses fewer computations than the RLS approach. For a version of RLS which uses a rectangular sliding window, [57] demonstrates that estimation of $\mathbf{R}_{\mathrm{I+N}}$ and use of (22b) instead of **R** and (22a) results in more accurate tracking of time variations.

For high bit rate applications, where the multipath delay spread extends over tens or hundreds of symbol intervals, the total number of required coefficients M, and hence the complexity of temporal processing is high. In this case, frequency domain receiver space-time processing, including adaptation, can yield significant simplification[34] . The simplification is achieved by using FFT operations to replace time domain filtering, and through independent adaptation of each frequency component.

Note that the actual autocorrelation matrix $\mathbf{R}$, given by (16b), depends only on desired and interfering channel responses and noise, and its estimate is obtained by time-averaging products of delayed array element outputs. However the actual propagation vector $\mathbf{v}$, given by (17), depends only on the desired channel response, and its estimate is obtained by time-averaging products of array outputs with training data symbols; i.e. estimation of $\mathbf{v}$ requires training symbols, while $\mathbf{R}$ does not. In principle then, it seems that $\mathbf{v}$ could be estimated during the training period, while $\mathbf{R}$ could be estimated over longer periods to achieve a more accurate estimate, such as when new interferers appear, or old ones disappear. Unfortunately, using different data to estimate $\mathbf{R}$ and $\mathbf{v}$ turns out to produce poor estimates of the coefficients, unless the estimation periods are impractically long, and conditions do not vary with time; the reason is that this potential approach does not give a least squares solution for the given data. A partial solution to this problem is to take a subspace processing approach [58], in which $\mathbf{v}$ is estimated from the training symbols in combination with a subspace decomposition of the estimate of $\mathbf{R}$, which has been obtained by time-averaging from a longer block of received data. The result is a robust array adaptation algorithm which allows estimation of the optimal coefficients from an arbitrarily long sequence of array channel outputs and a relatively short training sequence.

The robustness to suddenly appearing and disappearing interferers, of adaptive space-time processing in CDMA systems is shown in [59]. In comparison to their effects on a CDMA receiver with adaptive temporal-only processing, newly appearing interferers cause a much lower mean squared error, and their cumulative effect is minimal as shown in Figure 9.

While it is likely that most applications of spatial-temporal processing of digital communications signals will employ known training sequences for adaptation and synchronization, <u>blind</u> adaptation (without the use of a training sequence) may also be useful, for example in broadcast situations, where receivers may have to recover from disruptions. Constant modulus blind algorithms are relatively simple and robust, but converge slowly. Faster convergence, at the expense of higher signal processing complexity, is achieved with oversampling, cyclostationarity-exploiting algorithms and subspace approaches. A good survey and comparison of these approaches is found in [10].

An example of a blind array adaptation approach which exploits cyclostationarity is found in [60]. In this approach, users' signals are separated by small frequency offsets (much less than signal bandwidth), and a reference signal used for DMI adaptation is just the complex output of one of the array elements. The adaptation uses a weighted sum of the results of several such reference signals. Adaptation is relatively fast, and can be accelerated to decision-directed adaptation once the array coefficients start to converge.

6. Transmitter Space-Time processing

Up to this point our discussion has mainly been relevant to space-time processing at a receiver. However, transmitter space and/or time processing can also be done at transmitters; an important application is to cellular wireless systems, where size and cost constraints limit or preclude array processing at mobile terminals, thus leaving the adaptive processing of downlink (base to mobile) signals to be done at the base transmitters. Assuming that transmission paths are linear, and that their vector impulse responses are known, or can be estimated, one can show that linear space and/or time processing can be done at transmitters, using zero-forcing or MMSE optimization criteria. If the channels are reciprocal, which means that their uplink (mobile to base) and downlink responses are identical, then the optimum transmitter space-time coefficients can be derived from (in fact they are identical to), the corresponding uplink coefficients. The assumption of reciprocity is valid if the same antenna elements are used for transmitting and receiving, and if uplink and downlink transmissions take place in the same frequency band, close enough in time that channels do not change significantly between uplink and downlink transmissions. Time division duplex (TDD) systems with burst durations smaller than the channel's coherence time satisfy reciprocity. FDD (frequency division duplex) systems generally do not, since widely separated frequency bands are used for uplink and downlink transmission.

A simple example can illustrate relationships between spatial processing at transmitters and receivers, under the assumption of reciprocity. Consider first two base stations, designated "1" and "2", which are receiving signals from two mobiles, also designated as "1" and "2"; mobile 1 is transmitting to base station 1, and provides interference at base station 2; mobile 2 is transmitting to base 2, and provides interference at base 1. The two base stations each have antenna N-dimensional array coefficient vectors $\mathbf{w}_1$ and $\mathbf{w}_2$, respectively. As shown in Figure 10, the complex vector responses among the two mobiles and two base stations are designated $\mathbf{h}_{11}$, $\mathbf{h}_{12}$, $\mathbf{h}_{21}$, and $\mathbf{h}_{22}$. If mobile 1 is transmitting data symbol a_1, and mobile 2 is transmitting a_2, the two array input vectors $\mathbf{r}_1$ and $\mathbf{r}_2$ at the two base stations are given in vector form as

$$\begin{pmatrix} \mathbf{r}_1 \\ \mathbf{r}_2 \end{pmatrix} = \begin{pmatrix} \mathbf{h}_{11} & \mathbf{h}_{12} \\ \mathbf{h}_{21} & \mathbf{h}_{22} \end{pmatrix} \begin{pmatrix} a_1 \\ a_2 \end{pmatrix} \tag{26}$$

The corresponding two array outputs, y_1 and y_2, are

$$\begin{pmatrix} y_1 \\ y_2 \end{pmatrix} = \begin{pmatrix} \mathbf{w}_1^H & \mathbf{0}^H \\ \mathbf{0}^H & \mathbf{w}_{12}^H \end{pmatrix} \begin{pmatrix} \mathbf{r}_1 \\ \mathbf{r}_2 \end{pmatrix} = \begin{pmatrix} \mathbf{w}_1^H & \mathbf{0}^H \\ \mathbf{0}^H & \mathbf{w}_{12}^H \end{pmatrix} \begin{pmatrix} \mathbf{h}_{11} & \mathbf{h}_{12} \\ \mathbf{h}_{21} & \mathbf{h}_{22} \end{pmatrix} \begin{pmatrix} a_1 \\ a_2 \end{pmatrix} \tag{27}$$

The following zero-forcing solution will eliminate interference, and yield the desired outputs $y_1=a_1$ and $y_2=a_2$ if

$$\mathbf{w}_1^H \mathbf{h}_{11} = 1;\ \mathbf{w}_1^H \mathbf{h}_{12} = 0;\ \ \mathbf{w}_2^H \mathbf{h}_{21} = 0 \ \text{ and } \mathbf{w}_2^H \mathbf{h}_{22} = 1. \tag{28}$$

Figure 11 shows the corresponding situation in which base stations 1 and 2 are transmitting to mobiles 1 and 2 respectively. The complex vector responses are the same, assuming reciprocity. If the same coefficient vectors are used, as shown in the figure, the two scalar outputs, z_1 and z_2 at the two mobiles are

$$\begin{pmatrix} z_1 & z_2 \end{pmatrix} = \begin{pmatrix} a_1 & a_2 \end{pmatrix} \begin{pmatrix} \mathbf{w}_1^H & \mathbf{0}^H \\ \mathbf{0}^H & \mathbf{w}_{12}^H \end{pmatrix} \begin{pmatrix} \mathbf{h}_{11} & \mathbf{h}_{12} \\ \mathbf{h}_{21} & \mathbf{h}_{22} \end{pmatrix} = \begin{pmatrix} a_1 & a_2 \end{pmatrix} \tag{29}$$

if the two transmitting coefficient vectors are the same as the optimal receiving coefficient vectors given by equation (28). While this example applies to zero-forcing optimization, MMSE optimization will behave in essentially the same fashion.

Thus when there is true reciprocity, as in a TDD system, base stations can adapt their arrays to minimize cochannel and intersymbol interference in the <u>received</u> signals, as in Figure 10, and then use the same array coefficient vectors when transmitting, as in Figure 11. As illustrated in the above example, this will shape the transmitting antenna arrays' beam patterns in such a way that downlink interference from base stations to mobiles is minimized or eliminated. Note that adaptation can be done during the receive (uplink) mode at each separate base station, without requiring coordination among base stations.

For non-reciprocal FDD systems, base station transmitter array processing, as in Figure 11 can also eliminate downlink interference if the coefficient vectors satisfy (28), but in this case, the complex vector responses $\{\mathbf{h}_{ij}\}$are different in the uplink and downlink directions. Application of (25) then requires separate estimation, using training sequences, of the downlink $\{\mathbf{h}_{ij}\}$, which can be complex, and require significant feedback signaling among base stations and mobiles[10,61]. In practice, transmitter-only spatial processing in such multi-base station systems can use of switching among directional transmit antenna beams at each base station. Alternatively, downlink adaptive approaches can be use averaging of data collected on the uplink. For example in [62], time averaging is used to remove uplink frequency-specific characteristics from estimates of the desired signal and interference covariance matrices. The results are used to determine downlink array coefficients which maximize average desired signal power while constraining average interference plus noise power to be below a fixed level. A related approach, involving a transformation of the uplink covariance matrix is described in [63].

7. Conclusions and Future Applications

Adaptive spatial and temporal linear combining both have the ability to minimize intersymbol interference arising from multipath, and co- and adjacent-channel interference in multiuser wireless digital communications. They can be used singly or together, to allow a number of users to share the same time, bandwidth, and space. We have seen that the number of interfering signals which can be separated at a receiver is proportional to the number of antenna elements and also to the received signals' excess bandwidth. We have mainly considered linear space-time processing methods here since their implementation and adaptation is relatively simple. However nonlinear interference cancellation techniques will also be useful, especially at cellular base stations for removing the strongest interferers.

Smart antennas have a capacity-multiplying effect in cellular systems that employ them. They make possible space division multiple access (SDMA), in which individual channels may be reused by different users within a cell, without performance loss due to interference. Smart antennas have been proposed for capacity and reliability enhancement in Third Generation wireless systems [64]. Indoor and outdoor broadband cellular systems will also benefit from the use of smart antennas, starting first with directive, switched-beam systems [65,66], and eventually incorporating full array adaptation and temporal processing [67] for interference and multipath mitigation. The interference-elimination properties of spatial-temporal processing will also likely find application in unlicensed wireless broadband systems, such as those expected to operate in three recently-allocated 100 MHz-wide U-NII (unlicensed national information infrastructure) bands in the USA [68]. Users of these bands will operate in an interference environment that is essentially uncontrolled, except for some basic rules on transmitted power, power spectral density and antenna gain. Array processing, coupled with other techniques such as coding and smart multiple access protocols, will be essential, once the U-NII bands become heavily used.

A significant remaining problem area is implementation complexity. For example the realization of spatial-temporal processing at a receiver generally requires a separate RF front end, together with a downconverter and A/D converter, to allow for digital processing of the baseband or IF output of each antenna element [69]. An alternative approach – direct weighting and combining at RF – is not supported by mature, cost-effective technology at this point. Thus the development of low cost, low power single chip radio realizations are of prime importance.

8. Acknowledgements

It is a pleasure to acknowledge very helpful discussions and suggestions from Lek Ariyavisitakul of Home Wireless Networks, Abdelgader Legnain and Sébastien Roy of Carleton University, Mansoor Shafi of Telecom New Zealand and John Thompson of the University of Edinburgh.

FIGURE CAPTIONS

Fig. 1 Receiving array channel model

Fig. 2 General linear space-time processor

Fig. 3 Illustration of the effect of excess bandwidth on interference suppression capability of a linear space-time processor

Fig. 4 Illustration of centralized decision feedback for K=2 signals, with full cancellation of signal “0” from delayed signal “1”.

Fig. 5 S-T processing of CDMA signal where filter memory=2 bit intervals (2T)

Fig. 6 Effect of the Number of Coefficients and Number of Interfering Signals. Number of antenna elements=4 [from [48]]. Curves from bottom to top represent numbers of interfering users from 2 to 18 in steps of 2.

Fig. 7 Number of Interferers and Number of Antenna Elements. Number of forward taps =14. [from [48]].

Fig. 8 Capacity comparison for 1, 4 and 9 antenna elements: linear S-T processor, decision feedback S-T processor and centralized decision feedback S-T processor. 8 forward taps and 5 feedback taps. [from [37]]

Fig. 9 Effect of Sudden Birth of Interference. The system starts with 6 users, after which a new user is added every 300 symbols. [from [59]]

Fig. 10 Mobile to base transmission. Processing at base receivers.

Fig. 11 Base to mobile transmission. Processing at base transmitters.

Fig. 1 Receiving array channel model

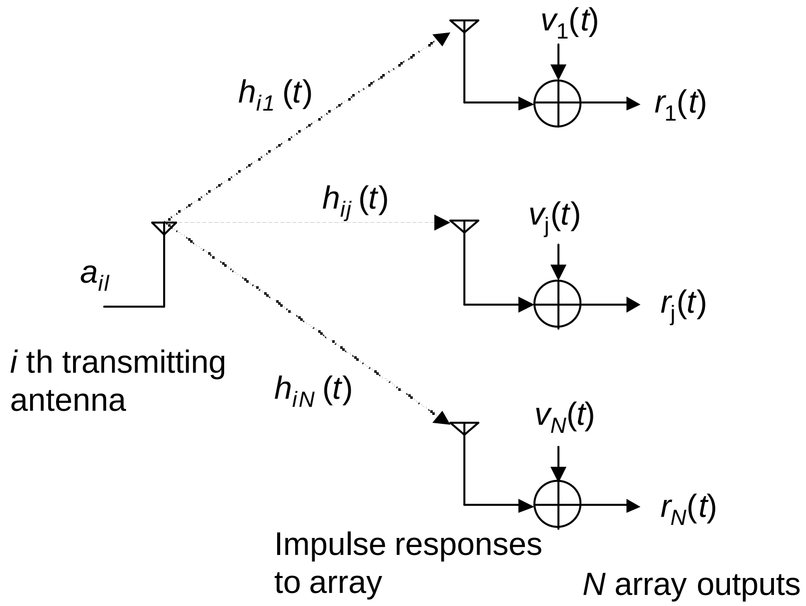


Fig. 2 General linear space-time processor

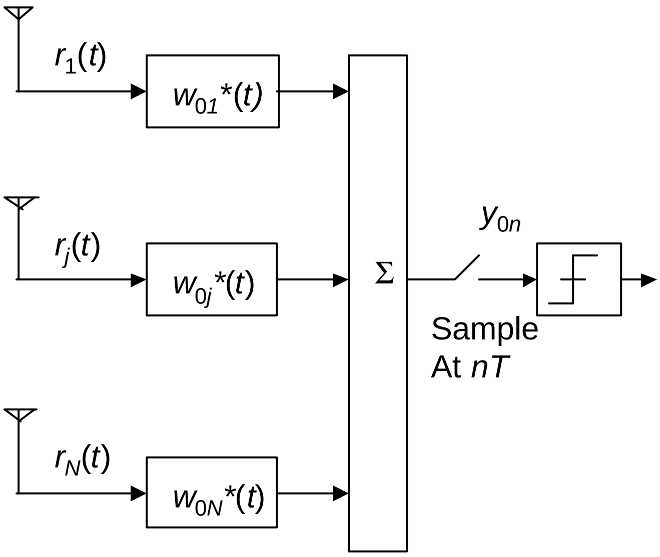


Fig. 3 Illustration of the effect of excess bandwidth on interference suppression capability of a linear space-time processor

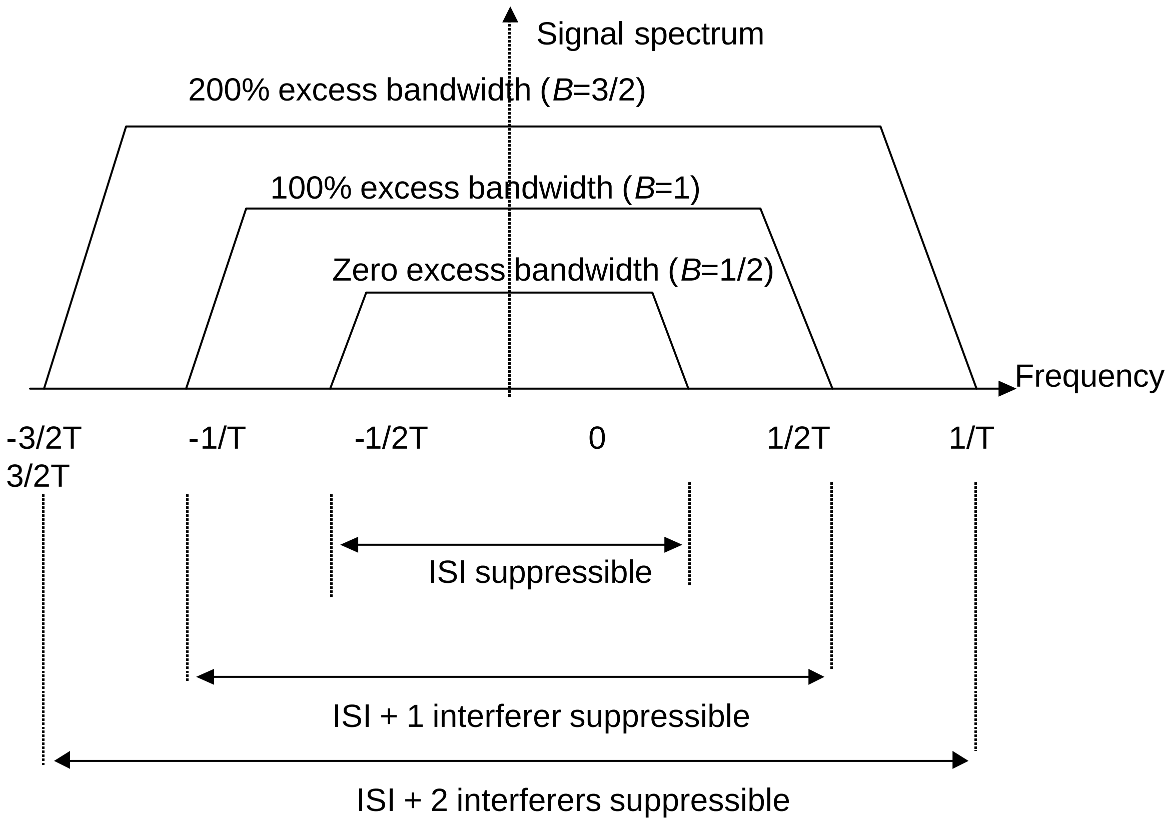


Fig. 4 Illustration of centralized decision feedback for K=2 signals

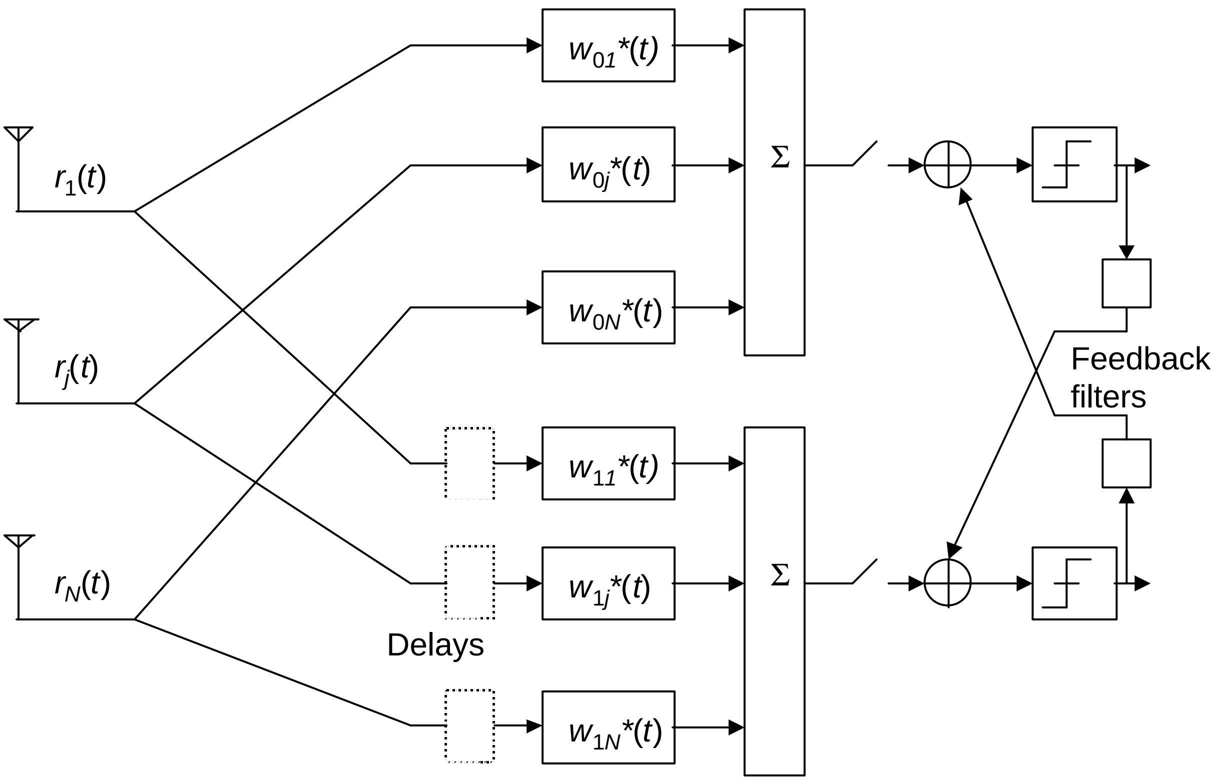


Fig. 5 S-T processing of CDMA signal where filter memory=2 bit intervals (2T)

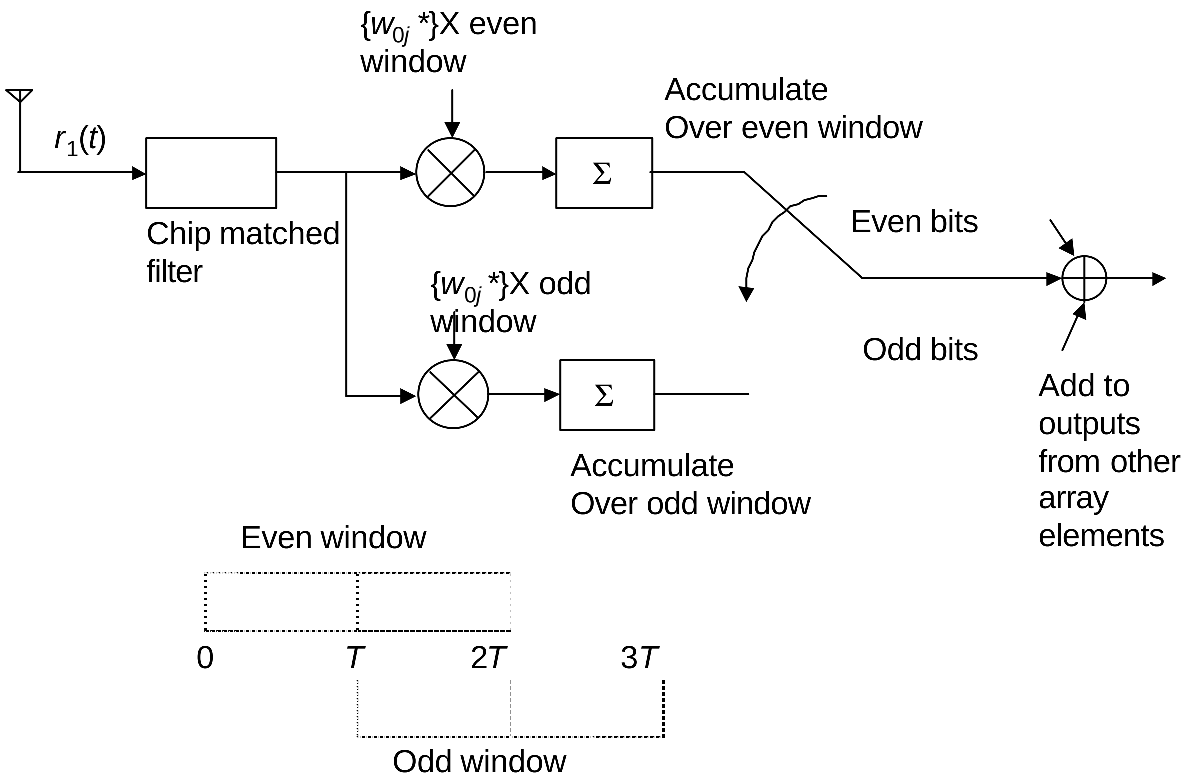


Fig. 6 Effect of the Number of Coefficients and Number of Interfering Signals

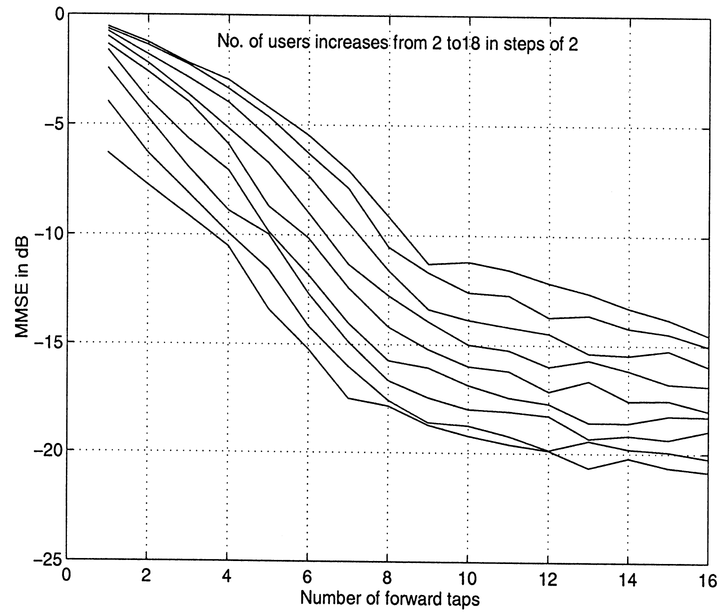


Fig. 7 Number of Interferers and Number of Antenna Elements

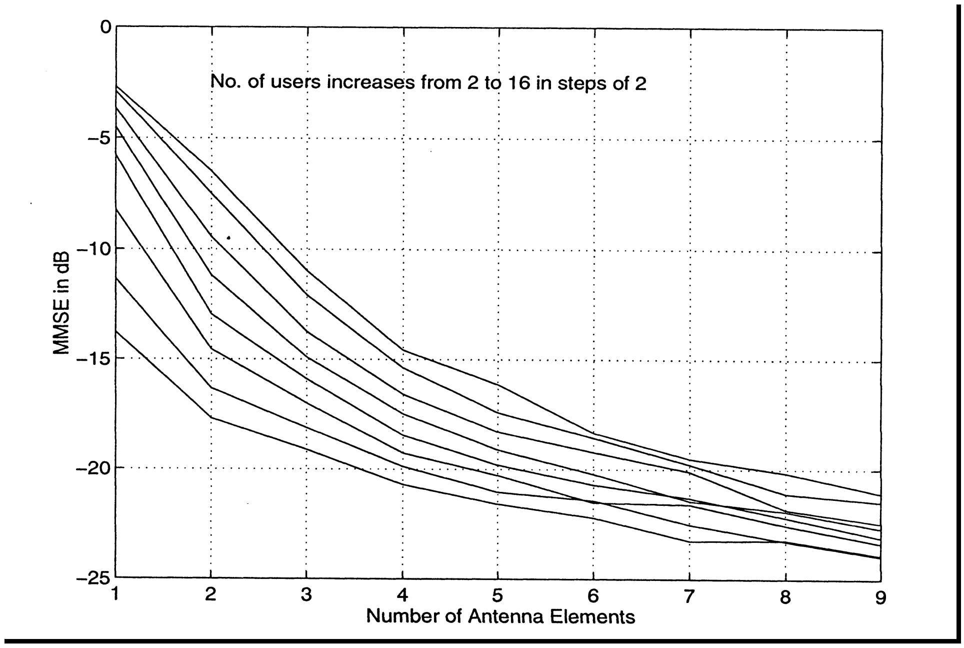


Fig. 8 Capacity comparison for 1, 4 and 9 antenna elements: linear S-T processor, decision feedback S-T processor and centralized decision feedback S-T processor. 8 forward taps and 5 feedback taps

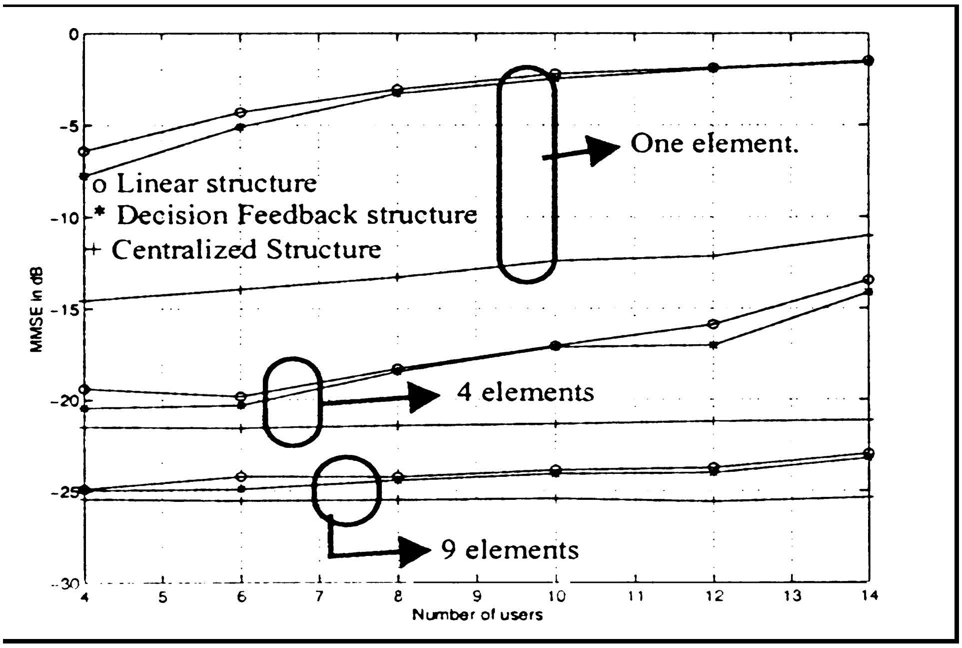


Fig. 9 Effect of Sudden Birth of Interference

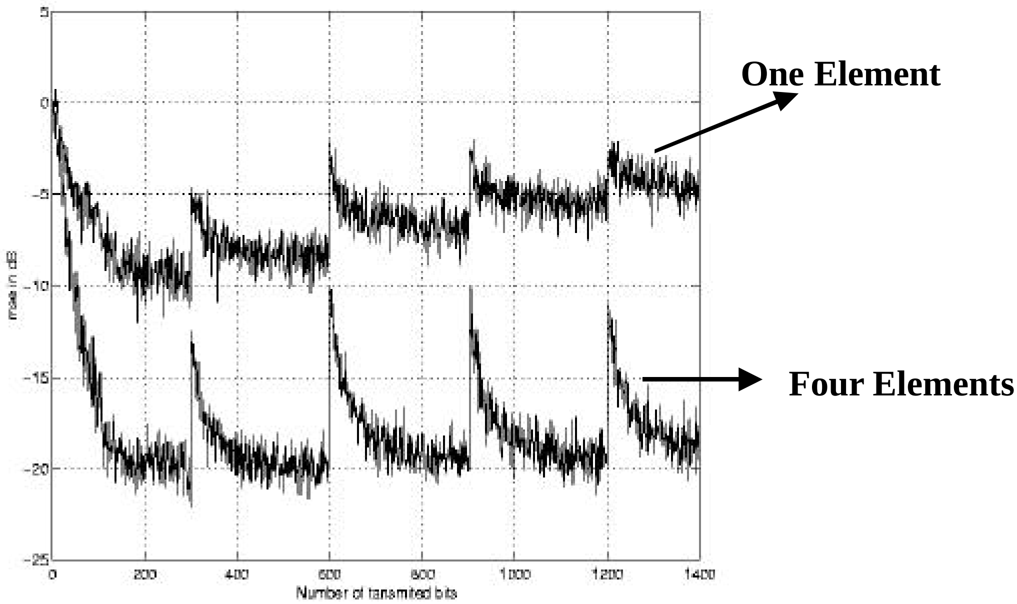


Fig. 10 Mobile to base transmission. Processing at base receivers.

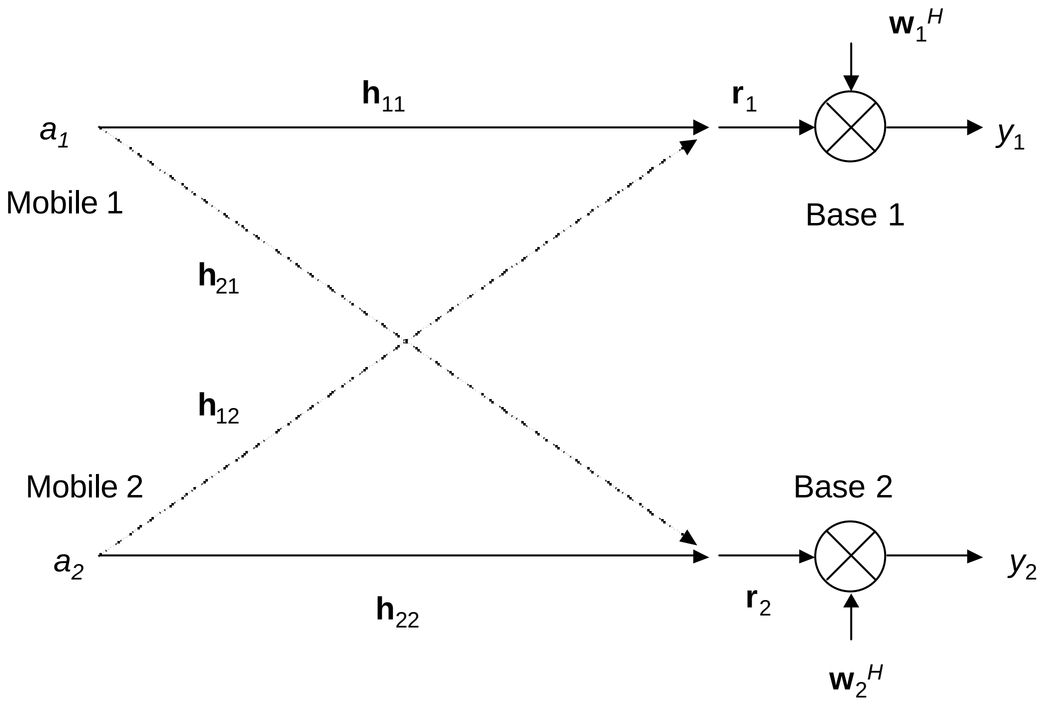
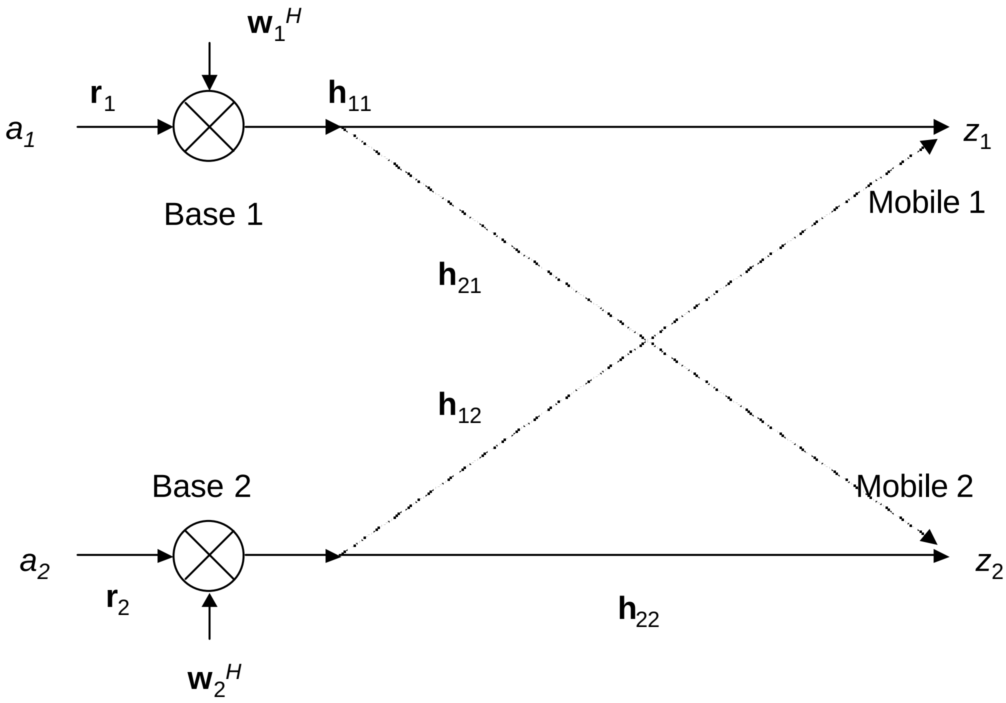


Fig. 11 Base to mobile transmission. Processing at base transmitters.



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