Subband Adaptive Antenna Array for Wideband Wireless Communications

Y. Zhang, S. Weiss and L. Hanzo

Communications Research Group
Department of Electronics and Computer Science
University of Southampton, S017 1BJ, UK
{yz, s.weiss, l.hanzo} @ecs.soton.ac.uk

Abstract: Subband array processing are considered a promising method in the fields of echo cancellation and sonar. With the growth of wideband communication service, computation load and real time performance of adaptive antenna array become one of the key issues in wideband mobile environment. In the paper we apply subband processing technology to wideband beamforming of mobile communications instead of conventional space-time processing. By using this scheme, The reduction in computation becomes viable with additional advantages such as parallelisation of processing tasks and reduced spectral dynamics.

1. Introduction

With the development of multimedia mobile communications, the bandwidth of the signals, such as wideband CDMA signals, become larger and larger . This is a great challenge to communication systems including array antenna systems. The bandwidth of signals incident on the array has a significant impact on the ability of the array to reject interference. A narrowband array is able to form nulls at a single frequency and make excellent performance. When the elements space is $\lambda_0/2 = c/(2f_0)$, it is well-known that an array steering vector $a(\theta)$ can be given by [1]

$$a(\theta) = \left[1, e^{-j\pi\frac{f}{f_0}\cos\theta}, \cdots\cdots, e^{-j\pi(M-1)\frac{f}{f_0}\cos\theta}\right]^T \qquad (1)$$

where, f_0 is center frequency of the signals, θ is incident angle of signal to the array and M is element number of the array. The objective of any array beamforming can be expressed as

$$w^H a(\theta_d) = 1 \quad and \quad w^H a(\theta_{other}) = 0 \qquad (2)$$

In equation (2), w is array weight vector, θ_d is DOA of the desired signal, θ_{other} is DOA of the other interference signals. Assuming that signal bandwidth B is big relative to the center frequency, namely, $f \in [f_0 - B/2 \cdot \ f_0 + B/2]$, it can be easily shown that it is not possible to achieve an arbitrary antenna pattern across all frequency. This is because the steering vectors associated with array are a function of both the frequency of the incident signal and the spacing of array elements. A traditional method for the problem is combining spatial filtering with temporal filtering which makes the response of the array the same across different frequencies[1]. A tapped-delay-line on each branch of the array allows each element to have a phase response that varies with frequency, compensating for the fact that lower frequency signal components have less phase shift for a given propagation distance, whereas higher frequency signal components have greater phase shift as they travel the same length. However, a key shortcoming of the approach is slow convergence and heavy computation load.

In addition, when signal bandwidth B and element number M is relatively big simultaneously so that the time difference that different signal frequency components travel across the whole array is closer to element space, this will cause serious beamforming problem, even aliasing.

In this paper, we offer an excellent method to overcome the above problems, namely, subband array beamforming method. The basic idea of the subband approach is to decompose a fullband signal by means of a filter bank into a number of frequency bands, which can be run at a lower sampling rate due to the reduced bandwidth. Any computational costly processing can then be performed in the decimated subbands at a lower update rate[2]. Subband array processing can removes the necessity of temporal equalization and realizes the same results as a space-time adaptive array. The reduction in complexity by the subband approach becomes viable due to decreased complexity by processing in decimated

0-7803-7486-X/02/$17.00 © 2002 IEEE.

subbands. Furthermore, the separation into frequency bands can bring additional advantages such as parallelisation of processing tasks and reduced spectral dynamics[2].

2. Subband adaptive array

2.1 Signal models

For wideband CDMA systems, when down-converting to baseband, we can write the complex baseband received signals at antenna array as

$$X(t)=\sum_{i=1}^{I}\sum_{l=1}^{L}p_{i,l}(t)b_i(t-\tau_{i,l})c_i(t-\tau_{i,l})a_{i,l}+n(t) \quad (3)$$

where, $X(t)=[x_0(t),x_1(t),\cdots,x_{M-1}(t)]^T$, I and L is respectively the number of users and multipath components per user, $p_{i,l}$ and $\tau_{i,l}$ is respectively amplitude gain and time delay for the ith user in lth multipath component, b_i is source signal bit of the ith user, c_i is spreading code for the ith user, $a_{i,l}$ is array steering vector for the ith user in lth multipath component defined in equation (1), $n(t)$ is additional white noise.

2.2 Subband adaptive filter (SAF)

The flow graph of an analysis filter bank implementing a subband decomposition is shown in Fig.1. the received baseband CDMA signal x_m(t) at mth array element is digitized into $x_m(n)$ and is decomposed into subbands by an analysis filter bank with K bandpass filters, and subsequently each subband is decimated by a factor of $N<=K$. When the decimation ratio N equals the number of uniform subbands K, that is called the case of critical decimation with some drawbacks[2]. Oversampled SAF systems, with a decimation ration $N < K$, are designed such that after decimation the alias level within the subbands is kept sufficient low. Therefore, we focus on an oversampled SAF system that are based on generalized discrete Fourier transform (GDFT) filter banks, performing a particular type of complex valued subband decomposition. A dual operation is performed by the synthesis banks which restores the original sampling rate by upsampling, and interpolates with appropriate bandpass filters. The summation over the various branches yields an output signal $y_m(n)$, which ideally is only a delayed version of the input $x_m(n)$.

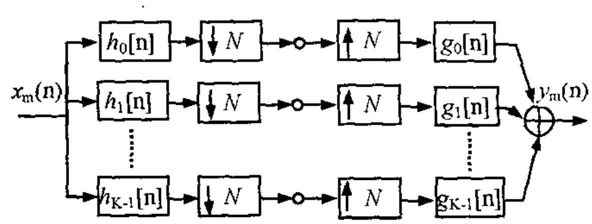

Fig.1 Decomposition and reconstruction of a signal x_m(n)

The analysis filter h_k[n] can be derived from a real valued lowpass prototype FIR filter $f[n]$ of length L_f by GDFT,

$$h_k[n]=e^{j\frac{2\pi}{K}(k+k_0)(n+n_0)}f[n], \quad k=0:1:L_f-1, \quad n=0:1:K-1 \quad (4)$$

the term GDFT stems from offsets k and n introduced into the frequency and time indices. With the choice of k_0 = 0.5 for even K, the frequency interval $\Omega\in[0;\pi]$ will be covered by exactly $K/2$ subband signals, while the rest is complex conjugate versions of these. On the synthesis side, the $K/2$ unprocessed subbands can be restored by a real operation Re{ }. At the same time, with the choice $n_0=-(L_f-1)/2$ and starting from a real valued linear phase prototype $f[n]$, both the real and imaginary part of $h_k[n]$ will separately satisfy linear phase conditions.
The synthesis filters $g_k[n]$ can be obtained by time reversion and complex conjugation of the analysis filters,

$$g_k[n]=h_k^*[L_f-n-1] \quad (5)$$

The modulation approach allows for both low memory consumption for storing filter coefficients and an efficient polyphase implementation. Through the modulation, the filter bank design reduces to an appropriate choice of the prototype filter with a passband width $2\pi/K$, which has to fulfill two criteria. Firstly, the filter's attenuation in the stopband, $\Omega\in[\pi/N;\pi]$, has to be sufficiently large. Every frequency of the input signal $x_m[n]$ lying within the interval $\Omega\in[\pi/N;\pi]$ will be aliased into the baseband after filtering and decimation, and cause a distortion of the subband signal. A second constraint on the design is the perfect reconstruction condition. If stopband attenuation of the prototype filter is high enough to sufficiently suppress aliasing, this condition reduces to the consideration of inaccuracies in power complementarity,

$$\sum_{k=0}^{K-1}\left|H_k(e^{j\Omega})\right|^2 \cong 1 \qquad (6)$$

A prototype filter approximating these constraints can be constructed by an iterative least-squares method and dyadically iterated halfband filters[2].

2.3 GSC_LMS adaptive algorithm

With only spatial processing, the weights defined by equation (2) can remove interference and receive the desired signal, but possibly amplify the noise simultaneously, so the optimum solution for choosing the suband weight vector can be formulated as,

$$w_{l,k} = \arg\min_{w_{l,k}} w_{l,k}^H R_{x_k x_k} w_{l,k} \quad subject\ to\ \ C_l^H w_{l,k} = 1 \quad (7)$$

where $R_{x_k x_k}$ is the array data covariance matrix for the kth subband, $w_{l,k}$ is the kth subband weight vector for the lth multipath signal, C_l is constraint vector for the lth multipath of the desired user and can be given by equation (1) [4].

It is well-known that the generalized sidelobe canceller (GSC) can transform the constrained minimization problem in (7) into an unconstrained form shown in Fig.2. w_c is a data independent beamformer that determines which spatial regions are to be preserved and/or eliminated. The blocking matrix B will block signal from the desired directions. The choice for w_c and B implies that the constraints are satisfied independently of w_a. Therefore, w_a is adjusted by some typical adaptive algorithms such that the variance of the output is minimized, it will try to suppress any signal components in other spatial regions.

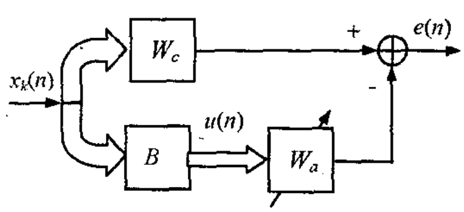

Fig.2 Structure of GSC_NLMS beamformer

GSC adaptive beamformer employing the NLMS algorithm is following as[3]

1. $u(n) = B^H x_k(n)$ (8.1)

2. $e(n) = w_c^H x(n) - w_a^H(n)u(n)$ (8.2)

3. $\mu = \mu'[u^H(n)u(n)]^{-1}$ (8.3)

4. $w_a[n+1] = w_a(n) + \mu e^*(n)u(n)$ (8.4)

2.4 Performance of CDMA subband array systems

Fig.3 is a system diagram for CDMA subband array processing. Firstly, the received data of each element is decomposed into K subband by analysis filters and decimation with a factor $N<K$. Then, GSC_NLMS algorithm perform adaptive subband beamforming for different subband components. Synthesis filters S will reconstruct fullband beamforming for a multipath signal of the desired user. If subband processing block in Fig.3 is performed repeatedly with different constraint vector C_l for different multipath signal, we will get all of L multipath signals $O_{0\sim L-1}$ of the desired user.

Finally, we combine all multipath signals to further enhance the SINR for offering strong resistance against fast fading in mobile communication. Through beamforming and matched filter (namely, depreading and integration), the signal power in y_l becomes significantly higher than the sum of interference and noise, this enables us to approximate the optimum combining vector using the principal eigenvector of R_{yy}, which can be obtained using standard decomposition techniques[5],

$$w_{opt} = R_{yy}^{-1} q \qquad (9)$$

the channel vector of the desired user q can be estimated by the principal eignvector of autocorrelation matrix R_{yy}. If we set $w_{opt,l} = 1$, equation (9) will decay to a equal gain combination.

3. Simulation example

Assuming the length of gold code c_i for CDMA system is 31 and The number of array element 5, there are 5 multipath signals for each user and time delay between adjacent multipaths is one chip, amplitude gain $p_{i,l}$ obey rayleigh frequency-selective fading, DOA of any multipath signals are randomly distributed over [0, 180^0].

The subband structure is characterized by the filter banks with K=16 channels decimated by N=14 based on a prototype filter of length L_f= 448. Fig 4 shows the convergence curve for different user numbers. Convergence curve can be changed by adjusting step size for GSC_NLMS. In actual calculation, we find that the time for runing fullband beamforming is much longer than that of subband beamforming. In fact, due to applying subband processing, this yields a reduction in computational complexity by a factor $\zeta(K/N^2)$ for NLMS-type algorithms[2]. Fig 5 demonstrates chip error ratio (CER) with different user number for subband beamforming when iterative times keep equal. If user number <= 6, BER (bit error ratio) is lower than 0.001.

4. Conclusions

A subband adaptive beamforming structure for wideband mobile communications has been proposed. Whereby the sensor signals are split into decimated but oversampled subbands. In each subband, a separate beamforming algorithm - just like in the fullband - can be operated. The advantages are the reduction in computational cost and an increase in convergence speed for subband adaptive filter schemes while performance ability of the system keep invariance.

REFERENCES

1. Jr. J C Liberti, T S Rappaport. *Smart Antennas for Wireless Communications*. Prentice Hall, 1999
2. S. Weiss and R. W. Stewart. *On Adaptive Filtering in Oversampled Subbands*. Shaker Verlag, Aachen, Germany, 1998
3. S.Weiss, R.W.Stewart, etc. An efficient scheme for broadband adaptive beamforming. In *Asilomar Conference on Signals, Systems, and Computers*, Monterey, CA, November, 1999.
4. S. Haykin. *Adaptive Filter Theory*(3rd Ed.). Prentice Hall, 1996
5. H. Liu, K. Li. A decorrelating RAKE receiver for CDMA communications over frequency-selective fading channels. *IEEE Trans. Commun.*, 1999, 47: 1036~1045

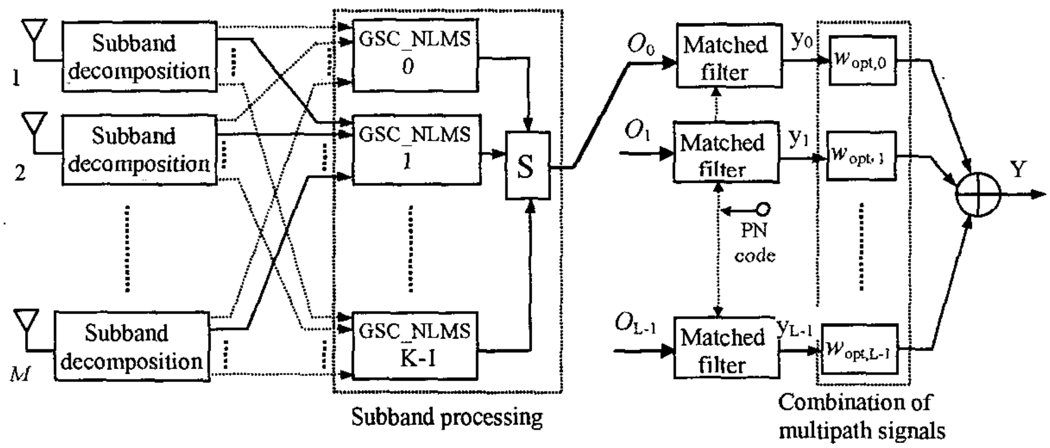

Fig.3 Structure of subband adaptive antenna array for WCDMA systems

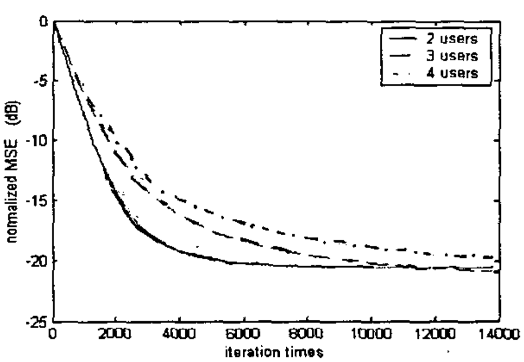

Fig.4 Convergence behaviour of subband beamforming for different user number

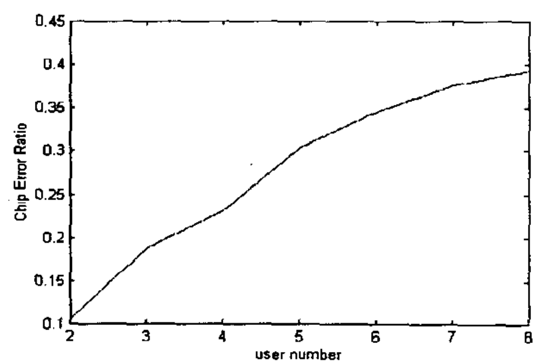

Fig.5 CER(chip error ratio) with user number for subband beamforming