Institute for Neural Computation

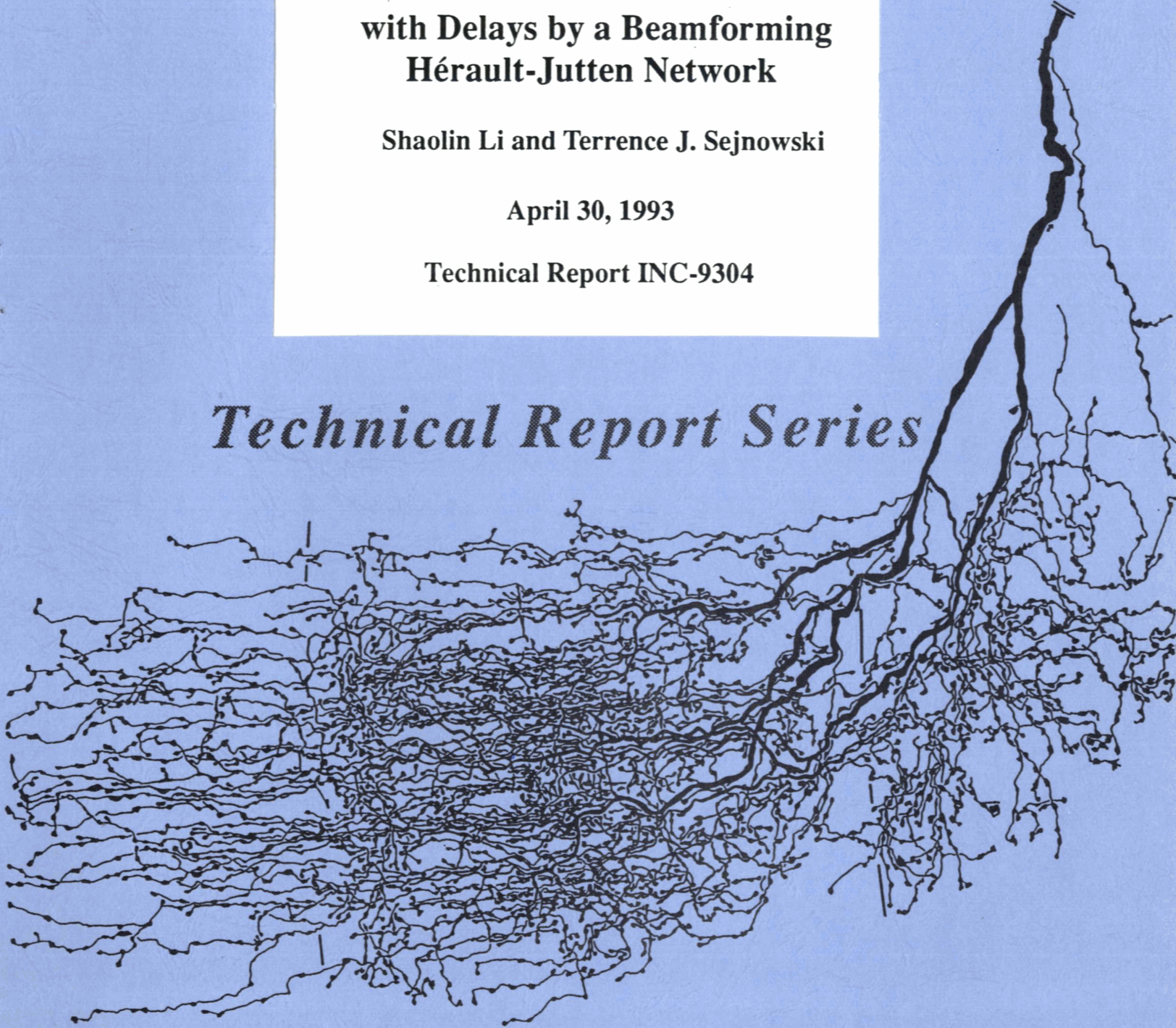
Adaptive Separation of Mixed Broadband Sound Sources with Delays by a Beamforming Hérault-Jutten Network

Shaolin Li and Terrence J. Sejnowski

April 30, 1993

Technical Report INC-9304

Technical Report Series



University of California, San Diego La Jolla, California 92093

Adaptive Separation of Mixed Broadband Sound Sources with Delays by a Beamforming Hérault-Jutten Network

Shaolin Li and Terrence J. Sejnowski

April 30, 1993

Technical Report INC-9304

Institute for Neural Computation
University of California, San Diego
9500 Gilman Drive
La Jolla, California 92093-0523

Shaolin Li is with the Howard Hughes Medical Institute and Terrence J. Sejnowski is with The Salk Institute, Howard Hughes Medical Institute, and the Institute for Neural Computation at the University of California, San Diego.

Correspondence concerning this article may be addressed to: Shaolin Li, Computational Neurobiology Laboratory, The Salk Institute, P.O. Box 85800, San Diego, CA 92186-5800.

Adaptive Separation of Mixed Broadband Sound Sources with Delays by a Beamforming Hérault-Jutten Network

Shaolin Li and Terrence J. Sejnowski
Computational Neurobiology Laboratory
The Salk Institute, P.O. Box 85800
San Diego, CA 92186-5800

Abstract

The Hérault-Jutten network has been used to separate independent sound sources that have been linearly mixed. The problem of separating a mixture of several independent signals in free field conditions or a signal and echoes in confined spaces is compounded by propagation time delays between the source(s) and the microphones because the conventional Hérault-Jutten network cannot tolerate time delays. In this paper, we combine a symmetrically balanced beamforming array with the conventional Hérault-Jutten network. The resulting system can adaptively separate signals that include delays introduced by the propagation medium. The proposed algorithm has been simulated in digital communication multipath channels where intersymbol interference exists. The simulation results show two clear advantages of the proposed method over the conventional adaptive equalization: 1) There is no penalty for very long impulse responses caused by long delays. 2) No training signals are needed for equalization. The design for a multi-beamformer to handle the source separation of multiple broadband signals is also presented.

1 Introduction

The problem of separating mixtures of independent signals or their delayed versions is encountered in many fields: in the "cocktail party" problem, a person wants to listen to a single sound source and filter out other interfering sources; in underwater acoustic digital communication, a receiver needs to filter out delayed versions of the transmitted signal in order to eliminate intersymbol interference (ISI). In both circumstances the locations of sources and/or their echoes are unknown and may vary over the time.

The Hérault-Jutten (HJ) network can deal with these types of problems because of its adaptive separation ability [1]-[4]. However, in many common circumstances, such as in the "cocktail party" problem, it is not possible to obtain N distinct linear combinations of N signals without delays or phase shifts required by the HJ network. In order to generate N full rank linear combinations of inputs to the HJ network, N microphones have to be placed at different locations for N signal sources located at different places. The propagating medium between the sound sources and the microphone produces different weights on the arrivals of different sound source signals as required, and introduces significant signal delays, which cannot be handled by the conventional HJ network. Platt and Faggin [7] have proposed an extension of the HJ network to estimate the matrix of delays, together with the mixing matrix $\mathbf{A}$.

We have investigated another method for coping with unknown delays that relies on beamforming. An array of microphones can be organized to produce N fixed beam lobes without phase shifts. This is equivalent to having N virtual directional microphones at a single spatial location each aimed at a different directions. If all of the sound sources have different directions, N full-rank linear combinations of N signals are formed, as required for the HJ network. A brief report of this method was given in [9]

2 An overview of the Hérault-Jutten network

The HJ network is an $N \times N$ network [1]-[4] which can be used to solve the following signal processing problem: given N observed data sequences $\mathbf{Y}(t) \triangleq [y_1(t), y_2(t), ..., y_N(t)]^T$ which are distinct linear combinations (full rank) of N physical independent signals [1] $\mathbf{X}(t) \triangleq [x_1(t), x_2(t), ..., x_N(t)]^T$ without time delays or phase shifts, the network can adaptively recover $\mathbf{X}(t)$ without a priori knowledge of the mixing matrix $\mathbf{A}$. In mathematical form,

$$\underbrace{\begin{bmatrix} y_1(t) \\ y_2(t) \\ \vdots \\ \vdots \\ y_N(t) \end{bmatrix}}_{\text{Observed Data } \mathbf{Y}(t)} = \underbrace{\begin{bmatrix} a_{11} & a_{12} & \cdots & a_{1N} \\ a_{21} & & \cdots & a_{2N} \\ \vdots & & & \vdots \\ \vdots & & & \vdots \\ a_{N1} & a_{N2} & \cdots & a_{NN} \end{bmatrix}}_{\text{Mixing Matrix } \mathbf{A}} \underbrace{\begin{bmatrix} x_1(t) \\ x_2(t) \\ \vdots \\ \vdots \\ x_N(t) \end{bmatrix}}_{\text{Original Signal } \mathbf{X}(t)} \tag{2.1}$$

where $\mathbf{Y}(t)$ is the known observed data, $\mathbf{A}$ is an unknown full rank matrix and $\mathbf{X}(t)$ is the signals to be recovered. It is important to note that Eq.(2.1) is a memoryless

[1] For $\mathbf{X}(t)$ to be independent signals is a sufficient but not a necessary condition. The HJ network can converge as long as $\int_0^T x_i(t)^{2k+1} x_j(t)^{2l+1} dt = 0, i \neq j$, see [1]-[4].

system. Thus, at a given time $t = t_1$, the output $\mathbf{Y}(t_1)$ depends only on $\mathbf{X}(t_1)$ and not om any other past or future values of $\mathbf{X}(t)$.

Introduce a matrix $(\mathbf{I} + \mathbf{C})$, where $\mathbf{I}$ is identity matrix and

$$\mathbf{C} \triangleq \begin{bmatrix} 0 & & c_{ij} \\ & \ddots & \\ c_{ji} & & 0 \end{bmatrix} \tag{2.2}$$

The output of HJ network is $\mathbf{S}(t)$

$$\mathbf{S}(t) \triangleq \begin{bmatrix} s_1(t) \\ \vdots \\ s_N(t) \end{bmatrix} = (\mathbf{I} + \mathbf{C})^{-1} \begin{bmatrix} y_1(t) \\ \vdots \\ y_N(t) \end{bmatrix} \tag{2.3}$$

The adaptive rule used in HJ network is

$$dc_{ij}/dt = -\mu f(s_i) g(s_j) \tag{2.4}$$

where $f(s)$, $g(s)$ are different odd functions. In later simulations we use $f(s) = s^3$ and $g(s) = s$. The quantity μ sets the rate of adaptation. When the network learning approaches an equilibrium point,

$$\begin{bmatrix} s_1(t) \\ \vdots \\ s_N(t) \end{bmatrix} \Longrightarrow \begin{bmatrix} x_1(t) \\ \vdots \\ x_N(t) \end{bmatrix} \tag{2.5}$$

At the equilibrium point, $\mathbf{I} + \mathbf{C}$ is equal to $\mathbf{PDA}$, where $\mathbf{A}$ is defined in Eq.(2.1), $\mathbf{D}$ is a diagonal matrix and $\mathbf{P}$ is a permutation matrix (see [1]-[4]).

3 Separating signals from different directions

In this section a beamformer is constructed to form multi-beams using inputs $x_i(t)$ $(i = 1, ..., N)$, which are independent signals or a signal with its delayed versions, and propagated from different directions. Each beam output of the beamformer is a distinct linear combination of $x_i(t)$ $(i = 1, .., N)$ as required in Eq.(2.1).

Assume $2M + 1$ sensors are placed in a straight line a distance δ apart. Let the signal received by the sensor at the center be $z_0(t)$. A beam is formed by adding delays to the output of each sensor and summing them up

$$y_n(t) = \sum_{m=-M}^{M} w_m z_m(t - m\tau_n) \tag{3.6}$$

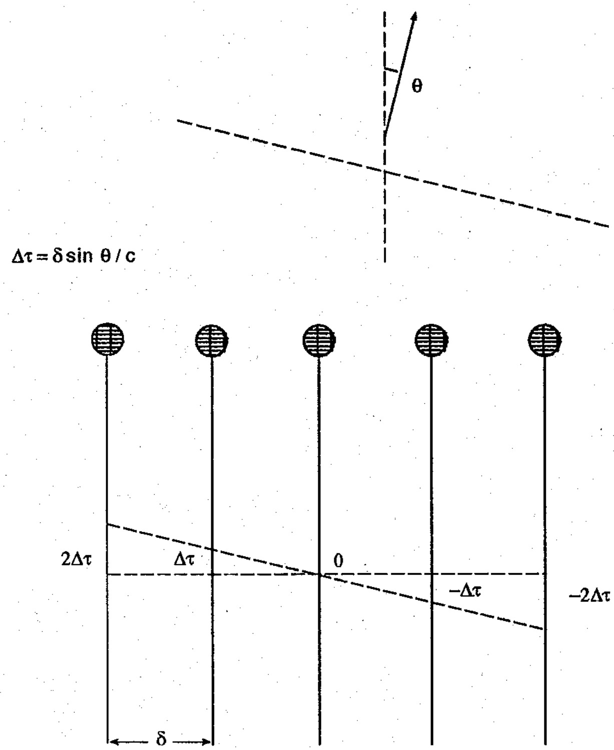


Figure 1: The array and arriving plane wave geometry

where $y_n(t)$ is the output of the n^{th} beam, $z_m(t)$ is the output of the m^{th} sensor and w_m is the weight to form a beam.

Assume that a plane wave $x_i(t)$ is arriving from an angle θ_i (see Fig. 1) . The output of sensor m is

$$z_m(t) = x_i(t - m\tau_i) \tag{3.7}$$

where $\tau_i = \frac{\delta \sin(\theta_i)}{c}$ is the time delay between neighboring sensors and c is the wave propagation speed in the medium. [2] Substituting Eq.(3.7) into Eq.(3.6), one obtains

[2] We use i as an index of incoming signals, n as an index of the beams formed and m as an index of sensors.

$$y_n(t) = \sum_{m=-M}^{M} w_m x_i \left[t - m(\tau_n + \tau_i)\right] \tag{3.8}$$

From Eq.(3.8) the beam direction can be steered by changing τ_n. If the weights are symmetric to the center $w_m = w_{-m}$, and $w_0 = 1$, the Fourier transform of Eq.(3.8) is

$$\begin{aligned} Y_n(\omega) &= X_i(\omega) \sum_{m=-M}^{M} w_m \exp^{-im\omega(\tau_n+\tau_i)} \\ &= X_i(\omega) \underbrace{\left[1 + 2\sum_{m=1}^{M} w_m \cos\left[m\omega(\tau_n + \tau_i)\right]\right]}_{\triangleq a_{ni}(\omega)} \end{aligned} \tag{3.9}$$

Since $a_{ni}(\omega)$ in Eq.(3.9) has a real value, there are no phase shifts or delays for any values of ω,τ_i and τ_n. The symmetry of the array with respect to the central sensor eliminates the phase shifts between the beam outputs. Thus, when exact delay lines are used to form the beam, the phase shifts of $-m^{th}$ and $+m^{th}$ sensor outputs exactly cancel. For narrowband signals centered at ω_0, $a_{ni}(\omega_0)$ can be considered approximately constant over the bandwidth. For I incoming signals from different angles θ_i, one can form $N = I$ beams with different τ_n aimed at N different but fixed directions, the narrowband sources will produce N linear combinations of signals without delays, which can then serve as inputs to an HJ network. This is equivalent to having N virtual directional microphones at a single spatial location (at the central sensor) each aimed at a different directions. If all of the sound sources have different directions, N full-rank linear combinations of N signals are formed, as required for the HJ network. In the time domain Eq.(3.9) has the same form as Eq.(2.1),

$$\underbrace{\begin{bmatrix} y_1(t) \\ y_2(t) \\ \vdots \\ \vdots \\ y_N(t) \end{bmatrix}}_{\text{Outputs of } N \text{ Beams}} = \underbrace{\begin{bmatrix} a_{11}(\omega_0) & a_{12}(\omega_0) & \cdots & a_{1N}(\omega_0) \\ a_{21}(\omega_0) & & \cdots & a_{2N}(\omega_0) \\ \vdots & & & \vdots \\ \vdots & & & \vdots \\ a_{N1}(\omega_0) & a_{N2}(\omega_0) & \cdots & a_{NN}(\omega_0) \end{bmatrix}}_{\text{Mixing Matrix } \mathbf{A}} \underbrace{\begin{bmatrix} x_1(t) \\ x_2(t) \\ \vdots \\ \vdots \\ x_N(t) \end{bmatrix}}_{\substack{\text{Original Signals} \\ \text{from different directions}}} \tag{3.10}$$

A conventional HJ network uses amplitude differences to form the different weight a_{ij} in **A**; our method relies on differences in beam magnitudes for signal arrivals from different directions. Passing the N outputs of beams through an $N \times N$ HJ network, the output should recover the original signals $x_i(t)$ when the HJ network has been reached equilibrium.

The separation of multiple broadband signal sources is not feasible with the beamformer described above because the directional sensitivity of the sensor array decreases from high frequency to low frequency. In that case, $a_{ni}(\omega)$ in Eq.(3.9) will change with frequency, and Eq.(3.10) is no longer correct for the broadband signals. However, this problem could be solved by replacing the weights w_m in Eq.(3.8) with filters on $x_i(t)$ before forming a beam. With this approach, it should be possible to construct a beamformer whose beam lobe width is frequency-invariant over a broad bandwidth. Therefore Eq.(3.8) becomes

$$y_n(t) = \sum_{m=-M}^{M} h_m(t) * x_i\left[t - m(\tau_n + \tau_i)\right] \tag{3.11}$$

where '*' stands for the time convolution. In frequency domain, Eq.(3.11) has the form

$$Y_n(\omega) = X_i(\omega)\left[1 + 2\sum_{m=1}^{M} H_m(\omega)\cos\left[m\omega(\tau_n + \tau_i)\right]\right] \tag{3.12}$$

Here we also assume the filters are symmetric to the center: $H_m(\omega) = H_{-m}(\omega)$.

We seek the filters with transfer functions $H_m(\omega)$ so that $\sum_{m=1}^{M} H_m(\omega)\cos(m\omega\tau)$ will be independent of ω over the signal bandwidth $\omega_a \leq \omega \leq \omega_b$. There are many ways to parameterize $H_m(\omega)$. The method we use to achieve this goal is as follows: First chose a frequency ω_0 where $\omega_a \leq \omega_0 \leq \omega_b$. At the frequency ω_0 the outputs of sensors are weighted by Gaussian window function to reduce the side lobes, i.e.

$$H_m(\omega_0) = e^{-\frac{m^2}{\alpha_0}} \tag{3.13}$$

At the frequencies around ω_0, we minimize the following cost function at each frequency bin ω over a range of steering directions from $\tau = 0$ to $\tau = \tau_0$.

$$[\alpha(\omega), \beta(\omega), \gamma(\omega)] = \arg\min_{\alpha,\beta,\gamma} \int_0^{\tau_0} \left[\sum_{m=1}^{M} e^{-\frac{m^2}{\alpha_0}}\cos(m\omega_0\tau) - (\gamma e^{-\frac{m^2}{\alpha}} + \beta)\cos(m\omega\tau)\right]^2 d\tau \tag{3.14}$$

Then the transfer function of the filter for the m^{th} sensor is given by

$$H_m(\omega) = \gamma(\omega)e^{-\frac{m^2}{\alpha(\omega)}} + \beta(\omega) \tag{3.15}$$

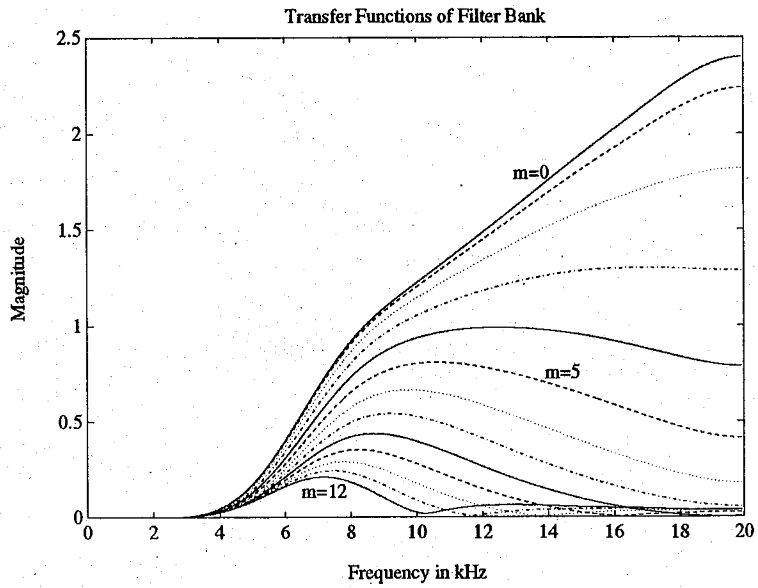


Figure 2: The magnitude of $H_m(\omega)$ for 25 sensors over a frequency band 5kHz to 20kHz. Each curve represents $H_m(\omega)$ of a different sensor (Because of symmetry, only 13 curves show here).

Fig. 2 shows a set of calculated transfer functions of $H_m(\omega)$ over frequency band 5kHz to 20kHz for various sensors. At lower frequencies $H_m(\omega)$ have Gaussian window weights (here $\alpha_0 = 1.5$). As the frequency increases, the $H_m(\omega)$ of the two end sensors go to zero and the central $H_m(\omega)$ increases. The resulting beam power over the given frequency band is plotted in Fig. 3; the directional sensitivity is approximately uniform over the bandwidth as required.

The relative bandwidth ratio is defined as

$$r \triangleq \frac{\omega_b - \omega_a}{\sqrt{\omega_a \omega_b}} \tag{3.16}$$

In the example above, $r = 1.5$. For underwater communication this ratio provides adequate bandwidth for practical purposes, as we will demonstrate. For the "cocktail party" problem, the frequency band usually is from 300Hz to 5kHz which corresponds to $r = 3.84$, twice that in the case above. To obtain a uniform beam over that frequency band, two arrays could be used to form the beams: one operating from 300Hz to 1.2kHz and the other one operating from 1.2kHz to 5kHz. The two set of beams can be superposed together to form the desired beams.

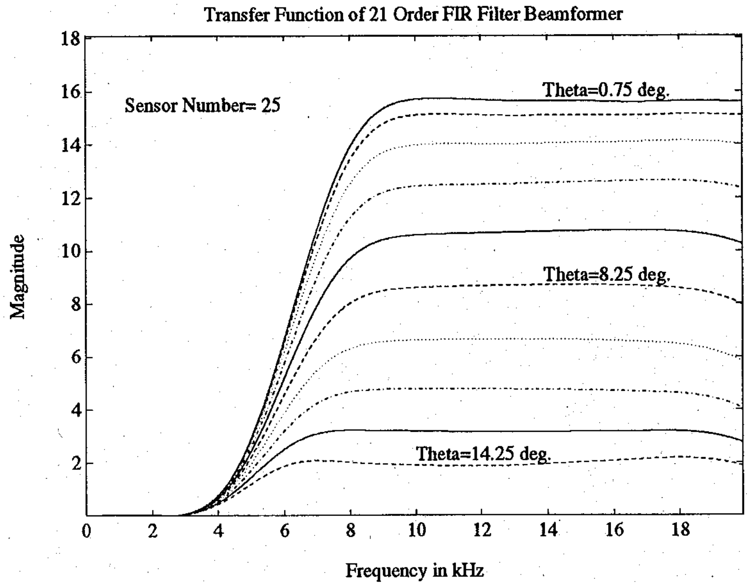


Figure 3: The beam power pattern obtained from a 25-sensor array using the filters shown in Fig.2. The filters are realized by 21-taps FIR filters with symmetric taps weights. The different curves correspond to different directions of arrivals from 0^o to 15^o. The beam power is approximately constant over the bandwidth 8kHz to 20kHz for a given direction of arrival.

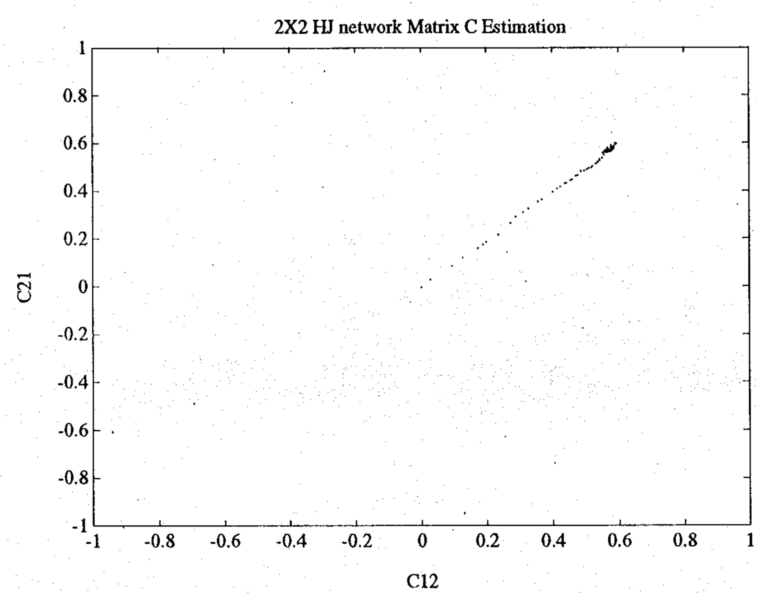


Figure 4: The coefficients c_{12} and c_{21} converge to the correct values during the adaptation. The initial condition is $c_{12} = c_{21} = 0$.

4 Simulations

A 2×2 HJ network with a two-beam array was simulated for a two-path channel. The array had $2M + 1 = 25$ sensors. One formed beam was aimed at 0^o and the other beam was aimed at 10^o. The simulated direct path came from 2^o and the reflected path came from 8^o with a delay of 12.75ms. The signal used in the simulation was a biphase shift keying (BPSK) signal and the carrier frequency was 14kHz. The bit rate was 5kbit/sec. With this carrier frequency and bit rate the main signal bandwidth ranged from $\omega_a = 5$kHz to $\omega_b = 20$kHz. The filter bank consisted of 21-taps finite impulse response (FIR) digital filters to realize $H_m(\omega)$ in Eq.(3.15). Fig. 3 shows the output beam power of such a beamformer. One can see that for a given direction of arrival, which is represented by one of curves, the output beam power is approximately constant over the signal bandwidth from 5kHz to 20kHz. The whole simulation was carried out in the time domain. The carrier synchronization was obtained using a matched filter which convolved the received signal with a time-reversed replica of the signal transmitted.

As the delay time between the different paths increases, unlike a conventional equalization filter which requires additional taps [12], there is no penalty in our algorithm because the signal and its delayed versions are treated as two uncorrelated signals. The information in the signal bandwidth and angle-of-arrival is enough to resolve the different paths. The estimation of time delays, which is difficult in underwater channels, can be performed after the signals from multichannels are detected. The signal and its delays can be considered as uncorrelated when the delay time is longer than a symbol length and there is no correlation between the adjacent bits transmitted. If the adjacent bits are correlated, decorrelation can also be achieved by modulating the transmitted bit train with a pseudo-random bit train which has the same bit rate as the data. When the delay time is smaller than the symbol length, which is 0.2ms in the simulation, the HJ network will not converge in general. For short delay times, the spread spectrum technique can be used to reduce the symbol length. In practice, the main concern usually is for very long delay times rather than very short delay times. A 0.2ms delay presents only 6.8 cm path length difference in air and 30 cm in water, which are very short lengths indeed, and the length can be further decreased as the signal bit rate increases.

Fig. 4 shows the adaptation process of the 2×2 HJ network which used the two beam outputs of the array mentioned above as two inputs. The two quantities c_{12} and c_{21} of the matrix $\mathbf{C}$ in Eq.(3.7) , which are plotted as x axis and y axis respectively, were set to be zero at the beginning of adaptation. We used $f(s) = s^3$ and $g(s) = s$ for c_{12} and c_{21} in the HJ network adaptation [see Eq.(2.4)]. The adaptation rate parameter was $\mu = 0.2$.

Two kinds of white noises were included in the simulation. One was the noise associated with the signal (If this directional noise is strong enough, it can be treated

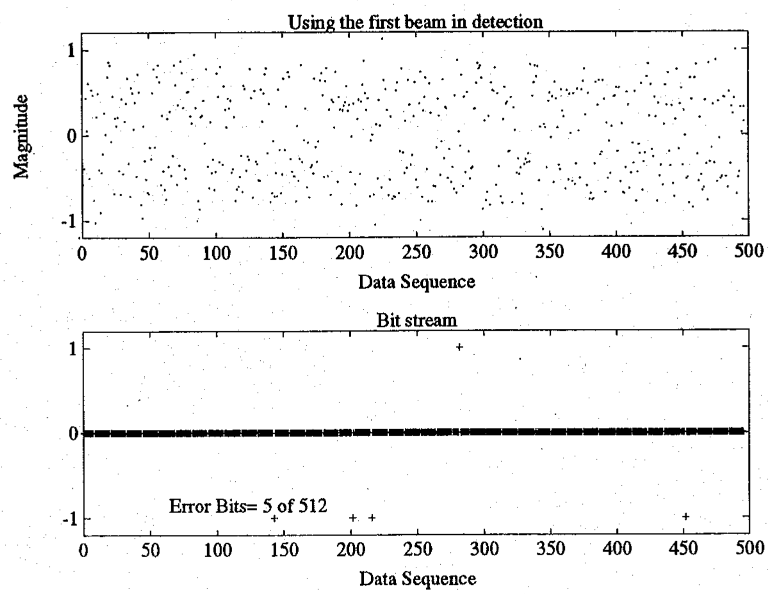


Figure 5: The detection of a BPSK data stream of 512 bits in 0.1 second using only the beam aimed at 0^o without equalization. When a bit was transmitted correctly, there was + located at zero. There were 5 bit errors as shown in the lower panel at +1 and -1.

as an independent signal.). The other is the nondirectional noise that includes the surrounding noise of the sensors and the electronic noise from the receiver circuit. With these noise sources, Eq.(3.6) becomes

$$\mathbf{Y}(t) = \mathbf{A}\left[\mathbf{X}(t) + \mathbf{W}_1(t)\right] + \mathbf{W}_2(t) \tag{4.17}$$

where $\mathbf{W}_1(t)$ and $\mathbf{W}_2(t)$ are independent white Gaussian random vectors with the power spectral densities S_{w1} and S_{w2} respectively. The signal to noise ratios which are defined as $SNR_1 \triangleq 10\log_{10}(S/S_{w1})$ and $SNR_2 \triangleq 10\log_{10}(S/S_{w2})$, are $SNR_1 =$ 7db and $SNR_2 = 27$db in the simulation.

The quantities c_{12} and c_{21} quickly converged to the correct values in simulation. Fig. 5 shows the detection of the bit train when only the output of the beam aimed at 0^o was used. There were several bit errors as shown in Fig. 5. Fig. 6 gives the detection of the same bit train using one of the HJ network outputs. The first 30 data bits were received with about the same reliability as before, but after the HJ network converged, the transmission was error free. Fig. 7 shows the detection when only the output of the beam aimed at 10^o is used. Fig. 8 gives the detection of bit train using the other HJ network output. When μ increases the adaptation speed increases, but the fluctuation of the final values of c_{12} and c_{21} around the correct values also increases. There is a trade off between the adaptation speed and the

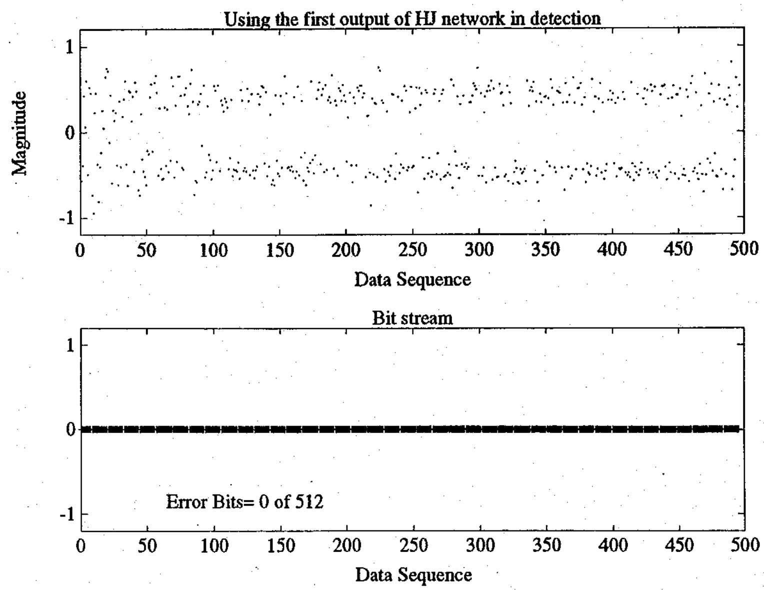


Figure 6: The detection of the same data stream as in Fig.6 of 512 bits in 0.1 second using one of the HJ network outputs. There were no errors. See Fig.8 for the performance of the second output channel.

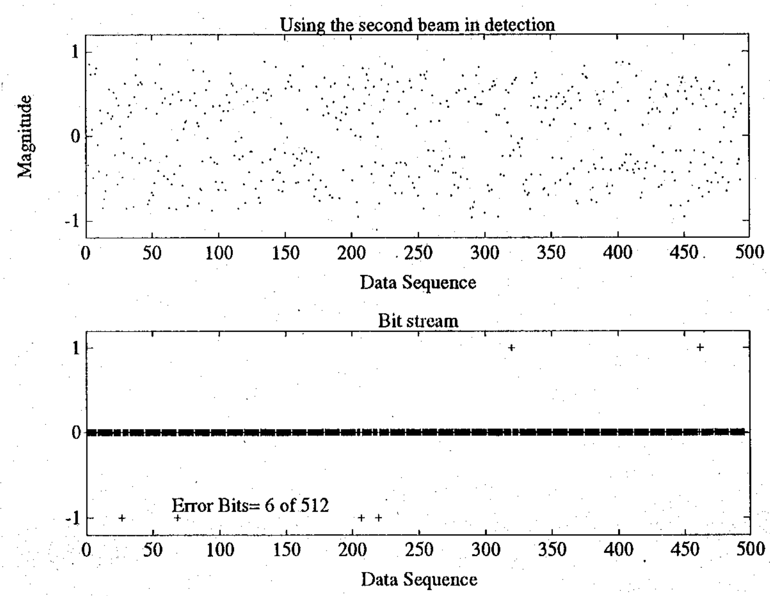


Figure 7: The detection of the data stream of 512 bits in 0.1 second using only the beam aimed at 10°. There were 6 bit errors, shown in the lower panel.

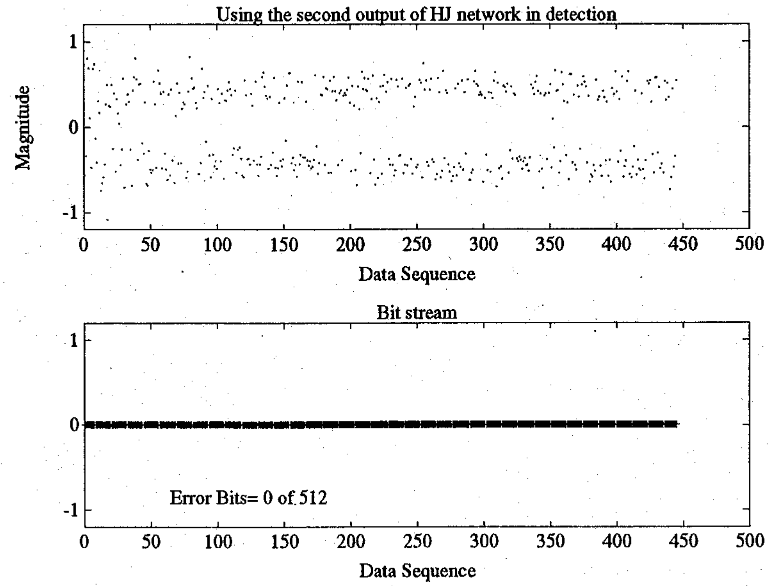


Figure 8: The detection of the data stream of 512 bits in 0.1 second using the second HJ network output. There were no bit errors. The gap at the end of the bit stream is due to the 12.75ms delay in the pathway.

separation accuracy. When the directions of signal arrivals in the underwater acoustic channels change rapidly, the degree of separation will suffer. In our simulation c_{12} and c_{21} adapted from random initial condition to their correct values within 10 to 20 ms. Thus the algrithm is fast enough to handle time-variant fading in underwater acoustic channels.

5 Implementation

Fig. 9 shows a diagram of a complete system for the proposed adaptive signal source separation. The filter bank, beamformer and HJ network can be implemented for real-time processing in several different ways.

5.1 Digital

FIR digital filters can be used to implement $H_m(\omega)$ of Eq.(3.15) using DSP chips. The motivation to use FIR filter is due to its linear phase property which is required by the beamformer at the next stage. Everything will be the same as in the simulation except the system is running in real time using DSP chips instead of a computer. A filter bank of transfer function as $H_m(\omega)$ in Fig. 2 can be realized by

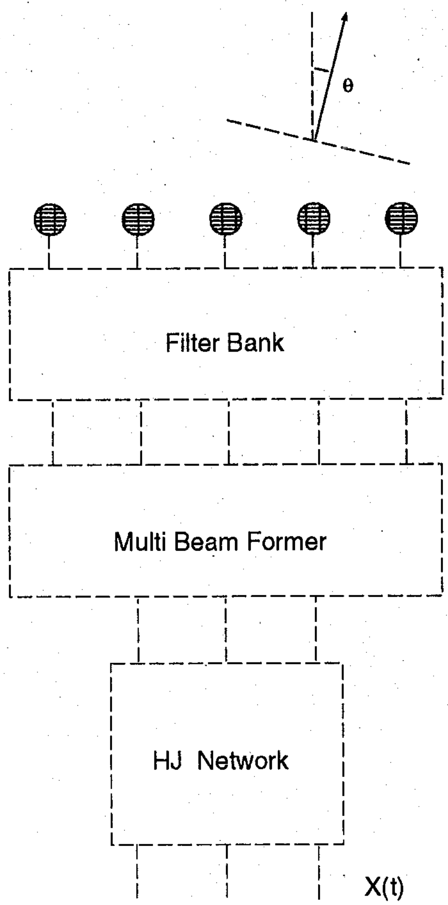


Figure 9: The implementation diagram of the proposed system.

$$y(n\Delta T) = B_0 x(n\Delta T) + \sum_{l=1}^{L} B_l \left[x(n\Delta T - l\Delta T) + x(n\Delta + l\Delta T)\right] \tag{5.18}$$

where B_l is the first $L+1$ coefficients of the inverse discrete Fourier transform of $H_m(\omega)$.

5.2 Analog

The main computational burden of the digital implementation is to solve N unknowns from the N equations in Eq.(2.3) in real time at least at the Nyquist sampling rate. When the number of sources N is large and the sampling frequency is high, the analog implementation has a great advantage over the general purpose DSP chip implementation. The analog implementation can pre-process enormous mount of data from the sensor array, separate the signals among the paths with a moderate accuracy, and leave the detection to the digital processing part.

Analog VLSI HJ network chips already exist ([5], [6]) and it should be possible to implement the entire system including the filter bank, beamformer and HJ network using CMOS analog VLSI technology on a single chip. The filter bank can be implemented by either the charge-coupled device (CCD) split-electrode transversal filter, the bucket brigade device (BBD) transversal filter or a liner phase transconductance capacitor filter. We have successfully implemented an CMOS 3-order Bessel transconductance capacitor filter bank and a BBD beamformer with 15 microphone or more inputs and 6 beams [10]. The next step is to integrate whole system on one chip.

6 Conclusion

We have presented a system for separating independent broadband sound sources and multipath delays using a conventional HJ network and beamforming. Our system relies on a symmetric multibeam-former with frequency equalized sensors to cancel time delays and frequency variations of its inputs to the HJ network.

Our technique can be used in underwater acoustic communication where usually several constantly changing paths exist and cause severe ISI in the data link. Unlike other methods [11] and [12] that have been proposed for reducing ISI in multipath fading channels, which either treats the other multipath signals as noises or results in a high computational complexity for long delays, our system can recover information from each path without using training signals. It is therefore possible to achieve faster signal transmission rates than with other methods. The simulations show that our method can significantly reduce the error rate and may be a practical approach to reliable high-speed acoustic telemetry. For application to the "cocktail party" problem, it could be used to separate various voice and music sources in video conferencing.

The analog implementation can be realized in a single analog VLSI chip, forming a very compact system. The digital implementation is suitable for the 'cocktail party' problem where the sampling rate required for the audio frequency is relatively low. At present our system is restricted to bandpass communication channels, but this limitation could overcome with several arrays.

7 Acknowledgements

We are grateful to Andreas G. Andreou, Weimin Liu and Marc H. Cohen for help and advice with our implementation of the BBD beamformer and transconductance-capacitor filter.

References

[1] J. Hérault and C. Jutten, "Space or time adaptive signal processing by neural network models," *Neural Networks for Computing*, AIP Conference Proceedings 151, pp.207-211, Snowbird, Utah, 1986.

[2] C. Jutten and J. Hérault, "Blind separation of sources, Part I: An adaptive algorithm based on a neuromimetic architecture." *Signal Processing*, 24(1),1-10, 1991.

[3] P. Comon, C. Jutten and J. Hérault, "Blind separation of sources, Part II: Problem statement." *Signal Processing*, 24(1),11-20, 1991.

[4] E. Sorouchyari, "Blind separation of sources, Part III: Stability analysis." *Signal Processing*, 24(1),21-29, 1991.

[5] M. H. Cohen, P. O. Pouliquen and A. G. Andreou, "Silicon implementation of an auto-adaptive network for real time separation of independent signals," *Proc. IEEE Intl. Symp. on Circuits and System*, pp.2971-2974, 1991.

[6] M. H. Cohen, P. O. Pouliquen and A. G. Andreou, "Analog VLSI implementation of an auto-adaptive network for real time separation of independent signals," *Advances in Neural Information Processing System*, V.4, Morgan-Kaufmann, San Mateo, 1992.

[7] J. C. Platt and F. Faggin, "Network for the separation of sources that are superimposed and delayed," *Advances in Neural Information Processing System*, V.4, Morgan-Kaufmann, San Mateo, 1992.

[8] J. Catipovic, "Performance limitations in underwater acoustic telemetry," *IEEE Journal of Oceanic Engineering*, Vol.15, pp. 205-216, July 1990.

[9] S. Li and T. Sejnowski, "Automatic Separation of Mixed Sound Sources by a Beamforming Hérault-Jutten Network," *Processing of the Annual Research Symposium* Institute for Neural Computation, U.C. San Diego, June 1992.

[10] S. Li, T. Sejnowski and A. Andreou, "Implementation of FIR filter and beamformer using analog CMOS VLSI," in preparation.

[11] J. Fisher, K. Bennett S. Reible, J. Cafarella and I. Yao, "A high data rate, underwater data-communication transceiver," *Oceans 92*, v.2 pp. 571-576, 1992

[12] J. Proakis, "Adaptive equalization techniques for acoustic telemetry channels," *IEEE Journal of Oceanic Engineering*, Vol.16, pp. 21-31, January 1991.