

BLIND MULTICHANNEL IDENTIFICATION FOR HIGH-SPEED TDMA [†]

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ABSTRACT

Transmission of multimedia data is the goal of third generation (3G) cellular radio systems necessitating high-speed data links where the multipath delay spread generally encompasses many symbol intervals. The channel impulse response for each antenna in a prototypical high-speed data link is many symbol intervals in duration and generally sparse. In order to reduce the number of parameters that need to be estimated in order to characterize the channel, researchers have recently proposed a number of schemes for direct estimation of the multipath parameters. However, this approach requires a relatively long observation interval in order to accurately estimate the complex gain of each dominant multipath at each antenna. We here propose an alternative approach based on subbanding and the Cross-Relation Method (CRM) of Xu, Liu, Tong, and Kailath. Basically, we apply the CRM in each subband using basis functions derived from the symbol waveform and the bandpass filter to blindly identify the frequency response of each channel in the given subband. A method is presented for proper phasing of each subband channel estimate in order to reconstruct the full channel impulse response associated with each antenna. Simulations are presented demonstrating accurate identification of multiple channels under adverse multipath conditions, moderate SNR, and with a relatively small number of symbols.

1. INTRODUCTION: SUBBAND MODEL

Employing linear modulation with a finite alphabet in a wireless digital communication system, the received signal at the i -th antenna, $i = 1, \dots, N_A$, may be expressed as

$$\mathbf{y}_i(t) = \sum_{m=0}^{N_b-1} b[m]h_i(t - mT_o) + n_i(t), \quad i = 1, \dots, N_A, \quad (1)$$

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where $b[m]$ is the m -th transmitted symbol, N_b is the number of symbols comprising the burst, $1/T_o$ is the symbol rate, and $n_i(t)$ is the (wide-sense stationary) noise process at the i -th antenna after front-end matched filtering with $p_{sr}(-t)$, the suboptimal demodulator that matches the transmit pulse symbol waveform, $p_{sr}(t)$, having a square-root raised cosine spectrum, for example.

The "channel" impulse response associated with the i -th antenna, $h_i(t)$, may be expressed as

$$h_i(t) = p_{rc}(t) * h_{RF,i}(t) \quad (2)$$

where "*" denotes linear convolution and $p_{rc}(t) = p_{sr}(t) * p_{sr}(-t)$ has a raised cosine spectrum and is referred to as the Nyquist symbol waveform with a roll-off factor β . $h_{RF,i}(t)$ is the RF multipath propagation channel that is approximated by a sum of Q dominant specular multipaths as

$$h_{RF,i}(t) = \sum_{k=1}^Q g_{ik} \delta_\alpha(t - \tau_k), \quad (3)$$

where g_{ik} and τ_k are the complex gain and delay, respectively, of the k -th multipath arrival at antenna i , and $\delta_\alpha(t)$ is the Dirac delta function. An upper bound on the delay spread is assumed known and without loss of generality selected to be an integer multiple of $T_o/2$:

$$\max\{\tau_k\} < \tau_{max} = J \frac{T_o}{2} \quad (4)$$

Note that we are dealing with the wideband case where the value of J may be large.

For equalization purposes, it is only necessary to accurately determine the frequency response of the RF multipath channel associated with each antenna over the signal bandwidth. In a narrowband setting, we showed in [1] that this may be achieved by employing a small number of delays equi-spaced across the delay spread, regardless of the actual multipath delay times. In order to utilize the techniques developed in [1] here for the wideband case, consider passing the signal received at each antenna through a bank of M analysis filters, as shown in Figure 1, where each filter passes a different narrow subband. We initially consider a uniform filter bank wherein the passband of each filter is

the same. For reasons to presented later, it is important that adjacent subbands have significant overlap. Although the subbanding will be done via digital filtering after A/D conversion at twice the symbol rate, it will be useful at times for conceptual reasons to also view the subbanding operation in continuous time wherein the impulse response of the ℓ -th bandpass filter is denoted $f_\ell(t)$, $\ell = 1, \dots, M$.

The output of the ℓ -th bandpass filter, $y_{i,\ell}(t)$, at the i -th antenna may be expressed as

$$y_{i,\ell}(t) = \sum_{m=0}^{N_b-1} b[m]h_{i,\ell}(t - mT_o) + n_{i,\ell}(t), \quad i = 1, \dots, N_A, \quad (5)$$

where the "channel" $h_{i,\ell}(t)$ may be expressed as

$$h_{i,\ell}(t) = p_{rc}(t) * h_{RF_i}(t) * f_\ell(t) \quad (6)$$

Following the lead of [1], for the ℓ -th subband, we would like to determine a set of channel tap values $h_{i,\ell}^{(c)}[k]$ equi-spaced across the delay spread ideally satisfying

$$\begin{aligned} h_{i,\ell}(t) &= \sum_{k=1}^Q g_{ik} \delta_a(t - \tau_k) * \{p_{rc}(t) * f_\ell(t)\} \\ &= \sum_{k=0}^{N_h} h_{i,\ell}^{(c)}[k] \delta_a(t - k \frac{T_{max}}{N_h}) * \{p_{rc}(t) * f_\ell(t)\}. \end{aligned} \quad (7)$$

That is, we wish to determine $h_{i,\ell}^{(c)}[k]$, $k = 0, 1, \dots, N_h$, so that the Fourier Transform of $\sum_{k=0}^{N_h} h_{i,\ell}^{(c)}[k] \delta_a(t - k \frac{T_{max}}{N_h})$ is ideally equal to the Fourier Transform of $\sum_{k=1}^Q g_{ik} \delta_a(t - \tau_k)$ over the passband of the analysis filter $f_\ell(t)$. This is sufficient for our equalization purposes here.

Again, the proposed system operates at twice the symbol rate. A model for the $y_{i,\ell}[n] = y_{i,\ell}(nT_o/2)$ under noiseless conditions may be obtained by substituting $nT_o/2$ into eqn. (5) and ignoring noise (in order to develop the Cross-Relation Method).

$$\begin{aligned} y_{i,\ell}[n] &= y_{i,\ell}\left(n \frac{T_o}{2}\right) \\ &= \sum_{m=0}^{N_b-1} b[m]h_{i,\ell}(t - mT_o) \Big|_{t=nT_o/2} \\ &= \sum_{m=0}^{N_b-1} b[m]h_{i,\ell}[n - 2m] \end{aligned} \quad (8)$$

where $h_{i,\ell}[n] = h_{i,\ell}(nT_o/2)$. Consider choosing J and N_h so that the ratio J/N_h is an integer so that, in turn,

$$\frac{T_{max}}{N_h} = I \frac{T_o}{2} \quad \text{where: } I = \frac{J}{N_h} = \text{integer}. \quad (9)$$

Sampling the expression for $h_{i,\ell}(t)$ (with the substitution above) at twice the symbol rate yields the following expression for the impulse response of the channel associated with

the ℓ -th subband and i -th antenna:

$$\begin{aligned} h_{i,\ell}[n] &= h_{i,\ell}\left(n \frac{T_o}{2}\right) \\ &= \sum_{k=0}^{N_h} h_{i,\ell}^{(c)}[k] \delta_a(t - k \frac{T_{max}}{N_h}) * \{p_{rc}(t) * f_\ell(t)\} \Big|_{t=nT_o/2} \\ &= \sum_{k=0}^{N_h} h_{i,\ell}^{(c)}[k] f_\ell(t - kI \frac{T_o}{2}) * p_{rc}(t) \Big|_{t=nT_o/2} \\ &= \sum_{k=0}^{N_h} h_{i,\ell}^{(c)}[k] f_\ell[n - Ik] * p_{rc}[n] \\ &= \left\{ \sum_{k=0}^{N_h} h_{i,\ell}^{(c)}[k] \delta[n - Ik] \right\} * \{p_{rc}[n] * f_\ell[n]\} \quad (10) \\ &= \sum_{k=0}^{N_h} h_{i,\ell}^{(c)}[k] p_\ell[n - Ik] \quad (11) \end{aligned}$$

where $\delta[n]$ is the Kronecker delta function,

$$p_{rc}[n] = p_{rc}\left(n \frac{T_o}{2}\right) \quad \text{and} \quad f_\ell[n] = f_\ell\left(n \frac{T_o}{2}\right) \quad (12)$$

and

$$p_\ell[n] = p_{rc}[n] * f_\ell[n] \quad (13)$$

Both the model for $h_{i,\ell}[n]$ in eqn. (10) and that in eqn. (11) will be useful for later developments.

2. BLIND CHANNEL IDENTIFICATION

In order to apply the Cross-Relation (CR) method, consider decimating $y_{i,\ell}[n]$ in (8) by a factor of 2.

$$y_{i,\ell}^{(0)}[n] = y_{i,\ell}[2n] = b[n] * h_{i,\ell}^{(0)}[n], \quad (14)$$

where

$$h_{i,\ell}^{(0)}[n] = h_{i,\ell}[2n]. \quad (15)$$

For the sake of simplicity, assume without loss of generality that there are only $N_A = 2$ antennas. Considering $i = 1$ and $i = 2$, eqn. (14) indicates that we have the same input, the symbol sequence $b[n]$, into two different FIR filters, $h_{1,\ell}^{(0)}[n]$ and $h_{2,\ell}^{(0)}[n]$. We thus have the cross-relation:

$$y_{1,\ell}^{(0)}[n] * h_{2,\ell}^{(0)}[n] - y_{2,\ell}^{(0)}[n] * h_{1,\ell}^{(0)}[n] = 0 \quad (16)$$

Decimating the model for $h_{i,\ell}[n]$ in eqn.(11) by a factor of 2 and substituting into the equation above yields

$$\begin{aligned} &\left\{ \sum_{k=0}^{N_h} h_{1,\ell}^{(c)}[k] p_\ell[2n - Ik] \right\} * y_{2,\ell}^{(0)}[n] - \\ &\left\{ \sum_{k=0}^{N_h} h_{2,\ell}^{(c)}[k] p_\ell[2n - Ik] \right\} * y_{1,\ell}^{(0)}[n] = 0. \end{aligned} \quad (17)$$

Invoking the distributive property of convolution, the above equation may be rewritten as

$$\sum_{k=0}^{N_h} h_{1,\ell}^{(c)}[k] y_{2,\ell}^{(0,k)}[n] - \sum_{k=0}^{N_h} h_{2,\ell}^{(c)}[k] y_{1,\ell}^{(0,k)}[n] = 0 \quad \forall n, \quad (18)$$

where

$$y_{i,\ell}^{(0,k)}[n] = p_\ell[2n - Ik] * y_{i,\ell}^{(0)}[n], \quad k = 0, 1, \dots, N_h, \quad i = 1, 2. \quad (19)$$

The relation in eqn. (18) may be expressed in matrix form as

$$\begin{bmatrix} \mathbf{Y}_{2,\ell}^{(0)} \\ -\mathbf{Y}_{1,\ell}^{(0)} \end{bmatrix} \mathbf{h}_\ell^{(c)} = \mathbf{0} \quad 1 \leq \ell \leq M, \quad (20)$$

where

$$\mathbf{h}_\ell^{(c)} = \left[h_{1,\ell}^{(c)}[0], \dots, h_{1,\ell}^{(c)}[N_h], h_{2,\ell}^{(c)}[0], \dots, h_{2,\ell}^{(c)}[N_h] \right]^T \quad (21)$$

and

$$\mathbf{Y}_{i,\ell}^{(0)} = \begin{bmatrix} y_{i,\ell}^{(0,0)}[0] & y_{i,\ell}^{(0,1)}[0] & \dots & y_{i,\ell}^{(0,N_h)}[0] \\ y_{i,\ell}^{(0,0)}[1] & y_{i,\ell}^{(0,1)}[1] & \dots & y_{i,\ell}^{(0,N_h)}[1] \\ \vdots & \vdots & \ddots & \vdots \\ y_{i,\ell}^{(0,0)}[N_s] & y_{i,\ell}^{(0,1)}[N_s] & \dots & y_{i,\ell}^{(0,N_h)}[N_s] \end{bmatrix} \quad (22)$$

$i = 1, 2$, where $N_s + 1 \leq N_b$ is the number of symbols used to estimate the channels.

Similar to the development in [1], the performance may be improved by generating additional equations by employing the other polyphase component of the ℓ -th subband output of the two antennas.

$$y_{i,\ell}^{(1)}[n] = y_{i,\ell}[2n + 1] = b[n] * h_{i,\ell}^{(1)}[n]. \quad (23)$$

where

$$h_{i,\ell}^{(1)}[n] = h_{i,\ell}[2n + 1]. \quad (24)$$

Similar to the previous development, we have the cross-relation:

$$y_{1,\ell}^{(1)}[n] * h_{2,\ell}^{(1)}[n] - y_{2,\ell}^{(1)}[n] * h_{1,\ell}^{(1)}[n] = 0 \quad (25)$$

The critical observation for improved performance is that the same coefficients $h_{i,\ell}^{(c)}[k]$ characterize both polyphase components of $h_{i,\ell}[n]$. Decimating the model for $h_{i,\ell}[n]$ in eqn.(11) by a factor of 2 starting at $n = 1$ and substituting into the equation above yields

$$\left\{ \sum_{k=0}^{N_h} h_{1,\ell}^{(c)}[k] p_\ell[2n + 1 - Ik] \right\} * y_{2,\ell}^{(1)}[n] - \left\{ \sum_{k=0}^{N_h} h_{2,\ell}^{(c)}[k] p_\ell[2n + 1 - Ik] \right\} * y_{1,\ell}^{(1)}[n] = 0, \quad (26)$$

which may ultimately be expressed as

$$\sum_{k=0}^{N_h} h_{1,\ell}^{(c)}[k] y_{2,\ell}^{(1,k)}[n] - \sum_{k=0}^{N_h} h_{2,\ell}^{(c)}[k] y_{1,\ell}^{(1,k)}[n] = 0 \quad \forall n, \quad (27)$$

where

$$y_{i,\ell}^{(1,k)}[n] = p_\ell[2n + 1 - Ik] * y_{i,\ell}^{(1)}[n], \quad k = 0, 1, \dots, N_h, \quad i = 1, 2. \quad (28)$$

The relation in eqn. (27) may be expressed in matrix form as

$$\begin{bmatrix} \mathbf{Y}_{2,\ell}^{(1)} \\ -\mathbf{Y}_{1,\ell}^{(1)} \end{bmatrix} \mathbf{h}_\ell^{(c)} = \mathbf{0} \quad 1 \leq \ell \leq M, \quad (29)$$

where

$$\mathbf{Y}_{i,\ell}^{(1)} = \begin{bmatrix} y_{i,\ell}^{(1,0)}[0] & y_{i,\ell}^{(1,1)}[0] & \dots & y_{i,\ell}^{(1,N_h)}[0] \\ y_{i,\ell}^{(1,0)}[1] & y_{i,\ell}^{(1,1)}[1] & \dots & y_{i,\ell}^{(1,N_h)}[1] \\ \vdots & \vdots & \ddots & \vdots \\ y_{i,\ell}^{(1,0)}[N_s] & y_{i,\ell}^{(1,1)}[N_s] & \dots & y_{i,\ell}^{(1,N_h)}[N_s] \end{bmatrix} \quad (30)$$

Appending eq.(29) to eq.(20) yields

$$\underbrace{\begin{bmatrix} \mathbf{Y}_{2,\ell}^{(0)} & -\mathbf{Y}_{1,\ell}^{(0)} \\ \mathbf{Y}_{2,\ell}^{(1)} & -\mathbf{Y}_{1,\ell}^{(1)} \end{bmatrix}}_{\mathbf{X}_\ell(N_s)} \mathbf{h}_\ell^{(c)} = \mathbf{0} \quad 1 \leq \ell \leq M \quad (31)$$

In the practical case where the channels are corrupted by noise, we follow the lead of Xu et al [3] and choose $\mathbf{h}_\ell^{(c)}$ as that right singular vector of $\mathbf{X}_\ell(N_s)$ associated with the smallest singular value.

3. MAXIMIZATION OF IN-BAND ENERGY

Recall the cross-relation in (16) and that the output of each analysis filter $f_\ell[n]$ is maintained at twice the symbol rate. For the narrowband case, we observed in [1] that oversampling leads to improved performance but also trivial solutions as well in which the cross-relation is satisfied by making both $y_{1,\ell}^{(0)}[n] * h_{2,\ell}^{(0)}[n]$ and $y_{2,\ell}^{(0)}[n] * h_{1,\ell}^{(0)}[n]$ individually zero for all n (approximately zero in practice.) This occurs when the matrix $\mathbf{X}_\ell(N_s)$ in eqn. (31) has a number of small singular values and the right singular vector associated with the smallest singular value yields $h_{1,\ell}^{(0)}[n]$ and $h_{2,\ell}^{(0)}[n]$ that are band-reject filters with respective DTFT's approximately equal to zero for $\omega_{1\ell} \leq \omega \leq \omega_{2\ell}$, where $(\omega_{1\ell}, \omega_{2\ell})$ is the passband of the analysis filter $f_\ell[n]$.

We propose a maximum in-band energy post-processing step to avert this phenomenon. Let \mathbf{V}_ℓ be a matrix containing the K right singular vectors of $\mathbf{X}_\ell(N_s)$ associated with the K smallest singular values. The vector $\mathbf{h}_\ell^{(c)} = \mathbf{V}_\ell \beta_\ell$ basically satisfies the cross-relation for any value of the $K \times 1$ vector β_ℓ . The proper linear combination β_ℓ is found by maximizing the in-band energy of $\sum_{k=0}^{N_h} h_{i,\ell}^{(c)}[k] \delta[n - Ik]$ via the solution to

$$\max_{\mathbf{h}_\ell^{(c)}} \frac{\sum_i \text{in-band energy for channel } i \text{ corresponding to subband filter } f_\ell[n]}{\sum_i \text{total energy for channel } i} \quad (32)$$

In mathematical terms,

$$\max_{\mathbf{h}_\ell^{(c)}} \frac{\sum_{i=1}^{N_A} \int_{\omega_{1\ell}}^{\omega_{2\ell}} \left| \sum_{k=0}^{N_h} h_{i,\ell}^{(c)}[k] e^{-jI\omega k} \right|^2 d\omega}{\sum_{i=1}^{N_A} \int_{-\pi}^{\pi} \left| \sum_{k=0}^{N_h} h_{i,\ell}^{(c)}[k] e^{-jI\omega k} \right|^2 d\omega} \quad (33)$$

where

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$$\left\{ \sum_{k=0}^{N_h} h_{1,\ell}^{(c)}[k] p_\ell[2n + 1 - Ik] \right\} * y_{2,\ell}^{(1)}[n] - \left\{ \sum_{k=0}^{N_h} h_{2,\ell}^{(c)}[k] p_\ell[2n + 1 - Ik] \right\} * y_{1,\ell}^{(1)}[n] = 0, \quad (26)$$

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$$\underbrace{\begin{bmatrix} \mathbf{Y}_{2,\ell}^{(0)} & -\mathbf{Y}_{1,\ell}^{(0)} \\ \mathbf{Y}_{2,\ell}^{(1)} & -\mathbf{Y}_{1,\ell}^{(1)} \end{bmatrix}}_{\mathbf{X}_\ell(N_s)} \mathbf{h}_\ell^{(c)} = \mathbf{0} \quad 1 \leq \ell \leq M \quad (31)$$

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$$\max_{\mathbf{h}_\ell^{(c)}} \frac{\sum_i \text{in-band energy for channel } i \text{ corresponding to subband filter } f_\ell[n]}{\sum_i \text{total energy for channel } i} \quad (32)$$

In mathematical terms,

$$\max_{\mathbf{h}_\ell^{(c)}} \frac{\sum_{i=1}^{N_A} \int_{\omega_{1,\ell}}^{\omega_{2,\ell}} \left| \sum_{k=0}^{N_h} h_{i,\ell}^{(c)}[k] e^{-jI\omega k} \right|^2 d\omega}{\sum_{i=1}^{N_A} \int_{-\pi}^{\pi} \left| \sum_{k=0}^{N_h} h_{i,\ell}^{(c)}[k] e^{-jI\omega k} \right|^2 d\omega} \quad (33)$$

Note that it is important for the subbands to have substantial overlap or the result of each of the above two convolutions will be negligible. Defining $\mathbf{z}_{\ell,\ell+1} = [z_{\ell,\ell+1}[0], \dots, z_{\ell,\ell+1}[N_\alpha]]^T$ and $\mathbf{z}_{\ell+1,\ell} = [z_{\ell+1,\ell}[0], \dots, z_{\ell+1,\ell}[N_\alpha]]^T$, the cross-relation for N_α successive values of n may be cumulatively expressed as

$$\mathbf{Z}_{\ell,\ell+1} \boldsymbol{\alpha}_{\ell,\ell+1} = \mathbf{0}, \quad \text{where: } \mathbf{Z}_{\ell,\ell+1} = \begin{bmatrix} \mathbf{z}_{\ell+1,\ell} \\ \vdots \\ \mathbf{z}_{\ell,\ell+1} \end{bmatrix} \quad (44)$$

$\boldsymbol{\alpha}_{\ell,\ell+1}$ is thus computed as the the right singular vector of $\mathbf{Z}_{\ell,\ell+1}$ associated with the smallest singular value.

The reconstructed signal is formed as in Figure 1 as

$$\hat{y}[n] = \sum_{\ell=1}^M \frac{1}{\alpha_\ell} \hat{y}_\ell^{eq} * e_\ell[n], \quad (45)$$

where $e_\ell[n]$, $\ell = 1, \dots, M$, represent a bank of synthesis filters designed so that

$$DTFT \left\{ \sum_{\ell=1}^M f_\ell[n] * e_\ell[n] \right\} = 1, \quad \text{for } |\omega| < \frac{1+\beta}{2} \pi \quad (46)$$

where DTFT denotes the Discrete Time Fourier Transform. The reconstructed signal $\hat{y}[n]$ is an estimate of $y[n] = \sum_{m=0}^{N_b-1} b[m] p_{rc}(n-D-2m)$; D is some integer delay.

6. SIMULATIONS

We here present a very simple proof-of-concept simulation for illustrative purposes. Although the RF bandwidth of the signal simulated, 135 KHz, is 4.5 times that of the current TDMA signal based on the IS-136 standard, it is only about one-tenth of the 1.6 MHz bandwidth proposed for third generation cellular based on a high-speed version of IS-136. The smaller bandwidth was chosen so that it could be well-encompassed by three overlapping subbands of width roughly equal to 60 KHz depicted by the passbands of the three analysis filters plotted in Fig. 2(a). The use of only three subbands averted "cluttered-looking" plots and eased the explanation of results within the space limitations.

Three-ray multipath was simulated. The multipath signals arrived at the angles 0° , 5° , and -5° , with complex amplitudes at the first antenna of 1, $0.9e^{j\pi/4}$, and $0.8e^{j2\pi/3}$, respectively, and corresponding time delays of 0, 5, and 10 micro-seconds. The magnitude of the frequency response of the respective multipath channels associated with each of the two antennas is plotted in Fig. 2(b); the dashed vertical lines demarcate the baseband bandwidth. The $N_A=2$ antennas were separated by three wavelengths. The symbol alphabet was QPSK. The symbol rate was 100 KHz so that $T_o = 10 \mu\text{s}$. The symbol waveform had a square-root raised cosine spectrum with $\beta = 0.35$. The analysis filters, $f_\ell[n]$, $\ell = 1, 2, 3$, were of length 33 at the sampling rate of 200 KHz and were designed via the Parks-McClellan algorithm.

The subbanding based blind multi-channel identification algorithm presented herein was applied to a block of $N_s = 70$ symbols at an SNR of 25 dB. Figure 2(c) plots the magnitude of the DTFT of

$$\hat{p}_\ell[n] = h_{1,\ell}[n] * \hat{g}_{1,\ell}[n] + h_{2,\ell}[n] * \hat{g}_{2,\ell}[n],$$

using the actual values of $h_{1,\ell}[n]$ and $h_{2,\ell}[n]$ and the estimated values of $g_{1,\ell}[n]$ and $g_{2,\ell}[n]$ for the center subband obtained for a typical trial run. The corresponding results for both the left and right subbands are plotted in Figure 2 (d). For each subband, the magnitude of the DTFT of the ideal value of $h_{1,\ell}[n] * g_{1,\ell}[n] + h_{2,\ell}[n] * g_{2,\ell}[n]$, $p_\ell[n]$ equal to $p_{rc}[n] * f_\ell[n]$ is also plotted in each of the figures. The estimated results $\hat{p}_\ell[n]$, $\ell = 1, 2, 3$, were combined via the synthesis filters, $e_\ell[n]$, and estimated coefficients, $\hat{\alpha}_\ell$, as

$$\hat{p}_{rc}[n] = \sum_{\ell=1}^M \frac{1}{\hat{\alpha}_\ell} \hat{p}_\ell[n] * e_\ell[n].$$

The result is plotted in Fig. 2(e) and is observed to match well with the ideal raised cosine spectrum. The 70 reconstructed symbols are plotted in Fig. 2(f) and are observed to cluster about four points commensurate with QPSK.

REFERENCES

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- [2] M. D. Zoltowski, D. Tseng, "A Weighted Energy Concentration Criterion for Improving the Performance of Deterministic Least Squares Blind Channel Identification," to appear *Proc. Asilomar Conf. on Signals, Systems, and Computers*, Pacific Grove, CA, Nov. 1998.
- [3] G. Xu, H. Liu, L. Tong & T. Kailath, "A Least-Squares Approach to Blind Channel Identification," *IEEE Trans. Signal Process.*, pp. 2982-2993, Dec. 1995.

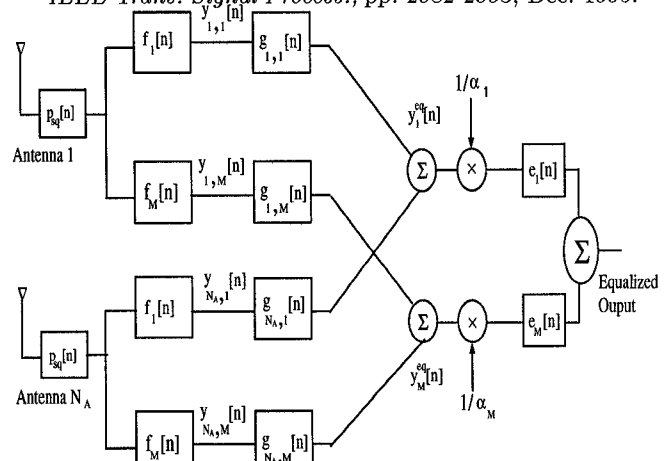
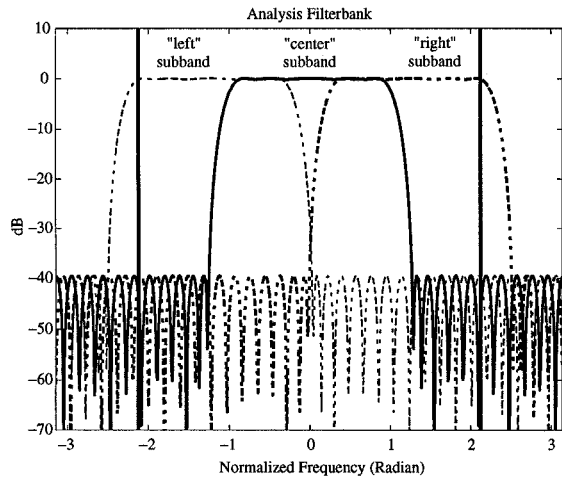
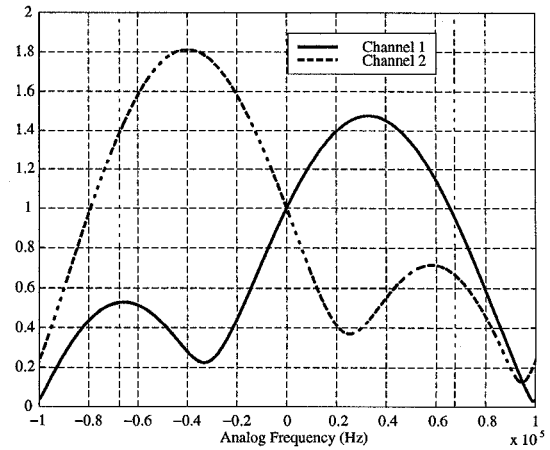


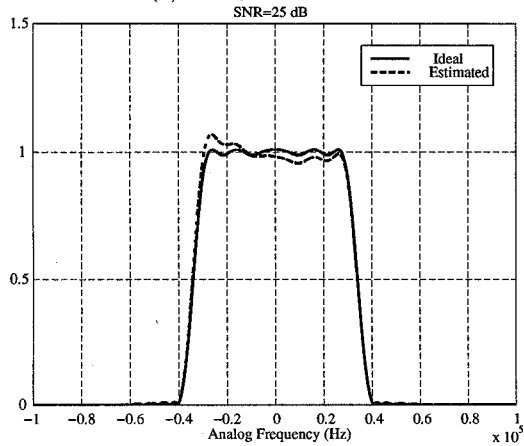
Fig. 1. Block Diagram of Space-Time Equalizer.



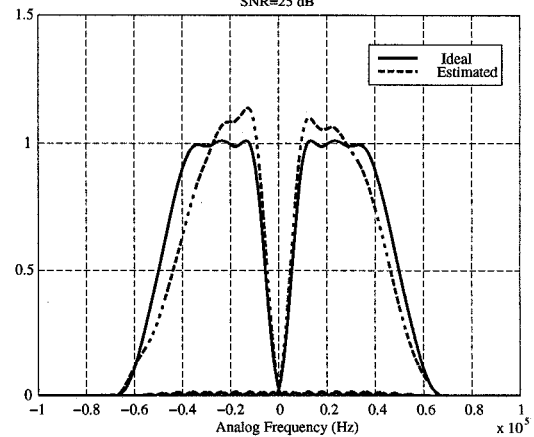
2(a) Analysis Filterbank



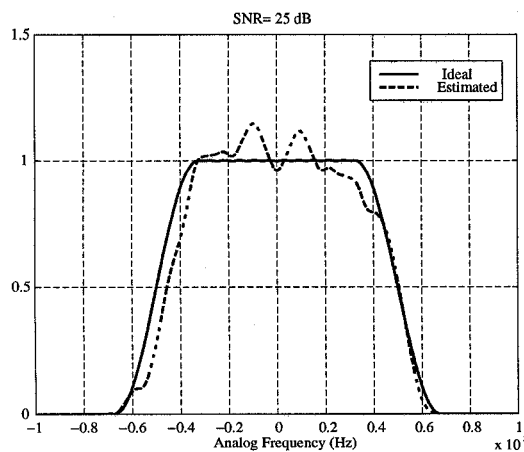
2(b) Frequency Response of Multipath Channels (Magnitude)



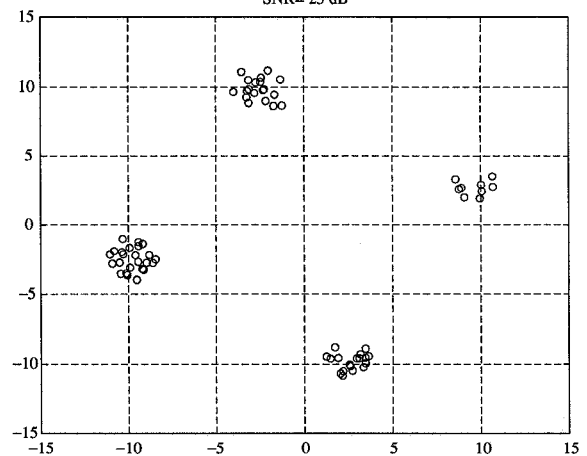
2(c). Equalized "center" subband



2(d). Equalized "left" and "right" subbands



2(e) Actual versus reconstructed raised cosine spectrum.



2(f). Equalized constellation - 70 symbols.