DECORRELATING COMPENSATION SCHEME FOR COEFFICIENT ERRORS OF A FILTER BANK PARALLEL A/D CONVERTER

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ABSTRACT

In this paper, a parallel A/D conversion scheme with a filter bank for Low-IF receivers is presented. The analysis filters of the filter bank divide the frequency components of the received signal, and achieve parallel A/D conversion. Therefore, the required conversion rates and the resolution of the A/D converters can be reduced and the receiver can demodulate wideband signals.

As the analysis filters consist of analog components, their coefficients include errors. These errors cause mutual interference between signals in orthogonal frequencies. In order to remove this interference, a decorrelating compensation scheme is proposed.

1. INTRODUCTION

Recently, software defined radio (SDR) technology has been receiving much attention among researchers working on wireless communications. This is because a number of wireless systems, including IEEE802.11a, IEEE802.11b, and IMT-2000, have been standerdaized, and the demand for multimode / multiband receivers has been increasing.

One of the research issues in SDR technology is its receiver architecture. The ideal structure for the SDR receiver includes a very high speed A/D converter following a LNA in order to convert the received signal in the RF region. However, it is not possible in practical to convert the received analog signal to digital as no such high speed A/D converters are available. Instead, a down conversion technique called direct conversion has been investigated [1]. With direct conversion, the received signal is down converted to a baseband signal directly by a quadrature mixer. The direct conversion receiver inherently has no image response, and thus the fixed-frequency image rejection filters can be eliminated. Furthermore, the anti-alias LPF can be designed with active variable bandwidth filters such as switched-capacitor filters. However, the performance of the direct conversion receiver is limited due to distortion such as the DC offset or 2nd-order distortion.

The alternative choice is the low-IF receiver in which the IF is set to be relatively than in conventional IF receivers. In this receiver it is possible to avoid the DC offset problem as the desired signal is not situated around DC with a high path filter (HPF). To realize the digital HPF for broadband signal reception, the resolution and the speed of A/D converters must be sufficiently high enough. However, improvement of A/D converters has been quite slow as compared to that of digital signal processing devices [2,3].

In order to improve the performance of A/D converters the concept of parallel A/D conversion with a filter bank has been proposed [4]. With this architecture it is possible to increase the total speed of A/D conversion and reduce the required resolution of each A/D converter for each subband. It is also possible to reduce power consumption [5]. Nevertheless, the analysis filters in the filter bank have to be constructed of analog filters such as switched-capacitor filters [6].

In this paper, a digital compensation scheme for the errors of the filter coefficients with a decorrelator is proposed. The proposed scheme first estimates the coefficients of the analysis filters. Based on the estimation, a cross correlation matrix between the subband signals is calculated. With multiplying the inverse of the cross correlation matrix, interference between the subband signals is removed.

2. SYSTEM MODEL

2.1 Filter Bank Parallel A/D Converter

A block diagram of a low-IF receiver with the proposed parallel A/D conversion scheme is shown in Fig. 1. The DFT filter bank is employed. It is assumed that the transmitted frequency components are denoted as $\mathbf{d} = \{d_0, d_1, ..., d_{N-1}\}^T$. Then the received signal is

$$x(n\Delta T) = \sum_{k=0}^{N-1} d_k \exp(j2\pi nk/N) \qquad (1)$$

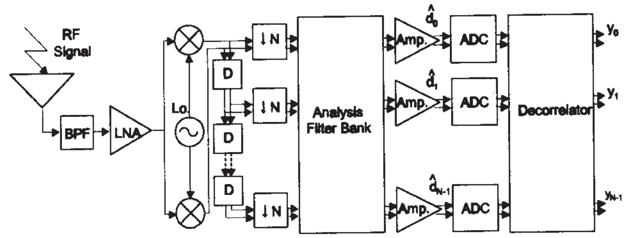

Fig. 1 System Model

where ΔT is the sampling interval and N is the number of input to the analysis filters.

The received signal then goes through the BPF and is converted to the low-IF signal. The low-IF signal is put into the delay line, the decimators, and the analysis filters. If there is no errors in the coefficients of the analysis filters, the output of the k -th filter is given as

$$\begin{aligned}\hat{d}_k(i) &= \frac{1}{N}\sum_{n=0}^{N-1} x(iN\Delta T + n\Delta T)\exp(-j2\pi nk/N) \\ &= \frac{1}{N}\sum_{n=0}^{N-1}\left\{\sum_{k'=0}^{N-1} d_{k'}(i)\exp(j2\pi nk'/N)\right\} \\ &\quad \times \exp(-j2\pi nk/N) \\ &= \frac{1}{N}\sum_{n=0}^{N-1}\sum_{k'=0}^{N-1} d_{k'}(i)\exp(j2\pi n(k'-k)/N) \\ &= \begin{cases} d_k(i) & k'=k \\ 0 & k'\neq k \end{cases} .\end{aligned} \tag{2}$$

Here, it is assumed that the amplification factor is 1 for simplicity. The outputs of the analysis filters are then fed to the N A/D converters. Since the received signal goes through the decimator, the conversion rate of each A/D converter is reduced with the factor of N .

The converted digital signal is put into the decorrelator in order to compensate the errors of the analysis filters. With the decorrelator, the interference between the subband signals is removed.

2.1 Decorrelating Compensation Scheme

In Eq. (2) the errors of the filter coefficients and the noise components are ignored. In the real system, these factors are included in the received signal.

$$\begin{aligned}\hat{d}_k(i) &= \frac{1}{N}\sum_{n=0}^{N-1}(x(iN\Delta T + n\Delta T) \\ &\quad + \eta(iN\Delta T + n\Delta T))c_{nk}\exp(-j2\pi nk/N) \\ &= \frac{1}{N}\sum_{n=0}^{N-1}\left\{\sum_{k'=0}^{N-1} d_{k'}(i)\exp(j2\pi nk'/N)\right. \\ &\quad \left. + \eta(iN\Delta T + n\Delta T))\right\}c_{nk}\exp(-j2\pi nk/N) \\ &= \frac{1}{N}\sum_{n=0}^{N-1}\sum_{k'=0}^{N-1} d_{k'}(i)c_{nk}\exp(j2\pi n(k'-k)/N) \\ &\quad + \frac{1}{N}\sum_{n=0}^{N-1}\eta(iN\Delta T + n\Delta T)c_{nk}\exp(j2\pi nk)/N) ,\end{aligned} \tag{3}$$

where c_{nk} indicates the error of the n -th filter coefficient of the k -th subband, and η is the thermal noise. The matrix form of (3) is

$$\hat{\mathbf{d}} = \hat{\mathbf{R}}\mathbf{d}(i) + \boldsymbol{\mu}(i) \tag{4}$$

where

$$\hat{\mathbf{R}} = \begin{bmatrix} \hat{r}_{0,0} & \hat{r}_{0,1} & \Lambda & \hat{r}_{0,N-1} \\ \hat{r}_{1,0} & \hat{r}_{1,0} & \Lambda & \hat{r}_{1,N-1} \\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} \\ \hat{r}_{N-1,0} & \hat{r}_{N-1,0}1 & \Lambda & \hat{r}_{N-1,N-1} \end{bmatrix} \tag{5}$$

$$\hat{r}_{k,k'} = \frac{1}{N}\sum_{n=0}^{N-1} c_{nk}\exp(j2\pi n(k'-k)/N) , \tag{6}$$

$$\boldsymbol{\mu}(i) = [\mu_0(i), \mu_1(i), \ldots, \mu_{N-1}(i)]^T , \tag{7}$$

$$\mu_k(i) = \frac{1}{N}\sum_{n=0}^{N-1}\eta(iN\Delta T + n\Delta T)c_{nk}\exp(-j2\pi nk/N) . \tag{8}$$

When the received signals go through the A/D converters, quantization noises occur. Suppose that the quantization noises are expressed as $\boldsymbol{\zeta}(i) = \{\zeta_0(i), \zeta_1(i), \ldots, \zeta_{N-1}(i)\}^T$, then

$$\mathbf{Y}(i) = \hat{\mathbf{R}}\mathbf{d}(i) + \boldsymbol{\mu}(i) + \boldsymbol{\zeta}(i) . \tag{9}$$

With the decorrelating compensation scheme the outputs of the A/D converters are multiplied with the inverse matrix of the estimation of $\hat{\mathbf{R}}$

$$\overline{\mathbf{R}}^{-1}\mathbf{Y}(i) = \overline{\mathbf{R}}^{-1}\mathbf{R}\mathbf{d}(i) + \overline{\mathbf{R}}^{-1}(\boldsymbol{\mu}(i) + \boldsymbol{\zeta}(i))$$

The outputs of the decorrelator include the subband signal component $d_k(i)$, thermal noise, and quantization noise.

2.3 Estimation of Filter Coefficients

In order to compensate the errors, the coefficients of the analysis filters have to be estimated. In the estimation mode, the input of the mixer is grounded through a resistance and no signal is input from the antenna. The known waveform is generated and then put in the analysis filters. From Eq.(3)

$$\begin{aligned} y_k(i) &= \sum_{n=0}^{N-1}\hat{h}_k(n)(x(iN\Delta T + n\Delta T) \\ &\quad + \eta(iN\Delta T + n\Delta T)) + \zeta_k(i) \end{aligned} \tag{11}$$

where

$$\hat{h}_k(n) = \frac{c_{nk}}{N}\exp(-j2\pi nk/N) . \tag{12}$$

Calculate $y_k(i)$ for $i = i, \ldots, i + N - 1$, then,

$$\mathbf{Y}_k(i) = (\mathbf{X}(i) + \boldsymbol{\eta}(i))\hat{\mathbf{h}}_k + \boldsymbol{\zeta}_k(i) \tag{13}$$

where

$$\mathbf{X}(i) = \begin{bmatrix} x(iN\Delta T) & x(iN\Delta T + \Delta T) & \cdots & x(iN\Delta T + (N-1)\Delta T) \\ x((i+1)N\Delta T) & x((i+1)N\Delta T + \Delta T) & \cdots & x((i+1)N\Delta T + (N-1)\Delta T \\ \vdots & \vdots & \ddots & \vdots \\ x((i+N-1)N\Delta T) & x((i+N-1)N\Delta T + \Delta T) & \cdots & x((i+N-1)N\Delta T + (N-1)\Delta T) \end{bmatrix}, \tag{14}$$

$$\boldsymbol{\eta}(i) = \begin{bmatrix} \eta(iN\Delta T) & \eta(iN\Delta T + \Delta T) & \cdots & \eta(iN\Delta T + (N-1)\Delta T) \\ \eta((i+1)N\Delta T) & \eta((i+1)N\Delta T + \Delta T) & \cdots & \eta((i+1)N\Delta T + (N-1)\Delta T \\ \vdots & \vdots & \ddots & \vdots \\ \eta((i+N-1)N\Delta T) & \eta((i+N-1)N\Delta T + \Delta T) & \cdots & \eta((i+N-1)N\Delta T + (N-1)\Delta T) \end{bmatrix}, \tag{15}$$

$$\hat{\mathbf{h}}_k = \begin{bmatrix} \hat{h}_k(0) \\ \hat{h}_k(1) \\ \vdots \\ \hat{h}_k(N-1) \end{bmatrix}, \tag{16}$$

$$\boldsymbol{\zeta}_k(i) = \begin{bmatrix} \zeta_k(i) \\ \zeta_k(i+1) \\ \vdots \\ \zeta_k(i+N-1) \end{bmatrix}, \tag{17}$$

$$\mathbf{Y}_k(i) = \begin{bmatrix} y_k(i) \\ y_k(i+1) \\ \vdots \\ y_k(i+N-1) \end{bmatrix}. \tag{18}$$

Assuming that $\mathbf{X}(i)$ are the known signal, the estimated filter coefficients, $\hat{\mathbf{h}}_k$, are then given by the following equation.

$$\begin{aligned} \mathbf{X}^{-1}(i)\mathbf{Y}_k(i) &= \hat{\mathbf{h}}_k + \mathbf{X}^{-1}(i)\boldsymbol{\eta}(i)\hat{\mathbf{h}}_k + \mathbf{X}^{-1}(i)\boldsymbol{\zeta}_k(i) \\ &= \hat{\mathbf{h}}_k + \mathbf{X}^{-1}(i)\boldsymbol{\mu}_k(i) + \mathbf{X}^{-1}(i)\boldsymbol{\zeta}_k(i) \end{aligned} \tag{19}$$

where

$$\boldsymbol{\mu}_k(i) = [\mu_k(i), \mu_k(i+1), \ldots, \mu_k(i+N-1)]^T \tag{20}$$

Due to the thermal noise and quantization noise, there are errors in the estimated filter coefficients.

3. NUMERICAL ANALYSIS

3.1 BER without Compensation

In this section the BER without compensation is derived. Assume that d_k is the modulated signal with BPSK. From Eq. (9) the desired signal and the interference from different subbands is given as

$$s_k(\mathbf{d}(i)) = \mathrm{Re}[\hat{\mathbf{R}}\mathbf{d}(i)]_k \tag{21}$$

where $\mathrm{Re}\{a\}$ is the real part of a, and $[\mathbf{A}]_k$ indicates the k-th element of a $N \times 1$ matrix $\mathbf{A}$. From Eq. (8) covariance of the thermal noise components is

$$\begin{aligned} &E\left[\mu_k(i)\mu_{k'}^*(i)\right] \\ &= E\left[\left(\frac{1}{N}\sum_{n=0}^{N-1}\eta(iN\Delta T + n\Delta T)c_{nk}\exp(-j2\pi nk/N)\right)\right. \\ &\times\left.\left(\frac{1}{N}\sum_{n=0}^{N-1}\eta^*(iN\Delta T + n\Delta T)c_{nk'}^*\exp(j2\pi nk'/N)\right)\right] \\ &= \sigma^2 p_{k,k'} \end{aligned} \tag{22}$$

where σ^2 is the variance of the thermal noise. Therefore,

$$E\left[\boldsymbol{\mu}(i)\boldsymbol{\mu}^H(i)\right] = \sigma^2\mathbf{P} \tag{23}$$

where

$$\mathbf{P} = \begin{bmatrix} p_{0,0} & p_{0,1} & \cdots & p_{0,N-1} \\ p_{1,0} & p_{1,1} & \cdots & p_{1,N-1} \\ \vdots & \vdots & \ddots & \vdots \\ p_{N-1,0} & p_{N-1,1} & \cdots & p_{N-1,N-1} \end{bmatrix}, \tag{24}$$

$$p_{k,k'} = \sum_{n=0}^{N-1}\frac{c_{nk}c_{nk'}^*}{N}\exp(j2\pi n(k'-k)/N). \tag{25}$$

It is also assumed that the absolute value of the maximum input to the k-th A/D converter is Am_k, the resolution of the A/D converters is B_q bits, and the first bit of the A/D converter is assigned for the sign of the input signal. The quantization noise, ζ_k, is then uniformly distributed between $\left[-(Am_k/2^{B_q}),(Am_k/2^{B_q})\right]$.

The conditional BER is given by the following equation.

$$Pc_k(\mathbf{d}(i)) = \frac{1}{2(Am_k/2^{B_q})} \times \int_{-Am_k/2^{B_q}}^{Am_k/2^{B_q}} \frac{1}{2}\mathrm{erfc}\left(\frac{|s_k(\mathbf{d}(i))|+\zeta_k}{\sqrt{\sigma^2[\mathbf{P}]_{k,k}}}\right)d\zeta_k \quad (26)$$

where $[\mathbf{A}]_{k,k}$ is the (k,k) element of the $N\times N$ matrix $\mathbf{A}$. With the condition, $t=(|s_k(\mathbf{d})|+\zeta_k)/\sqrt{\sigma^2[\mathbf{P}]_{k,k}}$, the error function can be rewritten as,

$$\begin{aligned} Pc_k(\mathbf{d}(i)) &= \frac{\sqrt{\sigma^2[\mathbf{P}]_{k,k}}}{2(Am_k/2^{B_q})}\int_{r_{lk}(\mathbf{d}(i))}^{r_{hk}(\mathbf{d}(i))}\frac{1}{2}\mathrm{erfc}(t)dt \\ &= \frac{\sqrt{\sigma^2[\mathbf{P}]_{k,k}}}{2(Am_k/2^{B_q})}\int_{r_{lk}(\mathbf{d}(i))}^{r_{hk}(\mathbf{d}(i))}\left(\frac{1}{2} - \frac{1}{\sqrt{\pi}}\sum_{i=i}^{\infty}(-1)^{i+1}\frac{t^{2i-1}}{(2i-1)(i-1)!}\right)dt \\ &= \frac{\sqrt{\sigma^2[\mathbf{P}]_{k,k}}}{2(Am_k/2^{B_q})}\left(\frac{1}{2}(r_{hk}(\mathbf{d}(i))-r_{lk}(\mathbf{d}(i))) - \frac{1}{\sqrt{\pi}}\sum_{i=i}^{\infty}(-1)^{i+1}\frac{1}{(2i-1)(i-1)!} \times\frac{(r_{hk}(\mathbf{d}(i))^{2i}-(r_{lk}(\mathbf{d}(i))^{2i}}{2i}\right) \end{aligned} \quad (27)$$

where

$$r_{lk}(\mathbf{d}(i)) = (|(s_k(\mathbf{d}(i))|-Am_k/2^{B_q})/\sqrt{\sigma^2[\mathbf{P}]_{k,k}}, \quad (28)$$

$$r_{hk}(\mathbf{d}(i)) = (|(s_k(\mathbf{d}(i))|+Am_k/2^{B_q})/\sqrt{\sigma^2[\mathbf{P}]_{k,k}}. \quad (29)$$

Then the BER without compensation is given as

$$Pc_k = \underset{\mathbf{d}(i)}{E}\left[Pc_k(\mathbf{d}(i))\right]. \quad (30)$$

3.2 Accuracy of the Filter Coefficient Estimation

From Eq. (19), the covariance of the noise elements is calculated. As for the thermal noise, the covariance is given as

$$E\left[(\mathbf{X}^{-1}\boldsymbol{\mu}_k(i))(\mathbf{X}^{-1}\boldsymbol{\mu}_k(i))^H\right] = \sigma^2\mathbf{X}^{-1}\mathbf{P}'(\mathbf{X}^{-1})^H, \quad (31)$$

where

$$\mathbf{P}' = \begin{bmatrix} p_{0,0} & 0 & \Lambda & p_{0,N-1} & \\ 0 & p_{1,1} & \Lambda & 0 & \\ \mathrm{M} & \mathrm{M} & \mathrm{O} & \mathrm{M} & 0 \\ 0 & 0 & \Lambda & p_{N-1,N-1} & \end{bmatrix}, \quad (32)$$

$$p_{k,k'} = \sum_{n=0}^{N-1}\frac{c_{nk}c_{nk'}^{*}}{N}\exp(j2\pi n(k'-k)/N). \quad (33)$$

Assuming that the variance of the quantization noise in each subband is equal to λ^2, the covariance of the quantization noise is

$$E\left[(\mathbf{X}^{-1}\zeta(i))(\mathbf{X}^{-1}\zeta(i))^H\right] = \lambda_e^2\mathbf{X}^{-1}(\mathbf{X}^{-1})^H \quad (34)$$

where λ_e is given as

$$\lambda_e^2 = 2\frac{Am_k^2}{12\times 2^{2B_q}}. \quad (35)$$

Thus, the variance of the noise to the estimated value of n-th filter coefficient is given by

$$\delta_n^2 = \sigma^2\left[\mathbf{X}^{-1}\mathbf{P}'(\mathbf{X}^{-1})^H\right]_{k,k} + \lambda^2\left[\mathbf{X}^{-1}(\mathbf{X}^{-1})^H\right]_{k,k} \quad (36)$$

where $[\mathbf{A}]_{k,k}$ indicates (k,k) element of a $N\times N$ matrix $\mathbf{A}$.

3.3 BER with the Proposed Compensation Scheme

In this section, the BER with the proposed compensation scheme is derived. From Eq. (10), the desired signal and the interference from different subband is given as

$$\hat{d}_k(\mathbf{d}(i)) = \mathrm{Re}[\overline{\mathbf{R}}^{-1}\hat{\mathbf{R}}\mathbf{d}(i)]_k. \quad (37)$$

From Eq. (10), the covariance of the noise elements is calculated. As for the thermal noise, the covariance is given as

$$E\left[(\overline{\mathbf{R}}^{-1}\boldsymbol{\mu}_k(i))(\overline{\mathbf{R}}^{-1}\boldsymbol{\mu}_k(i))^H\right] = \sigma^2\overline{\mathbf{R}}^{-1}\mathbf{P}(\overline{\mathbf{R}}^{-1})^H. \quad (38)$$

Assuming that the variance of the quantization noise in each subband is equal to λ^2, the covariance of the quantization noise is

$$E\left[(\overline{\mathbf{R}}^{-1}\zeta(i))(\overline{\mathbf{R}}^{-1}\zeta(i))^H\right] = \lambda^2\overline{\mathbf{R}}^{-1}(\overline{\mathbf{R}}^{-1})^H \quad (39)$$

where λ is given as

$$\lambda^2 = 2\frac{Am_k^2}{12\times 2^{2B_q}} \quad . \quad (40)$$

Though the quantization noise shows uniform distribution, it can be modeled as Gaussian noise after it is multiplied with the coefficients of the decorrelator and summed [7]. Therefore, the SNR in the k-th subband is given by

$$\mathrm{SNR}_k = \frac{(\hat{d}_k(\mathbf{d}(i)))^2}{\sigma^2\left[\overline{\mathbf{R}}^{-1}\mathbf{P}(\overline{\mathbf{R}}^{-1})^H\right]_{k,k} + \lambda^2\left[\overline{\mathbf{R}}^{-1}(\overline{\mathbf{R}}^{-1})^H\right]_{k,k}} \quad . \quad (41)$$

Suppose that the d_k is the modulated signal with BPSK, the BER is then

$$Pp_k = \frac{1}{2}\mathrm{erfc}(\sqrt{\mathrm{SNR}_k}) \quad . \quad (42)$$

If the received signal on the k'-th subband is significantly larger than the signals received on the other subbands, the quantization noise added on that subband does not follow Gaussian distribution. In this case, the BER can be rewritten as

$$Pp_{kk'} = \frac{1}{2(Am_k/2^{B_q})} \times \int_{-Am_{k'}/2^{B_q}}^{Am_{k'}/2^{B_q}} \frac{1}{2}\mathrm{erfc}\left(\sqrt{\mathrm{SNR}_k(\zeta_{k'})}\right)d\zeta_{k'} \quad , \quad (43)$$

where $Am_{k'}$ is the maximum value of the received signal on the k'-th subband, and

$$\mathrm{SNR}_k(\zeta_{k'}) = \frac{(\hat{d}_k(\mathbf{d}(i)) + \mathrm{Re}\{\left[\overline{\mathbf{R}}^{-1}\right]_{k,k'}\zeta_{k'}\})^2}{\left(\sigma^2\left[\overline{\mathbf{R}}^{-1}\mathbf{P}(\overline{\mathbf{R}}^{-1})^H\right]_{k,k} + \lambda^2\left(\left[\overline{\mathbf{R}}^{-1}(\overline{\mathbf{R}}^{-1})^H\right]_{k,k} - |\left[\overline{\mathbf{R}}^{-1}\right]_{k,k'}|^2\right)\right)} = \frac{(\hat{d}_k(\mathbf{d}(i)) + \mathrm{Re}\{\left[\overline{\mathbf{R}}^{-1}\right]_{k,k'}\zeta_{k'}\})^2}{\varepsilon^2} \quad , \quad (44)$$

$$\varepsilon = \sqrt{\left(\sigma^2\left[\overline{\mathbf{R}}^{-1}\mathbf{P}(\overline{\mathbf{R}}^{-1})^H\right]_{k,k} + \lambda^2\left(\left[\overline{\mathbf{R}}^{-1}(\overline{\mathbf{R}}^{-1})^H\right]_{k,k} - |\left[\overline{\mathbf{R}}^{-1}\right]_{k,k'}|^2\right)\right)} \quad . \quad (45)$$

By transforming the variable as

$$t = (\hat{d}_k(\mathbf{d}(i)) + \mathrm{Re}\{\left[\overline{\mathbf{R}}^{-1}\right]_{k,k'}\zeta_{k'}\})^2/\varepsilon , \quad (46)$$

Eq. (43) can be rewritten as

$$Pp_{kk'}(\mathbf{d}(i)) = \frac{\varepsilon}{2(|\mathrm{Re}\{\left[\overline{\mathbf{R}}^{-1}\right]_{k,k'}\}|Am_{k'}/2^{B_q})} \times \int_{r_{lk'}(\mathbf{d}(i))}^{r_{hk'}(\mathbf{d}(i))}\frac{1}{2}\mathrm{erfc}(t)dt$$

$$= \frac{\varepsilon}{2(|\mathrm{Re}\{\left[\overline{\mathbf{R}}^{-1}\right]_{k,k'}\}|Am_{k'}/2^{B_q})} \times \int_{r_{lk'}(\mathbf{d}(i))}^{r_{hk'}(\mathbf{d}(i))}\left(\frac{1}{2} - \frac{1}{\sqrt{\pi}}\sum_{i=i}^{\infty}(-1)^{i+1}\frac{t^{2i-1}}{(2i-1)(i-\right.$$

$$= \frac{\varepsilon}{2(|\mathrm{Re}\{\left[\overline{\mathbf{R}}^{-1}\right]_{k,k'}\}|Am_k/2^{B_q})}\left(\frac{1}{2}(r_{hk'}(\mathbf{d}(i)) - r_{lk'}(\right.$$

$$-\frac{1}{\sqrt{\pi}}\sum_{i=i}^{\infty}(-1)^{i+1}\frac{t^{2i-1}}{(2i-1)(i-1)!}\frac{(r_{hk'}(\mathbf{d}(i))^{2i} - (r_{lk'}(}{2i} \quad (47)$$

where

$$r_{lk'}(\mathbf{d}(i)) = (\hat{d}_k(\mathbf{d}(i)) - |\mathrm{Re}\{\left[\overline{\mathbf{R}}^{-1}\right]_{k,k'}\}|Am_k/2^{B_q})/\varepsilon \quad , \quad (48)$$

$$r_{hk}'(\mathbf{d}(i)) = (\hat{d}_k(\mathbf{d}(i)) + |\mathrm{Re}\{\left[\overline{\mathbf{R}}^{-1}\right]_{k,k'}\}|Am_k/2^{B_q})/\varepsilon \quad . \quad (49)$$

4. SIMULATION RESULTS

4.1 Simulation Conditions

The following results assume that the subband signal is modulated with BPSK, transmitted through an AWGN channel, and down converted to the IF frequency of $3\pi/N$ [rad/sample]. Accuracy of coefficient estimation is given as $1/\delta_n^2$. The maximum input value of the A/D converters is set to $Am_k = 2\hat{d}_k$ and $Am_k = \mathrm{Re}\{2s_k(\mathbf{d})\}$ with and without compensation cases. The interference signal is received with the IF frequency of $-3\pi/N$ [rad/sample] and the power of the interference is given by SIR. The maximum amplitude of the interference is set to $Am_{k'}/2$. The BER is calculated 10000 times with adding the coefficient errors which are uniformly distributed over $[-0.01, 0.01]$.

4.2 Simulation Results

The BER versus Eb/No is shown in Fig. 2. The resolution of the A/D converters is 8 [bit]. The BER without compensation shows the error floor. This is because the interference occurs due to the errors of the analysis filters. On the other hand, with the proposed compensation scheme, the BER is close to the theoretical performance of BPSK if estimation accuracy is more than 30 [dB]. This means that the proposed scheme is valid with appropriate coefficient estimation schemes.

5. CONCLUSIONS

In this paper, a parallel A/D conversion scheme with a filter bank for the low-IF receivers has been investigated. The analysis filters of the filter bank divide the frequency components of the received signal, andachieve parallel A/D conversion. Therefore, the required conversion rates and the resolution of the A/D converters can be reduced and the receiver can demodulate wideband signals.

As the analysis filters consist of analog components, their coefficients include errors. These errors cause mutual interference between signals in orthogonal frequencies. In order to remove this interference, the decorrelating compensation scheme has been proposed. With the proposed compensation scheme, it has been shown that the interference can be removed if estimation accuracy is more than 30 [dB] when SIR is less than 60 [dB].

9. ACKNOWLEDGMENT

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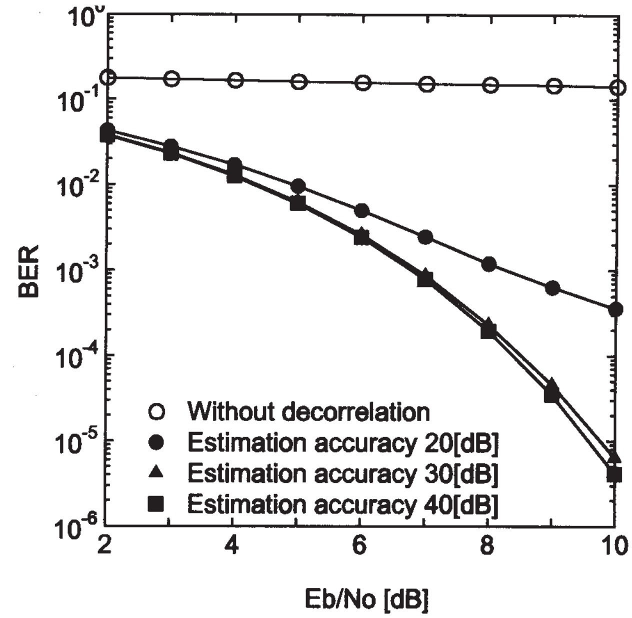

Fig.2 BER vs. E_b/N_0, N=8, Resolution of A/D Convereter=8 [bits], SIR=60 [dB]

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