PILOT ASSISTED COHERENT DS-CDMA REVERSE-LINK COMMUNICATIONS WITH OPTIMAL ROBUST CHANNEL ESTIMATION

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ABSTRACT

Optimal pilot assisted estimation of communication channels is considered for coherent cellular and PCS CDMA reverse link communications. Both pilot symbol and pilot channel based schemes are described and the optimal estimators for these two schemes are analyzed. Relative mean square estimation error (RMSEE) and optimal power allocation between data and pilot signals are derived based on the analysis. Finally, simulation results are given to show the reverse link performance can be significantly improved by using the pilot assisted coherent communication instead of non-coherent schemes for CDMA reverse link.

1. INTRODUCTION

Direct-sequence code division multiple access (DS-CDMA) technology, a form of spread spectrum communication, has been adopted as an interim standard (IS-95) for digital mobile cellular communications and personal communication systems (PCS) in the United States. It has also been implemented in a number of countries around the world. As is well known, CDMA communication systems are interference-limited. In order to increase the system capacity, i.e., to allow more users in the system, the receivers in subscriber units and base-stations should be able to operate at a lower signal to interference ratio (SIR), which is usually expressed in terms of the energy per information bit per interference density, i.e., E_b/I_0.

For the DS-CDMA cellular system specified by IS-95, the forward link, i.e, transmission from a basestation to subscribers, has a common pilot channel, which transmits an unmodulated signal, shared by all the users associated with the same sector. Any such user uses the pilot channel to establish the phase reference for performing coherent demodulation of the traffic signal. On the other hand the reverse link, i.e. transmission from a subscriber to a base station, uses a Walsh orthogonal code set for channel coding. For such signaling and using non-coherent demodulation, it is shown in [1] that to achieve a 1% frame error rate at 100 km/h an E_b/I_0 of 5.8 dB is required under flat fading with dual antenna diversity at 900 MHz. As was shown recently [2,3], it is possible to perform coherent detection of the IS-95 reverse link signal to obtain a performance improvement of 1.0 to 1.5 dB. However, the gain is achieved with increased complexity.

It was perceived by some researchers previously that pilot aided coherent communication is not appropriate for reverse link communication. It was assumed that it would be necessary to have a strong pilot signal to obtain a good channel estimate for coherent demodulation. Based on this argument, since, unlike the forward link, each subscriber must have its own pilot signal, which does not carry information, the power used by the pilot would offset the gain obtained by using coherent demodulation. Hence, the performance would be worse than non-coherent communication.

The application of pilot assisted coherent communication to CDMA reverse-link was first proposed in [4] in the form of a pilot symbol assisted scheme. It was shown there that the net gain due to coherent demodulation over non-coherent detection of the orthogonal Walsh code can more than compensate the loss due to the use of the pilot signal – the net gain can be over 2.5 dB. The criteria for parameter selection were also analyzed in [4]. A similar scheme for wideband CDMA communication was investigated in [5]. The equivalence between the pilot symbol[4,5] and pilot channel assisted [6] schemes were shown in [7]. Recently, the advantage of the coherent reverse link has been recognized by researchers and we expect a future CDMA system to feature a coherent reverse link.

In a pilot assisted coherent communication system, an estimate of the communication channel is obtained based on the pilot signal sent together with the information bearing data signal. Thus, its performance directly depends on the accuracy of the channel estimate. The accuracy of the channel estimate is especially important for CDMA reverse-link communication to reduced the overhead caused by the pilot signal. In this paper, various aspects of the pilot assisted channel estimation in such a CDMA communication system are investigated.

2. TWO PILOT ASSISTED COHERENT REVERSE LINK SCHEMES FOR CDMA WIRELESS COMMUNICATION

In a wireless multiple access communication system with a pilot assisted coherent reverse link, a mobile transmits a signal known to the base station, called a pilot signal, together with information bearing data (coded bits). The pilot signal known to the base-station is used by the base receiver to generate an estimate of the phase and amplitude of the communication channel. In this section, we described two such schemes.

Using the first scheme, known symbols are spread and inserted into a spread data sequence. Thus, it is called the *pilot, or reference, symbol* assisted method. In the second scheme, the known pilot and the unknown data symbols are spread with two spreading sequences, which have the same chip rate but are uncorrelated or orthogonal to each other. The spread signals of the pilot and data are then added together for transmission. The second method is called the *pilot channel* assisted method. A block diagram showing the generation of the signal with pilot symbols and its data format is shown in Figure 1. The generation of the signal in the pilot channel assisted method is depicted in Figure 2.

Optimal system performance can be achieved by properly setting the ratio of the data signal power to the pilot signal power. For the pilot symbol assisted method, we assume each unmodulated pilot symbol is spread to N_{cp} chips. Each of the data symbol is spread to N_{cd} chips. The pilot and data chip magnitudes, denoted as A_p and A_d, are both equal to 1. By inserting one pilot symbol for every M_d data symbols, the data to pilot power ratio, M, is equal to $M_d N_{cd}/N_{cp}$. The frequency of the pilot symbol, f_p, is equal to $f_c/[N_{cp}\times(M+1)]$, where f_c is the chip rate.

For the pilot channel assisted scheme, we set the magnitude of the pilot $A_p = 1/\sqrt{P+1}$ and the magnitude of the data sequence $A_d = \sqrt{P}/\sqrt{P+1}$. Since, the total chip RMS magnitude is 1, the average chip power is the same as that in the pilot symbol case. The power ratio of the data signal to the pilot signal is $P = A_d^2/A_p^2$. If P = M, the data to pilot power ratios of the two schemes are equal.

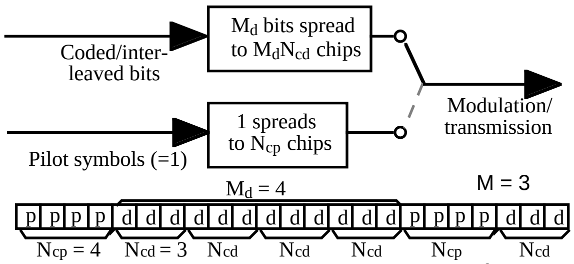

Figure 1 Pilot symbol assisted scheme and data format

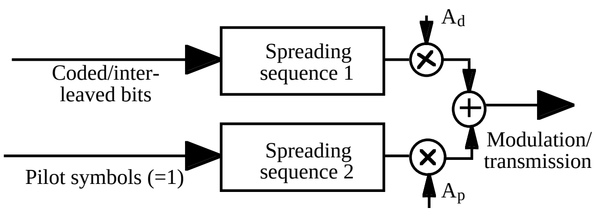

Figure 2 Pilot channel assisted scheme

In pilot assisted coherent CDMA communications, it is necessary to generate an adequately accurate channel estimate from the pilot signals in order to achieve effective coherent detection. In a base-station with a Rake receiver, the received signal passes through a front end filter that matches to the transmitted chip pulse waveform and the output of the matched filter is sampled at the chip-rate to generate chip-rate samples. In this paper, we assume that the integral of the squared magnitude of the filter frequency response is equal to unity. The sampling instant is selected to match the path delay of the communication channel such that the output sample energy is maximized. For a multipath channel, multiple sampled sequences, each of which corresponds to one of the paths, are generated. Each of such sequences is processed individually by a processor called a *Rake finger.*

For the pilot symbol-assisted scheme, each chip-rate sample in a sample sequence is multiplied by the conjugate of the complex chip values of the spreading sequence to generate a *descrambled sample*. Every N_{cd} descrambled samples associated with a coded bit are summed together to yield a data sample. Similarly, every N_{cp} descrambled samples associated with a pilot symbol are summed together to yield a pilot sample, which is a noisy channel estimate of the corresponding path.

For the pilot channel assisted scheme, chip-rate samples from the matched filter are multiplied by the conjugate of the corresponding complex values of the data spreading sequence to generate descrambled *data chip-samples*. The data chip samples associated with a coded bit are summed together to generate a data sample. On the other hand, the same chip-rate samples are multiplied by the conjugates of the complex values of the pilot channel spreading sequence to yield a descrambled *pilot chip-sample*.

An accurate channel estimate can be generated from a plurality of the pilot samples (symbol assisted case) or pilot chip-rate samples (pilot channel assisted case) as discussed below.

3. PILOT ASSISTED CHANNEL ESTIMATION

Below we describe the optimal minimum mean square error (MMSE) and optimal minimax robust channel estimators for these two schemes.

For the Rake receiver described above, the signal at the matched filter output can be modeled as the transmitted signal passing through a flat fading channel. Thus, in this paper, we shall only analyze the performance for flat Rayleigh fading channels. In a wireless communication environment, such a fading channel can be characterized by a low-pass random process. Hence, the k-th chip-rate sample, which is associated with the n-th pilot symbol $a_p(n)$, for Rake finger i, denoted by $r_i(k)$, can be expressed as

$$r_{pi}(k) = \alpha_i(k)\sqrt{E_c}c(k)A_p a_p(n) + z_c(k) \tag{1}$$

where $\alpha_i(k)$ is the complex channel gain of the i-th Rake finger, E_c is the energy per chip waveform, $c(k)$ is the complex value of the k-th chip, and $z_c(k)$ is the portion of the total interference including thermal noise, other user signal and portions of the user's own signal. The variance of $z_c(k)$ is equal to I_0, where I_0 is the power density of the total interference. Then, the SIR of $r_{pi}(k)$ is

$$\gamma_{ci} = |\alpha_i(k)|^2 |A_p|^2 E_c / I_0 . \tag{2}$$

Multiplying the pilot chip samples $r_{pi}(k)$ by the complex conjugate of $c_p(k)$, we have the unscrambled pilot chip sample

$$\tilde{r}_{pi}(k) = c_p{}^*(k) r_{pi}(k) = \alpha_i(k)\sqrt{E_c}A_p + z'(k) . \tag{3}$$

The SIR of $\tilde{r}_{pi}(k)$ is also equal to γ_{ci}. For a Rayleigh fading channel, the channel coefficient $\alpha_i(k)$ is a low-pass Gaussian random process with a power spectrum of $\phi(f)$, $\phi(f) = 0$ for $f \geq f_d$. It is called the *Doppler spectrum* and f_d is the maximum Doppler frequency. For mobile communications, the most commonly used channel model is described in [8] and it is often called the "Jakes model," which has the Doppler spectrum

$$\phi(f) = K[1 - (f / f_d)^2]^{-1/2} , \tag{4}$$

where K is a constant. This model characterizes the fading channel for vertical antenna responsive to the E_z field. However, it should be noted that even though the Jakes model is widely used, the fading process in real world may take other forms. For example, as shown in [8], the Doppler spectra of signals received by antennas responsive to H fields are different from (4).

The purpose of the channel estimator is to determine the complex channel coefficient $\alpha_i(k)$ up-to a fixed scaling factor. If the Doppler spectrum $\phi(f)$ is known, it is well known from the Wiener filter theory that, in white noise, the optimal MMSE estimator of $\alpha_i(k)$ is a linear phase filter, which has a frequency response equal to the square root of $\phi(f)$. On the other hand, since the Doppler spectrum is usually not known, or it may change with time, our best knowledge about $\alpha_i(k)$ is that it is a lowpass process with a maximum Doppler frequency f_d. In such a case, we may use an ideal low pass filter with a cut-off frequency greater than or equal to f_d. as the fixed (non-adaptive) estimator. Such an estimator, as shown in the Appendix, is an optimal robust estimator in the sense that it minimizes the worst case relative mean square estimation error (RMSEE). Note that the RMSEE for the robust estimator is determined only by the average channel gain and independent of its spectrum shape. This characteristic greatly simplifies our analysis. Practically, it is more realistic to implement the channel estimator using a low pass filter which has a flat frequency response up to f_d with a noise bandwidth $B_n > f_d$. Below we consider the implementation of such robust estimators and their RMSEE for the pilot symbol and pilot channel assisted schemes.

Let us first consider the following general case. The mobile transmitter sends pilot symbols at a frequency of f_p, and each pilot symbol is spread to N_{cp} chips. At the base receiver, the corresponding N_{cp} unscrambled pilot chip samples given by (3) are summed together (despread) to form a despread pilot sample

$$p_i(n) = \sum_k \tilde{r}_{pi}(k) = N_{cp}\alpha_i(n)\sqrt{E_c}A_p + z'(k) , \tag{5}$$

where we assume that the channel does not change during the N_{cp} chip intervals. The variance of $z'(k)$ is equal to $N_{cp}I_0$. By passing $p_i(n)$ through the channel estimator, the relative estimation error can be reduced.The low-pass filter will reduce the relative noise variance by a factor of $f_p/2B_n$. After normalization, the channel estimate obtained from such a robust channel estimator can be expressed as

$$\hat{\alpha}'_i(n) = \alpha_i(n) + \hat{z}'(k), \qquad (6)$$

where the variance of $\hat{z}'(k)$, which is zero mean, is

$$\sigma^2_{\hat{z}} = (2B_n/f_p)I_0/(|A_p|^2 N_{cp}E_c). \qquad (7)$$

For pilot symbol based estimation, $f_p = f_c/[N_{cp}\times(M+1)]$ and $A_p = 1$. Thus, we have

$$\sigma^2_{\hat{z}} = (M+1)(2B_n/f_c)(I_0/E_c), \qquad (8)$$

and its RMSEE, denoted as ζ_i, is

$$\zeta_i = 2(M+1)I_0B_n/(|\alpha_i(n)|^2 E_c f_c) \qquad (9)$$

RMSEE can also be written in a general form:

$$\zeta_i = 2B_nI_0/(|\alpha_i(n)|^2 P_p), \qquad (10)$$

where $P_p = E_cf_c/(M+1)$ and $|\alpha_i(n)|^2P_p$ is the received pilot power. In words, *RMSEE is equal to the noise density times twice of the estimator noise bandwidth divided by the received pilot power.*

For pilot channel assisted estimation, it is possible to directly pass the unscrambled pilot chip samples given by (3) through a channel estimator. However, in order to reduce computational complexity, it is desirable to sum N_{cp} chip-samples first to form despread pilot samples. A channel estimate is then obtained by passing these samples through a robust estimator with a noise bandwidth B_n. Noting that the frequency of the despread pilot sample is $f'_p = f_c/N'_{cp}$. and $A_p = 1/\sqrt{P+1}$, the normalized channel estimate can be expressed as

$$\hat{\alpha}''_i(n) = \alpha_i(n) + \hat{z}''(k). \qquad (11)$$

Using (7), we can show the variance of $\hat{z}'(k)$ is also given by (8) and its RMSEE is given by (9) with the understanding that P is used in the places of M, since the transmitted pilot power is simply $f_cE_c/(P+1)$. Thus, pilot symbol and pilot channel assisted schemes yield equivalent channel estimates if P = M.

4. COHERENT DEMODULATION BASED ON PILOT ASSISTED CHANNEL ESTIMATION

Based on the above analysis, we are now ready to analyze the performance of the pilot assisted coherent DS-CDMA reverse link communication. Let us assume that the information bearing data bits at the rate of f_{is} are first coded by a rate $1/r$ convolutional encoder. The coded data symbols are interleaved and organized in frames. At the receiver end, data samples, each of which corresponds to a coded bit, are generated from the descrambling and despreading operations as described above. The despread data samples corresponding to the n-th coded bit can be expressed as

$$d_i(n) = N_{cd}\alpha_i(n)\sqrt{E_c}A_d a(n) + \hat{z}_d(n), \qquad (12)$$

where $\hat{z}_d(n)$ is due to interference with a variance of $N_{cd}I_0$. To perform coherent demodulation, the despread data samples $d_i(n)$ are multiplied by the conjugate of the channel estimates given by (6) or (11) to generate coherent detected samples, which, for a static channel, can be expressed as

$$\hat{\alpha}'^*_i(n)d_i(n) = N_{cd}|\alpha_i(n)|^2\sqrt{E_c}A_d a(n) + \tilde{z}'(n), \qquad (13)$$

where $\tilde{z}'(n)$ is a zero mean noise term, which can be expressed as

$$\hat{z}'^*(k)N_{cd}A_d\alpha_i(n)\sqrt{E_c}a(n) + \alpha_i^*(n)\hat{z}_d(n) + \hat{z}'^*(k)\hat{z}_d(n). \quad (14)$$

The corresponding coherent detected samples from different fingers and antennas are combined. After deinterleaving, these combined samples are used to form decoding metrics in a Viterbi convolutional decoder to recover the information data bits.

In the Viterbi decoder, a decision is made based on the sum of many detected samples. As a result, we can assume that the noise term in the decision variable can be approximated by a Gaussian variable based on the argument of the Central Limiting Theorem. Thus, we can evaluate the performance of the pilot assisted coherent demodulator by evaluating the signal to noise ratio of the detected sample given by (13). Observing that $\hat{z}^*(k)$ and $\hat{z}_d(n)$ are uncorrelated and $A_d = 1$, we have the variance of (14) as

$$\sigma^2_{\tilde{z}'} = N_{cd}I_0|\alpha_i(n)|^2\left\{1 + \frac{2(M+1)B_nN_{cd}}{f_c} + \frac{2(M+1)B_nI_0}{|\alpha_i(n)|^2E_cf_c}\right\} \quad (15)$$

For the static channel, $\alpha_i(n)$ is a constant and the signal to noise ratio of the coherently detected samples given by (13) is

$$\gamma_i' = \frac{N_{cd}|\alpha_i(n)|^2(E_c/I_0)}{[\beta + 1 + \beta I_0/(|\alpha_i(n)|^2E_c)N_{cp}]}, \qquad (16)$$

where we have defined $\beta = 2(M+1)B_nN_{cd}/f_c$.

The performance of an ideal coherent receiver over a stationary channel or a Rayleigh fading channel with additive white Gaussian noise (AWGN) has been thoroughly analyzed in the literature. Hence, it is informative and sufficient to compare the performance of pilot assisted coherent demodulation with that of ideal coherent demodulation when the channel is perfectly known. In the latter case, the coherently detected sample is

$$\alpha_i^*(n)d_i(n) = N_{cd}|\alpha_i(n)|^2\sqrt{E_c}a(n) + \alpha_i^*(n)\hat{z}_d(n) \qquad (17)$$

The signal to noise ratio of $\alpha_i^*(n)d_i(n)$ is

$$\gamma_i = N_{cd}|\alpha_i(n)|^2E_c/I_0. \qquad (18)$$

Dividing (16) by(18), we obtain,

$$\gamma_i'/\gamma_i = 1/(1 + \beta + \beta/\gamma_i). \qquad (19)$$

It can be seen from (9) and (18) that $\beta = \zeta_i\gamma_i$.

It is more informative to express (19) in terms of the information bit rate f_{is} and the coding rate r. By noting that $f_{is} = rMf_c/[(M+1)N_{cd}]$ and then substituting this relation and (8) into (19), we obtain

$$\frac{\gamma_i'}{\gamma_i} = \left[1 + \frac{2rMB_n}{f_{is}}\left(1 + \frac{1}{\gamma_i}\right)\right]^{-1} \qquad (20)$$

For the pilot channel assisted demodulation, the spreading factor is $N_{cd}(1+1/M)$ and $A_d^2 = P/(P+1)$. letting M = P, the despread data samples have the same expression as (12). Thus, the two schemes have the same performance, if M = P.

For a Rayleigh fading channel, the average noise variance of the coherently detected data samples can be written in the same form as (15) with $|\alpha_i|^2$ replaced by its expectation, $E[|\alpha_i|^2]$. Its signal power will also has the same expression with $|\alpha_i|^2$ replaced by $E[|\alpha_i|^2]$. Thus, all the expressions remain the same with the understanding that such a replacement is made.

Equation (20) shows that the inaccuracy of the channel estimates degrades receiver performance relative to that of an ideal coherent receiver. In addition, the insertion of non-information bearing pilot signal effectively reduces E_b, the energy per information bit, by a factor of $(M+1)/M$ for the same total transmitted power. Thus, the total performance loss is

$$\eta_{total} = \left(\frac{M}{M+1}\right)/\left[1 + \frac{2rMB_n}{f_{is}}\left(1 + \frac{1}{\gamma_i}\right)\right] \qquad (21)$$

Because the power allocation of the pilot and data signals affects both factors in opposite ways, there exist an optimum M or P, which minimizes the loss. By taking the derivative of (21) with respect to M, letting it equal to zero and solving for M we obtain

$$M_{opt} = \sqrt{\frac{f_{is}\gamma_b}{2B_n(1+r\gamma_b)}}, \tag{22}$$

where γ_b is the required energy per information bit per noise density by an ideal coherent receiver to achieve a certain bit or frame error rate. By substituting (22) into (21), we can compute the performance loss of the pilot assisted scheme with M_{opt} relative to the optimal coherent demodulation. As an example, we consider the case in which $r = 1/4$, $B_n = 300$ Hz, $f_{is} = 9600$ Hz and $\gamma = 1.4$ dB, Using (22), we compute the optimal M is approximately equal to 4. From (21) we compute the loss is approximately 1.42 dB.

The pilot assisted coherent detection, even with the loss caused by pilot insertion and estimation error, is advantageous to the non-coherent detection when coding and diversity combining of signals from multiple Rake fingers and/or multiple antennas is used. To show the advantage of such schemes, simulation results of a pilot symbol assisted coherent communication are shown in Figure 3. A rate 1/4, K = 9, convolutional encoder is used in the simulation. The data at 9600 bps is organized as 20 ms frames. The data to pilot ratio is M = 4. For comparison, the performance of a non-coherent IS-95 reverse link receiver is also given. Two antenna diversity is used with independent flat fading channels. It can be seen that the coherent scheme has an advantage of 3 dB at 100 kmph with a 900 MHz carrier frequency. The simulation results also verify the theoretical predictions given above.

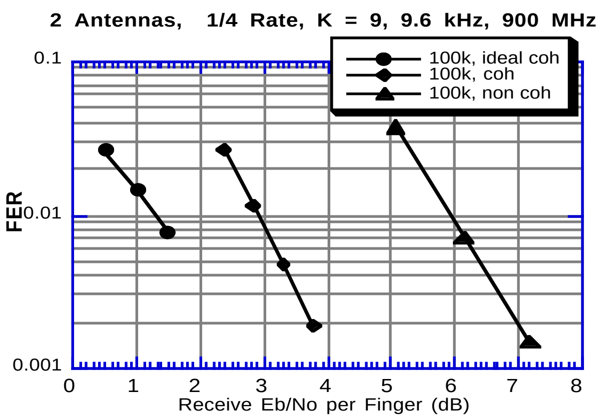

Figure 3. Frame error rate for different schemes

5. CONCLUDING REMARKS

In this paper, we consider the optimal pilot assisted estimation of communication channels for coherent cellular and PCS CDMA reverse link communication. Both pilot symbol and pilot channel based schemes are described and the optimal estimators for these two schemes are analyzed. An analysis of the relative mean square channel estimation error shows that it is determined by only the received pilot signal power, the noise-bandwidth of the estimator and the interference density. The performance loss of the pilot assisted schemes relative to the ideal coherent detection is also analyzed and a formula for optimal power allocation between the data and pilot signals is derived based on the analysis. Moreover, it is shown that, for either scheme, required SIR depends on the information bit rate. The performance will be closer to the ideal coherent demodulation at the higher data rate. Although the analysis was based on the mean square error criterion, these results agree with our recent analysis on Chernoff-bound and cut-off rate for Rayleigh fading channels. Although it has been shown that the two pilot assisted schemes are equivalent theoretically, we would like to note that the signal of the pilot symbol assisted scheme has a lower peak to average ratio and will potentially have less cross interference between the data and pilot. Finally, simulation results are given to show the pilot assisted coherent reverse link can achieve a performance that is 3 dB better than the IS-95 non-coherent reverse link at 9600 bps. It should be noted that such a gain is obtained with less receiver complexity.

This paper is intended to present pilot assisted coherent CDMA reverse-link communications from a signal processing point of view. Other aspects of such schemes related to digital communication theory will be discussed in a forthcoming paper.

APPENDIX

The RMSEE is defined to be the average of the squared estimation error divided by the average of the squared channel coefficient. When estimating the channel coefficient from the pilot symbol given by (5) using a linear filter with a frequency response H(f), the RMSEE is

$$\zeta = \frac{\sigma_{\tilde{z}}^2\int_{-1/2}^{1/2}|H(f)|^2 df}{\int_{-1/2}^{1/2}\phi(f)|H(f)|^2 df} = \frac{\sigma_{\tilde{z}}^2}{\int_{-1/2}^{1/2}\phi(f)|H(f)|^2 df}. \tag{23}$$

where, without loss of generality, we assume $f_d = 1/2$ and H(f) is normalized so that the integral in the numerator is equal to 1. To compare different fading spectra, we assume ϕ(f) is also normalized, i.e., the integral of ϕ(f) from -1/2 to 1/2 is equal to 1.

Since H(f) is ideally low-pass, $|H(f)|^2 = 1$, for$1/2<f<1/2$. Thus, the denominator is simply the integral of ϕ(f) that is equal to one. and the relative error is equal to $\sigma_{\tilde{z}}^2$

If $|H(f)|^2$ is not identically equal to 1, there must exist at least one non-trivial interval in [-1/2, 1/2] on which $|H(f)|^2$ is less than 1. Let $|H(f)|^2< 1$, on [a, b], $-1/2<b<f<a<1/2$. We construct a $\tilde{\phi}(f)$ such that $\tilde{\phi}(f) = 0$, for f < b and f > a, and $\tilde{\phi}(f) = 1/(a-b)$ for $b<f<a$. It is easy to verify that $\tilde{\phi}(f)$ satisfy the condition for ϕ(f). Then, the denominator of (22) is less than 1 and the relative error is greater than $\sigma_{\tilde{z}}^2$ for $\tilde{\phi}(f)$. Similarly, we can show that it is also true without the white-noise assumption.

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