CHANNEL IDENTIFICATION AND APPLICATIONS TO OFDM COMMUNICATION SYSTEMS WITH LIMITED BANDWIDTH

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ABSTRACT

A novel channel identification algorithm is presented for OFDM communication systems in which the signal bandwidth is limited severely. Since little information outside the signal band can be extracted from the received signal directly in conventional methods, it is usually difficult to identify the channel dynamics with signal band restriction; as a result, channel identification remains large uncertainty, which leads to fragile performance of adaptive systems based on channel estimation especially outside the signal band. In order to explore much information, some specified signals are constructed and their spectral properties are analyzed by using the distinctive feature of OFDM guard interval, furthermore the spectra based identification algorithm is developed. The main computation can be performed through fast Fourier transform so the computational complexity is low. The effectiveness of the proposed approach is demonstrated by numerical simulations of channel identification and coupling cancellation in an OFDM communication relay system.

1. INTRODUCTION

Persistent excitation (PE) is essentially required in system identification. To explore the dynamic characteristics of a system through identification techniques, one should excite sufficient dynamic modes especially in the frequency band dominating the system performance. This requirement can be fulfilled when the PE test signals can be used for identification. Nevertheless, the user's freedom in choosing input characteristics may not be allowed in many applications, e.g., some process industries, communication systems with specified restrictions on signal band. If the excitations are not informative enough, i.e., they do not contain sufficient distinct frequency components, some dynamic modes of the system cannot be excited; as a result, the identification problem becomes very difficult [1]. For example, in multi-carrier communication systems such as OFDM, DMT, some carriers far away from the central band frequency do not convey any information data in order to simplify the design of filter circuit and not to interference the adjacent communication channels, thus the transmitted signal is restricted within a specified frequency band. When one uses the estimation of dynamic transmission channel to determine the coefficients of equalizer or interference canceller [2], some coefficients are left unaltered since the dynamic modes are not excited at the carriers that do not convey data, as a result, the confidential performance of the equalizer or interference canceller cannot be guaranteed [3].

The solutions to this problem were considered in some previous works. A straightforward method was proposed to transmit dummy signal with low power level [4], however, the dummy signal was undesired in practical systems since it caused other problems such as increasing power of transmitted signal, requirement of extra circuit to deal with dummy signal, etc. Alternatively a bandpass filter with very steep characteristics was used to reduce the affection of the model uncertainty outside the signal band, but such a filter prolonged the channel impulse response to cause inter-carrier interference. Circulant decomposition canceller design in DMT receivers splitted the received data matrix into a circular and a skew-circular part, which leaded to the frequency domain and partial time domain echo cancelling respectively [3], however, its slow convergence restricted its practical applications. A time domain approach to identify the dynamics outside the signal band [5] was also proposed, it's computational complexity was still high for real-time implementation. Though extrapolation methods or regularization methods such as Levenberg-Marquardt procedure can estimate the dynamic characteristics outside the signal band to some extent, its uncertainty is too large to be used in system design when the width of outside signal band is not short enough. Consequently the development of an effective identification algorithm with low computational complexity and high convergence rate is a challenging work in OFDM systems.

In this paper, a novel channel identification algorithm is considered for OFDM communication systems. In order to extract much information of the dynamic modes for the identification, some specified signals are constructed from the received signal by considering the distinctive property of guard interval, and their relation with the channel dynamics is revealed through Fourier analysis techniques. Then a new identification algorithm is developed by using the relation. Compared with conventional methods of band-limited identification, the algorithm is implemented in frequency domain mainly through FFT so it has low computational complexity and can be applied for adaptive processing systems.

2. COMMUNICATION CHANNEL MODEL

For the simplicity of notation, the sampling instants of the OFDM signals are normalized and denoted as integer k. Furthermore, the normalized discrete instant in the l-th transmission symbol period is indicated as m, the normalized frequency of n-th sub-carrier is ω_n, $\omega_n = 2\pi n/M$, M is the window length of the Fourier transform.

2.1 OFDM Signal Band

In the l-th transmission symbol period, the baseband signal $d(l,m)$ is generated through the M-point IFFT

$$d(l,m) = \sum_{n=-M/2+1}^{M/2} D(l,n)e^{j\omega_n m}, \quad (1)$$

where $D(l,n)$ is the information symbol conveyed on the n-th sub-carrier, which is given as follows.

$$D(l,n) = \begin{cases} \text{Information data}(\neq 0), & \text{for } |n| \leq N_1 \\ 0, & \text{for } |n| > N_1 \end{cases} \quad (2)$$

where $N_1 = (N-1)/2$, and N is the number of active sub-carriers, $N < M$. It can be seen that the carriers whose distance from the central carrier is more than N_1 do not carry any information data, hence the spectrum of transmitted signal $d(l,m)$ in an effective symbol period, i.e. $m = 0,\cdots,M-1$ is limited to $|n| \leq N_1$, whereas the spectral density outside signal band, i.e. for $|n| > N_1$, becomes to 0. It implies that $d(l,m)$ and its corresponding received signal $r(l,m)$ for $m = 0,\cdots,M-1$ do hold little information about the channel dynamics outside the signal band.

Since $D(l,n)$ is the element of a specified constellation, the following Proposition holds.

Proposition 1 *If the information symbol $D(l,n)$ can be treated as a uniformly distributed random sequence with respect to l, n, then for any integers l_1, l_2, and any active sub-carrier n_1, n_2,*

$$\lim_{L\to\infty} \frac{1}{L}\sum_{l=1}^{L} D^*(l+l_1,n_1)D(l+l_2,n_2) = \bar{D}^2\delta(l_1-l_2)\delta(n_1-n_2) \quad (3)$$

holds, where $\bar{D}^2$ is a constant determined by the constellation, $|n_1| \leq N_1$ and $|n_2| \leq N_1$.

Furthermore, for the purpose of synchronization, the pilot symbols are assigned at some specified sub-carriers, whose frequencies are denoted as $\omega_{P,n}$. It implies that when the synchronization with the base station is achieved, the source information symbol carried by P_n-th sub-carrier at frequency $\omega_{P,n}$ is known as *a priori*.

2.2 OFDM Guard Interval

In OFDM systems, guard interval is attached at the header of every transmission symbol period. If the length of guard interval is m_{gi}, $d(l,m)$ for $m = -m_{\text{gi}},\cdots,-1$, which is inside the guard interval, is equal to the signal for $m = M - m_{\text{gi}},\cdots,M-1$ at the tailor of the symbol period respectively. Then, the practical transmission period including guard interval becomes to $M + m_{\text{gi}}$, and the relation between the normalized discrete time of k and m can be given by $k = (l-1)(M+m_{\text{gi}}) + m_{\text{gi}} + m$, i.e.,

$$d(k) = d(l,m), \ \text{for } k = (l-1)(M+m_{\text{gi}}) + m_{\text{gi}} + m \quad (4)$$

2.3 Channel Model

When the communication is suffered from multi-path interference, the received signal $r(k)$ is approximated by

$$r(k) = \sum_{\tau=0}^{L_h} h_\tau d(k-\tau) + v(k), \quad (5)$$

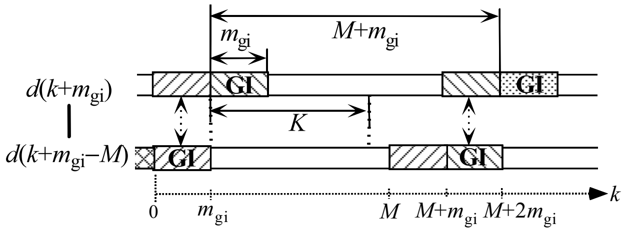

Figure 1: Illustration of signal $x(k)$

where the additive noise $v(k)$ is often assumed to be a white Gaussian noise with zero mean and finite variance, and independent of the source signal $d(k)$. $r(k)$ and $v(k)$ can also be arranged in the form of $r(l,m)$, $v(l,m)$ similarly as $d(l,m)$ in (4). h_τ is the coefficient corresponding to the τ-th delay interference wave, and the effective tap length is $L_h + 1$. Then the channel dynamics is expressed by

$$H(z) = h_0 + h_1 z^{-1} + \cdots + h_{L_h} z^{-L_h}. \quad (6)$$

3. EXPLORATION OF SIGNAL PROPERTY

Let $R(l,n)$ be the Fourier transform of the received signal $r(l,m)$ at the n-th carrier frequency in the l-th symbol period within the FFT window $m = 0,\cdots,M-1$. For $|n| \leq N_1$, (5) can be rewritten in the frequency domain as

$$R(l,n) = H(e^{j\omega_n})D(l,n) + V(l,n). \quad (7)$$

Nevertheless, following the fact that $D(l,n) = 0$ for $|n| > N_1$, the components remaining in the spectrum $R(l,n)$ are noise components and cannot provide sufficient information for identification, so extra information should be extracted.

3.1 Fourier Analysis of Specified Signals

Two signals are constructed from the original transmitted and received signals as follows:

$$x(k) = d(k+m_{\text{gi}}) - d(k+m_{\text{gi}}-M), \quad (8)$$

$$y(k) = r(k+m_{\text{gi}}) - r(k+m_{\text{gi}}-M). \quad (9)$$

As an example, the signal $x(k)$ is illustrated in Figure 1, where K is an integer satisfying $m_{\text{gi}} < K + m_{\text{gi}} \leq M$, e.g., $K = M/2$. Moreover, the formulation of $x(l,m)$ and $y(l,m)$ can also be constructed similarly as $d(l,m)$ or $r(l,m)$. From the feature of guard interval, $x(l,m) = 0$ holds for $-m_{\text{gi}} \leq m < 0$. Then substituting (1) into the expression of $x(l,m)$ yields that

$$x(l,m) = \sum_{n=-N_1}^{N_1} D(l,n)e^{j\omega_n(m-m_{\text{gi}})} - \sum_{n=-N_1}^{N_1} D(l-1,n)e^{j\omega_n m}$$
$$= \sum_{n=-N_1}^{N_1} \left(D(l,n)e^{-j\omega_n m_{\text{gi}}} - D(l-1,n)\right) e^{j\omega_n m}, \ \text{for } 0 \leq m < K \quad (10)$$

Next consider the signal $y(k)$. Let its component corresponding to tap delay τ be denoted by $y_\tau(k)$, then neglecting the noise term, $y_\tau(k)$ and $y(k)$ can be expressed by

$$y_\tau(k) = h_\tau x(k-\tau), \ \ y(k) = \sum_{\tau=0}^{L_h} y_\tau(k) \quad (11)$$

Moreover, in the l-th symbol period, $y_\tau(l,m)$ becomes to

$$y_\tau(l,m) = \begin{cases} h_\tau \sum_{n=-N_1}^{N_1} \left(D(l,n)e^{-j\omega_n m_{\mathrm{gi}}} - D(l-1,n)\right) e^{j\omega_n(m-\tau)}, & \text{for } \tau \le m < K \\ 0, & \text{for } 0 \le m < \tau \end{cases} \quad (12)$$

On the other hand, let the frequency properties of $x(l,m)$ in $0 \le m < K$ be defined by

$$X(l,n) = \sum_{m=0}^{K-1} x(l,m)e^{-j\omega_n m} \quad (13)$$

for $n = -M/2+1,\cdots,M/2$, and similarly $y_\tau(l,m)$, $y(l,m)$ respectively, then Lemma 1 holds.

Lemma 1 *The frequency properties of signals $x(k)$ and $y(k)$ satisfy the following equation*

$$H(e^{j\omega_n})X(l,n) = Y(l,n) + \sum_\tau h_\tau \left(E_{\tau,1}(l,n) + E_{\tau,2}(l,n)\right) \quad (14)$$

where $E_{\tau,1}(l,n)$ and $E_{\tau,2}(l,n)$ are distortion terms given by

$$E_{\tau,1}(l,n) = e^{-j\omega_n\tau} \sum_{\substack{\bar{n}=-N_1 \\ n-\bar{n}\neq 0,M}}^{N_1} \left(D(l,\bar{n})e^{-j\omega_{\bar{n}} m_{\mathrm{gi}}} - D(l-1,\bar{n})\right) \cdot \frac{e^{-j(\omega_n-\omega_{\bar{n}})(K-\tau)} - e^{-j(\omega_n-\omega_{\bar{n}})K}}{1-e^{-j(\omega_n-\omega_{\bar{n}})}} \quad (15)$$

$$E_{\tau,2}(l,n) = \tau e^{-j\omega_n\tau}\Big(D(l,n)e^{-j\omega_n m_{\mathrm{gi}}} - D(l-1,n) + D(l,n-M)e^{-j\omega_n m_{\mathrm{gi}}} - D(l-1,n-M)\Big). \quad (16)$$

3.2 Spectral Estimation

Let the spectral estimation of $\overline{XX}(l,n)$ be defined by

$$\overline{XX}(l,n) = \frac{1}{l}\sum_{l_1=1}^{l} X^*(l_1,n)X(l_1,n), \quad (17)$$

and similarly $\overline{XE}_{\tau,1}(l,n)$ and $\overline{XE}_{\tau,2}(l,n)$, then $\overline{XE}_{\tau,1}(l,n)$ and they can be determined by Lemma 2.

Lemma 2 *If l is large enough, $\overline{XX}(l,n)$ is approximated by*

$$\overline{XX}(l,n) \approx 2\overline{D}^2 \sum_{\substack{\bar{n}=-N_1 \\ n-\bar{n}\neq 0,\pm M}}^{N_1} \frac{1-\cos(\omega_n-\omega_{\bar{n}})K}{1-\cos(\omega_n-\omega_{\bar{n}})} + 2K^2\overline{D}^2 \quad (18)$$

for $|n| \le N_1$, and

$$\overline{XX}(l,n) \approx 2\overline{D}^2 \sum_{\bar{n}=-N_1}^{N_1} \frac{1-\cos(\omega_n-\omega_{\bar{n}})K}{1-\cos(\omega_n-\omega_{\bar{n}})} \quad (19)$$

for $N_1 < |n| < M/2$. Meanwhile, $\overline{XE}_{\tau,1}(l,n)$ satisfies

$$\overline{XE}_{\tau,1}(l,n) \approx \overline{D}^2 \sum_{\substack{\bar{n}=-N_1 \\ n-\bar{n}\neq 0,\pm M}}^{N_1} \left(\frac{1-e^{-j(\omega_n-\omega_{\bar{n}})K}}{1-\cos(\omega_n-\omega_{\bar{n}})}\left(e^{-j\omega_n\tau} - e^{-j\omega_{\bar{n}}\tau}\right)\right). \quad (20)$$

Furthermore, spectral distortion $\overline{XE}_{\tau,2}(l,n)$ is

$$\overline{XE}_{\tau,2}(l,n) \approx 2e^{-j\omega_n\tau}\overline{D}^2 K\tau \quad (21)$$

for $|n| \le N_1$, while for $N_1 < |n| < M/2$ it turns to

$$\overline{XE}_{\tau,2}(l,n) = 0. \quad (22)$$

Following Lemma 1, the relation between spectra of $x(k)$, $y(k)$ and the frequency property of the dynamic channel $H(e^{j\omega_n})$ is summarized in Theorem 1.

Theorem 1 *Spectra of $x(k)$, $y(k)$ and the channel satisfy*

$$H(e^{j\omega_n})\overline{XX}(l,n) \approx \overline{XY}(l,n) + \sum_\tau h_\tau\Big(\overline{XE}_{\tau,1}(l,n) + \overline{XE}_{\tau,2}(l,n)\Big). \quad (23)$$

In (23), $\overline{XX}(l,n)$ and $\overline{XY}(l,n)$ can be estimated from $x(k)$ and $y(k)$ directly. On the other hand, from Lemma 2, the spectral distortion terms $\overline{XE}_{\tau,1}(l,n)$ and $\overline{XE}_{\tau,2}(l,n)$ can be calculated beforehand without any observation data and the information of channel dynamics. Consequently, it is possible to estimate the channel property outside the signal band from signals $x(k)$ and $y(k)$ if $\overline{XY}(l,n)$ is compensated by $\overline{XE}_{\tau,1}(l,n)$ and $\overline{XE}_{\tau,2}(l,n)$. Though $\overline{XX}(l,n)$ can also be precalculated by (18) and (19), the approximation error is large for small l so the estimation of (17) is used especially in the first several iterations when l is small.

On the other hand, when $H(e^{j\omega_n})$ is time-varying, a forgetting factor λ, where $0 < \lambda < 1$, can be used to estimate $\overline{XX}(l,n)$ and $\overline{XY}(l,n)$

$$\overline{XX}(l,n) = \lambda\overline{XX}(l-1,n) + X^*(l,n)X(l,n), \quad (24)$$

$$\overline{XY}(l,n) = \lambda\overline{XY}(l-1,n) + X^*(l,n)Y(l,n) \quad (25)$$

respectively.

4. ALGORITHM IMPLEMENTATION

4.1 Identification Algorithm

Provided that the scattered pilot symbols are assigned at P_n-th sub-carrier with normalized frequency $\omega_{P,n}$. Then the information symbol $D(l,P_n)$ at the pilot sub-carrier P_n is known from the information of pilot symbols, consequently

$$\hat{H}(e^{j\omega_{P_n}}) = \frac{R(l,P_n)}{D(l,P_n)} \quad (26)$$

is obtained at pilot carrier frequency ω_{P_n}. As for the channel dynamics at non-pilot sub-carriers inside the signal band, interpolation is an effective choice since the scattered pilot symbols are distributed within $|n| \le N_1$ uniformly. To simplify the computation, a linear interpolation is considered as follows.

$$\hat{H}(e^{j\omega_n}) = \hat{H}(e^{j\omega_{P_{n,1}}}) + \frac{n-P_{n,1}}{P_{n,2}-P_{n,1}}\left(\hat{H}(e^{j\omega_{P_{n,2}}}) - \hat{H}(e^{j\omega_{P_{n,1}}})\right), \quad (27)$$

where $P_{n,1}$ and $P_{n,2}$ are the number of two adjacent pilot sub-carriers, $P_{n,1} \le n \le P_{n,2}$. Moreover, the source symbols

$\hat{D}(l,n)$ can be given by

$$\hat{D}(l,n) = \begin{cases} \dfrac{R(l,n)}{\hat{H}(e^{j\omega_n})}, & \text{for } |n| \le N_1 \\ 0, & \text{for } N_1 < |n| < \dfrac{M}{2} \end{cases}, \quad (28)$$

then $\hat{d}(l,m)$ can be calculated by IFFT easily.

By the virtue of Theorem 1, the estimation of channel model outside the signal band can be deduced iteratively.

$$\hat{H}^{(i+1)}(e^{j\omega_n}) = \bar{H}(e^{j\omega_n}) + \sum_{\tau} \hat{h}_{\tau}^{(i)} A_{\tau}(l,n) \quad (29)$$

where $\hat{h}_{\tau}^{(i)}$ is the coefficients of IFFT of $\hat{H}^{(i)}(e^{j\omega_n})$, and $\bar{H}(e^{j\omega_n}), A_{\tau}(l,n)$ are given by

$$\bar{H}(e^{j\omega_n}) = \hat{H}^{(0)}(e^{j\omega_n}) = \begin{cases} \hat{H}(e^{j\omega_n}), & |n| \le N_1 \\ \frac{\overline{XY}(l,n)}{\overline{XX}(l,n)}, & |n| > N_1 \end{cases}, \quad (30)$$

$$A_{\tau}(l,n) = \frac{\overline{XE}_{\tau,1}(l,n) + \overline{XE}_{\tau,2}(l,n)}{\overline{XX}(l,n)}. \quad (31)$$

4.2 Computational Complexity

The main numerical computation in the proposed algorithm is just FFT to estimate the signal spectra $\overline{XX}(l,n)$, $\overline{XY}(l,n)$ of $x(k)$ and $y(k)$; division of $\left(\overline{XE}_{\tau,1}(l,n) + \overline{XE}_{\tau,2}(l,n)\right)$ and $\overline{XX}(l,n)$, while the former two terms are pre-calculated, division of $\overline{XY}(l,n)$ and $\overline{XX}(l,n)$, IFFT to calculate $\hat{h}_{\tau}^{(i)}$. Furthermore, the computational complexity does not increase too much even though the interference delay taps get longer. So the proposed algorithm can be implemented easily.

5. APPLICATION TO COUPLING WAVE CANCELLATION

Consider an application of the proposed channel identification algorithm to coupling wave cancellation in OFDM relay station to enlarge the service area. Since the carrier frequency for transmission is the same as that of the received signals at the relay station, the radio wave from the transmitter antenna couples into the receiver antenna leading to coupling wave interference, which degrades the quality of the signal transmission, even causes the serious oscillation problem. Hence, the coupling cancellation schemes should be developed to reduce the coupling effects [6]. The diagram of the relay station with coupling interference is illustrated in Figure 2, where $C(z)$ denotes the model of coupling interference, $W(z)$ is the time domain canceller, g is the amplifier gain, whereas $s(k)$ is the re-transmitted signal.

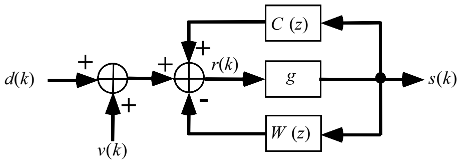

Figure 2: Schematic diagram of relay station

Then the channel model can be approximated by

$$r(k) = H(z)(d(k) + v(k)), \quad s(k) = gr(k) \quad (32)$$

where

$$H(z) = \frac{1}{1 + g(W(z) - C(z))}$$

Obviously $H(z)$ is a transfer function of closed-loop model, hence its poles must be located inside the unit circle to guarantee the stability of the cancellation system when updating canceller $W(z)$. Accordingly the dynamic characteristics of $H(z)$ over the full frequency band is required not only for updating $W(z)$, but also for the system stability.

When the estimate $\hat{H}(e^{j\omega_n})$ is obtained, the canceller updating in the frequency domain can be given by

$$\Delta W(e^{j\omega_n}) = \mu g \left(1 - \frac{1}{\hat{H}(e^{j\omega_n})}\right) \quad (33)$$

where μ is the step size. To guarantee the stability, following the small gain theorem, it should be satisfied that

$$\left|\Delta W(e^{j\omega_n}) H(e^{j\omega_n})\right| \le \alpha \quad (34)$$

where $0 < \alpha \le 1$ is a constant. Then μ can be determined by

$$\mu = \min_n \left\{ \alpha, \frac{\alpha}{g\left|\hat{H}(e^{j\omega_n}) - 1\right|} \right\} \quad (35)$$

Since the channel identification can be executed for every symbol period, the canceller updating can also be performed iteratively. Thus, when the channel model $H(e^{j\omega_n})$ outside the signal band cannot be obtained, neither $\Delta W(e^{j\omega_n})$ nor μ can be calculated so the cancellation stability might lose during canceller updating.

6. NUMERICAL SIMULATIONS

In the example, the OFDM information symbols $D(l,n)$ are 64QAM, the FFT/IFFT length is $M = 2048$, the guard interval is $m_{\text{gi}} = M/4$. Let the number of active sub-carriers be $N = 1201$, which implies that the signal band is only about $3/5$ of the full band width. There are 6 symbol transmission periods per sub-frame, and 200 scattered pilot subcarriers are distributed uniformly in the first and fourth symbol periods [7]. Let K be chosen as $K = M/2$, the signal-to-noise ratio is 15dB.

6.1 Channel Identification

The true frequency property of $H(e^{j\omega})$ and its estimate after 50 iterations are plotted in Figure 3. It illustrates that the estimate is very close to the true one even outside the signal band, though the transmitted signal $d(k)$ has severe band limitation. As a comparison, the channel is also identified by conventional methods using RLS or LMS, and the results show that both RLS and LMS cannot provide satisfactory identification outside the signal band.

As shown in Figure 4, the signal $x(k)$ has spectral power density of 10^{-3} outside the signal band. Compared with the original spectrum of $d(k)$, the information over the entire frequency band can be extracted from $x(k)$ though the spectral power density outside the band is a little lower than the inside one. It means that it is possible to identify the dynamics of channel even outside the signal band. On the other hand, since the nonlinear distortion of HPA increases the out of

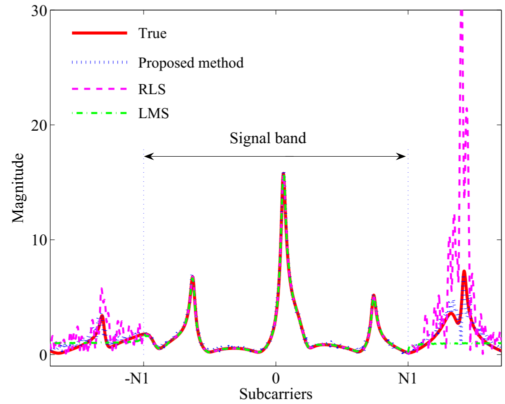

Figure 3: Frequency property of communication channel

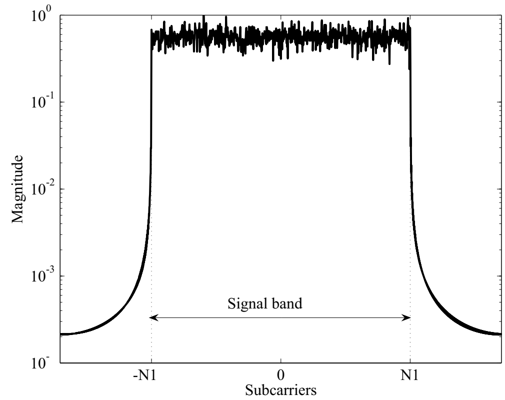

Figure 4: Spectrum of $x(k)$

band noise, some predistortion methods can be used to guarantee the channel identification performance.

The estimation errors outside the signal band under various noise environments are illustrated in Figure 5, where the error is evaluated by

$$E_{H,\mathrm{Out}} = \frac{\sum\limits_{N_1<|n|<\frac{M}{2}} \left|\hat{H}(e^{j\omega_n})-H(e^{j\omega_n})\right|^2}{\sum\limits_{N_1<|n|<\frac{M}{2}} \left|H(e^{j\omega_n})\right|^2}. \quad (36)$$

It can be seen that even for low SNR, the estimation error successfully reduces lower than 10^{-1} just by tens iterations.

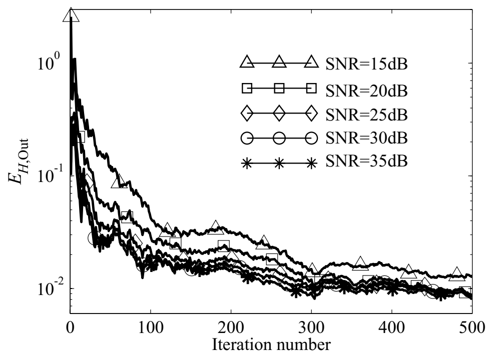

Figure 5: Estimation error vs. different SNR

6.2 Adaptive Coupling Cancellation

Let the canceller $W(z)$ is an FIR filter with 100 taps and its initial value is 0. The true channel model $H(z)$ for $W(z) = 0$ is the same as that in Figure 3.

The ratio of desired and undesired signal powers (DU) for 100 iterations is shown in Figure 6. The DU converges to noise level after 25 iterations and the DU ratio is much better than that of the conventional method without using the estimation outside the signal band, whereas the later falls into instability sometimes.

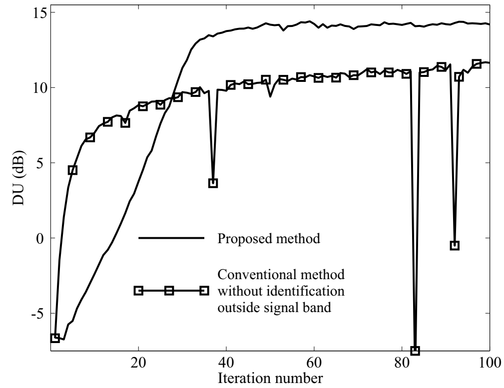

Figure 6: Convergence performance of cancellation

7. CONCLUSIONS

The algorithm for OFDM channel identification with severely limited signal band has been developed. By using the special properties of guard interval and pilot symbols, the approach to extraction of important information from the band-limited signals for channel identification is analyzed. The proposed algorithm is implemented in the frequency domain by the fast Fourier transform, consequently it has low computational complexity.

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