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行政院國家科學委員會專題研究計畫成果報告

下一世代無線行動接取技術(I)-子計畫四:同步、通道估計與內接收機設計(I)

Next Generation Mobile Radio Access Technologies (I): Synchronization, channel estimation and inner receiver design (I)

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中文摘要

我們探討在頻域正交多工（OFDM）系統中合併頻率、資料與通道的估測的可行性。由於載波頻率偏移（Carrier Frequency Offset, CFO）所造成的頻域信號的載波間干擾（Inter-Carrier Interference, ICI）與複徑通道所引起在時間軸上符碼間干擾（Inter-Symbol Interference, ISI）有類似的數學模式，後者的許多已知解決方案，如最大可能性序列估測（Maximum Likelihood Sequence Estimation, MLSE），便可用來解決前者。然而，不同於傳統的 MLSE 可直接套用(Trellis)圖。我們面對的是含有未知數（不確定性）的籬柵圖。我們採用遞迴式估合併檢測與估計（Iterative Joint Detection and Estimation）的原則來解決。

首先，我們針對單天線（SISO）系統的狀況來設計，並且針對效能及複雜度等主要課題提出改進方案。我們討論了利用接收端視窗的等效濾波及刪除較小載波干擾的鄰頻等降低籬柵狀態（trellis state）數目的方法的可行性以降低演算法的複雜度。接著我們將這套遞迴式估合併檢測與估計演算法推展到多天線（MIMO）的系統。

我們所得到的數值結果顯示，無論是 SISO 或 MIMO 環境，我們的解決方案都能提供令人滿意的表現。

關鍵詞：多天線，正交頻率多工，載波頻率偏移，遞迴式檢測與估計

Abstract

We consider the problem of joint carrier frequency offset (CFO)/channel estimation and data detection for Multiple Input Multiple Output-orthogonal frequency-division multiplexing (MIMO-OFDM) systems. An iterative approach in this article is employed. The iteration procedure consists of an inner CFO/data estimation loop and an outer CFO/data-channel estimation loop. First we discuss the single-input single-output (SISO) case and then extend to a multiple-input multiple-output (MIMO) scenario. Drawing an analogy between the inter-carrier interference corrupted frequency domain waveform and the inter-symbol interference limited time domain waveform, we apply maximum likelihood sequence estimation scheme to detect the data sequence in the presence of residual CFO. Related algorithms either for reducing the implementation complexity like the trellis state number or for improving the overall performance such as receiver time domain windowing and refined channel estimates are presented. Computer simulation results are given to predict the performance of the proposed algorithms.

Keywords: MIMO，OFDM、Carrier Frequency Offset ， ICI ， Iterative detection and estimation

1. Introduction

The main idea behind the orthogonal frequency division multiplexing (OFDM) waveform is to split a data block into multiple parallel sub-blocks so that each of them is transmitted by distinct orthogonal subcarriers. The orthogonality between those transmitted sub-streams is ensured by choosing a proper subcarrier spacing. By inserting a cyclic prefix in the (parallel-to-serial) combined time-domain signal before it is transmitted via the main carrier, inter-symbol interference (ISI) caused by a multipath channel can be eliminated at the receiving end if the corresponding cyclic prefix part in each OFDM frame is removed.

The cyclic prefix (also known as the guard interval) of an OFDM block is a copy of the last portion of the original block and if the maximum channel path delay is smaller than the guard interval duration, the received time-domain waveform will be a weighted sum of various delayed version of the transmitted blocks. Deleting the cyclic prefix, each data-bearing subcarrier experiences only flat fading. Such an arrangement enables an OFDM-based system to overcome frequency selective fading in broadband wireless transmission and is the major reason for its current popularity.

With all its merits, OFDM, unfortunately, is far more sensitive to synchronization errors, especially the carrier frequency offset (CFO), than single carrier systems. CFO is primarily caused by relative movement or channel-induced Doppler shifts and instabilities of and mismatch between transmitter and receiver oscillators [1]. In high dynamic environment, the received waveform is likely to encounter large Doppler shifts by antipodal frequencies. Residual CFO results in inter-carrier interference (ICI) among subcarriers due to the loss of orthogonality and will bring about serious performance degradation. Whence it is highly desirable to reduce the sensitivity of OFDM systems to CFO of any kind.

There have been a multitude of proposals for CFO compensation. But most of them require pilot symbols or training sequences [2], which lowers the effective data rate. Blind techniques that operate in the time domain [3] often require observation over a number of OFDM symbols, and are not suitable for tracking time-varying CFO. Decision-directed modifications of some pilot-assisted CFO estimation are more promising but require good initial estimate. On the other hand, the ultimate goal of a receiver is the detection of the transmitted data stream and, as the detection of OFDM signals takes place in the frequency domain, channel estimation must be obtained before data detection commences. Conventional OFDM channel estimators, however, are derived under a zero or negligible CFO assumption. We therefore face the dilemma of channel estimation in the presence of CFO and CFO estimation without pilots or reliable decisions.

It is thus the goal of this investigation to develop feasible solutions to the above joint estimation and detection problem. We adopt an iterative approach for joint CFO/channel estimation, tracking and data detection. To simply our problem, we shall assume that the initial CFO is to be less than one half of the subcarrier spacing. Drawing an analogy between the ICI effect on a frequency domain signal and the ISI effect on a time domain signal, we can easily apply the maximum likelihood sequence estimation (MLSE) technique to detect the data stream in the presence of ICI, assuming known CFO**.** Finally We extends our investigation to a MIMO environment.

2. Joint CFO Estimation and Data Detection for SISO-OFDM

In a real-world scenario, received OFDM

signals often suffer from residual frequency offsets, caused by oscillator instabilities and relative movement-induced Doppler spread, resulting in inter-carrier interference (ICI) among subcarriers. It is desirable to reduce the sensitivity of OFDM to carrier frequency offsets.

Shown in Fig. 1. is a block diagram of an OFDM transmitter. The mapper converts input data stream into complex valued constellation points, according to the selected constellation, e.g., BPSK, QAM. The complex valued data sequence is parallel-to-serial converted before being used to modulate multiple carriers via an inverse discrete Fourier transformation (usually through the inverse fast Fourier transform, IFFT). Each OFDM symbol has a "useful interval" of length T_u and a cyclic prefix (CP)(or "guard interval") with duration T_g; see Fig. 2. The CP is a periodical repetition of the useful interval so that precise timing recovery is not needed as long as the discrete Fourier transform frame of the receiver starts within the guard interval. In order to protect the OFDM signal against inter-symbol interference (ISI), the CP is set longer than the maximum channel memory, T_g / T_u <0.25$ being a practical value. The baseband signal is modulated by a Nyquist signaling pulse, up-converted to the radio frequency (RF) and then transmitted through the channel.

At the receiver, the signal is down-converted to the intermediate frequency (IF) and then demodulated (Fig. 3). After removing the cyclic prefix, the signal part of the received time-domain baseband waveform can be expressed as

$$s[n]=\frac{1}{N}\sum_{k=0}^{N-1}a_k e^{j2\pi nk/N}e^{j2\pi f_e n/N} \tag{1}$$

where a_k represents the symbol carried by the kth subcarrier. The i.i.d data sequence $\{a_k\}$ is such that $E[a_k]=0$ and $E[|a_k|^2]=1$, N=the number of subcarriers, and f_e is CFO normalized by the intercarrier spacing $f_t=1/T_u$ and T_u is an OFDM symbol period (without cyclic prefix).

This time domain signal is then passed through a serial-to-parallel converter and then an FFT demodulator. Assuming an AWGN channel, the signal at the FFT output becomes

$$r_k=\frac{1}{N}\sum_{n=0}^{N-1}s[n]w[n]e^{-j2\pi nk/N}+n_k \tag{2}$$

where $w[n]$ represents the receiver pulse shaping function and $n[k]$ is additive white Gaussian noise (AWGN). The received signal can be written as

$$\begin{aligned} r_k &= \sum_{m=0}^{N-1}a_m\left\{\frac{1}{N}\sum_{n=0}^{N-1}w_n e^{-j2\pi n(k-m)/N}e^{j2\pi f_e n/N}\right\}+n_k \\ &=\sum_m a_m W_{k-m}^{f_e}+n_k=a_k W_0^{f_e}+\underbrace{\sum_{m\neq k}a_m W_{k-m}^{f_e}}_{ICI}+n_k \\ &=y_k(f_e,a)+n_k \end{aligned} \tag{3}$$

where $W_k^{f_e}=\frac{1}{N}\sum_{n=0}^{N-1}w[n]e^{-j2\pi nk/N}e^{j2\pi f_e n/N}$

As shown in (3), the signal portion of r_k, can be viewed as the output of an ISI channel driven by the input **a** with the channel response $W_k^{f_e}$, dependent upon the given CFO, f_e. The choice of the window function $w[n]$ affects the extent of ISI (actually the ICI in this case). To control the complexity of the estimator, the length of $W_k^{f_e}$ must be kept reasonably short.

(i) Data detection:

With a proper window shaping, we can rewrite (3) as

$$r_k \approx a_k W_0^{f_e} + \sum_{m=k-L, m\neq k}^{m=k+L} a_m W_{k-m}^{f_e} + n_k \qquad (4)$$

where L is usually less or equal than 2 for $|f_e| < 0.5$.

Given $\hat{f}_e$, the data detection is equivalent to estimating the state of a discrete-time finite-state machine. The finite-state machine in this case is the equivalent discrete-time channel with coefficients $\{W_k^{\hat{f}_e}\}$ and its state at any instant in time is given by the *2L* most recent inputs. Even though we have noncausal term in (4), we can add

a delay function so that the resulting model is causal.

$$r_{K-L} \approx a_k W_{-L}^{\hat{f}_e} + a_{k-1} W_{-L+1}^{\hat{f}_e} + \cdots + a_{k-2L} W_L^{\hat{f}_e} + n_{k-L} \qquad (5)$$

The state at time k is

$$S_k = (a_{k-1}, a_{k-2}, \cdots, a_{k-2L}) \qquad (6)$$

If the information symbols are M-ary, the channel filter has M^{2L} states. Consequently, the channel is described by an M^{2L}-state trellis and the Viterbi algorithm may be used to determine the most probable path through the trellis.

(ii) CFO estimation:

Given $\hat{a}$ as well as **r**, we can get a CFO estimate by finding f_e that minimizes the cost function

$$C(f_e) = \sum_k \left| r_k - y_k(f_e, \hat{a}) \right|^2 \qquad (7)$$

The minimization of (7) is carried out base on a gradient descent search method as following:

$$\begin{aligned} \nabla_{f_e}(C(f_e)) &= \frac{\partial C}{\partial f_e} \\ &= \sum_k \{-A \cdot r_k^* - A^* r_k + BA^* + AB^*\} \end{aligned} \qquad (8)$$

where $()^*$ denotes the complex conjugate and

$$\begin{aligned} A &= \sum_m a_m \left(W_{k-m}^{f_e}\right)' \approx \sum_{m=k-L}^{k+L} a_m \left(W_{k-m}^{f_e}\right)' \\ B &= \sum_m a_m \left(W_{k-m}^{f_e}\right) \approx \sum_{m=k-L}^{k+L} a_m \left(W_{k-m}^{f_e}\right) \\ \left(W_{k-m}^{f_e}\right)' &= \frac{\partial W_{k-m}^{f_e}}{\partial f_e} \end{aligned} \qquad (9)$$

A *learning rate u* is introduced to control the rate of change in each iteration

$$\begin{aligned} f_e^1 &\leftarrow f_e^0 - u * \nabla_{f_e}\left(C\left(f_e^0\right)\right) \\ f_e^2 &\leftarrow f_e^1 - u * \nabla_{f_e}\left(C\left(f_e^1\right)\right) \\ &\vdots \\ f_e^m &\leftarrow f_e^{m-1} - u * \nabla_{f_e}\left(C\left(f_e^{m-1}\right)\right) \end{aligned}$$

When f_e^m reaches a stable value with very little jitter, which indicates that we have arrived at a minimum, the algorithm

stops; see Fig. 4.

The joint frequency-data sequence (FD) estimation algorithm consists of the following steps:

A.1 (Frequency Estimate Initialization) Set the initial estimate $\hat{f}_e$ to zero.

A.2 (Starting the FD iteration loop) Using the CFO estimate $\hat{f}_e$, generate the corresponding frequency domain "channel response" $W_k^{f_e}$:

$$W_k^{f_e} = \frac{1}{K} \sum_{n=0}^{K-1} w[n] e^{-j2\pi n(k-m)} e^{j2\pi f_e n/K}$$

A.3 (Compute the tentative data estimate) Based on the observation baseband samples r, we run the Viterbi algorithm using the branch metric

$$\lambda_k = \left| r_k - y_k\left(\hat{f}_e, a\right) \right|^2$$

to obtain an estimate $\hat{a}$ of the transmitted data sequence.

A.4 (Update the frequency estimate) Given $\hat{a}$ from Step 3 as well as **r**, we obtain a new CFO estimate by searching for f_e that minimizes the cost function

$$C = \sum_k \left| r_k - y_k\left(f_e, \hat{a}\right) \right|^2$$

A.5 (Convergence check and end of the FD loop) If the new CFO estimate is close enough to the previous estimate, i.e. absolute difference is less than 10^{-6}, then stop the iteration and release the most recent estimate; otherwise, go to Step 2.

Simulation results show that this algorithm usually converges within just two or three iterations.

3. Joint Estimation for Frequency-Selective Channels

(i) System model for frequency-selective channels

For frequency-selective fading channels, we assume that the CP is longer than the channel impulse response whence there is no frequency domain self interference between successive OFDM symbols. With such an assumption we can concentrate on a single OFDM symbol scenario. Let the signal part of the received signal be given by

$$s[n] = \frac{1}{N}\sum_{k=0}^{N-1} a_k H_k e^{j2\pi nk/N} e^{j2\pi f_e n/N} \tag{10}$$

where H_k is the channel response at k/T. After taking FFT, the received signal becomes

$$r_k = \frac{1}{N}\sum_{n=0}^{N-1} s[n] w[n] e^{-j2\pi nk/N} + n_k \tag{11}$$

like the AWGN case, we have

$$\begin{aligned} r_k &= \sum_{m=0}^{N-1} a_m H_m W_{k-m}^{f_e} + n_k \\ &= y_k\left(f_e, \mathrm{a}, \mathrm{H}\right) + n_k \end{aligned} \tag{12}$$

(ii) Joint frequency-data sequence-channel (FDC) estimation algorithm

The estimator proposed in the previous section cannot be applied directly to (12) because the channel response is unknown. However, assuming the data sequence and the CFO are known, the channel impulse response can be easily estimated.

For every subcarrier n, the channel can be estimated as (using LS):

$$\hat{H}_n = \frac{1}{a_n}\left[\left(\mathrm{F}^H \mathrm{F}\right)^{-1} \mathrm{F}^H \mathrm{r}\right]_n, n = 0,1,\ldots,N-1 \tag{13}$$

where

$$\mathrm{F} = \begin{bmatrix} W_0^{f_e} & \cdots & W_{-N+1}^{f_e} \\ \vdots & \ddots & \vdots \\ W_{N-1}^{f_e} & \cdots & W_0^{f_e} \end{bmatrix}$$

$$\mathrm{r} = \begin{bmatrix} r_0 \\ r_1 \\ \vdots \\ r_{N-1} \end{bmatrix}$$

Based on this channel estimator and the joint CFO-data sequence estimator proposed in the previous section, we arrive at the following iterative joint CFO-data sequence-channel estimation algorithm.

B.1 (Channel estimate initialization) Let $\hat{H}_o$ to be the channel estimate from the previous OFDM symbol interval.

B.2 (FD iteration loop) Using this channel estimate, the CFO f_e and the symbol sequence **a** are estimated by the joint FD estimation algorithm described in the previous section.

B.3 (Channel estimate update) Compute a new channel estimate $\hat{H}_N$ using (13) based on the current estimates $\hat{f}_e$ and $\hat{a}$.

B.4 (Convergence check for the channel estimate) If the new channel estimate $\hat{H}_N$ is close enough to the previous estimate, more specifically, if

$$\left|\frac{\hat{H}_N - \hat{H}_o}{\hat{H}_N}\right|^2 < 10^{-2}$$

then stop iteration and release the most recent estimates. If not, go to Step 2.

We plot the overall receiver structure in Fig. 5.

(iii) Numerical Experiments

The computer simulation results reported in this section are obtained by using a pilot and data format which are the same as the Tgn Sync 802.11n proposal (52 data tones and 4 pilot tones). It is a slight modification of the IEEE 802.11a data format (48 data tones and 4 pilot tones).

From Fig. 6 we can see that the proposed joint FD estimator gives good BER performance in AWGN channels.

The mean squared estimation error (MSEE) performance of the frequency estimate for different f_e is plotted in Fig. 7, where SNR is defined as the ratio of the total transmit signal energy per bit to noise power density.

The frequency-selective fading channel is modeled as an linear FIR filter with impulse response given by

$$h(k) = \sum_{n=0}^{C_L-1} a_n e^{-j\Phi}\delta(k-n) \tag{14}$$

where Φ is uniformly distributed in $(0,2\pi]$ and α_n is Rayleigh distributed with an exponential power profile

$$\overline{\alpha}_n^2 = \left(1-e^{-T_s/T_{rms}}\right)e^{-nT_s/T_{rms}}$$

with C_L=10, T_{rms}=30ns and T_s=50ns. We use Jakes channel model with maximum Doppler shift of 500 Hz to simulate time-correlated Rayleigh fading α_n.

Fig. 8 and Fig. 9 present respectively the BER and MSEE performance of the CFO estimator in frequency-selective fading channels. And we show the MSEE of the channel estimator in Fig. 10. These performance curves indicate that the BER performance is dominated by the channel estimation error since the MSEE of the CFO estimator is relatively small, especially at high SNRs where the MSEE of our frequency estimate reaches an almost constant floor while the channel estimate's MSEE continues to decrease.

4. Joint Estimation for MIMO-OFDM Systems

(i) System Model

Consider a MIMO system with M_T transmit antennas and M_R receive antennas. Assuming a flat block-fading channel model, we can express the equivalent signal part of the received time-domain samples at qth receive antenna as

$$b^q[n] = \sum_{p=1}^{M_T} \tilde{s}^{p,q}[n], q = 1,2,\cdots,M_R \tag{15}$$

where

$$\tilde{s}^{p,q}[n] = \frac{1}{N}\sum_{k=0}^{N-1}\sqrt{\frac{E_s}{M_T}}a_k^p H_k^{q,p} e^{j2\pi nk/N} e^{j2\pi f_e n/N} \tag{16}$$

corresponds to the OFDM signal sent from the pth transmit antenna to the qth receive antenna. where

- a_k^p represents the symbol carried by the kth subcarrier at the pth transmit antenna,
- $H_k^{q,p}$ is the kth subcarrier's channel transfer function between the qth receive antenna and the pth transmit antenna,

. E_s is the average energy allocated to the kth subcarrier across the transmit antennas,

. f_e is the CFO normalized by the inter-carrier spacing, and

. N is the total number of sub-carriers for each transmit antenna.

After establishing proper frame (symbol) timing, removing the cyclic prefix and taking FFT on the time-domain samples, the received baseband frequency domain signal at the qth RX antenna reads

$$
\begin{aligned}
r_k^q &= \sum_{n=0}^{N-1} \tilde{b}^q[n] e^{-j2\pi nk/N} + n_k^q \\
&= \sum_{p=1}^{M_T} \sum_{m=0}^{N-1} a_m^p H_m^{q,p} W_{k-m_e}^{f_e} + n_k^q \qquad (17) \\
&= \sum_{p=1}^{M_T} y_k\left(f_e, a^p, H^{q,p}\right) + n_k^q, q = 1,2,\cdots,M_R
\end{aligned}
$$

where $\tilde{b}^q[n]$ is the "folding window signal" from all transmit antennas.

Fig. 11 plots the transmission channel model for the qth receive antenna with respect to M_T transmit antennas. Comparing the above equation with (12), we find that to apply the joint FDC method to solve (17), we will have a trellis with the number of trellis states equals to $M^{2L*M}{}_T$ since we now have to consider all transmitted data streams from all transmit antennas.

(ii) MIMO-OFDM Channel Estimation

Rewriting (17) in matrix form (for qth RX antenna)

$$
\begin{bmatrix} r_0^q \\ r_0^q \\ \vdots \\ r_{N-1}^q \end{bmatrix} = \begin{bmatrix} W_0^{f_e} & \cdots & W_{-N+1}^{f_e} \\ \vdots & \ddots & \vdots \\ W_{N-1}^{f_e} & \cdots & W_0^{f_e} \end{bmatrix} \begin{bmatrix} \sum_{p=1}^{M_T} H_0^{q,p} a_0^p \\ \sum_{p=1}^{M_T} H_1^{q,p} a_1^p \\ \vdots \\ \sum_{p=1}^{M_T} H_{N-1}^{q,p} a_{N-1}^p \end{bmatrix} + \begin{bmatrix} n_0^q \\ n_1^q \\ \vdots \\ n_{N-1}^q \end{bmatrix} \qquad (18)
$$

and invoking the substitutions,

$$\mathbf{r}^q = \begin{bmatrix} r_0^q & r_0^q & \cdots & r_{N-1}^q \end{bmatrix}^T \qquad (19)$$

$$\mathbf{n} = \begin{bmatrix} n_0^q & n_0^q & \cdots & n_{N-1}^q \end{bmatrix}^T \qquad (20)$$

$$
\begin{aligned}
\mathbf{H_A^q} &= \begin{bmatrix} (H_A^q)_0 & (H_A^q)_1 & \cdots & (H_A^q)_{N-1} \end{bmatrix}^T \\
&= \begin{bmatrix} \sum_{p=1}^{M_T} H_0^{q,p} a_0^p & \sum_{p=1}^{M_T} H_1^{q,p} a_1^p & \cdots & \sum_{p=1}^{M_T} H_{N-1}^{q,p} a_{N-1}^p \end{bmatrix}^T
\end{aligned}
$$

$$
\mathbf{F} = \begin{bmatrix} W_0^{f_e} & \cdots & W_{-N+1}^{f_e} \\ \vdots & \ddots & \vdots \\ W_{N-1}^{f_e} & \cdots & W_0^{f_e} \end{bmatrix}
$$

We obtain

$$\mathbf{r^q} = \mathbf{F H_A^q} + \mathbf{n} \qquad (21)$$

Assuming the data sequence and the CFO are known, we first compensate for the CFO through the LS estimate

$$
\begin{aligned}
\mathbf{H_A^q} &= \begin{bmatrix} (\hat{H}_A^q)_0 & (\hat{H}_A^q)_1 & \cdots & (\hat{H}_A^q)_{N-1} \end{bmatrix}^T \\
&= \mathbf{(F^H F)^{-1} F^H r^q}
\end{aligned} \qquad (22)
$$

The kth subcarrier over M_R receive antennas reads

$$\left(\hat{\mathbf{H}}_A\right)_{k,t} = \begin{bmatrix} \left(\hat{\mathbf{H}}_A^1\right)_{k,t} & \left(\hat{\mathbf{H}}_A^2\right)_{k,t} & \cdots & \left(\hat{\mathbf{H}}_A^{M_R}\right)_{k,t} \end{bmatrix}^T$$

$$\mathbf{A_{k,t}} = \begin{bmatrix} a_{k,t}^1 & a_{k,t}^2 & \cdots & a_{k,t}^{M_T} \end{bmatrix}^T$$

where the index $1 \le t \le M$ denotes the tth OFDM symbol of an M-symbol block that suffer from the same flat (block-)fading.

Stacking up the M received frequency domain symbols we have

$$\overline{\left(\hat{\mathrm{H}}_{\mathrm{A}}\right)_k} = \hat{\mathrm{H}}_k \overline{\mathrm{A}_k} \qquad (23)$$

where

$$
\begin{aligned}
\overline{\left(\hat{\mathrm{H}}_{\mathrm{A}}\right)_k} &= \begin{bmatrix} \left(\hat{\mathrm{H}}_{\mathrm{A}}\right)_{k,1} & \left(\hat{\mathrm{H}}_{\mathrm{A}}\right)_{k,2} & \cdots & \left(\hat{\mathrm{H}}_{\mathrm{A}}\right)_{k,M} \end{bmatrix} \\
\overline{\mathrm{A}_k} &= \begin{bmatrix} \mathrm{A}_{k,1} & \mathrm{A}_{k,2} & \cdots & \mathrm{A}_{k,M} \end{bmatrix}
\end{aligned} \qquad (24)
$$

and $\hat{\mathbf{H}}_k$ is the M_RxM_T complex channel matrix for the kth subcarrier. Finally, the channel matrix can be estimated by

$$\hat{\mathrm{H}}_k = \overline{\left(\hat{\mathrm{H}}_A\right)_k}\,\overline{\mathrm{A}}_k^H \left(\overline{\mathrm{A}}_k \overline{\mathrm{A}}_k^H\right)^{-1} \tag{25}$$

on the premise that the rank of $\overline{\mathbf{A}}_k$, the "over symbols data matrix at subcarrier k" is greater than M_T. In other words, one has to sent at least M_T OFDM symbols over the same flat fading block (a coherent time) in order to obtain a proper channel estimate [4].

(iii) An Iterative Joint FDC Estimation Algorithm

Based on the above discussion, we suggest an iterative joint FDC estimation algorithm as follows.

C.1 (Channel estimate initialization) Set $\hat{H}_o$ to be the channel estimate from the previous OFDM symbols.

C.2 (Frequency Estimate Initialization) Set the initial estimate $\hat{f}_e = 0$.

C.3 (Starting the FD iteration loop) Use current channel estimate for the frequency-data sequence iteration loop. Calculate the channel frequency response compensation matrix $F = \left[W_k^{f_e}\right]$ and current frequency estimate $\hat{f}_e$.

C.4 (Compute the tentative data estimate) Given the observed sample vector **r** defined by (4.5), perform the Viterbi algorithm with the branch metric

$$\lambda_k^q = \left| r_k^q - \sum_{p=1}^{MT} y_k\left(\hat{f}_e, a^p, \hat{H}^{p,q}\right) \right|^2$$

to find a conditional optimal data sequence estimate $\hat{a}$.

C.5 (Update the frequency estimate) Update the CFO estimate by minimizing the cost function :

$$C = \sum_q \sum_k \left| r_k^q - \sum_{p=1}^{MT} y_k\left(\hat{f}_e, a^p, \hat{H}^{p,q}\right) \right|^2$$

C.6 (Convergence check and end of the FD loop) If the new CFO estimate $\hat{f}_e$ is close to the previous estimate, i.e. absolute difference is less than 10^{-6}, then stop the current FD iteration and release the new estimates $\hat{a}$ and $\hat{f}_e$; otherwise go back to Step 3.

C.7 (Channel estimate update) Compute a new channel estimate $\hat{H}_N$ using (25) based on the current estimates $\hat{a}$ and $\hat{f}_e$.

C.8 (Convergence check for the channel estimate) If the new channel estimate $\hat{H}_N$ is close enough to the previous estimate, more specifically, if

$$\left| \frac{\hat{H}_N - \hat{H}_o}{\hat{H}_N} \right|^2 < 10^{-2}$$

then stop iteration and release the most recent estimates. Otherwise, go to Step 3.

(iv) Numerical Experiments

The simulation parameters and environments used in the following paragraphs are identical to those used in the SISO-OFDM system simulations, except that we now have M_T **x** M_R independent channels. The BER performance of the iterative joint FDC estimation algorithm for the 2x2 configuration is given in Fig. 12. For comparison purpose the SISO performance curve is shown in the same figure. Clearly, the spatial diversity does pay off--with a gain of more than 4 dB for BER<10^{-2}. The MSEE performance of the corresponding frequency estimate is plotted in Fig. 13. Similarly, we find that more receive antennas leads to better performance due to the MIMO diversity gain.

Fig. 14 shows the effect of the channel estimate on the BER performance. These curves indicate that, as expected, the model-based channel estimation still outperform the LS channel estimate for MIMO-OFDM systems.

5. Conclusion

We have presented iterative joint estimation and detection algorithms for OFDM systems in various scenarios. An iterative joint frequency estimation and data sequence detection algorithm is first given for AWGN channels. By incorporating a proper channel estimation method, we then provide an iterative joint FDC estimation algorithm for SISO-OFDM systems. Some refinements for enhanced performance and complexity reduction are suggested. Finally, we extend the our joint FDC estimation algorithm to MIMO-OFDM environments. Simulation based on a candidate IEEE 802.11n standard--the Tgn Sync (Task Group *n*) [10] proposal--is provided to validate various joint estimation and detection algorithms. The corresponding numerical results do prove that the proposed solutions are both practical under the assumed channel conditions. Our solution can operate in the blind mode if the initial CFO is less than half of a subcarrier spacing. Otherwise, we can use the preamble (like the 802.11n case) to reduce the CFO to within the desired range. Our proposal can then be used for subsequent data detection and CFO/channel tracking in time-varying channels.

6. Reference

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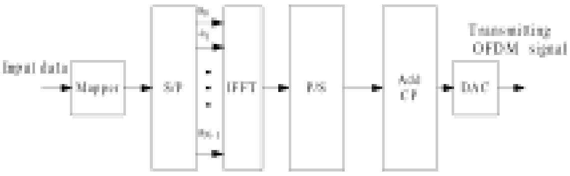

Figure 1：Block diagram of an OFDM modulator.

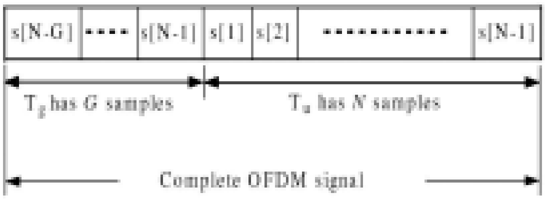

Figure 2：Structure of complete OFDM signal with the the guard period

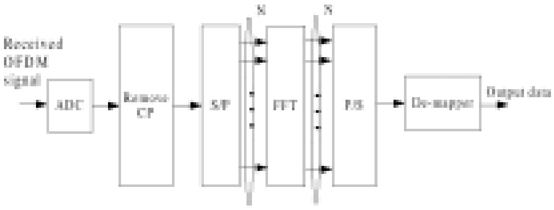

Figure 3：Block diagram of a typical OFDM demodulator

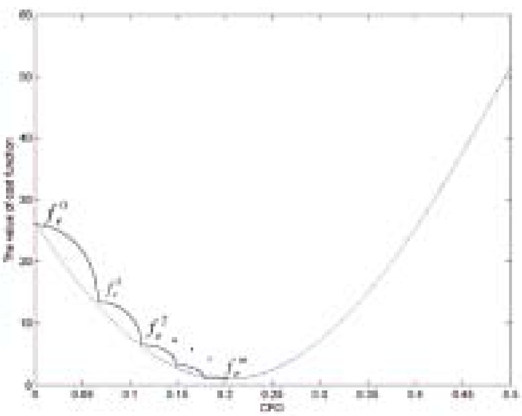

Figure 4：An example of cost value for CFO=0.2 without noise

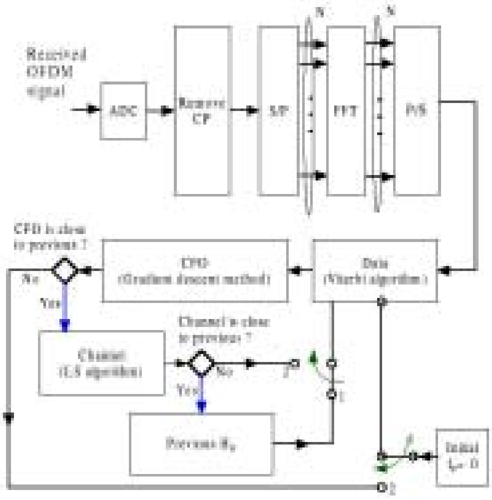

Fig. 5: Illustrating the proposed iterative joint Frequency-Data Sequence-Channel (FDC) estimation algorithm". Initially, the switch is in position 1; and for the tracking mode it is moved to position 2; It stops when the new channel estimation is close enough to the previous estimate .

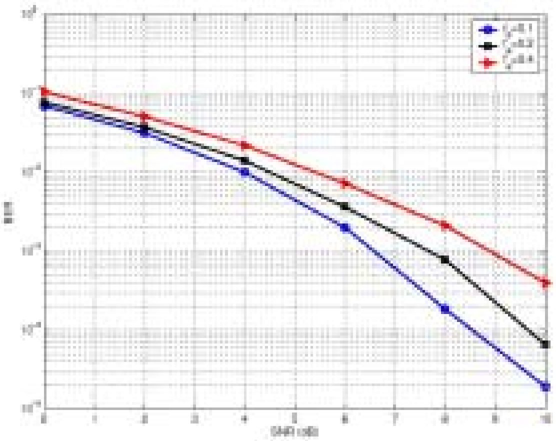

Fig. 6: BER performance of SISO-OFDM systems using iterative joint FD estimate in AWGN channels with different f_e.

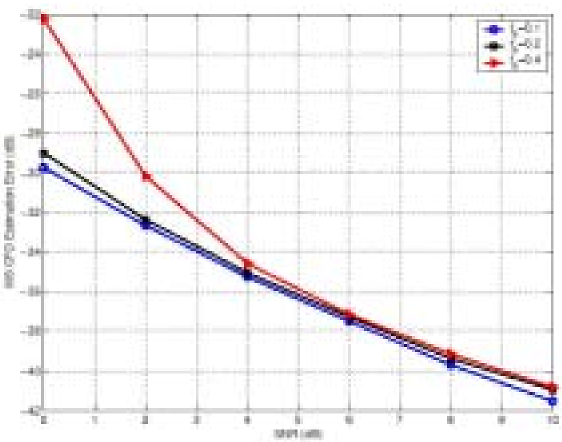

Fig. 7: SEE performance of CFO estimates for the joint FD estimate in AWGN channels.

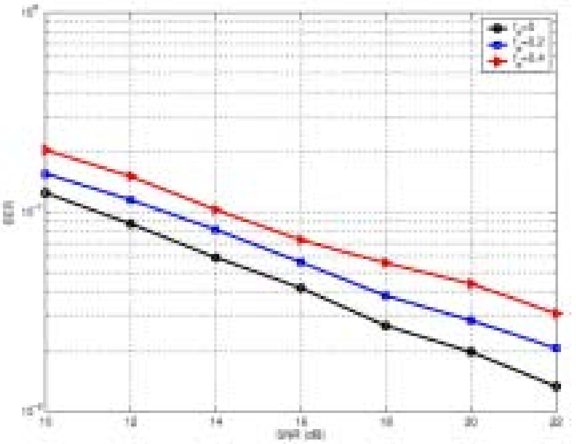

Fig. 8 BER performance of SISO-OFDM systems using iterative joint FDC estimate in Rayleigh fading channels with different f_e.

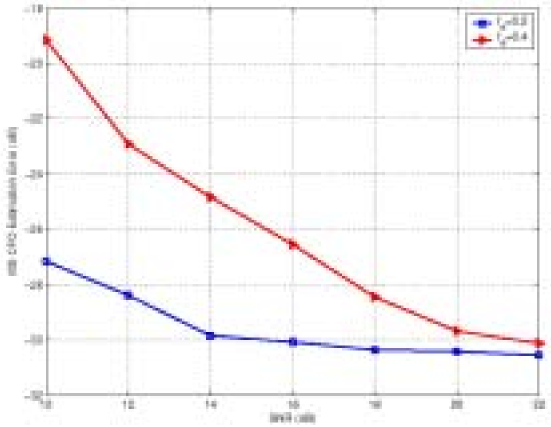

Fig. 9: MSEE performance of CFO estimates for the joint FDC estimate in Rayleigh fading channels.

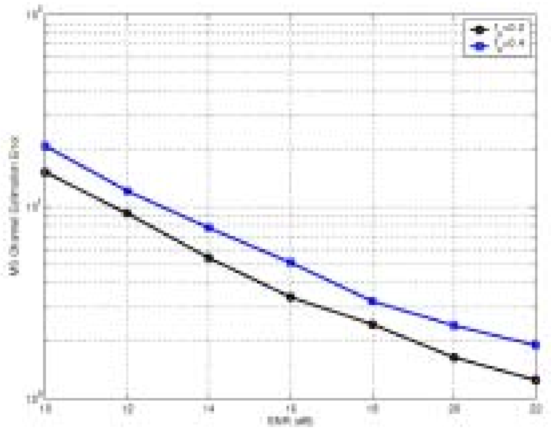

Fig. 10: MS channel estimation error of the joint FDC estimate in a Rayleigh fading as a function of f_e.

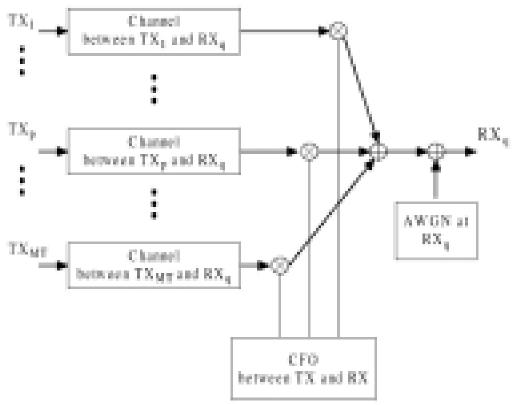

Fig. 11: Channel model for *q*th receive antenna.

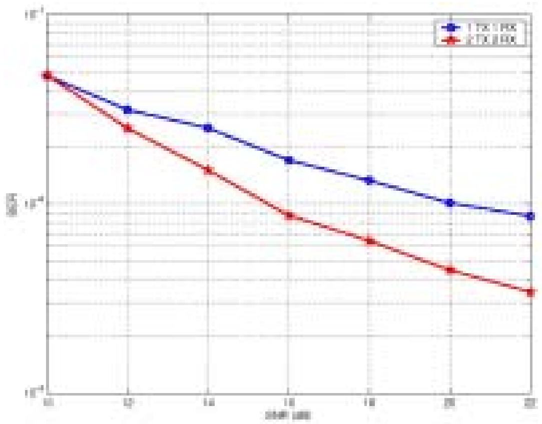

Fig.. 12: BER performance of MIMO- and SISO-OFDM systems using iterative joint FDC estimate (f_e =0.2).

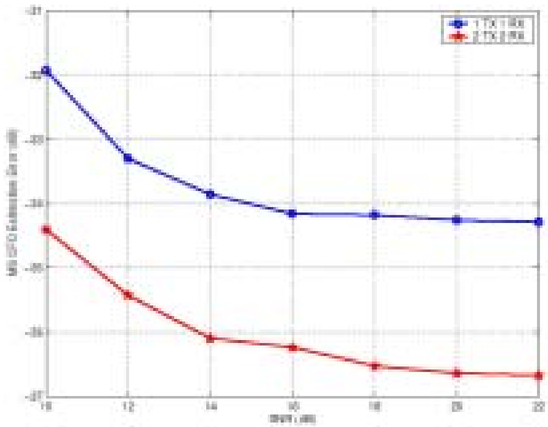

Fig. 13: MSEE performance of CFO estimates in MIMO-OFDM systems for 2 x2 and SISO configurations (f_e =0.2).

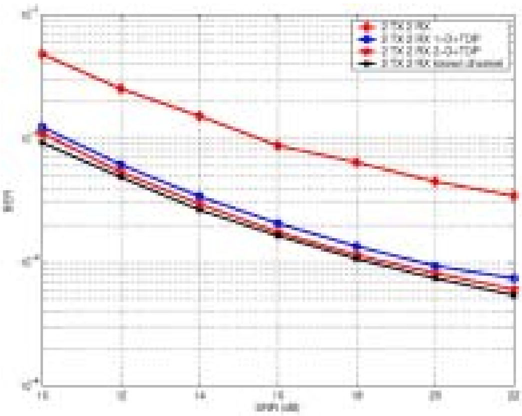

Fig. 14: Effect of the channel estimate on the MIMO-OFDM system's BER performance (f_e =0.2).