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New Preamble Structures for Synchronization and Cell searching in OFDM systems

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New Preamble Structures for Synchronization and Cell searching in OFDM systems

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Abstract

This thesis proposes two new preamble structures for synchronization and cell searching in cellular based Orthogonal Frequency Division Multiplexing (OFDM) systems. The proposed preamble structures are to impose Pseudo Noise sequences (PN-sequences) on the OFDM transmitted signal in time-domain for time and frequency synchronization. This scheme make it possible to achieve a good performance at a low SNR, therefore, it can guarantee a reliable communication to a mobile located at the cell edge. Herein, the PN-sequences occupies the whole or part of the preamble according to the two different preamble structures. The proposed preamble structures are compared each other in terms of frame and frequency synchronization, cell searching and mean acquisition time.

The simulation results elucidate that the proposed preamble structures are appropriate for cellular based OFDM systems.
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List of Abbreviations

AWGN  Additive White Gaussian Noise
BER   Bit Error Rate
BS    Base Station
DFT   Discrete Fourier Transform
FDM   Frequency Division Multiplexing
FFT   Fast Fourier Transform
ICI   Inter Carrier Interference
ISI   Inter Symbol Interference
ITU   International Telecommunication Union
OFDM  Orthogonal Frequency Division Multiplexing
PN    Pseudo Noise
PS    Preamble Structure
RMS   Root Mean Square
SISO  Single Input Single Output
SNR   Signal to Noise Ratio
I. Introduction

Orthogonal frequency-division multiplexing (OFDM) systems have recently gained increased interest. OFDM is used in the European digital broadcast radio system and is being investigated for other wireless applications such as digital broadcast television and mobile communication systems, as well as for broadband digital communication on existing copper network. OFDM systems have several advantages, firstly, its robustness against inter-symbol interference (ISI) and mitigating the effect of multi-path delay spread in wireless radio channel make OFDM systems to be one of the prospecting technique for the next generation. Secondly, OFDM systems divide the available spectrum into several sub-channels, so the data loaded at each sub-channel can be transmitted over flat fading channel. Thirdly all sub-channels are narrow-band and they experience flat fading, which makes equalization very simple. Finally the spectrum of the sub-channels are overlapping under interference is significantly reduced. It enhances the bandwidth efficiency.

However, it requires more precise synchronization than a single carrier system, therefore, the synchronization is one of the most important issue for OFDM systems. For example, carrier frequency offsets, due to the mismatch of oscillators between the transmitter and receiver, can cause Inter-Carrier Interference (ICI) [1]-[2], and also frame synchronization errors can cause Inter-Symbol Interference (ISI) [1].

Additionally, the technique which can discriminate the cell is vital in cellular systems. Moreover, in cellular systems, the mobiles which are located at the cell edge will be interfered from neighboring cells. For the communication in OFDM based cellular systems, we should guarantee a reliable communication at the cell edge, specifically we consider
synchronization and cell searching in this thesis.

1.1 OFDM systems

Orthogonal frequency division multiplexing (OFDM) is a parallel transmission scheme, where a high rate serial data stream is split up into a set of low rate substreams, each of which is modulated on a separate subcarrier. The concept of using parallel data transmission by means of frequency division multiplexing (FDM) was published in mid 60s. The idea was to use parallel data streams and FDM with overlapping subchannels to avoid the use of high speed equalization and to combat impulsive noise, and multipath distortion as well as to fully use the available bandwidth. Weinstein and Ebert applied the discrete Fourier transform (DFT) to parallel data transmission system as part of the modulation and demodulation process. In addition to eliminating the banks of subcarrier oscillators and coherent demodulators required by FDM, a completely digital implementation could be built around special-purpose hardware performing the fast Fourier transform (FFT).

In a conventional serial data system, the symbols are transmitted sequentially, with the frequency spectrum of each data symbol allowed to occupy the entire available bandwidth. But in a parallel data transmission system several symbols are transmitted at the same time, what offers possibilities for alleviating many of the problems encountered with serial systems.

In OFDM, the data is divided among large number of subcarriers. And the data which is loaded on each subcarrier are transmitted in a parallel way instead of serial way. Because only a small amount of the data is carried on each carrier, and by this lowering of the bit rate per carrier, the influence of intersymbol interference is significantly reduced.

And also the OFDM divides an entire channel bandwidth into many
narrow subbands, the frequency response over each individual subband is relatively flat. Since each subchannel covers only a small fraction of the original bandwidth, equalization is potentially simpler than in a serial data system.

OFDM can be simply defined as a form of multicarrier modulation where its carrier spacing is carefully selected so that each subcarrier is orthogonal to the other subcarriers. As is well known, orthogonal signals can be separated at the receiver by correlation techniques; hence, intersymbol interference among channels can be eliminated. Orthogonality can be achieved by carefully selecting carrier spacing, such as letting the carrier spacing be equal to the reciprocal of the useful symbol period. In result, it can enhance the bandwidth efficiency. Fig. 1.1 shows the spectrum of OFDM.

![OFDM spectrum](image)

Figure 1.1: OFDM spectrum

The orthogonality of subchannels in OFDM can be maintained and individual subchannels can be completely separated by the FFT at the receiver when there are no intersymbol interference (ISI) and intercarrier
interference (ICI) introduced by transmission channel distortion.

One way to prevent ISI is to create a cyclically extended guard interval Fig. 1.2, where each OFDM symbol is preceded by a periodic extension of the signal itself. The total symbol duration is $T_{total} = T_g + T_b$, where $T_g$ is the guard interval and $T_b$ is the useful symbol duration. When the guard interval is longer than the channel impulse response or the multipath delay, the ISI can be eliminated. The ratio of the guard interval to useful symbol duration is application dependent. Since the insertion of guard interval will reduce data throughput, $T_g$ is usually less than $T_b/4$. There are two reasons to use a cyclic prefix for the guard interval. The one is to maintain the receiver carrier synchronization. The other is that a cyclic convolution can still be applied between the OFDM signal and the channel response to model the transmission system.

![Figure 1.2: Cyclic prefix in OFDM symbol](image)

OFDM systems use discrete Fourier transform (DFT) to modulate and demodulate parallel data. The individual spectra are now sinc functions and are not band limited. The FDM is achieved, not by bandpass filtering, but by baseband processing. Using this method, both transmitter and receiver can be implemented using efficient FFT
techniques that reduce the number of operations from $N^2$ in DFT, down to $N \log N$.

### 1.2 Synchronization in OFDM systems

#### 1.2.1 Previous Work

In OFDM systems, a lot of studies has been done for achieving time and frequency synchronization. There are two main technologies, first is using the cyclic prefix [2]-[6] and the latter is using special OFDM training symbols[7]-[8].

Because of the cyclic prefix, the first $T_g$ seconds part of each OFDM symbol is identical to the last part. This property can be exploited for both time and frequency synchronization by using a synchronization system like depicted in Fig. 1.3. Basically, this device correlates a $T_g$ long part of the signal with a part that is $T_b$ seconds delayed in Fig. 1.2.

![Figure 1.3: Synchronization using the cyclic prefix](image)

Once symbol timing is known, the cyclic extension correlation output can be used to estimate the frequency offset. The phase of the correlation output is equal to the phase drift between samples that are $T$ second apart. Hence, the frequency offset can simply be found as the correlation phase divided by $2\pi T$. This method works up to a maximum absolute frequency offset of half the subcarrier spacing. To increase this
maximum range, shorter symbols can be used, or special training symbol with different PN sequences on odd and even subcarrier frequencies to identify a frequency offset of an integer number of subcarrier spacing.

For packet transmission, it is required more accurate synchronization and for high rate packet transmission, the synchronization time needs to be as short as possible. To achieve this, special OFDM training symbol can be used for which the data content is known to the receiver. In this way, the entire received training signal can be used to achieve synchronization, whereas the cyclic extension method only uses a fraction of each symbol.

![Synchronization using a OFDM training symbol](image)

**Figure 1.4: Synchronization using a OFDM training symbol**

Fig. 1.4 shows a block diagram of a matched filter that can be used to correlate the input signal with the known OFDM training signal. Here, $T$ is the sampling interval and $c_i$ are the matched filter coefficients, which are the complex conjugates of the known training signal. From the correlation peaks in the matched filter output signal, both timing and frequency offset can be estimated.

### 1.3 Motivation

For accurate time and frequency synchronization, the correlation technique using the cyclic prefix is not appropriate, because the length of correlation is short to find a fine timing point. And the synchronization
performance which use a special training OFDM symbol will be degraded when each OFDM subchannel has different fading. In addition, in cellular based OFDM system, we can expect that the signal will be interfered by adjacent cells.

Therefore, this thesis propose new preamble structures which are expected to work well at a low SNR, in cellular based systems. It can guarantee the cell edge mobiles to communicate reliably. Each preamble structure consists of synchronization part, cell group part and cell identification part. The principle is to impose PN-sequences on the OFDM transmitted signal in timedomain for time and frequency synchronization. And the PN-sequences occupies the whole or part of the preamble according to the two different preamble structures. Finally this thesis investigates the performance of synchronization, cell searching and mean acquisition time of the two preamble structures.

1.4 Organization

This thesis is organized as follows. Chapter II describes the system model and in chapter III, we propose two preamble structures. Chapter IV introduces frame and frequency synchronization process, cell identification method and mean acquisition of cell identification to each preamble structure. In chapter V, we presents the performance of each proposed preamble structure obtained by simulation. Conclusions and further work are given in chapter VI.
II. System Model

2.1 OFDM System Structure

The principle of OFDM is to split a high-rate datastream into a number of lower rate streams that are transmitted simultaneously over a number of subcarriers. The basic block diagram of typical OFDM system is shown in Fig. 2.1. There are two main incorporated principles. Firstly the IFFT and FFT are used for, respectively, modulation and demodulation technique. Secondly the cyclic prefix, whose length should exceed the maximum delay spread, is used as a guard interval and it can eliminate the inter-symbol interference.

Figure 2.1: OFDM system block diagram
2.2 OFDM System Model

2.2.1 Signal Model

The OFDM signal is expressed as a sum of the prototype pulses shifted in the time and frequency directions and multiplied by the data symbols. In equation form, the $k$-th OFDM symbol is defined as

$$s_k(t-kT) = \begin{cases} 
  \text{Re}\{w(t-kT)\sum_{i=-N/2}^{N/2-1} x_{i,k} e^{j2\pi(f_c+i\frac{f_{FFT}}{T_{FFT}})(t-kT)}\}, & kT-T_{\text{win}}-T_{\text{guard}} \leq t \leq kT + T_{FFT} + T_{\text{win}} \\
  0, & \text{otherwise} 
\end{cases}$$

(II.1)

where $w(t)$ denotes the transmitter pulse shape defined as

$$w(t) = \begin{cases} 
  \frac{1}{2}[1 - \cos(\pi(t + T_{\text{win}} + T_{\text{guard}})/T_{\text{win}})], & -T_{\text{win}} - T_{\text{guard}} \leq t < -T_{\text{guard}} \\
  1, & -T_{\text{guard}} \leq t < -T_{FFT} \\
  \frac{1}{2}[1 + \cos(\pi(t - T_{FFT})/T_{\text{win}})], & T_{FFT} \leq t < -T_{\text{win}} 
\end{cases}$$

(II.2)

A list of parameters is given in Table 2.1.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Explanation</th>
</tr>
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<tbody>
<tr>
<td>$T$</td>
<td>symbol length</td>
</tr>
<tr>
<td>$T_{FFT}$</td>
<td>FFT-time; effective part of OFDM symbol</td>
</tr>
<tr>
<td>$T_{\text{guard}}$</td>
<td>duration of cyclic prefix</td>
</tr>
<tr>
<td>$T_{\text{win}}$</td>
<td>window interval</td>
</tr>
<tr>
<td>$f_c$</td>
<td>center frequency</td>
</tr>
<tr>
<td>$N$</td>
<td>number of FFT-point</td>
</tr>
<tr>
<td>$k$</td>
<td>index of transmitted symbol</td>
</tr>
<tr>
<td>$x_{i,k}$</td>
<td>signal constellation point</td>
</tr>
</tbody>
</table>
finally, a continuous sequence of transmitted symbol is expressed as

\[ s(t) = \sum_{k=-\infty}^{\infty} s_k(t - kT) \] (II.3)

2.2.2 Time-Dispersive Channel

The received signal is expressed by the convolution between the channel impulse response and transmitted signal plus AWGN.

\[ r(t) = h(\tau, t) \ast s(t) + n(t) = \int_{0}^{\tau_{\text{max}}} h(\tau, t)s(t - \tau)d\tau + n(t) \] (II.4)

where \( h(\tau, t) \) and \( n(t) \) means the impulse response of channel and AWGN respectively. Maximum delay spread \( (\tau_{\text{max}}) \) is shorter than the cyclic prefix, therefore there is no interference between consecutive OFDM symbols.

2.2.3 OFDM Demodulation

The demodulation of the OFDM signal should be performed by a bank of filters, which are matched to the effective part \([kT, kT + T_{FFT}]\) of the OFDM symbol. From the received signal \( r(t) \), try to extract the transmitted signal constellation \( x_{i,k} \). The received signal constellations are denoted \( y_{i,k} \)

\[ y_{i,k} = \frac{1}{T_{FFT}} \int_{t=kT}^{kT+T_{FFT}} r(t)e^{-j2\pi i(t-kT)/T_{FFT}} dt \]

\[ = \frac{1}{T_{FFT}} \int_{t=kT}^{kT+T_{FFT}} \left[ \int_{0}^{\tau_{\text{max}}} h_k(\tau)s(t - \tau)d\tau + n(t) \right] e^{-j2\pi i(t-kT)/T_{FFT}} dt \] (II.5)
because of the integration ranges, $\tau_{\text{max}} < T_{\text{guard}}$, there is no interference of the consecutive OFDM symbols, so Eq.IV.2 can be represented as

$$ y_{i,k} = \frac{1}{T_{\text{FFT}}} \int_{t=kT}^{kT+T_{\text{FFT}}} \left[ \int_{0}^{\tau_{\text{max}}} h_k(\tau) \sum_{i'=-N/2}^{N/2-1} x_{i',k} e^{-j2\pi i' (t-kT-\tau)/T_{\text{FFT}}} d\tau \right] e^{-j2\pi i(t-kT)/T_{\text{FFT}}} dt $$

$$ + \frac{1}{T_{\text{FFT}}} \int_{t=kT}^{kT+T_{\text{max}}} n(t) e^{-j2\pi i(t-kT)/T_{\text{FFT}}} dt $$

(II.6)

the second integral in Eq.IV.3 leads to independent additive noise samples $n_{i,k}$ since the complex exponential terms represent orthogonal functions. Substituting $u = t - kT$,

$$ y_{i,k} = \sum_{i'=-N/2}^{N/2-1} \frac{1}{T_{\text{FFT}}} \int_{u=0}^{T_{\text{FFT}}} \left[ \int_{0}^{\tau_{\text{max}}} h_k(\tau) e^{j2\pi i' (u-\tau)/T_{\text{FFT}}} d\tau \right] e^{-j2\pi i' u/T_{\text{FFT}}} du + n_{i,k} $$

$$ = \sum_{i'=-N/2}^{N/2-1} \frac{1}{T_{\text{FFT}}} \int_{u=0}^{T_{\text{FFT}}} \left[ \int_{0}^{\tau_{\text{max}}} h_k(\tau) e^{j2\pi i' \tau/T_{\text{FFT}}} d\tau \right] e^{-j2\pi (i-i') u/T_{\text{FFT}}} du + n_{i,k} $$

(II.7)

the inner integral of the second expression represents the Fourier transform of $h_k(\tau)$ at the frequency instants $i'/T_{\text{FFT}} = i' F$, which is the sampled channel transfer function at time $kT$. It is expressed by the channel coefficients

$$ h_{i',k} = \text{FT}\{h_k(\tau)\} = \int_{\tau=0}^{\tau_{\text{max}}} h_k(\tau) e^{j2\pi i' \tau/T_{\text{FFT}}} d\tau = H(i' F, kT) $$

(II.8)

so the receiver of output is

$$ y_{i,k} = \sum_{i'=-N/2}^{N/2-1} x_{i',k} h_{i',k} \frac{1}{T_{\text{FFT}}} \int_{u=0}^{T_{\text{FFT}}} e^{-j2\pi (i-i') u/T_{\text{FFT}}} du + n_{i,k} $$

(II.9)
the integral in this equation has the value 1, only if \( i = i' \). For \( i \neq i' \), \( i \) and \( i' \) being integer values, the integral is zero. Thus we finally obtain

\[
y_{i,k} = x_{i,k} h_{i,k} + n_{i,k}
\]  

(II.10)

From this form it is seen that a perfectly synchronized OFDM system can be viewed as a set of parallel Gaussian channels as depicted in Fig.2.2. The multipath channel introduces an attenuation/amplification and phase rotation according to the channel coefficients \( h_{i,k} \).

![OFDM system block diagram](image)

Figure 2.2: OFDM system block diagram

Channel estimation is required in order to retrieve the data contained in these signal constellations, because the receiver must have a phase (and amplitude) reference to correctly detect the transmitted symbol. Differential detection can be used alternatively, in which case the decision is made by comparing the phases (and amplitudes) of symbols transmitted over adjacent sub-carriers or subsequent OFDM symbols.

Due to the attenuation/amplification, each sub-carrier typically has an individual signal- to-noise ratio (SNR). The SNR per subcarrier (after the FFT) is defined as
\[(E_c/N_0)_{i,k} = E\{|x_{i,k}|^2\}|h_{i,k}|^2/\sigma_N^2 \quad \text{(II.11)}\]

where \(\sigma_N^2 = E\{|n_{i,k}|^2\}\) is the noise variance.

### 2.2.4 Time Synchronization Error

The impact of an FFT-timing offset at the receiver can be analyzed mathematically by shifting the integration interval of the matched filter bank, Eq. IV.2. For a timing error of \(\delta t\), the ideal interval \(t \in [kT, kT + T_{FFT}]\) becomes \(t \in [kT + \delta t, kT + T_{FFT} + \delta t]\) and Eq. IV.2 can be written as

\[y_{i,k} = \frac{1}{T_{FFT}} \int_{t=kT+\delta t}^{kT+T_{FFT}+\delta t} r(t) e^{-j2\pi i(t-kT-\delta t)/T_{FFT}} dt \quad \text{(II.12)}\]

The phase rotation is zero at the center frequency and it linearly increases towards the edges of the frequency band. If coherent detection is utilized, the induced progressive phase rotation is detected implicitly by the channel estimation algorithm. The subsequent equalization will thus automatically correct for small timing-offsets. No performance degradation is thereby caused. However, if the timing offset is too large, ISI and ICI are introduced because energy is also collected from one of the adjacent OFDM symbols, leading to a partial loss of orthogonality.

\[y_{i,k} = x_{i,k}h_{i,k}e^{-j2\pi i\delta t/T_{FFT}} + n_{i,k} = x_{i,k}h_{i,k}e^{-j2\pi i\delta t'/N} + n_{i,k} \quad \text{(II.13)}\]
2.2.5 Frequency Synchronization Error

Frequency offsets are typically introduced by a frequency mismatch in the local oscillators of the transmitter and the receiver. Doppler shifts can be neglected in indoor environments.

The impact of a frequency error can be seen as an error in the frequency instants, where the received signal is sampled during demodulation by the FFT. Fig. 2.3 depicts this two-fold effect. The amplitude of the desired sub-carrier is reduced (+) and inter-carrier-interference ICI arises from the adjacent sub-carriers (o).

![Diagram](image.png)

Figure 2.3: ICI due to frequency synchronization error

A frequency offset can be accounted for by a frequency shift \( \delta f \) and a phase offset \( \theta \) in the lowpass equivalent received signal

\[
r'(t) = r(t)e^{j2\pi \delta ft + \theta} \tag{II.14}
\]

with Eq.IV.2 the received signal can be written as
\[ y_{i,k} = \frac{1}{T_{FFT}} \int_{t=kT}^{kT + T_{FFT}} r(t) e^{j(2\pi \delta ft + \theta)} e^{-j2\pi i(t-kT)/T_{FFT}} dt \]

\[ = e^{j2\pi \theta} \frac{1}{T_{FFT}} \int_{t=kT}^{kT + T_{FFT}} \left[ \int_{\tau=0}^{\tau_{max}} h(\tau) s(t-\tau) d\tau + n(t) \right] e^{j2\pi \delta ft} e^{-j2\pi i(t-kT)/T_{FFT}} dt \]

(II.15)

due to the frequency error, the integral is not equal zero for \( i \neq i' \), neither it is one for \( i = i' \), as in the idealized case above. I.e., the orthogonality between sub-carriers has been partly lost. The evaluation of this expression yields two terms. The first term accounts for equal phase rotation and attenuation of all sub-carriers, the second one describes the ICI.

\[ y_{i,k} = e^{j(\theta + 2\pi \delta ft)} x_{i,k} h_{i,k} \frac{1}{T_{FFT}} \int_{u=0}^{T_{FFT}} e^{j2\pi \delta fu} du \]

\[ + e^{j(\theta + 2\pi \delta ft)} \sum_{i'=-N/2,i' \neq i}^{N/2-1} x_{i',k} h_{i',k} \frac{1}{T_{FFT}} \int_{u=0}^{T_{FFT}} e^{-j2\pi(i + \delta f - \delta f_i)u} du + n_{i,k} \]

(II.16)

these expressions are valid for a frequency-offset \( \delta f < 0.5 \) subcarrier spacing. For larger offsets, the transmitted data symbols \( x_{i,k} \) would get shifted by one or more positions in the frequency direction. I.e., the data symbol of the \( i - th \) transmitted subcarrier would appear at the \( (i + \delta fi) \)-th subcarrier at the receiver, where \( \delta fi = round(\delta f/F) \) is the integer part of the frequency error in subcarriers.

In result, the phase rotation and attenuation due to a frequency error yields

\[ y_{i,k} = x_{i,k} h_{i,k} \text{sinc}(\delta f T_{FFT}) e^{j[\theta + 2\pi \delta f(kT + T_{FFT}/2)]} + n'_{i,k} \]  

(II.17)
2.3 Cellular and Asynchronous OFDM System

The considered OFDM system consists of 7 cells, each cell has one Base station which is located at a cell origin. And the frame of all cells are not necessarily synchronized. The mobiles are uniformly distributed in cell 3.

Fig. 2.4 describes the interference model in cellular systems. The received signal must be combined signal from adjacent base stations (BSs).

Figure 2.4: Interference model in cellular OFDM systems
III. Preamble Structures

This chapter describes the proposed preamble structures in detail. Each preamble structure consists of synchronization part, cell group part and cell identification part. The principle is to impose PN-sequences on the OFDM transmitted signal in time domain for time and frequency synchronization.

3.1 The First Preamble Structure

Three different preambles construct preamble structure 1 (PS1). One of them is designed for synchronization (S₁-preamble), another for cell group (G₁-preamble), and the last one for cell identification (C₁-preamble). Where 1 means the first preamble structure. The proposed preamble structure is shown in Fig. 3.1.

Figure 3.1: The first preamble structure
The $S_1$-preamble has the purpose of synchronization and composed of repeated PN-sequences. In Fig. 3.1, $P_{S1}$, $P_{G1}$ and $P_{C1}$ represent PN-sequences, cell group signal and cell identification signal respectively. The $G_1$-preamble and $C_1$-preamble produce a repeated signal pattern by only using the even number of indices of subcarriers in OFDM systems in Fig. 3.2 [6].

![Diagram](image)

Figure 3.2: The cell and cell group representation in PS1

The samples of transmitted baseband preamble can be given by

\[
s(n) = \begin{cases} 
  p_1(n) \mod N, & S_1\text{-preamble} \\
  \sum_{k=0}^{N_S-1} g_{1,j}(k)e^{j2\pi kn/N}, & G_1\text{-preamble} \\
  \sum_{k=0}^{N_S-1} c_{1,m}(k)e^{j2\pi kn/N}, & C_1\text{-preamble}
\end{cases}
\]

where $p_1(n)$ is composed of repeated PN-sequences of length $N$, $N_S$ denotes the number of samples per symbol and also the number of FFT points,
\[
g_{1,j}(k) = \begin{cases} 
p_{G1,j}(i)_{\mod N_{G1}}, & k = 2i \\
0, & \text{otherwise}
\end{cases} \quad (\text{III.2})
\]
\[
c_{1,m}(k) = \begin{cases} 
p_{C1,m}(i)_{\mod N_{C1}}, & k = 2i \\
0, & \text{otherwise}
\end{cases} \quad (\text{III.3})
\]

where \(p_{G1,j}(i)\) and \(p_{C1,m}(i)\) are repeated PN-sequences of length \(N_{G1}\) and \(N_{C1}\), which represent cell group \((j)\) and cell identification \((m)\) respectively. The \(j\) and \(m\) are the index of cell group and cell identification in PS1.

### 3.2 The Second Preamble Structure

Fig. 3.4 shows the other proposed preamble structure. The two different preambles which consist of cell group \((G_2\text{-preamble})\) and cell identification preamble \((C_2\text{-preamble})\) constitute preamble structure 2 (PS2). Fig. 3.4 shows that the \(G_2\text{-preamble}\) consists of 4 signal and PN-sequence parts. The signal parts \((P_{G2})\) are designed for cell group identification and the PN-sequence \((P_{S2})\) parts for frame synchronization. Fig. 3.3 describes the IFFT in PS2.

The length of each part is \(N_s/N_p\), so the first region that the range of samples is from 0 to \(N_s/N_p\) is occupied by signal part and the next region that the range of samples is from \(N_s/N_p\) to \(2N_s/N_p\) is occupied by PN-sequence part, and so on. For example, the samples of transmitted \(G_2\text{-preamble}\) are composed of \(P_{G2}\) and \(P_{S2}\), then, they are given by

\[
P_{G2}(n) = \sum_{k=0}^{N_s-1} g_{2,l}(k) e^{j2\pi nk/N_s}, \quad 0 \leq n < N_s/N_p \quad (\text{III.4})
\]
Figure 3.3: The cell and cell group representation in PS2

\[
g_{2,1}(k) = \begin{cases} 
    p_{G_{2,1}(i) \mod N_{G_{2}}, k = N_P i} & \text{if } k = N_P i \\
    0 & \text{otherwise}
\end{cases} \quad (I.5)
\]

\[
P_{s2}(n) = p_{2(n) \mod N}, \quad N_S/N_P \leq n < 2N_S/N_P \quad (I.6)
\]

Preamble structure 2 presents the same sample pattern between the first half and the latter half of the preamble such as PS1. So we can use this preambles for fractional frequency offset estimation. \(C_2\)-preamble can be described the same as \(G_2\)-preamble, the only difference is that \(C_2\)-preamble is designed for cell identification.
Figure 3.4: The second preamble structure
IV. Performance Analysis

This chapter presents the process of frame and frequency synchronization, and investigates the algorithm of cell identification and its mean acquisition time. Frame synchronization is derived from $S$-preamble, cell group and cell identification from $G$ and $C$-preamble. Frequency synchronization can be achieved from multiple preamble, since it is difficult to get a reliable frequency estimation by one preamble symbol.

4.1 Frame Synchronization

We begin frame synchronization with symbol timing. The symbol timing is obtained from cyclic prefix correlation technique [4]. Let the sampled received signal be $r(n)$, then, the symbol timing, $\tau_s$, is estimated as

$$\tau_s = \arg \max_n \sum_{u=0}^{N_{CP}-1} r^*(n+u)r(n+u+N_S)$$  \hspace{1cm} (IV.1)

where $N_{CP}$ denotes the number of samples in a guard interval, $N_S$ means the number of samples in a symbol, and $\tau_S$ is estimated in every symbol. For frame synchronization, we impose repeated PN-sequences on the OFDM transmitted signal in the whole ($S_1$-preamble) or part of the preamble ($G_2$ or $C_2$-preamble) and then, the received signal cross-correlated with the known PN-sequences. When the received signal coincides with the known PN-sequences, a correlation peak occurs. In this thesis, to find the frame synchronization means to search the sample at which the cross correlation value with the known PN-sequences exceeds the threshold. Hence, the frame timing, $\tau_f$, is estimated as
\[ \tau_f = \sum_{u=0}^{M \times N - 1} r(n + u)p(u)_{\text{mod}N} \geq \chi_{th} \]  

(IV.2)

where \( r(n) \) is the received signal and \( p(u) \) denotes the known PN-sequences. \( M \) and \( N \) represent the number of repetition and the length of PN-sequences respectively. \( \chi_{th} \) means the threshold value.

### 4.2 Frequency Synchronization

In preambles (including \( S, G \) and \( C \)-preambles), at the receiver, there will be a phase difference between the samples in the first half and their replica in the second half caused by the carrier frequency offset. To get a insignificant BER degradation, the tolerable frequency offset should be below 1 percent of the subcarrier spacing, which is not likely achievable within one preamble of our proposed preamble structures at a low SNR. Therefore, in this paper, we employ multiple preambles for frequency offset estimation. We can estimate the fractional frequency offset, \( \varepsilon \), as

\[ \varepsilon = \frac{1}{2\pi} \Im \left\{ \sum_{j=0}^{N-1} z(n + jN_S) \right\} \]  

(IV.3)

\[ z(n) = \sum_{u=0}^{N_S/2 - 1} r^*(n + u)r(n + u + N_S/2) \]  

(IV.4)

where \( N \) is the number of preambles which are employed for frequency offset estimation. The maximum frequency range is limited to \( \pm 1/2 \) of the subcarrier spacing. The integer frequency offset can be estimated by the short repeated PN-sequences. To evaluate the integer frequency offset, this thesis, firstly, detect the phase difference between adjacent PN-sequences using multiple preambles, and then multiply it by the
number that the number of samples of one symbol is divided by the length of a short PN-sequence.

### 4.3 Cell Identification

For cell identification, we first, search a cell group and then identify cell in that cell group. We can find a cell group in PS1 by the following method. For simplicity, we ignore the Doppler spread. In Eq. (III.1), the FFT output of $G_1$-preamble ($r$-th OFDM symbol, subcarrier $k$) and $C_1$-preamble($s$-th OFDM symbol, subcarrier $k$) are given by

$$R_r(k) = g_{1,j}(k) \cdot H(k) + w(k) \quad \text{(IV.5)}$$

$$R_s(k) = c_{1,m}(k) \cdot H(k) + w(k) \quad \text{(IV.6)}$$

where $w(k)$ is a white complex Gaussian noise. The channel frequency response at subcarrier, $f_k = k/T_s$, can be written as

$$H(k) = \sum_{l=0}^{L-1} h_l e^{-j2\pi k\tau_l} \quad \text{(IV.7)}$$

where $T_s$ is symbol duration. And cell group can be represented by

$$G_i = \arg \max_i \sum_{r=0}^{N_r-1} \sum_{k=0}^{N_s-1} R_r(k)g_{1,i}(k) \quad \text{(IV.8)}$$

where $N_r$ means the number of symbols between successive $G_1$-preambles. And cell identification is given as

$$C_p = \arg \max_p \sum_{s=0}^{N_r-1} \sum_{k=0}^{N_s-1} R_s(k)c_{1,p}(k) \quad \text{(IV.9)}$$
4.4 Mean Acquisition Time

Fig. 5.7 shows the circular state diagram of searching mean acquisition
time for cell identification and the abbreviations are illustrated as table 4.1,

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Explanation</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L$</td>
<td>Number of symbol state for frame synch.</td>
</tr>
<tr>
<td>$C$</td>
<td>Interval from S-preamble to G-preamble</td>
</tr>
<tr>
<td>$K$</td>
<td>Penalty time for false alarm in cell id.</td>
</tr>
<tr>
<td>$P_{faF}$</td>
<td>False alarm probability in frame detection</td>
</tr>
<tr>
<td>$P_{dF}$</td>
<td>Detection probability in frame detection</td>
</tr>
<tr>
<td>$P_{dfC</td>
<td>faF}$</td>
</tr>
<tr>
<td>$P_{dC</td>
<td>dF}$</td>
</tr>
<tr>
<td>$P_{faC</td>
<td>dF}$</td>
</tr>
<tr>
<td>$P_{dfC</td>
<td>dF}$</td>
</tr>
</tbody>
</table>

We define the generalized gains $U(z)$, associated with the various
branches of the circular state diagram, as follows. $H_D(z)$ corresponds
to the path from frame synchronization to cell identification, $H_M(z)$
models the total path gains that the one is missing the frame synchro-
nization, another is following, after correct frame synchronization, by
either a false alarm of cell with penalty time $K$ or missing the cell with
time $C$, and the last one is returning with penalty time $K$ after a false
frame synchronization. $H(z)$ is equivalent to the two path gains be-
tween any two successive states, that the one is the case without false
alarm of frame synchronization and the other is, by first a false alarm,
following by a return to the next state with time $C$.

$$H(z) = (1 - P_{faF})z + P_{faF}P_{dF|faF}z^{C+1}$$

$$H_D(z) = P_{dF}P_{dC|dF}z^{2C+1}$$

$$H_M(z) = (1 - P_{dF})z + P_{dF}P_{dF|dF}z^{C+1}$$
$$+ P_{dF}P_{faC|dF}z^{K+1}$$
$$+ P_{faF}(1 - P_{dF|faF})z^{K+1} \quad (IV.10)$$

$$U(z) = \frac{H_D(z)H_{L-1}(z)}{1 - H_M(z)H_{L-1}(z)} \quad (IV.11)$$

From Eq.(IV.11), the mean acquisition time can be derived as

$$T_{acq} = \frac{dU(z)}{dz} \bigg|_{z=1} \quad (IV.12)$$
Figure 4.1: ICI due to frequency synchronization error
V. Simulation Results

The system parameters for simulation are given in Table 5.1

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Carrier Frequency</td>
<td>2GHz</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>20MHz</td>
</tr>
<tr>
<td>Channel Model</td>
<td>Veh.Ch.A</td>
</tr>
<tr>
<td>Velocity</td>
<td>100km/h</td>
</tr>
<tr>
<td>SNR (For synch.)</td>
<td>-6dB</td>
</tr>
<tr>
<td>Symbol duration</td>
<td>100us</td>
</tr>
<tr>
<td>samples/ Symbol</td>
<td>256(CP), 2048(Data)</td>
</tr>
<tr>
<td>Symbols/ Frame</td>
<td>32</td>
</tr>
<tr>
<td>PN-length for frame synch.</td>
<td>128, 256</td>
</tr>
<tr>
<td>PN-length for cell group or cell id.</td>
<td>32</td>
</tr>
</tbody>
</table>

In our study, we have adopted averaged and normalized Vehicular A channel in ITU SISO channel model for our delay profile near 2 GHz bandwidth. The multipath power delay profiles are shown in following Tables 5.2.

5.1 PN sequence

This thesis superimposes PN sequence on the OFDM transmitted signal to achieve a synchronization, so we do some simulations about time and frequency synchronization with changing the length of PN-sequence and
Table 5.2: Normalized multipath power-delay profile for Veh.A

<table>
<thead>
<tr>
<th>Path</th>
<th>Power</th>
<th>Delay(µs)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.3856</td>
<td>0</td>
</tr>
<tr>
<td>2</td>
<td>0.0611</td>
<td>310</td>
</tr>
<tr>
<td>3</td>
<td>0.0485</td>
<td>710</td>
</tr>
<tr>
<td>4</td>
<td>0.0153</td>
<td>1090</td>
</tr>
<tr>
<td>5</td>
<td>0.0048</td>
<td>1730</td>
</tr>
<tr>
<td>6</td>
<td>100us</td>
<td>2510</td>
</tr>
</tbody>
</table>

with or without frequency offset. Each simulation scenario assumes signal-to-noise ratio is equal to $-6dB$ and delay spreads accord with Vehicular Channel A. The frequency offset is equal to $+/- 20ppm$.

Frame synchronization performance is characterized by the detection failure probability and false alarm probability. This thesis get the synchronization point which the detection failure probability is equal to the false alarm probability. The threshold ratio is written as

$$th = \frac{\left| \sum_{n=0}^{N-1} r(n)c(n) \right|^2}{\sum_{n=0}^{N-1} r(n)^2} \times N$$  \hspace{1cm} (V.1)

Fig. 5.1 and Fig. 5.2 show the false alarm probability and detection failure probability when the PN length takes 1/4 part of the symbol. The performance difference is slight with respect to the length of PN sequence(128, 256). In Fig. 5.1, the detection failure probability is 0.011 but the case with frequency offset, the detection failure probability is 2 times more higher than the case without frequency offset.

Fig. 5.3 represents the 128 PN length repetition has two times better performance than 256 PN length. The detection failure probability is below 0.005 in with or without frequency offset.
Table 5.3 and Table 5.4 represent the comparison of frame synchronization and frequency offset performance between signal repetition and PN sequence repetition. Table 5.3 describes PN sequence repetition is better for frame synchronization than the signal repetition. Because PN sequence repetition has a sharp correlation peak, it produces more accurate synchronization point. In Table 5.3, we can see that the RMS frequency offset error of PN repetition and signal repetition are 0.11 and 0.26 of subcarrier spacing, respectively, with frequency offset.

Table 5.3: Comparison of timing synchronization performance between PN repetition and signal repetition

<table>
<thead>
<tr>
<th>frequency offset</th>
<th>10ppm</th>
</tr>
</thead>
<tbody>
<tr>
<td>PN (128 x 16)</td>
<td>5e-4</td>
</tr>
<tr>
<td>signal</td>
<td>4e-3</td>
</tr>
</tbody>
</table>

Table 5.4: Comparison of frequency offset performance between PN repetition and signal repetition

<table>
<thead>
<tr>
<th>frequency offset</th>
<th>5ppm</th>
<th>10ppm</th>
</tr>
</thead>
<tbody>
<tr>
<td>PN</td>
<td>0.113473</td>
<td>0.113569</td>
</tr>
<tr>
<td>signal</td>
<td>0.259086</td>
<td>0.259614</td>
</tr>
</tbody>
</table>

In Fig. 5.4, we can see the performance of integer and fractional frequency offset varying the number of repetition which is used for average. The result of integer frequency offset error is the probability of detection error and the fractional frequency offset error is the RMS frequency offset error normalized with subcarrier spacing. As we can see, the repetition symbols which are used for averaging increase, the frequency offset errors decrease.
5.2 Frame Synchronization

Table 5.5 shows the simulation result for frame synchronization errors with frequency offset 10ppm. In the simulation, we assume optimal threshold value that the probability of false alarm of frame detection is the same as that of detection failure.

<table>
<thead>
<tr>
<th>PN-length</th>
<th>$P_e$</th>
</tr>
</thead>
<tbody>
<tr>
<td>128 x 16 ($T_s$)</td>
<td>5e-4</td>
</tr>
<tr>
<td>128 x 8 ($T_s/2$)</td>
<td>2e-3</td>
</tr>
<tr>
<td>256 x 8 ($T_s$)</td>
<td>2.5e-3</td>
</tr>
<tr>
<td>256 x 4 ($T_s/2$)</td>
<td>4.5e-3</td>
</tr>
</tbody>
</table>

From this table, the frame synchronization error probability is less than 4.5e-3, when the signal to noise ratio is -6dB. In addition, the result indicates that the shorter the length of PN-sequence is and the more repetition of PN-sequences is, the better the performance is. Therefore, we conclude that the insertion of PN-sequences in time-domain is efficient for frame synchronization at a low SNR.

5.3 Frequency Synchronization

In this paper, a frequency offset estimation is performed for fractional offset. Because the integer frequency offset error occurs negligibly when we employ multiple preambles. In chapter IV, we explain the scheme to find integer frequency offset using PN-sequence repetitions.

Fig. 5.5 shows the root-mean-square value of frequency error from the simulation. As we can see, if we increase the number of preambles which are used for frequency offset estimation, the performance could
be improved. To satisfy the target of frequency offset estimation error, 1 percent of subcarrier spacing, we should employ about 30 preambles at each preamble structure. Additionally, when the average frame length is short, we find out that the PN-sequences repetition in time domain is better for the estimate of frequency offset than the OFDM signal repetition, because OFDM signals receive more influence from noise than PN-sequences.

### 5.4 Cell Identification

Fig. 5.6 represents the cell identification error probability of PS1 and PS2 with varying SNR.

PS1 can use the half number of total subcarriers for cell group and cell identification, but PS2 can only use the one over eight of total subcarriers in simulation due to eight-time signal repetition. The graph says that PS1 gives the performance of 6dB gain compared to PS2, because PS1 can utilize the four-time number of subcarriers as that used in PS2 for cell and cell group identification.

### 5.5 Mean Acquisition Time

In our proposed preamble structure, the number of symbol states for searching synchronization and cell identification is different in PS1 and in PS2. For example, if we miss detection of a frame in PS1, we could find the $S_1$-preamble for frame synchronization after three frames passed, but in PS2 we could find it only after 1 frame passed. So, in the simulation, $L$ is 96 in PS1, 32 in PS2 because one frame consists of 32 symbols. What is more, after frame synchronization, to identify a cell group, the interval symbol state from S-preamble to G-preamble is 32 in PS1, but 0 in PS2. Because in PS2, one preamble includes both
synchronization parts and cell group or cell identification parts. If we evaluate the necessary cell searching time in this way, PS2 will need a shorter time than PS1 to identify the cell, however, PS2 has a higher failure probability of synchronization and cell identification than PS1. In result, Fig. 5.7 shows that at lower SNR, the mean acquisition time of PS1 is shorter than PS2 due to the higher error probability of synchronization and cell identification in PS2. However, when we increase the SNR, PS2 needs the shorter time for cell identification than PS1. Because at a high SNR, both PS1 and PS2 have very little error probabilities of frame synchronization and cell identification, so the delay time for searching frame synchronization and cell identification could be dominant for mean acquisition time.
Figure 5.1: Detection failure and false alarm probability with 512 PN length without frequency offset
Figure 5.2: Detection failure and false alarm probability with 512 PN length with frequency offset
Figure 5.3: Detection failure and false alarm probability with 1024 PN length
Figure 5.4: Frequency offset estimation using 1024 PN length
Figure 5.5: Frequency offset estimation
Figure 5.6: Cell identification error probability
Figure 5.7: Mean acquisition time (symbol)
VI. Conclusions and Further work

This thesis considers time and frequency synchronization, cell searching and mean acquisition time in cellular based OFDM systems. The synchronization performance of previous technique will be degraded when each subchannel experience different fading and the SNR is low. Therefore, this thesis propose new preamble structure which can overcome these problems and also make it possible to find cell identification.

The new preamble structures impose PN-sequence in time-domain for synchronization and cell searching in OFDM cellular based systems. The results explain that our proposed preamble structures achieve very robust synchronization and cell searching performance at a low SNR. The mean acquisition time for cell searching is from 0.01s to 0.1s below the 0dB SNR condition. In cellular systems, it is required to support a reliable communication of the cell boundary mobile, these proposed schemes must be appropriate for OFDM cellular based preamble structures.

This thesis assumes the perfect channel estimation and no Doppler spread. If the system has some problems about channel estimation, the performances will also be degraded. In practical cellular based OFDM systems, the channel estimation is an important issue, therefore, for further work we study the channel estimation in cellular system and present synchronization and cell searching performance based on channel estimation.
OFDM 다중셀 환경에서 동기와 셀탐색을 위한 선험자 구조

김종남

최근 멀티미디어 서비스에 대한 요구의 증가로 OFDM에 대한 연구가 활발히 진행되고 있다. 반면 OFDM 시스템은 데이터를 다중 반송파로 전송하므로 단일 반송파 전송의 경우보다 더욱 정확한 동기를 필요로 한다.

이 논문에서는 OFDM 다중셀 환경에서 동기와 셀탐색에 적합한 선험자를 제안하였다. 제안된 선험자는 동기, 셀그룹 그리고 셀탐색 부분으로 구분된다. 본 논문은 시간동기를 위해서 OFDM 선험자의 시간축 신호에 PN 코드를 삽입하고 셀그룹과 셀탐색은 OFDM 전송 신호에 의해 획득 할 수 있도록 한다. 또한 전체 선험자는 시간축에서 신호의 전반부와 후반부가 동일하게 구성되어 있으므로 주파수 동기를 얻을 수 있다.

시뮬레이션은 시간, 주파수 동기, 셀탐색 그리고 평균 엑춰지션 시간에 대해 진행하였는데 본 논문에서 제안된 선험자 구조가 다중셀 OFDM 시스템환경에서 우수한 성능을 보임을 확인 할 수 있었다.
References


