

Chapter 18:

Diversity and Error Compensation in OFDM Transceivers: Principles and Implementation

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18.1 Introduction

Digital modulation techniques can be largely divided into two categories. The one is single-carrier modulation, which utilizes a single radio frequency (RF) to transmit data. The other is multi-carrier modulation, which utilizes simultaneously modulated multiple RF carriers in order to combat inter symbol interference (ISI) while increasing communication bandwidth. This chapter focuses a particular type of multi-carrier modulation known as orthogonal frequency division multiplexing (OFDM). The idea of OFDM [1] was proposed very old and later Discrete Fourier Transform (DFT) based implementation algorithm was proposed [2]. Then, OFDM became practical and has been popular in a number of applications such as an asymmetric digital subscriber line (ADSL), wireless local area network (WLAN), digital TV broadcasting (DTV). It also becomes strong candidate for 4th generation cellular land mobile radio system.

It is famous that OFDM modulation and demodulation can be implemented by IDFT and DFT. But in actual implementation, not only those DFTs but also several error compensation mechanisms are indispensable because orthogonality between parallel transmitted subcarrier signals are easily destroyed by synchronization errors such as RF error, Clock sampling rate error and FFT window position shift. When applying OFDM technology for mobile communication, Doppler induced RF errors easily distort a reception performance. Then the main contents of this chapter are two fundamental techniques to realize high performance OFDM communication systems. Those are Diversity technologies and Synchronization error Detection and Compensation methods. After all, real hardware Implementation examples and their data are summarized.

The rest of this chapter is organized into five following sections. Section 18.2 is devoted to present the principles and design of OFDM transceivers. Section 18.3 focuses on adaptive antenna array and several diversity techniques. In Section 18.4, we will deal with the

detection and compensation of various error sources and practical factors. Section 18.5 presents the hardware implementation and related chip performance. In the final section, we will give a summary and make further discussions.

18.2 OFDM transceivers: Principles and design

18.2.1 Single-carrier vs. multi-carrier modulation

In single-carrier modulation, data are sent in serial by modulating one single carrier frequency at a baud rate of R symbols per second. Then symbol length ($1/R$) has to be shortened in order to increase data transmission bandwidth as shown in figure 18.1. However, wireless communication system usually suffers from multipath fading. When the multipath delay time becomes significant compared with the symbol length, ISI happens and degrades wireless communication performance. Then it is not easy to increase the communication bandwidth by just reducing the symbol length.

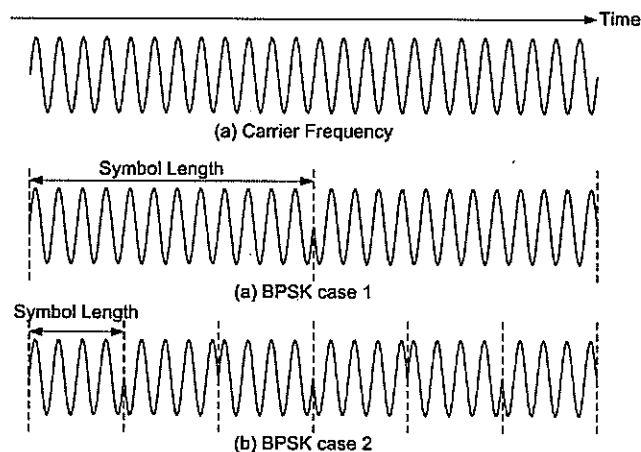


Figure 18.1 Single-carrier modulation

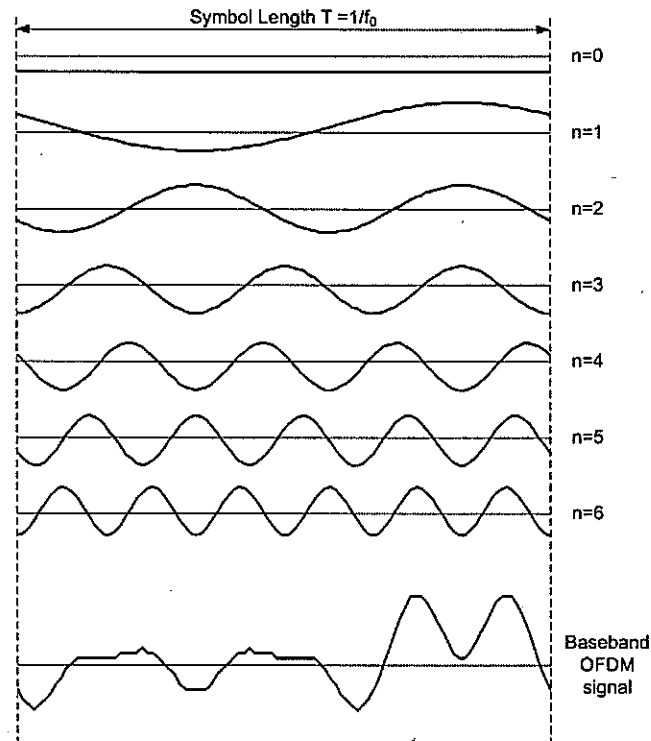


Figure 18.2 Subcarrier and Baseband OFDM signal

The idea of multi-carrier modulation is to use long symbol length against ISI but to use multiple carrier frequencies (called subcarriers) to increase the communication bandwidth. Figure 18.2 shows an example of Baseband OFDM signal, which consists of $n=0$ to 6 seven sinusoids of frequency $=n \cdot f_0$ when symbol length $T = 1/f_0$. By increasing the number of subcarriers, one symbol length can be kept long while achieving high bandwidth data transmission.

18.2.2 OFDM signal generation

OFDM signal consists of plural orthogonal carrier frequencies based on the following three orthogonal relations. In the equations, m and n are integers and $T=1/f_0$.

$$\int_0^T \cos(2\pi m f_0 t) \cdot \cos(2\pi n f_0 t) dt = \begin{cases} \frac{T}{2} & (m = n) \\ 0 & (m \neq n) \end{cases} \quad (18.1)$$

$$\int_0^T \sin(2\pi m f_0 t) \cdot \sin(2\pi n f_0 t) dt = \begin{cases} \frac{T}{2} & (m = n) \\ 0 & (m \neq n) \end{cases} \quad (18.2)$$

$$\int_0^T \cos(2\pi n f_0 t) \cdot \sin(2\pi n f_0 t) dt = 0 \quad (18.3)$$

One subcarrier signal, which is a basic component of OFDM signal, in period $[0, T]$ is expressed as equation (18.4). In one symbol length T , a precise n cycle of sinusoid signal is included. By summing the N component signals, baseband OFDM signal can be obtained as equation (18.5). Here, f_0 is neighboring subcarrier spacing.

$$a \cos(2\pi n f_0 t) - b \sin(2\pi n f_0 t) \quad (18.4)$$

$$s_B(t) = \sum_{n=0}^{N-1} \{a_n \cos(2\pi n f_0 t) - b_n \sin(2\pi n f_0 t)\} \quad (18.5)$$

In practical OFDM transceiver design, complex baseband OFDM signal is used in signal processing instead of real baseband OFDM signal $s_B(t)$, which can be shown by using complex baseband OFDM signal $u(t)$ as equation (18.6). In the equation (18.7), d_n is complex expression of digital modulation symbol such as $d_n = a_n + j b_n$.

$$s_B(t) = \text{Re}[u(t)] \quad (18.6)$$

$$u(t) = \sum_{n=0}^{N-1} d_n e^{j2\pi n f_0 t} \quad (18.7)$$

If $u(t)$ is sampled by N points during the one symbol period ($1/f_0$), the time index t can be substituted by $t = (k/Nf_0)$ as shown in equation (18.8).

$$\begin{aligned} u\left(\frac{k}{Nf_0}\right) &= \sum_{n=0}^{N-1} d_n e^{j2\pi n f_0 \frac{k}{Nf_0}} \\ &= \sum_{n=0}^{N-1} d_n e^{j\frac{2\pi n k}{N}} \\ &= \sum_{n=0}^{N-1} d_n \left(e^{j\frac{2\pi}{N}}\right)^{nk} \quad (k = 0, 1, 2, \dots, N-1) \end{aligned} \quad (18.8)$$

According to equation (18.8), it is clear that complex baseband OFDM signal can be generated by N -points Inverse Discrete Fourier Transform (IDFT) of complex digital modulation symbols d_n ($n=0, 1, \dots, N-1$).

Passband signal $s(t)$ which is actually transmitted from an antenna is shown in equation (18.9). Here, f_c is minimum frequency of passband. Then, passband signal consists of $[f_c, f_c+1f_0, f_c+2f_0, \dots, f_c+(N-1)f_0]$ frequencies.

$$\begin{aligned}
 s(t) &= \sum_{n=0}^{N-1} [a_n \cos\{2\pi(f_c + nf_0)t\} - b_n \sin\{2\pi(f_c + nf_0)t\}] \\
 &= \operatorname{Re} \left[\sum_{n=0}^{N-1} d_n e^{j2\pi(f_c + nf_0)t} \right] \\
 &= \operatorname{Re} \left[e^{j2\pi f_c t} \sum_{n=0}^{N-1} d_n e^{j2\pi n f_0 t} \right] \\
 &= \operatorname{Re} \left[(\cos 2\pi f_c t + j \sin 2\pi f_c t) \sum_{n=0}^{N-1} d_n e^{j2\pi n f_0 t} \right] \\
 &= \cos 2\pi f_c t \cdot \operatorname{Re} \left[\sum_{n=0}^{N-1} d_n e^{j2\pi n f_0 t} \right] + (-\sin 2\pi f_c t) \cdot \operatorname{Im} \left[\sum_{n=0}^{N-1} d_n e^{j2\pi n f_0 t} \right] \quad (18.9)
 \end{aligned}$$

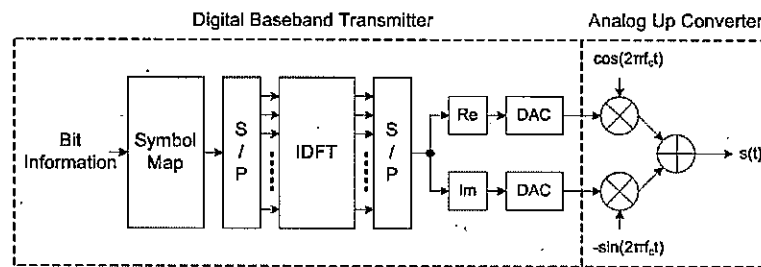


Figure 18.3 OFDM Transmitter Block Diagram

According to equation (18.9), OFDM transmitter block diagram is shown in figure 18.3. First, bit information is mapped onto complex modulation symbols d_n . Then serial-to-parallel converter makes N -points symbol vector d_n ($n=0, \dots, N-1$). By Inverse Discrete Fourier Transform, OFDM complex baseband signal $u(k/Nf_0)$ ($k=0, \dots, N-1$) is generated. After converting the digital baseband signal to analog, passband OFDM signal $s(t)$ is generated by some up-converters.

18.2.3 OFDM signal reception

In this section, OFDM demodulation from passband signal in equation (18.10) is explained. Since passband signal is high frequency signal, analog down conversion process is needed before OFDM digital demodulation process.

$$s(t) = \sum_{n=0}^{N-1} [a_n \cos\{2\pi(f_c + nf_0)t\} - b_n \sin\{2\pi(f_c + nf_0)t\}] \quad (18.10)$$

In typical direct down conversion process multiplies f_c frequency sinusoids to passband signal and applies low-pass filtering as follows.

$$\begin{aligned} s(t)2\cos(2\pi f_c t) &= 2 \sum_{n=0}^{N-1} [a_n \cos\{2\pi(f_c + nf_0)t\} \cos(2\pi f_c t) - b_n \sin\{2\pi(f_c + nf_0)t\} \cos(2\pi f_c t)] \\ &= \sum_{n=0}^{N-1} \{a_n \cos(2\pi n f_0 t) - b_n \sin(2\pi n f_0 t)\} \\ &\quad + \sum_{n=0}^{N-1} [a_n \cos\{2\pi(2f_c + nf_0)t\} - b_n \sin\{2\pi(2f_c + nf_0)t\}] \quad (18.11) \end{aligned}$$

Then, by applying the low pass filter to equation (18.11), $2f_c$ component can be removed and $S_I(t)$ signal is obtained as equation (18.12). This signal is equivalent to the baseband OFDM signal in equation (18.5). In this process, precise f_c frequency sinusoid is used. In general down converter the local oscillator's frequency includes frequency error or frequency drift on time. Then additional RF error detection and compensation is required. These topics on RF error detection and compensation will be explained in the following related sections.

$$s_I(t) = \sum_{n=0}^{N-1} \{a_n \cos(2\pi n f_0 t) - b_n \sin(2\pi n f_0 t)\} \quad (18.12)$$

In order to obtain complex baseband OFDM signal as equation (18.7), another down conversion process is needed as follows.

$$\begin{aligned} s(t)2\{-\sin(2\pi f_c t)\} \\ &= \sum_{n=0}^{N-1} \{a_n \sin(2\pi n f_0 t) + b_n \cos(2\pi n f_0 t)\} \\ &\quad + \sum_{n=0}^{N-1} [-a_n \sin\{2\pi(2f_c + nf_0)t\} - b_n \cos\{2\pi(2f_c + nf_0)t\}] \quad (18.13) \end{aligned}$$

Again, by low pass filtering, $S_Q(t)$ signal is obtained as equation (18.14).

$$s_Q(t) = \sum_{n=0}^{N-1} \{a_n \sin(2\pi n f_0 t) + b_n \cos(2\pi n f_0 t)\} \quad (18.14)$$

By combining $S_I(t)$ and $S_Q(t)$, complex baseband signal $u(t)$ can be obtained as follows.

$$u(t) = s_I(t) + js_Q(t) = \sum_{n=0}^{N-1} (a_n + jb_n) \{ \cos(2\pi n f_0 t) + j \sin(2\pi n f_0 t) \} = \sum_{n=0}^{N-1} d_n e^{j2\pi n f_0 t} \quad (18.15)$$

By sampling process such as N -points during the one symbol period ($1/f_0$), this equation can be converted as follows.

$$u\left(\frac{k}{Nf_0}\right) = \sum_{n=0}^{N-1} d_n \left(e^{j\frac{2\pi}{N}} \right)^{nk} \quad (k = 0, 1, 2, \dots, N-1) \quad (18.16)$$

As described in previous section, this is IDFT operation. Then the complex digital modulation symbol d_n ($n=0, 1, \dots, N-1$) can be calculated by Inverse IDFT operation, that is, Discrete Fourier Transform (DFT) as shown in equation (18.17).

$$d_l = \frac{1}{N} \sum_{k=0}^{N-1} u\left(\frac{k}{Nf_0}\right) \left(e^{-j\frac{2\pi}{N}} \right)^{lk} \quad (l = 0, 1, 2, \dots, N-1) \quad (18.17)$$

In figure 18.4, simple OFDM receiver block diagram is shown. Here, it is assumed that the received OFDM signal is completely same as the transmitted OFDM signal. However, in real wireless communication, the received signal is distorted. Then the OFDM receiver needs additional signal processing block to remove the distortion such as an equalizer.

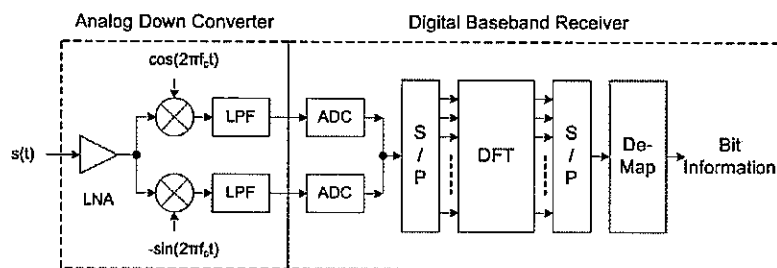


Figure 18.4 Simple OFDM Receiver Block Diagram

18.2.4 Multipath channel

In real wireless communication channel, RF wave propagates in space where refraction and reflection exist. Multiple propagating paths (multipath) exist between transmitter antenna and receiver antenna as shown in figure 18.5. Reflecting path2 and path3 are longer

propagation paths compared with the direct path. Then at the receiver antenna, multiple delayed and phase-rotated signals are overlapped and cause ISI. Since OFDM is especially sensitive to ISI, Guard Interval (GI) is usually inserted between OFDM symbols.

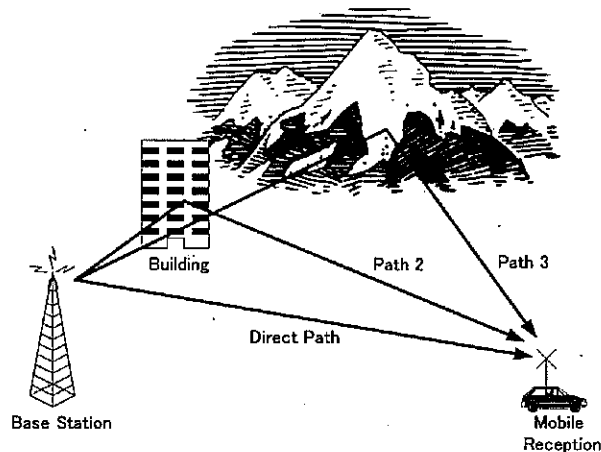


Figure 18.5 Multipath channel

Typical Guard Interval between OFDM symbols is a copy of the tail part of the OFDM symbol as shown in figure 18.6. Please note that the connection point between GI and OFDM symbol is smoothly connected because the all components of OFDM symbol have cycle times of n/f_0 . This type of guard interval is called as cyclic prefix (CP). Since the insertion of guard interval T_g reduces data rate, shorter GI length is preferable. Typical GI length is the 1/16 to 1/4 of OFDM symbol length $T=1/f_0$.

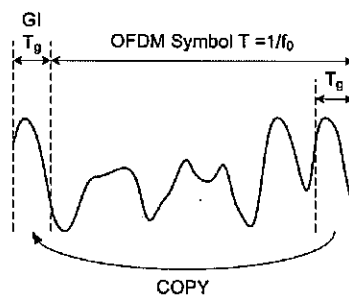


Figure 18.6 Guard Interval

Figure 18.7 shows how GI mitigate the ISI. OFDM receiver performs FFT operation at the FFT window position. If CP type GI is used and a maximum multipath delay path τ is shorter than T_g , ISI can be removed perfectly and the orthogonality between subcarriers is maintained. Then equalizer which removes the multipath distortion becomes very simple.

However If zero signals (zero padding) are used in GI period instead of CP, the orthogonality between subcarriers is destroyed and inter carrier interference (ICI) is generated. Then complicated equalizer is required [3]. In this chapter, only CP type GI is assumed.

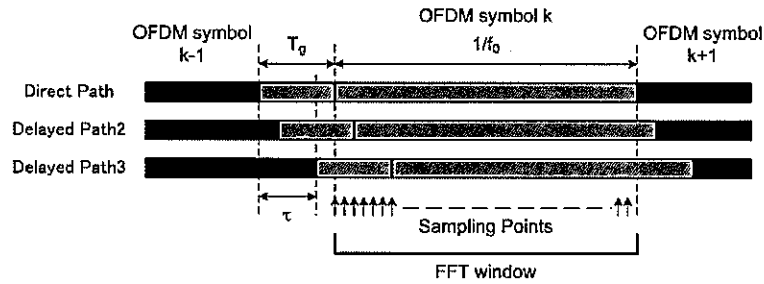


Figure 18.7 Multipath with GI

18.2.5 Channel Estimation and Equalizer

Assuming an impulse response of OFDM signal $s(t)$ transmission channel is $h(t)$, the reception signal $s_R(t)$ can be expressed as follows using convolution.

$$s_R(t) = \int_{-\infty}^{\infty} h(u)s(t-u)du + n_c(t) \quad (18.18)$$

The first term is distortion caused by multipath channel and the second term is noise addition in the OFDM communication frequency band. If the width of channel impulse response $h(t)$ is shorter than GI length T_g , channel transfer function $H(f)$ is given as follows.

$$H(f) = \int_0^{T_g} h(t)e^{-j2\pi ft} dt \quad (18.19)$$

Since convolution operation can be replaced by multiplication after DFT, the following equation can be obtained.

$$x(k,l) = H(f_c + lf_0)d(k,l) + N_B(k,l) \quad (18.20)$$

where, k and l are symbol number and subcarrier number, respectively. The $d(k,l)$ and $x(k,l)$ is complex digital modulation symbol of transmitter and receiver at symbol number of k and subcarrier number of l . $N_B(k,l)$ is the frequency domain complex noise. Since $H(f_c+lf_0)$ is complex number, equation (18.20) means that received symbol $x(k,l)$ is distorted in amplitude and phase from $d(k,l)$ and noise is added. Once channel transfer function $H(f_c+lf_0)$

is obtained, original transmission symbol can be estimated by the following equation. This process is called equalization. Therefore, if the noise is enough small, the equalization process can recover original symbol $d(k,l)$ from received symbol $x(k,l)$.

$$\begin{aligned} \hat{d}(k,l) &= \frac{x(k,l)}{H(f_c + lf_0)} \\ &= d(k,l) + \frac{N_B(k,l)}{H(f_c + lf_0)} \end{aligned} \quad (18.21)$$

18.2.6 OFDM communication and broadcasting systems

Table 18.1 summarizes the state-of-the-art OFDM or OFDMA based communication and broadcasting systems. Wide variety of symbol lengths is used from 3.2us to 1008us for variety of applications. As explained in the previous section, GI length is relating to the multipath delay profile. Then, the larger service area becomes, the longer OFDM symbol length is used. WLAN and three digital TV broadcasting are one-to-one data transfer at a time based on OFDM. XGP [4], LTE[5] and WiMAX are one base station to multiple user terminal communication based on Orthogonal Frequency Division Multiple Access (OFDMA).

Table 18.1 OFDM applications and Effective symbol length

Application of OFDM	Effective OFDM symbol length $1/f_0$
WLAN (802.11g/a/n)	3.2us
XGP (eXtended Global Platform)	26.67us
3.9G mobile phone LTE	66.67 us
WiMAX (802.16e)	102.4us
3.9G mobile phone LTE (Extended mode)	133.33 us
China Digital TV Broadcasting DTMB	500.0us
Europe Digital TV Broadcasting DVB-T	896us (8K mode)
Japan Digital TV Broadcasting ISDB-T	1008us (8K mode)

More detail OFDM communication and broadcasting system block diagram is shown in figure 18.8. For transmitter side, GI insertion is added to figure 18.3. Since receiver side needs synchronization to the transmitter, more blocks are added. Resampler represents

sampling frequency error compensation block. De-rotator rotates baseband OFDM signal to compensate for radio frequency error. The sampling frequency error is caused by system sampling clock accuracy. Then, in the case that enough high accuracy crystal oscillators are used in the system design, resampler might not be needed. However, radio frequency error is caused by the frequency difference between up-conversion and down-conversion. Since the conversion frequency is large such as mega to giga Hz order, the frequency error usually exists and de-rotator is always needed. Those synchronization error compensation issues will be covered in later sections. FFT block needs to know FFT window position as shown in figure 18.7. Then synchronization block also generate FFT window position timing.

- Radio frequency (RF) synchronization
- Sampling rate synchronization
- Symbol timing (FFT window position) synchronization

These three synchronizations have to be considered to implement a real OFDM system. Channel transfer function (CTF) has to be estimated to remove the distortion by multipath channel. In the figure 18.8, following synchronization functions are assumed as

- Time domain RF and sampling rate error detection,
- Time domain symbol timing (FFT window position) detection,
- Frequency domain RF and sampling rate error detection,
- Frequency domain channel estimation.

In addition, time domain channel estimation and frequency domain symbol timing (FFT window position) detection are also possible. Suitable detection and estimation algorithms have to be applied with depending on the applications.

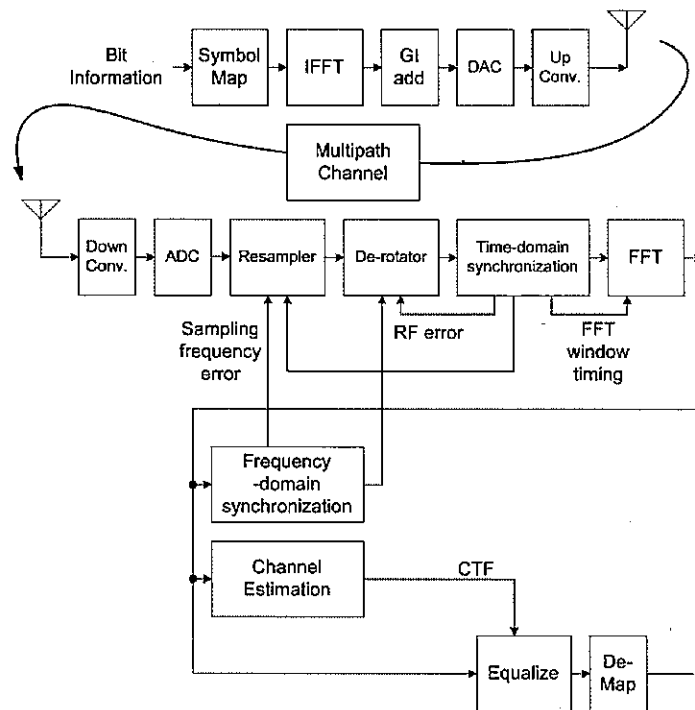


Figure 18.8 OFDM communication and broadcasting system

18.2.7 OFDM WLAN design

18.2.7.1 WLAN system overview

IEEE 802.11 standard WLAN is successfully utilized in market. The first generation IEEE 802.11 standard was made by the end of 1996 to provide 1 or 2 Mbps in the 2.4GHz Industry-Science-Medical (ISM) band. Later, in order to provide higher data rate, OFDM based physical layer standard was established as IEEE 802.11a/g. Since 802.11a/g is based on the distributed carrier sense multiple accesses with collision avoidance (CSMA/CA), in wireless channel only one packet of OFDM signal is transmitted from one to another. Table 18.2 shows IEEE 802.11a/g OFDM based WLAN parameters. It supports from 6Mbps to 54Mbps data rates. Effective OFDM symbol length is 3.2us. Since sampling frequency is 20MHz, 64 sampling points exists in one effective OFDM symbol. Then demodulation process needs 64 point FFT.

Table 18.2 OFDM WLAN parameters (802.11a/g)

Data rate (Mbps)	6	9	12	18	24	36	48	54
Modulation	BPSK	BPSK	QPSK	QPSK	16QAM	16QAM	64QAM	64QAM
Code rate	1/2	3/4	1/2	3/4	1/2	3/4	2/3	3/4
# of subcarriers	52							
# of pilots	4							
OFDM symbol length	4us (effective symbol=3.2us)							
Guard Interval length	0.8us							
Subcarrier spacing	312.5kHz = 1/3.2us							
Sampling rate	20MHz (50ns)							

18.2.7.2 WLAN system architecture

Figure 18.9 shows the frame format of transmission packet. In addition, receiver tasks are indicated in the lower side of the figure. First portion of the frame is short preamble, which is composed of 0.8us pre-determined signals, is 10-times repeated. During the former half periods of short preamble, auto gain control (AGC) value to the low noise amplifier (LNA) in the analog front-end has to be determined. Since channel estimation is performed only once in the long preamble period in WLAN system, AGC control value has to be fixed in the short preamble period and has to be kept until the end of frame. Then during the latter half period of the short preamble, RF error has to be measured and compensated by de-rotator. Detail of the RF error detection will be explained in section 18.4. Then the end point of short preamble has to be detected to get FFT window position for later OFDM symbols. In long preamble, there are two identical 3.2us OFDM symbols T1, T2 and 1.6us GI. Using T1 to T2 rotation, more precise RF error can be estimated (see section 18.4.1). Since the T1 and T2 is predetermined pilot OFDM symbol, CTF can be estimated from FFT output of T2. This CTF value will be used in the following OFDM symbol equalization until the end of frame since WLAN assumes a NOT moving application such as fixed channel.

As for sampling frequency error, it is not easy to detect the error and compensate for it during the preambles since the preamble duration is limited. However, if less than 100ppm error crystal is used, maximum $64 \times 100 \text{ppm} = 0.0064$ points sampling error is caused during 3.2us one OFDM symbol. This 0.0064 points error in one OFDM symbol is enough small than 1 % error, then this sampling clock error does not cause ICI problem. The 0.0064 points error will be accumulated in following maximum 4K of OFDM data symbols. Then the accumulation of point shift will be $4K \times 0.0064 = 26$ points. Since the GI length is 16 points, this FFT window shift has to be taken care. Then, the receiver needs to detect FFT window shift and compensate for it. Since the CTF value is estimated by T1, T2 at the long preamble and it is used in equalization process, the equalized output will be rotated by the FFT

window position shift. Then, after conventional equalizer (EQ), FFT window shift detector and equalizer output compensator (EQC) is needed as shown in figure 18.10. If high precise crystal is used, this special detector and compensator are not needed. However, analog down conversion LO frequency might drift during the 4K OFDM symbols, and then RF error tracking mechanism should be needed.

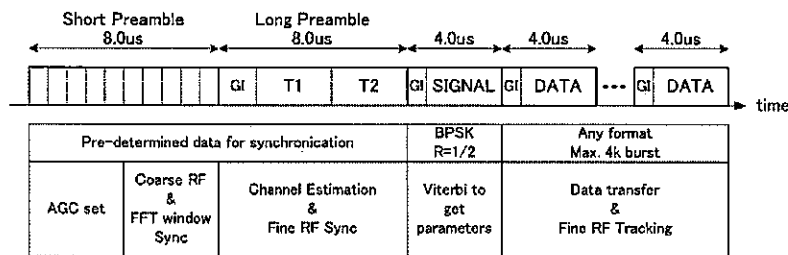


Figure 18.9 WLAN frame format and receiver tasks

Figure 18.10 shows the block diagram of WLAN transceiver. RX is an analog frontend. By calculating the short preamble power, AGC feedback value to the LNA is determined. In the diagram, decimate and resample block is assumed to be fixed rate. Since FFT size is small as 64, sampling clock frequency error does not have significant impact on ICI in one OFDM symbol. But an accumulated sampling error causes the FFT window position shift, which causes the equalizer output constellation rotation depending on the subcarrier number. Then in this design, the rotation compensation block EQC is included. The EQC process is further explained in section 18.4.6. After the de-mapping process, forward error correction for convolutional code is performed using famous Viterbi algorithm.

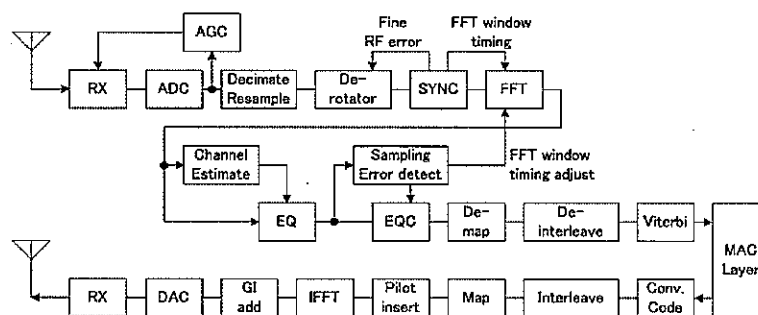


Figure 18.10 WLAN

18.2.8 OFDMA transceiver design

18.2.8.1 OFDMA transceiver system overview

OFDMA represents Orthogonal Frequency Division Multiple Access. Then multiple users have to share OFDM subcarriers in a symbol. Figure 18.11 shows an example of wireless multiple access for mobile phones. First, Up-Link (UL) and Down-Link (DL) have to be separated. This separation is usually implemented by Frequency Division Duplex (FDD) or Time Division Duplex (TDD). FDD needs two separate frequency bands although TDD has to share one frequency band by time slot. Secondly, multiple user communications have to be separated. This second separation also can be done by FDD or TDD method. However, this inter-user separation can be realized by OFDM technology, which is so called OFDMA.

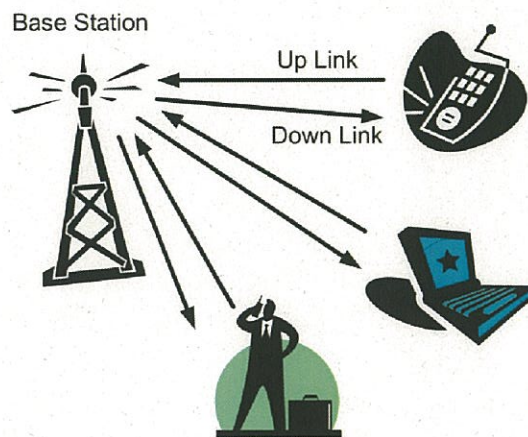


Figure 18.11 Multiple access

In OFDMA communication system, many subcarriers are divided into plural clusters as shown in figure 18.12. Those clusters are assigned to each user. During DL communication, all clusters are transmitted from base station to users. Each user receives all clusters and demodulates assigned clusters. Instead, in UL communication each user generates pre-assigned cluster subcarriers and base station receives summed cluster subcarriers from all users.

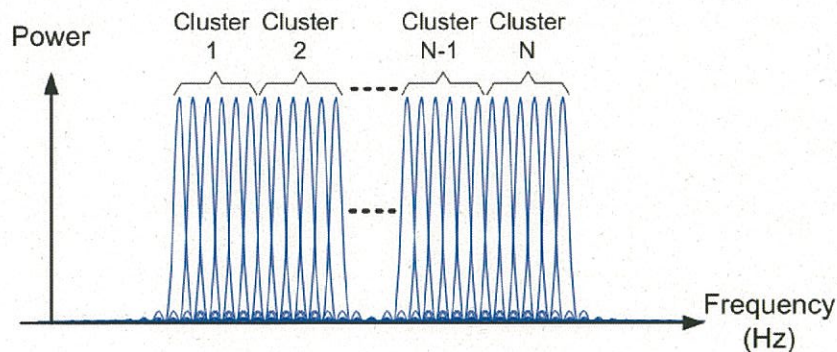


Figure 18.12 OFDMA power spectrums

Figure 18.13 shows TDD Frame structure for one base station (BS) and two user terminals U1, U2. Delay time1 and 2 are wireless propagation delays between BS and U1, U2 respectively. Since UL OFDM symbol has to reach BS with same timing, each user has to transmit UL signal in advance timing as shown in the figure. BS measures the UL arrival signal timing from each user and gives feedback to each user. This timing measurement will be further explained in section 18.4.5.

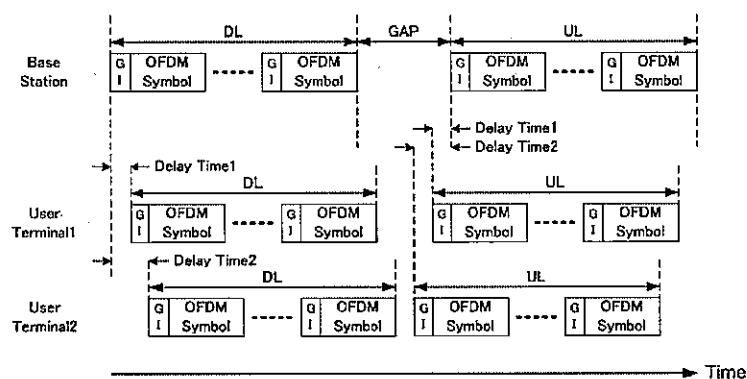


Figure 18.13 TDD Frame and GAP

18.2.8.2 SC-FDMA transceiver overview

One famous problem of OFDM/OFDMA is large Peak to Average Power Ratio (PAPR), since the transmit waveform is generated by IFFT of many complex digital modulation symbols. The large PAPR needs a wide range of linearity in transmission power amplifier and causes higher power dissipation at transmitter. Especially, this will be a problem in battery operated handsets. SC-FDMA (Single-Carrier Frequency Division Multiple Access) system tries to reduce PAPR while using same OFDMA frequency domain equalization technique. Figure 18.14 shows a comparison between OFDMA and SC-FDMA system. In OFDMA system, bit information is mapped to complex digital modulation symbols such as QPSK / 16QAM / 64QAM constellation. Then the output of IFFT might have large peak value. In SC-FDMA system, bit information is mapped to complex digital modulation symbols and the N symbols are processed by N-points FFT. The N-point FFT outputs are bundled to a block. Then the K blocks are connected and mapped to M-points IFFT inputs. Here, assume $M=N \times K$. In this mapping process, each block corresponds to different subcarrier frequencies. Consequently, at the output of M-points IFFT is a sum of frequency shifted N-FFT input sets. Since the input of N-FFT is determined by complex digital modulation symbols, the PAPR of M-IFFT output is reduced. Please note that GI is also used in SC-FDMA although it is not indicated in figure 18.14.

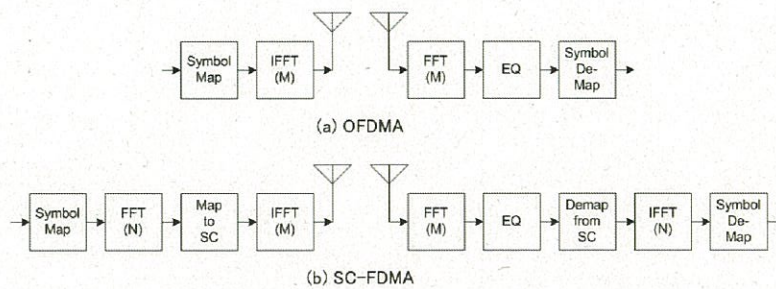


Figure 18.14 simplified OFDMA and SC-FDMA

In another word, SC-FDMA uses Spread Spectrum method by N-FFT. Then the spread spectrum is assigned onto many subcarrier frequencies and transmitted parallel in channel. At the receiver, frequency domain channel estimation and equalization is performed. Then the spread spectrum is re-converted to original complex digital modulation symbols by N-IFFT. The mapping process between N-FFT and M-IFFT is equivalent to assign frequency bands to the spread spectrums.

18.2.8.3 OFDMA transceiver architecture

In this section, OFDMA downlink and SC-FDMA uplink are assumed since SC-FDMA transmitter is good for battery operated mobile terminals. Figure 18.15 shows a Base Station Transmitter (TX) and Receiver (RX) block diagram. In TX, OFDMA transmitter is used and in RX, SC-FDMA receiver is used. Since all user terminals communicating with the same BS have to synchronize to the Base Station, a Base Station receiver is not required to have sampling clock frequency error compensation and RF error compensation. Since FFT window position timing is determined in a BS, all user terminals have to adjust the UL transmission timing by the control of the BS. Then the BS needs to measure the UL arrival signal timing from each user and gives feedback to each user. This timing measurement will be further explained in section 18.4.5.

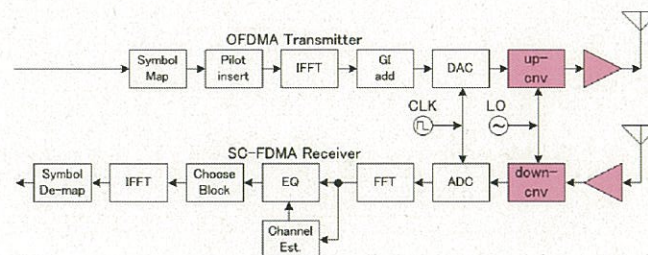


Figure 18.15 Base Station TX and RX

Figure 18.16 shows a user terminal transmitter and receiver block diagram. In the receiver

side, first AGC block controls the LNA in analog front end. There are two local oscillators. The one is for RF conversion and the other is for ADC/DAC. In order to compensate for the RF error and sampling clock frequency error, resampler and de-rotator are required. In the figure, NCO represents Numerical Control Oscillator, which is digital form of Voltage Controlled Oscillator (VCO). The NCO frequency corresponds to RF error. In the transmitter side, the reverse operation of the resampler and the de-rotator are required and those are interpolator and rotator, respectively. Control signals for those blocks are generated in receiver side. Power amplifier (PA) gain in analog front end is controlled by the Base Station. One more Base Station control item is UL start timing.

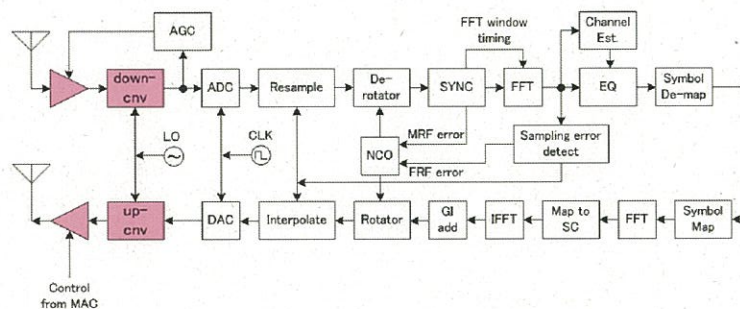


Figure 18.16 User Terminal TX and RX

We have discussed various OFDM based transceivers and design. In the following sections, related error detection and compensation (see 18.4) together with diversity techniques (see 18.3) will be addressed.

18.3 Adaptive Antenna Array and Diversity

18.3.1 Introduction to multiple antenna receivers

When an OFDM receiver is used in moving vehicles such as automobiles etc., radio frequency wave suffers from Doppler frequency shift as the automobile is moving as shown in figure 18.17. The Doppler shift destroys the orthogonality between OFDM subcarrier signals and increases an inter carrier interference. Longer multipath might cause inter symbol interference and weak reception signal power might decrease Signal to Noise ratio (SNR) so as to degrade the reception performance. Therefore, it is a severe challenge to maintain the reception quality in mobility circumstances.

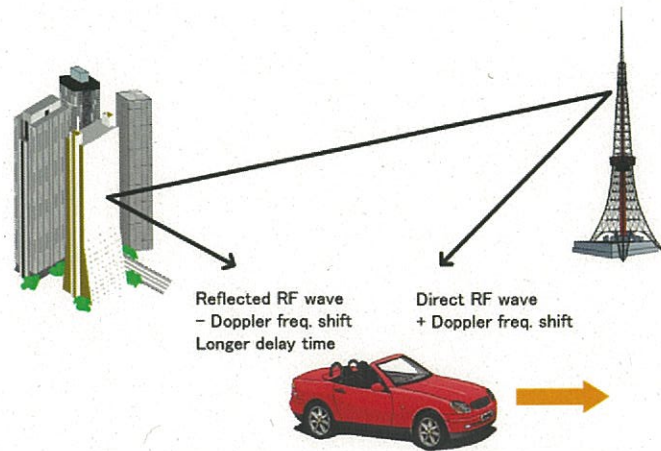


Figure 18.17 Problems in mobile reception

One well-known way to improve the performance of OFDM receiver is to exploit a spatial diversity by utilizing multiple antenna elements. There are two approaches to utilize the array antenna for multi-carrier transmission method, which are pre-FFT adaptive array (AA) antenna and post-FFT carrier diversity (CD) combining scheme. The AA scheme is a conventional method to employ the array antenna in which inputs from the array antenna are combined before OFDM demodulation [6-9] as shown in figure 18.18(a). Since this approach uses the one set of coefficients such as $w_1(t)$ and $w_2(t)$ for each OFDM symbol, it is an attractive solution in term of low computation requirement. However, in order to have a good performance, received antenna signals should have higher correlation.

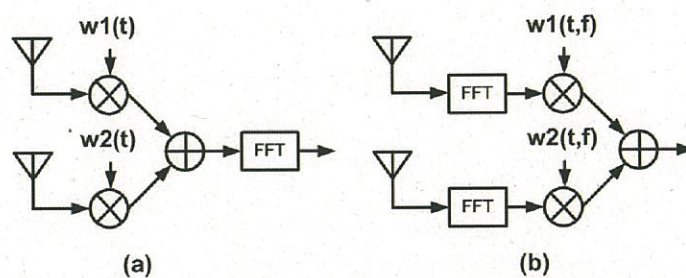


Figure 18.18 Two types of space diversity for OFDM communication

On the other hand, post-FFT scheme is an advanced method to utilize the array antenna for multi-carrier system as shown in figure 18.18(b). Instead of combining before OFDM demodulation, inputs are demodulated using multiple OFDM demodulations. Subcarriers are then combined accordingly in frequency domain using diversity combining method [10]. Post-FFT scheme therefore is an optimum approach in term of maximizing SNR for each OFDM subcarriers. However, its computation complexity grows drastically as more antenna

elements associated. The feature of two diversity methods is summarized in Table 18.3.

Table 18.3 Feature of two diversities

Type	Circuit size	Reception performance	Restriction
Adaptive Array (AA)	SMALL	MODERATE	High antenna signal correlation
Carrier Diversity (CD)	LARGE	HIGH	NONE

18.3.2 Time-domain diversity

18.3.2.1 OFDM-directed array antenna system

An adaptive array antenna is considered as an attractive solution to suppress interferences and to enhance the quality of reception in the multipath environment [11, 12]. It usually requires in advance the knowledge of the characteristics of the desired signals to distinguish between them and undesired signals. However, due to their high complexity, these approaches cost lots of computations and efforts from a viewpoint of hardware implementation. In this section, OFDM-directed two array antenna Digital Beam-Forming (DBF) algorithms with lower computation requirements are introduced. The DBF algorithms make use of cyclic prefix's periodic property. Figure 18.19(a) and (b) illustrate the periodic property of OFDM signal and the principle of the time-domain adaptive array antenna system. For convenience, here, GI with duration of T_g is referred as "HeadGI" and the last part of the effective OFDM symbol with duration of T_g is referred as "TailGI".

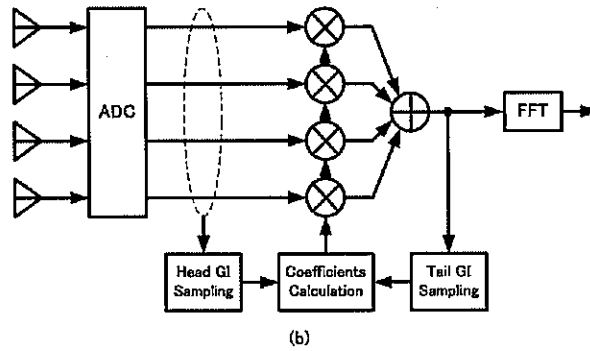
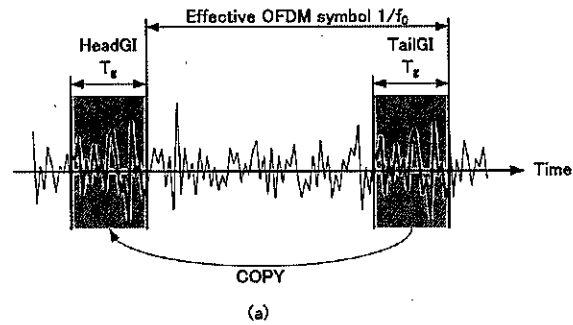


Figure 18.19 (a) Cyclic properties of OFDM signal and (b) Adaptive array using GI

Suppose that an array antenna is equipped with M elements. The vector input signals of the array antenna can be expressed as

$$\begin{aligned} \mathbf{X}(t) &= [x_1(t) \ x_2(t) \ \dots \ x_M(t)]^T \\ &= \mathbf{V}_s s(t) + \mathbf{V}_i i(t) + \mathbf{N}(t) \end{aligned} \quad (18.22)$$

where superscript $[\cdot]^T$ denotes a transpose of matrix. $s(t)$ and $i(t)$ are desired signal and interference, respectively. \mathbf{V}_s and \mathbf{V}_i are the array response vectors of desired signal and interference, respectively. $\mathbf{N}(t)$ is AWGN vector of array antenna. Hence, the output of the array antenna system is calculated as follows

$$\begin{aligned} y(t) &= \mathbf{W}^H \mathbf{X}(t) \\ &= \mathbf{W}^H \mathbf{V}_s s(t) + \mathbf{W}^H \mathbf{V}_i i(t) + \mathbf{W}^H \mathbf{N}(t) \\ &= a_s s(t) + a_i i(t) + n(t) \end{aligned} \quad (18.23)$$

where superscript $[\cdot]^H$ denotes Hermitian. $\mathbf{W} = [w_1 \ w_2 \ \dots \ w_M]^T$ is coefficient vector of the array antenna. $a_s = \mathbf{W}^H \mathbf{V}_s$ and $a_i = \mathbf{W}^H \mathbf{V}_i$ are constant during one OFDM symbol.

In References [13-14], the conventional Maximum Ratio Combining (MRC) algorithm is

derived directly from the cross-correlation of inputs and outputs of the array antenna as follows

$$\begin{aligned}
\mathbf{W}_{\text{convMRC}} &= E[\mathbf{X}(t)y^*(t)] \\
&= E[(\mathbf{V}_s s(t) + \mathbf{V}_i i(t) + \mathbf{N}(t))(a_s s(t) + a_i i(t) + n(t))^*] \\
&= a_s^* P_s \mathbf{V}_s + a_i^* P_i \mathbf{V}_i
\end{aligned} \tag{18.24}$$

where $P_s = E[|s(t)|^2]$ and $P_i = E[|i(t)|^2]$. Superscripts $(.)^*$ denotes complex conjugate. Apparently from equation (18.24), the array antenna utilizing the conventional MRC combines not only the desired signal but also the interference. Under the circumstances that the power of interference is stronger than that of the desired signal, the interference will be emphasized at the output of the array antenna.

18.3.2.2 OFDM selective Maximum Ration Combining (MRC)

In order to overcome the limitation of the conventional MRC method described above, by making use of HeadGI of input and feedback of TailGI of output, the cross correlation of input and output is derived as

$$\begin{aligned}
\mathbf{r}_{xy} &= E[\mathbf{X}_h(t) y_t^*(t)] \\
&= E[(\mathbf{V}_s s_h(t) + \mathbf{V}_i i_h(t) + \mathbf{N}_h(t))(a_s s_t(t) + a_i i_t(t) + n_t(t))^*] \\
&= a_s^* \mathbf{V}_s E[s_h(t) s_t^*(t)] + a_i^* \mathbf{V}_i E[i_h(t) i_t^*(t)]
\end{aligned} \tag{18.25}$$

where subscripts $(.)_h$ and $(.)_t$ denote HeadGI and TailGI of signal, respectively. Since $i(t)$ is interference and does not have cyclic prefix GI, the second term of equation (18.25) can be neglected. Employing the periodic property of OFDM signal, equation (18.25) can be expressed as follows.

$$\mathbf{r}_{xy} = a_s^* P_s \mathbf{V}_s \tag{18.26}$$

Hence, equation (18.26) presents the information of AOA of desired signal without the reference signals. Therefore, the coefficients vector of array antenna using MRC algorithm is derived as follows. In another word, this OFDM selective MRC algorithm is beam-forming only to OFDM signals.

$$\mathbf{W}_{\text{MRC}} = \text{normalize}(\mathbf{r}_{xy}) \tag{18.27}$$

18.3.2.3 Minimum Mean Square Error (MMSE) and Sampled Matrix Inversion (SMI)

Although the OFDM selective MRC algorithm emphasizes the desired OFDM signal, it does not minimize the interference. In this section, Minimum Mean Square Error (MMSE) algorithm is derived. Assume a reference signal is $r(t)$. The error signal is given as follows.

$$e(t) = r(t) - \mathbf{W}^H \mathbf{X}(t) \quad (18.28)$$

MSE (Mean Square Error) between the output of the array antenna and the reference signal is given as follows.

$$\text{MSE} = E[|e(t)|^2] = E[|e(t)r(t) - \mathbf{W}^H \mathbf{X}(t)|^2] \quad (18.29)$$

By minimizing MSE with well-known Wiener-Hoff equation, optimum coefficients are derived as

$$\mathbf{W}_{\text{MMSE}} = \text{normalize}(\mathbf{R}_{\text{xx}}^{-1} \mathbf{r}_{\text{xy}}) \quad (18.30)$$

where $\mathbf{R}_{\text{xx}} = E[\mathbf{X}_h \mathbf{X}_h^H]$ is auto correlation matrix of inputs. In real applications, \mathbf{R}_{xx} can be obtained by a simple averaging scheme [15]. According to the equation (18.30), additional calculation of inverse of \mathbf{R}_{xx} makes MMSE possible and then this algorithm is also called as Sampled Matrix Inversion (SMI).

18.3.3 Frequency-domain diversity

18.3.3.1 MRC for uniformly distributed branch noise

Figure 18.20 illustrates the post-FFT Carrier Diversity (CD) combiner for OFDM receiver. This system assumes there are M antennas and M OFDM demodulations. The combined symbol at k -th subcarrier of i -th OFDM symbol is given as follows

$$Z(i, k) = \sum_{l=1}^M C_l(i, k) Y_l(i, k) \quad (18.31)$$

where $C_l(i, k)$ and $Y_l(i, k)$ are combining coefficient and demodulated symbol corresponding to l -th branch of combiner at k -th subcarrier of i -th OFDM symbol, respectively. Here, the following equation is assumed

$$Y_l(i, k) = H_l(i, k)d(i, k) + N_{Bl}(i, k) \quad (18.32)$$

where $H_l(i, k)$, $N_{Bl}(i, k)$ are channel transfer function and additive noise corresponding to l -th branch of combiner at k -th subcarrier of i -th OFDM symbol, respectively and $d(i, k)$ is the transmitted complex digital modulation symbol at k -th subcarrier of i -th OFDM symbol. According to Reference [10], the combining coefficients for MRC algorithm can be obtained as follows

$$C_l(i, k) = \frac{H_l^*(i, k)}{\sum_{l=1}^M |H_l(i, k)|^2} \quad (18.33)$$

where assuming that the mean of $N_{Bl}(i, k) = 0$ for all l, i, k and the deviation of $N_{Bl}(i, k)$ are all the same for all l, i, k . By using MRC algorithm, the optimum combined output SNR can be achieved for each subcarrier by post-FFT CD scheme.

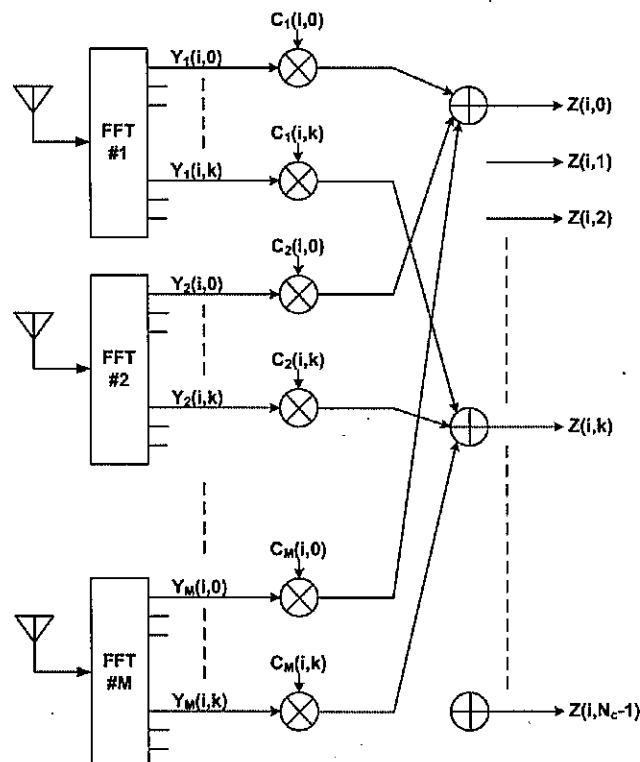


Figure 18.20 Carrier diversity combiner for OFDM receiver

18.3.3.2 MRC for independently distributed branch noise

When one antenna of the receiver is broken or hidden, the noise of the branch increases

and has impact on the combined output. Figure 18.21 shows a typical 4 space diversity antenna configuration for an automobile. Two antenna elements are set in the front and the two are set in the rear. However as the body of the automobile is made from metal, the directional characteristics of the front antenna is obviously distorted and concentrates in the forward direction. Then, the uniform noise power distribution assumed in previous section sometimes cannot be applied to these kinds of unbalanced antenna situations.

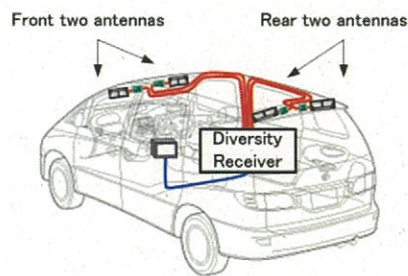


Figure 18.21 Multiple antennas of a car

Equation (18.33) can be rewritten as follows.

$$\begin{aligned}
 C_1(i, k) &= \frac{H_1^*(i, k)}{|H_1(i, k)|^2} \cdot \frac{|H_1(i, k)|^2}{\sum_{l=1}^M |H_l(i, k)|^2} \\
 &= \frac{1}{H_1(i, k)} \cdot \frac{|H_1(i, k)|^2}{|H_1(i, k)|^2 + |H_2(i, k)|^2 + \dots + |H_M(i, k)|^2} \quad (18.34)
 \end{aligned}$$

The first division by $H_l(i, k)$ corresponds to the equalizing operation. The second multiplication term means how much percent contribution of the branch l to combining. Since $H_l(i, k)$ is CTF, then it corresponds to the signal gain in the transmission channel. When each branch ($l=1 \dots M$) average noise power is shown by σ_{nl}^2 . The equation can be modified to the following equation (18.35), which is taking noise distribution between branches into account.

$$C_1(i, k) = \frac{1}{H_1(i, k)} \cdot \frac{|H_1(i, k)|^2 / \sigma_{n1}^2}{|H_1(i, k)|^2 / \sigma_{n1}^2 + |H_2(i, k)|^2 / \sigma_{n2}^2 + \dots + |H_M(i, k)|^2 / \sigma_{nM}^2} \quad (18.35)$$

According to equation (18.35), when noise power at branch l increases, the combining coefficients $C_l(i, k)$ becomes small to have lesser contribution to the combining. When using the equation (18.35) in real application, additional noise power estimation is needed.

18.3.4 Hybrid diversity

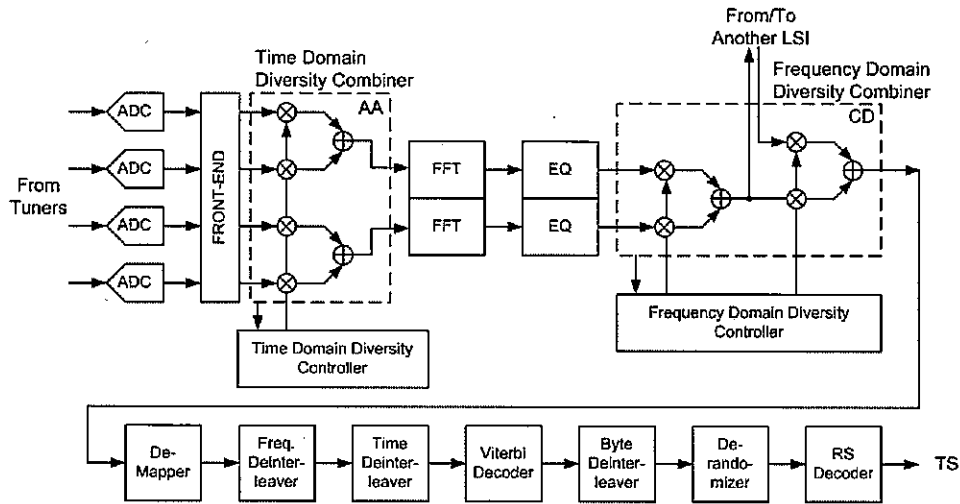


Figure 18.22 4 antennas Hybrid diversity combining receiver

Figure 18.22 shows a block diagram of 4 antenna diversity combining receiver LSI for Japan OFDM based Digital TV [16]. In the first stage, two sets of two antenna MRC combiners are used. Then there are two sets of FFT and Equalizer circuits. In the second stage, the two EQ outputs are combined as a carrier diversity method. The final combined output is de-mapped and error correction processing is applied. The two stage diversity combiners halve the number of FFT/EQ circuits from 4 to 2 and sustain similar mobile reception performance comparing with 4 FFT/EQs case. In the second CD stage, another LSI interface is prepared then totally 4 kinds of diversity configurations are possible as shown in figure 18.23.

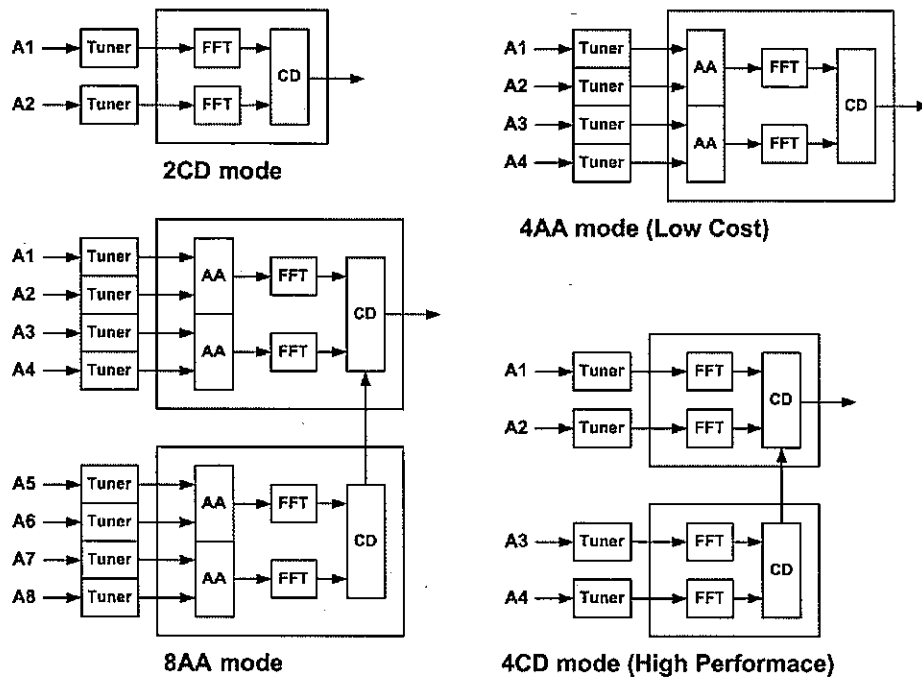


Figure 18.23 Four possible diversity configurations

Mobile reception performance was measured by field experiment at three severe test courses in Osaka Japan for 4CD, 4AA, 2CD configurations. The experimented UHF channel is 13ch which is broadcasted from the top of IKOMA Mountain between Osaka and Nara prefecture. 64QAM, code rate of 3/4, 12 segments HDTV broadcasting service are used. The three test courses are suburban area 50km from IKOMA broadcasting station (BS), mountain area 40km from BS, and urban area but shadow of the hill 20km from BS. Figure 18.24 shows the reception rate (error free reception duration to the total experiment duration) of those test courses. One chip solution 4AA mode showed a little lower performance to the two chip solution 4CD mode. Hybrid method shows a little lower reception performance with lower system cost. More comparison results are available in [16].

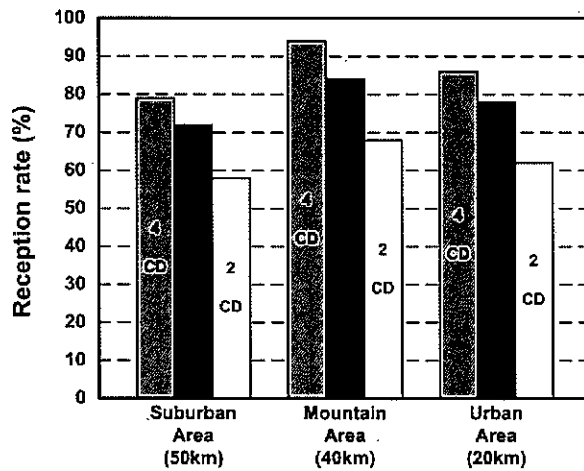


Figure 18.24 Field experimental results

Various diversity techniques for OFDM receivers are addressed in this section. In the next section, related error detection and compensation techniques for OFDM based transceivers will be discussed.

18.4 Error Detection and Compensation

18.4.1 Time domain RF error detection and compensation

Figure 18.25 shows WLAN frame format. At the short preamble period, there are 10 repeating $1/(4f_0)$ length signals and at the long preamble, T1 and T2 are $1/f_0$ length repeating signals. At the data burst, GI is a cyclic prefix, that is, a copy of tail part of OFDM symbol. By making use of these repeating or copy signals, RF error can be computed as follows.

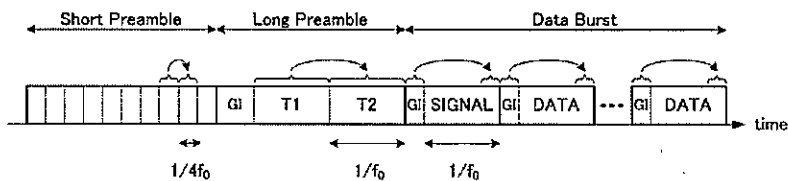


Figure 18.25 Time domain RF error detection

If the time interval of identical signal $g(t)$ and $g(t+T)$ is T , RF error Δf causes the rotation as follows.

$$g(t + T) = g(t)e^{j2\pi\Delta f T} \quad (18.36)$$

From equation (18.36), Δf can be solved as follows.

$$\Delta f = \frac{1}{2\pi T} \tan^{-1}[g(t + T)g^*(t)] \quad (18.37)$$

$$-\frac{1}{4T} < \Delta f < \frac{1}{4T} \quad (18.38)$$

At the short preamble, $T=1/(4f_0)$. Then, RF error Δf in the region $[-f_0, f_0]$ can be detected. Similarly at the preamble region, Δf in the region of $[-f_0/4, f_0/4]$ can be estimated. Since T1, T2 has 64 points whose average reduces the Δf detection error, the high accuracy Δf can be obtained. At the data burst region, Δf should be very small because of preceding Δf detection and compensation. Then equation (18.37) can be approximated as follows.

$$\Delta f \cong \frac{1}{2\pi T} \text{Im}[g(t + T)g^*(t)] \quad (18.39)$$

During the data burst region, RF error can be compensated by feedback loop as shown in figure 18.26. In the loop, loop gain α and accumulator are used. When the RF error detector output = 0, the accumulator holds the same value since the input = 0. The accumulator output corresponds to RF error Δf .

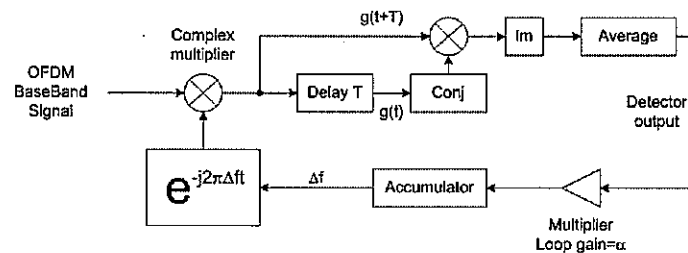


Figure 18.26 Time domain RF error compensation feedback

18.4.2 Time domain FFT window position detection and RF error detection

FFT window position can be found by cyclic prefix GI. According to Reference [17], Head and Tail GI correlation can be calculated as follows.

$$A_s(k) = \sum_{i=0}^{M-1} \{|r(k+i)|^2 + |r(k+i+N_{PD})|^2\} - 2 \left| \sum_{i=0}^{M-1} r(k+i+N_{PD}) \cdot r^*(k+i) \right| \quad (18.40)$$

where $r(k)$ is the baseband OFDM signal input. Index i is used to average M points. If time index k is not in GI period, $r(k+N_{PD})$ and $r(k)$ are independent. Then the second term will be close to zero. Instead, if time index k is in GI period, $r(k+N_{PD})$ and $r(k)$ should be close. Then the output of equation (18.40) will be close to zero. Consequently, the correlation $A_s(k)$ waveform is negative pulse at GI period as shown in figure 18.27 (a). Here, N_{PD} is the OFDM symbol length.

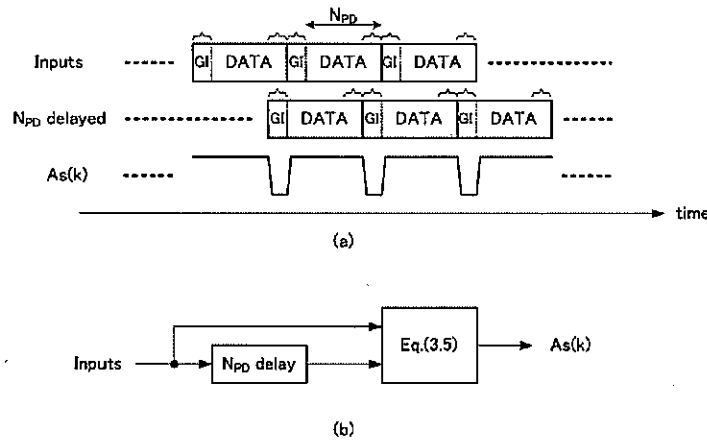


Figure 18.27 Cyclic prefix GI time domain correlation

The negative pulse will be repeatedly generated. By detecting the pulse rising edge, OFDM starting position can be detected and then FFT window position is obtained. Since the negative pulse period corresponds to the GI period, phase rotation φ in one OFDM symbol can be obtained as follows.

$$\varphi = \text{angle} \left\{ \sum_{i=0}^{M-1} r(k+i+N_{PD}) \cdot r^*(k+i) \right\} \quad (18.41)$$

Since the phase rotation φ happens in OFDM symbol period ($1/f_0$), RF error Δf can be calculated as follows.

$$\Delta f = \frac{\varphi}{2\pi} f_0 \quad (18.42)$$

18.4.3 Time domain RF error, sampling error, and symbol timing detection and compensation

One subcarrier signal with index of n can be expressed as follows.

$$x_n(t) = d_n \cdot e^{j2\pi n f_0 t} \quad (18.43)$$

where d_n is n -th digital modulation symbol and f_0 is carrier spacing. Then $t=0$ to $1/f_0$ rotation with RF error Δf and sampling clock error β is can expressed as follows.

$$\begin{aligned} \frac{x_n\left(\frac{1}{f_0}\right)}{x_n(0)} &= e^{\frac{j2\pi(nf_0 + \Delta f)(1-\beta)}{f_0}} \\ &\cong e^{j2\pi(\alpha - n\beta)} \end{aligned} \quad (18.44)$$

Here α is normalized RF error of $\Delta f/f_0$. According to the equation (18.44), RF error causes common phase rotation but sampling clock error β causes the phase error in proportional to subcarrier index number n as shown in figure 18.28.

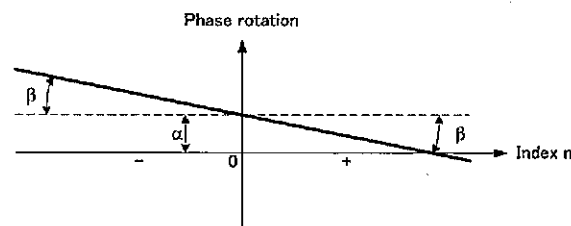


Figure 18.28 Phase rotation with RF error α and sampling clock error β

Then by averaging all phase rotation for all subcarriers, RF error α can be obtained. By calculating phase error in positive frequency ($n>0$) – phase error in negative frequency ($n<0$), sampling clock error β can be calculated.

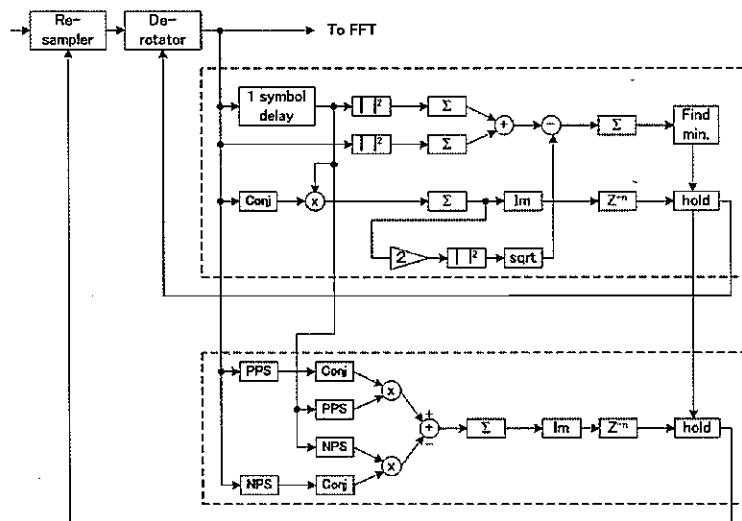


Figure 18.29 Time domain RF error and sampling clock error compensation

Figure 18.29 shows one example of time domain GI correlation, RF error detection and sampling clock error detection circuit diagram. The "Find min." block detects minimum points of GI correlation given in equation (18.40). The upper dashed block outputs RF error and the lower dashed block outputs sampling clock error. PPS and NPS represent a positive and a negative pass FIR filters, respectively. Since they form feedback loop and the detection error should be close to zero in converged stable state, the imaginary components of the phase errors are used to feedback outputs. The simplest implementation of PPS and NPS FIR filters' tap coefficients are given as $[-j, 1, j]$ and $[j, 1, -j]$ respectively. Here, j represents a square root of (-1) .

18.4.4 Frequency domain RF error and sampling error detection

Figure 18.30 shows one Physical Resource Unit (PRU) of OFDMA based XGP communication system [4]. It corresponds to 19 OFDM symbols and 24 subcarriers. In one PRU, there are 24 pilots, which transmit pre-determined value. For the same subcarrier number such as F3, F7 and so on, the pilot values are identical and then the phase rotation shown in solid or dashed arrows can be detected. In one PRU, there are 18 phase rotations. The averages of the phase rotation in lower frequency side shown in dashed arrows and in higher frequency side shown in solid arrows can be expressed as follows.

$$\text{LowRotMean} = \frac{p(F3, S5)^* \times p(F3, S9) + \dots + p(F11, S13)^* \times p(F11, S17)}{9} \quad (18.45)$$

$$\text{HighRotMean} = \frac{p(\text{F15}, \text{S5})^* \times p(\text{F15}, \text{S9}) + \dots + p(\text{F23}, \text{S13})^* \times p(\text{F23}, \text{S17})}{9} \quad (18.46)$$

where $p(\text{F}, \text{S})$ means pilot subcarrier value at subcarrier of F and symbol number of S .

If we define LV and HV as

$$\text{LV} = \frac{\text{imag}(\text{LowRotMean})}{\text{real}(\text{LowRotMean})} \quad (18.47)$$

$$\text{HV} = \frac{\text{imag}(\text{HighRotMean})}{\text{real}(\text{HighRotMean})} \quad (18.48),$$

RF error FRF in equation (18.49) and sampling clock error CLKD in equation (18.50) can be calculated as follows.

$$\text{FRF} = \frac{m \cdot \text{LV} - n \cdot \text{HV}}{\frac{9}{4}\pi(m - n)} \quad (18.49)$$

$$\text{CLKD} = \frac{\text{LV} - \text{HV}}{\frac{9}{4}\pi(m - n)} \quad (18.50)$$

where n and m are the center carrier index numbers at lower frequency side and higher frequency side, respectively. The idea of the calculation is based on the property that RF error rotates the same phase for all subcarriers and that sampling clock error rotates each subcarrier phase in proportional to the carrier number index as shown in figure 18.28.

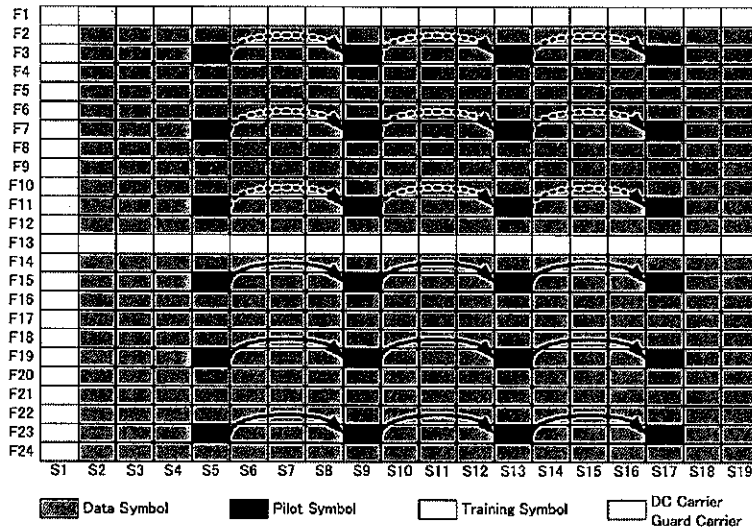


Figure 18.30 XGP PRU structure

18.4.5 Delay profile detection and frequency domain FFT position detection

In frequency domain, from series of pilot or training symbols, delay profile can be easily obtained. S1 symbols of PRU shown in figure 18.30 are training symbols and their values are all pre-determined. Then by IFFT calculation, delay profile can be obtained as follows.

Delay Profile

$$= \text{IFFT} \left[0, \frac{X(F2)}{T(F2)}, \frac{X(F3)}{T(F3)}, \dots, \frac{X(F12)}{T(F12)}, 0, \frac{X(F14)}{T(F14)}, \frac{X(F15)}{T(F15)}, \dots, \frac{X(F24)}{T(F24)}, 0, \dots, 0 \right] \quad (18.51)$$

where $X(F)$ is FFT output of subcarrier F and symbols $S1$, and $T(F)$ is the pre-determined training value at subcarrier F and symbols $S1$. Then power profile can be obtained as shown in figure 18.31. In this profile, the power of the main wave is at index=0 and then FFT window position is placed at the main wave position. Because of cyclic property of IFFT and FFT, the tail profile is connecting to the head profile as shown in figure 18.31.

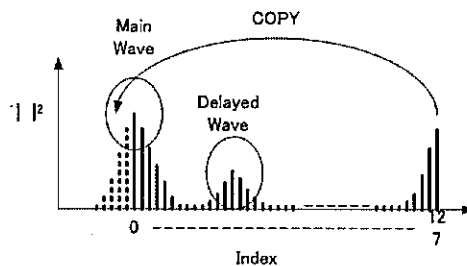


Figure 18.31 Power delay profile

If FFT window position shifts ahead, the delay profile shift backward as shown in figure 18.32. According to the index number of the peak point of delay profile, FFT window shift can be detected. Detectable maximum delay is the OFDM symbol length divided by training subcarrier interval. Since S1 symbol has all training symbol and the interval=1, the detectable maximum delay is the same as OFDM symbol length. If pilots in S5 are used to calculate delay profile, the detectable maximum delay is 1/4 of OFDM symbol because the pilot interval = 4.

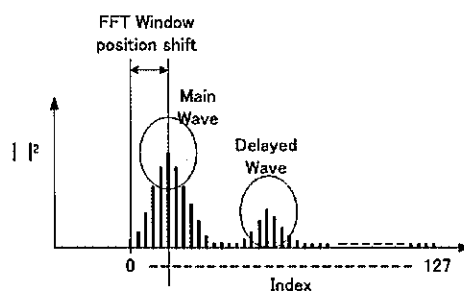


Figure 18.32 FFT window position shift ahead

18.4.6 FFT window shift detection by constellation rotation and compensation

Subcarrier signal of index= n can be expressed as equation (18.52) in one OFDM symbol $[0, 1/f_0]$. The FFT output of the OFDM symbol with index= n is d_n .

$$x_n(t) = d_n \cdot e^{j2\pi n f_0 t} \quad (18.52)$$

When FFT window shifts k points ($-0.5 < k < 0.5$) forward direction, the equivalent time shift Δt is as follows.

$$\Delta t = \frac{k}{N f_0} \quad (18.53)$$

Then the FFT output of the OFDM symbol with index= n will become as follows.

$$d_n \cdot e^{j2\pi \frac{nk}{N}} \quad (18.54)$$

where N is the size of FFT. This means by FFT window shift causes the constellation rotation which is proportional to the index number n . Then by detecting the equalizer

output's rotation, FFT window shift can be detected. Figure 18.33 shows a block diagram of equalizer output phase rotation detection and compensation. Since the equalizer output should be pre-determined constellation based, phase rotation can be detected with many sample average. If averaging all the points is performed, common phase error CPH in the figure can be detected. If the difference between positive and negative index averages is calculated, the phase rotation which is proportional to index will be obtained and this is corresponding to the FFT window shift WSH in the figure. By using CPH and WSH, the constellation can be compensated as shown in figure 18.33.

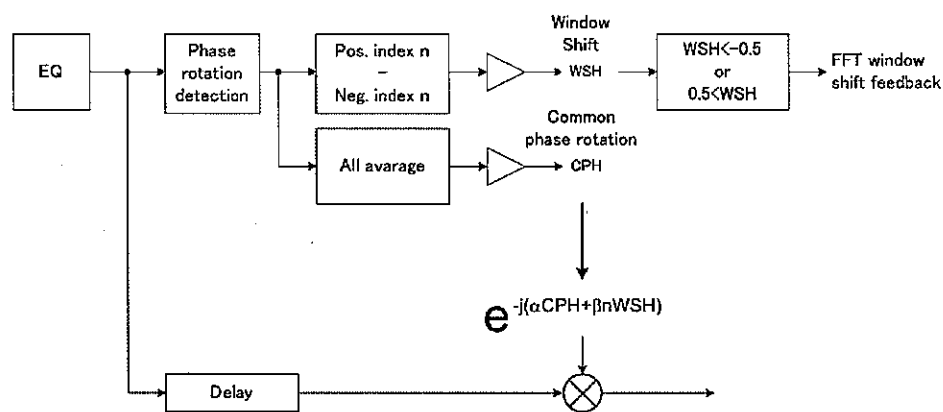


Figure 18.33 Equalizer output rotation detection and compensation

We have discussed various OFDM based transceivers and their design, various diversity techniques, and related error detection and compensation techniques. In the final section, the hardware implementation and related chip performance will be addressed.

18.5 Hardware Implementation and System Integration

18.5.1 Four antenna adaptive array combiner LSI [7]

In this section, some implementation results by ASIC, FPGA and DSP devices will be shown. The first implementation result is ASIC implementation of 4 antennas adaptive array combiner with MRC algorithm for digital TV receivers. This LSI can be used for 6MHz bandwidth receivers such as NTSC, ATSC and ISDB-T. One feature of this LSI is that three independent 4 antennas MRC combiners are included because the 6MHz bandwidth is wide. The 6MHz bandwidth is divided into three 2MHz bandwidth and MRC combining process is performed. Then 3-band division and combination circuit are also integrated as shown in the

figure 18.34.

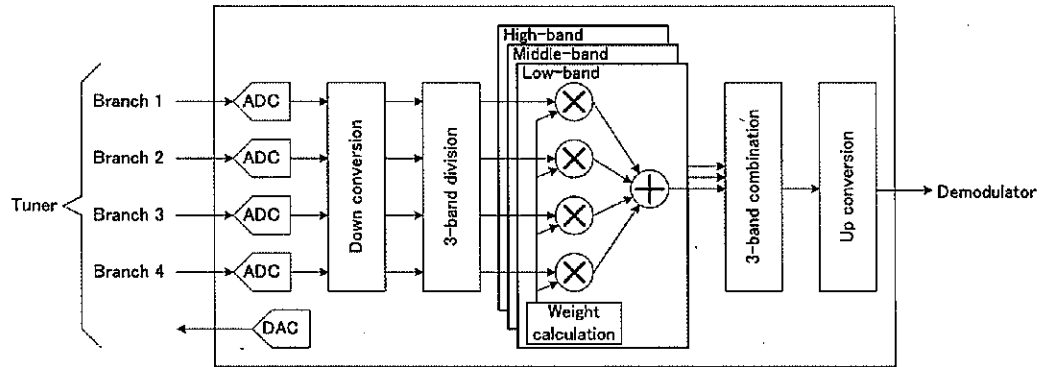


Figure 18.34 Block diagram of 4 antennas adaptive combiner LSI

Table 3.1 summarizes the features of adaptive combiner LSI. Total logic gate size is 700K gates and 4 analog to digital converters are integrated in 25mm² LSI chip with 0.18um CMOS technology. The LSI needs 1.8V voltage supply for internal circuits and 3.3V voltage for I/O buffers.

Table 18.4 Features and Data of LSI

Die size	25mm ²
Process technology	0.18um CMOS 6-metal
Supply voltage	1.8V for ADC & Logic 3.3V for IO buffer
Number of gates	700K gates
ADC	Four ADCs 10bits 33Msps
Power consumption	600mW

In this implementation, four ADC needs approximately 25% of die area is used in 0.18um CMOS technology. Chip photomicrograph is shown in figure 18.35.

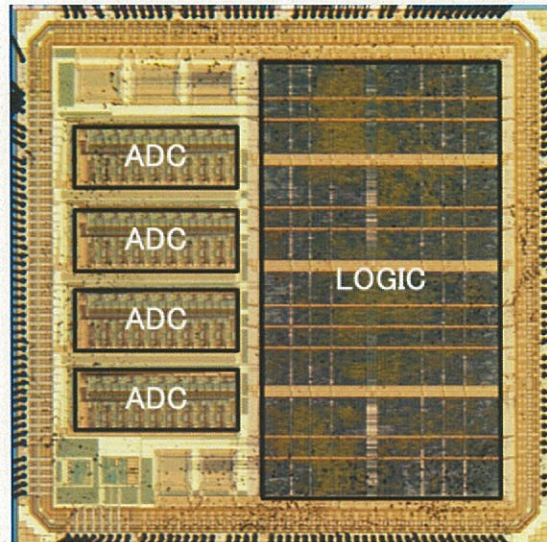


Figure 18.35 Chip photomicrograph of adaptive combiner LSI

18.5.2 A 2/4/8 Antennas Configurable Diversity OFDM Receiver LSI [16]

The second implementation result is a 2/4/8 Antennas Configurable Diversity OFDM Receiver LSI for Japan digital TV specification of ISDB-T. The LSI is composed of time domain adaptive array and frequency domain carrier diversity as shown in figure 18.22. Using 1 or 2 LSIs, 2/4/8 antennas diversity configurations are possible as shown in figure 18.23. Not only OFDM demodulation, the LSI integrates Viterbi and Reed-Solomon Error correction circuits.

Table 18.5 Features and Data of Diversity OFDM Receiver LSI

Process Technology	90nm 7M1P CMOS
Logic	1.8M gates
Memory and ADC	18.4M bit / 4 ADC
Supply Voltage	1.2V core, 3.3V I/O
Active Power	310mW typical
Die Size	7.0mm x 7.0mm
Package	12mm x 12mm 144FBGA

The feature of diversity receiver LSI is shown in table 18.5. The LSI integrates 4 ADCs, 18M gates and 18.4Mbit SRAM memory in 49mm² with 90nm CMOS technology. The largest

memory is time de-interleave RAMs because of ISDB-T requirement. The core logic supply voltage is 1.2V and I/O circuit needs 3.3V supply voltage. Chip photomicrograph is shown in figure 18.36.

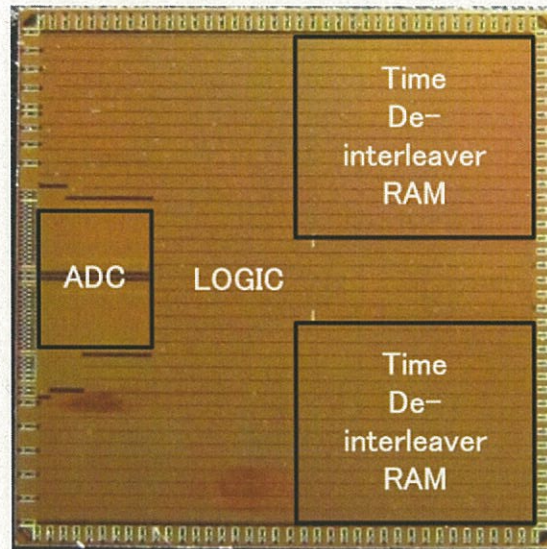


Figure 18.36 Chip photomicrograph of Diversity OFDM receiver LSI

18.5.3 Joint Hardware-Software Implementation of Adaptive array antenna OFDM receiver by FPGA [9]

Figure 18.37 illustrates a block diagram of the array antenna concatenating with OFDM demodulator. Digital beam-forming (DBF) algorithm to generate the coefficients of the array antenna is executed prior to FFT. The output of the array antenna is fed to OFDM demodulator to retrieve the transmitted data. Joint hardware-software (HW-SW) solution is chosen to provide an easily debugged and flexible platform for adaptive array antenna processing. The supported adaptive array antenna algorithms are OFDM selective Maximum Ration Combining (MRC) in section 18.3.2.2 and Sample Matrix Inversion (SMI) in section 18.3.2.3.

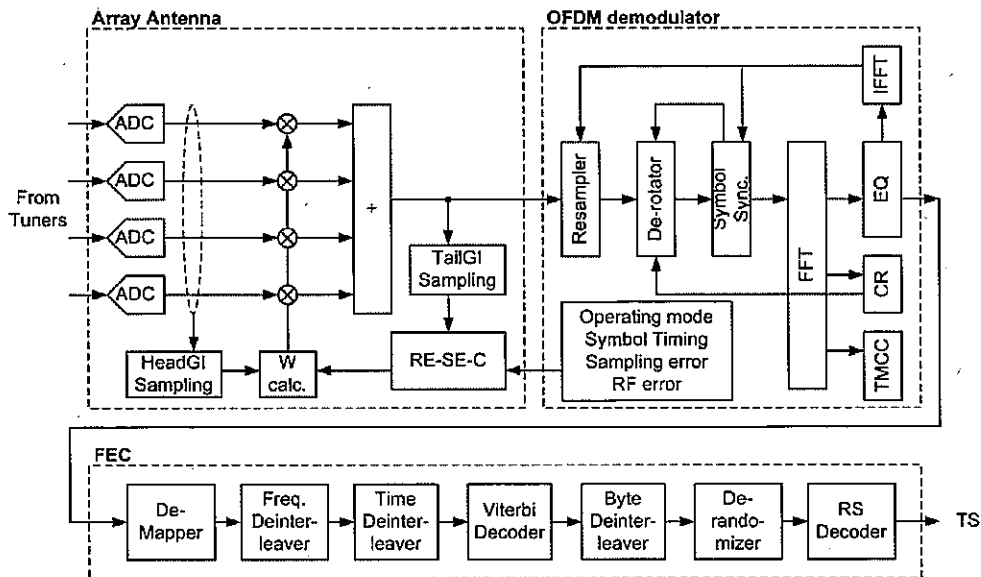


Figure 18.37 Block diagram of adaptive array antenna OFDM receiver (where “RE-SE-C” stands for “RF error and Sampling error compensator”)

Since RF error and sampling rate error are inherently unavoidable, the compensation feedback signals of RF error and sampling error are provided from OFDM demodulator to array antenna signal processing. Figure 18.38(a) shows an impact of sampling rate error on a sampled TailGI. Due to the drifting of sampling portion of TailGI, the samples of TailGI do not remain precisely. Therefore, it obviously deteriorates the cross correlation between HeadGI of the inputs and TailGI of the output. The block diagram of the compensator of sampling rate error is shown in figure 18.38(b). The error is estimated in OFDM demodulator and accumulated on each OFDM symbol to measure the exact duration of effective symbol length T_e . TailGI is then fed to a 13-tap poly-phase FIR filter to eliminate the impact of sampling rate error. The coefficients of 13-tap FIR filter are determined by the value of the T_e using LUT method (Look-Up Table). Not only RF error and sampling rate error, but also the operation mode of ISDB-T and the synchronization of OFDM symbol are provided from OFDM demodulation and they are fed back to the adaptive array antenna.

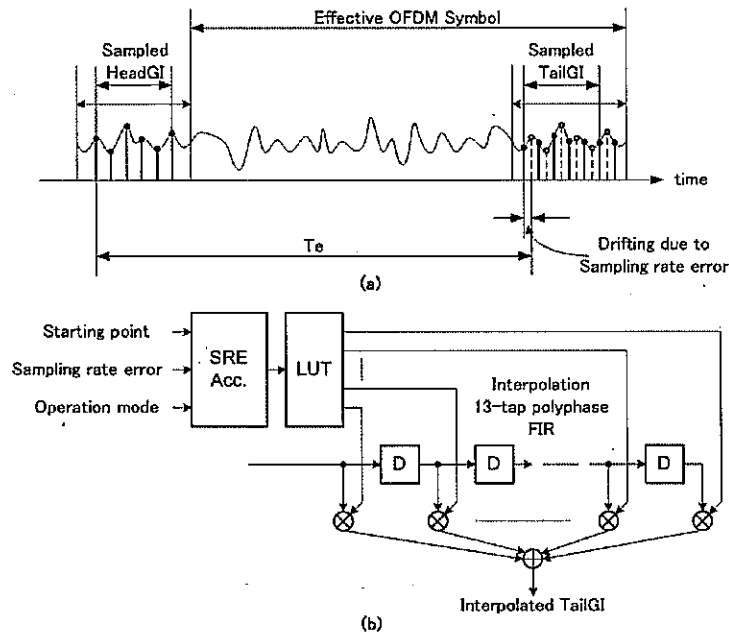


Figure 18.38 (a) Impact of sampling error rate and (b) block diagram of sampling rate error compensator (where "SRE Acc." Stands for "Sampling Rate error Accumulator")

Figure 18.39 illustrates the functional block diagram of the array antenna, which comprises of HW and SW components. The HW side consists of prefixed functions of the array antenna including four complex multiplications and a complex addition of four-element array antenna. It also takes responsibility for accessing interface memory, timing in the SW side, I/O and so on. Meanwhile, the SW side delivers the flexible functions corresponding to DBF algorithms in C programming language. In detail, signals coming from 4-element array antenna are firstly analog-to-digital converted. Using the information of the starting point of OFDM symbol provided by OFDM demodulation, samples of HeadGI of the inputs and TailGI of the outputs are extracted and are stored into the interface memory. The compensation of RF error and sampling rate error are conducted in the HW side. The HW side also generates timing control signals to the SW side to trigger CPU. Then, the coefficients generated by the SW side are used to combine 4 branches of the array antenna. In the SW side, two banks of block memory storing samples of two successive OFDM symbols directly link to CPU. Using these samples, CPU performs DBF algorithm to generate the coefficients. They are stored back into the corresponding bank of memory and are delivered to the HW side.

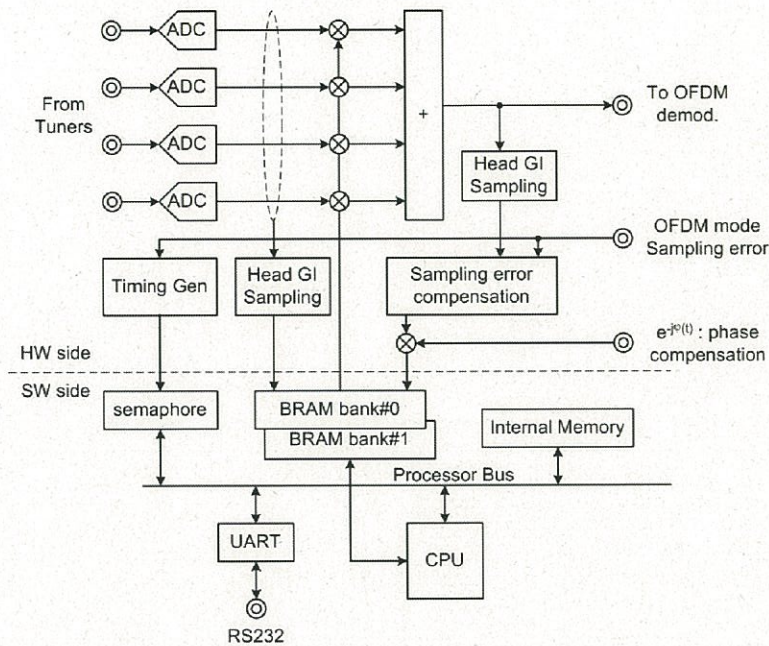


Figure 18.39 Functional block diagram of joint HW-SW solution for array antenna

Figure 18.40 and Table 18.6 show the photograph of the prototype and hardware features, respectively. 4 canned-type tuner devices are used and down-convert the RF signal to IF signals. The four ADCs convert the analog IF signals to digital signals. The lower left boards carries two FPGAs and a Power PC microprocessors embedded in one of the FPGAs is used for the array antenna system. ISDB-T OFDM demodulation is performed in the right board which carries three FPGAs. Two VirtexII V4000 FPGAs are used to implement OFDM demodulation and one VirtexII V3000 FPGA is used for FEC processing.

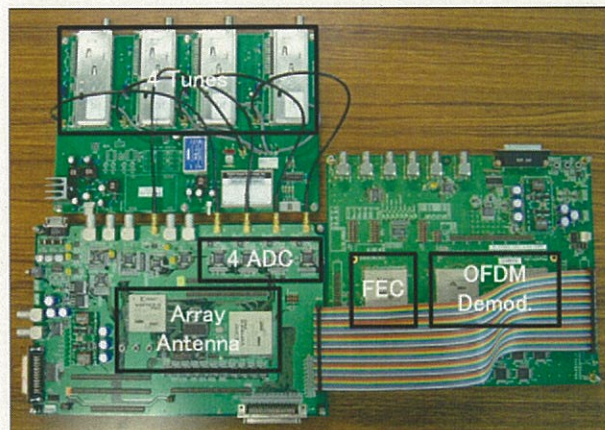


Figure 18.40 Prototype of adaptive array antenna OFDM receiver

Table 18.6 Features and Data of Prototype Implementation

Array Antenna	ADC	4 channel, 10bit 32MHz sampling rate
	FPGA	Xilinx Virtex-II Pro VP70 x1 Xilinx Virtex-II Pro VP20 x1
	CPU	PowerPC405 260MHz in FPGA
OFDM	Demod.	Xilinx Virtex-II V4000 x2
	FEC	Xilinx Virtex-II V3000 x1

3.2.4 Hybrid Pre-FFT Adaptive Array Antenna and Post-FFT Space Diversity Combining for Mobile ISDB-T Receiver by FPGA [19]

Figure 18.41 shows a block diagram of hybrid pre-FFT adaptive array antenna (AAA) and post-FFT space diversity combiner (SDC) system. In this implementation, prefixed processes such as FFTs, CTF estimation, pre-FFT and post-FFT estimations are realized by gates in FPGA. In contrast, DBF algorithms are realized in Digital Signal Processor (DSP) such as TI C6713 floating point DSP running at 225MHz. The supported adaptive array antenna algorithms are OFDM selective MRC and SMI. SMI needs the 4X4 matrix inversion and then the C6713 floating point calculation can support complicated DBF algorithms well. Figure 18.42 shows supported configurations. The system configuration A needs two independent 2-element antenna DBF calculation by DSP and the system B needs 4-element antenna DBF calculation by DSP. In order to change these configurations, control signals AAA_opt and SDC_opt have to be controlled while DBF software in DSP has to be controlled.

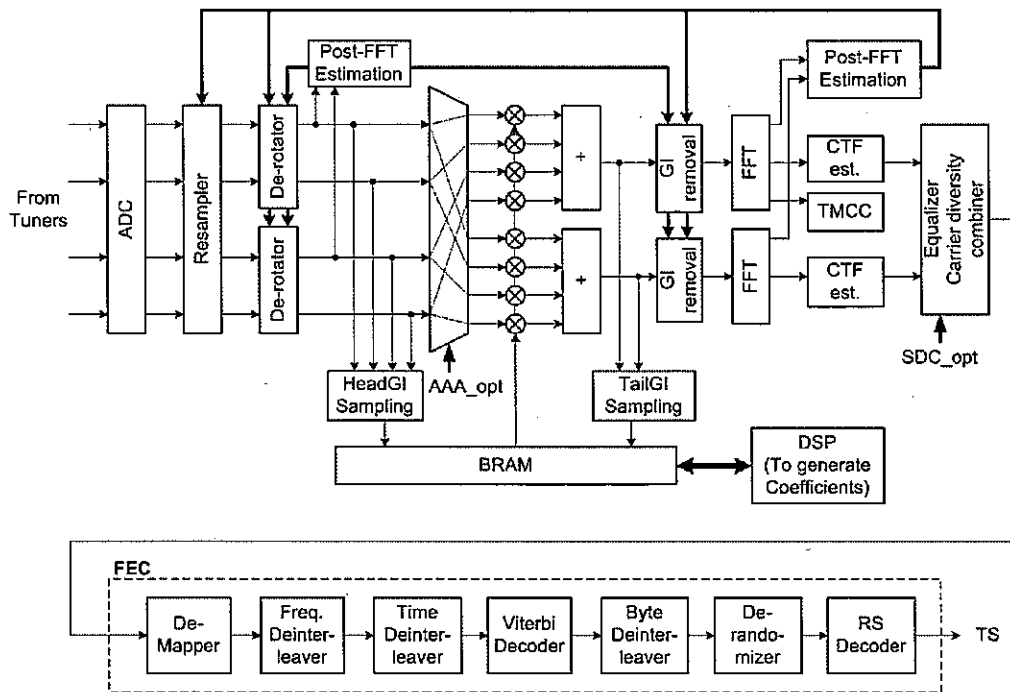


Figure 18.41 Block diagram of AAA-SDC OFDM receiver

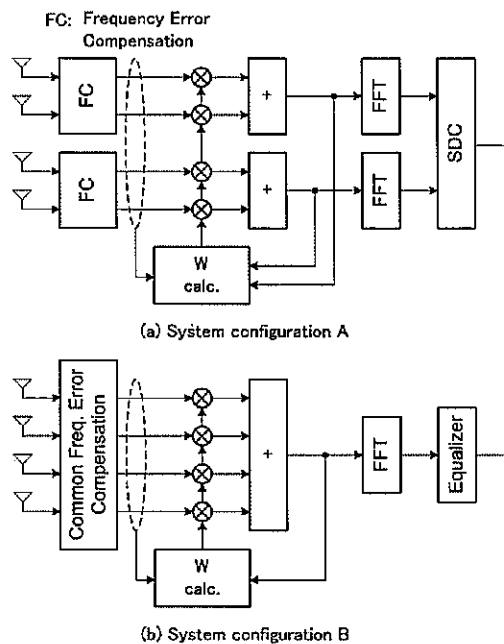


Figure 18.42 System configurations A and B

Figure 18.43 shows the photograph of the prototype, which is composed of two large FPGA boards as same as figure 18.40 and small DSP daughter board on the top. The DSP daughter board carries TI C6713 floating point DSP running at 225MHz and DRAMs for

main memory for DSP. The external 32bit memory bus is connected to the other FPGA board by flat cable as shown in the photo. Table 18.7 summarizes the system features.



Figure 18.43 Prototype of AAA-SDC OFDM receiver

Table 18.7 Features of AAA-SDC system

Array Antenna	ADC	4 channel, 10bit 32MHz sampling rate
	FPGA	Xilinx Virtex-II Pro VP70 x1 Xilinx Virtex-II Pro VP20 x1
	DSP	TI DSP C6713 TMS320C6712 225MHz
OFDM Carrier Diversity	Demod.	Xilinx Virtex-II V4000 x2
	FEC	Xilinx Virtex-II V3000 x1

18.6 Conclusions

In this chapter, we have discussed various OFDM based transceiver architecture and their designs such as DTV receiver, OFDM WLAN and OFDMA transceivers. Since the OFDM based applications are still expanding, much more OFDM based communication systems will be developed in future. However, the two key techniques such as Error Detection and Compensation, and Diversity covered in this chapter are fundamental techniques to achieve OFDM based successful communication system designs. Then those fundamental signal

processing should also be further researched in future. Since semiconductor technology progress pushes system integration level higher and higher, further performance increase will be achieved by heavy signal processing enhancements for both Single Input Single Output (SISO) and Multiple Input Multiple Output (MIMO) applications. Then let's enjoy the digital signal processing researches for future communication systems.

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