Compensation Techniques for Signal Distortion in Multipath Radio Channel

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Fangwei Tong
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Chapter I

Introduction

1.1 Backgrounds - A Brief Review

The confirmation of the existence of electromagnetic waves by H. R. Hertz in 1888 opened the prologue to radio communications. Now, radio communications have become a part of human's life. Television, radio, mobile telephone, and so on are perhaps the most often used words. The rapid development of radio communications in recent years lies on the fact that wireless communications made it possible to call or receive information anywhere. Today, communicating anywhere, anytime, and with anyone has been considered as the next development target. However, it is not always plain sailing to realize such ambitious target. There are many obstacles that need to be surmounted.

One problem we encounter is the deterioration of the radio wave propagation. In particular, when the signal is transmitted through a "out-of-sight channel", such as land mobile radio channel, multipath fading including flat fading and frequency-selective fading will occur due to the superposition of the reflected, refracted and scattered signals at receiver. These two kinds of fadings affect both the audio/video broadcasting and mobile radio communications, such as cellular telephone, portable telephone, and PHS (personal handy phone system, or the second generation cordless phone). Flat fading causes random fluctuation in signal envelope and phase. Such fluctuation deteriorates the error rate performance of a digital system and decreases the signal to noise power ratio of an analog system. Furthermore, if multipath signals arrive with the
different delays that cannot be simply ignored, or if signal bandwidth is relatively wide, which usually means the data rate is high for a digital transmission system, waveform distortion will occur due to frequency-selective fading. At this time, system performance will be worsen further. On the other hand, for digital transmission systems, the demand for uninterruptedly increased data transmission rate is an inexorable trend. For example, the channel speed of Japanese digital cellular system is 42 kbps [1], [2]; American digital cellular (IS-54) is 48.6 kbps [1], [3]; GSM (group special mobile, or global system for mobile) is 270.833 kbps [1], [4]. In addition, FPLMTS (future public land mobile telecommunication systems) [1], [4], [5] based on the recommendations by ITU (International Telecommunication Union) are being developed to provide all services generally available through the fixed wireline networks. It is conceivable that higher data rate is necessary in order to provide all these services. A data rate as high as 1920 kbps was ever taken into consideration by CCIR (international radio consultative committee). Now, ITU-Radio) in 1989. Under such a situation, compensating the signal for frequency-selective fading becomes an important issue. For analog transmission system, especially for broadcasting systems, high signal quality is essential so as to satisfy the requirements of people pursuing the melodious audiovisual entertainment. Any noise and/or distortion in signal will greatly disappoint listeners (Perhaps, everybody has this experience when he or she uses a car radio). Therefore, for both digital mobile communication system and analog broadcasting system, it is necessary to compensate the signal for multipath fading.

Many compensation techniques have been developed for compensating signal in multipath propagation environments. Some of them are only effective for digital transmission systems, for example, decision feedback equalizer [6], the Viterbi equalizer [7]-[9], and so on. All these compensation techniques for digital system make use of the features of a digital signal, i.e., digital signal has discrete value in amplitude and/or phase at sampling instant. Moreover, these techniques compensate signal at receiving terminal. Finding a technique that compensates the signal at transmitting terminal (base station) is also seriously considered, since all receivers (mobile stations) can benefit from this compensation technique without employing any additional
circuits. For analog systems, especially for an FM broadcasting system, the available compensation techniques are inferior to that for digital systems in both quality and quantity. This is perhaps due to that no obvious features, such as discrete amplitude and/or phase, can be used for an analog system. Main compensation methods are diversity [10], equalizer [11], and adaptive array antenna (or echo canceller) [12]. All these techniques are also valid for digital systems. Diversity reception has been widely applied, but it is not of sufficient compensation capability for frequency-selective fading. The equalizer proposed by [13], [14] for an FM broadcasting receiver did not show good performance for echoes with small excess delay. An adaptive array antenna has not been applied in an FM broadcasting system yet as far as we know. Even if applied for a analog transmission system, whether an adaptive array can work well or not in a dynamic channel is also a problem. Therefore, there are a lot of works need to be done to compensate signals for multipath fading.

1.2 The Aim of This Study

In mobile communication systems, mobile station (handy terminal) with small size is one of the key requirements. If we can compensate the signal for frequency-selective fading at base station, the burden on the compensation circuits in mobile station can be reduced, or with the same compensation capability at mobile stations, the total system performance can be increased. A technique called beam tilting was proposed for decreasing the interference between two adjacent frequency-reuse cell. According to its working mechanism, we think that it should be effective to reduce the echo power and thereby compensate the signal for frequency-selective fading. The analysis of the effect of beam tilting on system performance is one of the aims of this study. Since signal may be reflected on different obstacles with different types of surface roughness, different echo propagation models need to be taken into consideration. In this thesis, two models, namely, perfect reflection and diffusely scattered reflection are considered.

A feedback type equalizer for use in an FM broadcasting receiver to compensate for
frequency-selective fading was proposed by [14]. The compensation performance achieved for echoes with small excess delay, less than 5 μs for example, was not so ideal. Considering that the excess time delay is usually 2 - 4 μs, sometimes it is 4 - 8 μs, improvement of the performance of the equalizer is necessary. Moreover, the performance of this equalizer was examined only at a few excess delay values. Thus, further examination of the performance of such a feedback type equalizer is perhaps necessary. Improving the performance of the feedback type equalizer is another aim of this study. On the basis of the theoretical analysis, we find out the reason affecting the performance of the feedback type equalizer. In order to improve the performance, an adaptive feedback equalizer (echo distortion canceller) is proposed.

Adaptive array antenna is effective to cancel interference (echo), but in a dynamic channel, its compensation performance may be degraded. To remedy this defect, we propose an adaptive switching echo cancellation/diversity reception in this thesis. The proposed adaptive switching reception monitors the channel conditions and selects either an adaptive array antenna or a diversity receiver accordingly to compensate the signal for instantaneous channel conditions. The aim of this work is to realize the advantage of both echo cancellation (adaptive array antenna) and diversity reception. For the examination of the performance of the proposed adaptive switching reception, two propagation models are evolved. One is that antenna branches are not correlated. The other one is that antenna branches are fully correlated for each of the direct wave and reflected wave (echo).

Summarizing the paragraphs above, this study is aimed to compensate the signal for frequency-selective fading for digital mobile communication system and FM broadcasting system.

1.3 Structure and Organization of This Thesis

This thesis is composed of six chapters.

Chapter 1, that is this chapter, describes the backgrounds and aim of this study. The organization of this thesis is also introduced in this chapter.
As a reference of the contents of the later chapters, Chapter 2 introduces the mechanism of the occurrence of flat and frequency-selective fadings and their properties related to this study. Moreover, some current compensation techniques common to digital and analog transmission systems for multipath fading are also briefly described in this chapter.

Chapter 3 analyzes the effects of beam tilting on system bit rate selection in a digital mobile communication system. For wave reflection due to obstacles with different types of surface roughness, two echo propagation models, namely, perfect reflection and diffusely scattered reflection, are taken into consideration. System performance measure methods are also evolved and explained.

Chapter 4 studies on the feedback type echo distortion canceller applied to an FM broadcasting receiver. The feature of the compensation performance of such a echo distortion canceller is analyzed. The effect of the tap interval of the transversal filter, which is used in the echo distortion canceller, on the performance is investigated. In order to improve the performance of the feedback type echo distortion canceller, an adaptive feedback type echo distortion canceller is proposed. The proposed echo distortion canceller controls taps of the transversal filter with a dynamic adaptive algorithm. The performance of the proposed adaptive feedback echo distortion canceller is investigated.

In Chapter 5, we propose an adaptive switching echo cancellation/diversity reception for an FM broadcasting receiver in multipath mobile channel. The adaptive switching reception monitors channel conditions and selects either an echo canceller or a diversity receiver accordingly to compensate the signal for instantaneous channel conditions. Two schemes of the proposed switching reception are considered. One is termed pilot switching, which monitors the channel conditions by observing the pilot signal included in FM broadcasting signal. The other one is termed BF switching, which observes the beat level to short-term average envelope fading ratio (BF ratio). For investigation of the performance of the proposed adaptive switching reception, two propagation models, namely, uncorrelated antennal branches and correlated branches (for each of direct wave and reflected wave), are presented. The performance of the proposed adaptive
switching reception is examined for flat fading, frequency-selective fading, and as a function of the excess delay of the echo, respectively.

Chapter 6 summarizes the study in this thesis. The possibility of further study is also indicated in this chapter.
References


Chapter 2

Land Mobile Multipath Channel and Compensation Techniques

2.1 Introduction

One of characteristics of the land mobile communication channel is multipath fading. It usually results from the out-of-sight transmission characteristics of the mobile communication channel and the motion of receiver. Because of multipath fading, we must face many problems that do not exist in static (motionless) communication systems, such as violently fluctuation in the received field intensity, distortion in signal waveform caused by frequency-selective fading, random FM effect. Therefore, mobile communication equipments are of their own features so as to adapt themselves to the multipath propagation environments, for example, one or the combination of diversity reception, adaptive interference/echo cancellation, various adaptive equalizers, multicarrier transmission, anti-multipath modulations, error control techniques, and so on are usually applied in a mobile communication transmitter and/or receiver. At the same time, counter measures against multipath fading have been also attracting considerable interests of researches including us and industries.

In this chapter, we briefly introduce the mechanism of the occurrence of multipath fading and some current compensation techniques. The purpose of this chapter is to offer a simple
reference to the later chapters rather than to describe the current situation of compensation techniques. For the detailed description of fading and compensation techniques, please refer to [1]-[5].

2.2 Flat Fading

Land mobile radio channel is characterized by out-of-sight communication to/from a moving terminal. Many random signals that propagated through different signal paths from the transmitter are superposed at the receiver as shown in Fig. 2.1 and thereby standing waves are produced.

Transmitter antenna

Fig. 2.1 Mobile radio communication environment. Reflected and scattered waves are superposed at the receiver.
While a receiver and/or a transmitter moves in the standing wave field, the received signal suffers random fluctuation in signal level and phase, and a Doppler shift in received signal frequency. If the difference between the arrival times of the incident waves can be neglected compared to the reciprocal of the signal bandwidth, then, the multipath channel will not cause signal waveform distortion or intersymbol interference (for digital signal), and the random fluctuation phenomenon is then called flat fading.

### 2.2.1 Distributions of Received Signal Amplitude and Phase

To analyze flat fading, we model the multipath reception as shown in Fig. 2.2. Consider the simplest case for simplicity of argument, where a unmodulated carrier signal is transmitted. The i-th incident wave \( e_i(t) \) can be expressed as

\[
e_i(t) = Re[z_i(t)e^{j2\pi f t}]
\]  

(2.1)
where $\text{Re} \{ \cdot \}$ means real part; $z_i(t)$ is the complex envelope of the $i$-th incident waves; $\omega_c$ is the carrier frequency. Assume that the wavelength is $\lambda$; the vehicular velocity is $v$. The complex envelope is then expressed as

$$z_i(t) = a_i e^{i \left( 2\pi \frac{v \cos \phi_i}{\lambda} + \theta_i \right)}$$

$$= x_i(t) + j y_i(t)$$

where $a_i$ and $\theta_i$ are the amplitude and phase of the $i$-th incident wave, respectively, both $a_i$ and $\theta_i$ are random variables; $x_i(t)$ and $y_i(t)$ are the inphase and quadrature components of $z_i(t)$, respectively. $v/\lambda \cos \phi_i$ represents Doppler shift. When $\phi_i = 0$ (or $180^\circ$), Doppler shift has its maximum value $v/\lambda$ (or $-v/\lambda$), and $v/\lambda$ is usually called maximum Doppler frequency. Note that the difference between the arrival times of the incident waves is here neglected, since we assume this difference is very small.

Assuming the number of incident waves is $N$, the received signal $e(t)$ is then expressed as

$$e(t) = \sum_{i=1}^{N} e_i(t)$$

$$= \text{Re} \left[ \sum_{i=1}^{N} z_i(t) e^{i 2\pi f_c t} \right]$$

$$= \text{Re} \left[ z(t) e^{i 2\pi f_c t} \right]$$

(2.3)

where

$$z(t) = \sum_{i=1}^{N} z_i(t)$$

$$= \sum_{i=1}^{N} x_i(t) + j \sum_{i=1}^{N} y_i(t)$$

(2.4)
We then express $e(t)$ as

$$e(t) = x(t)\cos(2\pi f_c t) - y(t)\sin(2\pi f_c t) \quad (2.5)$$

Assume that all $x_i(t)$ and $y_i(t)$ obey the same distribution. As a consequence of the central limit theorem, $x(t)$ and $y(t)$ are independent Gaussian random processes with expected value (mean) being equal to zero and same variance [6], [7]. Then, the joint probability density function $p(x, y)$ of $x=x(t)$ and $y=y(t)$ can be expressed as

$$p(x, y) = \frac{1}{2\pi b} e^{-\frac{x^2 + y^2}{2b}} \quad (2.6)$$

where $b$ is the average received signal power. To analyze the distribution of the envelope and phase of the received signal, we express the received signal $e(t)$ as

$$e(t) = r(t)\cos(2\pi f_c t + \Theta(t)) \quad (2.7)$$

where

$$\begin{align*}
\left\{ 
\begin{array}{ll}
  r^2(t) &= x^2(t) + y^2(t) & r(t) \geq 0 \\
  \Theta(t) &= \tan^{-1}\left[\frac{y(t)}{x(t)}\right] & -\pi \leq \Theta(t) < \pi
\end{array}
\right.
\end{align*} \quad (2.8)$$

or

$$\begin{align*}
\left\{ 
\begin{array}{l}
  x(t) = r(t)\cos\Theta(t) \\
  y(t) = r(t)\sin\Theta(t)
\end{array}
\right.
\end{align*} \quad (2.9)$$

Using Eqs. (2.6) and (2.9), we can yield the joint probability density function $p(r, \Theta)$ of $r=r(t)$ and $\Theta=\Theta(t)$ as

$$p(r, \Theta) = \frac{r}{2\pi b} e^{-\frac{r^2}{2b}} \quad (2.10)$$

Then, the marginal probability density functions $p(R)$ and $p(\Theta)$ can be obtained as
respectively. Equation (2.11) indicates Rayleigh distribution as shown in Fig. 2.3(a) and Eq. (2.12) represents the uniform distribution as shown in Fig. 2.3(b). Samples of the received signal amplitude and phase are shown in Fig. 2.4(a) and (b), respectively. Observing Fig. 2.4(a), we can find out that the minimum distance between the signal level drops is about one half of the wavelength. Thus, for a carrier frequency of 1 GHz, this minimum distance becomes as short as about 15 cm and the minimum distance between the local maximum and minimum is about 7.5 cm.

Flat fading deteriorates the CNR (Carrier to Noise power Ratio) of the received signal due to signal level dropping and causes random FM effects because of the random fluctuation in the signal phase. Diversity reception, especially space diversity, is most often used to compensate for flat fading. The correlation of the received signal envelope, especially the spatial correlation, strongly affects the performance of diversity systems. In the next section, we cite some results related to the correlation of the signal envelope.

2.2.2 Spatial Correlation of the Envelope

The autocorrelation $R(\tau)$ of the received signal envelope is expressed as
(a) pdf of the amplitude (Rayleigh distribution).

(b) pdf of the phase (Uniform distribution).

Fig. 2.3 Probability density functions (pdf) of fading signal amplitude and phase.
Fig. 2.4 Random fluctuation in received signal amplitude and phase. $\Delta d$ is the increment in distance; $\lambda$ is the wavelength.
\[ R_s(\tau) = E \left[ r(t)r(t+\tau) \right] \]
\[ = \int_{0}^{\infty} \int_{0}^{\infty} r_1 r_2 p(r_1, r_2) dr_1 dr_2 \tag{2.13} \]

where \( E[ \cdot ] \) means assemble average; \( r_1 = r(t), \ r_2 = r(t+\tau) \); \( P(r_1, r_2) \) is the joint probability density function of \( r_1 \) and \( r_2 \), we denote

\[
\begin{align*}
x_1 &= x(t) = r(t) \cos \theta(t) \\
y_1 &= y(t) = r(t) \sin \theta(t)
\end{align*}
\]
\[
\begin{align*}
x_2 &= x(t+\tau) = r(t+\tau) \cos \theta(t+\tau) \\
y_2 &= y(t+\tau) = r(t+\tau) \sin \theta(t+\tau)
\end{align*}
\tag{2.14}
\]

Using the general expression for the joint density of the four Gaussian variables \( x_1, x_2, y_1, \) and \( y_2 \) [7] and applying the transformation of variables of Eq. (2.14), we can obtain the joint density of the envelopes and phases \( p(r_1, r_2, \theta_1, \theta_2) \) as [8]

\[
p(r_1, r_2, \theta_1, \theta_2) = \frac{r_1 r_2}{(2\pi \mu)^2 (1-\lambda^2)} e^{-\frac{r_1^2 + r_2^2 - 2r_1 r_2 \cos(\theta_1 - \theta_2)}{2\mu (1-\lambda^2)}} \tag{2.15}
\]

where \( \theta_1 = \theta(t) \), and \( \theta_2 = \theta(t+\tau) \); \( \tan \phi = \mu / \mu_2 \), \( \lambda^2 = (\mu_1^2 + \mu_2^2) / \mu^2 \), and \( \mu = E(x_1^2) \), \( \mu_1 = E(x_1 x_2) \), \( \mu_2 = E(x_1 y_2) \). The joint probability density function \( P(r_1, r_2) \), can be then obtained as

\[
p(r_1, r_2) = \int_{-\pi}^{\pi} \int_{-\pi}^{\pi} p(r_1, r_2, \theta_1, \theta_2) d\theta_1 d\theta_2
\]
\[
= \frac{r_1 r_2}{\mu^2 (1-\lambda^2)} e^{-\frac{r_1^2 + r_2^2}{2\mu (1-\lambda^2)}} I_0 \left( \frac{r_1^2 + r_2^2}{\mu (1-\lambda^2)} \right) \tag{2.16}
\]

where \( I_0(\cdot) \) is the modified Bessel function of zero order. Substituting Eq. (2.16) into Eq. (2.13) and performing trivial mathematical processing under the assumption that the incident power is uniformly distributed with respect to the incident angular, it can be shown [8]

\[
R_s(\tau) = \frac{\pi}{2} b \left[ J_0(\omega_m \tau) \right] \left[ 1 + \frac{J_0(\omega_m \tau)}{4} \right] \tag{2.17}
\]
where \( J_0( \cdot ) \) is the zeroth-order Bessel function of the first kind; \( \omega_m \) is maximum Doppler frequency. The autocovariance function of the signal envelope \( L(\tau) \) is defined for a stationary stochastic process as
\[
L(\tau) = R_x(\tau) - E^2(r) \tag{2.18}
\]
Using Eq. (2.11), we have
\[
E(r) = \int_0^\infty r p(r) dr
= \frac{1}{b} \int_0^\infty r^2 e^{-\frac{r^2}{2b^2}} dr
= \sqrt{\frac{\pi}{2b}} \tag{2.19}
\]
Substituting Eqs. (2.17) and (2.19) into Eq. (2.18), we have
\[
L(\tau) = \frac{\pi}{8} b J_0^2(\omega_m \tau) \tag{2.20}
\]
Equation (2.20) shows the envelope correlation as a function of time separation. Considering that a vehicular moves at a speed \( v \), the correlation function in terms of the time difference \( \tau \) is equivalent to that in terms of the distance \( d = v \tau \). Substituting \( \tau = d/v \) and \( \omega_m = 2\pi v/\lambda \) into Eq. (2.20), we obtain the spatial correlation of the signal envelope as
\[
L(d) = \frac{\pi}{8} b J_0^2 \left( \frac{2\pi d}{\lambda} \right) \tag{2.21}
\]
The normalized spatial correlation is
\[
L_n(d) = \frac{L(d)}{L(0)} = J_0^2 \left( \frac{2\pi d}{\lambda} \right) \tag{2.22}
\]
Equation (2.22) shows that the correlation of the signal envelope as a function of distance
separation decreases with the increase in $d$ and becomes zero for the distance separation of $0.38\lambda$, $0.88\lambda$, ... . In practical application, the correlation can be thought as zero for the distance separation of $0.5\lambda$ [9].

There are also other properties of the fading signal, such as power spectra, inphase and quadrature moments, random frequency modulation, the correlation at a base station, and so on. Since all these properties are not so closely related to the contents of later chapters, we here omit the description of them.

2.3 Frequency Selective Fading

The mechanism of the occurrence of frequency-selective fading is the same as that of flat fading, i.e., many reflected, refracted, and/or scattered signals with different time delay are superposed at the receiver. However, the difference in arrival times can no longer neglected now compared to the reciprocal of the signal bandwidth. This is the only difference between flat fading and frequency-selective fading, but they will result in different impacts on signal.

2.3.1 Channel Transfer Function

For simplicity of description, let us consider a two-path case, where one wave arrive at the receiver with time delay $\tau_1$, and the other one with time delay $\tau_2$. The channel transfer function $H(\omega)$ is then expressed as

$$H(\omega) = r_1 e^{j\omega \tau_1} + r_2 e^{j\omega \tau_2}$$

(2.23)

where $r_1$ and $r_2$ are the amplitude of the two waves, respectively. The amplitude of transfer function $|H(\omega)|$ is then

$$|H(\omega)| = \sqrt{|r_1|^2 + |r_2|^2 + 2Re[r_1^* r_2 e^{j\alpha(\tau_2 - \tau_1)}]}$$

(2.24)
where $*$ is complex conjugate. If $r_1 = r_2 = 1$, Eq. (2.24) becomes

$$|H(\omega)| = \sqrt{2 + 2\cos[\omega(\tau_2 - \tau_1)]}$$

Equation (2.25) is shown in Fig. 2.5. We can see that $|H(\omega)|$ shows frequency selective characteristics. If a modulated signal propagates through such a channel, signal waveform distortion will occur, since the signal is filtered by channel transfer function. It can be also seen that if $|\tau_1 - \tau_2|$ is very small compared to the reciprocal of the signal bandwidth, the signal will not be filtered and thereby no distortion occurs. This is in fact the case of flat fading.

![Fig. 2.5 Transfer function (amplitude) of a two-path frequency-selective fading channel.](image)
2.3.2 Impacts of Frequency-Selective Fading on Signal

A constant envelope signal can be expressed as

\[ s(t) = \cos(\omega_c t + \theta(t)) \]  

(2.26)

Assume this signal is transmitted through a two-path frequency-selective fading channel. Then, the received signal \( v(t) \) is

\[ v(t) = \cos(\omega_c t + \theta(t)) + r \cos(\omega_c (t-\tau) + \theta(t-\tau)) \]  

(2.27)

where \( r \) and \( \tau \) are the amplitude and the excess delay of the reflected wave (echo), respectively. By performing some mathematical processing, Eq. (2.27) can be expressed as [20]

\[ v(t) = \sqrt{1 + r^2} \cos[\theta(t)-\theta(t-\tau)+\omega_c \tau] \cos[\omega_c t + \theta(t)+\tan^{-1}\left(\frac{r \sin[\theta(t)-\theta(t-\tau)+\omega_c \tau]}{1 + r \cos[\theta(t)-\theta(t-\tau)+\omega_c \tau]}\right)] \]  

(2.28)

It can be seen that the signal has no longer constant envelope. There is fast fluctuation in signal envelope. This fast fluctuation is usually called beat. We also can see that there is a \( \tan^{-1} \) term in the phase of the second cosine function. This \( \tan^{-1} \) term causes waveform distortion/intersymbol interference in the signal. Samples of received signal envelope and distortion signal waveform are show in Figs. 2.6 and 2.7.

2.4 Compensation Techniques for Flat Fading

For a digital transmission system, flat fading causes thermal noise error due to signal level drop and irreducible error due to random FM modulation effects. Both of them result in the deterioration of BER (Bit Error Rate) performance of the system. From the viewpoint of decreasing the system BER, any error control coding techniques [10] - [12] can be thought as the compensation techniques for flat fading. Trellis-coded modulation [13], [14] is also such a
Fig. 2.6 A sample of received signal envelope. The original signal has constant envelope.

Fig. 2.7 A sample of distorted signal waveform. The original signal is a sine signal.
compensation method, since coding and modulation are integrated in trellis-coded modulation so as to achieve better error rate performance. Moreover, counter measures against random FM to improve system BER performance are effective compensation methods for flat fading too, for example, adaptive carrier tracking demodulation [15], QPSK coherent detection with linear mean square estimation [16], pilot signal insertion method to compensate 16 QAM signal for Rayleigh fading [17], [18], and so on. For analog and digital transmission systems, the simple and effective compensation method for flat fading is diversity reception [1], [3]. From the point of the view of preventing signal level drop, diversity reception is most typical and most often applied compensation method for flat fading.

2.4.1 Diversity Systems

Diversity reception systems receive different signals travelling through different propagation paths and combine them with some criteria in order to improve the signal quality. Here, the "different propagation paths" not only means the difference in space, but also in time, angle, frequency, polarization, and so on. According to the difference in propagation path, diversity systems can be classified as space diversity with multiple spaced antennas, polarization diversity receiving differently polarized waves, frequency diversity with multiple frequencies, angle diversity using directive antennas aimed in different direction, time diversity by transmission of the same signal in different times, and multipath diversity (Rake receiver). Of these, the last two have been primarily considered only for digital data transmission. Except for polarization diversity, there is generally no limit to the number of diversity branches, the number of diversity branches can be simply increased by multiplication of equipment. Considering the signal combining methods/criteria, we have maximal ratio combining, equal gain combining, and selection combining. These are further classified into pre-demodulation and post-demodulation combining. Pre-demodulation requires signal synchronization (or cophasing) so that the modulated signals are in phase with each other. Without synchronization, the diversity effect is lost for maximal
ratio combining or equal gain combining and switching noise occurs for the selection diversity. This problem can be overcome by using post-demodulation combining.

Diversity effect or diversity gain is related to the correlation between diversity branches [1], [3], [19]. With the increase in diversity branch correlation, diversity effect decreases. Therefore we should properly set the diversity branches. For space diversity, antennas should be spaced 0.5λ apart, where antenna branches are not correlated with each other as described in section 2.2.2. In the following, we assume the antenna branches are independent of each other.

2.4.2 Probability Density Function of SNR for Diversity Systems

Assume the same mean signal power and mean noise power for all branches. Then mean SNR (Signal to Noise power Ratio) is also the same for all branches. Denote the SNR of the i-th branch as γ_i. The pdf (probability density function) of the SNR of the i-th branch, p(γ_i), is given as [1]

\[
p(γ_i) = \frac{1}{γ_0} e^{-\frac{γ_i}{γ_0}} \quad (0 \leq γ_i < \infty)
\]  

(2.29)

where γ_0 is the average SNR. The probability that γ_i takes values less than γ is

\[
P(γ_i \leq γ) = \int_0^γ p(γ_i) dγ_i = 1 - e^{-\frac{γ}{γ_0}}
\]

(2.30)

Selection Diversity

An M branch space diversity system with selection combining is shown in Fig. 2.8. At each instant, selection combining selects the branch that has the highest SNR, γ_max, for this instant. The SNR of the other branches are less than γ_max at this instant. Then, the probability that
Fig. 2.8 Space diversity selection combining system.

Fig. 2.9 Maximal ratio combining diversity system.
the SNR of the output of the selection diversity system is less than \( \gamma \); \( P_M(\gamma) \), can be yielded as

\[
P_M(\gamma) = P(\gamma_1, \gamma_2, \ldots, \gamma_M \leq \gamma) = \left(1 - e^{-\frac{\gamma}{\gamma_0}}\right)^M
\]  

(2.31)

The pdf \( p_M(\gamma) \) is then

\[
p_M(\gamma) = \frac{d}{d\gamma} P_M(\gamma)
\]

\[
= M \frac{\gamma}{\gamma_0} \left(1 - e^{-\frac{\gamma}{\gamma_0}}\right)^{M-1}
\]

\[
= M \frac{\gamma^{M-1}}{\gamma_0} \left(1 - e^{-\frac{\gamma}{\gamma_0}}\right) \quad (\gamma << \gamma_0)
\]

(2.32)

Maximal Ratio Combining

An \( M \) branch maximal ratio combining diversity system is shown in Fig. 2.9. Received signals at each branch are weighted by a factor and coherently summed. The pdf of SNR at the output of the combiner is given as [1],[3]

\[
p(\gamma) = \frac{1}{(M-1)!} \frac{\gamma^{M-1}}{\gamma_0^M} e^{-\frac{\gamma}{\gamma_0}}
\]

\[
= \frac{1}{(M-1)!} \frac{\gamma^{M-1}}{\gamma_0^M} \quad (\gamma << \gamma_0)
\]

(2.33)

The SNR at the output of the combiner can be calculated as follows. The received signal of the i-th branch, \( v_i \), can be expressed as (omit the time variable)

\[
v_i = r_i u + n_i
\]

(2.34)

where \( n_i \) is the noise of the i-th branch; \( r_i \) is the amplitude of the received signal, which represents the impact of flat fading on signal envelope; \( u \) is the normalized transmitted signal. The output
of the combiner $v$ is then

$$v = \sum_{i=1}^{M} a_i v_i = u \sum_{i=1}^{M} a_i r_i + \sum_{i=1}^{M} a_i n_i$$  \hspace{1cm} (2.35)$$

where $a_i$ are the weighting factors of the maximal ratio combing diversity. The noise power $N$ at the output of the combiner is

$$N = E\left[ \left( \sum_{i=1}^{M} a_i n_i \right)^2 \right] = \sum_{i=1}^{M} a_i^2 E(n_i^2) = \sum_{i=1}^{M} a_i^2 N_i$$  \hspace{1cm} (2.36)$$

where $N_i$ is the noise power of the $i$-th branch. Considering $u$ has been normalized, the SNR at the output of the combiner $\gamma$ is then expressed as

$$\gamma = \frac{1}{2} \frac{\left( \sum_{i=1}^{M} a_i r_i \right)^2}{\sum_{i=1}^{M} a_i^2 N_i}$$  \hspace{1cm} (2.37)$$

Applying the Schwarz inequality

$$\left| \sum_{i=1}^{M} \alpha_i^* \beta_i \right|^2 \leq \left( \sum_{i=1}^{M} |\alpha_i|^2 \right) \left( \sum_{i=1}^{M} |\beta_i|^2 \right)$$

for Eq. (2.37), we have

$$\gamma \leq \frac{1}{2} \sum_{i=1}^{M} \frac{|r_i|^2}{N_i} = \sum_{i=1}^{M} \gamma_i$$  \hspace{1cm} (2.38)$$

The equality holds if and only if

$$a_i = K \sqrt{\gamma_i}$$  \hspace{1cm} (2.39)$$
where $K$ is an arbitrary constant.

**Equal Gain Combining**

The equal gain combining diversity system is the same as the maximal ratio combining system but $a_i$ being equal to the same constant for all diversity branches. The approximate pdf for $\gamma \ll \gamma_0$ is given as [1], [3]

$$p(\gamma) = \frac{2^{M-1}M^M}{(2M-1)!} \frac{\gamma^{M-1}}{\gamma_0^M} \quad (\gamma \ll \gamma_0) \quad (2.40)$$

Among three combing methods above, the maximal combining system has the best compensation performance, then equal gain and selection combining in turn under the same conditions. In addition, when reception involves linear demodulation, it is in principle irrelevant whether the signals are combined before or after demodulation. However, for nonlinear demodulation, for example FM system, there may be a performance difference between pre-demodulation and post-demodulation combining. This is because the maximal ratio combining can benefit for pre-demodulation combining from that each individual branch SNR is below FM threshold, while the sum of the SNR's is above, but for post-demodulation combining it cannot. Selection combining will not show difference in compensation performance between pre- and post-demodulation combining, since the same choice will be made in either pre- or post-demodulation combining. It is pointed out that for FM signal dual diversity, ideal post-demodulation maximal ratio combining has only 0.5 dB advantage over selection combining [3].

**2.5 Compensation Techniques for Selective Fading**

For digital data transmission systems, any impact of the channel on signal will be finally reflected in system error rate performance. Thus, error control coding can be again thought as a
compensation techniques for frequency-selective fading. Diversity system is effective too as a counter measure against frequency-selective fading, since the waveform distortion becomes serious while the signal level drops. Some other compensation techniques are briefly introduced as follows without detailed explanation.

**Anti-multipath fading modulation**; Some digital modulation methods show immunity against frequency-selective fading, for example, π/2-TFSK (π/2 transition frequency shift keying) [21], DSK (double shift keying) [22], and so on. These modulation methods generate a phase transition that is not sensitive to, or immunity against, the intersymbol interference (ISI) at the sampling instant so as to avoid the impact of frequency-selective fading.

**Multicarrier Transmission** [23], [24]; It makes a serial to parallel convert on data. The parallel data are transmitted with several or more carriers. Since the data rate is decreased due to serial to parallel convert, the signal bandwidth is narrow for each carrier. Thus, the impacts of frequency-selective fading can be reduced.

**RAKE receiver** [1]; It uses a filter matched to the channel transfer characteristics. The matched filter outputs, at the sampling instant, a signal made by coherently combining the multipath signal components.

**Adaptive equalizer** [11], [25], [26]; It generates a proper transfer function to compensate the channel transfer function in order to cancel ISI at the sampling instant. Some equalizers are considered only for digital data transmission systems, for example, decision feedback equalizer [11], Viterbi equalizer [27]-[30], and so on.

**Adaptive array antenna (adaptive interference canceller)** [31]-[34]; It cancels the interference/echo by directing nulls of its directivity pattern to the interference/echo.

For analog signal transmission systems, the main compensation techniques for signal distortion are equalizer and adaptive array antenna (echo canceller). In the following, we briefly introduce these two techniques.
2.5.1 Equalizer

2.5.1.1 Feedback Type Equalizer

The block diagram of the feedback type equalizer is shown in Fig. 2.10. The transfer function of the equalizer is expressed as

\[ H(\omega) = \frac{1}{1 + G(\omega)} \]  

(2.41)

So called compensating the channel means letting \( H(\omega)C(\omega) = 1 \). Then, we have

\[ G(\omega) = C(\omega) - 1 \]  

(2.42)

Assume \( C(\omega) \) as

\[ C(\omega) = 1 + \sum_{i=1}^{M} r_i e^{-j\omega \tau_i} \]  

(2.43)

where the first term "1" represents the direct wave; \( r_i \) and \( \tau_i \) are amplitude (or reflection coefficient)
and excess delay of the $i$-th echo, respectively; $M$ is the number of echoes. Then $G(\omega)$ becomes

$$G(\omega) = \sum_{i=1}^{M} r_ie^{-j\alpha_i},$$

(2.44)

$G(\omega)$ is usually realized with a transversal filter. For the simplest case, where there is only one echo, $G(\omega)$ can be realized by a one tap transversal filter with tap delay and tap coefficient being equal to the excess delay $\tau$ and the amplitude $r$ of the echo respectively. Generally, several or tens of taps are enough for a feedback type equalizer used in an FM broadcasting receiver [35]. Therefore, one prefers to adopt the feedback type equalizer [35]-[39]. The feedback type equalizer will be analyzed in detail in Chapter 4.

2.5.1.2 Feedforward Type Equalizer

The block diagram of the feedforward type equalizer is shown in Fig. 2.11. To compensate the channel, we should let

$$H(\omega) = \frac{1}{C(\omega)}$$

(2.45)

Using Eq. (2.43), we have

$$H(\omega) = \frac{1}{1 + \sum_{i=1}^{M} r_ie^{j\omega \tau}},$$

(2.46)

![Fig. 2.11 Block diagram of the feedforward type equalizer.](image-url)
It can be seen from Eq. (2.46) that if $H(\omega)$ is realized with a transversal filter, infinite number
taps are needed even though there is only one echo. A 400 taps feedforward type equalizer is
proposed in [40].

### 2.5.2 Adaptive Array Antenna

We also often call it interference canceller or echo canceller if the adaptive array antenna
is applied to cancel the interference or echo. The structure of an adaptive array antenna is shown
in Fig. 2.12. The directivity pattern of the array can be controlled by adjusting its weights $w_i$. If
the nulls of the directivity pattern is directed to the interference, the interference will not be
received, that is, interference is cancelled in the received signal. An adaptation procedure
(algorithm) is necessary to adjust automatically the array weights and thereby form the required

![Fig. 2.12 Adaptive array antenna.](image)
A well-known weights control procedure is LMS (least mean squares) algorithm [41]-[43]. Using complex vector notation, we can express the array output $y(t)$ as

$$y(t) = W^T(t)X(t) \quad (2.47)$$

where the superscript $^T$ denotes the transpose of a vector; $X(t) = [x_1(t), x_2(t), ..., x_N(t)]^T$ is the vector of array element output signals; $W(t) = [w_1, w_2, ..., w_N]^T$ is the vector of array weights; $N$ is the number of the array elements. In order to measure the interference in the compensated signal, we consider a mean squared error expressed as (omit the time variable)

$$E(\varepsilon^2) = E(d - y)^2 \quad (2.48)$$

where $d$ is the desired signal, i.e., the a priori knowledge. Substituting Eq. (2.47) into Eq. (2.48), we get

$$E(\varepsilon^2) = E(d^2) + W^T \Phi(x, x) W - 2 W^T \Phi(x, d) \quad (2.49)$$

where

$$\Phi(x, x) = E(xx^T) \quad \text{and} \quad \Phi(x, d) = E(xd)$$

Differentiating Eq. (2.49) with respect to $W$ yields the gradient $\nabla E(\varepsilon^2)$ given as

$$\nabla E(\varepsilon^2) = 2 \Phi(x, x) W - 2 \Phi(x, d) \quad (2.50)$$

When the choice of the weights is optimized, i.e., the mean squared error achieves its minimum value, the gradient should be equal to zero. Letting Eq. (2.50) zero, we get the optimum weights vector that gives the least mean squared error, denoted as $W_{LMS}$,

$$W_{LMS} = \Phi^{-1}(x, x) \Phi(x, d) \quad (2.51)$$

Equation (2.51) is generally straightforward, but presents serious computational problems when the number of weights $N$ is large and when data rates are high. In addition to the necessity of inverting an $N \times N$ matrix, Eq. (2.51) may require as many as $N(N+1)/2$ autocorrelation and crosscorrelation measurements to obtain the elements of $\Phi(x, x)$. Therefore, the so called steepest
descent method is actually used to find approximate solutions to Eq. (2.51). This method adjusts the weights vector along the direction of the estimated gradient of the mean squared error function. The weights vector is then adjusted as

$$W(n+1) = W(n) - k_s \nabla e^2(n) \quad (2.52)$$

where we have returned to the use of time variable denoted as index $n$; $W(n)$ is the weights vector before adaptation (or at time index $n$); $W(n+1)$ is the vector after adaptation (or at time index $n+1$); $k_s (>0)$ is a scalar constant (often called step size) controlling rate of convergence and stability; $\nabla e^2(n)$ is the estimated gradient vector of the mean squared error function with respect to $W$. $\nabla e^2(n)$ can be solved as

$$\nabla e^2(n) = 2\varepsilon(n) \nabla \varepsilon(n)$$

$$= -2\varepsilon(n)X(n) \quad (2.53)$$

Substituting Eq. (2.53) into Eq. (2.52), we have

$$W(n+1) = W(n) + 2k_s \varepsilon(n)X(n) \quad (2.54)$$

It can be shown that the mean of the weights vector determined by Eq. (2.54) converges to $W_{LMS}$ indicated by Eq. (2.51) when $n \to \infty$ under the condition $0 < k_s < 1/e_{max}$ [31], where $e_{max}$ is the maximum eigenvalue of $\Phi$. 

2.6 Summary

In this chapter, the mechanism of the occurrence of flat fading and frequency-selective fading and their impacts on transmitted signal were described, some current compensation techniques common to digital and analog transmission systems for flat and frequency-selective fadeings were introduced.

Flat fading and frequency-selective fading occur due to the same reason, that is, the existence of reflected, refracted, and/or scattered signals. The distinction of these two fadeings lies on
whether the difference in arrival times of incident waves can be neglected or not compared to the reciprocal of the signal bandwidth. However, they have different impacts on transmitted signal. Flat fading causes the random fluctuation in signal envelope and phase, while frequency-selective fading results in distortion in signal waveform.

The most often used compensation method common to digital and analog systems for flat fading is diversity reception technique. It combines signals received at each diversity branches to avoid signal level drop. Equalizer and adaptive array antenna are introduced as the compensation methods for frequency-selective fading. The equalizer for analog system generates a inverse channel transfer function to compensate the signal for frequency-selective fading. Adaptive array antenna cancels the interference (echo) by directing the nulls of its directivity pattern to the interference.
References


Chapter 3

Effects of Beam Tilting on Bit-rate Selection in Mobile Multipath Channel

3.1 Introduction

When a signal is transmitted through a multipath mobile channel, due to frequency-selective fading, it suffers from multipath distortion, which results in intersymbol interference. As the system bit rate increases, the impacts of frequency-selective fading becomes serious, since the bandwidth of the signal is wider when system bit rate is increased. If the system bit rate is too high, intersymbol interference may be so serious that the system is rendered inoperative. Therefore, there are limits in the selection of system bit rate for data transmission through multipath mobile fading channel [1].

Various equalizers and adaptive antenna arrays can be used to compensate for multipath distortion. Equalizers and adaptive antenna arrays and their applications have been well investigated [2] - [11]. All these techniques compensate the signal for multipath distortion usually at receiving side. Each signal receiving terminal needs to be equipped with an equalizer or an adaptive array.

On the other hand, a technique called beam tilting has been adopted in order to decrease the interference between two adjacent frequency-reuse cells for cellular mobile communication
system. As a transmitted wave beam is tilted down at base station, the received power level will decrease relative to that without beam tilting at the same propagation distance as pointed out in [12] and [13]. The same technique is also expected to be applied to improve the system performance characterized by the maximum achievable bit rate by decreasing the power level of the echo signals with long delay times [14], [15], [26]. This technique compensates for multipath distortion at base station. All mobile terminals can share its benefits without additional equipments.

In this chapter, we analyze the effects of beam tilting on system bit-rate selection in mobile multipath channel under the conditions of the existence of diffusely scattered reflection in addition to the perfect reflection. The reduction of the multipath distortion by beam tilting at a base station is theoretically analyzed using multipath propagation models. The system performance, which is characterized by the critical bit rate and the probability that multipath distortion can be handled by the system, is evaluated. The results show that the capability of the system to handle multipath distortion is improved; the system bit rate can be selected over a wider range with the introduction of beam tilting than that without beam tilting.

3.2 Mechanism of Beam Tilting to Counter against Multipath Distortion

With the introduction of beam tilting in a base station, the received signal power level at mobile terminal decreases relative to that without beam tilting. Meanwhile, the power level of the echo, which causes the multipath distortion, is also decreased due to beam tilting. Since the signal and the echo have different propagation paths, the degrees of the effects of beam tilting on the signal and echo are different. This is the point that beam tilting can be used as a counter measure against multipath distortion.

In this section, we investigate the effect of beam tilting on received power level and the mechanism of beam tilting to counter against multipath distortion.
3.2.1 Effect of Beam Tilting on Received Power

The power propagation loss has been well analyzed in [13], [16]-[21] for various propagation environments. Here, the propagation loss is simply modeled as

\[ P = Ar^{-\alpha} \]  

(3.1)

where

- \( A \) a constant related to the transmitted power, antenna gain, etc.;
- \( r \) propagation distance;
- \( \alpha \) equals 2 for free space and 3 to 4 for urban terrain. We assume \( \alpha = 3.4 \) in this chapter.

If a directive transmitting antenna is used in a base station, the effect of the directivity pattern of this antenna should be taken into account. Then, the received power level at distance \( r \) can be expressed as

\[ P = AG_r(\theta)r^{-\alpha} \]  

(3.2)

where \( G_r(\theta) \) is the directivity pattern of the base station antenna. Equation (3.2) implies that an omnidirectional antenna is used in the mobile station. In the following, we assume that the base station antenna has an omni directivity pattern in horizontal plane, the antenna is directive only in vertical plane. If the transmitting wave beam is tilted downward with an angle \( \theta_0 \) in vertical plane as shown in Fig. 3.1, the \( \theta \) in Eq. (3.2) can be then expressed as

\[ \theta = \tan^{-1}(h/r) - \theta_0 \]  

(3.3)

where

- \( h \) the base station antenna height;
- \( \theta_0 \) the tilting angle as shown in Fig. 3.1;

Obviously, the effects of beam tilting are related to the directivity pattern of the transmitting antenna used in the base station. A sharp antenna directivity pattern is expected to markedly reduce the multipath distortion. Figure 3.2 shows an antenna directivity pattern, which is an
Fig. 3.1 Tilting angle $\theta_0$. The distance between the base station (BS) and mobile station (MS) is $r$. Base station antenna height is $h$.

Fig. 3.2 An antenna directivity pattern, which is an approximated pattern of an antenna currently used in Japan.
approximated pattern of an antenna practically used in Japan [22]. The half-power angle of this antenna is 3°, which is the same as the practical one. We will use this antenna pattern for all calculations in this chapter. (To consult the practical antenna's directivity pattern, refer to reference [22].)

In order to investigate the effect of beam tilting on received power level, the ratio of the received power level with a given tilting angle to that without beam tilting is calculated. The results are shown in Fig. 3.3, where the Y axis is the power ratio defined above; the X axis is the propagation distance. An omnidirectional antenna is used for the case without beam tilting. It can be seen that relative to the case without beam tilting, the received power level is decreased with the increase in the propagation distance. Moreover, compared to the case that tilting angle is equal to 3°, the received power level is more rapidly decreased when tilting angle is equal to 7.5°. That is, a little large tilting angle is needed for rapidly power level decreasing.

3.2.2 Mechanism of Beam Tilting to Counter against Multipath Distortion

Multipath distortion, which causes intersymbol interference, results from the echo with excess time delay relative to the direct wave. If the echo power can be decreased, multipath distortion can be also reduced. Causing an extra attenuation on the received echo power is the basic mechanism of beam tilting to be used as a counter measure against multipath distortion.

Assume that the propagation distances of the direct wave are \( r \). Using Eqs. (3.2) and (3.3), the direct wave power level received by the mobile station, \( P_m \), can be expressed as

\[
P_m = Ar^{-\alpha}G_1\left[\tan^\alpha\left(\frac{h}{r}\right) - \theta_0\right]
\]

(3.4)

In order to calculate the echo power, it is needed to be known that the effect of the antenna directivity pattern on the echo propagation. We will discuss this effect in section 3.3. It is clear, however, the antenna has different effects on the direct wave and the echo, since they propagate through different paths. Let us now simply assume that the echo is equivalently received by the
Fig. 3.3 The effect of beam tilting on received power level. The Y axis, power ratio, is defined as the ratio of the received power level with a given tilting angle to that without beam tilting. Antenna height is 100 m, $\alpha$ is 3.4.
mobile station at a propagation distance \( r+x \), where \( x \) is called excess distance. Then, the echo power level \( P_e \) can be expressed as

\[
P_e = A(r+x) \alpha G_r \left[ \tan^{-1} \left( \frac{h}{r+x} \right) - \theta_0 \right]
\]

As a result, the echo power level with the excess distance \( x \) will be attenuated with respect to the direct wave power level by a factor \( \gamma \), called (received) signal to echo power ratio. \( \gamma \) is defined as

\[
\gamma = \frac{P_m}{P_e} = \frac{G_r \left[ \tan^{-1} \left( \frac{h}{r} \right) - \theta_0 \right]}{G_r \left[ \tan^{-1} \left( \frac{h}{r+x} \right) - \theta_0 \right]} \left( \frac{r+x}{r} \right)^\alpha
\]

The signal to echo power ratio \( \gamma \) versus the excess delay \( x \) curves are shown in Fig. 3.4. It can be seen that as the excess distance \( x \) increases, the signal to echo power ratio \( \gamma \) also increases. When \( \gamma \) reaches a certain value, 10 dB for example, the echo can be disregarded because of the propagation loss. In this situation, the corresponding value of \( x \) is called the maximum excess distance denoted as \( x_m \). This \( x_m \) can be determined from Eq. (3.6) if the \( \gamma \) value at which the echo can be neglected is given.

Comparing Fig. 3.4 (a) and (b), we can also see that the rate of the increase of \( \gamma \) with respect to the increase in the excess distance \( x \) is different for different tilting angle values. The increase of \( \gamma \) for tilting angle = 7.5° is much faster than that for tilting angle = 3°. The beam tilting technique is originally introduced to decrease the interference between two adjacent frequency-reuse cells. For this purpose, the value of the tilting angle is small, 3° for example. The radius of a cell is much smaller than the distance between two adjacent frequency-reuse cells. Therefore, if we expect that the signal to echo power ratio \( \gamma \) could increase fast to suppress echo power in the area of a cell so that the system can be operated at a higher bit-rate by using the beam tilting technique, a slightly larger value of the tilting angle is needed. We select the tilting angle of 7.5°. A larger tilting angle value may make \( \gamma \) increase faster, but at too large a
Fig. 3.4 Signal to echo power ratio $\gamma$ versus the excess distance $x$ curves. Antenna height is 100 m, propagation distance of the directive wave, $r$, is 2 km, and $\alpha$ is 3.4.
tilting angle, the sidelobe effects of the transmitting antenna will offset the echo suppression by beam tilting.

### 3.3 Echo Propagation Models

Different types of surface roughness will produce different effects on wave reflection as shown schematically in Fig. 3.5 [23]. In particular, if the signal reflection occurs at a rough surface, diffusely scattered reflection [16] will happen and the reflected wave will disperse more of its energy before it reaches the mobile station antenna. As a result, the echo power levels at the mobile station will be different for different types of the reflection surface.

In the following, we establish the multipath propagation models for two cases where the transmitted wave is reflected on a smooth surface and on a rough surface, respectively.

![Figure 3.5 Effects of different types of surface roughness on wave reflection.](image)

(a) Smooth surface. (b) Rough surface.

3.3.1 Reflection on Smooth Surface

Figure 3.6 shows the case that the wave is reflected on a smooth surface. Here, it is assumed that the reflection is perfect, or say, specular reflection (unity reflection coefficient and no diffusely scattered reflection). The locus of echoes that have the same excess distance $x$ is an ellipse as
shown in Fig. 3.7. For this case, we can think that the echo is received at the location of the image of the mobile station. Since the base station antenna has an omni directivity pattern in the horizontal plane, Fig. 3.6 can be reduced to Fig. 3.8. According to Fig. 3.8, we can obtain the received direct wave power $P_m$ and the echo power $P_e$ as

$$P_m = A r^{-\alpha} G_i \left[ \tan \left( \frac{h}{r} \right) - \theta_0 \right]$$  \hspace{1cm} (3.7)$$

and

$$P_e = A (r+x)^{-\alpha} G_i \left[ \tan \left( \frac{h}{r+x} \right) - \theta_0 \right]$$  \hspace{1cm} (3.8)$$

respectively, where the impact of the base station antenna height $h$ on the calculation of the propagation distance is neglected, since $h$ is much less than $r$ or $r+x$. The signal to echo power

![Fig. 3.6 Echo occurs on a smooth surface.](image)
Fig. 3.7 The Locus of echoes with the same excess distance.

\[ r_1 + r_2 = r + x \]

Fig. 3.8 The equivalent propagation path of the direct wave and the echo.

\[ r_1 + r_2 = r + x \]
ratio $\gamma$ can be then expressed as

$$\gamma = \frac{P_m}{P_e} = \frac{G_t \left[ \tan^{-1} \left( \frac{h}{r} \right) - \theta_0 \right]}{G_t \left[ \tan^{-1} \left( \frac{h}{r+x} \right) - \theta_0 \right]} \left( \frac{r+x}{r} \right)^\alpha$$  \hspace{1cm} (3.9)

### 3.3.2 Reflection on Rough Surface

As shown in Fig. 3.5, when the transmitted wave is reflected on a rough surface, the so-called "diffusely scattered reflection" [16] will occur. The propagation of the direct wave and the echo for this case is shown in Fig. 3.9. Since the base station antenna has a omni directivity pattern in the horizontal plane, Fig. 3.9 can be reduced to Fig. 3.10. Different from the case that the echo occurs on a smooth surface, where the base station antenna directivity pattern shows its effect on the echo between the base station and the image of the mobile station as shown in Fig. 3.8, in this case, the base station antenna directivity pattern is effective only between the base station and the echo.
station and the point at which the transmitted wave is reflected, whose distance is \( r_1 \) as shown in Fig. 3.10. The locus of echoes which are reflected at the same distance away from the base station is a circle as shown in Fig. 3.11. Moreover, due to the diffuse reflection, the reflected wave will disperse more of its energy before it reaches the mobile station antenna compared to the perfect reflection case. Considering such additional energy dispersion occurrence after the wave is reflected and the distance between the reflection point and the mobile station is \( r_2 \) as shown in Fig. 3.11, we use a function of \( r_2 \) to simulate this energy dispersion.

Considering the reasons mentioned above and assuming that the reflected wave can be always received by the mobile station, the echo power \( P_e \) can be expressed as

\[
P_e = A (r+x)^a G_i \left[ \tan^{-1} \left( \frac{h}{r_1} \right) - \theta_0 \right] R(r_2)
\]  

(3.10)
Reflection point

Fig. 3.11 The locus of echoes which are reflected at the same distance away from the base station is a circle.

where \( r_1 + r_2 = r + x \); \( R(r_2) \) is a function of \( r_2 \), which is used to simulate the additional energy dispersion. It may be appropriate to assume that \( R(r_2) \) has the form as

\[
R(r_2) \propto \left( \frac{1}{r_2} \right)^\beta
\]

(3.11)

where \( \beta \) is a parameter related to the degree of additional energy dispersion. The larger value of \( \beta \) means the more additional energy dispersion. Here, it is implicitly assumed that the reflection coefficient is equal to one. The direct wave power received by the mobile station is the same as that expressed by Eq. (3.7). The signal to echo power ratio \( \gamma \) can be then expressed as

\[
\gamma = \frac{P_m}{P_e} = \frac{G_t \left[ \tan^{-1} \left( \frac{h}{r} \right) - \theta_0 \right]}{G_t \left[ \tan^{-1} \left( \frac{h}{r_1} \right) - \theta_0 \right]} \left( \frac{r+x}{r} \right)^\alpha R^{-1}(r_2)
\]

(3.12)
Equations (3.9) and (3.12) express the signal to echo power ratio $\gamma$ for the smooth surface case and rough surface case, respectively. As pointed out in section 3.2.2, if the $\gamma$ value at which the echo can be neglected is given, we can calculate the maximum excess distance $x_m$ with Eqs. (3.9) and (3.12) for the smooth surface and rough surface cases, respectively.

### 3.4 Effects of Beam Tilting on Bit-rate Selection

Due to beam tilting technique, the echo power level received at the mobile station has additional attenuation. Thus, the impacts of the echo on the received signal can be mitigated. That is, the multipath distortion can be suppressed to a certain extent. As a result, the system can be operated at a higher bit rate compared to the case without beam tilting.

In this section, we first introduce the measure of the system performance and then evaluate the effect of beam tilting on bit-rate selection.

#### 3.4.1 System Performance Measures

The probabilities of occurrence of the following events are considered to significantly affect communication system performance. These events can be expressed as [1]:

- **R**: The multipath distortion caused by echoes can be handled by the receiver. Assume that a receiver can handle all echoes whose excess time delay relative to the direct wave, denoted as $\tau$, is less than $\nu T$, where $T$ is the data symbol duration; $\nu$ is a parameter determined by the capability of the receiver handling the multipath distortion, for example, $\nu=3.3$ or larger for a receiver with an adaptive equalizer, and $\nu=0.2$ for a diversity receiver [24]. Then, within the range $\tau < \nu T$, the echo will not cause serious multipath distortion, or exactly speaking, the impact of the echo can be handled by the receiver. The probability of occurrence of event $R$ is...
\[ \text{Prob}(R) = \text{Prob}(\tau < vT) \] (3.13)

\text{D:} The echo can be disregarded because of propagation loss. The received power level of the echo with the excess distance \( x \), or equivalently the excess time delay \( \tau = x/c \), where \( c \) is the speed of light, is less than the direct wave power level by a factor \( \gamma \) as indicated in Eq. (3.9) or (3.12). If \( \gamma \) reaches a certain value, the echo can be neglected. The corresponding excess distance is called \( x_m \). That is, if \( x > x_m \), or equivalently \( \tau > x_m/c \), the echo can be neglected. \( x_m \) can be determined by Eqs. (3.9) and (3.12) for smooth surface case and rough surface case, respectively, if \( \gamma \) is given. The probability of occurrence of event \( D \) is

\[ \text{Prob}(D) = \text{Prob}(\tau > x_m/c) \] (3.14)

\text{S:} Serious intersymbol interference occurs and the system is rendered inoperative. If the excess time delay of the echo is greater than \( vT \) and less than \( x_m/c \), the echo can neither be handled by the receiver, nor can be disregarded due to the propagation loss. As a result, serious intersymbol interference will occur. The probability of occurrence of event \( S \) is

\[ \text{Prob}(S) = \text{Prob}(vT < \tau < x_m/c) \] (3.15)

The events defined above are illustrated in Fig. 3.12. From the definitions of the events or Fig. 3.12, it can be seen that event \( R \) and \( D \) have the same meaning: echo will not cause problems in the receiver. Therefore, we use two probabilities to measure the system performance. One is the probability of occurrence of event \( S \), denoted as \( P_S \); the other one is the probability of occurrence of event \( R \) or \( D \), denoted as \( P_{RorD} \), which is equal to \( 1 - P_S \).

It can be also seen from Fig. 3.12 that the range of event \( S \) is related to the data symbol duration \( T \) besides \( v \) and \( x_m \). Thus, the range of event \( S \) will change as data symbol duration \( T \) changes. We of course do not expect the occurrence of event \( S \). Therefore, the selection of
Fig. 3.12 Events for different excess time delay. The X-axis is the excess delay of the echo relative to the propagation delay of the direct wave. \( c \) is the speed of light; \( x \) is the excess distance; \( T \) is the data symbol duration; \( v \) is a parameter representing the performance of the receiver to handle the echo.

The system bit rate is limited as

\[
vT \geq x_m/c \tag{3.16}
\]

that is

\[
\frac{1}{T} \leq \frac{vc}{x_m} \tag{3.17}
\]

Under the condition that Eq. (3.17) holds, event S will not occur and the system performance can be maintained at a high level. This is because the echo can be handled by the receiver when excess delay \( \tau < vT \) and echo can be disregarded when \( \tau > vT \) due to \( vT \geq x_m/c \). The system bit rate when the equality of Eq. (3.17) holds is called critical rate. The system bit rate should be less than the critical rate to avoid the occurrence of event S. Otherwise, serious intersymbol interference will occur and the system tends toward failure. The critical rate is also used as a measure of the system performance.
3.4.2 Effects of Beam Tilting on Bit-rate Selection

Using the propagation models illustrated in section 3.3 and the system performance measures described in section 3.4.1, we investigate the effect of beam tilting on system bit rate selection.

3.4.2.1 Smooth Surface

Assume the mobile is moving around slowly, the reflection is perfect, and the probability of occurrence of an echo is proportional with the area which is illuminated. Then the probability that the excess distance of the echo, $X$, is greater than a given value, say $x$, is given by Maseng in [1] and can be expressed as

$$\text{Prob}(X > x) = \text{Exp} \left[ -\frac{\Lambda}{4} \left( \frac{x}{r} + 1 \right) \sqrt{\left( \frac{x}{r} \right)^2 + \frac{2x}{r}} \right]$$  (3.18)

where $r$ is the distance between the base station and the mobile station, $\Lambda = \rho \pi r^2$, and $\rho$ is the echo density. According to [1] and [25], $\Lambda$ is around 10 for an urban environment and smaller for less irregular terrains. $\Lambda$ is assumed here to be equal to 10. Since the excess delay of the echo, $\tau$, can be expressed as $\tau = X/c$, the probability that the excess delay $\tau$ is greater than a given value, say $\tau_0$, can be derived from Eq. (3.18) as

$$\text{Prob}(\tau > \tau_0) = \text{Prob}(X > c \tau_0)$$

$$= \text{Exp} \left[ -\frac{\Lambda}{4} \left( \frac{c \tau_0}{r} + 1 \right) \sqrt{\left( \frac{c \tau_0}{r} \right)^2 + \frac{2c \tau_0}{r}} \right]$$  (3.19)

Using Eqs. (3.15) and (3.19), $P_S$ can be expressed as

$$P_S = \text{Exp} \left[ -\frac{\Lambda}{4} \left( \frac{vc T}{r} + 1 \right) \sqrt{\left( \frac{vc T}{r} \right)^2 + \frac{2vc T}{r}} \right] - \text{Exp} \left[ -\frac{\Lambda}{4} \left( \frac{x_m}{r} + 1 \right) \sqrt{\left( \frac{x_m}{r} \right)^2 + \frac{2x_m}{r}} \right]$$  (3.20)
and $P_{\text{RorD}}$ can be expressed as

$$P_{\text{RorD}} = 1 - P_S$$

$$= 1 - \text{Exp}\left[-\frac{\Lambda}{4} \left(\frac{vcT}{r} + 1\right)\sqrt{\left(\frac{vcT}{r}\right)^2 + \frac{2vcT}{r}}\right] + \text{Exp}\left[-\frac{\Lambda}{4} \left(\frac{x_m}{r} + 1\right)\sqrt{\left(\frac{x_m}{r}\right)^2 + \frac{2x_m}{r}}\right]$$

(3.21)

Using the propagation loss models and Eqs. (3.20) and (3.21), the probabilities defined in section 3.4.1 as well as the critical rate with beam tilting are calculated and compared to that without beam tilting as shown in Figs. 3.13, 3.14, and 3.15, respectively. Here, Figs. 3.13 and 3.14 are for the case $v = 3.3$ and 5.3, respectively, and Fig. 3.15 is for the case $v = 0.2$. From these figures, it can be seen that at the same system bit-rate, $P_{\text{RorD}}$ with beam tilting is greater than that without tilting, while $P_S$ with beam tilting is less than that without beam tilting. These results indicate that the capability of the system in handling multipath distortion has been improved. In particular, compared to the case without beam tilting, the system bit-rate at which $P_S$ begins to appear is increased with beam tilting, which means that the system can operate at a higher bit rate. The critical rates are increased from 511, 821, and 31 kb/s to 2174, 3492, and 131 kb/s for the cases that $v = 3.3$, $v = 5.3$, and $v = 0.2$, respectively. The increase in critical rate is over 4 times for any case.

As pointed out in section 3.4.1, when the system bit rate is below the critical rate, serious intersymbol interference will not occur, otherwise, it will occur and the system tends toward failure in handling the multipath distortion. Such a situation is obviously shown in Figs. 3.13 and 3.14. The points at which $P_S$ curves intersect the X axis (probability = 0.001) give the approximate critical rate (the exact value of the critical rates are given by the points where $P_S$ curves intersect the line of probability = 0). It can be seen that when the system bit rate is less than the critical rate, which is the intersection of $P_S$ curve and X axis as explained above, $P_S$ is very small and can be neglected. This means that the system can handle the multipath distortion. In contrast, when bit rate is greater than the critical rate, $P_S$ increases rapidly. This means that the system tends toward failure to operate.
Fig. 3.13 $P_{\text{RorD}}$ and $P_S$ versus normalized system bit rate $r/T_c$. $r = 2$ km, $\alpha = 3.4$, $v = 3.3$, $\Lambda = 10$, $\gamma = 10$, antenna height = 100 m, and tilting angle = 7.5°. The critical rates are 511 kb/s without beam tilting and 2174 kb/s with beam tilting, respectively.
Fig. 3.14 $P_{\text{RorD}}$ and $P_5$ versus normalized system bit rate $r/Tc$. $r = 2$ km, $\alpha = 3.4$, $v = 5.3$, $\Lambda = 10$, $\gamma = 10$, antenna height = 100 m, and tilting angle = 7.5°. The critical rates are 821 kb/s without beam tilting and 3492 kb/s with beam tilting, respectively.
Fig. 3.15 $P_{RorD}$ and $P_S$ versus normalized system bit rate $r/Tc$. $r = 2$ km, $\alpha = 3.4$, $\nu = 0.2$, $\Lambda = 10$, $\gamma = 10$, antenna height = 100 m, and tilting angle = 7.5°. The critical rates are 31 kb/s without beam tilting and 132 kb/s with beam tilting, respectively.
Figure 3.16 shows the critical rate versus tilting angle curve for $v = 3.3$ case. It can be seen that the critical rate increases with the increase in the tilting downward angle if the tilting angle is not greater than $7.5^\circ$. When the tilting angle is greater than $7.5^\circ$, the critical rate does not continue to increase, on the contrary, it decreases. This result is due to that the sidelobe of the base station antenna pattern offsets the echo suppression by beam tilting when the tilting angle is too large.

Figure 3.17 shows the critical rate versus $r$ curves. For tilting angle $= 6.5^\circ$, the critical rate is increased by about 3 times with the introduction of beam tilting in the range of $1 \text{ km} < r < 3.5$ km. For tilting angle $= 7.5^\circ$, the critical rate is increased by 4 times in the range of $1 \text{ km} < r < 2$ km. When $r > 2$ km, the effect of beam tilting is reduced for case tilting angle $= 7.5^\circ$. This is also because the side lobe of the base station antenna offsets the echo suppression when the tilting angle is too large.

Fig. 3.16 The critical rate versus tilting angle curve. $r = 2 \text{ km}$, $\alpha = 3.4$, $v = 3.3$, $A = 10$, $\gamma = 10$, antenna height $= 100 \text{ m}$. 
Fig. 3.17 The critical rate versus $r$ curves. Tilting angle=6.5°, $\alpha = 3.4$, $\nu = 3.3$, $\Lambda = 10$, $\gamma = 10$, antenna height = 100 m.

angle is large. According to different cell radii, we can select different tilting angles to get the best effects of beam tilting. For example, if cell radius is less than 2 km, we can use the tilting angle of 7.5°; if the radius is less than 3.5 km, the tilting angle of 6.5° may be used.

3.4.2.2 Rough Surface

The excess distance is defined as

$$r_1 + r_2 = r + x$$  \hspace{1cm} (3.22)

By inspection of Fig. 3.11, it can be seen that

$$r_1 + r > r_2$$  \hspace{1cm} (3.23)

$$r_2 + r > r_1$$  \hspace{1cm} (3.24)

Rearranging and combining (3.22)-(3.24) yields
\[
x/2 < r_2 < r + x/2
\]
(3.25)

We assume \( R(r_2) = 1 \) when \( r_2 = x/2 \), which is the minimum value of \( r_2 \) as indicated by Eq. (3.25).

That is, \( R(r_2) \) can be expressed as

\[
R(r_2) = \left( \frac{x/2}{r_2} \right)^\beta = \left( \frac{x/2}{r + x - r_1} \right)^\beta
\]
(3.26)

Substituting Eq. (3.26) into Eq. (3.12), we can solve the maximum value of the excess distance, \( x_m \), if \( \gamma \) is given. In the rough surface case, \( x_m \) is a function of \( r_1 \), the distance between the base station and the reflection point (see also Fig. 3.11). We then work out the critical rate from \( x_m \).

The calculation results are shown in Fig. 3.18. It can be seen that when the reflection point is not so close to the base station, that is, when \( r_1 \) is not so small, the critical rate is usually large, while

![Diagram showing critical rate versus \( r_1 \) curves. \( r_1 \) is the distance between the base station and the reflection point. \( r = 2 \) km, \( \alpha = 3.4 \), \( v = 3.3 \), \( \Lambda = 10 \), \( \gamma = 10 \), and antenna height = 100 m.](image)
the critical rate is low when $r_i$ is small. This result can be interpreted that the transmission antenna of the base station causes an extra attenuation on the echo power due to beam tilting when $r_i$ is not so small, while the reflection point will be strongly illuminated also due to beam tilting when $r_i$ is small as shown in Fig. 3.19, which results in the decrease in the critical rate.

The comparison between tilting and no tilting cases is shown in Fig. 3.20. It can be seen that the critical rate for tilting case is greater than that for no tilting case when $r_i$ is greater than 2000 m, which is just equal to $r$, the distance between the base station and the mobile station.

Fig. 3.19 The situation when $r_i$ is less than $r$. The reflection point is strongly illuminated, and the beam tilting cannot cause an extra attenuation on the echo.
Fig. 3.20 Comparison between tilting and no tilting cases. $r = 2 \text{ km}$, $\alpha = 3.4$, $\nu = 3.3$, $A = 10$, $\gamma = 10$, antenna height = 100 m, $\beta = 1$, and tilting down angle = 7.5°.
3.5 Summary

In this chapter, we have explained the mechanism of the beam tilting to counter against multipath distortion, established the multipath propagation models, and evaluated the effect of beam tilting at a base station on the system bit rate selection for two cases of perfect reflection and diffusely scattered reflection.

The system performance, which is characterized by the probability that multipath distortion can be handled by the system, is improved if the beam tilting technique is introduced under the condition of perfect reflection (unity reflection coefficient and no diffusely scattered reflection). In particular, the critical rate is increased by several times. Considering the existence of diffusely scattered reflection, the system critical rate is related to $r_i$, the distance between the base station and the reflection point. When $r_i$ is not too small, beam tilting technique can also show fairly good effects.

In short, if the potential reflection points are not quite close to the base station, especially in mountainous areas, where $r_i$ is usually greater than $r$, the system bit rate can be selected in a wider range with the introduction of beam tilting.
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Chapter 4

Feedback Type Multipath Distortion Canceller for an FM Broadcasting Receiver

4.1 Introduction

It has been essential to provide the high quality audiovisual effects so as to satisfy the requirements of people pursuing the melodious audiovisual entertainment. Compensating the multipath channel by cancelling the multipath distortion caused by frequency-selective fading is one of the several ways to improve the audiovisual effects for an FM broadcasting receiver used in a multipath channel. Some schemes and products have been presented to cancel the multipath echo in the FM broadcasting receiver, TV receiver, video recorder, and so on [1]-[9].

A multipath echo distortion canceller or say an equalizer usually consists of a transversal filter with taps. Considering their structures, echo distortion cancellers can be classified into two categories: feedforward type and feedback type. A feedforward echo distortion canceller consisting of a transversal filter needs a lot of taps, for example a 400 taps equalizer was ever proposed in [10], while several or tens of taps are enough for a feedback type equalizer [1]. This is the reason why we prefer a feedback type echo distortion canceller, since small number of taps can decrease not only the size of the echo distortion canceller, but also the amount of calculation to control the tap coefficients. However, all feedback equalizers presented in [1]-[4]
need to measure the channel transfer function to make the equalizing algorithm work correctly. Since the channel transfer function can be measured only in a certain range from the received signal, the performance of such an equalizer may degrade, for example work in [2] showed that the performance achieved when the excess time delay of the echo is small, less than 5 μs for example, is not ideal. On the other hand, the excess time delay is usually 2-4 μs, and sometimes it is 4-8 μs [2], [11]. Therefore, it is necessary to improve the performance of the feedback type echo distortion canceller. For this purpose, some improved feedback type multipath distortion cancellers for an FM broadcasting receiver are proposed [12], [14].

In this chapter, firstly we analyze the operating principle of a feedback type echo distortion canceller for use in an FM broadcasting receiver; study its compensating features and the effects of the practical operating environment on its performance. On the basis of the analysis, we then propose an adaptive feedback type echo distortion canceller operating with Constant Modulus Algorithm (CMA) and examine its performance.

4.2 Operation Principle of Feedback Type Multipath Distortion Canceller

For an analog transmission system, the multipath distortion canceller, which is also called echo distortion canceller, equalizer, and so on, generates an inverse transfer function of the channel transfer function so that the total transfer function can become flat in order to compensate the channel for multipath distortion. In this section, we will discuss that under what condition the channel can be compensated with a feedback type multipath distortion canceller; work out the values of the parameters of such a canceller realized with a transversal filter; analyze the feature of the compensation performance of the feedback type echo distortion canceller; investigate the effect of the tap interval of the transversal filter on the performance of the multipath distortion canceller.
4.2.1 Compensation Condition for a Feedback Type Multipath Distortion Canceller

A feedback type multipath distortion canceller and the echo model are shown in Fig. 4.1, where $r_i$, $\theta_i$, and $\tau_i$ are the reflection coefficient, additional phase shift, and the excess time delay of the i-th echo, respectively.

According to Fig. 4.1, the transfer function of the multipath channel, denoted as $C(\omega)$, can be expressed as

$$C(\omega) = 1 + \sum_{i=1}^{M} r_i e^{j(\omega \tau_i + \theta_i)}$$

$$= 1 + E(\omega)$$

(4.1)

where

$$E(\omega) = \sum_{i=1}^{M} r_i e^{j(\omega \tau_i + \theta_i)}$$

(4.2)

and $M$ is the number of echoes. To compensate the channel we can let

$$G(\omega) = E(\omega)$$

(4.3)

Fig. 4.1 Echo model and the principle diagram of a feedback type echo distortion canceller.
and then we have

\[
H(\omega) = \frac{1}{1 + E(\omega)} = C^{-1}(\omega)
\]

(4.4)

The total channel transfer function will be then \(C(\omega)H(\omega) = 1\). Equation (4.3) represents the condition for a feedback type echo distortion canceller so that channel can be compensated. In this chapter, we assume this condition and will not distinguish between \(G(\omega)\) and \(E(\omega)\) for simplicity of description.

### 4.2.2 Realization of Feedback Type Multipath Distortion Canceller

Before beginning the discussion, we assume that the channel characteristics expressed by parameters \(r_p, \theta_p\), and \(\tau_p\), or equivalently the channel transfer function over the whole frequency range, are unknown. If these parameters are known, the echo distortion can be completely cancelled with a programmable delay line feedback echo distortion canceller [12]. In implementing, we assume an equalizer with digital transversal filter: that is a sampled systems.

For an echo distortion canceller, \(G(\omega)\) is usually realized in time domain. According to the relationship between frequency and time domain, the corresponding impulse response \(g(t)\) is the inverse Fourier transform of \(G(\omega)\). Since delay never introduces distortion, \(g(t)\) can be the inverse Fourier transform of \(G(\omega)e^{-j\omega T_0}\), where \(T_0\) is a time constant usually being equal to zero. Then the impulse response \(g(t)\) can be expressed as

\[
g(t) = F^{-1}\left[ G(\omega)e^{-j\omega T_0} \right] \\
= \frac{1}{2\pi} \int_{-\infty}^{\infty} G(\omega)e^{-j\omega T_0} e^{j\omega t} \, d\omega \\
= \frac{1}{2\pi} \int_{-\infty}^{\infty} G(\omega) e^{j\omega(1-T_0)} \, d\omega
\]

(4.5)
where $F^{-1}$ represents inverse Fourier transform. Time constant $T_o$ means that a time shift for the corresponding impulse response is permitted. For some special cases, the introduction of the time constant $T_o$ can simplify the echo distortion canceller and improve its compensation performance. We will show the effects of $T_o$ on the echo distortion canceller in sections 4.2.3 and 4.3.2.

Usually, it is not necessary to compensate the channel over the whole frequency range. Compensating the channel in a certain range, say $B$, is enough. Therefore, we truncate $G(\omega)$ within the range $B$. Furthermore, to get a discrete impulse response function so that the echo distortion canceller can be realized with a transversal filter, we expand the truncated $G(\omega)$ into a periodic function, denoted as $G_p(\omega)$, as shown in Fig. 4.2. $G_p(\omega)$ can be mathematically expressed as

$$
G_p(\omega) = \sum_{n=-\infty}^{\infty} G_t(\omega-nB) \\
= G_t(\omega) * \sum_{n=-\infty}^{\infty} \delta(\omega-nB) \tag{4.6}
$$

where

$$
G_t(\omega) = \begin{cases} 
G(\omega) & \omega \in B \\
0 & \text{Otherwise}
\end{cases}
$$

Since the inverse Fourier transform of $G_p(\omega)$ has equally spaced discrete components due to that $G_p(\omega)$ is a periodic function, $G_p(\omega)$ can be easily realized in time domain with a transversal filter. The impulse response of the transversal filter is the inverse Fourier transform of $G_p(\omega)$. Thus, the weights of the discrete components and the discrete component spacing of the inverse Fourier transform correspond to the tap coefficients and the tap interval of the transversal filter, respectively. Calculating the inverse Fourier transform of Eq. (4.6), denoted as $g_p(t)$, we can easily obtain the tap coefficients, denoted as $w_i$, and the tap interval, denoted as $T_d$. Due to the same reason mentioned above, a time shift for $g_p(t)$ is permitted. Therefore, $g_p(t)$ can be expressed...
truncating within a range B

expanding into a periodic function

Fig. 4.2 The relationship among $G(\omega)$, $G_t(\omega)$ and $G_p(\omega)$. 
as

\[ g_p(t) = F^{-1}[G_p(\omega)e^{-j\omega T_0}] \]  

(4.7)

Using [15]

\[ \sum_{n=-\infty}^{\infty} \delta(t-nT) \leftrightarrow \omega_0 \sum_{n=-\infty}^{\infty} \delta(\omega-n\omega_0) \]

and

\[ 2\pi f_1(t)f_2(t) \leftrightarrow F_1(\omega)F_2(\omega) \]

where \( \omega_0 = 2\pi/T \); "\( \leftrightarrow \)" represents Fourier transform pair; \(*\) means the convolution integral; and \( f_1(t) \leftrightarrow F_1(\omega) \) \( f_2(t) \leftrightarrow F_2(\omega) \), we have

\[ g_p(t) = F^{-1}[G_p(\omega)e^{-j\omega T_0}] \]

\[ = \frac{2\pi}{B} F^{-1}[G_1(\omega)] \sum_{i=-\infty}^{\infty} \delta(t - \frac{2\pi i}{B} - T_0) \]  

(4.8)

Here, \( \sum_{i=-\infty}^{\infty} \delta(t - \frac{2\pi i}{B} - T_0) \) is an impulse train with period being equal to \( 2\pi/B \). Thus, \( T_0 \) is limited in the range \( 0 < T_0 \leq 2\pi/B \). The impulse train can be also expressed as

\[ \sum_{i=-\infty}^{\infty} \delta\left(t - \frac{2\pi(i+1)}{B} + \frac{2\pi}{B} - T_0\right) \]  

(4.9)

Letting \( i' = i+1 \) and \( T_0' = 2\pi/B - T_0 \) and still denoting \( i' \) as \( i \) and \( T_0' \) as \( T_0 \), then, the impulse train can be expressed as

\[ \sum_{i=-\infty}^{\infty} \delta\left(t - \frac{2\pi n}{B} + T_0\right) \]  

(4.10)

Using the above equation in Eq. (4.8), the impulse response of the transversal filter can be expressed as

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Then, the tap coefficients $w_i$ and the tap interval $T_d$ of the transversal filter are

$$w_i = \frac{2\pi}{B} \mathcal{F}^{-1} [G_i(\omega)] \bigg|_{t = \frac{2\pi i}{B} - T_0} \quad (4.12)$$

and

$$T_d = \frac{2\pi}{B} \quad (4.13)$$

respectively. Here $t = 2\pi i / B - T_0$ corresponds to the delay of tap $w_i$. We should pay attention to Eq. (4.13). It shows the relationship between the tap interval and $B$, the bandwidth that needs to be compensated. Without loss of generality, assume the bandwidth that needs to be compensated is $(-B/2, B/2)$. Substituting Eq. (4.3) into Eq. (4.12) and using Eqs. (4.2) and Eq. (4.13), we obtain

$$w_i = \frac{2\pi}{B} \mathcal{F}^{-1} [G_i(\omega)] \bigg|_{t = \frac{2\pi i}{B} - T_0}$$

$$= \frac{2\pi}{B} \int_{B/2}^{B/2} \sum_{k=1}^{M} r_k e^{-j(\omega_\xi_k + \theta_k)} e^{j\omega_d t} d\omega \bigg|_{t = \frac{2\pi i}{B} - T_0}$$

$$= \frac{1}{B} \sum_{k=1}^{M} r_k e^{j\theta_k} \left[ \frac{e^{jB(t-\tau_k)}}{B/2} - e^{jB(t-\tau_k)} \right] \bigg|_{t = \frac{2\pi i}{B} - T_0}$$

$$= \frac{1}{B} \sum_{k=1}^{M} r_k e^{j\theta_k} \left[ \frac{e^{jB(t-\tau_k)}}{j(t-\tau_k)} - e^{jB(t-\tau_k)} \right] \bigg|_{t = \frac{2\pi i}{B} - T_0}$$

$$= \sum_{k=1}^{M} r_k e^{j\theta_k} \frac{\sin \left[ \pi \left( (i-1)T_d + T_{ini} - \tau_k \right) / T_d \right]}{\pi \left[ (i-1)T_d + T_{ini} - \tau_k \right] / T_d}$$

(4.14)
where $T_{in} = 2\pi/B - T_0$ (0 < $T_{in}$ ≤ $T_d$), which is the time delay of tap 1 ($w_1$), called initial delay. For the simplest case, where there is only one echo, the tap coefficients $w_i$ can be shown from Eq. (4.14) as

$$w_i = r e^{-j\theta} \frac{\sin \left[ \frac{\pi ((i-1)T_d + T_{in} - \tau)}{T_d} \right]}{\pi \left[ \frac{\pi ((i-1)T_d + T_{in} - \tau)}{T_d} \right]}$$

(4.15)

where $\tau$, $\theta$, and $r$ are the excess time delay, additional phase shift, and the reflection coefficient of the echo, respectively; The result indicated by Eq. (4.15) is the same with that obtained in [1], though complete different methods are used in [1] and here. The feedback type echo distortion canceller realized with a transversal filter is shown in Fig. 4.3.

Fig. 4.3 Feedback type echo distortion canceller realized with a transversal filter.
4.2.3 Feature of Feedback Multipath Distortion Canceller

The transversal filter used in the feedback type should realize the impulse response indicated by Eq. (4.11). Without loss of generality, assuming $T_0 = 0$, and using Eqs. (4.12), (4.13), Eq. (4.11) can be then expressed as

$$g_p(t) = \sum_{i=\infty}^{\infty} w_i \delta(t - iT_d)$$  \hspace{1cm} (4.16)

The output of the feedback echo distortion canceller at time $t$ is then expressed as (also refer to Fig. 4.3)

$$y(t) = x(t) - \sum_{i=1}^{\infty} w_i y(t - iT_d) - w_0 y(t) - \sum_{i=\infty}^{\infty} w_i y(t - iT_d)$$  \hspace{1cm} (4.17)

If the third term at the right hand side of the equality sign in Eq. (4.17) is moved to the left side, we can see that the third term will not affect the output of the echo distortion canceller except for a proportional ratio. That is, the third term does not play any part in compensating the signal. Therefore, the third should be discarded. The fourth term of the right side uses the future output of the echo distortion canceller. Thus, the fourth term must be discarded so as to make the echo distortion canceller be a causal system. Comparing Eq. (4.17) to Eq. (4.16), we can know that the fourth term corresponds to the impulse response part located in the range $t < 0$. Thus, the discard of the fourth term is equivalent to that the impulse response is truncated to positive value of $t$. As a result, a truncation error will occur even though the number of taps is infinite. Due to this truncation error, it is conceivable that the performance of the feedback type echo distortion canceller will degrade.

Considering the simplest case, where there is only one echo, the tap coefficients are determined by Eq. (4.15). For a certain value of $i$, say $i_0(\geq 1)$, if the following equation holds,
\[ \tau = (i_0 - 1)T_d + T_{ini} \]  

(4.18)

that is, the excess time delay of the echo is just equal to the delay of a tap \( w_{i_0} \), by substituting Eq. (4.18) into Eq. (4.15), we can see that only \( w_{i_0} \) has non-zero value; all the other terms are equal to zero as shown in Fig. 4.4(a). That is, though the fourth term of Eq. (4.17) need to be discarded, there is no truncation error since the fourth term is equal to zero if Eq. (4.18) holds. At this time, the channel can be compensated perfectly. On the other hand, if

\[ \tau = (i_0 - 1)T_d + T_{ini} - T_d/2 \]  

(4.19)

at this time, since the absolute value of the numerator in Eq. (4.15) is equal to 1, which is the maximum value, the absolute value of all \( w_i \) except \( w_{i_0} \) will achieve their local maximum value as shown in Fig. 4.4(b). As a result, the discard of the fourth term in Eq. (4.17) will result in local maximum truncation error occurring. That is, if Eq. (4.19) holds, the echo distortion canceller has local maximum performance degradation. If the excess time delay of the echo varies in a range, zero and local maximum truncation error will take place by turn. Thus, a wavelike pattern of the truncation error will occur. Due to this wavelike pattern, the compensation performance of the feedback type echo distortion canceller will also show a wavelike curve with respect to the variation of the excess time delay of the echo.

For a special case, where the FM receiver does not move and the echo has fixed excess time delay, we can make Eq. (4.18) hold by adjusting \( T_{ini} \), i.e., by adjusting \( T_0 \). Then, we can perfectly compensate the channel. This is the reason why \( T_0 \) is introduced in Eqs. (4.5) and (4.7). In section 4.3.2, we will show the effect of \( T_{ini} \) on the echo distortion canceller performance with computer simulation.

For general case, where there is more than one echoes, the tap coefficients \( w_i \) are expressed by Eq. (4.14). It can be seen from Eq. (4.14) that \( w_i \) are determined by a sum of \( M \) terms. Each of these \( M \) terms corresponds to an echo. If there is a main echo that dominates the signal, we can neglect all the other echoes. Then, same to the one echo case, the echo distortion canceller will
(a) No truncation error position \( \tau = (i_0 - 1)T_d + T_{\text{ini}} \).

(b) Local maximum truncation error position \( \tau = (i_0 - 1)T_d + T_{\text{ini}} - T_d/2 \).

Fig. 4.4 No truncation error position and local maximum truncation error position.
show a wavelike performance curve with respect to the variation of the excess time delay of the main echo. For the case that all echoes have about the same effect on the signal, we cannot simply get the result, but, it is clear that, due to the truncation error, the performance of the echo distortion canceller will degrade.

At last of this section, we would like to point out that tap coefficients given by Eq. (4.14) are only determined by the feedback structure and not related to the channel transfer function estimation method and tap coefficients calculation method. Therefore, the wavelike performance curve cannot be avoided as long as the feedback structure is adopted. We will show the wavelike performance curves in sections 4.3.2 and 4.4.2.

4.2.4 Effect of Tap Interval

Equation (4.13) shows the relationship between the tap interval and \( B \), the bandwidth that needs to be compensated. In other words, for a given tap interval \( T_d \), the channel transfer function in the range \( B = 2\pi T_d \) needs to be known, since the calculation of Eq. (4.12) needs it. In this section, we consider the effect of tap interval on echo distortion canceller under the condition that the channel characteristics is known in the range of \( B' \) and \( T_d < 2\pi B' \). Here, we assume that the channel transfer function is necessary to calculate the tap coefficients [1]-[4].

Without loss of generality, assume only one echo and \( T_o = 0 \). Using Eq. (4.12) and considering that \( G_a(\omega) \) is limited in the range of \( B' < 2\pi T_d \), we have

\[
 w_i = \frac{2\pi}{B} \int_{-\frac{B}{2}}^{\frac{B}{2}} \frac{B'}{r} e^{-j(\omega \tau + \theta)} e^{j\omega r} d\omega \bigg|_{r = \frac{2\pi}{B}}
\]

\[
= \pi \frac{\sin \left[ \frac{B'}{2} (iT_d - \tau) \right]}{\pi (iT_d - \tau)/T_d}
\]

(4.20)
Comparing Eq. (4.20) with Eq. (4.15) for $T_0 = 0$, it can be seen that the numerators of them are different from each other. We intuitively conjecture that the difference between Eqs. (4.20) and (4.15) may result in the difference in compensation performance between case $T_d = 2\pi B'$ and case $T_d < 2\pi B'$. Note that if $T_d (= 2\pi B) > 2\pi B'$, where $B$ is the bandwidth needing to be compensated and $B'$ is the observable bandwidth, we should set $T_d$ to $2\pi B'$ so as to mitigate the effect of the truncation error on compensation performance. We will explain the reason in section 4.4.1. The effect of the tap interval on compensation performance of the echo distortion canceller will be examined with computer simulation in section 4.3.3.

4.3 Performance of Feedback Multipath Distortion Canceller

On the basis of the analysis in section 4.2, we will investigate the performance of the feedback echo distortion canceller with computer simulation in this section. Firstly, we examine the echo distortion canceller performance as a function of the excess delay of the echo. Then, the effect of tap interval on the performance is examined.

4.3.1 Performance Measure and Simulation System

The block diagram of computer simulation system is shown in Fig. 4.5, where the modulating signal is a 1 kHz sinusoidal signal; the maximum frequency deviation of the modulated FM signal is $\pm 75$kHz; a DSP-TAN type demodulator [3, 16, 17] is used for the demodulation of FM signal; two waves model (one echo and one direct wave) is assumed in the simulation.

The compensation performance of the echo distortion canceller is evaluated in terms of the distortion rate defined as

$$DR = \sqrt{\frac{E_2^2 + E_3^2 + \ldots + E_{15}^2}{E_1}}$$

(4.21)
Fig. 4.5 Block diagram of computer simulation system.

\[ E_i = \sum_{j=0}^{K-1} s(j\Delta t)\sin(i\omega_0 j\Delta t) \]  

(4.22)

where DR means the distortion rate; \( s(t) \) is the signal to be analyzed; \( K \) is set to 100; \( \Delta t \) is a time step set to 10 \( \mu s \); \( \omega_0 / 2\pi = 1 \) kHz (corresponding to the modulating signal). Here the 15-th harmonic frequency is 15 kHz. The higher frequency components are usually filtered out with a postdetection low-pass filter. Therefore, the harmonic components higher than the 15-th order are not taken into account.

In the examination of the echo distortion canceller performance, we focus on its compensation feature and the effects of its parameters on the performance. Therefore, noise is not taken into consideration and all filters related to noise are omitted for the simplicity of simulation. Moreover, without loss of generality, the complex zero IF expression of the modulated signal is used in our computer simulation.
4.3.2 Compensation Performance as a function of the excess delay

It was reported that the feedback echo distortion canceller compensation performance achieved when the excess time delay of the echo is small, less than 5 µs for example, is not so ideal [2], [3]. Here, we will analyze the reason why the compensation performance is not good for the echo with small excess time delay.

Without loss of generality, we consider the simplest case, where only one echo exists. For this case, the tap coefficients are given by Eq. (4.15). According to Eq. (4.15), the tap coefficients sequence is shown in Fig. 4.6. As explained in section 4.2.3, taps numbered 0, -1, -2, ... should be discarded so as to make the canceller be a causal system. Then, it can be seen from Fig. 4.6 that when the excess delay \(\tau\) is small, especially when \(\tau\) is less than the initial delay \(T_{ini}\), the truncation error is serious and cannot be neglected. The truncation error is perhaps the reason
why the compensation performance of the feedback echo distortion canceller for the echo with small excess delay is not good. We have also known that the truncation error has a wavelike pattern. Truncation error becomes zero when \( \tau = (i_0 - 1) T_d + T_{ini} \), while truncation error has the local maximum value when \( \tau = (i_0 - 1) T_d + T_{ini} - T_d / 2 \). Therefore, we can adjust the initial delay \( T_{ini} \) to locate the zero truncation error point to where we want. In the following, we investigate the compensation performance of the echo distortion canceller as a function of the excess delay of the echo.

The channel parameters \( r, \theta, \) and \( \tau \) are actually unknown. Therefore, Eq. (4.15) cannot be used for the calculation of tap coefficients. An approximate expression of Eq. (4.15) is [1], [2]

\[
\hat{w}_i = \frac{1}{N} \sum_{n=0}^{N-1} \left| 1 + G_i(f_n) \right|^2 e^{j 2 \pi f_n T_d} \quad i = 1, 2, \ldots, L
\]  

(4.23)

where \( f_n = f_0 + n B / 2 \pi N - B / 4 \pi \); \( f_0 \) is the center frequency; \( N \) is the number of points for DFT; and \( L \) is the number of taps, which is set to 10 [1]. As explained later, the bandwidth within which the channel transfer function (amplitude) can be measured from the received signal is 150 kHz (±75 kHz). Thus, \( B / 2 \pi = 150 \) kHz and the tap interval \( T_d \) is set to \( 1 / (150 \) kHz\) = 6.667 μs. Furthermore, considering the case that \( T_{ini} \neq T_d \), we modify Eq.(4.23) as

\[
\hat{w}_i = \frac{1}{N} \sum_{n=0}^{N-1} \left| 1 + G_i(f_n) \right|^2 e^{j 2 \pi f_n (i(i-1)T_d + T_{ini})} \quad i = 1, 2, \ldots, L
\]  

(4.24)

Using Eqs. (4.21), (4.22) and (4.24) in the simulation, we obtained the computer simulation results as shown in Fig. 4.7. This figure shows that the distortion rate curves have several extrema. Considering the description in section 4.2.3, such phenomenon can be easily understood, i.e., wavelike truncation error results in such a wavelike performance curves. Also we can see from Fig. 4.7 that, compared to the case \( T_{ini} = T_d \), the compensation performance of the echo distortion canceller for small excess delay, 1 ~ 5 μs for example, has been improved if the initial delay \( T_{ini} \) is set to \( T_d / 2 \). However, the improvement is very small except for the case that the excess delay
is equal to 3.3 μs (=Td/2). If an FM receiver does not move and the echo has the fixed excess delay, we can let τ = (i0-1)Td + T_{ini} by adjusting T_{ini} to achieve the best performance. That is, for a fixed value of excess delay, even if it is very small, the echo distortion can be cancelled very well. Unfortunately, this case is not general.

4.3.3 Effects of Tap Interval on Compensation performance

The channel transfer function, say 1+G_f(ω) or its amplitude 1|1+G_f(ω)|, is measured from the received signal for the equalizer used in an FM broadcasting receiver. Reference [4] introduced a method to measure the channel transfer function, where abscissa value, say ω_f, is obtained by measuring the instantaneous frequency of received FM signal and the value of the channel transfer
function, say \( |1 + G(\omega)| \), is measured by means of an envelope detector. Let us assume that this method is used to measure the channel transfer function. Then the observable bandwidth of the channel transfer function, i.e., \( B' \), is determined by the maximum frequency deviation of the received FM signal. Considering the FM signal standard shown in table 4.1, we can see that the observable bandwidth \( B' \) is equal to 150 kHz (± 75 kHz). In section 4.2.4, it has been pointed out that tap interval \( T_d \) perhaps affects the compensation performance. Here, we examine the effect of the tap interval.

Table 4.1 FM broadcasting standard (Japan)

<table>
<thead>
<tr>
<th>Carrier frequency</th>
<th>76-90MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum frequency deviation</td>
<td>± 75kHz</td>
</tr>
<tr>
<td>Channel bandwidth</td>
<td>200kHz</td>
</tr>
</tbody>
</table>

In the simulation below, we assume that the channel characteristics in the range of ± 75 kHz is known for simplicity of simulation, since the estimation of the channel transfer function does not belong to the scope of this chapter. Initial delay \( T_{in} \) is set to \( T_d \). The tap coefficients are calculated with Eq. (4.23). The tap interval \( T_d \) is set to \( 1/(150 \text{ kHz}) = 6.667 \mu s \) and \( 1/(200 \text{ kHz}) = 5 \mu s \), respectively, to examine the effects of tap interval on the compensation performance. The computer simulation results are shown in Fig. 4.8. It can be seen that the echo distortion canceller shows different performance for different tap intervals. A obvious change in compensation performance for case \( T_d = 5 \mu s \) is that there is no longer special excess delay values at which the echo distortion can be perfectly cancelled. Some fundamental knowledge may explain the reason [13].
Fig. 4.8 Compensation performance for different tap intervals.
\[ \theta = \pi/6, \ r = 0.5, \ N=200, \ L=10. \]
4.4 Adaptive Feedback Type Echo Distortion Canceller

We have known that the echo distortion canceller has a wavelike performance curve. In particular, for the echo with small excess delay, the truncation error will seriously degrade the performance of the echo distortion canceller. In this section, we propose an adaptive feedback type echo distortion canceller to improve the compensation performance.

4.4.1 Method for Improving Compensation Performance

The point to improve the echo distortion canceller performance is to reduce the truncation error. For our case, we have two parameters that can be adjusted. One is the initial delay $T_{ini}$. The other one is the tap interval $T_d$. In section 4.3.2, we have shown that changing $T_{ini}$ is nothing else than shifting the performance curve as shown in Fig. 4.7. We cannot expect to improve the echo distortion canceller performance a lot by adjusting $T_{ini}$. Then the parameter left to be adjusted to improve the compensation performance is only the tap interval $T_d$. Fortunately, adjusting the tap interval $T_d$ can, potentially, reduce the truncation error, which gives us a hope to improve the compensation performance of the feedback echo distortion canceller.

According to Eq. (4.14), Fig. 4.9 shows the tap coefficients sequences for different values of the tap interval $T_d$ under the condition that the excess delay of the echo is the same. This figure in fact explains why adjusting $T_d$ can reduce the truncation error: as $T_d$ decreases, the corresponding sinc function becomes narrow; compared to the large $T_d$ case, both the number and amplitude of the discarded taps become small, which results in the reduction in the truncation error. This is the reason why we prefer a tap interval with small value. For example, as mentioned in section 4.2.4, if $T_d (=2\pi/B) > 2\pi/B'$, we should decrease the value of $T_d$ to $2\pi/B$ in order to mitigate the impact of the truncation error. Theoretically, when $T_d \to 0$, there is only one non-zero tap for each echo and the truncation error will be zero. In short, decreasing the tap interval $T_d$ can reduce the truncation error and thereby improve the compensation performance of the
Fig. 4.9 Tap coefficients sequences for different values of tap interval $T_d$ under the condition that the excess delay $\tau$ of the echo is the same.
echo distortion canceller. However, we here ignored a very important fact, that is, if Eq. (4.12) or its approximate expression Eq. (4.23) is used for the calculation of the tap coefficients, the channel transfer function in the range $2\pi/T_d$ is necessary. Simply decreasing the value of tap interval cannot improve the compensation performance as discussed in section 4.3.3. Thus, we need other methods to calculate the tap coefficients. Under such situation, an adaptive feedback echo distortion canceller is proposed. The proposed adaptive echo distortion canceller controls its tap coefficients by using a dynamic adaptive algorithm, which does not need to measure the channel transfer function [14].

The FM broadcasting signal is a constant envelope signal. Due to the echo, fast fluctuation known as beat in the received signal envelope will occur. If we adjust the weights of an echo distortion canceller to compensate for the fast fluctuation, the impacts of the echo on the demodulated signal is expected to be cancelled. CMA (Constant Modulus Algorithm) [18] - [21] is just such an algorithm, which measures the fluctuation in signal envelope and adjusts the weights to minimize the fluctuation. Moreover, no reference signal being needed is also a merit of the CMA for its application in an FM receiver, where we are difficult to find a reference signal for an adaptive algorithm. Hence, we choose the CMA to control the tap coefficients of the echo distortion canceller.

The CMA is briefly introduced as follows. Using complex vector notation and referring to Fig. 4.3, the compensated signal at time $t_n$, $y(t_n)$, can be expressed as

$$y(t_n) = X^T(t_n)W(t_n) \quad (4.25)$$

where the superscript $^T$ denotes the transpose of a vector; and

$$X(t_n) = \begin{bmatrix} x(t_n) \\ y(t_n - T_m) \\ y(t_n - T_m - T_d) \\ \vdots \\ y(t_n - L T_m - (L-1)T_d) \end{bmatrix} \quad \text{and} \quad W(t_n) = \begin{bmatrix} 1 \\ -w_1 \\ -w_2 \\ \vdots \\ -w_L \end{bmatrix}_{1 \times L}$$
$L$ is the number of taps. The CMA measures and then minimizes the following objective function, $J^{pq}$,

$$J^{pq} = E \left( \left| y(t_n) \right|^p . |I|^{q} \right)$$  \hspace{1cm} (4.26)

where $E( \cdot )$ represents the ensemble average; $q = 2$ and $p = 1$, denoted as CMA$_1$, or $q = 2$ and $p = 2$, denoted as CMA$_2$. CMA$_1$ and CMA$_2$ show almost the same convergence performance and CMA$_1$ is stabler [22]. We choose CMA$_1$ for the algorithm to control the tap coefficients.

Substituting $q = 2$ and $p = 1$ into Eq. (4.26) and using the steepest descent method, we can obtain the weights update equation for CMA$_1$ as

$$W(t_{n+1}) = W(t_n) + \mu e(t_n) X^*(t_n)$$ \hspace{1cm} (4.27)

$$e(t_n) = y(t_n) |y(t_n)| y(t_n)$$ \hspace{1cm} (4.28)

where $\mu$ is a step size parameter; the superscript $^*$ denotes the complex conjugate. The adaptive feedback echo distortion canceller is shown in Fig. 4.10.

![Fig. 4.10 An adaptive feedback type echo distortion canceller.](image-url)
4.4.2 Performance of the Adaptive Feedback Echo Distortion Canceller

In this section, the performance of the adaptive echo distortion canceller for one echo case is examined. We set the tap interval $T_d$ to 3 $\mu$s, set the initial delay $T_{ini}$ to $T_d$, and set the number of taps $L$ to 6 to cope with the excess delay range of 0 - 15 $\mu$s. For $T_d = 3$ $\mu$s, according to the description in sections 4.2.4 and 4.3.3, the channel transfer function in the bandwidth of $1/T_d \equiv 333$ kHz needs to be known. Otherwise, the performance cannot be improved. It is impossible to measure the channel transfer function in this range from the received FM broadcasting signal. Therefore, controlling the tap coefficients with an adaptive algorithm that does not need to measure the channel transfer function is necessary. Initial tap coefficients are all equal to zero and the step size $\mu$ is 0.01. The computer simulation results are show in Fig. 4.11.

![Fig. 4.11 The compensation performance of the adaptive feedback echo distortion canceller. $\theta = \pi/6$, $r=0.5$. $T_d = 3$ $\mu$s, $T_{ini} = T_d$, $L=6$.](image-url)
with Fig. 4.7, we can see that a compensation performance improvement of about 20 dB is achieved by the proposed adaptive feedback echo distortion canceller (solid line). Though the tap coefficients are now controlled by CMA, the echo distortion canceller is also a feedback one. The adaptive echo distortion canceller does not show a wavelike performance curve (solid line). This is due to the convergence performance of the tap control algorithm. If the algorithm can perfectly converge, the wavelike performance curve should appear. However, we found that when applied to a feedback type equalizer, CMA hardly achieves perfect convergence state. Therefore, we simply force the taps with small value to be zero. Thus, taps with small value will be neglected. The result is shown in Fig. 4.11 (dashed line). The performance shows again a wavelike curve.

In the adaptive echo distortion canceller case, the convergence performance of the tap control algorithm becomes an issue. More number of taps can cope with a wider excess delay range, a smaller tap interval can compensate for the echo with smaller excess delay, but the convergence performance will be degraded. About the examination of the convergence performance of CMA, please refer to [22]. Moreover, as mentioned above, when applied to a feedback type equalizer, CMA hardly achieves perfect convergence state. More study is needed to investigate the convergence performance of the adaptive algorithm for a feedback type equalizer.

4.5 Summary

The effects of tap interval on the performance of a feedback type automatic echo distortion canceller for use in an FM broadcasting receiver and the compensation feature of such an echo distortion canceller have been theoretically analyzed and examined with computer simulation. In order to improve the compensation performance, an adaptive feedback echo distortion canceller is proposed.

For the fixed observable bandwidth of the channel transfer function, Simply decreasing
the value of tap interval $T_d$ cannot improve the compensation performance of the echo distortion canceller. The feedback type echo distortion canceller shows a wavelike compensation performance curve with respect to the excess time delay of the echo. Computer simulation shows that about 20 dB performance improvement has been achieved by the proposed adaptive feedback echo distortion canceller.
References


Chapter 5

Adaptive Switching Echo Cancellation/Diversity Reception for an FM Broadcasting Receiver in Multipath Mobile Channel

5.1 Introduction

For an FM broadcasting receiver, the quality of received signal is important, since such a receiver is often used to receive a music signal. When an FM broadcasting signal is transmitted through a multipath mobile channel, the received signal suffers from the waveform distortion caused by the echo and the fluctuation in the received signal level known as flat fading. Both the waveform distortion and the level fluctuation result in the degradation of received signal quality. The compensation of the channel is necessary so as to optimize the received signal. So far, the compensation for the waveform distortion and for flat fading are separately considered. For example, to compensate for the waveform distortion, one often uses an equalizer, which generates an inverse channel transfer function to compensate the received signal [1]-[6], or an echo canceller realized with an adaptive array, which cancels the echo by directing nulls of its directivity pattern to the echo [7] - [9], while to compensate for flat fading, diversity reception using multiple antennas is an effective method [10]. However, in a practical multipath mobile channel, the
received signal suffers from both the waveform distortion and the fluctuation in the received field intensity at the same time; moreover, the waveform distortion and the fluctuation take place in an irregular manner since the radio propagation environment differs greatly depending on the instantaneous position of the receiver that is assumed to be mounted in a moving vehicular. Under such situation, the compensation methods considered separately for wave distortion or flat fading cannot be expected to optimize the quality of the received signal. How to compensate the received signal for both the waveform distortion and flat fading leaves over as a problem.

To optimize the received signal under the situation mentioned above, we should take the fact into consideration that the waveform distortion and flat fading exist simultaneously when an FM broadcasting signal is transmitted through a practical multipath mobile channel. That means not only investigating the performance of an adaptive echo canceller or a diversity receiver under such situation, but more importantly, compensating the signal with an appropriate method according to the instantaneous channel conditions. In this chapter, we propose a compensation method termed adaptive switching echo cancellation/diversity reception [11] - [13]. The proposed adaptive switching echo cancellation/diversity reception monitors the impacts of channel conditions on received signal and then selects either an adaptive echo canceller or a diversity receiver accordingly to optimize the received signal for the instantaneous channel conditions.

5.2 Switching Reception Schemes

The proposed adaptive switching echo cancellation/diversity reception is applied for an FM broadcasting receiver. In an analog system, either the amplitude, the phase or the frequency of a sinusoidal carrier can be continuously varied in accordance with the voice or the message. Such an application environment will make us face some problems that we have not in a compensation circuit for a digital transmission system. In an FM receiving system, no special signal is available as a reference signal. Moreover, due to signal transmission in analog waveform, in contrast to digital transmission systems, we cannot generate the replica of the transmitted
signal so as to be used as a reference signal. In contrast to digital signal, due to the continuous
variation of an analog signal, we cannot use the decision error, which is the difference between
the sampled value of the received signal and the decision value at the sampling timing, to monitor
the impacts of the channel conditions and/or control the compensation circuit. Therefore, finding
some methods else to monitor the impacts of channel conditions and thereby control the
compensating system is the key issue of the proposed adaptive switching echo cancellation/
diversity reception method. In this section, two schemes of the adaptive switching reception are
presented. One is termed BF switching and the other one is termed pilot switching.

5.2.1 BF Switching Scheme

In the case that no echo signal exists, an FM signal has a constant envelope, while if there
are echo signals, the FM signal shows a beating phenomenon, that is, the envelope of the signal
interfered by the echo fluctuates at a frequency of signal bandwidth. At the same time, the signal
suffers from flat fading if it is transmitted through a multipath mobile channel. That is, the beat
caused by echo and the slow fluctuation caused by flat fading simultaneously exist in the envelope
of the received signal show in Fig. 5.1. The beat is directly related to the waveform distortion in
the demodulated signal. Low beat level implies low waveform distortion. Moreover, the impacts
of thermal noise, the waveform distortion, and the random FM effect become significant when
the signal level drops. A relative high signal level is therefore expected. Therefore, we propose
to use beat level to short-term average envelope fading ratio, called BF ratio, to monitor the
impacts of channel conditions. The definition of the BF ratio is illustrated in Fig. 5.2.

The block diagram of the BF switching scheme and its signal processing procedure are
shown in Fig. 5.3 and Fig. 5.4, respectively. The calculation of the BF ratio is executed for each
of the signal compensated by an adaptive echo canceller, denoted as BF_c, and that compensated
by a diversity receiver, denoted as BF_d. The value of BF_c and BF_d reflects the impacts of channel
conditions on received signal. Comparing BF_c and BF_d with thresholds, we then select one from
Fig. 5.1 An example of the received signal envelope. There are simultaneously the beat caused by echo and the slow fluctuation caused by flat fading in the envelope.

Fig. 5.2 An illustration of the BF ratio.
an echo canceller and a diversity receiver accordingly to compensate the received signal. Three thresholds are introduced. This is because the BF ratio of the signal compensated with an adaptive echo canceller is usually less than that compensated with a diversity receiver. In order to make the selection properly, it is necessary to introduce some thresholds. The values of the thresholds should be set according to experiments. This scheme can be used for any signal that has constant envelope.

In an FM broadcasting, a limiter is often used. For this case, the BF ratio should be measured prior to the limiter in order to avoid the effect of limiter on the signal envelope. If the dynamic

![Diagram of the BF switching scheme.]

Fig. 5.3 The block diagram of the BF switching scheme. DUR means the desired to undesired signal power ratio.
Signal compensated by a diversity receiver (IF signal)  
Signal compensated by an echo canceller (IF signal)  

bf ratio calculation for each of the input signals  

If $BF_d < Th_1$  
Yes  
Switch to the output of diversity receiver  

If $BF_c > Th_2$  
Yes  
Switch to the output of echo canceller  

If $BF_d < Th_3$  
No  
Switch to the output of diversity receiver  

Th = threshold  
$BF_d = BF$ ratio of diversity  
$BF_c = BF$ ratio of echo cancellation  

Fig. 5.4 The flow chart of the BF switching scheme.
range of the received signal is too large to measure the BF ratio prior to the limiter, we recommend the pilot switching scheme introduced in the following.

### 5.2.2 Pilot Switching Scheme

Among various stereo broadcasting signal schemes, the AM-FM stereo signal scheme is widely adopted by a lot of countries including Japan. Figure 5.5 shows the frequency spectrum of the modulating signal of the AM-FM scheme [14], where the Y-axis is the frequency deviation relative to the carrier. In this AM-FM stereo signal scheme, there is a pilot signal, a sinusoid signal, located at the frequency of 19 kHz. This pilot signal is also interfered by multipath interference when the FM signal is transmitted through a multipath mobile channel. Therefore, we can monitor the impacts of the channel conditions on the received signal by observing this pilot signal. The pilot signal can be picked up by using a band pass filter. The interference and noise power included in the pilot signal is then measured so as to monitor the impacts of the channel conditions. Since the interference and noise power can be measured in frequency domain, we will monitor the channel condition by observing the spectrum of the pilot signal. We then select one, which can compensate for the channel better, from an echo canceller and a diversity receiver to optimize the signal.

The block diagram of the pilot switching scheme and its signal processing procedure are shown in Fig. 5.6 and Fig. 5.7, respectively. The interference and noise power measure is executed for each of the signal compensated by an adaptive echo canceller, denoted as $p_{ic}$, and that compensated by a diversity receiver, denoted as $p_{rd}$. Comparing $p_{ic}$ and $p_{rd}$, we then select one, whose output has the smallest interference and noise power, from the echo canceller and the diversity receiver to compensate the received signal. No threshold needs to be introduced in the pilot switching scheme.

This scheme will not affected by the use of a limiter, since it selects an echo canceller or a diversity receiver by monitoring the demodulated signal. Moreover, in a practical FM receiver,
there is a local pilot signal. If we use the local pilot signal, noise and interference power also can be measured in time domain. Thus, FFT in Fig. 5.6 is not necessary, which may reduce the magnitude of signal processing.

Fig. 5.5 The spectrum of the modulating signal of the AM-FM scheme (without subsidiary communications authorization). For monaural broadcasting, the L+R channel is 100% modulation (maximum 75 kHz frequency deviation relative to the carrier); for stereo broadcasting, the sum of L+R channel, L-R channel, and pilot signal has 75 kHz maximum frequency deviation. Among them, the pilot signal has 7.5 kHz frequency deviation (10% modulation).
Fig. 5.6 The block diagram of the pilot switching scheme.
Signal compensated by a diversity receiver (demodulated signal)

Signal compensated by an echo canceller (demodulated signal)

Pick up the pilot signal

Fourier transform

Measure interference and noise power

If $p_{ic} > p_{id}$

Yes

Switch to the output of diversity receiver

No

Switch to the output of echo canceller

$p_{ic} =$ Interference and noise power of echo cancellation

$p_{id} =$ Interference and noise power of diversity

Fig. 5.7 The flow chart of the pilot switching scheme.
5.2.3 Selection of Echo Cancellation and Diversity Algorithms

This work is not especially concerned with the echo cancellation and diversity algorithms. Any available algorithm can be used in the proposed adaptive switching echo cancellation/diversity reception. In view of the application in an FM broadcasting receiver, we simply explain the selection of these algorithms for their application in our case.

*Echo cancellation algorithm:* In this chapter, the so-called echo canceller means an adaptive array antenna which cancels echoes by adjusting its weights so as to direct nulls of its directivity pattern to echoes. For an FM broadcasting receiver, time delay in received signal will not result in any problem, while signal with high received signal power is expected since high received signal power usually implies high SNR (signal to noise power ratio). Therefore, the desired signal is such a one which has the highest received signal power among multiple incident waves and the others are treated as echoes (interference) that need to be cancelled. Moreover, as described above, it is impossible for an FM receiver to find a reference signal. Thus, the echo cancellation algorithm should need no reference signal and can automatically select the signal with the highest power intensity. The constant modulus algorithm (CMA) [15]-[17] is obviously a good choice, which needs no reference signal and can automatically select the signal with the highest received signal power from multiple incident waves [18]. The weights update equation of the CMA has been briefly introduced in section 4.4.1. In this chapter, we also choose CMA$_1$ for the control of the array weights since it is stabler than CMA$_2$ [25].

*Diversity algorithm:* The term of diversity algorithm in this chapter means the signal combining methods: selection, equal gain and maximal ratio combing. Of these, the last two need cophasing between signals that need to be combined for predetection combining. To avoid carrying out cophasing, postdetection combining is usually employed. On the other hand, for an FM system (not only for an FM broadcasting receiver), the performance of postdetection maximal ratio combining will degrade due to the FM threshold phenomenon, that is, postdetection combining cannot enjoy the benefit of events in which each individual branch SNR is below
threshold but the sum of the SNR’s is above. Reference [10] points out that ideal postdetection maximal ratio combining has only 0.5 dB advantage over selection combining for dual diversity. Note that predetection and postdetection selection combining have the same performance since the same choice of antenna branches is made in either predetection or postdetection selection combining. The performance of equal gain combining is intermediate between selection and maximal ratio combining’s. Therefore, there is no thing to which special attention is worth being paid for choosing a postdetection diversity algorithm. We select postdetection selection combing as the diversity algorithm.

5.3 Propagation Models

It is usually assumed for performance investigation of an echo canceller with an adaptive array that antenna branches are fully correlated with each other, while for a diversity receiver, uncorrelated branches are assumed but echoes are not considered. These two propagation models cannot directly apply for the investigation of the proposed adaptive switching echo cancellation/diversity reception, since in our case, flat fading, echoes, diversity reception and echo cancellation should be taken into consideration simultaneously. Therefore, we need to find common propagation models for performance investigation of both echo cancellation and diversity reception, and finally for the performance investigation of the adaptive switching echo cancellation/diversity reception. The common propagation models are based on such fact that practical propagation environment will not change by interchanging an adaptive array and a diversity antenna.

Propagation model 1: It is assumed that the FM receiver is mounted in a moving vehicle that is much lower than surroundings; there is no LOS (Line Of Sight) to the broadcasting station and the reflection points of echoes; each of the receiver antenna branches receives a sum of a lot of reflected and scattered waves of the direct wave and echoes as shown in Fig. 5.8. Under such assumptions, Rayleigh flat fading can apply to both the direct wave and echoes [19].
Fig. 5.8 Propagation model 1.
We further assume that the received signal envelopes are spatially uncorrelated between antenna branches, which can be achieved by properly spacing the antennas, for example $0.38\lambda$, $0.88\lambda$ ... apart ($\lambda$ is the wavelength) [20]. In practice, uncorrelated branches can be obtained by spacing antennas about a half wavelength apart [19]. The number of echoes is assumed to be one.

**Propagation model 2:** As a reference, we also take another propagation model into consideration, where only one plane wave, whose envelope and phase have a Rayleigh and uniform distribution respectively, of each of the direct wave and echoes arrives at the receiver array antenna with an angle of incidence as shown in Fig. 5.9. Antenna branches are fully correlated with each other for each of the direct wave and echoes. The number of echoes is also assumed to be one for computer simulation. The incidence angle of the direct wave and the echo is assumed to be $0^\circ$ and $30^\circ$, respectively.

![Fig. 5.9 Propagation model 2.](image)
5.4 Performance of the Adaptive Switching Echo Cancellation/Diversity Reception

The aim of the adaptive switching echo cancellation/diversity reception is to compensate the channel for both flat fading and the waveform distortion. In this section, we first consider the measures of the compensation performance. Then, we investigate the performance of the adaptive switching reception for the waveform distortion and flat fading under the two propagation models mentioned above, respectively.

5.4.1 Performance Measures

The most often used performance measure for analog signal is the signal to noise power ratio (SNR). In our case, besides noise, waveform distortion should be also taken into consideration in performance measures. We use the desired to undesired signal power ratio (DUR) as a performance measure. DUR in decibel is defined as

\[
DUR = 10 \log_{10} \frac{\sum_{i=1}^{N} d^2(t_i)}{\sum_{i=1}^{N} e^2(t_i)}
\]

where \(d(t_i)\) is the desired signal and \(e(t_i)\) is the difference between the signal to be analyzed and the desired signal. Equation (5.1) is actually an average DUR in the duration of \(N\Delta t\), where \(\Delta t = t_i - t_{i-1}\). In this chapter, \(N = 512\) and \(\Delta t = 3.2895\) μs. Since the word "average" will be used for another average calculation later, we will not call Eq. (5.1) average DUR to avoid confusing. If dividing the numerator and denominator of Eq. (5.1) by \(N\), respectively, we can see that the
numerator is (omit the time variable)

\[
\frac{1}{N} \sum d^2 = \text{Desired signal power} \tag{5.2}
\]

Due to the same reason mentioned above, we do not call Eq. (5.2) average desired signal power though it is actually an average power. Denote the signal that needs to be analyzed, say \( y \), as

\[
y = d + I + n \tag{5.3}
\]

where \( I \) and \( n \) represent the interference (waveform distortion) and noise in the signal, respectively. The denominator is then

\[
\frac{1}{N} \sum e^2 = \frac{1}{N} \sum (y - d)^2 = \frac{1}{N} \sum (I + n)^2 = \frac{1}{N} \sum (t^2 + n^2) + \frac{2}{N} \sum nI \tag{5.4}
\]

The second term of Eq. (5.4) can be thought as the correlation between the noise \( n \) and the interference \( I \). Since \( n \) is not correlated with \( I \), we have reason to believe that the second term is equal to zero. Therefore, Eq. (5.4) is the interference and noise power. Thus, the DUR defined by Eq. (5.1) is equivalent to the SINR (Signal to Interference and Noise power Ratio). Due to the presence of flat fading, DUR of the compensated signal is not a fixed value, which varies with time as shown in Fig. 5.10. Therefore, we use an average DUR to measure the performance of the adaptive switching echo cancellation/diversity reception. In the performance investigation below, we will mean the DUR of the demodulated signal if not stated when the word "DUR" is used.

Similar to the burst error caused by flat fading in the digital communication systems, the signal quality in our case drops within a short duration repeatedly as shown in Fig. 5.10. For such a signal, the average DUR is not a good measure, since it can hardly distinguish a signal whose quality does not drop from that whose quality drops within a short duration repeatedly.
Fig. 5.10 The DUR of the demodulated signal versus time curve.

For this reason, the compensation performance is also measured in terms of the probability that DUR is below a given threshold. Referring to reference [21] and considering the situation of our case, we select 20 dB as the threshold. The probability of DUR below the threshold of 20 dB is denoted as $P_{b20}$. One hand, we expect that the probability measure, $P_{b20}$, should be sensitive for the signal in our case. The other hand, $P_{b20}$ will reflect the performance of a compensation method to prevent signal quality from drop.

5.4.2 Compensation Performance

A series of computer simulations are carried out to investigate the performance of the proposed adaptive switching echo cancellation/diversity reception. The simulation parameters are shown in Table 5.1.
Table 5.1 Simulation parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Details</th>
</tr>
</thead>
<tbody>
<tr>
<td>Antenna</td>
<td>2 isotropic antennas</td>
</tr>
<tr>
<td>Antenna spacing</td>
<td>$0.5\lambda$</td>
</tr>
<tr>
<td>Initial weights of the echo canceller</td>
<td>$(1, 0)$</td>
</tr>
<tr>
<td>Modulating signal</td>
<td>AM-FM stereo signal (See Fig. 5.11)</td>
</tr>
<tr>
<td>Maximum frequency deviation of modulated signal</td>
<td>$\pm 75 \text{ kHz}$</td>
</tr>
<tr>
<td>FM demodulator</td>
<td>DSP-TAN type demodulator [22]-[24]</td>
</tr>
<tr>
<td>Amplitude of the echo</td>
<td>-3 dB relative to the direct wave</td>
</tr>
<tr>
<td>Additional phase shift of the echo</td>
<td>$\pi/6$</td>
</tr>
<tr>
<td>Weights update frequency of echo canceller</td>
<td>38 kHz</td>
</tr>
<tr>
<td>Doppler frequency</td>
<td>11 Hz</td>
</tr>
<tr>
<td>Bandwidth of postdetection low-pass filter</td>
<td>17 kHz</td>
</tr>
<tr>
<td>Bandwidth of BPF (to pick up the pilot signal)</td>
<td>4 kHz</td>
</tr>
<tr>
<td>Bandwidth of equivalent low-pass receiving filter</td>
<td>100 kHz</td>
</tr>
</tbody>
</table>
The weights update frequency of the echo canceller is set to 38 kHz. This is because a practical CMA adaptive array was ever reported in [26], whose weights update frequency is at this level. Moreover, the weights update frequency should be high enough compared to fading speed. Since the maximum Doppler frequency is assumed to be 11 Hz in the simulation for 100 MHz frequency and 120 km/h vehicular velocity, we think that the weights update frequency of 38 kHz is appropriate. For the demodulation of FM signal, a DSP-TAN type demodulator is used. A merit of this demodulator is that it does not need an AGC or a limiter to restrain the fluctuation in signal amplitude [23]. The bandwidth of equivalent low-pass receiving filter is 100 kHz, whose corresponding bandpass bandwidth is 200 kHz. All BPF and LPF are ideal for simplicity. Only the L+R signal (see also Fig. 5.11) is used to evaluate the performance of the adaptive switching reception. Other signals (pilot and L-R signals) are filtered out by a postdetection low-pass filter. Thresholds 1, 2 and 3, which are defined in Fig. 5.4, are -14.9 dB (0.18), -18.4 dB (0.12) and -13.2 dB (0.22), respectively. Without loss of generality, complex zero IF expression of the signal is used in the computer simulation.

5.4.2.1 Effects of Step Size $\mu$

The effects of step size $\mu$ of the CMA on compensation performance for propagation model 1 is shown in Fig. 5.12. The vertical axis is the probability that DUR is below 20 dB, i.e.,
Fig. 5.12 Effects of step size $\mu$ on compensation performance. Input SNR = 30 dB.

Propagation model 1.
Pb2o. Step size $\mu$ is set to 0.2, 0.1, 0.05, and 0.01, respectively. From the figure, it can be seen that the adaptive switching reception is effective for all values of step size $\mu$. When the excess time delay of the echo is equal to 0 $\mu$s, CMA shows step size dependence performance as shown in Fig. 5.12(a), while the adaptive switching reception including both pilot switching and BF switching is not sensitive to the step size $\mu$. This result can be interpreted as that the compensation performance of the adaptive switching reception is mainly related to that of the diversity reception when the excess delay is equal to zero. Only when the diversity fails to compensate the channel, CMA is then adaptively selected. When excess delay is equal to 10 $\mu$s, CMA, pilot switching, and BF switching all show step size dependence performance. This is because the compensation performance is mainly determined by that of CMA at this time. In the following, the step size $\mu$ is set to 0.1 at which CMA shows a good performance as shown in Fig. 5.12. Note that though CMA performance is affected by its step size and also related to its weights update frequency, finding the best parameters for CMA is not necessary for the investigation of the performance of the adaptive switching reception. Our aim is to compensate the signal for the instantaneous channel conditions instead of optimizing CMA. If CMA performance is improved by adjusting its parameters, the performance of the adaptive switching reception should be improved accordingly, since the switching reception selects one, which can compensate the signal better, from CMA and diversity.

### 5.4.2.2 Compensation Performance for Fading

The excess delay of the echo is set to 0 $\mu$s to examine the compensation performance for flat fading. For propagation model 1, computer simulation results are shown in Fig. 5.13. It can be seen from Fig. 5.13(a) that the compensation performance of both the pilot and BF switching schemes is not inferior to that of the diversity reception, while CMA is even worse than uncompensated case when input SNR > 30 dB. Compared to CMA, the average DUR is increased about 4 dB by using the switching receptions when input SNR > 30 dB, whereas the diversity reception is about 3 dB higher than uncompensated case. If the performance is measured in
(a) Performance measured in terms of average DUR.

(b) Performance measured in terms of the probability of DUR below 20 dB.

Fig. 5.13 Compensation performance for flat fading. Excess delay = 0 μs. Propagation model 1.
Fig. 5.14 $P_{b_{20}}$ versus SNR curves for propagation model 2. Excess delay = 0 µs.

Fig. 5.15 CMA performance versus Doppler frequency. Propagation model 2. Excess delay = 0 µs. 30, 35, and 40 dB in the figure are the value of input SNR.
terms of the probability that DUR is below 20 dB, $P_{b20}$, as shown in Fig. 5.13(b), the pilot switching scheme shows the best performance and the BF switching is better than CMA. At $P_{b20} = 10^{-4}$, the pilot switching has about 3.5 dB advantage over CMA.

For propagation model 2, the $P_{b20}$ versus SNR curves are shown in Fig. 5.14. Both the BF and pilot schemes of the switching reception show again better performance than that of CMA. In particular, the performance of the pilot scheme is even better than that of diversity. On the other hand, we can see that for different propagation models, CMA shows different performance as shown in Fig. 5.13(b) and Fig. 5.14. A field experiment also shows the similar phenomenon [27]. We first doubt that the different propagation models result in the difference in the change speed of the received signal envelope and thereby CMA shows different performance for different change speeds due to its convergence performance. Therefore, we examine the CMA performance with respect to Doppler frequency for propagation model 2 as shown in Fig. 5.15. It can be seen that the difference in the envelope change speed really results in the change in CMA performance when Doppler frequency is greater than 10 Hz. However, in the range of 1-10 Hz, CMA almost does not show the difference in its performance with the change in Doppler frequency. There may be some reasons else that also affect the performance of CMA. An explanation about this point is presented in [28] as described in Appendix 5.1.

5.4.2.3 Compensation Performance for Waveform Distortion

The excess delay of the echo is set to 10 μs to investigate the compensation performance of the adaptive switching receptions for waveform distortion in the presence of flat fading. The computer simulation results for propagation model 1 are shown in Fig. 5.16. It can be seen that the performance of both the pilot and BF switching schemes is as good as that of CMA. In particular, the $P_{b20}$ curve of pilot switching scheme shows the best performance though the average DUR of it is almost the same with that of CMA. Such results mean that the pilot switching scheme has better performance to prevent the signal quality from drop than the others. Also, we can see from Fig. 5.16(b) that $P_{b20}$ curves show floors when input SNR>30 dB. This phenomenon
Fig. 5.16 Compensation performance for waveform distortion. Excess delay = 10 μs.

Propagation model 1.
is due to that CMA convergence speed is not fast enough to track the change in received signal envelope caused by flat fading and thereby the signal cannot be completely compensated for waveform distortion. Thus, the performance improvement achieved by pilot switching is important, since the $P_{b20}$ value cannot be decreased by increasing input SNR. Figure 5.17 shows CMA performance measured in terms of $P_{b20}$ versus Doppler frequency curves. It can be seen that CMA can compensate the signal better when Doppler frequency is small. Please note that the signal quality is now dominated by waveform distortion when noise is not so large, which is different from the case of excess delay $= 0 \mu$s where there is not waveform distortion. The range of $P_{b20}$ curves in Figs. 5.15 and 5.17 are different from each other.

The $P_{b20}$ versus SNR curves for propagation model 2 are shown in Fig. 5.18. It can be seen that the adaptive switching receptions, especially the pilot switching scheme, have obviously improved the compensation performance compared to CMA. We also can see by comparing Figs. 5.16(b) and 5.18 that the performance improvement achieved by switching reception for propagation model 2 is more obvious than that for propagation model 1. We think that the reason resulting in such a result is due to that the adaptive switching reception can more sensitively monitor the channel conditions for propagation model 2.

5.4.2.4 Compensation Performance as a Function of the Excess Delay

We now examine the compensation performance of the adaptive switching echo cancellation/diversity reception in the excess delay range of 0 - 15 $\mu$s. The excess delay of most echoes is in this range [29], [30]. The input SNR is set to 30 dB relative to the average power of the direct wave. Computer simulation results for propagation model 1 are shown in Fig. 5.19. It can be seen from Fig. 5.19(a) that the compensation performance of both the pilot and BF switching schemes for flat fading (excess delay is near zero) is as good as that of the diversity reception, while the compensation performance for waveform distortion (excess delay is large) is as good as that of CMA. That is, the adaptive switching reception has realized the advantages of both CMA array antenna and diversity reception. In particular, the compensation performance
Fig. 5.17 CMA performance measured in terms of the probability of DUR below 20 dB versus Doppler frequency. Excess delay = 10 μs. Propagation model 1. 30, 35, and 40 dB in the figure are the value of input SNR.

Fig. 5.18 $P_{b20}$ versus SNR curves for propagation model 2. Excess delay = 10 μs.
Fig. 5.19 Compensation performance as a function of the excess delay. Propagation model 1.

Input SNR = 30 dB.
Table 5.2 Performance comparison for different compensation methods. Propagation model 2.

Input SNR = 30 dB.

<table>
<thead>
<tr>
<th></th>
<th>Excess delay = 0 µs</th>
<th>Excess delay = 10 µs</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Average DUR</td>
<td>Probability of DUR below 20 dB</td>
</tr>
<tr>
<td>Diversity</td>
<td>52.01 dB</td>
<td>$5.07 \times 10^{-4}$</td>
</tr>
<tr>
<td>CMA</td>
<td>48.41 dB</td>
<td>$1.11 \times 10^{-3}$</td>
</tr>
<tr>
<td>Pilot switch</td>
<td>52.03 dB</td>
<td>$2.77 \times 10^{-4}$</td>
</tr>
<tr>
<td>bf switch</td>
<td>52.01 dB</td>
<td>$4.76 \times 10^{-4}$</td>
</tr>
</tbody>
</table>

of the pilot switching scheme measured in terms of $P_{b20}$ shows the best performance as shown in Fig. 5.19(b), which means again that the pilot switching scheme is most effective to prevent the signal quality from drop. For propagation model 2, the compensation performance comparison for excess delay = 0 and 10 µs is shown in Table 5.2. We can see that the adaptive switching echo cancellation/diversity reception is also effective for propagation model 2.

5.5 Summary

An adaptive switching echo cancellation/diversity reception with two schemes, pilot switching and BF switching, to compensate for both waveform distortion and flat fading in an FM broadcasting receiver has been proposed. The compensation performance of both pilot switching and BF switching schemes has been investigated with computer simulation. The compensation performance is measured in terms of both average DUR and the probability of DUR below a given value.

The computer simulation results show that the compensation performance for flat fading of both the pilot and BF switching schemes is as good as that of diversity reception, while the
performance for waveform distortion is as good as that of a CMA echo cancellation, that is, the adaptive switching echo cancellation/diversity reception has realized the advantages of both diversity reception and echo cancellation. In particular, the pilot switching shows better performance than diversity reception for flat fading and than echo cancellation for waveform distortion if performance is measured in terms of the probability that DUR is below 20 dB. This result means that the pilot switching scheme is more effective to prevent the signal quality from drop than the others.
Appendix 5.1 Impact of Noise on CMA Performance

Theoretical Analysis

The CMA array output \( y(n) \) at time \( n \) can be written as

\[
y(n) = X^T(n) W(n) \tag{A.5.1}
\]

where \( X^T(n) \) is the vector of array element output signals; \( W(n) \) is the vector of array weights.

CMA cost function is

\[
J_{pq} = E\left[\left|y(n)\right|^p - 1\right]^q \tag{A.5.2}
\]

Taking noise into account, we can write the CMA array output \( y \) as (time variable \( n \) is omitted in the following)

\[
y = s + n \tag{A.5.3}
\]

where \( s \) is the array output signal; \( n \) is the noise. For CMA \( (p=1, q=2) \) we have

\[
J_{12} = E\left[|s|^2 - 2\sqrt{|s|^2 + 2Re(sn) + |n|^2} + 1 + |n|^2\right] + 2Re\left[E(sn)\right] \tag{A.5.4}
\]

Assuming that \( s \) is uncorrelated with \( n \), the second term above is then equal to zero. If noise is much smaller than the signal, the noise terms in the square root operation can be neglected. Equation A.5.4 is then approximately expressed as

\[
J_{12} \approx E\left[\left(|s| - I\right)^2\right] + E\left(|n|^2\right) \tag{A.5.5}
\]

Equation A.5.5 achieves its minimum value when \( |s| \) approaches 1 and the noise is minimum, i.e., CMA array is equivalent to the maximal ratio diversity after it converges. However, if the signal is much smaller than noise, neglecting the signal terms in the square root operation in Eq.
A.5.4, we have

\[ J^{12} = E \left[ (|n| - I)^2 \right] + E(|s|^2) \]  \hspace{1cm} (A.5.6)

It can be seen that CMA can not optimize the signal. That is, when noise is large relative to the signal, the performance of CMA will become bad. In the following, we examine the CMA performance with respect to SNR (signal to noise power ratio) with computer simulation. The number of array elements is 2. Initial weights are (1,0).

**Static performance**

Branches are all a 100 kHz sinusoid signal with amplitude being equal to 1. A 200 kHz band limited noise is added to the signal. SNR values of all antenna branches are simultaneously changed. CMA performance is measured in terms of DUR of the undemodulated signal. Simulation result is shown in Fig. A.5.1. It can be seen that CMA shows a threshold phenomenon.

![Graph showing static performance of CMA and MRC](image)

**Fig. A.5.1 Static performance of CMA. MRC = maximal ratio diversity. Step size \( \mu = 0.1 \).**
For CMA$_2$, we can get the similar result, i.e., CMA$_2$ also shows a threshold phenomenon. In a fading channel, SNR of signal probably drops down to a small value. At this time, this threshold phenomenon will cause impacts on CMA performance.

**Dynamic performance**

Simulation parameters are the same with that in section 5.4.2. Performance is measured in terms of the probability of the DUR of demodulated signal below 20 dB, i.e., $P_{b<20}$. Simulation is carried out for propagation model 2. For predetection MRC (maximal ratio combing), cophasing is needed. To avoid cophasing, we can use postdetection MRC. In an FM system, postdetection MRC performance is near to selection diversity. Thus, we compare CMA with selection diversity. The results are shown in Fig. A.5.2. It can be seen that CMA performance is different from the selection diversity for $\mu=0.1$, where $\mu$ is the step size of CMA. As $\mu$ decreases, the convergence speed of CMA becomes slow so that it cannot track the variation of noise. As a result, the impact

![Graph showing dynamic performance](image)

Fig. A.5.2 Dynamic performance. Selection = selection diversity. Propagation model 2.
of noise is mitigated or even eliminated. For propagation model 1, the performance of CMA is almost the same with selection diversity. This is because uncorrelated antenna branches are hardly fall down to a low value simultaneously and thereby the threshold phenomenon will not shows its impact.
References


[22] O. Takizawa and S. Kondo, “Performance of an FM broadcasting receiver using DSP -


Chapter 6

Conclusions

In mobile multipath propagation environments, flat fading and frequency-selective fading will occur. Both of these fadings affect the transmitted signal. In particular, frequency-selective fading causes distortion in signal waveform. Such distortion results in the deterioration of error rate performance for digital transmission system and the degradation of signal quality for analog system. The distortion becomes serious as signal bandwidth or the difference in the time delay of echoes increases. Therefore, compensation for the multipath channel distortion by cancelling echoes or reducing their power becomes important in order to improve the system performance. Many researchers have been engaged in studying in this field and have dedicated a lot of achievements that have played significant role in mobile communication systems. In this thesis, a study on some compensation techniques for signal distortion in multipath radio channel is presented.

The characteristics of a multipath mobile channel and current compensation methods for flat and frequency-selective fadings are briefly introduced in Chapter 2. Flat fading and frequency-fading have the same occurrence mechanism. The distinction of them lies on whether the difference in the arrival times of incident waves can be neglected or not compared to the reciprocal of the signal bandwidth. Flat fading causes random fluctuation in signal envelope and phase, while frequency-selective fading results in distortion in signal waveform. Some compensation techniques, which are common to digital and analog transmission systems, for flat and frequency-
selective fading have been also briefly described in this chapter. The purpose of this chapter is to offer a preparatory knowledge or reference to the next chapters instead of to describe the current situation of compensation techniques for flat and frequency-selective fadings.

The effects of beam tilting technique on bit rate selection in multipath channel is theoretically analyzed in Chapter 3. Beam tilting can reduce the power of echoes by tilting down the transmitting wave beam at base station and thereby decrease intersymbol interference (waveform distortion). Therefore, when applied to a digital transmission system, it can increase the maximum achievable bit rate. The critical rate, which is the maximum attainable system bit rate, is evolved as a measure of the system performance. Two propagation models, namely, perfect reflection case and diffusely scattered reflection case are considered. The analysis in this chapter has verified that the beam tilting technique, which was originally proposed to decrease the interference between two adjacent frequency reuse cells, is valid to reduce the multipath distortion and thereby improve the system performance. The analysis results show that, with the introduction of beam tilting, the critical system bit rate is increased by several times for perfect reflection case; for diffusely scattered reflection case, the critical bit rate is related to the distance between the base station and the reflection point, and the beam tilting can also increase the critical bit rate a lot if the reflection point is not so close to the base station.

In Chapter 4, we study on the feedback type echo distortion canceller applied to an FM broadcasting receiver. First, the feature of the compensation performance is analyzed. The results show that the feedback type echo distortion canceller has a wavelike compensation performance curve with respect to the excess delay of the echo. The wavelike performance curve results from the feedback structure and is not related to the tap controlling methods. We cannot eliminate this wavelike performance curve by changing the tap controlling method and/or channel characteristics measuring method. The effect of tap interval on compensation is also examined. We find that simply decreasing the tap interval value cannot improve the compensation performance, if the observable bandwidth of the channel transfer function is kept fix and the tap coefficients are calculated from the channel transfer function. Then, in order to improve the compensation
performance, an adaptive feedback type echo distortion canceller has been proposed. The adaptive echo distortion canceller controls its taps by a dynamic adaptive algorithm. Since the adaptive echo distortion canceller does not need to measure the channel transfer function for controlling its taps, a small tap interval value can be used to improve the compensation performance. We have analyzed the reason why decreasing the tap interval can improve the compensation performance. Computer simulation shows that a compensation performance improvement of about 20 dB in terms of signal distortion rate has been achieved by the use of the adaptive feedback type echo distortion canceller.

In Chapter 5, an adaptive switching echo cancellation/diversity reception for an FM broadcasting receiver in multipath mobile channel has been proposed. In the multipath mobile channel, the received signal suffers from both the fluctuation in the received field intensity caused by flat fading and waveform distortion caused by echoes. The fluctuation and the waveform distortion take place in an irregular manner since the radio propagation environment differs greatly depending on the instantaneous position of the receiver that is assumed to be mounted in a moving vehicular. The proposed adaptive switching echo cancellation/diversity reception monitors the impacts of channel conditions on received signal. Then, either an echo canceller or a diversity receiver is selected accordingly to compensate the channel for the instantaneous channel conditions. Two schemes are proposed and examined with computer simulation. One is termed pilot switching scheme, which monitors the channel by observing the pilot signal included in FM broadcasting signal. The other one is termed BF switching scheme, which observes the beat level to short-term average envelope fading ratio (BF ratio). The compensation performance is measured in terms of both DUR (Desired to Undesired signal power Ratio) and the probability that DUR is below a given value. For the compensation for flat fading, the results show that both the pilot and BF schemes of the adaptive switching reception have the same performance as the diversity in terms of the average DUR. Compared to CMA, the average DUR is increased about 4 dB when input SNR is greater than 30 dB by using adaptive switching reception. If the performance is measured in terms of the probability of DUR below 20 dB, P_{b20}, the pilot scheme
has about 3.5 dB advantage over diversity. For waveform distortion, the performance of both pilot and BF schemes are as good as CMA in terms of the average DUR. In terms of $P_{b2o}$, the pilot scheme shows the best performance. Summarizing the computer simulation results, we can conclude that the adaptive switching echo cancellation/diversity reception has realized the advantages of both adaptive echo cancellation and diversity reception.

The proposed adaptive switching echo cancellation/diversity reception indeed shows improved compensation performance, but it is nothing else than selecting one, which can compensate the received signal better for the instantaneous channel conditions, from an adaptive echo canceller and a diversity receiver. Since both the adaptive echo canceller and the diversity receiver may not be operating at their optimal state for a specific channel condition, improving the compensation performance further for mobile multipath channel is possible. This issue is attracting interests of some researches. We are expecting the achievements could be contributed soon.
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