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**Efficient Narrow-Band Speech Transmission
Over
Land Mobile Channels**

by

C) Rafik A. Goubran

A thesis submitted to the
Faculty of Graduate Studies and Research
in partial fulfilment of the requirements
for the degree of
Doctor of Philosophy

Department of Systems and Computer Engineering
Faculty of Engineering
Carleton University

September, 1986

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ABSTRACT

This thesis addresses the problems of speech quality, and channel bandwidth efficiency, for speech transmission in mobile communications.

In the first part, a flexible, real-time, programmable, mobile radio channel simulator is proposed. The conventional basic configuration of the simulator was augmented with several blocks, allowing the simulation of multipath fading, Doppler shifts, line of sight propagation and shadowing, with a wide range of parameters. This was achieved through the use of variable rate and multirate digital signal processing, combined with an appropriate system design allowing an efficient task partition between software and hardware. The simulator was built and evaluated. Its output was compared to several field tests; a good agreement is demonstrated.

In the second part, a system is proposed for reducing the background acoustic noise from the transmitted speech signal in mobile telephony, and, therefore, improve the signal to noise ratio at the transmitter. Several field tests were conducted in order to characterize the background acoustic noise in cars. Accordingly, an adaptive noise reduction system is proposed, using a transversal finite impulse response adaptive filter and the stochastic gradient adaptation algorithm; a 10dB noise reduction is achieved.

In the third part, a novel concept, the switched tone (ST), is proposed for speech and data transmission over mobile channels. The speech is transmitted using a modified single sideband (SSB) technique with a channel separation of 5kHz, instead of the 20 - 30kHz currently used in analog frequency modulation (FM).

Three main subsystems, required by the ST concept, are proposed. A novel differential dynamic speech companding system improves the signal to noise ratio.

of the low volume syllables, and, therefore, improves the speech quality. An appropriate *Automatic Gain Control (AGC)* system compensates for the amplitude fluctuations attributed to multipath fading. The tracking range of the *AGC* is programmable and can take any value according to the signal to noise ratio encountered. Finally, the *Automatic Frequency Control (AFC)* problem is solved in a systemwise approach through the *ST* concept. The problem is reduced to two simpler ones, the initial acquisition, performed by a mid-band tone, and the phase tracking, performed by a tone above the speech band, allowing a full transparency to all speech and data formats.

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Chapter 1

Introduction

Introduction

Mobile communications have been known since 1897 [21]. The use of the mobile channel was limited to few applications such as taxi dispatch systems, and police communications [1, 21]. Other applications such as mobile telephony, and various data applications are now emerging [3, 5], which increases the demand for new channel allocation, and better transmission quality [1, 5]. The available spectrum on the other hand is limited, and, therefore, a spectrum congestion problem is now arising and is expected to grow further in the future [1, 5, 65].

Speech transmission over mobile channels has been dominated by one modulation technique, the analog frequency modulation (FM) [1-3, 7, 21]. In analog FM, the speech signal is transmitted using 20 to 30 kHz channel spacing [1, 3, 6, 7, 15, 21], whereas it is possible to transmit the same signal with a 5 kHz channel spacing using modified single sideband (SSB) techniques [7, 10, 11, 15]. There are however some problems associated with these techniques which limit their com-

mercial spread [8, 10-15]. These problems are discussed in section 3.3. In chapter (6) of this thesis, a novel concept, the switched tone, is proposed for narrowband speech and data transmission over mobile channels, using a 5 kHz channel spacing.

In mobile communications, one of the major sources of speech corruption is the background acoustic noise. It is, therefore, important to reduce this noise from the transmitted speech signal, and consequently, improve the signal to noise ratio at the transmitter. The general problem of noise cancelling is discussed in [51], and specific applications are presented in [51-58]. In this thesis, a noise reduction system, appropriate to mobile telephony, is proposed [63].

The testing and evaluation phases for any new concept in mobile communications is a difficult and cumbersome task. A moving vehicle loaded with equipment, as well as a base station set-up are needed. Even then, the comparison between different techniques is not easy, due to the impossibility of repeating the same fading patterns. A mobile channel simulator is, therefore, a necessary and an indispensable piece of equipment, in any mobile communication research laboratory. Many simulators are discussed in [29-35]. There are, however, some limitations to each of them. A flexible real-time simulator [71], overcoming these limitations, is proposed in chapter (4).

More background material, necessary to introduce the various research topics covered in this work, is presented in chapters (2) and (3). This chapter gives an overview of the problem addressed in this thesis as well as the proposed solution approach.

In section 1.1, the thesis objectives are presented. Subsections 1.1.1 and 1.1.2 include the motivation behind the work, the problem statement, and the solution approach. An outline of the thesis and its organization is given in subsection 1.1.3.

Section 1.2 includes a description of the testing set-up used throughout the thesis. It also includes a discussion of the reasons for using real-time simulation and multiprocessing.

1.1 Thesis Objectives

1.1.1 Motivation

The motivation behind this thesis is based on two factors, the spectrum congestion problem and the advances achieved in the field of VLSI technology.

The first factor is the radio spectrum congestion. In recent years there has been an increasing demand for radio channel allocation, as well as for a better quality. This demand is expected to grow more and more in the future due to the emerging new applications such as mobile and portable communication systems as well as various remote control applications [1,5]. On the other hand, the available usable spectrum is limited. So, the only alternative for the future would be a more efficient use of the available spectrum.

The second factor is related to the recent advances achieved in the field of VLSI technology. Modern high performance components have altered the cost-performance criteria for all digital systems. Consequently, new schemes in digital systems design and digital signal processing are now emerging; which opens new perspectives and offers feasible solutions to many of the traditional problems.

1.1.2 Problem Statement

The general aim of this work is to explore the capabilities of digital systems design and digital signal processing in the field of mobile communications. The main problem addressed is the investigation of narrowband high performance speech transmission over mobile channels. The price paid in this case would be more complexity at the system level.

In other words, the problem addressed is whether it is possible to achieve a better spectrum efficiency while maintaining a good performance at the expense of system complexity alone. The investigation is restricted to the mobile channel.

In the area of modulation techniques there is a conventional tradeoff between the bandwidth, the power, and the performance. It is generally accepted that for the same information rate, the narrower the bandwidth, the lower the performance. There is in fact another factor governing this relationship: the complexity, which affects the cost, reliability, and development time. The rules of complexity have been affected by the recent advances in VLSI technology. The main factor affecting the complexity of a system is the feasibility of a VLSI implementation, or the possibility of using standard integrated circuits.

The adopted approach for solving the spectrum congestion problem is to select an inherently efficient modulation technique. New system concepts are then used to solve the major related problems. The technique is then modified to suit the new system configuration. The specific application examined in this thesis is the analog speech transmission over mobile channels. The approach can, however, be extended to any application.

There are other ways to achieve a better use of the spectrum; time multiplexing, and frequency reuse systems are some examples of these other approaches. It is

important to notice that many of the systems proposed in this work could be incorporated with these approaches. This would have an additive effect with respect to spectrum efficiency. SSB communications, for example, could be used within a cellular system, so, frequency reuse could be achieved. The use of time multiplexing for speech and data transmission in mobile systems is also discussed.

1.1.3 Thesis Organization

The thesis is organized in seven chapters. This subsection includes a brief description of the contents of each of these chapters.

The first chapter, the introduction, includes the thesis objectives and a description of the testing set-up used.

Chapter (2) is a brief description of the main characteristics of the mobile channel. A good understanding of the impairment mechanisms of a channel is essential to improve the utilization of the channel.

Chapter (3) includes a survey of the different techniques used for speech transmission over mobile channels. These techniques are evaluated and compared for different user environments. Some modulation techniques relevant to the approach adopted in this thesis are described in some details.

In order to test and evaluate the various algorithms proposed in this thesis, a programmable real-time mobile channel simulator compatible with the testing set-up was required. Since such a simulator is not available as a ready-made affordable product, it was necessary to design and implement one. Although the design of this simulator might look as a slight detour from the main stream of the thesis, it was an essential step for completing the work and getting realistic results. For

completeness purposes, chapter (4) starts by an overview of the various mobile channel simulation techniques. The second part of the chapter describes the real-time mobile radio channel simulator. The design of this simulator is also related to the main goal of the thesis as one more example of the potentials of digital systems design and digital signal processing, in the field of mobile communications.

Chapter (5) deals with the background acoustic noise in cars. Background acoustic noise represents one of the major sources of speech corruption in the mobile environment. In order to ensure a high speech quality while using a spectrally efficient modulation technique it is important to reduce the background acoustic noise from the transmitted speech. The first part of this chapter describes several experiments carried in order to characterize the background acoustic noise in cars. In the second part of the chapter several methods are proposed to reduce the background acoustic noise from the transmitted speech in hand-free mobile telephony.

Chapter (6) includes a detailed description of a modulation scheme proposed for speech and data transmission over mobile channels. The proposed scheme, the Switched Tone (ST) concept, is evaluated and its various subsystems are analysed. These subsystems are the speech companding, the automatic gain control (AGC), and the automatic frequency control (AFC). Although each one of these subsystems is designed to be appropriate to the ST concept, it could be slightly modified to fit other systems. The ST concept allows the transmission of analog speech as well as data. This thesis addresses the global system design of a ST based system. A detailed description is carried only for analog speech transmission, which uses a modified single sideband approach.

Finally, the different conclusions are presented in chapter (7). The chapter also includes some proposed future work.

1.2 Testing Setup

Introduction

Most of the schemes proposed in this work were evaluated and tested using real-time simulation. Several dedicated hardware boards were designed and implemented for this purpose. Some non real-time computer simulation was also carried to investigate a few concepts. The reasons for using real-time simulation are outlined in section 1.2.1. The main processor used, the TMS-320, and its associated systems and testing equipments are briefly presented in section 1.2.2. The need for a multi-microprocessor system is discussed in section 1.2.3. This system is described in section 1.2.4.

1.2.1 The Need for Real-Time Simulation

Real-time simulation requirements are more complicated than non real-time computer simulation. Nevertheless real-time simulation was chosen for most of the tests. Real-time simulation offers many advantages such as the possibility of testing over long time periods, the short testing time, and the flexibility of changing the test conditions during testing. Real-time testing also helps in getting a better understanding of the feasibility of the proposed system for final implementation. There are on the other hand some difficulties such as the need for sophisticated test equipments and dedicated hardware, and the long initial setup time required.

The first advantage of real-time testing is the possibility of testing over long time periods. This is essential for a fair evaluation of the speech quality. Most of the systems to be evaluated involve speech processing. The background noise cancellation and the speech companding are examples of the functions that need

long-term testing. In real-time testing, no digital speech storage is required, therefore, no limits are imposed on the speech length to be tested. In non real-time testing on the other hand, the speech segment to be tested is usually stored digitally in the computer. The storage restrictions limit this speech segment to very short time periods.

The possibility of testing over long periods of time is very important in the mobile environment. It allows for evaluating an algorithm over long periods of shadowing and fading, hence providing statistically valid results. Shadowing is very slow compared to multipath fading. The shadowing period can reach several minutes which needs a very large storage if non real-time simulation is used.

The second advantage of real-time testing is the time required to perform different tests. The initial system setup time for a real time testing system is very long. Whereas, once the system is in operation, testing becomes very fast. It is, therefore, possible to try different parameters for a specific algorithm and observe their effects on the spot. No time and resource consuming operations, such as recompilation and rerunning are needed. This gives the possibility of trying a large number of alternatives in a very short time. Repeatability is also possible by storing the speech using a tape recorder.

The third advantage is the possibility of changing the parameters of an algorithm while it is running and observe the effects on the fly. This feature could be very important in subjective speech quality tests.

The fourth advantage of real-time testing is the possibility of having a prototype for demonstration and field tests. The algorithms used in the prototype are the ones used for evaluation with slight modifications. The high correlation between the testing system and the final product gives a better understanding of the complexity required, and helps in choosing the most appropriate algorithms

for the actual implementation.

There are other factors affecting the choice of real-time implementation such as the availability of equipments and computing facilities, the previous experience, and the flexibility of the testing setup for other purposes.

There are however some drawbacks to real-time testing. The main disadvantage is the long time required to design, implement, and test the necessary dedicated hardware. The initial programming is also complicated and time consuming. The programming is done in assembly language. No routine libraries are available, therefore, all the required routines have to be written. The program has to satisfy the restrictions imposed by the execution time, the program memory size, and the data memory size. Another drawback is the complexity of performing some specific quantitative evaluations of the results. Plotting the results and making charts is not as easy as in non real-time testing where all the results are stored in a file that could be accessed any time at any rate.

1.2.2 TMS-320 and Associated Systems

The real-time simulation is performed using the Texas Instruments TMS-320 chip. The TMS-320 is a general purpose microprocessor aimed at digital signal processing applications. The only version of the chip, available at the start of the work, was the TMS-32010. It uses NMOS technology and is clocked at 20 MHz. The execution time of most of the instructions, including a 16 by 16 bits multiplication is 200 nanosec. It has serious limitations regarding the data memory space available. Some hardware was designed to overcome this problem.

Two TMS-320 utility boards, the evaluation module and the analog interface board were also used. These boards are manufactured by Texas Instruments. The

evaluation module (EVM) is a development system for the TMS-320. It can accept text files from a host CPU. The resident assembler converts the incoming text into executable code. The EVM's operating system firmware can be divided into four segments, the debug-monitor, the assembler-reverse assembler, the text editor, and the PROM utility. The EVM board is based on two processors. The TMS-320 is used to execute the user's code in real time, and the TMS-9995 processor functions as a master in executing the operating firmware. The other board, the AIB includes several input and output digital ports, a 12 bits A/D, a 12 bits D/A, and their necessary filters.

Many TMS-320 based circuits have been designed and implemented. These circuits were used as part of the testing setup. The most important one is the channel simulator described in detail in section 4.2. An interface board has also been designed and implemented to connect the different TMS-320 based boards together and to the outside world.

The testing system is controlled by a desk top computer, the Texas Instruments 'Professional Computer', equipped with a 10 Megabytes hard disk, an 8087 numeric co-processor, a high resolution graphics board, and a dot matrix printer. Many testing equipments were used, an audio band spectrum analyser, a plotter, several oscilloscopes and signal generators.

1.2.3 The Need for Multiprocessors

The TMS-32010 chip has a serious limitation with respect to the data memory size. Most of the programming time is spent in optimising the use of data memory locations. In order to fulfill the real time constraints, inline programming has to be used to avoid the 'call' instructions. These two factors imply the use of

unmodular and tricky programming techniques. The debugging of such programs is time consuming, and any modification becomes almost impossible, which defies the main design objective of the testing system, the flexibility.

To overcome the data memory size restriction a circuit has been designed to expand the data memory of the chip externally. The external data memory in this case was configured as a stack. This circuit was useful for testing short algorithms.

A more radical approach was necessary. The computational power of the system had to be increased to allow modular, flexible, and straightforward programming. Testing of more complex algorithms was also needed. This could only be achieved through adding more processors.

1.2.4 Multiprocessor System Description

The testing system designed and implemented can have one of two configurations. The first is suited for evaluating a transmission scheme over mobile channels. The second is suited for testing numeric intensive algorithms such as very long transversal adaptive filters.

The first configuration is shown in figure 1.1. Two TMS-320 evaluation modules (EVM) and the channel simulator described in section 2.3 are used. The configuration includes an EVM acting as transmitter, the TMS-320 based channel simulator, and an EVM acting as receiver. The interface board generates the necessary clocks and control signals to synchronize the processors. It is equipped with two output ports connected to two D/As and their necessary antialiasing filters. These ports are accessed by the EVMs; therefore, any two signals within the tested system can be sent to the required port. These signals can be connected to the oscilloscope or the spectrum analyser. Two other analog ports are available

Figure 1.1: Testing Setup, First Configuration

on the interface board for the input output handling of the speech signal to be tested. A pseudo-random number generator was also built for data testing. This configuration is suitable for testing any scheme under fading conditions.

Figure 1.2 shows a block diagram of the second configuration of the testing system. It is a more flexible architecture, allowing easy expansion and easy debugging. Parallelism is achieved through array processing as well as pipeline processing. This architecture has been optimised for adaptive filtering-like processing. It was used successfully for other purposes as well.

All the processing units forming the pipeline are similar. Each unit includes one input and one output port connected to the bus of the master processor, and one input port connected to the previous stage. Its bus is connected to the next stage and can write into its input port. The architecture is very modular, and the processing power can be increased very easily by adding more processing units to the pipeline. No extra control is needed.

To describe this configuration it is better to consider a specific algorithm. Let us consider the case of a long adaptive transversal filter using the stochastic gradient algorithm for adaptation. Figure 1.3 shows a block diagram of the filter. The coefficient adaptation algorithm for stage i is:

$$c_i(t+1) = c_i(t) + y_i(t) \delta e \quad (1.1)$$

where:

- $c_i(t+1)$ is the updated coefficient of stage i
- $c_i(t)$ is the old coefficient of stage i

Figure 1.2: Multiprocessor Testing Setup, Second Configuration

- $y_i(t)$ is the output at stage i
- b is the step size
- e is the error term

The information required to update one tap includes local information $c_i(t)$ and $y_i(t)$ and global information $(b \cdot e)$. The tasks are distributed as shown in figure 1.3. The master processor handles the input/output operations as well as the necessary scaling. It calculates the product $(b \cdot e)$ and broadcasts it. The other processors, 1 to N , form a pipeline. Consider a middle stage unit. It gets the updating factor $(b \cdot e)$ from the master processor and its input data sample from the previous unit. The unit updates everyone of its coefficients, calculates its partial sum, shifts its data samples. Then it sends the partial sum to the master unit and the last data sample to the next unit.

In the case considered, all the units in the pipeline execute the same software. In general this is not the case. A complicated algorithm can be partitioned between the different processors.

A system including a master processor and four processing units in the pipeline was built to run a 256 taps adaptive transversal filter.

Chapter 2

The Mobile Channel

Introduction

In order to improve the utilization efficiency of a channel, a good understanding of the impairment mechanisms of the channel is essential. This chapter is an overview of the main characteristics of the mobile channel.

A radio signal transmitted between a fixed base station and a moving vehicle exhibits many impairments. These impairments can be sorted into two categories. The first one consists of the multiplicative impairments, and the second consists of the additive impairments.

The different multiplicative impairments such as fast fading and slow fading are presented in section 2.1.

The various additive impairments are presented in section 2.2. These additive impairments include background acoustic noise, thermal noise, and man-made noise such as interference.

Finally, a summary of the main characteristics of the mobile channel are presented in section 2.3.

2.1 Multiplicative impairments

Introduction

The multiplicative impairments encountered in the mobile environment can be classified into fast fading and slow fading. The fast amplitude and phase modulation attributed to multipath propagation is discussed in subsection 2.1.1. The discussion includes the effect of Doppler shifts. The slow variations of the local mean signal strength are discussed in subsection 2.1.2. These variations are caused by the terrain and environment. The description presented in this section is based on analytical models as well as field tests. A detailed description can be found in [21-27, 72].

In field tests, it is necessary to separate the fast fading and the slow fading. If a continuous carrier wave on a mobile channel, the received signal envelope is given by:

$$R(t) = m(t)r(t) \quad (2.1)$$

where:

- $R(t)$ is the received carrier envelope
- $m(t)$ represents the slow fading

- $r(t)$ represents the fast fading

Since only $R(t)$ can be measured, both $m(t)$ and $r(t)$ have to be estimated. An estimate of $m(t)$ is obtained by averaging $R(t)$ over a long time interval. Then, $r(t)$ and $m(t)$ are analysed in terms of amplitude statistics, autocorrelation, and frequency response.

2.1.1 Fast Fading

Many discussions of the multipath propagation problem can be found in [21-27]. Consider a signal transmitted from a fixed base station to a moving vehicle. This signal exhibits many reflections from the buildings and other structures in the vicinity of the mobile unit. The signal field at any point in space is described as a linear superposition of plane waves having random magnitude and random phase. A mobile receiver moving through this interference pattern experiences a random signal fading. Fades of 40 dB or more below the mean level are common in the mobile environment.

Many statistical models have been proposed to characterize multipath propagation. The most widely accepted model is the one derived by Clarke [22]. It assumes that the energy arrives at the receiver by way of a number of indirect paths. The phase and angle of arrival of the waves are completely random and statistically independent. According to Clarke's model the amplitude statistics of the received signal envelope are Rayleigh, and the phase is uniformly distributed.

Many modifications have been proposed to Clarke's model. Turin was one of the first to report experimental results on Rayleigh fading combined with lognormal fluctuations in the local mean [24, p.18]. These fluctuations will be discussed in section 2.1.2. Work by Rice was used to extend the model to cover the cases

when a line of sight path exists between the transmitter and the receiver. In this case, the probability density function becomes Rician. All these previous results have been confirmed by several experimental results [24,26,27,72].

Besides the random phase, each plane wave component reaches the mobile receiver with a different Doppler shift. The Doppler shift depends on the mobile speed, the carrier frequency, and the angle the propagation vector makes with the mobile velocity vector. This implies that the apparent power spectrum of each of the received field components is broadened to occupy a narrow band about the carrier frequency. This phenomenon results in having the received signal frequency slightly different from the transmitted frequency. The Doppler shift is given by:

$$f_D = \pm(v/\lambda) \cos \theta \quad (2.2)$$

where:

- f_D = Doppler frequency (Hz)
- v = Vehicle speed (m/sec)
- λ = Wavelength (m)
- θ = Angle of arrival

In the $900MHz$ band, for example, a vehicle speed of $50km/hr$ introduces a Doppler shift of about $42Hz$. If the vehicle speed reaches $100km/hr$, the Doppler frequency shift is about $84Hz$. Therefore, in land mobile communications, the Doppler spread is usually bandlimited to about $\pm 100Hz$.

2.1.2 Slow Fading

Apart from the fast fading described in the previous section, slower variations of the mean signal strength are encountered in mobile communications.

These slow variations depend on the terrain and the environment. It is generally observed to obey a lognormal probability density function [24], and the variations occur at a rate about 200 times slower than the Doppler frequency. Therefore, typical rates are a fraction of one Hz .

2.2 Additive Impairments

Apart from the fast and slow fading, which have a multiplicative effect, there is another source of impairment; the noise, which has an additive effect. The noise in the mobile environment can be sorted into several categories, acoustic noise, thermal noise, interference, and man-made noise.

The acoustic noise in cars is due to mechanical friction, air flow, and street noise. It is coupled to the transmitted speech through the microphone. Chapter (5) of this thesis addresses the problem of acoustic noise in mobile communications. The various sources of background acoustic noise in cars are analysed, and several approaches are proposed in to reduce this noise from the transmitted speech signal.

Thermal noise in mobile communications is similar to any other communication environment. It can be represented as a zero mean white Gaussian noise added to both the in-phase and the quadrature components of the transmitted signal.

Interference can be classified into co-channel interference and adjacent channel interference. The co-channel interference is due to the use of the same frequency

in a close geographical area. It is affected by the modulation technique, the power, and the frequency reuse scheme used. The adjacent channel interference depends also on the modulation technique, the power, the guard bands between the channels, and the receiver selectivity.

Finally, the man-made interference is due to the ignition system and the other electrical systems in the mobile. This type of noise is usually impulsive.

Reference [28] includes a detailed analysis and description of the various types of noise encountered in the mobile environment.

2.3 Summary

When a signal is transmitted on a mobile channel it exhibits multiplicative impairments as well as additive impairments. The multiplicative impairments can be classified into fast fading and slow fading. The fast fading is characterized by a Rayleigh amplitude distribution and a uniform phase distribution. When a line of sight is present, the amplitude distribution becomes Rician. The rate of change of the fast fading is bandlimited to the Doppler frequency, which is a function of the carrier frequency and the vehicle speed, as well as the angle between them. The slow variations have a lognormal amplitude distribution. They occur at a rate about 200 times slower than the Doppler frequency.

The additive impairments are due to the acoustic noise, the thermal noise, and the interference. The acoustic noise is coupled to the transmitted speech through the microphone. The thermal noise is zero mean, white, and Gaussian. The interference is due to co-channel and adjacent channel interference, as well as engine interference. The engine interference generates impulsive noise.

Chapter 3

Speech Transmission over Mobile Channels

Introduction

The mobile channel has been used since 1897 [21,p.1]. Nevertheless, its use is dominated by one modulation technique, analog FM. The actual increasing demand for mobile channel allocation and the limited spectrum available imply a more efficient use of that channel. Therefore, the feasibility of different modulation techniques in mobile communications has to be investigated.

Section 3.1 includes a brief discussion of the spectrum congestion problem. A classification of the various possible techniques for improving the utilization of the frequency spectrum is also presented.

Section 3.2 is an overview of the different modulation techniques used for speech transmission over mobile channels. An attempt is made to evaluate each one of

these techniques. The discussion includes analog and digital speech transmission.

Section 3.3 includes a more detailed overview of single side band (SSB) modulation techniques and its various modified versions. This special emphasis is due to the fact that the modulation technique proposed in this thesis is an SSB derivative.

3.1 Mobile Radio Spectrum Utilization

In recent years there has been an increasing demand for mobile radio channels allocation. This demand is expected to grow more and more in the future. On the other hand, the available frequency spectrum is limited. This section is an overview of the various techniques used for increasing the frequency spectrum utilization in mobile communications [3].

The first approach is to increase the number of radio channels in the available spectrum. This is achieved by narrowing the frequency spectrum of each channel. An example of this approach is the use of single sideband modulation (SSB) instead of frequency modulation (FM). A 3 KHz speech signal can be transmitted using a 5 KHz channel spacing using SSB techniques. The same signal requires a 30 KHz channel spacing using FM techniques, which represents a saving ratio of 6:1 as far as the channel bandwidth efficiency is concerned.

The second approach is to improve the spatial frequency spectrum reuse. This is achieved through a cellular system. In this case, the same channel can be reused in a different cell; therefore, the number of available channels in a given area can be increased. The spectrum efficiency here is a function of the channel bandwidth as well as the limitations regarding frequency reuse. These limitations are implied

by the power requirements and the sensitivity of the modulation technique used regarding co-channel interference. The spectrum efficiency in this case could be defined as the number of calls per hour per square mile per Hz.

The third approach is to improve the timewise frequency spectrum reuse. This is achieved by dynamic channel assignment in a cellular system. Other techniques such that time assigned speech interpolation (TASI) can also be used to improve the timewise spectrum reuse.

The fourth approach is to reduce the invalid frequency spectrum used. This could be achieved by using variable bit rate control, where the bit rate or the spectrum is assigned according to the service class, and the speech quality required.

A fifth possible approach is to reduce the traffic without losing information. This can be achieved by using the information compression technique appropriate to the specific application.

Finally, it is concluded that the spectrum efficiency could be improved using a large number of approaches. The best spectrum efficiency for a given application depends on the quality of service required, and can be achieved using the proper combination of the previous approaches.

3.2 Speech Transmission Techniques

Introduction

This section is an overview of the various speech transmission techniques used in the mobile environment. In subsections 3.2.1 and 3.2.2 the analog FM and the analog AM modulation techniques are presented. Section 3.2.3 is a overview of

the digital speech transmission techniques. Finally, in section 3.2.4, these various transmission techniques are compared and some conclusions are presented.

3.2.1 Analog Frequency Modulation (FM)

At present time, analog frequency modulation (FM) is the preferred modulation technique for speech transmission over land mobile channels. The main reasons for that are technical as well as economical.

The main technical advantage of analog FM is its immunity to amplitude fading and interference. FM can capture the signal from interference even when the signal to interference ratio is small. Another advantage is that in FM, the baseband signal to noise ratio can be improved without necessitating an increase in the transmitted power. This is done by increasing the frequency deviation.

The main economical advantage of analog FM was its simplicity. Both the FM transmitter and the FM receiver are very simple, which gave FM a cost advantage over other modulation techniques. But now, with the recent advances in VLSI technology, this cost advantage is not anymore valid. On the other hand, for security and privacy reasons the FM simplicity could be considered a drawback.

Despite its numerous advantages, analog FM has a serious disadvantage. Actual analog FM systems use 30, 25, 20, or 12.5 KHz channel spacing [1, 2, 28]. This spectrum waste may not be tolerable in the future with the spectrum congestion and the expected growth in demand. Therefore, for telephone quality speech transmission over land mobile channels, analog FM is not expected to remain the dominant modulation technique. For other applications, such as broadcast quality speech transmission, analog FM may remain unchallenged.

3.2.2 Analog Amplitude Modulation (AM)

In any amplitude modulation system, the information resides in the envelope of the carrier. In the land mobile environment the envelope of the carrier is susceptible to fading. Hence, recovery of the information becomes impossible unless some technique is used for channel equalization.

The main analog amplitude modulation techniques are double sideband (DSB) and single sideband (SSB). In a DSB modulator, the baseband signal is multiplied by a carrier frequency, which is a frequency translation operation. The frequency spectrum occupied by a DSB signal is double the spectrum of the modulating signal. The spectrum of any real-valued signal must exhibit the symmetry condition, and hence, the sidebands of DSB are uniquely related to each other by symmetry. Given the amplitude and phase of one, it is always possible to reconstruct the other; this leads to SSB modulation. In SSB, only one of the two sidebands that result from the multiplication of the message signal with a carrier is transmitted. Therefore, the spectrum occupied by a SSB signal equals the baseband spectrum of the modulating message signal. The main advantage of SSB is, therefore, its spectrum efficiency.

In this thesis a new concept, the switched tone concept, is proposed for speech and data transmission over mobile channels. In the speech transmission mode, the switched tone concept uses a variation of SSB. Therefore, a detailed evaluation of SSB and its various derivations is carried in a separate section, section 3.3.

The various SSB based systems described in section 3.3 try to overcome the main deficiencies of SSB while maintaining its principal advantage, spectrum efficiency. For toll quality speech transmission, some of these systems may be good candidates to replace existing analog FM. For other applications, requiring a better

speech quality, the SSB derivatives might not fulfill the application requirements.

3.2.3 Digital Transmission

Digital speech transmission is an attractive approach for all communication systems. It is possible to combat most of the impairments introduced by a channel through the use of more robust modulation techniques as well as error correction schemes. The price paid for that would be more bandwidth, more power, or more complexity. Digital speech transmission also makes radio transmission compatible with some standard networks such as the ISDN.

Direct pulse code modulation (PCM) is not efficient for coding digital speech transmission. It does not explore the inherent redundancy of the speech signals. Two main approaches are used to explore this redundancy. The first approach is to use waveform coding techniques such as delta modulation or adaptive predictive PCM (ADPCM). The second approach is to use vocoders. At present time, extensive research is being done in these areas [76]. Nevertheless, at least for the near future, given the high error rate encountered in the hostile mobile environment, it seems difficult to achieve a high quality digital speech transmission within a 5 KHz bandwidth for use in mobile communications.

3.2.4 Conclusions

Analog FM is the most widely used modulation technique for speech transmission over land mobile channels. The current radio spectrum congestion and the expected future demand imply the use of more spectrally efficient modulation techniques. Several variations of the analog SSB techniques have been proposed

lately. These techniques overcome most of the weaknesses of standard SSB while maintaining its spectrum efficiency. Some of these techniques are now mature enough to replace existing analog FM for speech transmission over mobile channels, commercially.

Many new digital techniques suitable for toll quality speech transmission over narrowband mobile channels are under development in different research laboratories. Some of these techniques offer solutions to most of the existing problems. Their complexity, however, is far beyond the VLSI capabilities. Digital speech transmission, however, is likely to take over sometimes in the far future.

3.3 SSB Communication

Introduction

Single sideband (SSB) modulation was introduced in section 3.2.2. In this section, an evaluation of SSB in the mobile environment is presented. In section 3.3.1 the advantages and disadvantages of SSB communication are discussed. The different existing SSB-based systems are presented in section 3.3.2.

3.3.1 Advantages and disadvantages of SSB

In this subsection, the advantages and disadvantages of SSB are presented. Channel bandwidth efficiency, and sensitivity to fading, noise, and interference are discussed.

The main advantage of SSB is its channel bandwidth efficiency. Consider the case of a speech signal, ranging from 0.3 to 3.3 KHz, to be transmitted in the

mobile environment. Theoretically, this signal needs a bandwidth of only 3 KHz to be transmitted. Practically, using the modified SSB technique proposed in this thesis, it needs 5KHz. Whereas, in the analog FM system presently used in North America, the same signal is transmitted using a (20-30) KHz channel spacing. Therefore, the proposed SSB system represents a saving of up to 6:1 in terms of channel bandwidth efficiency.

There is another advantage of the proposed SSB schemes. Many subsystems for analog scrambling and descrambling could be added without too much extra complexity. There is even some implicit scrambling-descrambling inherent in the basic system. This is a very important factor for future communication trends.

Despite its spectrum efficiency, SSB has not spread yet in commercial applications. The reason for that is that there are many problems associated with it. Its sensitivity to fading and to noise, the lack of any capture effect, interchannel interference, and complexity of the equipment are some examples of these problems. Therefore, the use of SSB without any corrective action leads to a very low speech quality, which is unacceptable in any commercial application.

The first disadvantage of SSB is its sensitivity to amplitude and phase fluctuations. In a mobile environment, motion in a multipath field generates severe amplitude and phase fluctuations. SSB is, therefore, unusable in the mobile environment, unless some corrective action is taken. A reference signal, known by the receiver has to be transmitted along with the speech signal. The receiver can, therefore, retrieve the amplitude and phase fluctuations, and use this information to correct the received signal. This reference signal is usually a pilot tone inserted within the transmitted signal. Several pilot-aided SSB systems have been proposed, the difference between them being the location of the pilot tone within the speech band. These different pilot-aided systems are presented and evaluated in

section 3.3.2. In this thesis, a novel concept concerning the pilot insertion scheme is proposed.

The second disadvantage of SSB is its sensitivity to amplitude and phase noise. The amplitude noise is directly added to the demodulated baseband signal. This is reflected in a poor speech quality due to low signal to noise ratio. This phenomenon is more serious in low volume syllables where the signal power is very low while the noise power remains the same. In order to reduce the effect of this problem, the low volume syllables are amplified at the transmitter, and restored to their original level at the receiver. This technique is known as speech companding. The various classes of speech companding are discussed in section 6.2 of this thesis, and a novel dynamic speech companding algorithm is proposed.

The third disadvantage is the initial frequency acquisition and synchronization. An initial frequency offset between the transmitted carrier and the receiver's demodulating carrier can be as large as a few kHz. As an illustrative example, consider the case of transmission at 500 MHz using a crystal with 10^{-6} precision, a frequency offset up to 1000 Hz can occur. If this offset is not taken care of, the speech quality, using conventional SSB techniques becomes severely degraded and hence unacceptable. The frequency offset can reach a few KHz in the 900 MHz band, specially if a less accurate crystal is used. The reference pilot tone used to retrieve the amplitude and phase fluctuations could be used for the initial frequency acquisition.

The fourth disadvantage of SSB is its susceptibility to interference. There are mainly two types of interference, the co-channel interference, due to the use of the same frequency in a close geographical area, and the adjacent channel interference. In pilot-aided SSB, a third type of interference is manifested; the interference, within the band between the pilot tone and the audio signal. This

problem is reduced by decreasing the pilot level as much as possible and by ensuring a sufficient guardband between the pilot tone and the voice signal. The interference problem is not explicitly analysed in this work. The various aspects of interference in SSB systems used in the mobile environment are discussed in [6, 8, 9, 18].

The fifth disadvantage is the nonlinearity introduced by the high power amplifier (HPA) in the SSB modulator. This nonlinearity is severe at the edges of the band. In order to reduce this problem, some preemphasis techniques are proposed in reference [16]. The detailed analysis of this problem is out of the scope of this work.

Finally, the sixth and last disadvantage of the modified SSB schemes is the complexity of the equipment. Traditionally, complexity was exponentially linked to cost and reliability. Now, the recent VLSI advances have altered this criterion. Also, VLSI advances made digital signal processing a feasible solution. The complexity involved in the new proposed schemes is believed to be within the capabilities of present VLSI technology.

As a conclusion, it is believed that conventional SSB modulation is unusable in the land mobile environment, unless some corrective actions are taken. The insertion of a pilot tone within the speech band can provide the necessary information required for automatic gain control, automatic frequency control, and initial frequency acquisition. This way, it is possible to overcome most of the drawbacks associated with conventional SSB, while maintaining its main advantage, the bandwidth efficiency. Speech companding is also necessary in any SSB system to overcome the problem of sensitivity to noise.

3.3.2 Pilot Aided SSB

Introduction

In order to overcome the deficiencies of conventional SSB modulation, it is necessary to send a reference pilot tone along with the speech signal. This pilot is used by the receiver to estimate the frequency offset, the amplitude fluctuations, and the phase fluctuations. The location of the pilot tone within the speech band has been the subject of many studies. Many systems, such as the partially suppressed carrier, the tone above band (TAB), the tone in band (TIB), the transparent tone in band (TTIB), and the switched tone (ST) have been proposed. These different schemes, shown in figure 3-1, are presented in the next subsections.

Partially Suppressed Carrier Concept

The partially suppressed carrier system has been developed by Philips Research Laboratories [10]. According to this concept, the suppressed carrier acts as a pilot tone and is used by the receiver for automatic gain control and automatic frequency control. There are several ways to generate a partially suppressed carrier single sideband signal. It is possible to generate a SSB signal using a Hilbert transform and then add the attenuated carrier to it. Another way is to generate a double sideband signal with carrier and filter out one of the sidebands.

The main advantage of this scheme is its transparency with all speech and data formats; the baseband signal is transmitted as is, it is not split in the frequency domain or distorted by pilot injection. Therefore, no restrictions are imposed on the transmitted baseband signal and, theoretically at least, no amplitude or phase distortion is introduced in it. The concept is very simple and achieves reasonable

Figure 3.1: Different Pilot Aided SBB Systems
(Baseband Spectrum)

results.

There are several disadvantages to the partially suppressed carrier SSB systems. These disadvantages come from the fact that the pilot is at the edge of the band. The first disadvantage is the pull-in range of the AFC circuit, the second is the distortion of the pilot, the third is the adjacent channel interference, and the fourth is the correlation between the pilot fluctuations and the signal fluctuations.

The first disadvantage is the pull-in range of the AFC circuit. This problem arises only in the initial synchronization between the transmitter and the receiver. The frequency offset can reach few KHz, which requires an asymmetric pull-in behaviour from the AFC circuitry. When the pilot is at the edge of the band, it may go completely out of band, due to this initial frequency offset. It is, therefore, attenuated and distorted by the various filters and amplifiers in the receiver. In this case it becomes very difficult for the receiver to recover the pilot and the transmitted signal. A simple way to reduce this problem is to use better crystals and to increase the pilot level. This is very costly in terms of cost, power efficiency, and interchannel interference.

The second disadvantage is the distortion of the pilot tone. The pilot tone is positioned in a region of high differential group delay with respect to the majority of the audio band. This is attributable to the IF crystal filter. It can degrade the performance of the AGC [12] and AFC [13].

The third disadvantage is the adjacent channel interference. This problem is reduced by reducing the pilot level and by shifting it as much as possible from the edge of the band. This is not a simple solution since reducing the pilot level affects the performance of the AGC and AFC, specially during the initial acquisition period. Shifting the pilot is not possible in the partially suppressed carrier system.

The fourth disadvantage is the correlation between the pilot fluctuations and the signal fluctuations. The pilot fluctuations are extracted and inversely applied to the signal. Therefore, there should be a very high correlation between the pilot and the signal fluctuations for the system to work properly. According to the results obtained so far, this does not seem to create any problems. In the SSB system under consideration the signal is bandlimited to 5 KHz. In such a narrow band, it is reasonable to assume that the fading is flat and not frequency selective.

The Tone Above Band Concept (TAB)

The tone above band (TAB) concept involves sending the pilot tone above the voice band. Reference [7] describes the different aspects of the TAB concept. Some commercial systems such as the amplitude companded single side band (ACSB) use the TAB concept. The exact location of the pilot tone in the TAB concept is not fixed, which gives some flexibility in the design not provided in the partially suppressed carrier concept.

Any TAB system has, in general, the same advantages and disadvantages of a partially suppressed carrier system. Its main advantage is its transparency to all speech and data formats. Its disadvantages lie in the initial acquisition of the offset frequency shift, the distortion introduced to the carrier, the adjacent channel interference, and the correlation between the pilot fluctuations and the signal fluctuations. However, the extra flexibility in choosing the exact pilot location helps in overcoming some of these disadvantages. A modification is proposed in this thesis to overcome the initial acquisition problem encountered in the TAB concept. The proposed concept, the switched tone (ST), is introduced later in this chapter; it is also described in detail in chapter 6.

Tone In Band (TIB) Concept

The tone in band (TIB) concept is described in [11]. A part of the audio spectrum is removed from the central region of the audio band by a notch filter for the tone to be inserted.

TIB achieves a very good spectrum efficiency. It offers a large symmetrical pull-in range for the frequency control circuit. It also offers the greatest degree of adjacent channel protection, and the highest correlation between fades on the pilot and fades on the audio signal.

The main disadvantage of TIB is the irreparable distortion introduced to the audio signal. In removing the central region of the audio signal, some information is lost and it cannot be recovered. In order to apply the TIB concept to data, the data has to be precoded to introduce a null at midband.

In summary, the TIB solves most of the problems encountered in partially suppressed carrier systems and in TAB systems. However, it introduces more serious distortion in the speech signal. The speech quality, specially in VHF is very poor. TIB has not received universal acceptance because of its non-transparency with all data formats [14].

Transparent Tone In Band (TTIB) Concept

The transparent tone in band (TTIB) is a modification to the TIB concept. As explained in the previous section, in TIB, a notch filter is used to remove the central region of the audio signal. The pilot tone is then inserted in this gap. In TTIB, the spectrum of the audio signal is split and shifted to create a null in the middle. The tone is then inserted in this null. A detailed description of the TTIB

concept is found in [15].

Figure 3-2 shows the different steps used to transmit a signal using the TTIB concept. The upper half of the audio signal spectrum is filtered out. It is shifted upward in frequency. The lower half of the audio signal spectrum is then added to the shifted version of the upper half. This way, a gap is generated between the upper and the lower bands of the audio signal. A pilot tone is inserted in this gap. In the receiver, the reverse operations are performed. The pilot is filtered out, the upper half of the audio spectrum is downconverted to its original location.

The main difference between TIB and TTIB is that in TIB some of the signal information is destroyed. Whereas in TTIB many complicated operations are performed to avoid this loss of information. With ideal filters, it is possible to avoid all the problems of the previous methods. Whereas, when practical filters are considered, many problems arise. The slope of the cutoff edge of the filters, the phase distortion, the group delay, and the small frequency drifts between the transmitter and the receiver introduce many distortion in the audio signal, specially at midband.

Switched Tone (ST) Concept

The switched tone (ST) concept is introduced in this thesis as a compromise between the TAB and the TTIB concepts. In this section, the motivation behind this new approach is presented. The switched tone concept is then described and its implications on channel tracking are discussed.

One of the main disadvantages of TAB lies in the initial frequency offset between the transmitter and the receiver. If this offset is too high, the pilot tone is out of band and cannot be recovered by the receiver. After the initial acqui-

a) Filter Upper Half of the spectrum

b) Apply a Frequency Shift to Upper Half

c) Add Lower Half and Pilot

Figure 3.2: Transparent Tone in Band (TITB) Concept

tion, the system's performance is satisfactory. In TTIB, on the other hand, the initial acquisition does not constitute a problem, whereas the fact that the audio spectrum is processed introduces some distortion. In the proposed switched tone concept, a midband tone is transmitted for initial acquisition. It is followed by a normal TAB transmission.

The proposed switched tone concept is conceived for the transmission of analog speech as well as digital data. The speech is transmitted on a talkspurt by talkspurt basis. When a speech burst is detected using a speech detector, a midband tone is transmitted followed by a 'start speech' message. The speech is then transmitted using analog SSB TAB techniques. When the talkspurt ends, an 'end speech' message is transmitted. Similarly, when a data burst is to be transmitted, a midband tone is transmitted, followed by a 'start data' message. At the end of the data session an 'end data' message is sent. The receiver has two modes of operation, the speech mode and the data mode. According to the start message it switches to the mode designated by the message. When an end message is detected, it waits for a new start message. Figure 3-3 shows the different steps performed to transmit an audio signal using the switched tone concept.

The switched tone concept offers many advantages. The first is its flexibility in treating the speech on a speech-spurt by speech-spurt basis, allowing the possibility of time multiplexing of speech and data. The second advantage is the large symmetric pull-in range for initial frequency offset. The pull-in range is one half of the bandwidth of the baseband signal. This allows the use of less accurate crystals and ensures a better initial acquisition. The third advantage is its transparency for any analog or digital format. The baseband signal is transmitted as is and no specific spectral shape is required, unlike TTIB. Another advantage is the possibility of reducing the level of the pilot tone after the end of the initial acquisition

Ampl.

a) Transmit Midband Pilot Tone

Ampl.

b) When Speech Detected
 Send 'Start Speech' Message
 (If Data is To Be Sent,
 Send 'Start Data' Message)

Ampl.

c) To Transmit Speech
 Send TAB Reduced Carrier

Figure 3.3: Switched Tone (ST) Concept, Typical Speech Transmission Session

period, and therefore reducing the effect of adjacent channel interference. It is also possible to incorporate scrambling techniques to ensure a better secrecy and privacy.

There are however some disadvantages in using the switched tone concept. At first look, the complexity of the control system, specially in the AFC subsystem, seems to be a big disadvantage. The complexity here is not in processing the speech itself like the TTIB. It is only in the control system and does not require extensive arithmetic operations.

Chapter 4

Mobile Channel Simulation

Introduction

The testing and evaluation phases of any new concept in mobile communications is a difficult and cumbersome task. A moving vehicle, loaded with equipment, as well as a base station set-up are needed. Even then, the comparison between different techniques is not easy due to the impossibility of repeating the same fading patterns. A mobile radio channel simulator is, therefore, a necessary and an indispensable tool in any mobile communications research laboratory.

During the evaluation process of the various algorithms presented in this thesis, it was necessary to acquire a flexible real-time mobile radio channel simulator, compatible with the testing set-up. Since such a simulator is not available as a ready-made product, it was necessary to design and implement one. Although the design of the simulator might seem as a slight detour from the main stream of the thesis, it was an essential step towards completing the work, and getting realistic results. Moreover, it can be considered as one more example of the potentials of

digital signal processing, and system design, in the field of mobile communications.

The first part of this chapter is an overview of the approaches commonly used for mobile channel simulation. The second part of the chapter deals with the description and evaluation of the simulator developed for this work.

4.1 Channel Simulation

Many new applications are now emerging in the fields of mobile and portable communications [5]. Consequently, many coding and modulation techniques are under development. There is, therefore, an increasing interest in mobile radio channel simulators [29-35].

The basic configuration of the simulators has not changed over the years. The implementation technology, however, has evolved from mechanical gears, to electronic tubes, to discrete transistors, to discrete integrated circuits, to VLSI programmable chips. The latest transition has opened new prospects for channel simulators. It has allowed the addition of many new features as well as a high degree of flexibility, which has almost eliminated the gap between simulation results and actual results.

Figure 4.1 shows the block diagram of a basic fading simulator. This configuration is commonly used to simulate the multiplicative impairments, such as multipath fading, and Doppler shifts [29-30]. The incoming RF signal is split into its in-phase and quadrature components using a delay module and a Hilbert transformer. Each of these components is modulated by a low pass signal generated by the baseband section of the simulator, shown by the dotted lines. The two signals are then added to form the simulated RF received signal.

Figure 4.1: Basic Fading Simulator

All the multiplicative impairments of the channel can be simulated in the baseband section of the simulator, and carried in its output signals, I and Q. Therefore, the main innovations and new features concerning the multiplicative impairments lie in the baseband section

The most important multiplicative impairment in mobile channels is the multipath fading. Clark's model [22] is a widely accepted model for multipath propagation [29, 30 35, 71]. In this model, the received signal envelope fluctuates about the local mean with Rayleigh statistics, and the RF phase has a uniformly distributed random component.

The corresponding baseband section of the simulator is the one shown by the dotted lines of figure 4.1. Two independent, zero mean Gaussian noise generators modulate both the in-phase and the quadrature components of the RF signal. The fading spectrum appropriate to mobile radio is obtained by properly shaping the spectrum of the noise source. The spectrum depends on the antenna pattern being used in the mobile radio. For an omnidirectional antenna, which is the most common type, the spectral density of the envelope of the received signal is given by:

$$S(f) = \begin{cases} E^2 / (2\pi f_D \sqrt{1 - (f/f_D)^2}) & f < f_D \\ 0 & f \geq f_D \end{cases}$$

and

$$f_D = \pm v/\lambda \quad (4.1)$$

where:

- E is the rms value of the signal

- f_D is the Doppler frequency shift (MAX)
- v is the vehicle speed
- λ is the wavelength of the signal

This theoretical spectral density, representing the frequency response of the shaping filters, is shown in figure 4.2. The effects of having a practical filter instead of this theoretical one are discussed in [29,30]. It is important to notice that the cutoff frequency of the filter is at f_D , which is dependent on the vehicle speed and the wavelength. This means that in order to simulate different vehicle speeds and different wavelengths, it is necessary to have a variable cutoff frequency filter. These implementation aspects are described section 4.2.

4.2 Completed Channel Simulator

4.2.1 Motivation

In order to test and evaluate the proposed schemes, a channel simulator was required, with the following features:

- Real-Time
- Variable Vehicle Speed (0-120 km/hr, steps of 1 km/hr)
- Variable RF Frequency (1 MHz-999 MHz, steps of 1 MHz)
- Variable line of sight ratios (0-20 dB)
- Variable Shadowing values (0-50 dB)

Figure 4.2: Theoretical Frequency Response of the Shaping Filter

- Easy to use (User friendly, keyboard-display interface)
- Stand Alone (To reduce the complexity of the test bed)
- Interfaceable with the existing testing system (TMS-320)
- Flexible regarding RF or Baseband Complex Configurations
- Flexible for adding noise with various distributions

Such features are not available in any affordable ready product. Therefore, a channel simulator featuring the previous specifications was designed, implemented, and tested [71]. It was then validated, and used to evaluate different transmission schemes.

In section 4.2.2, a general description of the channel simulator is given. A detailed description of each block is given in the subsequent sections. The Gaussian noise source, multipath, shadowing, and line of sight blocks are described in sections 4.2.3, 4.2.4, 4.2.5 and 4.2.6 respectively.

4.2.2 General Description

The simulator is based on the TMS-320 DSP chip. It has two modes of operation, the RF mode and the baseband mode. In the RF mode, the simulator expects an RF input signal and generates an RF faded output signal, according to the desired parameters. In the baseband mode, the simulator generates two low-pass signals carrying the required multiplicative impairments. The baseband mode is described in details in section 4.2.7.

At power on, the simulator prompts the user for the carrier frequency, the vehicle speed, the line of sight ratio, and the shadowing ratio. These parameters

are entered using the keyboard. The simulator then goes into a free running mode, simulating the desired channel until the reset button is activated by the user. In such case, new parameters can be programmed. During a simulation session, the different simulation parameters are displayed on the simulator's display.

The basic block diagram of the channel simulator in the RF mode was shown in figure 4.1. The incoming RF signal is split using a Hilbert transformer, generating the in-phase and the quadrature components. The baseband section, shown by the dotted lines, generates two low-pass signals that multiply both components of the RF signal. All the multiplicative impairments of the channel are embedded in these two baseband signals. The two branches are then added to form the fading RF signal. The additive impairments could then be added to the RF signal.

A detailed block diagram of the baseband section is shown in figure 4.3. This section generates two low-pass signals, the I and the Q channels, carrying all the multiplicative impairment information. The multipath propagation and the Doppler effect are represented by two independent, spectrally shaped, zero mean, white, Gaussian distributed signals. The shadowing is represented by a scalar lognormal distributed signal multiplying both the I and Q channels. The line of sight is represented by a dc value added to the I channel.

4.2.3 Gaussian Noise Source

A source of white Gaussian noise is required at various stages of the baseband section. It is also required for the additive impairments section which will be introduced later on. The function of the Gaussian noise source block is to generate a sequence of uncorrelated numbers having a Gaussian amplitude distribution. This block is described in detail in this section.

Line of Sight
SectionMultipath Propagation
and Doppler SectionShadowing
Section

Figure 4.3: Baseband Section of the Channel Simulator

Two different algorithms were used to implement this block. The first consists of generating several uniformly distributed sequences. The outputs of these sequences are added together to form a Gaussian sequence. The second algorithm generates a Gaussian sequence directly.

In the first algorithm, the first step is to generate several uniformly distributed white random variables. This was done using the linear congruential method [36, p.9]:

$$Z(i+1) = [Mult \cdot Z(i) + Incr] \text{ Mod}(m) \quad (4.2)$$

where:

- Z is the generated number, the seed
- $Mult$ is a multiplier
- $Incr$ is an increment
- m is the modulus

Only several values for these parameter insure the uncorrelation and the uniformness of the sequence. In our case the choice of m is a very important factor. Any arbitrary value for m would require a division operation, which is time consuming. Whereas if m is chosen as a power of 2, the mod operation is reduced to a simple shift. Therefore, the following choices were made:

$$\text{for } m = 2^{16} \quad Mult = +25173 \quad Incr = +13849 \quad (4.3)$$

$$\text{for } m = 2^{31} \quad \text{Mult} = +314,159,269 \quad \text{Incr} = 453,806,245 \quad (4.4)$$

The first choice gives uniformly distributed numbers ranging between 0 and 65536, or if a two's complement notation is adopted, the numbers range between -32768 and +32767. This means that their variance will be [38, p.267]:

$$\sigma_{R_1}^2 = T^2/12 = 18918.6$$

The number sequence repeats after 2^{16} . Therefore, if a sampling rate of 8 KHz is used, the sequence repeats after about 8.19 seconds.

The second choice gives a considerably longer sequence. The sequence repeats after 2^{31} . Therefore, with a sampling rate of 8 KHz, the sequence repeats after about 74 hours. On the other hand, the computational requirements in this case are far more involved; double precision arithmetic has to be used.

In order to get a Gaussian distribution, several of these numbers have to be added. The proof of that is in the central limit theory [38, p.266]. If we add N independent, zero mean, uniformly distributed variables, having the same variance (σ_{R_1}), we get:

$$R = R_1 + R_2 + R_3 + \dots + R_N \quad (4.6)$$

$$\sigma_R^2 = \sigma_{R_1}^2 + \sigma_{R_2}^2 + \sigma_{R_3}^2 + \dots + \sigma_{R_N}^2 \quad (4.7)$$

$$= N\sigma_{R_1}^2 \quad (4.8)$$

If we divide by N we get:

$$X = R/N \quad (4.9)$$

$$\sigma_X^2 = \sigma_{R_1}^2/N^2 \quad (4.10)$$

$$= \sigma_{R_1}^2/N \quad (4.11)$$

All the previous computations are performed on a fixed point 16 bits processor; therefore, it is important to ensure that no overflows occur during the various operations. The division by N is needed to decrease the variance of the output, and consequently reduce the possibility of overflows.

The previous algorithm was used in the various noise generation blocks. The specific choice of the value of the parameters Mod and N was made for every block on an individual basis, depending on the rate at which this specific block is called as well as the system requirements. A compromise was made between the following factors: a minimum execution time, a distribution as close as possible to Gaussian, no overflows, and a full use of the dynamic range of the processor.

Figure 4.4 shows the amplitude distribution of the generated number sequence, compared to an exact Gaussian distribution.

The second algorithm for generating two independent Gaussian sequences is described in [37, p.258]. It can be summarized in the following steps:

Figure 4.4: Amplitude Distribution of the Noise Signal

1. Generate a uniformly distributed sequence [0 to +1], $U1$
2. Generate a uniformly distributed sequence [-1 to +1], $U2$
3. Calculate $Z1 = \sqrt{-2 \ln U1} \cdot \sin(\pi U2)$
4. Calculate $Z2 = \sqrt{-2 \ln U1} \cdot \cos(\pi U2)$

The sequences $Z1$ and $Z2$ are independent, zero mean, Gaussian sequences. A scaling operation has to be performed in order to ensure that no overflows occur, and at the same time to ensure that the numbers are big enough to maintain a reasonable dynamic range.

4.2.4 Multipath Propagation

The multipath propagation block is shown in figure 4.3. It includes two independent zero-mean, white Gaussian sequence generators. The output of each generator is fed to a shaping filter. The noise generation blocks were described in section 4.2.2. The frequency response of the filter used in our channel simulator is shown in figure 4.4.

The cutoff frequency of the shaping filter is at the maximum Doppler frequency. The Doppler frequency is dependent on the wavelength and the vehicle speed. Therefore, if a programmable channel simulator is desired, the shaping filter has to be programmable. If the desired frequency range is between 1 MHz and 1000 MHz, and the vehicle speed ranges between 0 and 120 km/hr, the Doppler frequency range will be between 0 and 111 Hz. This means that the shaping filter must have a cut off frequency ranging between 0 and 111 Hz with a resolution of at least one Hz. This is a difficult problem, specially that the passband of the filter has a special shape.

Figure 4.5: Frequency Response of the Shaping Filter

The approach used to solve this problem is based on varying the sampling rate of the filter and the Gaussian noise generator. The problem is, therefore, reduced to a problem of filter design using variable rate digital signal processing techniques. Increasing the sampling rate of a digital filter keeps the shape of the filter unchanged while expanding the frequency scale, hence increasing the cutoff frequency.

A programmable counter was used in the fading simulator to achieve the desired sampling rate. According to the desired vehicle speed and RF frequency, the Doppler frequency is calculated. The corresponding division factor of the sampling rate is also calculated and sent to a programmable counter. The output of the programmable counter is used as the clock of the multipath propagation block.

There is still a problem with the previous approach. The aliasing terms at the output of the D/A converter have to be cancelled. This is usually done using a simple analog filter. In this case, given the wide range of the cutoff frequency, a fixed filter cannot be used. A possible solution to this problem is to have a cascade of digital filters, each of them being sampled at a higher sampling rate. These sampling rates are proportional to the sampling rate of the original filter. This multirate digital signal processing solution is very costly in terms of the computational power required. A simpler solution was adopted using switched capacitor filters. These filters were used as a first stage anti-aliasing filter. The sampling rate of the switched capacitor filters was obtained from the same programmable counter used to generate the digital filter sampling rate. A simple analog filter was used to suppress the spikes occurring at the sampling frequency of the switched capacitor filter.

Figure 4.6 illustrates the aliasing problem in question. Figure 4.7 shows a block diagram of the hardware used to solve the problem. The block diagram includes

Figure 4.6: Aliasing Problem

Figure 4.7: Block Diagram of the Anti-Aliasing Filtering System

the digital signal processor, the programmable counter, the switched capacitor filters, and the analog filters.

4.2.5 Shadowing

The shadowing, or slow fading, was described in section 2.1.2. It has a lognormal distribution [24, p.22]. The shadowing rate is proportional to the speed of the mobile. It is also dependent on the RF frequency.

The shadowing block consists of a Gaussian noise generator followed by an antilog block. The output of the shadowing block multiplies both outputs of the multipath propagation block described in section 4.2.4.

The shadowing rate is at least one hundred times slower than the multipath fading process. If the sampling rate of the shadowing block is the same as the one for multipath fading, the filter required would be very complicated. It is therefore useful to use multirate digital signal processing and have the shadowing block sampled less frequently. This would save some computations. But, again, it causes an anti-aliasing problem. It is impossible in this case to solve the problem using switched capacitor filters. Both signals in this case will have to be processed digitally. It is therefore inefficient to convert the shadowing signal to the analog domain, filter it, and then reconvert it to the digital domain.

In order to solve this problem, a multirate digital signal processing approach was adopted. A cascade of digital filters was used, each filter being sampled at a higher sampling rate. This way the role of each filter is to cancel a few aliasing terms of the previous filter. All the sampling rates required were generated digitally inside the processor without any extra hardware. Figure 4.8 illustrates the problem and the solution approach.

Figure 4.8: Multirate Digital Signal Processing Approach to Solve Aliasing Problem

The shadowing level is changed by changing the variance of the of the output shadowing signal. This is done by changing the variance of the Gaussian noise generator. The variance of the generated signal is between 0 and 31 dB.

4.2.6 Line of Sight

The line of sight is simulated by adding a dc component to one of the quadrature outputs of the baseband section. The ratio of the dc component to the rms value of the generated signal represents the ratio of the line of sight received signal to the reflections and refractions received. In the implemented channel simulator, this ratio can be changed from 0 to 9.9 in increments of 0.1. This limitation was implied by our need. It is very simple to increase this range if there is a need for that. This could also be desired in the case of mobile satellite channel simulation.

There is a practical problem encountered in the design of the line of sight block. The dynamic range of the signal is limited by the number of bits in the D/A converter used at the output of the digital section. If a dc signal is to be added to one of the channels, this means that the dynamic range has to accommodate that. As an illustrative example, a line of sight ratio of 9 means that the dc value is nine times bigger than the rms value of the multipath multiplicative signal. This implies that less than one tenth of the dynamic range can be used for the multipath multiplicative signal. This restricts the performance of the system, specially if the other factors affecting the dynamic range are taken into consideration. These factors are due to the fact that the signal is a product of two functions, the first is Gaussian and the second is lognormal.

A simple solution was adopted. The multipath signal is generally scaled to occupy the whole dynamic range. When a line of sight component is to be added

the signal is rescaled such that the sum of the dc component and the maximum value of the signal never reaches the maximum allowable number. Therefore, both quadrature components of the multiplicative signal have to be multiplied by a factor k . This factor is calculated as follows:

1. Maximum allowable number : 32767
2. Maximum value of the incoming signal : 32767
3. Signal after adding the dc value : $v_{dc} + 32767$

This signal has to be multiplied by a factor k to insure that its maximum value never exceeds 32767. So,

$$32767 = k(v_{dc} + 32767) \quad (4.12)$$

$$k = 32767 / (v_{dc} + 32767) \quad (4.13)$$

Figure 4.9 illustrates the line of sight block, using this simple approach, which does not require any extra hardware. Its main disadvantage, though, is that the rms value of the multiplicative multipath signal varies with the line of sight ratio. This can be viewed as a change in the receiver gain. Another disadvantage is a decrease in the dynamic range of the multipath signal proportional to the line of sight ratio. When a small line of sight ratio is needed such as in the mobile environment, this problem is minor. Whereas, when very large line of sight ratios are required such as in the satellite case, this approach is not efficient.

Figure 4.9: Line of Sight Section

Another solution to the dynamic range problem is to add a dc value to the multiplicative baseband signal after the D/A converter. This is done by analog means. In this approach, an extra D/A converter has to be added as well as an analog adder. In this case another dynamic range problem may arise, the dynamic range of the RF mixer might imply some limitations.

A third solution can be adopted. It is possible to simulate the line of sight in the RF section. Figure 4.10 shows a block diagram of a possible system. The in-phase RF signal is added to the in-phase output signal after being multiplied by the dc value. This solution was rejected due to its cost and complexity.

It is important to notice that, if the need arises in the future for large line of sight ratios, one of the previous solutions will be implemented.

4.2.7 Baseband Channel model

The channel simulator described in this chapter generates two lowpass signals that multiply both quadrature components of an RF signal. The RF section requires costly components and very costly measurement equipments. It does not add any extra information when coding and baseband modulation techniques are tested. The implementation of the RF section is also straightforward and well known. It is possible to avoid the use of any RF components and to simulate the transmitter, the channel and the receiver at baseband frequencies.

Figure 4.11 shows a block diagram of the RF section of the transmitter, the mobile channel simulator, and the RF section of a quadrature receiver. The input T to the transmitter has two components, T_i and T_q . The output R of the receiver has also two components, R_i and R_q . This configuration is used in many systems. If T_i is a speech signal and T_q is a 90 degrees delayed version of T_i , we have a

Figure 4.10: Other Line of Sight Alternatives

Figure 4.11: Block Diagram of the Transmitter, Channel, and Receiver

single sideband system, with R , the received signal. The same configuration can be used for most quadrature modulation techniques such as M-ary QAM, M-ary PSK, or M-ary APK. The signals at the different points are:

$$(A) T_i \cos(\omega_c t)$$

$$(B) -T_i \sin(\omega_c t)$$

$$(C) T_i \cos(\omega_c t) - T_i \sin(\omega_c t) = \Re[T \exp(j\omega_c t)]$$

$$(M) -T_i \sin(\omega_c t) - T_i \cos(\omega_c t) = -\Im[T \exp(j\omega_c t)]$$

$$(D) C_i T_i \cos(\omega_c t) - C_i T_i \sin(\omega_c t)$$

$$(E) -C_i T_i \sin(\omega_c t) - C_i T_i \cos(\omega_c t)$$

$$(F) C_i T_i \cos(\omega_c t) - C_i T_i \sin(\omega_c t) - C_i T_i \sin(\omega_c t) - C_i T_i \cos(\omega_c t) = \Re[CT \exp(j\omega_c t)]$$

$$(G) C_i T_i + C_i T_i \cos(2\omega_c t) - C_i T_i \sin(2\omega_c t) - C_i T_i \sin(2\omega_c t) - C_i T_i - C_i T_i \cos(2\omega_c t)$$

$$(K) C_i T_i - C_i T_i = \Re[CT]$$

$$(H) -C_i T_i \sin(2\omega_c t) + C_i T_i - C_i T_i \cos(2\omega_c t) + C_i T_i - C_i T_i \cos(2\omega_c t) + C_i T_i \sin(2\omega_c t)$$

$$(L) C_i T_i + C_i T_i = \Im[CT]$$

Therefore the output signals are:

$$R_i = C_i T_i - C_i T_i + N_i = \Re[CT] + \Re[N] \quad (4.14)$$

$$R_q = C_q T_q + C_q T_q + N_q = \Im[CT] + \Im[N] \quad (4.15)$$

Therefore it is possible to express R as follows:

$$R = TC + N \quad (4.16)$$

where R, T, C, and N are complex numbers.

Equation 4.16 shows that it is possible to simulate the transmitter, the channel, and the receiver at baseband; there is no need to translate the signal to RF. The signals in this case are complex. Figure 4.12 shows a block diagram of a complex baseband channel simulator. The baseband multiplicative signal C is the same signal used in the RF channel simulator to modulate the RF signal. The different parameters of this signal were described in detail in section 4.2.4 in this chapter, and a block diagram of the baseband section was shown in figure 4.3.

The additive noise, in this case, is also generated at baseband, therefore, it is possible to generate the noise using digital signal processing techniques. Several algorithms for noise generation were described in section 4.2.3. The noise power can be varied to simulate different signal to noise ratios.

The baseband channel simulator shown in figure 4.12 was implemented using the TMS-320 digital signal processor. It was used, within the testing set-up shown in figure 1.1, to evaluate the various algorithms proposed in this thesis.

Figure 4.12: Baseband Channel Model

4.2.8 Results and conclusions

The mobile radio channel simulator described in this chapter was evaluated and proved to be very reliable. Most of the experiments performed on the data obtained from field measurements, [24], were repeated using the data from the simulator; a perfect agreement is demonstrated.

Three sets of experiments were conducted on the envelope of the received signal. The first set of experiments deals with the amplitude statistics, the second set deals with the power spectral density, and the third set deals with the autocovariance function.

Figure 4.13 shows the amplitude distribution of the envelope of the received signal; only multipath fading was applied, no line of sight or shadowing was added. According to the experimental results and the various models presented in section 4.2.3, this distribution is supposed to be Rayleigh. Figure 4.13 also shows the best fitting Rayleigh curve. It is clear that both curves are perfectly similar. The field measurements described in [24] show that the distribution becomes Rician when a line of sight component is present. Figure 4.14 shows a number of distribution curves obtained from the simulator with different line of sight ratios. Comparing these curves to the one obtained from the field tests [24], it is possible to conclude that the line of sight ratio in these field tests is about 1 to 1.5.

Figure 4.15 shows the power spectral density of the envelope of the received signal. Again, several line of sight ratios were tried. When no line of sight is present, the spectral density has a large peak at double the Doppler frequency. Another small peak is present at the Doppler frequency. When the line of sight ratio is increased, the small peak, at the Doppler frequency tends to increase whereas the other peak at double the Doppler frequency decreases. Again,

Figure 4.13: Amplitude Distribution of the Faded Envelope (LSR=0)

and Best Fitting Rayleigh Distribution (LSR: Line of Sight Ratio)

Figure 4.14: Amplitude Distribution of the Faded Envelope (Various LSRs)

LSR: Line of Sight Ratio

Figure 4.15: Power Spectral Density of the Faded Envelope

Figure 4.16: Autocovariance Function of the Faded Envelope

Line of Sight Ratio = 1.5

comparing these results to the ones obtained in [24] it is possible to confirm the results of the the first set of experiments.

The third set of experiments deals with the autocovariance function of the envelope of the received signal. The same experiments described in [24] were repeated on the simulator's data instead of the field measurements. The same set-up and the same programs were used. Figure 4.16 shows the results. The solid and the dotted lines represent the field tests obtained from different areas in the Ottawa-Hull region, and the crosses represent the simulator's autocovariance function. In this specific plot, the line of sight ratio in the simulator was set to 1.0.

From the previous experiments, it is clear that the results obtained from the mobile radio channel simulator described in this chapter are very similar to the results obtained from the field measurements [24]. The simulator is therefore, reliable and reflects the actual conditions.

Chapter 5

Background Acoustic Noise Reduction

Introduction

Background acoustic noise represents one of the major sources of speech corruption in mobile communications. This problem is more serious in the newly proposed hand-free mobile telephone systems.

The background acoustic noise in cars comes from different sources: the engine's acoustic noise, the air flow noise, street noise, and internal man made noise. These various sources are examined in section 5.1.

Some experiments have been conducted in order to characterize the background noise inside the car cabin, and to determine the correlation between the outside and the inside noises. The experiments and their results are briefly presented in section 5.2.

It is possible to filter out the acoustic background noise from the transmitted speech, and thus obtain a better signal to noise ratio at the transmitter. Some of these methods are evaluated in section 5.3. All the methods described use adaptive filtering techniques. The various filter structures and adaptation algorithms appropriate to these filters are discussed.

Finally, in section 5.4, some results and conclusions are presented.

5.1 Sources of Background Acoustic Noise in Cars

Introduction

The background acoustic noise in cars can be classified into two main categories: noise generated outside the cabin and noise generated inside the cabin. In each case, the noise can be either localized or distributed. This means that the source of the noise can be either known and static or can be dynamic.

5.1.1 Noise from Outside-Cabin Origin

The first source of outside noise is generated within the engine compartment. The noise in this case is due to mechanical friction and due to air flow. Some examples of the sources for mechanical friction noises are the engine pistons, the belts, the generator, the air conditioning compressor, and the various electric motors such as the ventilation motor and the wipers motor. These different noises are also characterized by being very localized and very slow varying. The second class of noise generated within the engine compartment is the one due to air flow. The noise in this case is characterized by being distributed. It does not have a precise

localized source.

The second source of noise generated outside the cabin is from the interaction of the car with the outside. Again, in this case also the noise can be either localized or distributed. The friction of the tires with the road is an example of the localized origin. The air flow noise, on the other hand, depends on a variety of factors: the shape of the car and its speed, the direction of the wind and its speed, the leaks in the windows and doors, and many other factors.

The third source of noise generated outside the cabin is due to external sources, such as street noise.

5.1.2 Noise from Inside-Cabin Origin

Apart from the noise penetrating to the cabin from outside, there is some noise generated inside the cabin. This background noise is mainly due to vibrations, air flow and man-made noises due to the audio system or due to other passengers.

5.1.3 Conclusion

From the above discussion it is clear that most of the background acoustic noise inside a car cabin is generated outside the car and penetrates to the inside. There is therefore a high correlation between the outside noise and the inside noise. This correlation is manifested as a transfer function of the car with respect to acoustic noise penetration. It is also concluded that most of the noise has a localized origin, and is very slow varying.

The estimation of the acoustic transfer function of the car is a very important factor in determining the feasibility of any noise cancelling scheme. It is important

to notice that this transfer function varies from one car to the other. It also varies for the same car depending on the origin of the noise, the status of the ventilation system and the windows, the speed of the car, and many other factors.

5.2 Experimental Results

Several experiments have been conducted in order to analyse and characterize the background acoustic noise in cars. Several car models have been considered. The results varied from one model to the other, but a general pattern was found. The results presented in this section were obtained from a typical mid-size north American car.

The first experiment was to measure the noise spectral density inside and outside the car. Figures 5.1 and 5.2 show the power spectrum of the engine noise, outside and inside the cabin, while the car is not moving. The figure shows that the car acts as a low-pass filter with cutoff frequency around 250 Hz. Noises due to vibrations are also concentrated at very low frequencies.

The second set of experiments was to measure the transfer function of the car, or the impulse response of the car with respect to acoustic noise penetration. A gun shot was fired outside the car and the resulting noise inside the car was recorded. Figure 5.3 shows the recorded noise inside the car. The experiment was conducted in an open area to minimize as much as possible the effects of reflections of the gun shot pulse from the surrounding walls. It was repeated several times from various angles. The length of the pulse response is due to reflections within the cabin. Therefore, it was important to conduct a third set of experiments to determine the reflections within the cabin.

Figure 5.1: Noise Power Spectrum Outside The Cabin

2 of /de 2

MET

Ampl.

Transmit Midband Pilot Tone

Ampl.

Transmit Speech

Send TAB Reduced Carrier

Figure 6.2: Typical Speech Transmission Session (Frequency Domain)

Using the ST Concept (Speech-Only Mode)

Figure 5.3: Impulse Transfer Function of the Car

Figure 5.4: Impulse Response of the Cabin

The third set of experiments was to measure the reflections within the cabin of the car. An impulse is generated inside the cabin and the corresponding noise is recorded. This experiment helps in determining many parameters such as the reverberation time and the cancellation depth required for a specific cancellation time. These parameters will be used in determining the length of the noise cancelling filter described in the next sections. They also determine the need for echo cancelling. Figure 5.4 shows the impulse response of the cabin. The reverberation times were found to be in the order of 25 ms.

5.3 Background Acoustic Noise Reduction

5.3.1 Approaches for Background Noise Reduction

There are three different approaches for reducing the background acoustic noise in speech signals. The first approach is to use a proper directional microphone. The second approach is to take advantage of the quasi periodic nature of the speech signal to form an estimate of the clear speech. The third approach is to estimate the noise corrupting the speech and subtract that estimate from the original corrupted signal. In this section each of these approaches is presented and evaluated with respect to the mobile environment.

The first approach deals with the type of microphone used. The more the microphone is directional, the less it is going to pick up background noise. This solution is a very simple one, and could be used in conjunction with the other methods described later in this section. However, this is not a radical solution in noisy environments, specially in the cases where the microphone has to be far from the speaker.

The second approach is to take advantage of the quasi periodic nature of the speech signal. An estimate of the clean speech signal at time t is formed from the value of the signal at time t and the estimated pitch period. Reference [52] indicates that for additive white noise distortion, the signal to noise ratio was increased by 7 dB in a 0 dB environment using this approach. Many other single microphone approaches for reducing the background noise in speech signals have been developed [53]. However, these methods become ineffective when the noise power is equal to or greater than the signal power or when the noise spectral characteristics change rapidly with time [58].

The third approach assumes the existence of two input sources: a primary input and a reference input. Figure 5.5 illustrates this approach. The primary input consists of a speech signal plus noise ($S+N$). The reference signal consists of noise alone (N_1). If the reference noise (N_1) is highly correlated with the noise on the primary input (N), it is possible to design an adaptive filter to estimate the primary noise. A noise-free speech signal can, therefore, be obtained by subtracting the estimated primary noise from the primary input. Reference [51] includes a detailed analysis of the general concept and theory of adaptive noise cancelling.

5.3.2 Adaptive Noise Reduction Using Two Microphones

In this thesis, the third approach for background noise reduction, described in the previous section, is adopted. This two microphone approach, shown in figure 5.5, is investigated for hand-free mobile telephony. The primary input is connected to a directional microphone in front of the speaker. The secondary input is connected to a microphone located outside the cabin of the car, close to the engine compartment. That way ensures that the secondary input is a speech-free, pure noise signal.

Figure 5.5: Adaptive Noise Cancellation Concept

The output signal of the noise cancelling system can be written as:

$$Z = S + N - Y \quad (5.1)$$

Squaring:

$$Z^2 = S^2 + (N - Y)^2 + 2S(N - Y) \quad (5.2)$$

Taking the expectations of both sides:

$$E[Z^2] = E[S^2] + E[(N - Y)^2] + 2E[S(N - Y)] \quad (5.3)$$

But the signal S is uncorrelated with both N and Y :

$$E[Z^2] = E[S^2] + E[(N - Y)^2] \quad (5.4)$$

The signal power is not affected as the filter is adapting. Accordingly, the minimum output power is:

$$\min E[Z^2] = E[S^2] + \min E[(N - Y)^2] \quad (5.5)$$

Therefore, minimizing $E[Z^2]$ results in $E[(N - Y)^2]$ being also minimized.

The most straightforward implementation of the previous algorithm is to use a finite impulse response (FIR) transversal filter, and to use the stochastic gradient algorithm to adapt the coefficients. This algorithm, commonly known as the least mean square (LMS) algorithm is described in [59]. A 128 taps transversal filter

was implemented and tested for noise cancelling in the mobile environment. A 10dB noise reduction was achieved.

5.4 Results and Design Considerations

In this section, the various factors affecting the performance of the noise reduction scheme are evaluated. The location of the reference microphone, the filter length, and the filter configuration are discussed.

The first major design factor is the location of the reference microphone. This location implies the required number of taps for the adaptive filter, which affects its settling time. The possibility of locating the reference microphone inside the cabin, under the dash board, was also examined. In this case, the number of taps can be reduced; however, if some speech is present at the reference microphone, the filter may tend to cancel the speech instead of the noise. In order to avoid this problem, the adaptation process has to stop while the speech is present. This is achieved by having a speech detector and by updating the filter coefficients only during the silence intervals. References [19, 20] discuss the behaviour of speech detectors in the mobile environment. The possibility of having multiple reference microphones is another design factor that could enhance the performance of the system; this factor, however, was not investigated in this thesis.

Another design consideration is the filter causality. The performance of the adaptive noise reduction system may deteriorate if the noise reaches the primary microphone before reaching the reference microphone. This requires the use of a noncausal filter. Noncausal adaptive filters are easily generated by delaying the primary input. Long delays may be annoying to the users and may require some extra hardware.

The number of filter taps required in noise cancelling is generally large. This is due to the multiple reflections, or echoes encountered. In a typical room, with the primary and reference microphones 12 ft apart, 1500 taps are required [58]. In a fighter cockpit environment, with the reference microphone attached to the outside of the pilot's oxygen facemask, only 100 taps are required [57]. In our application of noise reduction in mobile telephony, with the car stopped and the engine on, a 10 dB was achieved using a 128 taps filter. This relatively small number of taps is due to two factors. The first is the small distance between the two microphones (about 2 ft). The second factor is the sound absorbing materials inside the cabin which reduces considerably the echos. The effect of the sound absorbing materials is clear from figure 5.4.

A longer filter is required to improve the performance of the system. There are many problems associated with long adaptive filters such as implementation difficulties, large settling time, and misadjustment problems. The first problem is related to the implementation aspects. Large filters usually require a multi-processor system with its associated handshaking and extra hardware. This in turn increases the complexity and the cost, and decreases the reliability. The second problem is the settling time which increases as the number of taps increases. The third problem is the misadjustment, or the ratio of excess to minimum mean square error, which increases as the number of taps increases. Large misadjustment manifests itself as pronounced echo in the speech waveform. This echo is removed by reducing the adaptation step, which increases the settling time [58]. A compromise has to be reached.

The function of the adaptive filter is to simulate the path between the reference microphone and the primary microphone. The transversal FIR filter described previously models that path as an all zero model. Many pole-zero models have

been recently proposed for modelling acoustic signal paths, and have been used in applications such as acoustic echo cancellers [61]. The advantage of using a pole-zero model is a significant reduction in the number of taps required, this number is nearly $1/5$ the number of taps in a conventional transversal filter [62]. Pole-zero adaptive filters are, however, less stable and more susceptible to truncation errors.

The adaptation algorithm considered for the adaptive filter is the stochastic gradient algorithm. It was chosen for its simplicity and robustness. Other adaptation algorithms such as the recursive least squares algorithm and the fast Kalman algorithm were rejected, despite their fast adaptation, due to their complexity.

5.5 Conclusions

It is possible to reduce the background acoustic noise in the transmitted speech in mobile telephony, and thus increase the signal to noise ratio at the transmitter. This can be achieved using adaptive filtering techniques. The approach proposed is to use two microphones. The primary one captures the speech signal plus some background noise. The second microphone is located so that it captures a speech-free noise signal. This noise signal is fed to the adaptive filter which estimates the noise at the primary microphone. A noise-free speech signal is then obtained by subtracting the estimated primary noise from the primary input.

The filter proposed for this application is an adaptive FIR transversal filter, using the stochastic gradient algorithm for adaptation. A noise reduction of 10 dB was achieved using a 128 taps filter. The system performance depends on several factors such as the car model, the status of the windows, and the wind friction. The reason for that is that the filter is not capable of cancelling the various noise sources equally. The longer the filter, the better is its performance regarding the

noise sources with a static origin. On the other hand, a long filter implies a larger adaptation time, which deteriorates the system performance regarding the noise sources with dynamic origin such as the airflow noise. Faster algorithms are not yet feasible for practical implementations. IIR filters seem to be very attractive for this application; however, their stability problems and high complexity represent a major obstacle.

Chapter 6

Switched Tone Concept and Associated Subsystems

Introduction

Speech transmission over mobile channels has been dominated by a single modulation technique, the analog frequency modulation (FM) [1-3, 7, 21]. In analog FM, a speech signal, bandlimited to 3 kHz, is transmitted using a 20 to 30 kHz channel spacing [1, 3, 6, 7, 15, 21], whereas, the same signal could be transmitted with a 5 kHz channel spacing using modified single sideband (SSB) techniques [7, 10, 11, 15, 60, 78]. There are, however, some problems associated with the modified SSB techniques, which limit their wide commercial use [8, 10-15].

In this chapter, a modified SSB transmission scheme, the switched tone (ST) concept, is proposed for speech and data transmission over mobile channels.

Section 6.1 is a general description of the switched tone concept. The main sub-

Figure 6.1: Typical Speech Transmission Session (Time Domain)

Using the ST Concept

band; it is used by the receiver for automatic gain control (AGC) and automatic frequency control (AFC). In order to reduce the effect of noise, speech companding has to be used. In dynamic speech companding, as described in section 6.2, some information needs to be transmitted along with the speech signal; this companding information is modulated on the tone above the speech band.

A data transmission session starts, also, with a midband pilot tone, followed by a 'start data' message to switch the receiver into data mode. The modulation technique used for data transmission was not investigated in this thesis. However, several techniques, such as M-ary phase shift keying (PSK) and differential PSK (DPSK) have been proposed for similar situations [14, 15, 64, 65, 73]; any of these tone-calibrated techniques could be used within the ST concept. If the data transmission scheme used requires a tone above band [73], it is suggested to send the 'start speech' and 'start data' messages on the tone above band rather than on the midband tone, in order to simplify the design of the receiver.

In a speech only system, it is possible to simplify the protocol used for initial acquisition, and, therefore, simplify the design of both the transmitter and the receiver. In that case, the 'start speech' message can be omitted, and a speech session is reduced to a midband pilot tone followed by speech transmission using SSB with a tone above the speech band. Figure 6.2 shows a typical speech transmission session using the speech only switched tone concept.

The ST concept overcomes most of the deficiencies encountered in the other modified SSB transmission schemes, shown in figure 3.1. The midband pilot tone offers a symmetric pull-in range for the initial frequency acquisition phase; therefore, it solves a problem frequently encountered in conventional tone above band (TAB) systems. In a TAB system, the pilot tone is always at the edge of the band; an initial frequency offset could, therefore, be such that the tone is out of

Ampl.

Transmit Midband Pilot Tone

Ampl.

Transmit Speech
Send TAB Reduced Carrier

Figure 6.2: Typical Speech Transmission Session (Frequency Domain)

Using the ST Concept (Speech-Only Mode)

the IF filter passband range. In this case the receiver can never synchronize to the transmitter. The ST concept also avoids the problem of distorting the speech signal encountered in the transparent-tone-in-band (TTIB) systems, where the spectrum of the audio signal is split and shifted to create a null in the middle for pilot insertion.

6.2 Speech Companding

Introduction

When a speech signal is processed and transmitted through a channel, it exhibits many forms of degradation, and, in general, the various signal components are not affected in the same way. In radio transmission using frequency modulation (FM), for example, the signal exhibits a frequency dependent noise degradation, whereas, in radio transmission using amplitude modulation (AM), the signal exhibits an amplitude independent noise degradation. The effect of having a particular component of the speech signal severely degraded is, usually, a deterioration in the global speech quality and intelligibility. It is, therefore, necessary to protect some of the signal components more than others, depending on the particular application. This process is called speech companding. The basic concept of speech companding is, therefore, to modify a particular component of the speech signal at the transmitter, and, to restore the signal back to its original state at the receiver; this is done in order to protect this particular component, which is affected the most by the channel degradation. The different possible classes of speech companding are presented in section 6.2.1, and a detailed analysis of the amplitude based amplitude companding is carried in section 6.2.2. A speech companding system, appropriate to speech transmission over mobile channels using the switched

tone concept, is, then, evaluated in the subsequent sections.

6.2.1 Different Companding Classes

It is possible to modify the frequency location, or the amplitude of a particular component of the speech signal, based on either its location in the frequency spectrum, or its amplitude level. The different companding classes are, therefore, frequency based frequency companding, frequency based amplitude companding, amplitude based frequency companding, and amplitude based amplitude companding. These four basic classes of speech companding will be presented in this section.

The first class of companding is the frequency companding based on the frequency location. In this class, the system takes advantage of the tendency in human speech for either low sounds (vowels), or high sounds (consonants), to be used, but never both at the same time. The system folds the higher frequencies down over the lower frequencies and transmits both simultaneously. Expansion at the receiver could give a high quality signal; that has been transmitted in about 60 percent of the bandwidth normally required. Nevertheless, this frequency based frequency companding is very complicated and its results, so far, are not very promising [7]. Therefore, it will not be considered in this work.

The second class of companding is the amplitude companding based on the frequency location. This is necessary when a signal exhibits a frequency dependent noise degradation. In the output of an FM demodulator, for example, the noise power-spectral density rises parabolically with frequency [74]. Taking into account that most of the speech energy is concentrated at low frequencies, it is clear that the noise power is largest when the signal is smallest. This results in a rather

inefficient communication system. To remedy this situation, the high frequency components of the input signal are emphasized at the transmitter. The inverse operation is performed at the output of the demodulator to deemphasize the high frequency components. The signal is, therefore, restored to its original state. Its high frequency components are less susceptible to noise, which improves the global signal to noise ratio.

Another example of this second class of companding is the Dolby system used in the audio tape recording area. When a signal is transmitted from the recording head to the tape, and back from the tape to the reading head, it exhibits a frequency dependent noise degradation. The Dolby system is used to remedy this situation. In the Dolby-A, for example, the audio spectrum is divided into four frequency bands, and a boost of 10 to 15 dB is applied to individual bands when their signal level falls below a preset level. The net signal to noise ratio is increased by 10 to 15 dB. Several other systems such as Dolby-B, Dolby-C, and DBX are commercially used.

The third class of companding is the amplitude companding based on the amplitude level of a particular syllable. This class of companding is particularly useful for signals, with large dynamic range, exhibiting a level-independent noise degradation, such as the case of speech transmission using amplitude modulation.

The dynamic range of a speech signal is very large; it can reach 70 dB. This large dynamic range is due to several factors: the different speakers, the distance from the microphone, and the volume variations of the different speech syllables. An automatic gain control system is usually used to compensate for the slow variations in the signal strength, which takes care of the variations due to the different speakers and the distance from the microphone. The actual dynamic range of the speech, in this case, is reduced to approximately 40 dB.

In most amplitude modulation systems, the thermal noise is directly added to the recovered baseband signal, and, therefore, the low volume syllables are severely affected; the noise power is the same, while the signal power is minimum. The signal to noise ratio is, therefore, very low during the low volume syllables and during the silence intervals, which reduces considerably the speech quality and the speech intelligibility. In order to remedy this situation, it is necessary to expand the low volume syllables at the transmitter, and compress them at the receiver. Most amplitude modulation transmission systems, such as AM and SSB, exhibit amplitude independent noise degradation, and, therefore, in order to achieve a high quality transmission using amplitude modulation, the use of amplitude based amplitude companding becomes essential.

The reduced dynamic range of the speech signal, achieved through amplitude companding, has some additional positive effects on the whole transmission system. The performance of the transmitting high power amplifier is enhanced with respect to saturation. The ratio of the maximum power to the average power required by the system is reduced, which enhances the power efficiency.

The amplitude companding based on the amplitude level will be investigated in more detail in the next sections. It will be termed amplitude companding for simplicity. Different techniques used for speech companding are presented in section 6.2.2. Many design parameters such as the choice of the companding interval and the companding gain, the coding algorithm used and the synchronization between the transmitter and the receiver are discussed in the subsequent sections.

Finally, the fourth class of companding is frequency companding based on the amplitude of the signal. This type of companding is not generally used; it does not seem to have any practical use.

6.2.2 Various Techniques for Amplitude Companding

Speech companding is proposed in this thesis, in order to improve the speech quality for SSB transmission over mobile channels. From the four classes of companding, presented in the previous section, the most appropriate one for our specific application in the third class, the amplitude based amplitude companding. The various techniques used for performing amplitude companding, the fixed and the dynamic companding, are discussed in this section.

Fixed compandors are presently used in many commercial applications, such as amplitude companded single side band (ACSB) systems [7]. A fixed compandor, as the name shows, performs a fixed ratio companding to the input signal. The reverse operation is performed at the receiver. No extra information is transmitted, and, therefore, there is no penalty paid in terms of bandwidth or power. Typical compandors give a 1 dB output variation for every 2 dB of input variation. This can usually lead a 15 dB signal to noise improvement with no change in bandwidth [7]:

Dynamic compandors can achieve a better performance. In dynamic companding the signal is observed for a period (T). An estimate is made for the optimum companding gain required for this particular interval. This companding gain is then applied to the signal and is digitally transmitted along with it. In the switched tone concept, the companding gain is transmitted, using digital modulation, on the pilot tone above the speech band. The companding gain, therefore, occupies a portion of the spectrum; this is the price paid for improving companding. There are many parameters to be evaluated in a dynamic companding algorithm. The choice of the companding interval (T), the choice of the companding gain and its coding algorithm, the synchronization between the compression and the expansion processes, the effect of errors and deep fades, and the effect of companding on the

speech spectrum will be discussed in the subsequent sections.

The fixed amplitude companding process have been known for a long time. It has been tested extensively in many experimental and practical systems [4-15]. Therefore, it is considered a solved problem and will not be investigated any further within this work. Dynamic amplitude companding, on the other hand, is relatively new; it has not been investigated enough, especially in the mobile environment. The interaction between some of its various parameters is still unknown, and needs to be tested and evaluated. These various aspects will be investigated in this thesis. Nevertheless, the restriction of the research to the dynamic companding does not imply that the dynamic companding is more appropriate than the fixed one in all cases.

6.2.3 Choice of the Companding Interval

Dynamic companding can be seen as extracting some information from the speech signal and sending it separately, using digital modulation. Speech transmission, in this case, can be seen as a hybrid transmission scheme, combining analog and digital transmission. The companding interval (T) is inversely proportional to the digital information extracted from the signal. As an extreme case, if the companding gain is updated every sample and is allowed to take any value, all the speech information would be buried in the companding gain and the signal itself will contain no information at all. The transmission, in this case, becomes purely digital. The companding interval, therefore, represents the ratio of the analog to digital portions of the transmitted signal. A compromise has to be made in order to achieve the best performance, while maintaining a good spectrum efficiency.

It is more convenient to view the companding interval in a different way. Given

that the sampling rate of the speech signal is fixed to 8 kHz (sampling time = 0.125 msec), changing the companding interval means changing the number of speech samples used by the companding algorithm for the gain calculation. A companding interval of 1.25 msec, for example, means that the companding gain is updated every 10 speech samples.

A speech companding scheme should estimate and track the syllabic volume change. An optimum compandor should, therefore, update its gain at every syllable, which implies a variable companding interval. There are two problems associated with this approach; the first is the synchronization between the compression interval and the actual speech syllables, which is a very difficult task, and the second is the synchronization between the transmitter and the receiver. This synchronization implies a higher transmission rate, which affects the spectrum efficiency of the whole system. Moreover, the computation complexity of the companding gain algorithm is also increased; the gain should be updated based on a different number of samples every time. Accordingly, it was concluded that, although this approach might lead to optimum companding, its extreme complexity and vulnerability to errors make it inappropriate for our specific application. Therefore, the companding was tested only for fixed intervals.

The choice of the fixed companding interval is governed by several factors. A short interval improves the performance of the compandor; it requires, however, a large bandwidth to be transmitted, which affects the main advantage of the ST concept, the channel bandwidth efficiency. In the ST concept, the channel spacing is limited to 5 kHz, which implies a lower limit on the companding interval. Given that the speech signal occupies 3 kHz, and taking into account the various guardbands, the bandwidth of the companding information should be limited to about 0.8 kHz. The corresponding companding interval depends on the number

of bits to be transmitted at every interval, as well as the modulation technique used, which will be discussed in the subsequent sections. If a single bit is to be transmitted every time, the minimum companding interval becomes about 2.5 msec, which corresponds to 20 speech samples. A large companding interval, on the other hand, degrades the performance of the compandor and increases the processing delay.

Several subjective tests have been conducted in order to evaluate the effect of the companding interval, as well as other parameters such as the companding incremental gain. The results will be presented in section 6.2.7. As far as the companding interval is concerned, it was surprisingly observed that increasing it up to 12 msec is not noticeable by the listener. The effect of the delay introduced, however, has to be evaluated with the other delays introduced by the system. This result implies that the companding interval should be maximized in order to increase the guardbands, and, therefore, reduce the adjacent channel interference. The upper bound should be dictated by the maximum tolerable delay introduced by the transmission system as a whole.

6.2.4 Companding Gain Setting Approaches

The second parameter, to be evaluated in dynamic speech companding, is the companding gain. The incoming signal is observed during one companding interval, and, according to a statistical analysis, the value of the companding gain is calculated. If every speech interval is treated independently, the companding setting algorithm is absolute, otherwise, it is relative. Several gain setting approaches are discussed in this section. The differential coding algorithm, proposed for the ST concept, is described in more detail in the next section, 6.2.5.

The first gain setting approach presented in this section is an absolute one. Every speech interval is stored, and its maximum absolute value is detected. The companding gain is maximized, so that, when multiplying the stored speech samples, no overflows occur. In this approach, the absolute value of the companding gain has to be transmitted along with every speech interval. Assuming that the companding interval is 10 msec and that the absolute gain is an 8 bit word, an 800 bits/sec transmission rate will be required, which is not an acceptable bit rate; the spectrum required by the modulated pilot will reduce the overall spectrum efficiency, which is our main objective.

Another approach for gain setting would be to limit the gain to few levels. Assuming that the input speech signal has a dynamic range of 48 dB, 16 gain levels 3 dB apart could be allowed. This way the number of bits needed to transmit each gain level would be four bits. This reduces the bit rate required to 400 bits/sec, which is still high for a good spectrum efficiency.

This bit rate can be reduced further by exploiting some of the speech characteristics. It was noticed that the speech rarely changes by more than 3 dB within the companding period. Therefore, it is possible to transmit the differential gain instead of the absolute one. In this case, only one bit needs to be transmitted every companding interval. Therefore, the bit rate required could drop to 100 bits/sec. A detailed analysis of the differential gain setting algorithm is carried in section 6.2.5.

The other problem is the algorithm used to calculate this gain setting. A statistical parameter of the samples within a companding period has to be observed and accordingly the gain has to be adjusted. It is proposed here to base our decision on the maximum value of the incoming samples. There are two reasons for this choice; the first is based on the companding objective in getting the maxi-

mum possible level while avoiding overflows and saturation, and the second is the computational complexity. Therefore, in this work, the companding gain selection will be based on the maximum absolute value of the incoming samples within one interval.

It is, therefore, concluded that the most appropriate gain setting approach, for speech companding within the switched tone concept, is the differential one. The most appropriate statistical parameter, according to which the companding gain is calculated, is the maximum absolute value of the speech samples within a companding interval.

6.2.5 Differential Coding Algorithm of the Companding Gain

Two differential coding algorithms of the companding gain are discussed in this section. An evaluation of each of them is presented.

Figure 6.2 (A) shows a flow chart of the first coding algorithm. The gain is assumed, for simplicity, to be only a multiple of two. In general, it is possible to view the gain as being a set of discrete levels. When a speech interval is to be companded, the gain is always set to the next higher level, unless an overflow occurs; in this case, the gain is reduced to the next lower level. The gain is, therefore, never the same for two consecutive speech intervals. Figure 6.2 (B) shows the corresponding decoding algorithm. If a one is received, the gain is increased by one level, and if a zero is received, the gain is reduced by one level. Figure 6.4 shows an example of a received pattern and the corresponding gain.

The first coding algorithm is very simple, and its response is very fast. However, a problem arises when the gain is to be constant for several consecutive

Figure 6.3: First Differential Companding Coding Algorithm
(A)

Figure 6.3: First Differential Companding Decoding Algorithm
(B)

systems required in an ST-based system, the speech companding, the automatic gain control (AGC), and the automatic frequency control (AFC), are described and evaluated in sections 6.2, 6.3, and 6.4 respectively. Finally, section 6.5 includes the conclusions.

A switched tone system also requires a speech detection subsystem. Speech detectors are not explicitly analysed in this work. A description of speech detectors in the mobile environment is found in [19, 20]. It is also possible to improve the signal to noise ratio at the transmitter by using a background noise reduction filter. Background acoustic noise reduction was described in chapter 5.

6.1 General description

The switched tone (ST) concept, proposed in this thesis, is a novel concept for speech and data transmission over mobile channels. The speech is processed on a talkspurt by talkspurt basis. Data can be transmitted, intermittently, between talkspurts, or continuously, when no speech is present. In the speech transmission mode, the ST concept uses a modified single sideband (SSB) transmission scheme. It overcomes most of the deficiencies of the other modified SSB schemes. These various SSB systems were presented and evaluated in section 3.2.2, and figure 3.1.

Figure 6.1 shows a typical speech transmission session using the ST concept. When the speech starts, a control signal is activated by a speech detector, indicating that a speech transmission session is to take place. A pilot tone is transmitted at the center of the allocated band. This midband tone is used by the receiver for initial frequency acquisition. The midband pilot tone is followed by a 'start speech' message, used to switch the receiver into speech mode. Speech is then transmitted using SSB modulation. A pilot tone is transmitted above the speech

Figure 6.4: Companding Example using First Algorithm

intervals. The only way to achieve a nearly constant gain is to send a pattern of 101010... . In this case, the gain fluctuates between two levels, which generates some background noise.

In order to avoid the previous problem, another algorithm is proposed. In this second algorithm, the gain could be stabilised, if this is needed. The algorithm is more complex and its tracking capability is inferior to the first one. In this algorithm the gain is increased to the higher level when two consecutive ones are received. It is decreased by one level if two consecutive zeros are received. Otherwise, the gain remains unchanged. Figure 6.5 shows a flow chart of the coding and decoding algorithms, and figure 6.6 shows a companding example using this second algorithm.

The main parameters to be evaluated in the previous coding and decoding algorithms are the incremental gain value and the number of gain levels to be allowed. The larger the incremental gain value, the better is the tracking performance of the companding algorithm and its ability to regain synchronization. A large incremental gain, on the other hand, reduces the precision of the companding algorithm. Assume, for example, that the incremental gain is set to the ratio 2; in this case the next gain level is not selected unless the maximum speech sample within the sampling interval is less than one half of the maximum allowable value. This, in turns, reduces the ability of the speech companding system to improve the signal to noise ratio.

The second coding-decoding algorithm shown in figure 6.5 was implemented using the testing set-up described in section 1.2. Several incremental gain values, and several number of levels have been tried. With the companding interval fixed to 5 msec, it was noticed that incremental gain values less than 1.5 dB (ratio: 1.2) limit the tracking ability of the system; the speech quality in this case

Figure 6.5: Second Differential Companding Coding Algorithm

(A)

Figure 6.5: Second Differential Companding Decoding Algorithm

(B)

Figure 6.6: Companding Example using Second Algorithm

becomes unacceptable. It was also observed that when the incremental gain value is increased to more than 6 dB (ratio: 2.0), the system starts to become inefficient in noise reduction. The number of gain levels is dependent on the incremental gain value. When the incremental gain value is 6 dB, for example, 10 gain levels are sufficient to cover a 60 dB dynamic range, whereas, when the incremental gain value is 2 dB, 10 gain levels cover only 20 dB. The best performance was achieved when the incremental gain value was between 2 and 5 dB, and when the number of discrete gain level was between 16 and 32.

The selected value of the incremental gain and the number of levels is also tied to other parameters in the system, such as the companding interval. The interaction between the various parameters and their effect on the system performance will be discussed in section 6.2.7.

6.2.6 Synchronization Between Transmitter and Receiver

Another problem which needs to be addressed is how to maintain the synchronization between the transmitter and the receiver. In this section, several synchronization schemes are proposed. Many related problems such as the effect of errors and deep fades are also discussed.

In the speech transmission scheme proposed in this work, the voice is transmitted burst by burst. This was proposed in order to accommodate data transmission between speech bursts. A start message is sent at the beginning of every speech burst. This message informs the different subsystems of the receiver that a speech signal will be transmitted. The receiver then goes into the speech mode and the AGC-AFC subsystems are initialized. It is possible to use this start message to initialize the companding gain to some preset midrange level. It is also possible

for the transmitter to send the absolute companding level after this start message. These two approaches are briefly discussed in this section.

Setting the companding gain at some preset level at the start of every speech burst is a very simple solution. If the companding interval is relatively small, the tracking capability of the system is high, therefore, the system performs well. However, if the companding period is relatively large this tracking performance will be slow. Therefore, the companding algorithm will not be optimum for the the first portion of the speech burst. On the other hand, this scheme is very simple and does not impose any extra function at the transmitter or the receiver.

The other approach is to send the absolute companding gain at the beginning of every speech burst. This will complicate both the transmitter and the receiver. Its benefit will be apparent in the first portion of every speech burst, specially when the companding period is large. The subjective tests show that the speech quality obtained by the first approach is satisfactory. Therefore, the extra complexity involved with the second approach is not recommended.

It is possible to simplify the system further and, at the same time, obtain a better performance when errors occur. This is achieved by slightly modifying the differential companding algorithm shown in figure 6.5. If an upper and a lower bound are imposed in the receiver, the system becomes adaptive. Suppose that, for some reason, the gain at the receiver is off by several levels. As soon as a silence period occurs, the transmitter gain level will go to its maximum value, and a one will be transmitted continuously. If the receiver had a higher gain level at the start, it will remain at its upper level, or ceiling, while the transmitter is catching up. If it had, at the start, a lower gain, it will keep getting ones and, therefore, will catch up to the transmitter. The same phenomenon will happen at the lower bound, the floor, when a high level speech is present at the transmitter.

This modification was implemented on the experimental set-up, and the system was tested under various conditions. The performance, as far as synchronization is concerned, is improved by decreasing the companding interval, increasing the companding incremental gain, and limiting the upper and lower bounds of the gain. The performance also depends on the nature of the speech signal encountered: its dynamic range, the presence of silence intervals, and the background acoustic noise. It was observed that, with a 5 msec companding interval, a 1.8 companding gain, and by limiting the gain to 32 levels, the system was performing reasonably well. It was able to regain synchronization after a complete loss. It is, therefore, very appropriate to mobile communications, where deep fades often occur, and cause a loss of synchronization.

6.2.7 Experimental Results and Conclusions

Several speech companding approaches have been proposed and evaluated in the previous sections. The evaluation was based on their suitability for use within the switched tone concept proposed in this thesis. It is concluded that the most appropriate approach is to use dynamic speech companding with differential coding. The proposed coding/decoding algorithm is the one shown in figure 6.5. It is also proposed to augment the system with the modification described in section 6.2.6; in this way, the system is fully adaptive and is able to regain synchronization, even after complete loss of synchronization occurs.

The various parameters of the companding algorithm, the companding interval, the incremental companding gain, and the number of gain levels required have been discussed, independently, in the previous sections. There are, however, some global compromises to be taken.

The performance of a speech compandor is measured by its ability to improve the signal to noise ratio. There are, however, other factors affecting the evaluation of a specific compandor. These factors are the frequency spectrum occupied by the companding information, the tracking performance of the compandor, and its ability to regain synchronization when this is required. It is, therefore, necessary to consider all these factors in the selection of the companding parameters. Increasing the companding interval, for example, improves the spectrum aspects but degrades the tracking performance, the ability of the system for resynchronization, and the signal to noise performance.

Most of the previous performance criteria are not easily indentifiable in a standard quantitative way; they also depend on the characteristics of the input speech signal. Therefore, it was concluded that an attempt to relate the various parameters, analytically, would lead to inaccurate results. Instead, a more appropriate approach would be to examine every one of the parameters, evaluate its effect on the performance of the system, and set a lower and an upper bound to its value. Subjective tests are, then, used for the global compromise. This approach was adopted, and, since every parameter have already been evaluated separately, this section will only include the results of the final subjective tests.

The experimental set-up used to evaluate speech companding was described in section 1.2. The adaptive differential speech companding algorithm shown in figure 6.5 was implemented in real-time. The input speech signal is sampled at 8 kHz, digitized, and compressed. White Gaussian noise is added to the compressed speech signal according to any desired signal to noise ratio. The signal is then fed to the expander, along with the companding gain information. It is possible to introduce errors in the companding gain bit, with any desired rate and any desired distribution. The output of the expander is then converted to analog, filtered, and

fed to a speaker. It is possible to vary the companding interval, the incremental gain value, and the number of gain levels.

It was observed that the optimum performance is achieved with a companding interval of 10 msec, an incremental companding gain of 3 dB, and with 24 discrete gain levels. There was no apparent distortion in the recovered signal, and an average 25 dB of improvement in signal to noise ratio was achieved.

6.3 Automatic Gain Control (AGC) Subsystem

Introduction

In the mobile environment, the signal exhibits amplitude and phase fluctuations due to multipath fading and shadowing. In amplitude modulated (AM) signaling the information is contained in the envelope, therefore, it is necessary to have an automatic gain control (AGC) system to compensate for the amplitude fluctuations. This section describes and evaluates different AGC approaches appropriate to amplitude modulation in mobile communications, and an AGC system is proposed for use within the switched tone concept.

Section 6.3.1 classifies the different types of AGC required in an SSB system. A feedforward quadrature AGC system is proposed in section 6.3.2. An alternative combined AGC/AFC simple scheme commonly used in SSB systems is presented and evaluated in section 6.3.3. The experimental testing set-up used to evaluate the quadrature AGC algorithm is described in section 6.3.4. Some design considerations and implementation aspects are discussed in section 6.3.5. Finally, the various results and conclusions are presented in section 6.3.6.

The feedforward quadrature AGC system proposed in this section is appropriate for the switched tone (ST) concept proposed in this chapter. Nevertheless, it could be optimized for any other pilot-based AM scheme used in the mobile environment, even if the pilot is phase modulated.

6.3.1 Different Types of AGC Required in an SSB System

There are two types of AGC required to compensate for slow and fast variations in the received signal level. The slow variations are very large, up to 120 dB, and are a result of changing the distance between the transmitter and the receiver as well as gross variations in the terrain. Feedback AGC techniques are usually used to compensate for these variations [45]. The fast variations are smaller in magnitude and are due to multipath propagation. Feedforward AGC techniques are usually used to compensate for these variations [39-44].

The feedback slow AGC needed is conventional. It is used in most AM receivers such as full carrier AM, double sideband diminished carrier, and single sideband receivers. Slow AGC is usually implemented using analog techniques. Fast AGC, on the other hand is not very common. New digital signal processing techniques are proposed to overcome the fast amplitude variations in the received signal. Therefore, this work will be focussed only on fast feedforward AGC.

6.3.2 Feedforward Quadrature AGC Algorithm

The feedforward quadrature AGC scheme proposed in this section is suitable for any pilot-aided system. The proposed scheme removes the amplitude fluctuations, only; it is, therefore, capable of tracking the amplitude of a constant pilot tone

as well as a phase modulated tone, which makes it appropriate for switched tone systems incorporating dynamic speech companding. The scheme requires the extraction of both the in-phase and quadrature components of the received signal, which is done at RF using a conventional quadrature receiver. The proposed algorithm does not require any critical filtering, and, therefore, very promising results have been achieved.

Figure 6.7 shows the proposed AGC system as well as the front end of the quadrature receiver. The received signal R is multiplied by the in-phase and the quadrature components of the carrier. Each branch is then low pass filtered.

Assume that a signal T is transmitted in the mobile environment. The signal exhibits the multiplicative impairments described in chapter (2), and according to the baseband channel model and the derivations described in chapter (4), section 4.2.5, the complex baseband received signal at points (A) and (B) can be written as:

$$(A) \quad R_i = C_i T_i - C_q T_q$$

$$(B) \quad R_q = C_i T_q + C_q T_i$$

where T_i and T_q are the in-phase and the quadrature components of the transmitted signal. C_i and C_q are the in-phase and quadrature components of the complex multiplicative fading. They are slowly varying signals, and their bandwidth depends on the carrier frequency and the vehicle speed. Typically, C_i and C_q are bandlimited to less than 100 Hz.

Assuming that the transmitted signal consists of a single tone at ω_s , representing the speech signal and a pilot tone at ω_p , T_i and T_q are:

Figure 6.7: Feedforward Quadrature AGC System

$$T_i = \cos(\omega_s t) + \cos(\omega_p t) = T_{is} + T_{ip}$$

$$T_q = -\sin(\omega_s t) - \sin(\omega_p t) = T_{qs} + T_{qp}$$

The signals at the various points of the receiver shown in figure 6.7 are as follows:

$$(C) \quad C_i T_{ip} - C_q T_{qp}$$

$$(D) \quad C_i T_{qp} + C_q T_{ip}$$

$$(E) \quad C_i^2 T_{ip}^2 - 2C_i C_q T_{ip} T_{qp} + C_q^2 T_{qp}^2$$

$$(F) \quad C_i^2 T_{qp}^2 + 2C_i C_q T_{ip} T_{qp} + C_q^2 T_{ip}^2$$

$$(G) \quad C_i^2 [T_{ip}^2 + T_{qp}^2] + C_q^2 [T_{qp}^2 + T_{ip}^2]$$

$$(G) \quad [T_{ip}^2 + T_{qp}^2] [C_i^2 + C_q^2]$$

$$(G) \quad [\cos^2(\omega_p t) + \sin^2(\omega_p t)] [C_i^2 + C_q^2]$$

$$(G) \quad [C_i^2 + C_q^2]$$

$$(H) \quad \sqrt{C_i^2 + C_q^2}$$

$$(I) \quad C_i T_{is} - C_q T_{qs}$$

$$(K) \quad [C_i / \sqrt{(C_i^2 + C_q^2)}] T_{is} - [C_q / \sqrt{(C_i^2 + C_q^2)}] T_{qs}$$

Let

$$[C_i / \sqrt{(C_i^2 + C_q^2)}] = \cos(y)$$

Then

$$[C_q / \sqrt{(C_i^2 + C_q^2)}] = \sin(y)$$

$$(K) \cos(y) \cos(\omega_s t) + \sin(y) \sin(\omega_s t)$$

$$(K) \cos(\omega_s t - y)$$

From the previous equation it is clear that the recovered signal at (K) is free of amplitude fluctuations. The phase fluctuations, on the other hand are not affected. This is useful in the case of dynamic companding, where the pilot tone is phase or frequency modulated. The circuit does not require critical filters, therefore, very good results have been achieved.

The constant (ϵ) is added to the circuit to avoid a division by zero when the signal is too weak. It limits the noise enhancement problem encountered during deep fades. Increasing the value of ϵ limits the dynamic range of the automatic gain control system.

In a switched tone system, the pilot filter is tuned to the pilot tone above band. During the acquisition period, the system is disabled.

6.3.3 Alternative AGC Approach

Figure 6.8: AGC-AFC System, First Configuration

The alternative AGC approach presented in this section has been proposed in [39-44]. Figure 6.8 shows a block diagram of this simple AGC/AFC system. Assume that the speech signal is a single tone at (ω_s) and that the pilot tone is at (ω_p)

The input signal at point (A) is:

$$(A) \quad 2EX(t) \cos(\omega_p t + \omega_e t + y(t)) + 2SX(t) \cos(\omega_s t + \omega_e t + y(t))$$

where:

E represents the pilot level

S represents the signal level

X(t) is the amplitude fading

ω_p is the pilot radian frequency

ω_s is the radian frequency of a tone within the band

ω_e is a fixed offset

y(t) is the phase fading

The signals at the different points are:

$$(B) \quad EX(t) \cos(2\omega_p t + \omega_e t + y(t)) + EX(t) \cos(\omega_e t + y(t)) + SX(t) \cos(\omega_s t + \omega_e t + \omega_e t + y(t)) + SX(t) \cos(\omega_s t - \omega_p t + \omega_e t + y(t))$$

$$(C) \quad EX(t) \cos(2\omega_p t + \omega_s t + y(t)) + S X(t) \cos(\omega_s t + \omega_p t + \omega_s t + y(t))$$

$$(F) \quad 2EX(t) \cos(\omega_p t + \omega_s t + y(t))$$

$$(G) \quad 2E^2 X^2(t) [1 + \cos(2\omega_p t + 2\omega_s t + 2y(t))]$$

$$(H) \quad 2E^2 X^2(t)$$

$$(I) \quad [\cos(\omega_p t + \omega_s t + y(t))] / [EX(t)]$$

$$(D) \quad \cos(3\omega_p t + 2\omega_s t + 2y(t)) + \cos(\omega_p t) + S \cos(\omega_s t + 2\omega_p t + 2\omega_s t + 2y(t)) / E + S \cos(\omega_s t) / E$$

$$(E) \quad \cos(\omega_p t) + S \cos(\omega_s t) / E$$

This output signal at point (E) is the recovered signal. The amplitude fluctuations $X(t)$, the phase fluctuations $y(t)$, and the frequency offset (ω_s) are compensated for. If non ideal filters are considered instead of ideal filters, many problems arise [39,40]. Some extra delays have also to be added to synchronize the system.

This configuration has been implemented on the TMS-320 testing system. The performance of the system was poor due to some inherent limitations in the configuration and due to the critical filtering required. An example of the limitations of this configuration is the LPF prior to point (E). For the system to function properly, the following conditions always have to be satisfied:

$$3\omega_{pt} + 2\omega_{st} + 2y(t) > \min(\omega_{st}, \omega_{pt})$$

$$\omega_{st} + 2\omega_{pt} + 2\omega_{st} + 2y(t) > \min(\omega_{st}, \omega_{pt})$$

These two conditions are not always satisfied, specially at the initial acquisition where ω_{st} could reach $\pm$ few kHz.

6.3.4 AGC Experimental Set-Up Description

Figure 6.9 shows a block diagram of the AGC experimental set-up. The experimental set-up was designed to model as close as possible the feedforward quadrature AGC algorithm described in section 6.3.2. This section describes the various parts of this set-up as well as the inputs and the outputs.

The inputs of this system consist of three pairs of signals, the transmitted pilot tone, the multiplicative impairment signal, and the additive noise.

The first pair of inputs is the pair (T_i, T_q) which represents the in-phase and the quadrature components of the baseband pilot tone used for AGC. This pilot tone is generated internally at the desired frequency. It is possible to modulate this pilot tone using an external or internal pseudo random binary sequence generator at any desired bit rate. Several modulation techniques could be used, such as continuous phase frequency shift keying (CPFSK) with any desired modulation index, and phase shift keying (PSK).

The second pair of inputs (C_i, C_q) represents the multiplicative impairments introduced by the channel. It is generated by the mobile radio channel simulator described in chapter (4). The level of the input pair (C_i, C_q) could be varied by the attenuation factor FAD, in order to simulate any desired average received signal power.

Figure 6.9: AGC Experimental Setup

The third pair of inputs (N_i , N_q) is the additive white Gaussian noise. It was generated either internally or externally. The signal level could be adjusted by the multiplier NOISE. Therefore, any desired noise variance could be obtained, which enables the testing under any desired signal to noise ratio (SNR).

According to the complex baseband channel model described in chapter (4), section 4.2.6, the input complex baseband pilot is multiplied by the complex multiplicative impairment input. A noise component is added to each of the real and the imaginary components of the product. The resulting signal pair (R_i , R_q) represents the in-phase and the quadrature component of the received pilot.

The AGC algorithm described in section 6.3.2 is then performed on the pair (R_i , R_q). Each branch is squared, then both are added. A square root operation is then performed to obtain the envelope of the incoming signal. R_i and R_q are then divided by this envelope to form the recovered in-phase and quadrature signals, I and Q. Before the division operation, a block is added to ensure that a division by zero never occurs.

6.3.5 Implementation Aspects and Design Considerations

In this subsection, the various blocks shown in figure 6.9 are described. Some design considerations related to each block are discussed. Finally, some problems related to the signal to noise ratio (SNR) are discussed and a clear definition of SNR is presented.

The mobile channel simulator block is external. It was described in detail in chapter (4).

The second block is the white Gaussian noise generation block. The function

of this block is to generate two independent zero mean white noise signals having a Gaussian amplitude distribution. This block was used as one of the building blocks of the channel simulator. It was described in chapter (4), section 4.2.2.

The SQRT block accepts a 32 bits number and generates a 16 bits number representing the square root of the input. Several algorithms have been considered for this block. The successive approximation algorithm was selected due to the low number of data and program memory locations required, as well as its fast execution time.

Figure 6.10 shows a block diagram of the algorithm. It is based on a trial and error scheme. The most significant bit of a temporary register is set to one and the register is squared. If the result is larger than the input number, this means that setting the most significant bit to one is wrong; therefore, it is reset to zero. If the result is smaller than or equal to the initial number, this means that setting the most significant bit is the right approach. Therefore, the most significant bit of the square root result is determined. The same approach is repeated for the successive significant bits.

The division block uses the division routine described in reference [77], page 5-6. It is based on a series of conditional subtractions. A block is added before the division block to ensure that a division by zero never occurs. This block limits the value of the denominator to a small constant: ϵ .

The AGC system shown in figure 6.9 removes the amplitude fluctuations in the received signal; the basic operation of the system is to amplify the incoming signal up to some predetermined level. When a deep fade is encountered, the received signal is extremely small, and its signal to noise ratio is also very small. In this case, the amplification factor becomes very high, and a noise enhancement problem arises. This problem can deteriorate the speech quality as well as the

Figure 6.10: Square Root Algorithm

performance of the other associated subsystems such as the AFC. It is, therefore, very important to limit the dynamic range of the AGC system by limiting the value of the denominator of the AGC division to the value ϵ . Several experiments have been conducted in order to characterize the effect of the constant ϵ under different signal to noise ratios.

The same noise enhancement problem arises if the transmitter stops transmitting. A higher level system is required to control the AGC subsystem. The AGC starts operating only when a start message is received and stops as soon as a stop message is received.

It is important in the evaluation process to define in a clear way the signal to noise ratio. In a fading environment, the received signal power varies due to multipath fading and shadowing. Furthermore, in the set-up used here, there are many scaling operations used to reach the maximum dynamic range of the fixed-point processor. There are also some A/D, D/A, and filtering operations that affect the scaling. Therefore, it was concluded that it would be difficult and very approximate to determine the SNR analytically; instead, the signal to noise ratio was calculated experimentally. The energy in a given signal is proportional to the sum of the square of its samples. A program was developed to calculate the average energy in a given signal for any given time period up to 30 minutes. Therefore, if the time period is fixed, the program output represents the signal power.

In the set-up shown in figure 6.9, the signal to noise average power ratio can be adjusted by the factors FAD and NOISE. First, in order to obtain a desired average signal power, the noise attenuation factor, NOISE, is set to zero, and the signal (R_i) is fed to the average power calculation program. The required average signal power is obtained by adjusting the signal attenuation factor FAD. Similarly,

it is possible to set the signal attenuation factor FAD to zero and adjust the noise attenuation factor NOISE to obtain the desired average noise power. Therefore, any signal to noise average power ratio could be obtained very accurately.

6.3.6 Results and conclusions

The results and conclusions presented in this section were obtained from the feed-forward experimental quadrature AGC system described in section 6.3.4.

The input of the system was a 500 Hz pilot generated internally. The mobile channel simulator parameters were set to an RF carrier frequency of 850 MHz, a vehicle speed of 25 km/hr, no line of sight, and no shadowing. These parameters generate a Doppler frequency shift of 19.68 Hz. The additive noise was set to zero. Therefore, the noise in the system is mainly due to the truncation error in the various blocks.

Figure 6.11 shows a time domain plot obtained from the oscilloscope. The upper trace shows the faded pilot tone before the AGC; it is clear that the envelope is fluctuating due to the multipath fading. The lower trace shows the same signal after the AGC, the envelope is almost constant.

Figure 6.12 shows some frequency domain plots obtained from the spectrum analyser. The original pilot is a unit impulse at 500 Hz. The faded pilot has a wider spectrum, having two peaks at about 480 Hz and 520 Hz. This is due to the phase fluctuations due to the Doppler effect which is about 20 Hz in this case. The amplitude fluctuations on the other hand cannot be identified in this frequency domain plot. The recovered constant envelope signal after the AGC is also shown in figure 6.12. It has the same spectrum as the faded pilot in addition to some white noise. This noise depends on the value of the constant ϵ , which

Figure 6.11: Faded Pilot Tone Before and After AGC (Time Domain)

Figure 6.12: Faded Pilot Tone Before and After AGC (Frequency Domain)

represents the dynamic range of the AGC system. It is noticed that the smaller ϵ , the larger this background noise is. It is also noticed that the spectrum of the recovered signal has the same shape as the faded pilot, including the spread and the two peaks. This is due to the fact that the AGC system tested here corrects the amplitude fluctuations only, and does not affect the phase fluctuations.

The performance of the AGC system was also tested under several SNR, or several average signal power to average noise power ratios. These ratios were set according to the description in section 6.3.5. The effect of noise on the factor ϵ was also tested.

Figure 6.13 shows a time domain plot of the faded noisy pilot and the recovered signal after the AGC. The only information from this plot is that the signal after the AGC has a constant envelope.

Figure 6.14 is a frequency domain plot of the faded pilot as well as the recovered signal after the AGC. The average signal power to the average noise power ratio is set to +20 dB. When $\epsilon = +10$, which represents an AGC dynamic range ratio of up to 3000, it is noticed that the average background noise level is about 7 dB higher than the original faded pilot. When the dynamic range of the AGC is reduced to 30, this background noise level is only about 3 dB higher than the original faded pilot.

Figure 6.15 shows a frequency domain plot of the recovered signal after the AGC. The dynamic range of the AGC system is set to the ratio 30, and several average signal power to average noise power ratios are plotted.

The AGC subsystem described in this section is a general AGC scheme that could be used with any pilot-aided amplitude modulated system; its dynamic range is programmable. It was successfully tested for pilot-aided SSB transmission; it

Figure 6.13: Faded Pilot Tone Before and After AGC (Time Domain)

SNR = 0 dB

Amplitude
10 dB/div

Figure 6.14: Faded Pilot Tone Before and After AGC (Frequency Domain)

Amplitude
10 dB/div

Figure 6.15: Recoved Signal After AGC

was also tested for some pilot-aided data transmission techniques such as pilot-aided M-ary QAM.

6.4 Automatic Frequency Control Subsystem (AFC)

Introduction

Automatic frequency control (AFC) is the last subsystem required for the switched tone approach proposed in this chapter. Several AFC configurations, presented in [48], are suitable to our specific application. In this section, the different types of AFC required within the switched tone concept are described in subsection 6.4.1. Several AFC discriminator configurations are presented in subsection 6.4.2. Finally, some conclusions are presented in section 6.4.3.

6.4.1 Different Types of AFC Required

The automatic frequency control subsystem (AFC) has two main functions, to remove the frequency offset between the transmitter and the receiver and to track the phase fluctuations due to fading.

The first function of the AFC is to remove the frequency offset between the transmitter and the receiver during the initial acquisition period. In the 900 MHz band, if a crystal with 10^{-6} accuracy is used, this offset could be up to 1,800 Hz.

Many problems associated with the initial acquisition were partially solved in a system-wise approach by the switched tone concept described in sections 3.3.2, and 6.1. One of the main problems of other systems such as the tone above band (TAB) is that the pilot tone is located above the speech spectrum, close to the edge

of the band. Therefore, during the initial acquisition phase the pilot tone could be completely out of band and, consequently, the whole transmission session fails. In a switched tone system, on the other hand, during the initial acquisition period, the pilot is located at midband, which increases the AFC acquisition range.

The second function of the AFC subsystem is to track the phase fluctuations due to fading. These are characterized by being relatively small in magnitude but quite fast. In a switched tone system, in the analog speech transmission mode, these functions have to be performed on the tone above band. If fixed companding is used, this tone is an unmodulated carrier. If dynamic companding is used, this carrier tone could be modulated by the companding information. The modulation technique used could be either FSK or DPSK.

6.4.2 Different AFC Configurations

Figure 6.16 shows a block diagram of a typical AFC loop configuration. The input signal to this block, S could be either at IF or at RF. The signal is translated into the in-phase and the quadrature baseband form, R_i and R_q . It is then fed to an AFC discriminator. The output of the AFC discriminator, V_e is proportional to the frequency deviation. V_e is filtered and fed to the voltage controlled oscillator (VCO). The various design considerations concerning the loop filter and the VCO will not be addressed in this thesis. Several configurations for the AFC discriminator will be presented and evaluated.

The main function of the AFC discriminator is to generate a voltage V_e proportional to the frequency error. The simplest configuration of the frequency discriminator is a differentiator (FD), which could be implemented by a finite impulse response (FIR) filter. This configuration was implemented in an experi-

Figure 6.16: Typical AFC Loop

mental system including both AGC and AFC. The results will be presented in the next section.

It is possible to improve the performance of the AFC system and consequently the whole ST system by using a more precise frequency discriminator (FD). Three basic FD configurations are briefly presented in this section, the differentiate and multiply discriminator (DIFAFC), the cross product discriminator (CPAFC), and the discrete Fourier transform discriminator (DFTAFC).

The first configuration, the differentiate and multiply discriminator (DIFAFC) is shown in figure 6.17. Reference [49] includes a detailed description of the DIFAFC. It is particularly suitable for analog implementation in a low signal to noise ratio environment. The DIFAFC is useful for tracking CW, DPSK, or the average of an FSK signal.

The second configuration, the Cross Product AFC configuration (CPAFC) is shown in figure 6.18. It is described in detail in reference [48]. The CPAFC is similar to the DIFAFC with the differentiators replaced by delays. This is very convenient for digital implementation. CPAFC is suitable for tracking CW, or the average value of an FSK signal. Several variations of the CPAFC suitable for use with DPSK signal are described in [48]. The CPAFC with sample excising, the sample doubling CPAFC, the cross dot product CPAFC, and the CPAFC with decision feedback are some examples of these variations.

The third configuration, the Discrete Fourier Transform AFC configuration (DFTAFC) is shown in figure 6.19. It is described in detail in reference [48]. The DFT can be viewed as a spectrum analysis algorithm which provides the equivalent of a filter bank. The DFTAFC can be used with DPFSK and FSK signaling. It is very suitable for digital implementation. A DFTAFC was implemented on the experimental set-up described in chapter (1), using an FFT. It was tested

Figure 6.17: Differentiate and Multiply AFC Discriminator (DIF AFC)

Figure 6.18: Cross Product AFC Discriminator (CPAFC)

Figure 6.19: Discrete Fourier Transform AFC Discriminator (DFTAFC)

independently, but not with other switched tone subsystems such as the speech companding and the AGC.

According to the performance comparisons of the previous discriminator configurations shown in reference [48], and according to the experiments conducted in this work, it is concluded that the most appropriate AFC configuration for switched tone systems would be the DFTAFC. The decision to replace the simple frequency discriminator by a DFTAFC depends on the specific application. The factors affecting this decision are the frequency of operation, the cost involved, and the speech quality required.

6.4.3 Conclusions

The automatic frequency control problem was solved in a systemwise approach through the switched tone concept. The initial acquisition is performed by a mid-band tone, and the tracking of the phase fluctuations introduced by the multipath fading is performed by a tone above the speech band. The system performance is acceptable using a simple differentiator as frequency discriminator. It is possible to improve the system performance by using a better AFC configuration such as the DFTAFC described in this chapter. A detailed analysis of the simple frequency discriminator was not performed. The initial acquisition problem was reduced to the acquisition of a continuous pilot at midband, which is a very conventional problem. The tracking of the phase fluctuations also becomes relatively simple once the initial acquisition is performed properly.

6.5 Conclusions

In this chapter, a novel concept, the switched tone (ST), was proposed for speech and data transmission over mobile channels. In the ST concept, the speech is treated on a spurt by spurt basis. It is possible to transmit data, intermittently, between talkspurts, or continuously, when no speech is present. If speech only is to be transmitted, it is possible to simplify the ST concept, and have a speech-only system.

In the speech-only mode, the ST concept can be viewed as a modified single sideband (SSB) transmission system. During the initial acquisition phase, a mid-band tone is transmitted, then the speech is transmitted using SSB modulation. A pilot tone is transmitted above the speech spectrum for automatic gain control (AGC) and automatic frequency control (AFC) purposes. The channel spacing is set to 5 kHz, which represents a saving of up to 6:1 over conventional FM transmission. The various subsystems required by the ST concept, speech companding, AGC, and AFC, were evaluated.

Speech companding is necessary to enhance the signal to noise ratio in the low volumes syllables of the speech signal. An adaptive dynamic speech companding system was proposed in section 6.2. An improvement of about 25 dB in signal to noise ratio is achieved.

ACC is necessary for removing the amplitude fluctuations caused by multipath propagation and shadowing. A quadrature AGC system is proposed in section 6.3. The system is capable of tracking a constant pilot tone as well as a phase modulated pilot.

The AFC problem was not explicitly analysed in this work, nevertheless, the different types of AFC required within the ST concept were discussed, and a survey

of several existing AFC configurations suitable for the ST concept was presented in section 6.4.

Chapter 7

Conclusions and Proposed Future Work

This chapter includes a summary of the general conclusions derived from the work carried throughout this thesis. A detailed conclusion subsection follows every major section.

It is possible to classify the main contributions achieved in this thesis within three main research problems, the mobile radio channel simulator, the background acoustic noise in mobile telephony, and the switched tone concept along with its associated subsystems. Each one of these research problems will be presented, along with the possible extensions, suitable for future research work.

7.1 Mobile Radio Channel Simulator

The first problem addressed in this thesis is the design of a mobile radio channel simulator. The basic configuration of the simulator is conventional. There were,

however, several implementation difficulties such as the design of the shaping filters, which have a variable cut-off frequency. These problems were solved in the thesis through the use of advanced techniques such as multirate and variable rate digital signal processing. The architecture of the simulator, including the tasks partition between software and hardware, and the variable rate antialiasing filters, was, also, a major step toward solving these problems. The basic conventional configuration of the simulator was then augmented with several sections allowing the simulation of a line of sight ratio and shadowing, with a wide range of parameters.

The simulator was implemented and tested. The results obtained from the simulator were compared to some published field experiments. A good match can be observed, which proves that the simulator is reliable and simulates the actual conditions. The simulator was used throughout the thesis to test the various proposed concepts under fading condition.

It is possible to add new features to the simulator, such as frequency selective fading, and variable vehicle speeds. It is also possible to modify the baseband section in order to simulate other channels such as the satellite and the in-building channels.

7.2 Background Acoustic Noise Reduction

The second research topic addressed in this thesis is the problem of background acoustic noise in mobile communications. This noise is coupled to the transmitted speech signal through the microphone. First, the various sources of the background acoustic noise were classified. Then, some experimental tests were conducted in order to characterize this noise. The main conclusion from these experiments is that the noise outside the cabin of the car is highly correlated with the noise

inside the cabin. Finally, a system is proposed to reduce the background acoustic noise from the transmitted speech signal, and, therefore, improve the signal to noise ratio at the transmitter.

The proposed system is based on the two microphones adaptive noise cancelling approach. In this approach, a primary microphone captures the speech signal plus noise. A secondary microphone is placed so that it captures a speech-free noise signal, highly correlated with the primary noise. The output of the secondary microphone is fed to an adaptive filter which estimates the noise corrupting the speech signal. Ideally, a noise-free speech signal could then be obtained.

The filter proposed for this application is an adaptive FIR transversal filter using the stochastic gradient algorithm for adaptation. A noise reduction of 10 dB was achieved using a 128 taps filter. The system performance depends on the car model, the status of the windows, and the wind friction; this is due to the fact that the proposed filter is not capable of cancelling all the noise sources equally. The longer the filter the better is the performance regarding noise sources with a static origin. On the other hand, a long filter implies a longer adaptation time, which deteriorates the system performance regarding the noise sources with dynamic origin such as the air flow noise.

Other filter configurations such as adaptive IIR filters using fast adaptation algorithms could be potential candidates for this application; IIR filters are capable of achieving the same performance with 1/5 the number of taps. There are, however, several outstanding problems associated with these configurations, such as stability and high complexity, which are still to be solved.

Other extensions of this research topic include the investigation of the possibility of having several reference microphones. It is also possible to extend the concept, and investigate the possibility of having several speakers and try cancel

the background noise acoustically.

7.3 Switched Tone (ST) Concept

A novel concept, the switched tone (ST) is proposed in this thesis for speech and data transmission over mobile channels. In the ST concept, the speech is treated on a spurt by spurt basis. It is possible to transmit data intermittently, between the speech spurts, or continuously, when no speech is present. In the speech transmission mode, a switched tone based system uses a modified SSB technique. A mid band pilot tone is transmitted for the initial acquisition phase. The pilot is then transmitted above the speech band to provide the receiver with the necessary information for AGC and AFC. The performance of a switched tone system depends on the performance of its associated subsystems, the speech companding, the AGC, and the AFC.

Speech companding is necessary to enhance the signal to noise ratio in the low volume syllables of the speech signal. A dynamic differential speech companding technique is proposed in this thesis. The companding information modulates the tone transmitted above the speech band.

AGC is essential in any amplitude modulation scheme such as the SSB used here. Two types of AGC are required, a slow AGC that compensates for the shadowing and the distance between the transmitter and the receiver, and a fast AGC to compensate for the multipath propagation. The slow AGC is a conventional analog one. A fast AGC subsystem is proposed in this thesis, using a quadrature configuration. The performance of the AGC subsystem is evaluated for various signal to noise ratios and various AGC tracking ranges.

The AFC problem is solved in a systemwise approach through the switched tone concept. The problem is divided into two relatively simpler problems, the initial acquisition phase and the fast fading tracking. The initial acquisition is performed through a mid-band tone, which provides a tracking range of one half the baseband bandwidth. The tracking of the phase fluctuations is performed by the tone above band. The AFC subsystem was tested using simple frequency discriminators, and an acceptable performance was achieved. It is possible to improve the AFC performance by using better AFC configurations such as the DFTAFC presented in the thesis.

The ST concept provides a solution for the radio spectrum congestion problem. It provides a bandwidth efficiency saving of up to 6:1 compared to analog FM, while maintaining an acceptable toll quality speech. The complexity of a switched tone system is well within the range of the present VLSI technology.

The work presented within the ST concept could be extended to several future research topics. The speech transmission aspect has been analysed; nevertheless, several areas such as the effect of using a more complex AFC, and the modulation scheme used for the companding gain, are still open to future investigation. The data transmission, on the other hand, has not been addressed in this work. The possibility of using a pilot-aided data transmission scheme, such as pilot-aided M-ary APK, with the same AGC and AFC proposed in this work is a proposed future research topic.

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