A Feasibility Study on the Adaptation of SYNCOMPEX Concept to the VHF/UHF Land Mobile Communication Systems

by

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Chapter 1

INTRODUCTION

1.1 Scope

The objective of this research is to study the feasibility of adapting the SYNCOMPEX concept, which is being used successfully in the HF band, to the VHF/UHF land mobile radio systems in urban and rural areas.

The SYNCOMPEX (Synchronous Compression Expansion) system is similar, in principles, to the ACSB (Amplitude Compandored Single SideBand) modulation. Both systems attempt to improve the performance of the basic SSB modulation by: (i) compressing the speech signals at the transmitter and expanding them at the receiver in order to maintain a near constant signal-to-noise ratio at all syllabic levels, and (ii) adding a constant amplitude signal (pilot tone or a digitally modulated signal) to the speech to be used as a reference to correct the frequency and amplitude of the received signals.

The difference between SYNCOMPEX modem and other ACSB modems (tone-based transceivers which are now available commercially) is in the way the voice compression information is communicated between the transmitter and the receiver. Pilot-tone-based modems use fixed voice compression ratios. SO, two reciprocal circuits are included in the modem; one for the transmit mode and one for the receive mode. The circuit in the transmitter compresses the high amplitude portions of the speech by a fixed factor. The same portions are then expanded at the receiver by the same factor, hence restoring the amplitude distribution of the speech. In contrast, SYNCOMPEX modems use variable compression/expansion ratios, hence allowing for more flexibility in maintaining a near constant signal-to-noise ratio for all syllables. The variable companding ratio, however, gives rise to the need for sending the gain information to the receiver. This is accomplished by sending a digitally modulated signal, carrying the compression information, along with the analog voice signal. This technique has two basic advantages over the fixed ratio method: (1) It provides a great deal of flexibility for implementing better voice compression algorithms; and (2) The frequency spectrum of the combined analog/digital signal does not have spectral lines; this may ease the problem of in-band intermodulation as well as the out-of-band spurious radiation.

The SYNCOMPEX system may have decisive advantages over the tone-ACSB system. However, the introduction of the SYNCOMPEX system into the land mobile environment faces two problems. The first is an economical one, since the complexity and sophistication of the system may result in high cost modems. The second is a technical one which is related to the effect of fading on the synchronization between the analog voice signal and the digital signal which describes the compression parameters.

Recently, however, a new generation of programmable digital processing chips have emerged. These chips have the capabilities of high speed real time implementation of fairly complicated algorithms. A primitive version of these chips (Intel 2920) was employed successfully to replace some of the critical circuits in the HF SYNCOMPEX modem at CRC. This suggests that the now-available powerful versions can be used to solve some of the tougher problems facing the SYNCOMPEX at high frequency bands.

The work reported here represents an initial feasibility study on the adaptation of the SYNCOMPEX concept to the mobile communications at VHF and UHF bands. In particular, the work is geared towards implementing the SYNCOMPEX modem on the TI-TM320 programmable chips, and addresses the following points:

(1) A review of the SYNCOMPEX system and the various algorithms for voice compression/expansion.

(2) A study of the AGC circuit under fading conditions.

(3) A study of the AFC design problems under fading conditions.

(4) Propose modifications to the existing SYNCOMPEX technology to suit the VHF/UHF mobile environment; and

(5) Evaluation of the possible impact of SYNCOMPEX system on the spectral utilization efficiency.

1.2 Report Outline

The report is divided into three sections. Chapter 2 constitutes the first section which describes the existing SYNCOMPEX and ACSB systems and highlights the differences and similarities between the two concepts. The second section address the problems related to the AGC and AFC under fading conditions and is covered in Chapters 3 and 4. In these chapters a description of a real-time fading simulator implemented on the TMS320 is given. The fading simulator was used as a test bed for

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testing the AGC and AFC circuits.

Chapter 4 concludes the reports by giving a general evaluation of the SYNCOMPEX technique based on the result obtained so far. Proposals for some modifications to the modem are described. The chapter also contains a section on the possible impact of the SYNCOMPEX technique on the overall spectrum utilization efficiency.

Chapter 2

THE AMPLITUDE COMPANDORED SIDE BAND TECHNIQUES

2.1 Introduction

FM systems have dominated the mobile communications area for three good reasons: (1) simplicity, (2) capture effect and (3) insensitivity to amplitude fluctuations. In the all analog-all discrete components era (1920-1970) simplicity was one of the prime design considerations, since the cost of the equipment was directly proportional to its sophistication. So, in that regard, FM techniques had a decisive edge over SSB techniques. The capture effect and the insensitivity to amplitude fluctuations have put the FM system a distance ahead of all AM methods (SSB or double side-band). This is because the mobile communications are dominated by interference and signal fading. These three advantages of FM systems have overshadowed its basic disadvantage of the radio spectrum inefficiency.

Since 1970, the situation has changed dramatically. First, the demand for VHF/UHF communications have grown to a very high level. Spectrum efficiency has become as important or even more important than any other technical considerations, and the pressure to replace the FM system by a more efficient one has grown considerably. Second, the drastic advances in digital technology have resulted in a situation where the circuits complexity became less of a factor in determining the cost of the equipment. Moreover, the same advances have provided the necessary means to solve many of the problems associated with the spectrally efficient modulations such as SSB.

As a result of these changing conditions, SSB has emerged again as a strong candidate for urban VHF/UHF communications. The following sections describe the technical problems associated with SSB in the mobile environment and the two basic approaches that were developed to combat these problems.

2.2 Speech Companding

The first basic problem of AM techniques, in general, is the very large dynamic range of the speech signal. Speech dynamic range could be as large as 70 dB. This large dynamic range leaves the low level portions of the speech more vulnerable to noise than the high amplitude portions. Both ACSB and SYNCOMPEX techniques are motivated by the need to protect the low level segments of the speech from getting buried in the noise, and both are based on the same idea of reducing the dynamic range of the transmitted signal by compressing the speech signal at the source. Figure 2.1 illustrates the essence of the speech companding process.

The ACSB uses what is called Fixed Ratio Companding as shown in Figure 2.2. Two compression ratios are illustrated in the figure; 2:1 and 4:1. In both cases the original signal level is divided into three ranges (0 to -10 dB), (-10 dB to -20 dB) and less than -20 dB, with the 0 dB representing the peak level. Each range is reduced by a factor of 2 or 4 depending on the

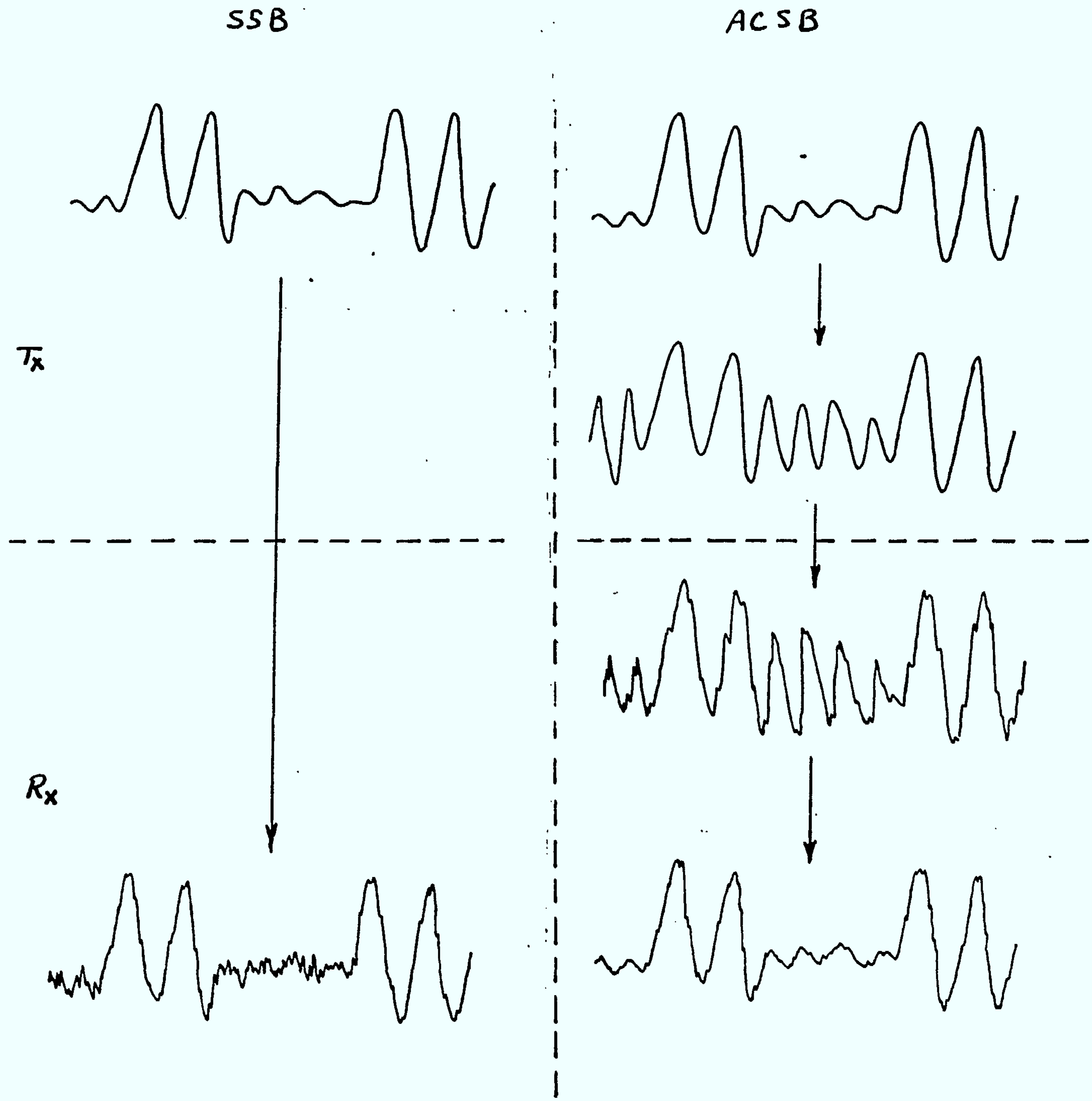


Fig. 2.1 - Noise Reduction Mechanism in ACSB Modems

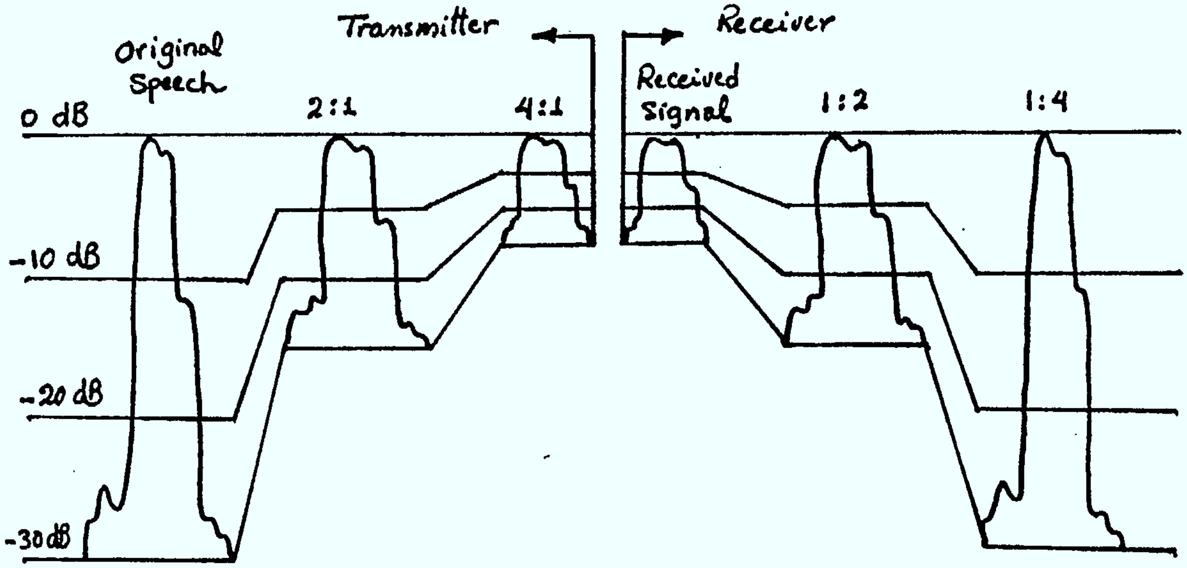


Fig. 2.2.- 2:1 and 4:1 Fixed Ratio Companding

ratio being used. For example, a speech signal whose magnitude spans the range (0 to -30 dB) will be compressed to the (0 to -7.5 dB) range. The operation is then reversed in the receiver to restore the speech to its original dynamic range. Since the compression and expansion ratios are fixed; then there is no need for the transmitter to inform the receiver about the original levels of the signal, and the process can then be implemented by means of two reciprocal circuits at the two ends of the communication link.

In contast, SYNCOMPEX uses a variable ratio compression/expansion scheme. In this case the compander ratio is adjusted for each syllable so that all syllables are transmitted with nearly the same volume. This variable ratio companding requires a control channel. Over this channel the actual companding ratio is transmitted to the receiver so that it will know by how much to attenuate the received voice signal. The control channel should then be combined with the companded voice signal for transmission.

2.2.1 Implementation of Variable Ratio Companding

The implementation of variable ratio companding requires more signal processing than the simple fixed ratio companding used in ACSB. The basic processing blocks needed to adapt the compression/expansion ratio to the syllabic volume are shown schematically in Figure 2.3. The speech signal is segmented into small portions of 10-50 msec each. A power estimation circuit

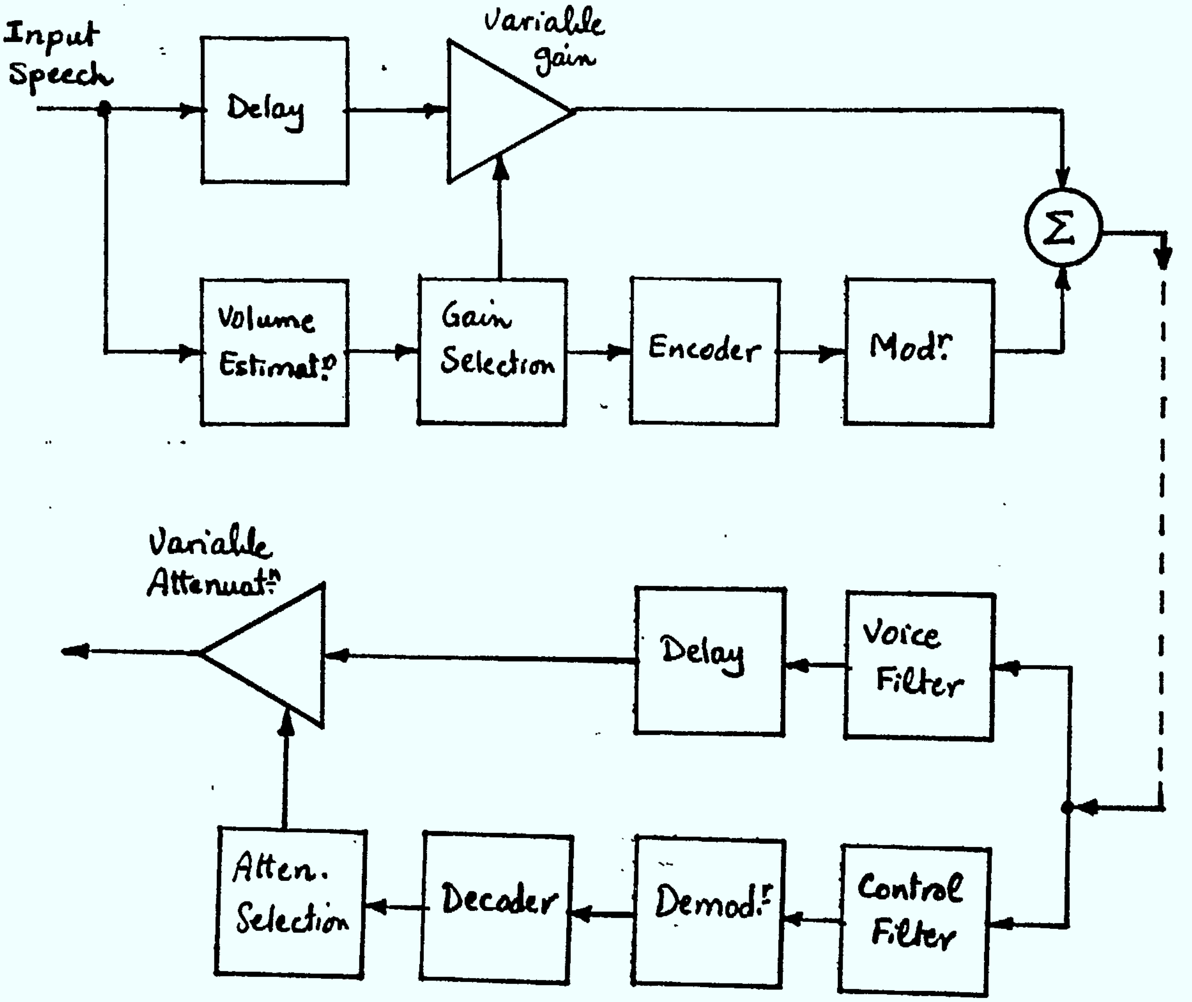


Fig. 2.3 - The Compression / Expansion Process in SYNCOMPEX MODEMS

estimates the signal volume for each speech segment and selects the appropriate gain level. The selected gain, which is a quantized binary number, is then encoded and passed to the digital modulator forming the digitally modulated control signal. The compressed speech and the control signal are then combined and passed to the rest of the transmission equipment. On the receiver side, the control signal is demodulated, decoded and converted back to the appropriate gain level which expands the speech back to its original dynamic range.

2.2.2 Gain Coding

There are several schemes for mapping the gain levels into a small number of binary digits. The objective of these methods is to achieve a large dynamic range and good gain tracking with the smallest possible data rate on the control channel, and to ensure that transmission errors do not lead to disasterous gain errors. A comparison between these methods is given in [1] and [2] and such a comparison will not be repeated here. For our purpose it suffices to say that a bit rate of 100-200 bps seems to satisfy a reasonable set of design objectives.

Assuming that the data rate on the control channel is 200 bps the combined voice/data spectrum would be as shown in Figure 2.4.

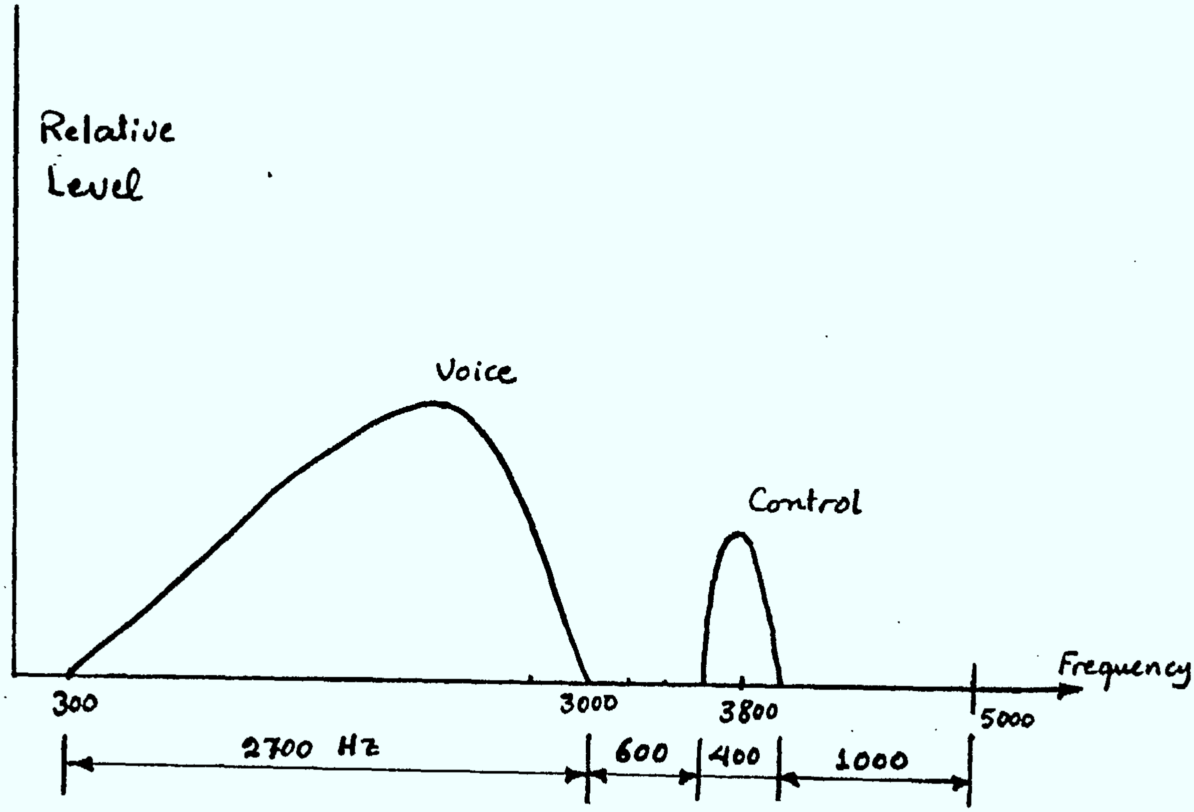


Fig. 2.4. – Combined Voice / Control Baseband spectrum of SYNCOMPEX Signals

2.3 Automatic Gain Control (AGC)

The second problem facing AM techniques in the mobile environment is the design of the AGC circuit. The amplitude fluctuations of the AM received signal come from two sources: (1) the information impressed on the carrier, and (2) the fading processes. The AGC circuit should be designed to track and correct the fading fluctuations while leaving the information-carrying modulation intact. At low carrier frequencies, the design is relatively simple since the fading process would be much slower than the lowest frequency component of the speech signal. As the carrier frequency increases the design of an effective AGC circuit becomes more complicated.

ACSB and SYNCOMPEX systems solve this problem in two different ways. In a typical ACSB system a pilot tone is sent, along with the compressed speech, at reduced power. At the transmit side the pilot tone has a constant magnitude; so, the received pilot magnitude will carry the fading pattern and can be used to correct the magnitude of the voice signal. The SYNCOMPEX system, on the other hand, uses a digitally modulated signal to convey the compression data to the receiver. If a constant envelope modulation technique is used (e.g., FSK), the same signal could be used to recover the fading envelope at the receiver, in much the same way as the pilot tone method.

2.4 Automatic Frequency Control

SSB modulation is known to be sensitive to phase error. The generation of SSB signals relies on combining the real and imaginary versions of the modulating signal accurately. A phase error will upset that balance and results in a distorted spectrum. Therefore, one of the main requirements of SSB system is the coherent demodulation, and since it is, in general, difficult to extract the carrier frequency from a SSB modulated signal, the system should be aided by an auxillary signal for the carrier recovery process. The pilot tone in ACSB and the control signal in SYNCOMPEX provide such an auxillary signal.

A complete block diagram of SYNCOMPEX system showing the AGC, AFC and the compression/expansion mechanism is shown in Figure 2.5.

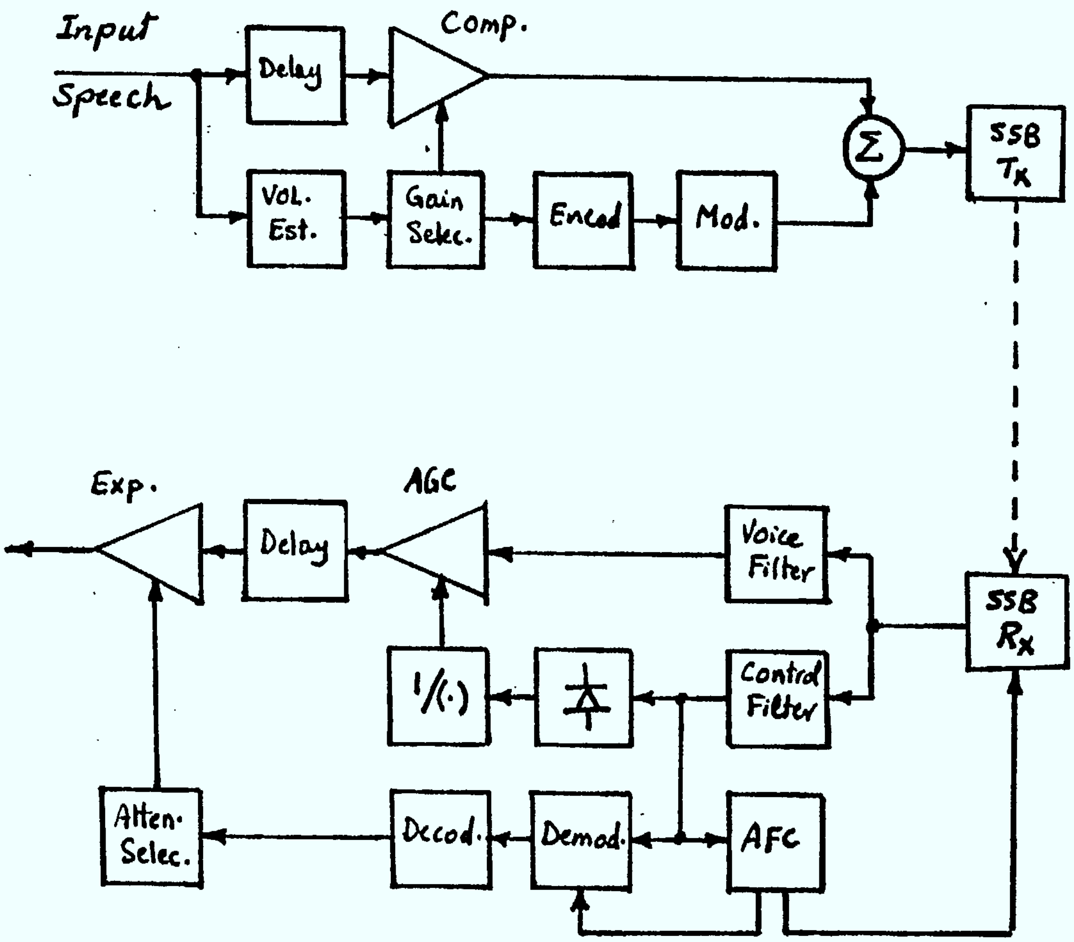


Fig. 2.5. - Functional Block Diagram of SYNCOMPEX System

Chapter 3

REAL-TIME CHANNEL SIMULATOR

3.1 Introduction

The performance of any system or sub-system, which is designed to work in the mobile environment, should be evaluated under realistic fading conditions. Therefore, it was decided that the first step into building a SYNCOMPEX system is to construct a reliable and flexible fading simulator. One that is capable of simulating a variety of signal impairments in real-time. The availability of such a simulator enables us to quickly evaluate the performance of any proposed circuit for AGC or AFC. The condition that the simulator should be a real-time one is important, since it would allow for a real-time testing of the system without the need for a cumbersome set-up to sample, store and process the signals on digital computers. It also allows for testing over long periods of fading and shadowing hence providing statistically valid results.

In this chapter, the theory, implementation and testing of a real time mobile channel simulator are described.

3.2 Theory

The transmitted signal, S(t), is assumed to be a bandpass signal centered at a carrier frequency, f_o, and can be represented by its low-pass equivalent signal, u(t), as follows:

$$S(t) = \mathrm{Re}\,[u(t) \exp(j2\pi f_o t)]$$

$$= u_i(t) \cos 2\pi f_o t - u_r(t) \sin 2\pi f_o t \qquad (1)$$

The low-pass equivalent signal, $u(t)$ represents the complex-valued envelope of the bandpass signal and can be expressed in its real and imaginary parts as:

$$u(t) = u_i(t) + j\, u_r(t) \qquad (2)$$

The channel model includes the following impairments:

(i) Additive Gaussian noise $n(t)$

(ii) Constant delay τ, and attenuation factor, α

(iii) Complex fading process, $g(t)$

(iv) Shadowing, $d(t)$

(v) Phase jitter, $\theta(t)$

(vi) Frequency offset and drift, $\Delta f(t)$

The channel model is shown in Figure 3.1 as a simplified block diagram that describes the relation between the transmitted signal, $S(t)$, and the received signal, $r(t)$. The fading process, $g(t)$, is a low pass complex envelope that gets multiplied by $u(t)$ to produce the low pass equivalent fading signal which could have either Rayleigh or Ricean distribution depending on the receiving conditions being simulated.

$n(t)$ is a bandpass noise process that can be expressed as:

$$n(t) = \mathrm{Re}\,[v(t) \exp(j2\pi f_o t)] \qquad (3)$$

Then, the received signal $r(t)$ is related to the transmitted

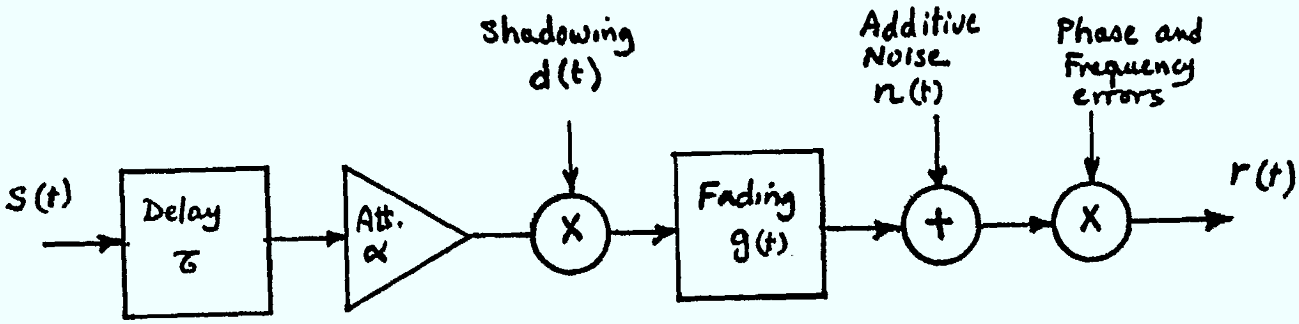


Fig. 3.1. – A Channel Model for Land Mobile Communications.

signal S(t) by the following equation:

$$r(t) = \alpha \cdot d(t) \cdot Re\left[g(t)\, u(t-\tau)\, e^{j[2\pi f_0(t-\tau) + \theta(t) + 2\pi\Delta f(t)]}\right] + Re\left[v(t)\, e^{j2\pi f_0 t}\right] \quad (4)$$

The constant time delay, τ, can be absorbed in the complex function g(t) without affecting its statistical characteristics. Also, θ(t) and $2\pi\Delta f$(t).t can be combined in one phase error term ϕ (t); then:

$$r(t) = \alpha \cdot d(t) \cdot Re\left[g(t) \cdot u(t)\, e^{j(2\pi f_0 t + \phi(t))}\right] + Re\left[v(t)\, e^{j2\pi f_0 t}\right] \quad (5)$$

$$= w_r(t) \cos 2\pi f_0 t - w_i(t) \sin 2\pi f_0 t$$

where

$$w_r(t) = Re\,[\alpha\, d(t) \quad g(t) \quad u(t) + v(t)]$$

and

$$w_i(t) = Im\,[\alpha\, d(t) \quad g(t) \quad u(t) + v(t)] \quad (6)$$

Equations (1-6) suggest the channel simulator model shown in Figure 3.2.

signal S(t) by the following equation:

$$r(t) = \alpha\, d(t)\, Re\left[g(t)\, u(t-\tau)\, e^{j\{2\pi f_o(t-\tau)+\theta(t)+2\pi\Delta f(t)\}} \right] + Re\left[v(t)\, e^{j2\pi f_o t} \right] \tag{4}$$

The constant time delay, , can be absorbed in the complex function g(t) without affecting its statistical characteristics. Also, (t) and 2 f(t).t can be combined in one phase error term $\phi(t)$; then:

$$\begin{aligned} r(t) &= \alpha\, d(t)\, Re\left[g(t)\cdot u(t)\, \exp\{ j(2\pi f_o t + \phi(t))\} \right] + Re\left[v(t)\, \exp(j2\pi f_o t) \right] \\ &= w_r(t) \cos 2\pi f_o t - w_i(t) \sin 2\pi f_o t \end{aligned} \tag{5}$$

where

$$w_r(t) = Re\,[\alpha \cdot d(t) \cdot g(t) \cdot u(t) + v(t)]$$

and

$$w_i(t) = Im\,[\alpha \cdot d(t) \cdot g(t) \cdot u(t) + v(t)] \tag{6}$$

Equations (1-6) suggest the channel simulator model shown in Figure 3.2.

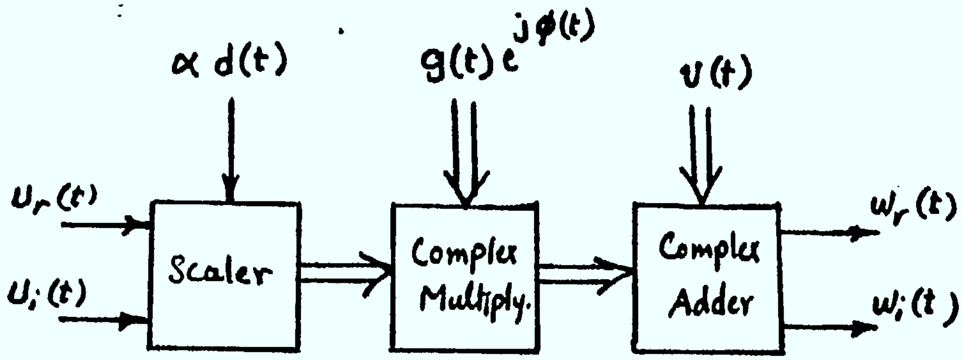


Fig. 3.2 – Channel Simulator Model

The input to the simulator is the real and imaginary components of the signal u(t), and the output is the real and imaginary components of the signal $w(t) = w_r(t) + j\, w_i(t)$. The first block ("The scaler") in the channel simulator performs the multiplication of the complex signal u(t) with the real signal $\alpha \cdot d(t)$, which represents the effects of shadowing (d(t)) and field strength (α). The second block ("The Complex Multiplier") simulates the effects of rapid fading (g(t)), phase jitter and frequency drift ($\phi(t)$). The last block simulates the additive noise impairment (v(t)).

The signals u(t) and w(t) are equivalent to the real signals S(t) and r(t), which means that processing these signals is exactly equivalent to processing the actual real signals. Moreover, the conversion from the low pass equivalent signal to the real signal and vice versa is quite straight forward as illustrated in Figure 3.3.

The implementation of the shadowing d(t), fading g(t) and noise v(t) requires five independent Gaussian Noise Generators (GNG), four filters (F) and one anti-log processor as shown in Figure 3.4.

The shadowing process is the slowest of the three processes followed by the fading, then the additive noise which is the fastest. Since the intent is to implement these processes using programmable digital signal processing chips, then the question of speed differences can be handled easily by adjusting the frequency of calling the various subroutines. The implementation details are given later.

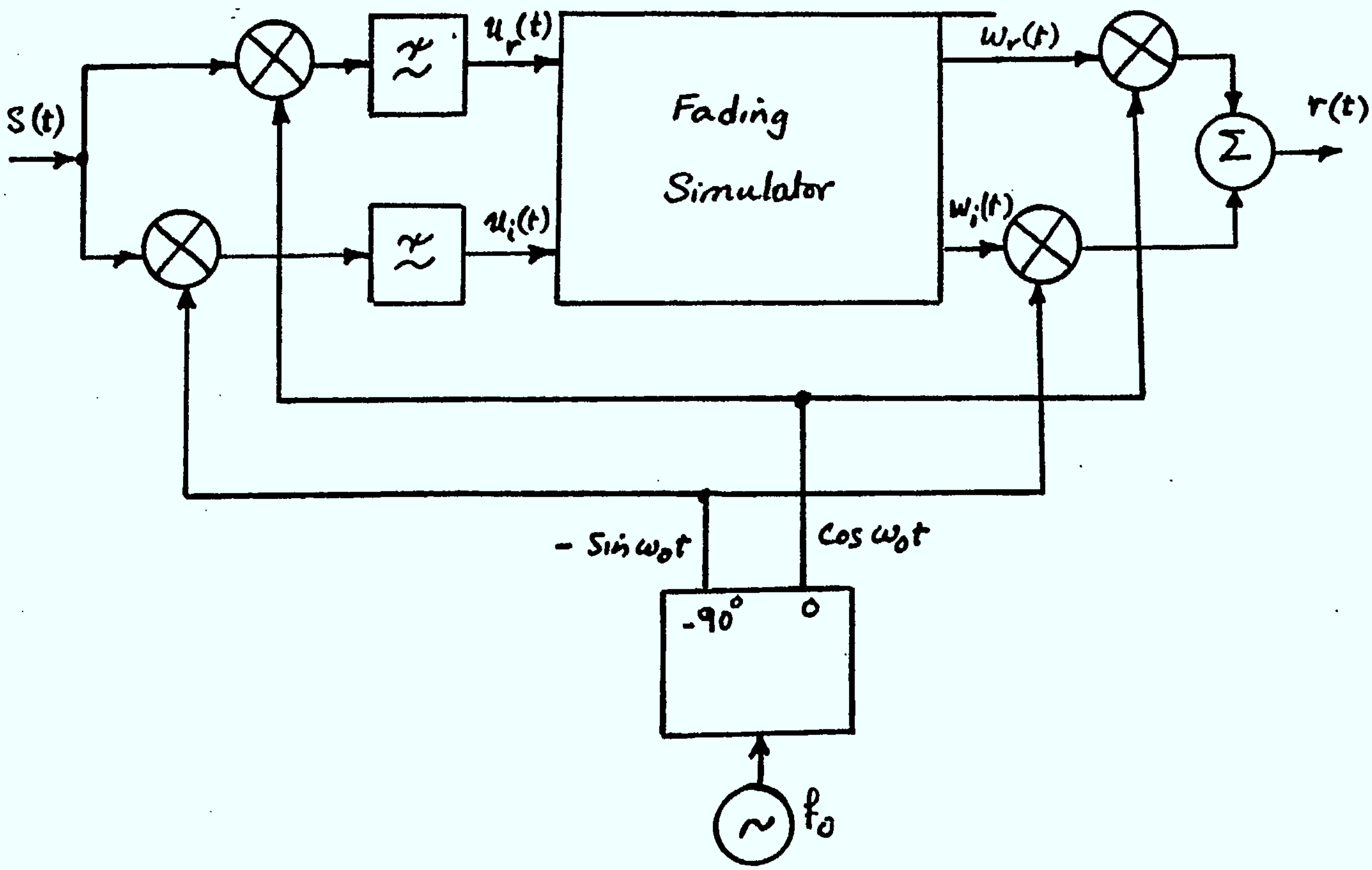


Fig. 3.3 - Conversion of Real Modulated Signals {S(t) and r(t)} into their Low Pass Equivalent Signals { u(t) and w(t) }

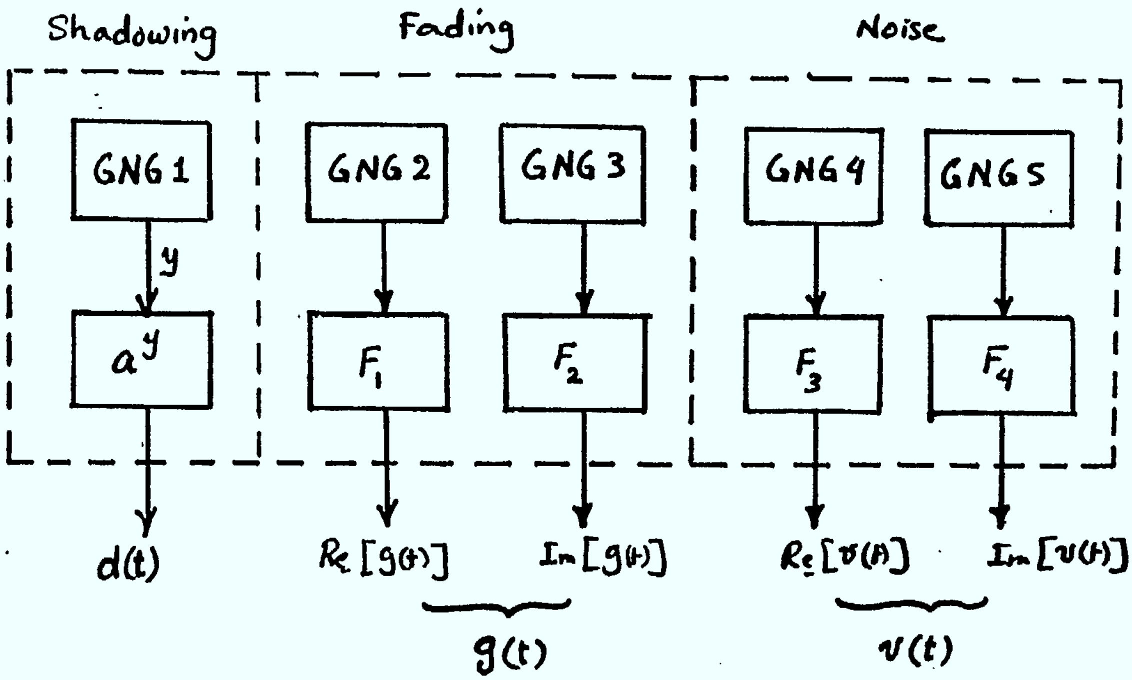


Fig. 3.4 - The Basic Blocks Needed to Simulate Shadowing, Fading and Additive Noise.

The shadowing generation routine involves two steps: first; a random number, y, is generated at regular intervals (say T_1). That number has a Gaussian probability density function and is equivalent to the logarithm of the shadowing level. So, to obtain the shadowing signal level an anti-log circuit (or program) is needed. The basis for the logarithm "a" does not change the statistical nature of the signal d(t). Changing the logarithm base is equivalent to multiplying d(t) by a constant.

The complex fading envelope generator consists of two identical and independent paths. Each path consists of a GNG and a low pass filter having the following transfer function

$$S(f) = 1/\left\{1-(f/f_m)^2\right\}^{1/2} \tag{7}$$

where f_m is the maximum Doppler frequency = λ/v

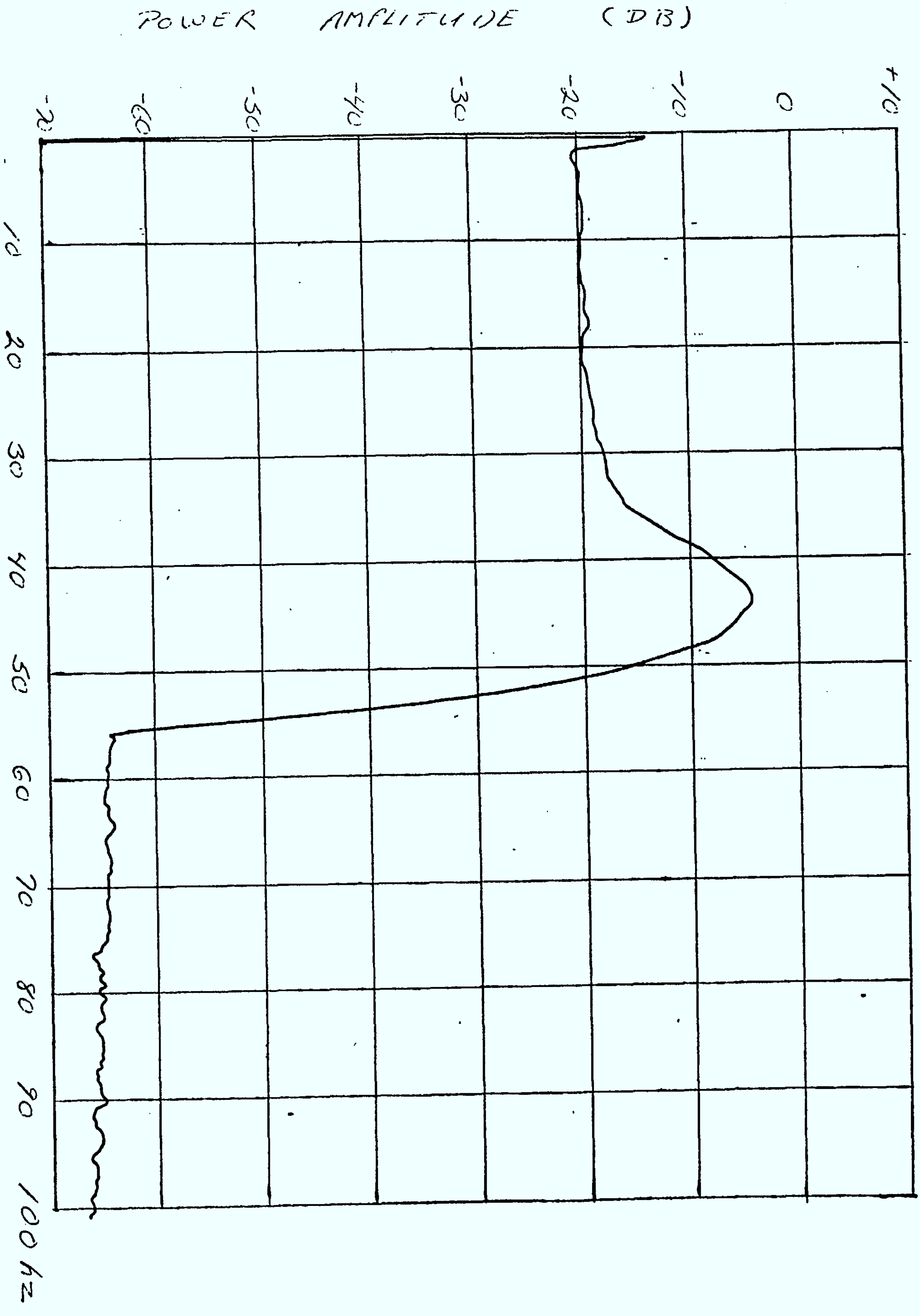
is the carrier wave length

and v is the car speed

Obviously such a filter is unrealizable and must be approximated. The approximated transfer function used in this study is shown in Figure 3.5. The selection of this approximated filter response was based on field measurement data collected under a previous contract [3].

3.3 Channel Simulator Parameters

The control channel of a SYNCOMPEX modem could be placed either in the mid-voice band or near the upper edge of the band. Each of the two locations has its advantages and disadvantages. The impact of the location on the design of the fading simulator


POWER AMPLITUDE (DB)
-70
-60
-50
-40
-30
-20
-10
0
+10
10
20
30
40
50
60
70
80
90
100 Hz
FREQ (HZ)

is minimal since the simulator is fully programmable. For the purpose of this study, we have selected the follwoing parameters for the control channel

(i) Carrier frequency: f_o = 3.1 KHz

(ii) Data rate: R = 100 bps

(iii) Modulation method: Minimum frequency shift keying (MSK)

3.4 Implementation of the Channel Simulator On TMS320

3.4.1 Introduction

Modern component technology has recently opened a floodgate of high performance components that have dramatically altered the cost/performance criteria for all digital systems. The new generation of high performance microprocessors and digital signal processors, such as the TMS 320, has made digital signal processing an affordable alternative to analog design.

Digital signal processing offers several advantages over analog signal processing. The flexibility, stability and reliability of digital systems are some of those advantages. Digital systems also eliminate the need of costly precision components and production returning so often required in analog systems. The rapid and continuous decrease in the cost of VLSI chips will make the digital alternative more economic in the near future.

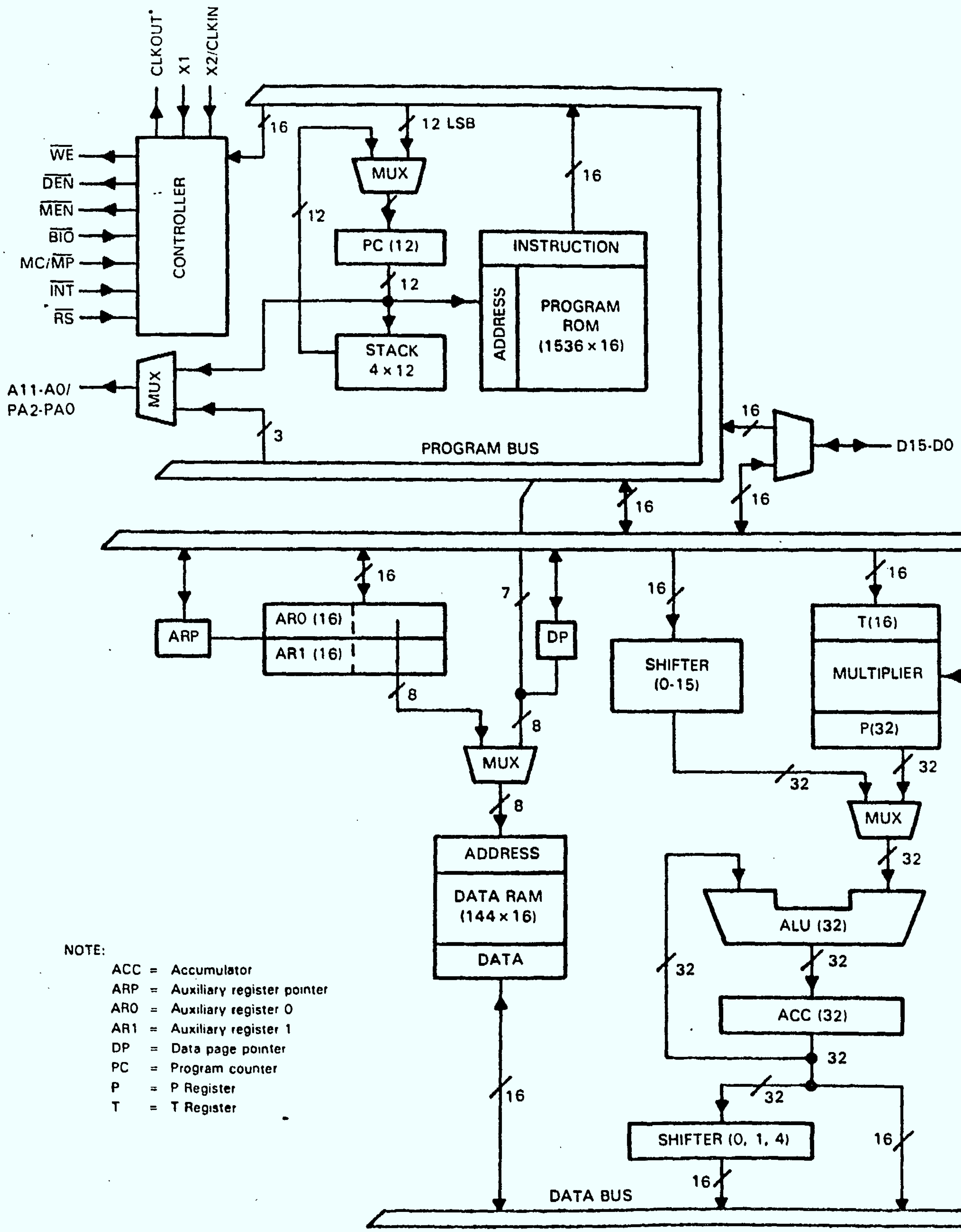
The new generation of digital signal processors includes the Intel 2920, the NEC UPD 7720, the DEC J-11, the Hitachi chip and the TI TMS320.

3.4.2 Texas Instrument TMS320 Digital Signal Processor

The TI TMS320 is a high performance microprocessor aimed at digital signal processing applications. It uses NMOS technology, clocked at 20 MHz. It achieves a 200-nanoseconds instruction execution time for more than 90% of the instruction set.

A block diagram of the internal architecture of the TMS320 is shown in Figure 3.6. It is a modified single-accumulator Harvard type architecture. Unlike a strict Harvard architecture, the TMS320 allows crossovers between the separate program and data memories. The program memory consists of 4K words, 16 bits each. 1.5K words of this program memory is an on-chip ROM in the micocomputer mode, whereas the remaining 2.5K is an external memory. In the microprocessor version of the TMS320, all the 4K program memory is external. The data memory is limited to 144 words, 16 bits each. It is external and is organized in two pages, the first contains 128 words while the second contains only 16 words in the present chips.

There are four basic arithmetic elements in the TMS320: the ALU, the accumulator, the multiplier, and the shifters. All arithmetic operations are performed using two's complement arithemtic. The ALU operates with a 32-bit data word; it adds, subtracts and performs logical operations. The 32-bit accumulator stores the output of the ALU and is also often an



NOTE:

ACC	=	Accumulator
ARP	=	Auxiliary register pointer
AR0	=	Auxiliary register 0
AR1	=	Auxiliary register 1
DP	=	Data page pointer
PC	=	Program counter
P	=	P Register
T	=	T Register

FIGURE [illegible] — BLOCK DIAGRAM OF THE TMS320M10

input to the ALU. The 16x16 bit parallel multiplier consists of three units: the T-register, a 16-bit register that stores the multiplicand, the P-register, a 32-bit register that stores the product, and the muliplier array. There are two shifters available for manipulating data: a Larret shifter for shifting data from the data RAM into the ALU and a parallel shifter for shifting the accumulator into the data RAM. The Larret shifter performs a left-shift of 0 to 15 places, zero-fills the low-order bits and sign-extends the 16-bit data memory word to 32 bits. The parallel shifter can execute shifts of 0, 1, or 4 only.

There are three addressing modes for the data memory: Direct, indirect and immediate. There are two auxillary registers, AR0 and AR1, which are used as the data memory address in the indirect mode. The program counter and stack enable the user to perform branches, subroutine calls, and interrupts. The stack is 12 bits wide and four layers deep.

A single interrupt system is implemented on the TMS320. An interrupt is generated by applying a signal to the INT pin. The interrupt system can be enabled or disabled by specific instructions (EINT, DINT). Another important external control is the BIO pin. It is an external pin which supports bit test and jump operations. It is very useful for monitoring peripheral device status. It is especially useful as an alternative to using an interrupt when it is necessary not to disturb time-critical loops.

The instruction set of the TMS320 is very powerful, specially for numeric-intensive operations such as digital signal processing. The instruction set contains a full set of branch instructions. Combined with the Boolean operations and shifters, these instructions permit bit test capability. Double-precision operations are also supported by the instruction set. Two other instructions, the IN and OUT are implemented to allow the access of eight external peripherals.

The main limitations of the current TMS320 chip are the size of the data RAM, the lack of some arithemtic and Boolean immediate instructions and the limitation of the auxillary registers to two. Another limitation is the parallel shifter which can execute shifts of 0, 1 or 4 only.

Nevertheless, the instruction set is capable of supporting up to 7 auxillary registers and up to 256 pages of data memory. New versions of the TMS320 overcoming some of these limitations, will be available in the near future.

3.4.3 Evaluation Module (EVM)

The TMS320 Evaluation Module (EVM) is a single board development system for the TMS320 signal processor. The EVM can stand alone as a development system, using the on-board text editor and audio cassette tape inerface as a mass storage media. Or, the EVM can accept text files from a host CPU. In both situations, the resident assembler will convert the incoming text into executable code. This object code is stored in a 4K word memory space.

The EVM operating system firmware resides in EPROM and can be divided into 4 main segments: debug monitor, assembler/reverse assembler, text editor, and 2764 EPROM utility.

The EVM contains two processors configured in a master-slave configuration. The on-board TMS320 is used to execute the user's code in real time. The TMS9995 processor, functioning in the role of master, executes the operating firmware.

3.4.4 Analog Interface Board (AIB)

The TMS320 Analog Interface Board (AIB) includes a 12-bit analog to digital converter, a 12-bit digital to analog converter, two 16-bit input ports and one 16-bit output port. The board also includes two reconfigurable blow pass filters, an audio amplifier and a wire-wrap area. A programmable counter is also available for generating the sampling clock from the 20 MHz chip clock. The counter divides by the powers of two, only.

3.4.5 Test Signal Generation

The test signal is a complex FSK signal, multiplied by a complex fading process. Gaussian noise is then added to both real and imaginary parts.

3.4.5.1 FSK Signal Generation

There are several ways by which an FSK signal could be generated. The following section describes some of these methods. Each method is evaluated in order to choose the most appropriate one for our specific application.

Analog Generation

In this method, two analog oscillators are used to generate sin f_1 and sin f_2. Two single tone 90 degrees phase shift circuits are used to generate cos f_1 and cos f_2. Two analog multiplexers select either f_1 or f_2 to be connected to the A/D converters. The selection is controlled by a pseudo random sequence generator. A block diagram of this method is shown in Figure 3.7.

The circuits required for this method are relatively complex due to the A/D converters and their control circuits. There are many sources of error, mainly the 12-bit limitation of the A/D. The main advantage is the flexibility in th choice of f_1 and f_2. They can also be varied while the program is running to allow the test of the AFC circuits.

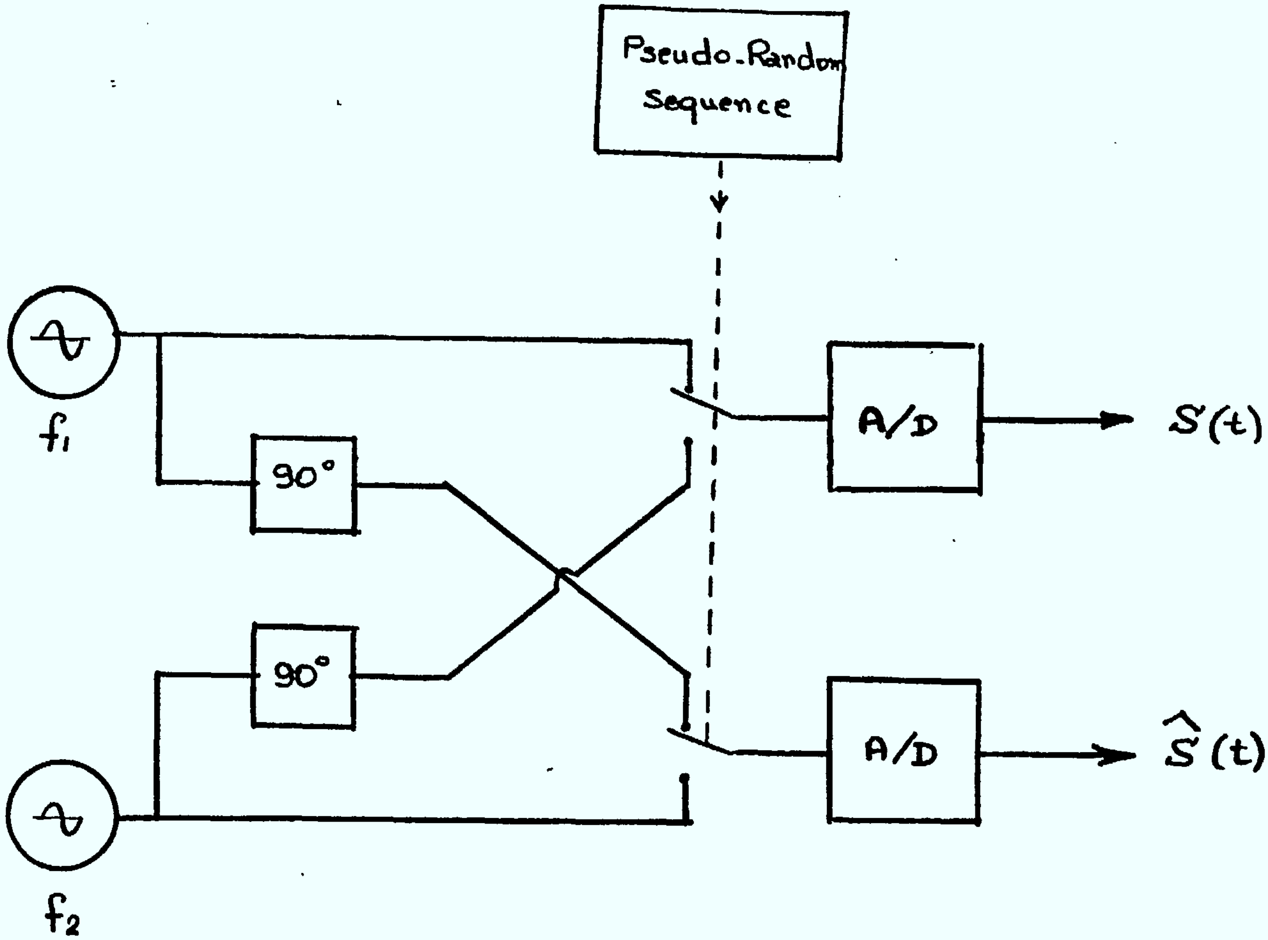


Fig. 3.7- Hardware Implementation of FSK Modulation in Quadrature Form

Digital Generation

The other alternative is to generate the FSK signal, digitally, inside the TMS320. This eliminates the requirement for external circuitry, on the other hand, it consumes some of the TMS320 resources.

In order to generate one sample of the FSK signal, an angle has to be calculated, as well as its Sine and Cosine. The angle is the product of two terms: "w", representing the frequency, and "t", the time. "w" in our case can take two values, one representing f_1 and the other representing f_2. The selection between them is done according to a pseudo random sequence applied to the BIO pin of the TMS320. The time is generated using an external clock connected to the interrupt pin. Every time an interrupt occurs, the previous time t is incremented by the clock period, multiplied by "w", then the sine and cosine of the angle are calculated.

The sine and cosine could be calculated on the fly using the power series or could be read from a lookup table. A 65 instructions program was developed to calculate both the sine and the cosine of an angle, using the in-line programming technique to avoid the time consuming branch instructions. The execution time of the program is 12 microseconds, and it calculates the sine and cosine with a precision of 0.05%. Given that our sampling frequency is 10 KHz, the time allowed for calculations is 100 microseconds. So, the program consumes 12% of the time resource, it uses 2% of the program memory. The program also uses 10 data memory locations, which represents 7% of the data memory resource. These locations are used just as scratch-pad memory, i.e., they can be reused by the other programs.

The other alternative for generating the sine and cosine is by using a look-up table. The angle is sent to an output port connected to the address bus of a memory. The sine and cosine are prestored in the appropriate locations. The total time needed is 2.3 microseconds which represents a saving of 9.7 microseconds over the previous alternative.

3.4.5.2 Noise Generation

AWGN is the key process to the channel impairments of the bandpass model. Several methods for obtaining a sample sequence of this process were considered.

The most straightforward is to simply prestore samples of the WGN sequence, obtainable by any number of methods, in ROM and subsequently read back the samples as required, during execution time. Memory and cost limitations restrict this approach. Allowing for program space, approximately 3K words are available for sample storage. For a noise process white to 4KHz, an 8KHz plus sample rate is required. This would exhaust the stored noise samples in 3/8 = 0.37 seconds maximum.

The noise process would constantly be repeating itself every 0.37 seconds, an inadequate mode. This is shown in Figure 3.8.

The address space limitation may be circumvented by interfacing an external stack of arbitrary lengthened poping

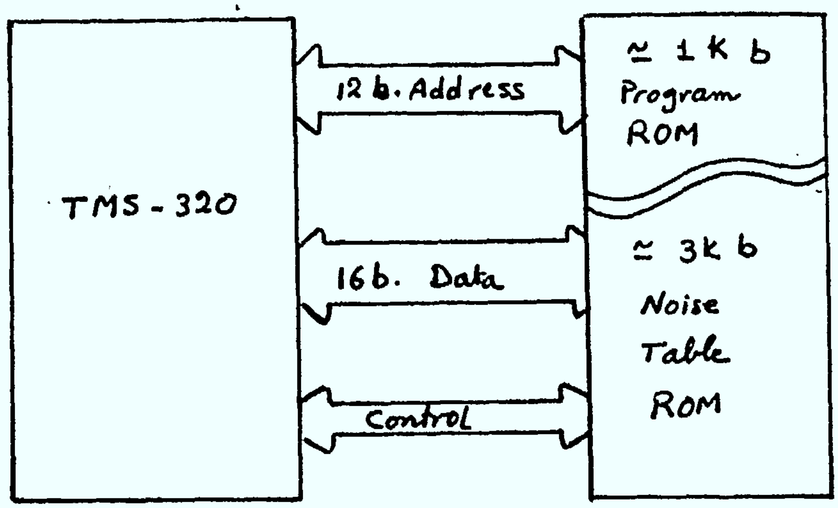


Fig. 3.8 - Prestored Noise Samples Method
(Internal Memory)

samples as required (Figure 3.9). Trivial circuitry can provide automatic addressing for use in conjunction with one of the post input instructions. Unfortunately, a desirable sequence duration of one-half hour requires 30x60x3K/.37 sec. 50 megawords of ROM.

The alternative to the stored sample approaches outlined above is real-time generation of the samples as required. Two sub-categories suggest themselves: (1) analog and digital combinations; (2) completely digital.

A couple of possibilities for A/D combinations are shown in Figure 3.10. In the top half it is assumed an analog recording has been made by empirical means. This is then digitized and fed to the processor. Alternately, use can be made of the fact that many real word processes generate Gaussian distributed noise, such as diode shut noise. This is shown in the lower part of Figure 3.10, where A is a highly stable (with temperature, etc.) amplifier. Since the stability of A is a difficult point, and the facilities to realize the recording method were unavailable, PN sequence generators were considered as a less straightforward but possibly more easily implementable method of generating white Gaussian noise.

A PN sequence generator is easily constructed from a shift register with a couple of feeback taps combined by moduls 2 additions.

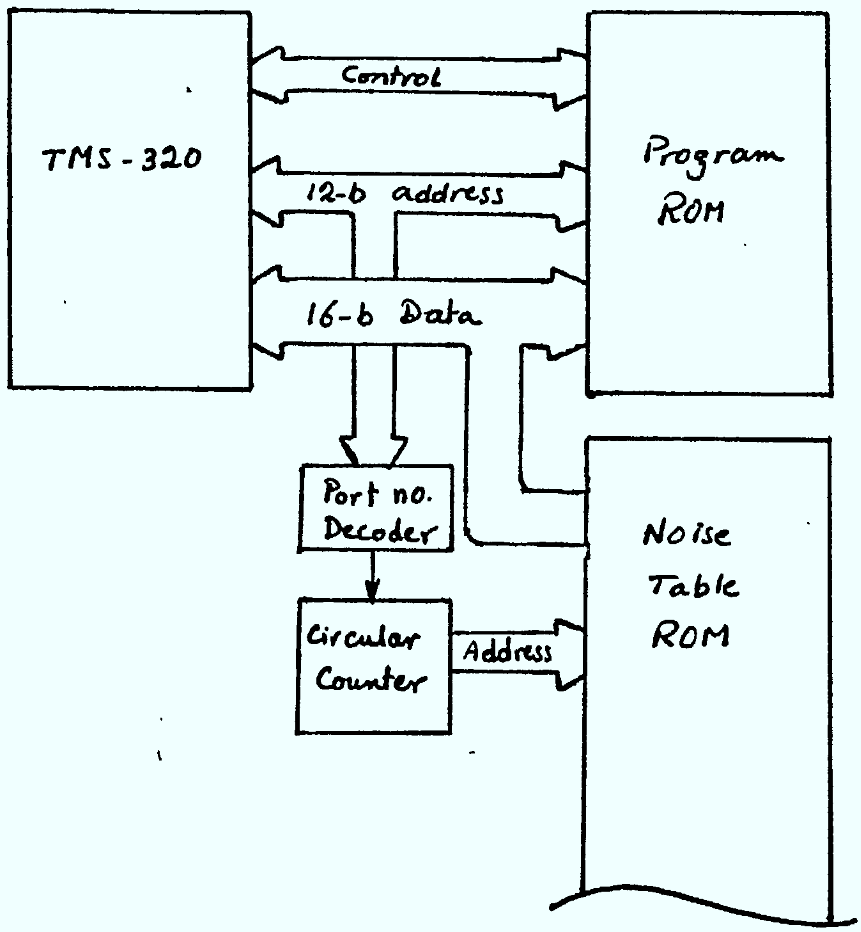


Fig. 3.9 - The Use of External Memory for Noise Table

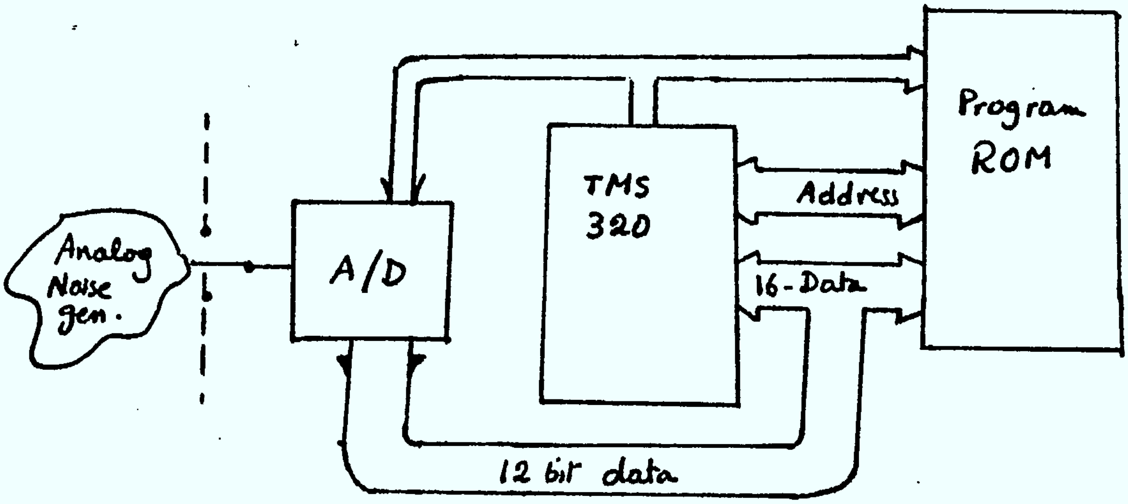


Fig. 3.10. Analog Noise Generation

In practice a long shift register (say 30 stages) is required. The spectrum of the output is fairly white to one half the clock frequency. Twelve such generators operating independently (similar but not identical construction producing different sequences) will yield a good approximation to the desired Gaussian distribution if their outputs are summed. This is by virtue of the Central Limit Theorem.

In Figure 3.11 one obvious method of performing this summation is shown. The analog output of this circuit would be digitized as per Figure 3.9. Many variations of this circuit to numerous to delineate present themselves the general idea being to change the dividing line between the analog and digital portions of the circuit. For example shift registers could collect binary words from the PN sequence generators, then the filters and op amp could be replaced by a digital adder and one digital filter.

The all digital circuit need not actually be constructed since the TMS320 instruction set includes all the necessary logical operations to simulate the circuits operation. (Of course it is also possible to simulate analog implementations in a less direct manner).

The idea of an all software impelementation has particular appeal because it is completely repeatable without additional construction efforts. Furthermore, a much less obtuse method of generating a WN and hence WGN sequence is available. This is the Linear Congenetial Generator* which has the form:

$$Z = \{Z_0,\ Z_i = A\ Z_{i-1} + C\ (\mathrm{mod}M);\ i=1,2,\ldots\}.$$

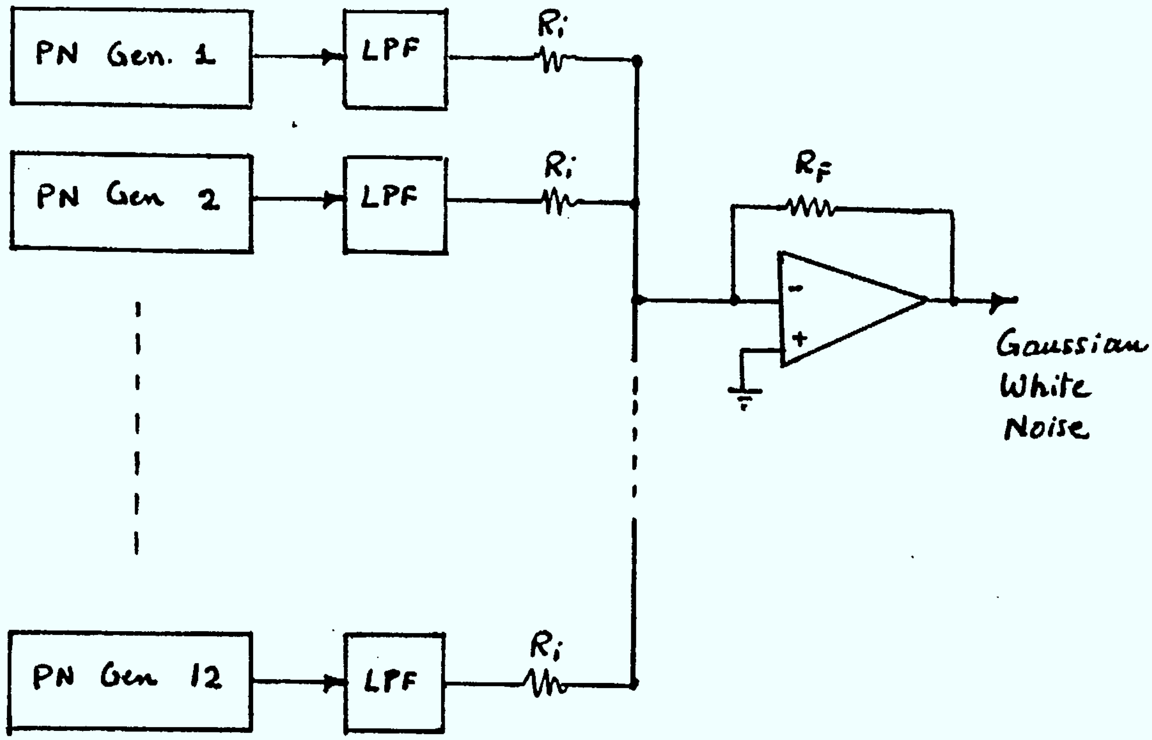


Fig. 3.11 - Generation of Noise with Gaussian Distribution

Z is a sequence of non-negative integers. The properties of Z are determined by the multiplier A, the seed Z_0, and the increment C. Their proper selection is critical, having been repeatedly discussed in the literature for a period of over twenty years. The fundamental problem is that the completely deterministic sequence Z must appear random. Fishman and Moore present in their paper, [4] an "empirical evaluation of 16 selected values for the multiplier A in the prime-modules multiplicative congruential random number generator" and also a comparison to Marsaglia's generator.*

Marsaglia's generator is particularly interesting because its period $T=2_{30}$ and it may be implemented conveniently on a 32 bit machine. This is important because the mod M** operation then reduces to shifts. This is not true in the case of a less carefuly chosen modulus M, and a time-consuming divide routine would be required.

Although the TMS320 is not a true 32 bit machine, it can with a little effort be made to appear so. There is a difficulty in that the hardware of the 320 treats its 16 bit operands as inherently two's complement numbers. This means there are really only 15 bits in each word to represent positive numbers. The 16th bit is considered by the hardware as a sign bit. Thus, straightforward concatenation of two 16 bit words to achieve a 32 bit double-word is not possible. What are required are 32 bit true binary double words for Z and A. This is contrary to the case of more general purpose processors (e.e.g, 8085A) where the number representation is solely the programmer's choice.

However, these apsects were dealth with and the algorithm implemented. It takes 3.4 microseconds to produce each noise sample. These samples ae white noise and summing twelve of them gives Gaussian white noise.

Testing

The sequence Z is over 1 billion samples long. At an 8 KHz rate, it would take over 37 hours to exhaust the sequence. Manual examination of all the Zi's was not feasible. The 320 programming language is low level assembly code and hence errors can be subtle. It was decided to first implement Marsaglia's generator in Fortran on a micro to get output that need not be suspect. A file was created of the first 100 output Z's and stored for later comparison. A couple of other Fortran programs were written to examine the distribution of the generator program and print out plots. This was done to provide a further basis for comparison. Typically the distribution plots would be prepared by examining several hundred thousand Z's and take all night to run on the TI Professional Computer (PC).

Next the 320 assembly code version was optimized for execution speed and the first 100 Z's stored in the on chip data RAM. The PC, which had been interfaced to the EVM board through a RS232 link, was able to take these data RAM contents and store them in a file accessible by high-level language programs.

More Fortran programs were written to perform the necessary format conversions before the raw data could be printed out as 32 bit true numbers represented in base 10. This final file was compared to the original Fortran noise generator and found to be identical.

Distribution tests were still in order. Another 320 assembly program was written to create a distribution file in 320 data RAM, identical in function to the Fortran distribution program. This was required because the Fortran program could not possibly keep up with the high speed 320 output. To prepare the distribution file of several hundred thousand samples took approximately 10 minutes execution time for the assembly code version. File transfer to the PC and suitable format conversions allowed a plot of the distribution files. The familiar white and Guassian distributions are shown in the plots of Figure 3.12 and Figure 3.13 respectively.

Application

The Fading Simulator is a complex process whose I and Q channels require the addition of independent Gaussian white noise signals. One generator was used to provide the I and Q signals by noting that consequtive sample are essentially uncorrelated, and since their distribution is Gaussian, they are also independent. Every second generator output is arbitrarily assigned to the I channel and every other assigned to the Q channel. To maintain the 4KHz bandwidth the generator must operate twice at 16 KHz.

S/N Ratio

S/N ratio is controlled by scaling the WGn samples before they are added to the I and Q signals by a constant. This constant may be changed as desired in the source code or at execution time.

Fading Process

The fading process is the most significant natural detriment to the mobile radio channel. It is due to the vector addition of multiple wavefronts arising from reflections off streets, buildings, other vehicles, etc. The power and rate of the fading depends on the local geometry, transmission frequency, and the vehicles velocity.

The power spectrum of the Rayleigh fading process is described by

Assuming linear phase, this may be converted to a magnitude specification of the form

which will have the general appearance of Figure 8. Figure 9 shows a practical approximation to this response.

Chapter 4

AGC AND AFC CIRCUITS

4.1 Introduction

The performance of SSB modulation systems under fading conditions depends critically on the performance of the AGC and AFC subsystems. The AGC circuit is required not only to adjust the receiver gain on a long term basis, but rather to track and correct the rapid amplitude fluctuations due to short term fading. The AFC is required to remove any frequency offset between the transmit and receive carriers.

In a SYNCOMPEX system, the control signal (which is a constant-envelope digitally-modulated signal) is used to perform both functions. The combined voice/control received signal is first down-converted to baseband; then the control signal is extracted by a BPF and passed through two parallel processors; one for the AGC and the other for AFC.

In the AGC circuit, the envelope of the control signal is to be recovered and fed-forward to correct the amplitude of the information signal; while in the AFC circuit the control signal is first hard-limited and then fed to a frequency discriminator followed by a processor that determines the frequency offset and generates an error signal to control the IF-VCO.

The approach taken in this study is to implement both the AGC and AFC on the TMS320, and then test them using the fading simulator described in Chapter 3.

4.2 The AGC Subsystem

In the SYNCOMPEX system (where the control signal is placed above the voice information signal), there are three important considerations regarding the AGC operation:

(i) The correlation between the fading envelope impressed on the control signal and the fading envelope of the information signal.

(ii) The differential delay between the information and control processings.

(iii) The dynamic range that can be tracked by the AGC subsystem.

4.2.1 Control/Information Envelopes Correlation

Since the control signal and the information signal cover two slightly different frequency ranges, one would expect a small difference between the fading processes impr~~ov~~essed on the two signals by the channel. The correlation degree between the two fading processes (in the control channel and in the information channel) depends on the delay spread encountered in receiving the combined signal. Most of the time, the signal arrives at the receiver from scatterers in the immediae vicinity of the receiver. This will result in a very small delay spread (in the order of $0.2\ \mu$sec). Under these conditions the fading process will be flat over the narrow bandwidth allocated for the SSB channel. This means that the fading envelope imposed on the control signal will be almost identical to that envelope imposed on the information channel. Sometimes, however, part of the received signal comes from objects located far from the receiver; this will lead to a larger delay spread and increases the decorrelation of the two envelopes. In order to quantitatively estimate the correlation between the control and information fading envelopes we develop the following simplified model: Refer to Figure 4.1. The control channel is very narrow compared to the information channel and the carrier frequency so that it can be considered as a tone with frequency f_o. The information channel spans the range (f_1, f_2) which is about 2.8 KHz. Under multi-path interference conditions the fading envelopes of f_o and f_2 will have a stronger correlation than those of f_o and f_1. Then the worst correlation case can be presented by the correlation between two tones at f_o and f_1 (with a frequency difference of Δf) at the output of a delay line of T second delay as shown in Figure 4.2. T represents the RMS delay spread in the channel.

The figure represents a very simple model of the channel where the receiver intercepts only two components of the signal which are equal in magnitude and separated in time by a oonstant differential delay of T seconds. The envelope of the received signal at either f_o or f_1 has the following form:

$$E_i = \text{envelope} \left[\cos 2\pi f_i t + \cos 2\pi f_i (t-T)\right]$$

or

$$E_i = \left[2\,(1+\cos 2\pi f_i T)\right]^{1/2} \quad , \quad i= 0 \text{ or } 1 \tag{4.1}$$

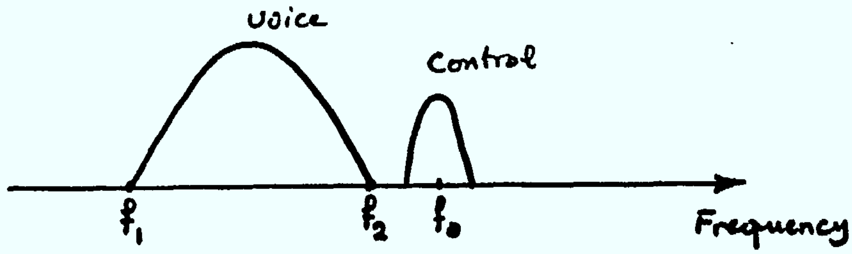


Fig. 4.1. - The Combined Spectrum of the Received Signal

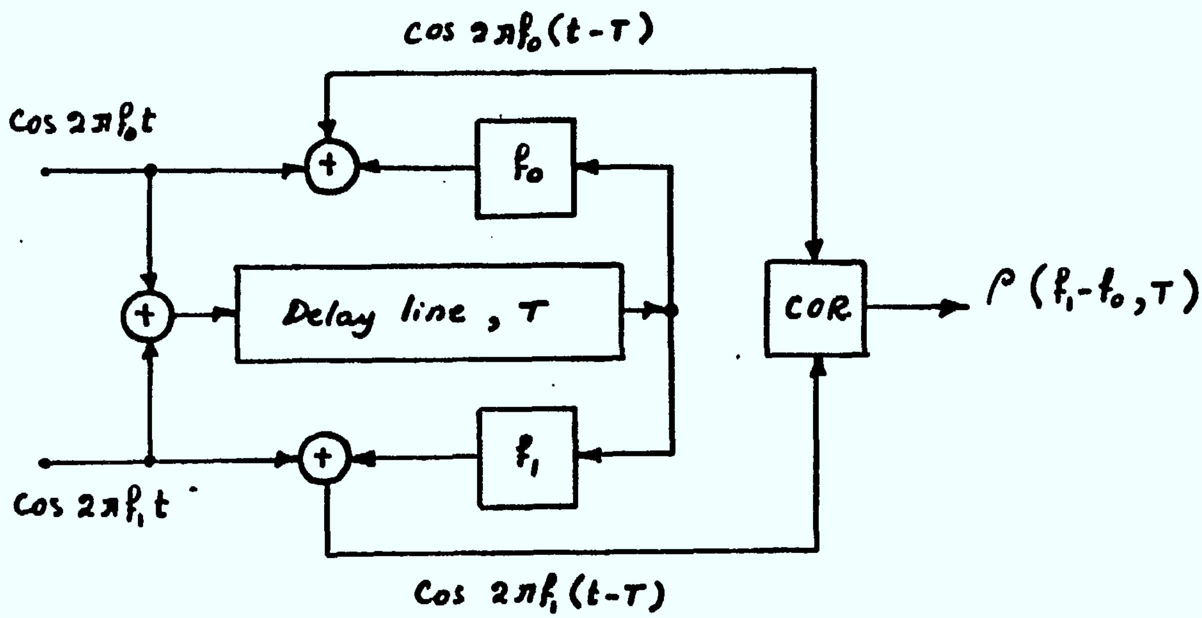


Fig. 4.2. - A Model Representing Worst case Envelope Decorrelation

The derivative of E_i with respect to the frequency f_i is:

$$\frac{\partial E_i}{\partial f_i} = \frac{1}{2} \cdot \frac{-4\pi T \sin 2\pi f_i T}{[2(1+\cos 2\pi f_i T)]^{1/2}} \tag{4.2}$$

Then, the ratio between the envelope of the control channel E_o and that of the information channel E_1 is:

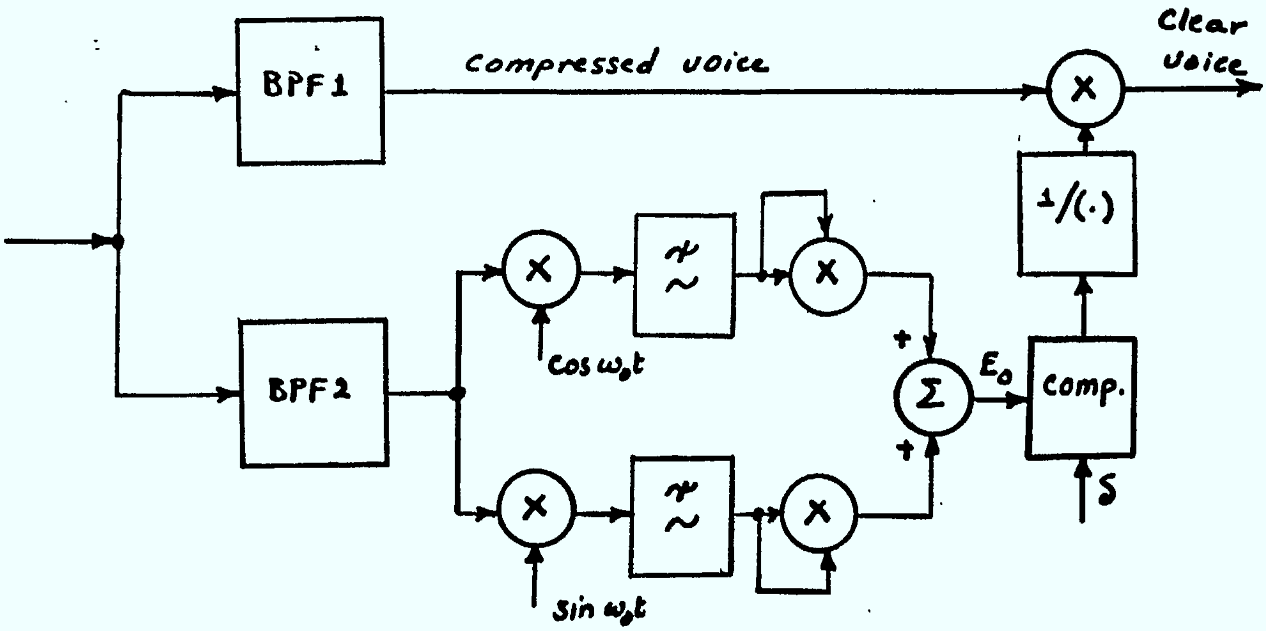
$$\frac{E_1}{E_o} = \frac{E_o + \Delta E}{E_o} = 1 + \frac{\Delta E}{E_o}$$

or

$$\frac{E_1}{E_o} = 1 - \left(\frac{\sin 2\pi f_o T}{1+\cos 2\pi f_o T}\right)(\pi T \Delta f) \tag{4.3}$$

Equation (4.3) points to an important problem in using the control signal envelope to control the gain of the information signal. Since the two envelopes are slightly different (although the difference is very small) and since the envelope of the control signal goes to zero for "f_oT" odd integer. Then, the ratio E_1/E_o could go to infinity. This can be corrected by limiting the minimum value of E_o to a small value δ. Then Equation (4.3) will be valid for the entire range of "f_oT" except when $E_o < \delta$.

Figure 4.3 illustrates the general block diagram of the AGC operation. The input signal is the combined information/control signal which is assumed to have no frequency offset. The information and control signals are separated by means of the two parallel filters, BPF1 for information and BPF2 for control. The envelope of the control signal, E_o, is then detected. E_o is a function of time since the details of the propagation delay spread varies as the receiver moves in the multipath field. $E_o(t)$ is then compared with δ and the largest one is fed to the inverse circuit (denoted in the figure by 1/(.)). The output of



f_o = Center frequency of the FSK signal

Fig. 4.3. - A Complete Block Diagram of the AGC Circuit

the inverse circuit ($1/E_o(t)$ or $1/\delta$) is then multiplied by the information signal in synchronization. Except for the times when E_o is small (in the order of δ or less) the amplitude fluctuations due to fading will be removed from the information signal.

4.2.2 Differential Delay

The differential delay between the information and control signals is caused by:

(i) Different group delay in the receiver IF filter, and

(ii) Different processing delays in the information and control processing circuits.

The first point is a serious one, especially if the IF bandwidth is narrow compared to the combined voice/control baseband frequency bandwidth. This problem can be minimized if the control signal is placed in the mid-band location. This would require notching out part of the speech spectrum and placing the control signal there. The subjective effect of such a technique on the speech quality is not clear and can only be tested experimentally. An alternative technique is to keep the control signal out of the speech spectrum and to estimate the differential delay at the beginning of each talk spurt transmission. The details of these two techniques will be discussed with some details in later sections.

The differential delay due to different processing stages for voices and control signals can be easily accounted for during the design phase. Digital implementation of the ACSB modem makes that task particularly easy.

4.2.3 The AGC Dynamic Range

In the mobile environment the fading envelope could sweep a dynamic range of more than 60 dB. In fact the signal strength could vary from +15dB above the RMS value down to almost complete interference (i.e., well below the noise level). Obviously, we should not be interested in signal level below the minimum detectable level, and that would limit the practical dynamic range to approximately 15 dB plus the average signal to noise ratio. Therefore, the practical dynamic range would be in the order of 40dB or less.

In order to analyze the required dynamic range in a quantitative manner we should start by examining the probability distribution of the signal level.

Taking into account both shadowing and fading, the AGC operation should be performed in two stages. In the first stage the slow shadowing process is to be corrected by a slow acting AGC circuit. This could be an ordinary analog AGC circuit using long term integration (in the order of seconds say) and having a dynamic range compatible with the variance of the log normal shadowing distribution (6-10dB). In the second stage a fast acting forward AGC circuit is to be used to correct the fast Rayleigh fading. The cumulative probability distribution of a Rayleigh envelop R is:

$$F_R(r) = 1-e^{-(r_1/r_o)^2} \quad (4.4)$$

where r_o is the RMS signal level. If we take r_o as a reference voltage for the level of the received signal (i.e., the 0dB level) then the fast acting AGC dynamic range around that level can be calculated as in Equation (4.4). In particular, a signal level exceeding the +10dB mark occurs with a very low probability (4.54×10^{-5}), so the AGC circuit does not have to track levels higher than 10dB above the RMS level. On the -ve dB side, the probability that the signal level drops to the practical range of the noise level (-15 to -30dB) is significant. However, these levels represent noisy signals and tracking them down will result in a very little improvement in the signal quality, so a reasonable lower limit on the AGC dynamic range could be (-10 to -15dB). Thus, a somewhat conservative selection of the AGC dynamic range would be (+10 to -15) dB, which would cover range of received levels nearly 97% of the time. A dynamic range of 25dB can be handled easily using a 16 bit processor such as the TMS320.

The fast acting AGC circuit shown in Figure 4.3 which operates on the basis of feed-forward gain correction should be aided by a slow acting AGC circuit placed at an early stage in the receiver. This slow acting AGC would be similar to any ordinary AGC circuit and would operate on a long term averaging basis. The purpose of this second circuit is to reduce the overall dynamic range of the received faded signal by adjusting the gain to counteract the slow fading and shadowing.

4.3 The AGC Circuit

One of the main disadvantages of SSB is the difficulty encountered in extracting the carrier frequency and phase information from the modulated signal. To illustrate the effects of phase and frequency errors on the demodulation process we consider the following simple analysis:

Assume that the received SSB signal is given by

$$S(t) = g(t)\cos(2\pi f_0 t) - g'(t)\sin(2\pi f_0 t) \tag{4.5}$$

where $g(t)$ and $g'(t)$ are the real and imaginary parts of the baseband signal, and f_0 is the carrier frequency. Let the locally generated carrier signal be

$$l(t) = \cos(2\pi f_0 t + 2\pi \Delta f t + \Theta) \tag{4.6}$$

where Δf is the frequency error and Θ is the phase error. Multiplying Equations (4.5) and (4.6) gives

$$r_2(t) = 1/2\,[g(t)\cos(2\pi\Delta f t) - g'(t)\sin(2\pi\Delta f t)] \tag{4.7}$$

where the double-carrier frequency terms have been omitted. To examine the effect of phase error along we set the frequency error to equal zero, then we have the following equation

$$r_1(t) = 1/2\,[g(t)\cos\Theta - g'(t)\sin\Theta] \tag{4.8}$$

The first term in Equation (4.8) represents the required output multiplying by $\cos(\Theta)$, and the second term is the distortion due to the phase error. Notice that if the phase error is exactly 90 degrees the real part of the signal disappears and only the imaginary part remains. This, however, does not affect the perception of the voice signal since the real and the imaginary parts have the same information. But, the mixture of the two parts represents a phase distortion in the baseband signal.

It turned out that the human ear is relatively insensitive to phase changes in the speech signals and thus this phase distortion in SSB demodulation is quite tolerable for voice communications. In fact, even slowly varying phase error are tolerable for voice.

Proceeding in the same way to see the effects of frequency error, we set the phase error equal zero and Equation (4.7) becomes

$$r_2(t) = 1/2\ [g(t) \cos (2\pi\Delta ft) - g'(t) \sin (2\pi\Delta ft)] \qquad (4.9)$$

Thus frequency errors give rise to spectral shifts as well as phase distortion in the demodulated output. As long as these spectral shifts are small they can be tolerated in voice communications. Large frequency shifts, however, lead to a perceptual distortion in the received speech, and hence should be avoided.

We conclude from the previous discussion that the demodulation of a speech SSB signal requires that the difference in frequency between the transmit and the receive carriers should be kept as small as possible. We also conclude that phase coherency is not a strict requirement in SSB voice communications.

In VHF/UHF radio communications, the frequency error due to carrier instability could be as large as one or two KHz. Such a large difference in frequency between the transmit and receive carriers would lead to a considerable distortion in the received signals. Moreover, in a SYNCOMPEX modem, with a control channel placed above the voice baseband spectrum such a spectral shift may push the control channel out of the IF filter passband resulting in a total loss of the control data. Therefore, a mechanism should be devised to:

(i) Ensure that during the frequency acquisition mode the received signal remains within the IF filter passband;

and (ii) Remove the frequency offset between the receive and the transmit carriers.

In this section we shall only concentrate on the frequency tracking of the received signal, i.e., we shall assume that the control channel falls within the IF filter passband, and that the frequency error between the transmit and the receive carriers is small.

4.3.1 Frequency Discriminators

Since we are only interested in tracking the carrier frequency rather than the carrier phase, then the circuit required could be relatively simple. Frequency discriminators, which are being used widely in demodulating FM signals, can be used to estimate the frequency difference between a locally generated carrier and the center frequency of the control channel. This could be done in the baseband using the TMS320 processor. A signal proportional to the frequency error could, then, be fed back to the IF stage to adjust the local IF carrier. This requires that the IF carrier be tunable over a range comparable to the expected frequency offset in the radio channel.

The AFC system is shown schematically in Figure 4.4. The received IF signal is down-converted using a local carrier tuned approximately at the received IF carrier of the combined voice/data signal. The down-converted signal is then lowpass-filtered, sampled and passed to the TMS320 processor. The control signal is then filtered and split into two paths; one for the AGC circuit and the other for the AFC circuit. The part that is fed to the AFC circuit is to be first band-limited then fed to the frequency discriminator to calculate the frequency error. A dc signal proportional to the frequency error is then to be produced at the discriminator output for controlling the IF VCO frequency.

There are several approaches for implementing the frequency discriminator, the simplest one is a linear filter with phase differentiation characteristics. This can be implemented either directly as a differentiator, or as two BPF's tuned at two different frequencies. The two methods are equivalent and are shown in Figure 4.5.

The advantage of the frequency discriminator described in the previous paragraph is its simplicity. The performance of the

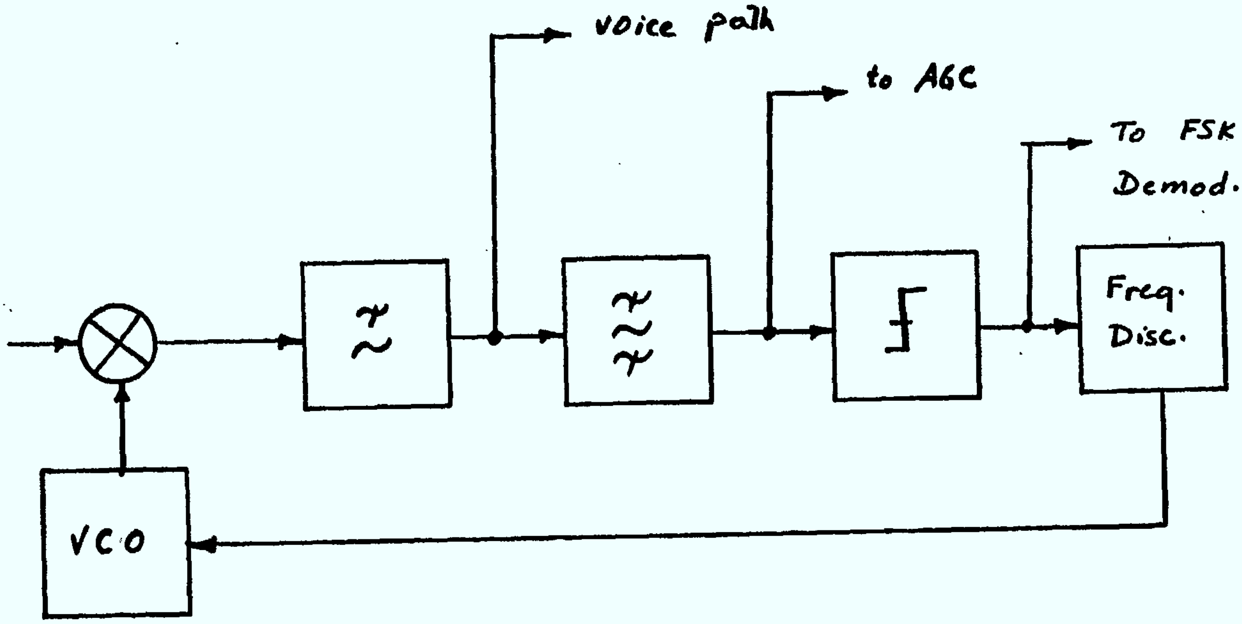


Fig. 4.4 - Block Diagram of the AFC circuit

scheme under fading condition should be tested using the fading simulator described in the previous chapter. However, one can make few predictions on the effect of fading using the following analysis:

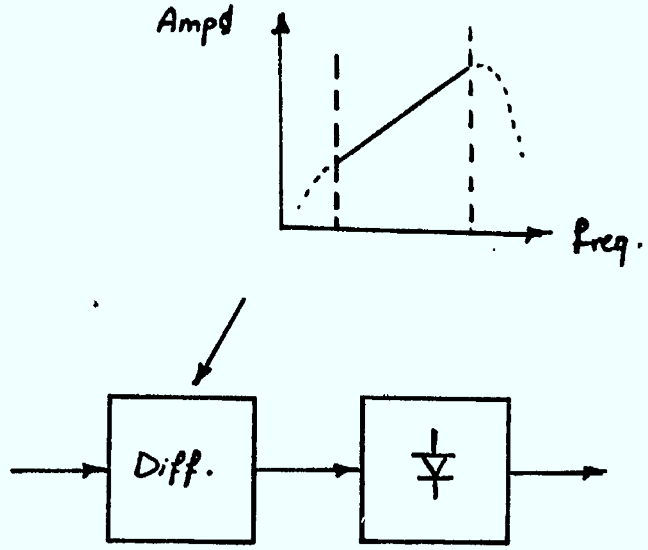
Under fading conditions the received signal can be expressed as:

$$x(t) = R(t)\cos(2\pi f_o t + 2\pi\Delta f t + \phi(t) + \Theta(t)) + n(t) \qquad (4.10)$$

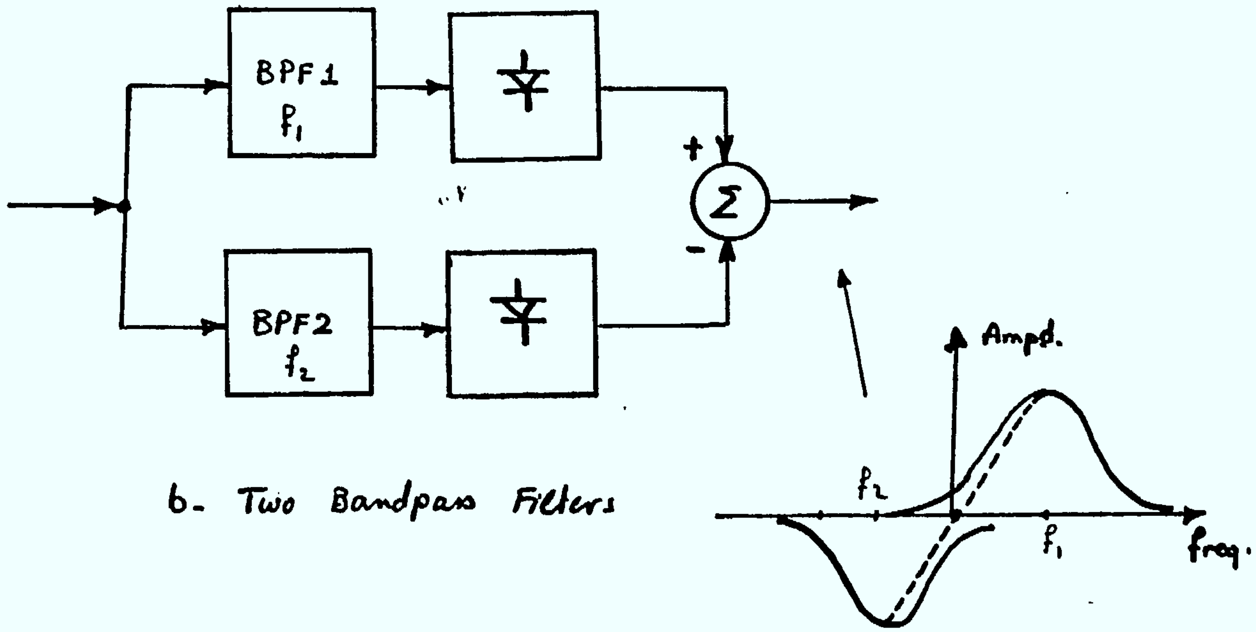
where R(t) and Θ(t) are the amplitude and phase of the fading envelope, and n(t) in the bandpass additive noise. When this signal is band-limited the amplitude fluctuations will be converted into phase noise and gets added to the information phase and the fading phase. The frequency discriminator will simply given an output protportion to the phase derivative as indicated in the following equation:

$$e(t) = 2\pi f_o + 2\pi\Delta f + \dot{\phi}(t) + \dot{\Theta}(t) + d/dt\left[\frac{n_q(t)}{R(t)}\right] \qquad (4.11)$$

$n_q(t)$ is the quadrature component of the noise. If we neglect the noise, the modulation, the fading and the frequency error, we find that the discriminator output is proportional to the carrier frequency. Under no fading and no modulation conditions (i.e., an unmodulated carrier in a Gaussian channel) the discriminator output will be proportional to the sum of the carrier and the frequency offset. Then it would be simple to determine the value and the sign of the frequency error. The noise and the fading add two terms to the discriminator output. Both terms are the



a. Differentiator



b. Two Bandpass Filters

Fig. 4.5. Two Approach for Implementing the Frequency Discriminator

time derivative of phase noise induced by the fading process. The term that includes the Gaussian noise has a significance only under deep fading situation, and since deep fading will make the synchronization virtually impossible, then we should ignore that term in the analysis and assume that synchronization can only be achieved when the signal level is relatively high.

So far we have managed to reduce the produce into studying the effect of two terms; namely, the phase of the fading envelope and the modulation phase. The modulation phase cannot be eliminated by the discriminator, and this is one of the main disadvantages of this method. So, if a discriminator is used to detect the frequency error, either the signal be an unmodulated carrier or a long term average of the discriminator output should be applied. Assuming that an unmodulated carrier is used to achieve initial synchronization, then we are left with the effect of fading noise only.

The derivative of the fading phase will look like a slowly varying frequency in the order of the Doppler frequency shifts. At VHF and lower UHF frequencies where the maximum Doppler frequency is in the order of 10-50 Hz. As it was argued in the previous section such a small frequency drift does not affect the detection of voice signals. At the upper part of the UHF, however, the Doppler frequency will start to affect the speech detection, and these drifts should be checked by an averaging scheme.

Chapter 5

System Aspects and Conclusions

5.1 Introduction

In this chapter we first discuss some of the system aspects that are related to the efficient use of the radio spectrum and the integration of voice and data traffics and the effect of using ACSB techniques on these issues. We then conclude the report by giving the general conclusions of this work, and outlining the necessary follow-up work needed to develop a complete ACSB prototype.

The radio spectrum efficiency is the prime reason for using ACSB techniques. A well designed ACSB system could increase the number of users by a factor of five. Moreover, since a SYNCOMPEX modem includes a built-in digital control channel, then this channel can be used not only to carry the voice compression information, but to also carry signalling control messages in systems that allow for integration of voice and data. This point will be discussed in more detail in section 5.3. Another point is that the availability of a built-in digital modem in the SYNCOMPEX tranceiver which provides an opportunity to design an efficient integrated voice/data modem by having the digital control channel to be switchable between two modes; a low data rate mode for the ACSB voice transmission and a high data rate mode for transmitting data. The availability of the powerful programmable DSP chips makes the implementation of such a modem both feasible and inexpensive.

The two points mentioned in the previous paragraph could lead to a system where the integration of voice and data can be achieved with a high degree of flexibilty and efficiency and would eventually result in a better use of the available radio spectrum.

5.2 Integrated Voice/Data System Based On SYNCOMPEX Tranceivers

The existing mobile systems are mostly FM analog voice where a single FM channel is assigned to a certain user for his exclusive use. The standard configuration of the FM system consists of a base station and a number of mobile units. The FM channel is shared by all units within the system by push-to-talk access mechanism. Every active unit in the system listens to all conversations on the single radio channel. Some systems are designed for data communications, and they also use a configuration similar to that of the analog systems except that the transmissions from the mobile units to the base station are multiplexed using a random access scheme such as the ALOHA protocol.

In systems that require both voice and data communications the voice and data traffic are carried on separate radio channels. It is possible, however, to integrate the two traffics on the same radio channel if the proper channel control is provided. Such integration would lead to a considerable saving of the scarce radio spectrum.

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The current forms of mobile radio systems as described in the previous paragraphs represent a considerable waste of the radio spectrum resources. First, the FM channel bandwidth is at least five times the voice baseband requirements. Second, since the voice and the data are carried on separate radio channels, no advantage is taken of their statistical variations. In order to arrive at a spectrally efficient modem for voice and data the aspects of integration should always be considered. Moreover, it would be more economical to design an integrated modem, on that is easily switchable between the voice and data modes, and one that utilizes minimum bandwidth to transmit a high quality voice and the maximum possible data rate.

The concept of SYNCOMPEX constitutes all the ingredients needed to design an efficient voice/data modem. To start with, the modem utilizes SSB modulation for voice. This is <u>the most</u> spectrally efficient modulation technique, one that not only utilizes small RF bandwidth, but also provides a high quality speech. The second important point is that the modem has a built-in digital modem which facilitates the exchange of control messages between the transmitter and the receiver, and makes the switching between the voice and data modes rather easy. The built-in digital modem can be designed to be switchable between two data rates; a low rate for the control function and a high rate for the data transmission.

5.3 Conclusions

In this work we have addressed the following issues:

(i) Comparison between the ACSB and SYNCOMPEX systems

(ii) Implementation of real time fading simulator to be used for testing the various SYNCOMPEX subsystems

(iii) Analysis and design of AGC subsystem

(iv) Analysis and design of AFC subsystem

(v) The impact of ACSB techniques on the efficient use of the radio spectrum.

In comparing the ACSB to the SYNCOMPEX system it was indicated that the SYNCOMPEX system offers the following advantages:

(1) It allows for the implementation of better compression/expansion algorithms hence providing a better protection for the low volume segments of the speech against the channel noise.

(2) It provides a flexible design of an integrated voice/data modem.

(3) The control channel employs a digitally modulated signal with its energy spread over a certain bandwidth hence minimizing the intermodulation product problem encountered in the tone-based ACSB systems.

The disadvantages of SYNCOMPEX modem in comparison with the tone-based ACSB systems are:

(1) More complicated design

(2) The phase and frequency tracking processes are more difficult to implement.

The main design philosophy adapted in this work calls for the implementation of all control functions (AGC and AFC) at baseband and usable programmable DSP chips (namely; the Texas Instrument TMS320 Chip). It was felt that the first step in studying the feasibility of the SYNCOMPEX concept is to design a flexible real time channel simulator. The advantages of such a simulator would be:

(1) Real time simulation allows for the testing to run over long period of time hence giving statistically valid results. This is in contrast with the computer-based simulation which would limit the simulation time due to the speed of the program's execution and the storage requirements.

(2) The real time simulation also allows for testing channel impairments which are slow by nature such as shadowing and slow frequency drifts.

(3) The implementation of the channel simulator on the TMS320 provides a unique opportunity to study the phase noise of the channel.

The AGC subsystem was designed and tested under fading conditions and gave excellent results. Basically, we have assumed that the AGC function will be carried on two steps:

(1) A slow averaging AGC to be implemented in the analog section of the receiver prior to the entry to the DSP chip. Such a slow acting circuit will take care of the slow variations of the average signal power due to shadowing.

(2) A fast acting AGC to be implemented on the DSP chip and use a feed forward technique for gain correction. Such an AGC

should be able to track the amplitude variations due to fast fading.

The AFC subsystem is being implemented on the TMS320 chip. Several advanced techniques for phase tracking have been studied with the general conclusion that it would be very difficult to maintain phase coherency under fast fading conditions. It was then decided that the best scheme would be to implement a simple frequency discriminator to detect the magnitude of carrier detuning. The locally generated IF carrier would then be adjusted according to the error signal at the discriminator output. The frequency tracking of FD is almost nonexistent when the channel carries information. Yet, for a simple FSK modulation the frequency deviation can be monitored at the discriminator output and can be compared to what it should be. An average process can then be devised to track any consistent frequency drift during the communication period. It should be pointed out that small and rapid frequency fluctuations due to the Doppler effect cannot be tracked by this technique.

Finally, the impact of the ACSB technology on the efficient use of the radio spectrum was briefly examined. It is very obvious that a tremendous saving in the frequency spectrum can be achieved by using the ACSB techniques. Such a saving can be further enhanced by integrating voice and data over the same radio channels. It was pointed out tha the SYNCOMPEX modem is particularly suited for these applications.

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