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entitled

Study on the Simulation and Analysis of an FH/FDMA OBP Satellite Based Mobile Communication System Under Critical Channel Impairment

by

Mike Orra

Submitted to the Graduate Faculty as partial fulfillment of the requirements for the Doctor of Philosophy Degree in Engineering

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The University of Toledo

August 2010
An Abstract of

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Fully regenerative satellite On-Board Processing (OBP) systems are endorsed in literature as being the most effective architecture for maintaining signal quality under jamming circumstances. Dehop-Rehop Transponder (DRT) systems have been proposed as economical alternatives, bridging the gap between passive repeaters and full OBP architectures; however, no openly published literature quantifies the performances of either DRT or OBP payloads through simulation or closed form analysis. The objectives of this study are to provide modeling and simulation of a frequency hopped, frequency division multiple access (FH-FDMA) DRT and OBP satellite based tactical mobile communications system under critical channel impairment. Analyses of the resulting end-to-end BER performances are provided for both architecture types. Two variants of phase shift keying (PSK) modulation are considered for the system waveforms: convolutional coded non-coherent Gaussian Minimum Shift Keying (GMSK) (1-bit differentially detected) and convolutional coded Symmetric Differential PSK (SDPSK). SDPSK and GMSK modulation schemes have been commonly cited in tactical satellite applications wherein bandwidth efficiency and immunity towards adjacent channel interference (ACI)
and inter-symbol interference (ISI) are desirable. While some literature has been published quantifying the performance of SDPSK modems under critical impairment, no such study considering GMSK in this context has been published. Consequently, this study also seeks to determine the feasibility of using 1-bit differentially detected GMSK modems in satellite-based tactical mobile communications systems.

Channel impairment is modeled as partial band noise jamming (PBNJ), and band multi-tone jamming (BMTJ) with AWGN. Degrading factors pertaining to the system hardware including quantization and nonlinear travelling wave tube amplifier (TWTA) are also considered.

Simulations were conducted to illustrate the end-to-end BER for the described system and waveforms of interest. Results show that SFH/SDPSK exhibits excellent immunity towards PBNJ and BMTJ with AWGN channel impairments which can be further enhanced by low rate convolutional codes. OBP processing gains range from 2 - 6.5 dB at a BER of $10^{-3}$ over corresponding DRT systems, depending on jamming intensity and coding rate used. Results further show that using OBP architectures with SFH/GMSK ($BT = 0.5$) waveforms with code rate $1/3$ under uplink PBNJ can realize power efficiency gains between 11.5 dB - 15.5 dB at a BER of $10^{-3}$ when compared to DRT systems. Increasing the BT product to 1 can provide gains of 3.5 dB - 7.5 dB for DRT system architectures (over BT = 0.5). While the increased BT product also results in improved performances for OBP architectures, it is not as pronounced.

SFH/GMSK with convolutional coding cannot realize sufficient performance to be considered for practical application under PBNJ and BMTJ with AWGN impairments, irrespective of the BT product values; in order to use SFH/GMSK modems in tactical
communications systems (both DRT and OBP based architectures), powerful concatenated coding or iterative decoding schemes are required.

Consequently, a theoretical analysis of the performance of 1-bit differential detected GMSK under AWGN is developed herein, so that turbo coding can be applied. Modem level simulations of turbo-coded GMSK under AWGN exhibit an approximate 2 dB gain over convolutional coded GMSK for a BER of $10^{-3}$ with further gains realized for additional decoding iterations. Substantial improvements in power efficiencies were also realized when subjecting turbo coded GMSK to the effects of PBNJ interference, particularly for code rate 2/3. Both empirical investigations into differential GMSK BER performance under AWGN and PBNJ interferences effectively demonstrate that greater bandwidth efficiencies can be realized by using high code rates turbo codes with modest BER performance degradation. These results strongly support use of turbo coding with differential GMSK under AWGN and PBNJ interferences, and in turn application in satellite-based tactical mobile communications systems. The results also warrant further investigation into the feasibility of using differential GMSK under tone jamming conditions.
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Chapter 1

Introduction

The complexity of satellite communications systems is ever increasing as a result of the need for greater data throughput in systems wherein limited bandwidth is available. Consequently, designers must give significant consideration to payload architecture and waveform design to ensure that bandwidth efficiency is maximized while minimizing overall bit error rate (BER). Many currently operating satellites possess the ability to perform transponder level switching, that is, they possess the ability to perform switching at the channel level. While this does represent an improvement in flexibility over traditional passive repeaters, it offers little for many next generation applications, particularly mobile satellite communications. Considering the enormous overhead involved in launching a satellite payload, it is desirable to have commercial systems operating at or near full link capacity to maximize revenues. In order to achieve this, substantial signal processing and networking flexibility is required.

Satellite payload architectures generally can be classified as either passive, partial-processing, or full-processing [1]. A passive repeater does not perform switching in principle, although some variants exist [2]. Regardless of signal format, passive payloads simply receive uplink carrier signals and translate them to another set of carrier
signals for downlink transmission. A partial-processing payload can perform transponder level switching and perhaps demodulation/remodulation of the signal, which can leverage various modulation schemes to accommodate the differences between uplink and downlink channel conditions. While the uplink and downlink are decoupled, no gain can be realized without the inclusion of forward error correction (FEC) decoding and re-encoding on-board the satellite payload. A full-processing payload, or on-board processing (OBP) payload, allows the transmitted uplink signal to be fully recovered on-board the satellite payload (through demodulation and decoding), routed toward the appropriate downlink carrier and formatted for downlink transmission. Recovering the data packets aboard the satellite payload effectively decouples the uplink and downlink channels and the on-board FEC decoding and re-encoding bolsters signal integrity against channel impairments. Furthermore, OBP allows switching to take place at the packet level, maximizing the bandwidth efficiency of the system. Although this architecture is the most complicated, it provides the necessary flexibility needed to serve the data throughput and dynamic requirements of next generation applications.

Previously, the cost of implementing OBP based systems was prohibitive in the commercial sector, and development and implementation was limited to certain Government agencies [3], [4], [5]; however, advances in technologies including digital signal processors (DSPs) and FPGAs are drawing OBP based satellite communication systems closer towards economic viability. Potential applications for OBP based satellite communication include tactical mobile communications, satellite based internet, direct video broadcast and basic telecommunications for rural areas wherein terrestrial networks are non-existent. Many of these services are typically reserved for operation in stationary
contexts; however, their application in mobile environments is also desirable, given current industry emphasis on mobile computing and communications, e.g. 3G/4G networks, netbooks, etc. In order to satisfy the high data throughput requirements of these applications as well as the dynamics of mobile communications, on-board processing (OBP) satellite architectures coupled with bandwidth efficient modulation schemes must be considered.

In addition to application demands, signal impairment represents a significant problem for satellite communications systems. Compromised signal integrity is the result of many factors relating to channel conditions and system hardware limitations. In this research, channel impairment factors including additive white Gaussian noise (AWGN), partial-band noise jamming (PBNJ), and band-multitone jamming (BMTJ) must be considered. Furthermore, the effects of system inherent degradation factors such as quantization noise, nonlinear amplification, and adjacent channel interference (ACI) must also be taken into account.

The primary objective of this study is to provide simulation and analysis to determine the performance of a frequency hopped, frequency division multiple access (FH-FDMA) OBP satellite based mobile communication system under critical channel impairment. It is widely accepted that OBP based systems will outperform partial-processing payloads and transparent repeaters; however, no openly published literature provides a quantitative measure of the difference in architectural performances with respect to overall bit error rate (BER). Thus, for comparative purposes, simulation and performance analysis will also be conducted for Dehop Rehop Transponder (DRT), a variant of a partial-processing payload. Based on frequency hopping scenarios, multiple
modulation schemes, forward error correcting (FEC) coding schemes, channel models and the source-to-sink system hardware model (comprised of transmitting earth-terminals, DRT and OBP satellite payloads and receiving earth-terminals) are considered. Their respective performances under various impairments will be assessed, primarily through simulation, with some consideration for mathematical modeling.

A secondary objective of this study is to assess the feasibility of using two variants of PSK modulation for the system waveforms: convolutional coded non-coherent GMSK (1-bit differentially detected) and convolutional coded SDPSK. SDPSK and GMSK modulation schemes have been commonly cited in tactical satellite applications, wherein bandwidth efficiency and immunity towards adjacent channel interference (ACI) and inter-symbol interference (ISI) are desirable [3], [6], [7], [8], [9], [10]. Some literature has been published quantifying the performance of SDPSK modems under critical impairment (i.e. naturally occurring or targeted jamming interferences) [11], [12], [13], [14]; no such study considering GMSK in this context has been published. Consequently, this study also seeks to determine the feasibility of using 1-bit differentially detected GMSK modems in tactical mobile communications systems as compared to previously used SDPSK modems.

The remainder of this chapter will present a detailed overview of existing satellite payloads incorporating on-board processing. This survey will serve as the primary basis from which an architectural topology will be drawn for this research.
1.1 Tactical Communication Satellite Payloads

1.1.1 Milstar I Low Data Rate (LDR) Payload

The Milstar I payload was designed to carry tactical low data rate (LDR) communications ranging from 75 bps to 2.4 kbps via a four satellite constellation with intersatellite crosslinks in the 60 GHz frequency range \[3\], \[15\]. The LDR waveform itself is defined by the military’s satellite data link standards for uplinks and downlinks in document MIL-STD-1582C \[16\]. The emphasis in the design of the system was placed upon implementing worldwide, anti-jam, low-probability of detection (LPD) and low probability of intercept (LPI), communications services. The system was designed to operate in the extremely high frequency (EHF) band (uplink) and super high frequency (SHF) band because of the improved resistance to jamming and interference \[3\]. The payload uplink signal employs FDMA and non-coherent M-ary frequency shift keying (NC-MFSK) modulation with fast frequency hopping (FFH). The maximum symbol transmission rate is equal to that of the hopping rate \[5\]. The user signals are received on the nine extremely high frequency (EHF) receive antenna beams and satellite crosslinks. The EHF signals are frequency hopped across the 43.5 – 45.5 GHz frequency band \[3\]. The LDR payload supports 192 MFSK uplink signals using frequency division multiple access (FDMA). Extensive on-board processing is used to dehop, demodulate, and decode signals to recover the transmitted data sequences. Signals are then reformatted and rehopped accordingly for downlink transmission. It should be noted that use of multiple demodulators was impractical due to the large number of user channels. Therefore, a multicarrier demodulation (MCD) approach was adopted \[5\].
The downlink data stream is TDM formatted with each user occupying a certain number of time slots. Milstar I has a single downlink mechanism which provides service to all users on a time-share basis. The downlink data stream is either MFSK or differential phase-shift keying (DPSK) modulated. It is possible for the system to dynamically alter modulation format from one time slot to another, as needed. The downlink signal is then frequency hopped in the 20.2 to 21.2 GHz SHF band and amplified by a single 25 Watt traveling wave tube amplifier (TWTA). No explicit specifications were given for supported channel coding schemes. Table 1.1 summarizes the Milstar I LDR payload operating parameters.

Table 1.1: Summary of Milstar I operating parameters.

<table>
<thead>
<tr>
<th>Satellite Payload</th>
<th>Milstar I</th>
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<tbody>
<tr>
<td>Data Rates</td>
<td>75 – 2400 bps (LDR)</td>
</tr>
<tr>
<td>High Power Amplifier</td>
<td>Single 25 W TWTA</td>
</tr>
<tr>
<td>Number of Channels</td>
<td>192 LDR</td>
</tr>
<tr>
<td>Multiple Access Scheme (up/down link)</td>
<td>FDMA/TDM</td>
</tr>
<tr>
<td>Modulation (up/down link)</td>
<td>MFSK / MFSK, DPSK</td>
</tr>
<tr>
<td>Channel Coding</td>
<td>Unknown</td>
</tr>
<tr>
<td>Frequency Band (up/down link)</td>
<td>EHF, UHF / SHF, UHF</td>
</tr>
<tr>
<td>On Board Channelization Scheme</td>
<td>Multicarrier Demultiplexer</td>
</tr>
<tr>
<td>End-to-End BER</td>
<td>Unknown</td>
</tr>
<tr>
<td>OBP Capabilities</td>
<td>Full packet and beam level switching</td>
</tr>
</tbody>
</table>

### 1.1.2 Milstar II Medium Data Rate (MDR) Payload

The Milstar II payload applies much of the Milstar I (LDR) payload to a medium data rate context, namely 4.8 kbps to 1.544 Mbps [3], [15]. Support is provided for 32 uplink channels, where each channel supports a single user at 1.544 Mbps or a combination of several users at lower data rates whose total exceeds 2 Mbps. The MDR waveform is governed by the US military’s satellite data link standards, as defined by MIL-STD-188-
The EHF uplink waveforms are modulated using a symmetric differential phase-shift keying (SDPSK) scheme and are frequency hopped. Support for the SDPSK modulation scheme was implemented because of its immunity towards adjacent channel interference (ACI). MFSK modulation (used for the LDR payload) was not possible here because it could not support MDR throughputs [3]. The multiple access schemes for the MDR up and down links are FDMA and TDM, respectively. Each of the 32 uplink channels time division multiplexed and formatted for FDMA [15].

The payload demodulators are partitioned into four sets of eight SDPSK demodulators each [3]. Each set of demodulators frequency division demultiplexes eight channels and then performs demodulation. A per-channel demultiplexing scheme based on SAW filters was employed [5] wherein one demodulator unit was assigned for each uplink channel. The MDR transmit sub-system DPSK modulates downlink signals and frequency hops them across a 1 GHz span of the SHF band. The datastream itself is TDM formatted, with each user being assigned a certain number of time slots. A single 60 Watt TWTA provides signal amplification. The higher power output amplification (relative to LDR) is necessary to achieve the downlink with smaller user terminals pushing higher data rates. Table 1.2 summarizes the Milstar II operating parameters.

Table 1.2: Summary of Milstar II operating parameters.

<table>
<thead>
<tr>
<th>Satellite Payload</th>
<th>Milstar II</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Data Rates</strong></td>
<td>75 bps – 1.544 Mbps (LDR/MDR)</td>
</tr>
<tr>
<td><strong>Max Data Throughput</strong></td>
<td>40 Mbps</td>
</tr>
<tr>
<td><strong>High Power Amplifier</strong></td>
<td>Single 60 W TWTA</td>
</tr>
<tr>
<td><strong>Number of Channels</strong></td>
<td>192 LDR, 32 MDR</td>
</tr>
<tr>
<td><strong>Multiple Access Scheme (up/down link)</strong></td>
<td>FDMA/TDM</td>
</tr>
<tr>
<td><strong>Modulation (up/down link)</strong></td>
<td>MFSK, SDPSK / MFSK, DPSK</td>
</tr>
<tr>
<td><strong>Channel Coding</strong></td>
<td>Unknown</td>
</tr>
<tr>
<td><strong>Frequency Band (up/down link)</strong></td>
<td>EHF / SHF</td>
</tr>
<tr>
<td><strong>On Board Channelization Scheme</strong></td>
<td>Unknown</td>
</tr>
</tbody>
</table>
1.1.3 Advanced Extremely High Frequency (AEHF), Milstar III Payload

The AEHF payload incorporates both LDR and MDR payloads and extends their functionalities to offer higher combined data rate throughputs ranging from 75 bps to 8.192 Mbps. The payload is completely backward compatible with both LDR and MDR payloads. Aside from the increase in data throughput capacity, the AEHF payload design considers the use of an on board packet routing processor that can process ATM or TCP/IP data packets. Advances in technology since implementation of previous payloads naturally translates into a smaller overall integrated package for the system as a whole.

As with the LDR and MDR payloads, the use of the EHF/SHF frequency bands provide excellent terminal mobility, high anti-jam (AJ) performance and low probabilities of both detection and interception [4], [5]. Much of the improved resistance to jamming claimed by the AEHF payload results from an uplink hopping bandwidth that is over 30 times as wide as SHF frequency hopping systems [4]. Table 1.3 summarizes the key operating parameters of AEHF.

Table 1.3: Summary of AEHF operating parameters.

<table>
<thead>
<tr>
<th>Satellite Payload</th>
<th>AEHF</th>
</tr>
</thead>
<tbody>
<tr>
<td>Data Rates</td>
<td>75 bps – 8.182 Mbps (LDR/MDR)</td>
</tr>
<tr>
<td>Max Data Throughput</td>
<td>≈200 Mbps</td>
</tr>
<tr>
<td>High Power Amplifier</td>
<td>Unknown</td>
</tr>
<tr>
<td>Number of Channels</td>
<td>≈256 LDR, 56 MDR</td>
</tr>
<tr>
<td>Multiple Access Scheme (up/down link)</td>
<td>FDMA/TDM</td>
</tr>
<tr>
<td>Modulation (up/down link)</td>
<td>MFSK, SDPSK, SDQPSK, GMSK / MFSK, DPSK, GMSK</td>
</tr>
<tr>
<td>Channel Coding</td>
<td>Unknown</td>
</tr>
<tr>
<td>Frequency Band (up/down link)</td>
<td>EHF / SHF</td>
</tr>
</tbody>
</table>
1.1.4 The Mobile User Objective System (MUOS)

The Navy’s Ultra High Frequency (UHF) Follow-On (UFO) satellite constellation (launched in 1993) provides narrowband tactical communications to Department of Defense (DOD). As the system is nearing the end of its life, the Navy has implemented a strategy outlining its replacement, which consists of a UHF gapfiller satellite (launched in 2003), the use of commercial satellite systems (to the extent practical), and development of the Mobile User Objective System (MUOS) [17], [18]. MUOS will provide unprotected narrowband tactical satellite communications (in the UHF band) to the Navy and Department of Defense but with expanded capabilities for future services and needs. The need for communications in the UHF band is of particular importance to the department of defense as signals in this band are able to pass through inclement weather, foliage, and concrete structures [18]. The most significant enhancements MUOS will provide are increased data throughput (up to 39.2 Mbps) and smaller, light-weight handheld user terminals (perhaps weighing as little as one pound).

On-board processing as well as bandwidth and power efficient modulation schemes have been proposed for MUOS to support the flexibility and data throughput requirements [19]. Traditional modulation schemes such as BPSK and DPSK are spectrally inefficient and produce significant sidelobes which can result in performance degradation due to adjacent channel interference (ACI). As a result, MacMillan et. al. [19] proposed a continuous phase modulation (CPM) approach for MUOS handheld
terminals, such as shaped offset quadrature phase shift keying (SOQPSK) or Gaussian minimum shift keying (GMSK), and MPSK or MQAM for larger terminals. Also recommended was the use of turbo coding for improved power efficiency [19], although it is possible that the system will employ convolutional and RS codes. The summarized MUOS operating parameters (as suggested by the available literature) are presented in Table 1.4.

Table 1.4: Summary of MUOS operating parameters.

<table>
<thead>
<tr>
<th>Satellite Payload</th>
<th>MUOS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Data Rates</td>
<td>Up to 64 kbps</td>
</tr>
<tr>
<td>Max Data Throughput</td>
<td>~40 Mbps</td>
</tr>
<tr>
<td>High Power Amplifier</td>
<td>Unknown</td>
</tr>
<tr>
<td>Number of Channels</td>
<td>~2000</td>
</tr>
<tr>
<td>Multiple Access Scheme</td>
<td>WCDMA</td>
</tr>
<tr>
<td>Modulation</td>
<td>MFSK, SOQPSK, GMSK, MPSK, MQAM</td>
</tr>
<tr>
<td>Channel Coding</td>
<td>Turbo, Convolutional, RS</td>
</tr>
<tr>
<td>Frequency Band</td>
<td>UHF (290 – 320 MHz / 240 – 270 MHz)</td>
</tr>
<tr>
<td>On Board Channelization</td>
<td>Unknown</td>
</tr>
<tr>
<td>End-to-End BER</td>
<td>$10^{-5}$</td>
</tr>
<tr>
<td>OBP Capabilities</td>
<td>Full</td>
</tr>
</tbody>
</table>

1.2 Commercial and Research Satellite Payloads

1.2.1 The Multimedia Demonstration Experiment with Datalink to the ISS (MEDIS) Payload

The Multimedia Demonstration Experiment with Datalink to the ISS (MEDIS) system is a German project consisting of two Medium Earth Orbit (MEO) satellites whose primary objectives are to provide an in-orbit test (IOT) bed for optical terminals, OBP, and Ka-band equipment. The satellite payload is composed of three main components: the Ka-
band subsystem, OBP subsystem, and laser communication terminal (LCT) subsystem [20].

The antenna system design provides independent paths for receiving and transmitting the five FDMA uplink and five TDM downlink signals. The signals are QPSK modulated. Received signals have a bandwidth of 160 MHz and are received on a 28 GHz carrier uplink. They are low-noise amplified and downconverted (in a single step process) to IF at 475 MHz. Typically the anti-aliasing filter (AAF) is used just prior to performing A/D conversion, however, due to implementation constraints, it is preferable to have the AAF inserted just after downconversion. The total power of the incoming signal to be digitized should be fall within the operating range of the ADC; therefore, an automatic gain control (AGC) module is used to ensure operation is compliant with the allowable ADC limits [20].

The OBP subsystem is comprised of a multicarrier demultiplexer demodulator (MCDD), modulator assembly, LCT interfaces and OBP switch. These components were implemented using high-speed digital signal processors. Digitized signals from the Ka-band subsystem are demultiplexed, demodulated and decoded (by five individual demodulators and decoders – one for each carrier). The data and coding rates for each demodulator are established independently from the telecommand center. The OBP switch performs ATM data routing from the appropriate input and output ports with a throughput of 300 Mbits/s. The switch then routes signals to one of five independent modulators which consist of a baseband processor and analog direct Ka-modulator. The baseband processors are responsible for framing, interleaving coding and pulse shaping. The signals are then passed to either the downlink transmit antenna or the optical
terminal. The LCT differs from the earth terminal uplink and downlink carriers (each comprised of up to five 30 Mbps carriers) in that its interface is a single high-speed (150 Mbps) ATM cell stream connected to the OBP switch [20].

The transmit assembly processes up to five QPSK modulated carriers occupying a 235 MHz band for downlink transmission at 18 GHz. The payload’s transmit power module encompasses a TWTA, high-voltage supply, electronic power conditioner (EPC) and channel amplifier and linearizer [20]. Table 1.5 summarizes the operating parameters of the MEDIS payload.

Table 1.5: Summary of MEDIS operating parameters.

<table>
<thead>
<tr>
<th>Satellite Payload</th>
<th>MEDIS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Data Rates</td>
<td>30 Mbps per carrier (up to five)</td>
</tr>
<tr>
<td>Max Data Throughput</td>
<td>150 Mbps</td>
</tr>
<tr>
<td>High Power Amplifier</td>
<td>TWTA (27 - 55 W)</td>
</tr>
<tr>
<td>Number of Channels</td>
<td>5</td>
</tr>
<tr>
<td>Multiple Access Scheme (up/down link)</td>
<td>FDMA/TDM</td>
</tr>
<tr>
<td>Modulation</td>
<td>QPSK</td>
</tr>
<tr>
<td>Channel Coding (up/down link)</td>
<td>Convolutional, RS</td>
</tr>
<tr>
<td>Frequency Band (up/down link)</td>
<td>28 GHz / 18 Ghz</td>
</tr>
<tr>
<td>On Board Channelization Scheme</td>
<td>MCDD</td>
</tr>
<tr>
<td>End-to-End BER</td>
<td>$10^{-9}$</td>
</tr>
<tr>
<td>OBP Capabilities</td>
<td>Full</td>
</tr>
</tbody>
</table>

1.2.2 Hotbird 4

Among the very first satellites to use OBP in a commercial context was Eutelsat’s Hotbird 4. The OBP features were developed at Eutelsat in order to provide easy, real-time access to satellite broadcast resources (i.e. DVB and derivative standards) from any location (within the satellite coverage area) [21]. This system incorporates the Skyplex transponder, which processes a 33 MHz bandwidth comprised of up to 8 carriers in 6 channels, each occupying a 5.5 MHz bandwidth. The functions performed at the ground
terminals include transport packet generation and multiplexing, RS outer encoding, scrambling and shaped (root raised cosine mask with rolloff = 0.35) QPSK modulation.

The Skyplex transponder is responsible for performing carrier demultiplexing and demodulation, SED descrambling, remultiplexing, scrambling, convolutional channel coding and remodulation (shaped QPSK). Channelization and recovery of the baseband data is achieved by using an ASIC implemented polyphase filter (PPF) and 8 point complex FFT operation. The MCD approach was selected, as opposed to a per-channel processing scheme, to minimize the power and mass of Skyplex. Table 1.6 summarizes the operating parameters of Hotbird 4 [22].

Table 1.6: Summary of Hotbird 4 operating parameters.

<table>
<thead>
<tr>
<th>Satellite Payload</th>
<th>Hotbird 4</th>
</tr>
</thead>
<tbody>
<tr>
<td>Data Rates</td>
<td>2 - 7 Mbps per channel</td>
</tr>
<tr>
<td>Max Data Throughput</td>
<td>36 Mbps</td>
</tr>
<tr>
<td>High Power Amplifier</td>
<td>Unknown</td>
</tr>
<tr>
<td>Number of Channels</td>
<td>6</td>
</tr>
<tr>
<td>Multiple Access Scheme (up/down link)</td>
<td>SCPC &amp; TDMA / TDM</td>
</tr>
<tr>
<td>Modulation</td>
<td>QPSK</td>
</tr>
<tr>
<td>Channel Coding</td>
<td>Convolutional, RS - rates 1/2, 2/3, 3/4</td>
</tr>
<tr>
<td>Frequency Band (up/down link)</td>
<td>13.75 – 18.4 GHz / 10.7 – 12.75 GHz</td>
</tr>
<tr>
<td>On Board Channelization Scheme</td>
<td>Polyphase Filter and 8-pt Complex FFT.</td>
</tr>
<tr>
<td>End-to-End BER</td>
<td>$10^{-7}$</td>
</tr>
<tr>
<td>OBP Capabilities</td>
<td>Full</td>
</tr>
</tbody>
</table>

1.2.3 The Advanced Communications Technology Satellite (ACTS)

The ACTS system was developed by the National Aeronautical and Space Administration (NASA) to advance satellite communications technology, particularly dynamic beam hopping, and advanced on board traffic switching and processing. It was
the first all digital, gigabit capacity GEO satellite and had the ability to carry digital communications at standard fiber-optic rates with the same quality of service (QoS). Perhaps more importantly, ACTS showed that satellite systems could provide communications at fiber optic rates, integrate with existing terrestrial fiber networks and still have cost and performance advantages. The data rates supported in this platform range from kbps to hundreds of Mbps. The raw BER achieved by the system at 622 Mbps is $10^{-6}$; however, RS coding can lower this error floor to $10^{-11}$. The ACTS payload operates in the Ka-band with a 2.5 GHz bandwidth with full on-board processing and switching [23]. It also incorporates adaptive rain fade compensation, which allows for increased link margins for uplink and downlinks experiencing rain fade effects [23]. The types of services carried by the ACTS system include on-demand integrated voice, video and data, high-speed data networks, broadband T1 video for aircraft and ships, aeronautical voice and low data rate mobile voice, video and data, and interactive multimedia [24]. Table 1.7 summarizes the operating parameters of the ACTS payload [24], [25].

Table 1.7: Summary of ACTS operating parameters.

<table>
<thead>
<tr>
<th>Satellite Payload</th>
<th>ACTS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Data Rates</td>
<td>kbps – Hundreds of Mbps</td>
</tr>
<tr>
<td>Max Data Throughput</td>
<td>Hundreds of Mbps</td>
</tr>
<tr>
<td>High Power Amplifier</td>
<td>46 W per channel</td>
</tr>
<tr>
<td>Number of Channels</td>
<td>3 (900 MHz BW per channel)</td>
</tr>
<tr>
<td>Multiple Access Scheme</td>
<td>DAMA &amp; TDMA</td>
</tr>
<tr>
<td>(up/down link)</td>
<td>BPSK, QPSK, SMSK</td>
</tr>
<tr>
<td>Modulation</td>
<td>RS</td>
</tr>
<tr>
<td>Frequency Band</td>
<td>30 GHz / 20 GHz</td>
</tr>
<tr>
<td>(up/down link)</td>
<td></td>
</tr>
<tr>
<td>On Board Channelization</td>
<td>Unknown</td>
</tr>
<tr>
<td>Scheme</td>
<td></td>
</tr>
<tr>
<td>End-to-End BER</td>
<td>$10^{-11}$ (with RS coding)</td>
</tr>
<tr>
<td>OBP Capabilities</td>
<td>Full</td>
</tr>
</tbody>
</table>
1.2.4 The Wideband InterNetworking Engineering Test and Demonstration Satellite (WINDS)

The Wideband InterNetworking engineering test and Demonstration Satellite (WINDS) is an experimental satellite developed by the Japan Aerospace and Exploration Agency (JAXA). Its primary objective is to establish a domestic and international ultra-high speed internet network using experimental (non-field tested) technologies. In addition to internet services, the platform will be used for information management as well as disaster relief applications.

WINDS operates in the Ka frequency band, which favors high data throughputs as well as concentrated multi-beam topologies. The system provides bent-pipe (transponder level) as well as baseband ATM switching. Access to the satellite is achieved through a satellite switched- TDMA (SS-TDMA) scheme. Areas of service coverage include Japan, southeast Asia and Pacific region. Downlink signal transmission is facilitated through use of a multiport amplifier which provides linear power output ranging from 0 – 280 Watts per port. This allows for dynamic uplink and downlink power control which serves to efficiently compensate for signal fading as a result of foliage, rain, etc. Overall, the system is expected to achieve maximum throughputs of 155 Mbps downstream and 6 Mbps upstream for users of 45 cm apertures (same as currently used for satellite communications). Throughput for users employing 5 meter terminal antennas will be 1.2 Gbps duplex [26]. Table 1.8 summarizes the operating parameters of the WINDS payload.
Table 1.8: Summary of WINDS operating parameters.

<table>
<thead>
<tr>
<th>Satellite Payload</th>
<th>WINDS</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Data Rates</strong></td>
<td>6 Mbps / 155 Mbps (45 cm aperture)</td>
</tr>
<tr>
<td></td>
<td>1.2 Gbps (5 m aperture)</td>
</tr>
<tr>
<td><strong>Max Data Throughput</strong></td>
<td>Gbps</td>
</tr>
<tr>
<td><strong>High Power Amplifier</strong></td>
<td>0 - 280 W per port</td>
</tr>
<tr>
<td><strong>Number of Channels</strong></td>
<td>Unknown</td>
</tr>
<tr>
<td><strong>Multiple Access Scheme (up/down link)</strong></td>
<td>SS-TDMA</td>
</tr>
<tr>
<td><strong>Modulation</strong></td>
<td>QPSK</td>
</tr>
<tr>
<td><strong>Channel Coding</strong></td>
<td>RS (255,233)</td>
</tr>
<tr>
<td><strong>Frequency Band (up/down link)</strong></td>
<td>30 GHz / 20 GHz</td>
</tr>
<tr>
<td><strong>On Board Channelization Scheme</strong></td>
<td>Unknown</td>
</tr>
<tr>
<td><strong>End-to-End BER</strong></td>
<td>$10^{-11}$ (with RS coding)</td>
</tr>
<tr>
<td><strong>OBP Capabilities</strong></td>
<td>Full</td>
</tr>
</tbody>
</table>
Chapter 2

Waveform Design

The term waveform refers to the specific manner in which communication through a given medium is achieved. With respect to this study, it defines the mechanisms through which the earth terminal transmitters and receivers communicate with the satellite payload, including (but not limited to) modulation/demodulation schemes, FEC encoding/decoding schemes, multiple-access schemes, interference mitigation and frequency plan.

This chapter focuses on some of the aforementioned aspects of waveform design considered in this study. Specifically section 2.1 discusses multiple access schemes, while sections 2.2 and 2.3 are concerned with modulation/demodulation schemes of particular interest and FEC coding/decoding schemes, respectively.

2.1 Multiple Access Schemes

One of the fundamental considerations in the design of a satellite communication system is the means through which the payload will “simultaneously” link to multiple earth stations. The most commonly used techniques to implement this task are time division
multiple access (TDMA), code division multiple access (CDMA), and frequency division multiple access (FDMA).

A TDMA based scheme allocates “time slots” to terminal users during which they have exclusive access to a given transponder. Due to this exclusivity of use, intermodulation products (IMPs), which result from passing sinusoids through highly nonlinear devices such as amplifiers, are not a primary concern, contrary to FDMA systems; however, implementing an efficient and effective sharing technique requires substantial design effort in network timing and synchronization [27].

The CDMA technique assigns each earth terminal an orthogonal pseudo-random code which is used to spread signal power over a large bandwidth. This scheme allows multiple signals to overlap in time and frequency without interfering with each other. Thus, users effectively have on-demand access to the entire shared system bandwidth; however, CDMA suffers from limitations in the ability to generate large numbers of orthogonal codes.

An FDMA scheme partitions the allocated bandwidth into a number of non-overlapping frequency bands, each of which constitutes a channel. Uniformity of the channels is dependent on the particular application at hand. Each earth terminal is assigned an access channel. In the context of multiple access schemes, the terms FDMA and single channel per carrier (SCPC) can be used interchangeably; both reflect the fact that a single carrier frequency is allocated to a single channel – i.e. an non-time multiplexed input signal. Figure 2.1 illustrates the spectral topology of a two-sided two-channel FDM formatted waveform. It is important to note that in this scheme multiple carrier signals are simultaneously received and processed by the payload. As a result of
imperfections in technology, the signals are subject to non-linearities introduced by filters, amplifiers, hard-limiters, etc. Despite the degradations introduced through IMP, FDMA remains a commonly used multiple access scheme for specific applications, as indicated in the literature survey in Chapter 1. Therefore, this study will be concerned only with the FDMA based multiple access scheme.

![Spectral topology of 2-channel FDM formatted signal.](image)

**Figure 2.1:** Spectral topology of 2-channel FDM formatted signal.

### 2.2 Modulation Schemes

The process of selecting appropriate modulation schemes is dependent upon the specific needs of the application, including power and bandwidth efficiencies, immunity towards interference and jamming, and implementation complexity. M-ary Phase Shift Keying (MPSK), M-ary Quadrature Amplitude Modulation (MQAM), and M-ary Frequency Shift Keying (MFSK) represent the three generic classes of modulation schemes from which a selection can be made. Herein, $M$ refers to the size of the alphabet (typically a power of 2) and each symbol represents $\log_2 M$ bits. The bandwidth efficiency of these schemes is improved by increasing the value of $M$; in the case of MFSK this translates into increased dimensionality and hence, decreased power efficiency. For mobile tactical communications, substantial consideration must be given to a waveform’s immunity towards jamming and interference. The modulation scheme should be chosen for spectral
compactness, which can mitigate the effects of ACI and inter-symbol interference (ISI). The spectral compactness is typically characterized by measuring main-lobe width, suppression ratio of side-lobes, and decaying speed of side-lobes’ power spectral density (PSD) as a function of frequency. Aside from spectral compactness, a modulation scheme alone has little or no interference-resistant properties, and must be coupled with coding and spectral shaping elements. Given the implications of mobile communications, the implementation complexity of a waveform is just as important as interference rejection and bandwidth efficiency. The design objective here is to implement the signal recovery (synchronization, estimation, decoding, demodulation) using as simple a scheme as possible while guaranteeing the minimum level of desired performance.

In light of these design considerations and the literature survey of Chapter 1, symmetric differential PSK (SDPSK) and differentially detected Gaussian minimum shift keying (GMSK) modulation schemes will be considered for application in the mobile communication system under study. In addition, non-coherent binary FSK (BFSK) modulation will be considered for comparative purposes, as MFSK based schemes have historically been used in tactical communications systems. Note that for this study all of the modems under consideration are non-coherently detected. Non-coherent demodulation techniques do not rely on phase information (which is easily corrupted) to demodulate signals; rather, non-coherent demodulation can be established through the use of correlators and square-law detectors that determine which symbols were transmitted based on corresponding signal energies. Sections 2.2.1 - 2.2.3 provide a detailed discussion of SDPSK, GMSK and BFSK, respectively.
2.2.1 Symmetric Differential Phase Shift Keying

Symmetric differential phase shift keying is a non-coherent PSK technique which is a variation on differential phase shift keying (DPSK). Differential phase shift keying (DPSK) or differential encoded PSK (DEPSK) is a non-coherent PSK technique often used when carrier phase synchronization is difficult to realize or impossible. In DPSK modulation, symbols are encoded by rotating the carrier phase by \( \pi \), if \( I_k = '0' \) is transmitted, and 0 (unchanged) if \( I_k = '1' \) is transmitted, where \( I_k \) is the \( k^{th} \) information bit in the transmission sequence [9]. Hence, the phase difference values of consideration in DPSK are in the set \([0, \pi]\). The DPSK signal can then be defined as [28],

\[
S(t) = A \cos(2\pi f_c t + \Theta_k) \quad \text{for} \quad kT_b \leq t \leq (k + 1)T_b \quad \text{and} \quad k = 0,1,2,...,n
\]  

(1)

where \( A \) is the carrier amplitude, \( T_b \) is the bit duration, and \( f_c \) is the carrier frequency, and \( \Theta_k \) is the differential encoded phase value. Differential encoding of the data sequence can be achieved through use of a simple logical XOR function [29],

\[
d_k = I_k \oplus d_{k-1}
\]

(2)

where \( d_k \) represents the encoded data sequence and \( I_k \) represents the information sequence. The encoded information sequence is then simply mapped to the corresponding phase values as shown in Table 2.1. Note that the modulated DPSK signal is antipodal and can only assume the values shown in equation (3).

Table 2.1: DPSK Phase Assignment.

<table>
<thead>
<tr>
<th>( d_k )</th>
<th>( \Theta_k )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>1</td>
<td>( \pi )</td>
</tr>
</tbody>
</table>

21
\[ S_1(t) = A\cos(2\pi f_c t) \]
\[ S_2(t) = A\cos(2\pi f_c t + \pi) = -A\cos(2\pi f_c t) \quad (3) \]

SDPSK is a variation of DPSK which uses phase shifts of ±π/2 to encode the transmitted bit information prior to performing PSK modulation. Hence, if \( I_k = '0' \) the phase is rotated by -π/2, and if \( I_k = '1' \) the phase is rotated by + π/2, where \( I_k \) is the \( k^{th} \) information bit in the transmission sequence. The SDPSK signal can also be defined using equation (1); however, the encoding of the phase is altered as follows [6]:

\[ d_k = d_{k-1} - 1, \text{ for } I_k = '0' \quad (4) \]
\[ d_k = d_{k-1} + 1, \text{ for } I_k = '1' \quad (5) \]

After generating the encoded data sequence \( d_k \in [-3,3] \), it can be used to generate SDPSK signal with phase,

\[ \theta_k = d_k \frac{\pi}{2} \quad (6) \]

Thereafter, the encoded data sequence can be mapped to the phase values shown in Table 2.2. Note that the SDPSK modulated signal can assume the signal values as defined in equation (7).

**Table 2.2: SDPSK Phase Assignment.**

<table>
<thead>
<tr>
<th>( d_k )</th>
<th>( \Theta_k )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>1, -3</td>
<td>( \pi/2 )</td>
</tr>
<tr>
<td>2, -2</td>
<td>( \pi )</td>
</tr>
<tr>
<td>-1, 3</td>
<td>-( \pi/2 )</td>
</tr>
</tbody>
</table>
Notice here that SDPSK holds an advantage over DPSK when considering long runs of the same bit. Because SDPSK continuously rotates the phase of the symbol either clockwise or counterclockwise, the symbol beginning and ending and is relatively easy to resolve; however, with DPSK transmission of the same bit does not yield any change in phase and therefore, synchronization in the receiver may be compromised. It should be noted that the bit error rate for both DPSK and SDPSK is identical under ideal conditions [9]; however, under non-negligible impairments (such as jamming, fading, nonlinear TWTA, etc.), the two modulation techniques will perform differently. For conventional DPSK with intersymbol interference (ISI) (caused by the repeater’s filters or timing error at the receiver), the bit error probability differs depending on the value of the data bit because the phase shift lacks symmetry for these bits. For a zero phase shift (corresponding to an information bit ’0’), the ISI from adjacent bits will have no effect, whereas a π radian phase shift (corresponding to ‘1’) will produce the maximum effect; however, under SDPSK, adjacent bits are always in quadrature with the preceding and succeeding bits [6]. In light of its superior immunity, the remainder of this section will present further details regarding the SDPSK modulation and demodulation schemes.

The complex envelope representation of an SDPSK modulated signal is given by:

\[
\tilde{S}(t) = \sum_{k=-\infty}^{\infty} \sqrt{\frac{2E_b}{T_b}} \exp(j\theta_k)P(t-kT_b)
\]  

(8)
\[
= \sum_{k=-\infty}^{\infty} \sqrt{\frac{2E_b}{T_b}} \cos(\theta_k) P(t - kT_b) + j \sum_{k=-\infty}^{\infty} \sqrt{\frac{2E_b}{T_b}} \sin(\theta_k) P(t - kT_b)
\]
where
\[
\theta_k = \theta_{k-1} - \frac{\pi}{2}, \text{ for } I_k = '0' \quad (9)
\]
\[
\theta_k = \theta_{k-1} + \frac{\pi}{2}, \text{ for } I_k = '1' \quad (10)
\]
and \( P(t) \) is the pulse train function. If we are to consider an SDPSK signal as a real bandpass signal, its expression can be given as:
\[
S(t) = \sqrt{\frac{2E_b}{T_b}} \cos(2\pi f_c t + \theta_k(t)) \quad (11)
\]

The SDPSK receiver used to recover data from the modulated signal can be categorized as a suboptimum receiver (in which the reference signal is the previous symbol), or an optimum noncoherent (i.e. differentially coherent) receiver. The reference signal for suboptimum receiver is the previous symbol, which is subject to impairment from noise and other interference. Consequently, the performance of the suboptimum receiver is inferior to that of optimum receiver and hence, its name. The optimum noncoherent receiver does not have any knowledge of the transmitted signal’s phase nor does it estimate its value to perform demodulation. Rather, it leverages correlators to perform demodulation as shown in Figure 2.2.

In order to illustrate operation of the optimum non-coherent SDPSK receiver, let us assume transmission of two symbols in the data sequence (\( k \) and \( k+1 \)) and that we are interested in resolving the phase value of the \((k+1)^{th}\) symbol. The received signal \( r(t) \) is mixed with carrier frequencies (which are in quadrature) generated by the local oscillator,
and then integrated over one symbol duration, $T_b$. The instantaneous signals $\varepsilon_i$ and $\varepsilon_q$ are expressed in equations (12) and (13).

\[
\varepsilon_i(t) = \cos(2\pi f_c t + \phi(t)) \cos(2\pi f_c t) = \frac{1}{2} \cos(4\pi f_c t + \phi(t)) + \frac{1}{2} \cos(\phi(t))
\]  \hspace{1cm} (12)

\[
\varepsilon_q(t) = \cos(2\pi f_c t + \phi(t)) \sin(2\pi f_c t) = \frac{1}{2} \sin(4\pi f_c t + \phi(t)) - \frac{1}{2} \sin(\phi(t))
\]  \hspace{1cm} (13)

Figure 2.2: Optimum non-coherent SDPSK demodulator.

The high frequency contributions in (12) and (13) can easily be low pass filtered out, leaving only the desired phase information. In practical applications, the correlated signals are integrated over one symbol duration, as shown in Figure 2.2. Thus, the total summed signal energies $E_i$ and $E_q$ over symbol duration $T_b$ can be expressed as in (14) and (15):

\[
E_i = \int_{kT_b}^{(k+1)T_b} r_k(t) \cos(2\pi f_c t) dt
\]  \hspace{1cm} (14)

\[
E_q = \int_{kT_b}^{(k+1)T_b} r_k(t) \cos \left(2\pi f_c t + \frac{\pi}{2}\right) dt = \int_{kT_b}^{(k+1)T_b} r_k(t) \sin(2\pi f_c t) dt
\]  \hspace{1cm} (15)

where $E_i$ represents the energy resulting from in-phase component of the modulated signal and $E_q$ represents the energy resulting from the quadrature phase component. Each of these energies is then multiplied by the $T_b$-delayed quadrature component, summed
and passed through the sign function to resolve the symbol value. Thus, the \((k+1)^{th}\) output symbol value can be computed as follows:

\[
d_{k+1} = \text{sgn}\left[ \cos(\phi_k)\sin(\phi_{k+1}) - \cos(\phi_{k+1})\sin(\phi_k) \right]
\]

(16)

To validate the modeling of the SDPSK modem, it is necessary to assess its performance under AWGN. Additive white Gaussian noise is simply a stochastic process, having a Gaussian amplitude distribution that exhibits a constant power spectral density over the entire span of the frequency domain, as shown in Figure 2.3. In reality, such a signal having constant power over the entire frequency domain does not exist, but serves as a convenient noise model over a finite bandwidth that is typically much smaller than the carrier frequency. AWGN is an ideal model for satellite and deep space links.

![Figure 2.3: Power Spectral Density of White Noise][30]

The average probability of bit error for conventional DPSK under AWGN is given by [31]:

\[
P_b = \frac{1}{2} \exp\left(-\frac{E_b}{N_0}\right)
\]

(17)

where \(E_b\) represents the bit energy and \(N_0\) is the one-sided noise spectral density. When there is no additional ISI, the error probability of SDPSK is the same as DPSK’s [28]. Figure 2.4 shows the simulation results of BER performance for SDPSK under AWGN along with the corresponding theoretical BER. The trends in Figure 2.4 indicate the
modeled performance agrees well with the theoretical results, validating the modem model.

![SDPSK BER Performance without ISI](image)

Figure 2.4: BER performance of SDPSK under AWGN.

### 2.2.2 Gaussian Minimum Shift Keying

Continuous phase modulation (CPM) schemes are a subset of the constant envelope modulation schemes. As the name implies, CPM schemes require the phase modulation be continuous. CPM based schemes are more resistant to adjacent channel interference (ACI) because of their compact power spectrum. That is, the majority of the transmitted signal energy lies in the main spectral lobe or channel bandwidth. It should be noted that because the phase modulation is continuous, each transmitted symbol is dependent on the previous symbols, thereby requiring memory in the modulator. GMSK is a type of CPM technique which has been adopted as the digital modulation scheme for the European Global System for Mobile communications (GSM) standard because of its spectral...
efficiency and constant envelope. These two properties translate into good performance in the presence of ACI and nonlinear amplifiers for GMSK [7]. GMSK requires use of a Gaussian pre-modulation filter which spreads the bit sequence out over longer time by deliberately causing ISI. In exchange for inducing ISI, a higher spectral efficiency can be realized by compressing the data into a smaller bandwidth [32].

The general expression for a CPM signal can be expressed as [33]:

$$S(t) = \sqrt{\frac{2E_s}{T_s}} \cos[2\pi f_c t + \varphi(t, a)]$$  \hspace{1cm} (18)

where $E_s$ is the symbol energy, $T_s$ is the symbol duration, $f_c$ is the carrier frequency, and $a$ represents the NRZ converted data sequence. The phase information represented by the term $\varphi(t, a)$ can be created by applying the sequence $a$ to a filter having an impulse response, $g(t)$, which spreads each data bit over multiple bit intervals. The resulting signal is multiplied by $2\pi h q(t)$, where $h$ represents the modulation index, and $q(t)$ is the time response of the pre-modulation filter. The signal phase is then expressed as [33]:

$$\varphi(t, a) = 2\pi h \sum_{k=-\infty}^{\infty} a_k q(t - kT_s)$$  \hspace{1cm} (19)

and

$$q(t) = \int_{-\infty}^{t} g(\tau) d\tau$$  \hspace{1cm} (20)

By choosing different pulses and varying the modulation index numerous CPM schemes can be obtained.

It is beneficial to first examine minimum shift keying (MSK) to see how the filter characteristics influence the accumulated phase $q(t)$. For MSK, the filter impulse response is given as [33]:
The resulting expression for the accumulated phase value is found by evaluating equation (20) for the filter response in equation (21), as shown in equation (22) [33].

\[
q(t) = \int_{-\infty}^{t} g(\tau) d\tau = \begin{cases} 
\frac{t}{2T_s} & 0 \leq t \leq T_s \\
\frac{1}{2} & t \geq T_s 
\end{cases}
\]

Finally, the signal phase value for MSK modulation is then obtained by substituting \(q(t)\) back into equation (18), yielding [33]

\[
\varphi(t,a) = \pi \left[ \sum_{i=\infty}^{k-1} a_i + a_k \frac{1}{2T_s} (t - kT_s) \right] = a_k \frac{\pi}{2T_s} \left( 1 + \frac{\pi}{2} \sum_{i=\infty}^{k-1} a_i - ka_k \right), \quad kT_s \leq t \leq (k+1)T_s
\]

Figure 2.5 shows the impulse response \(g(t)\) and accumulated phase \(q(t)\) of an MSK signal over one symbol duration, \(T_s\).

As indicated previously, GMSK is a variation of MSK which makes use of a pre-modulation Gaussian filter to improve spectral efficiency. The impulse response of the pre-modulation filter, \(g(t)\), for GMSK is expressed as [33]:

\[
g(t) = \frac{1}{2T_s} \left[ Q \left( 2\pi B_b \frac{t - T_b}{2} \right) - Q \left( 2\pi B_b \frac{t + T_b}{2} \right) \right], \quad 0 \leq B_b T_b \leq 1
\]

where

\[
Q(t) = \int_{-\infty}^{t} \frac{1}{\sqrt{2\pi}} \exp \left( -\frac{\tau^2}{2} \right) d\tau
\]

and \(B_b T_b\) represents the time-bandwidth product. Figure 2.6 shows the impulse response \(g(t)\) and accumulated phase \(q(t)\) of a GMSK signal over one bit duration, \(T_b\), for various
values of $B_c T_b$. Note that the horizontal axis denotes the time duration wherein $T$ is the bit duration $T_b$, and an integer parameter, $L$, is used to control number of symbol lengths over which pulse shaping takes place. This in turn creates ISI over multiple symbols and degrades the performance of GMSK, as compared to MSK.

Figure 2.5: Impulse response and accumulated phase of a MSK signal over one symbol duration, $T_s$. 
Figure 2.6: Impulse response and accumulated phase of a GMSK signal over one symbol duration for various values of $B_b T_b$.

Figure 2.7 illustrates the difference in accumulated phases from Figure 2.5 and Figure 2.6. Therein, GMSK produces ISI over an approximate 4 symbol duration (i.e. $L=4$). Evaluating this using equation (19) yields,

$$
\varphi(t,a) = \pi \left[ \sum_{i=-\infty}^{k-3} \frac{a_i}{2} + \sum_{i=k-3}^{k} a_i q(t-iT_s) \right] = \sum_{i=-\infty}^{k-3} \frac{\pi}{2} a_i + ISI(t), kT_s \leq t \leq (k+1)T_s \quad (26)
$$

From Equation (26), it can be seen that the each ISI term is always less than $\pi/2$, which explains why GMSK has ISI, but MSK does not. Thus, a notable tradeoff between spectral compactness (and in turn, ACI) and ISI is realized for GMSK. Figure 2.8 shows the spectra of an MSK signal as well as several GMSK signals versus the normalized frequency difference from the carrier center frequency $fT_s$, wherein the normalized 3 dB bandwidth-time product of the pre-modulation Gaussian filter is denoted as $B_b T_b$. 

31
Figure 2.7: Comparison of accumulated phases for MSK and GMSK with various values of $B_h T_h$.

Figure 2.8: Power spectra of MSK and GMSK signals having various $B_h T_h$ values.
In order to modulate a signal using GMSK, the data sequence must first be converted from a binary input data stream $I_k \in \{0,1\}$ to an NRZ sequence $a_k \in \{-1,1\}$ and then to a stream of rectangular pulse train $r(t)$ [34].

$$r(t) = \sum_k a_k P(t - kT_s)$$  \hspace{1cm} (27)

where

$$P(t) = \begin{cases} 1, & 0 \leq t \leq T_s \\ 0, & \text{otherwise} \end{cases}$$  \hspace{1cm} (28)

A Gaussian filter with impulse response $g(t)$ as defined in equation (24) is then applied to generate the shaped spectrum $m(t)$

$$m(t) = \sum_k a_k g(t - kT_s)$$  \hspace{1cm} (29)

The carrier frequency must then be modified by $m(t)$ around a center frequency $f_c$. This implies modulating its phase by $\int m(t)$ [34].

$$\varphi(t) = \omega_c t + 2\pi f_m \int_0^t m(\tau) d\tau$$  \hspace{1cm} (30)

The instantaneous frequency of the modulated signal is defined as in equation (31) so that continuity of the GMSK waveform’s phase can be maintained [34]

$$f(t) = \frac{1}{2\pi} \frac{d\varphi}{dt} = f_c + f_m m(t)$$  \hspace{1cm} (31)

where $f_c$ is the carrier frequency, and $f_m$ is the peak frequency deviation. The peak frequency deviation is a function of the bit rate. Because MSK contains $N$ cycles of frequency $f_c - f_m$ and $N+1/2$ cycles of $f_c + f_m$ over one bit duration $T_b$, the peak frequency deviation can be defined as [34]

$$f_m = \frac{1}{4T_b} = \frac{R_b}{4}$$  \hspace{1cm} (32)
The modulated carrier, \( S(t) \) can then be generated as in equation (33) [34].

\[
S(t) = \cos[\varphi(t)]
\]  

The block diagram for the GMSK modulator is shown in Figure 2.9. Based on equation (30), the GMSK modulated signal can be represented in complex envelope form by setting \( \omega_c = 0 \) and taking the real part of the expression. This result is expressed as

\[
\tilde{S}(t) = \exp\left[ j2\pi f_m \int_0^t m(\tau)d\tau \right]
\]  

From equation (18), modulated GMSK signals can be represented in discrete complex envelope form as [35],

\[
\tilde{S}(t) = \exp\left[ j2\pi h \sum_{k=0}^{\infty} a_k q(t - kT_s) \right] = \exp\left[ j2\pi f_m \sum_{k=0}^{\infty} a_k g(t - kT_s) \right]
\]  

where the modulation index \( h = 0.5 \).

Figure 2.9: Block diagram of a GMSK modulator [34].

GMSK modulated signals can be demodulated coherently or non-coherently; however, the additional complexity involved in using a coherent detection scheme is not justified by the marginal improvement gained in BER performance. Hence, the GMSK modem in this study will use non-coherent differential demodulation. Moreover, only one-bit differential detection will be considered.

The block diagram for the one-bit differentially detected GMSK demodulator is shown in Figure 2.10. According to [36], the received AWGN corrupted signal \( x(t) \) at the input of the demodulator is given as
\[ x(t) = \sqrt{2S} \cos[\omega_c t + \phi(t)] + \hat{n}(t) \]  

where \( S \) is the signal power, \( \omega_c \) is the frequency carrier, \( \hat{n}(t) \) is the AWGN term, and \( \phi(t) \) is the transmit filter data phase after frequency modulation (see equation (30)). The IF filter band-limits the input signal \( x(t) \) and produces a time-varying envelope \( \sqrt{2S} \alpha(t) \) along with additional phase corruption. The resultant signal is then [36]

\[ x_{IF}(t) = \sqrt{2S} \alpha(t) \cos[\omega_c t + \phi(t)] + n(t) \]

Let \( n_c(t) \) and \( n_s(t) \) be independent in-phase and quadrature low pass Gaussian random processes, respectively, having mean zero and variance \( \sigma_n^2 \) defined as [36]

\[ \sigma_n^2 = N_0 \int_{-\infty}^{\infty} |H(f)|^2 df = N_0 B_{rn} \]

where \( B_{rn} \) is the two-sided noise bandwidth of the IF filter. The noise term can then be rewritten as [36]

\[ n(t) = n_c(t) \cos[\omega_c t + \phi(t)] - n_s(t) \sin[\omega_c t + \phi(t)] \]

Substitution of (39) into (37) yields [36]

\[ x_{IF}(t) = R(t) \cos[\omega_c t + \phi(t) + \eta(t)] \]

where the signal envelope is represented by [36]

\[ R(t) = \sqrt{[\sqrt{2S} \alpha(t) + n_c(t)]^2 + n_s^2(t)} \]

and

\[ \eta(t) = -\tan^{-1} \frac{n_s(t)}{\sqrt{2S} \alpha(t) + n_c(t)} \]

For a one-bit differential detector, \( x_{IF}(t) \) is multiplied by a one bit duration \( (T_b) \) delayed copy of itself that is phase shifted by \( \pi/2 \) yielding [36]
\[ y(t) = \frac{R(t)R(t-T_b)}{2} \sin[\omega_c T_b + \Delta \phi(T_b)] \]  

(43)

where

\[ \Delta \phi(T_b) \equiv \phi(t) - \phi(t-T_b) + \eta(t) - \eta(t-T_b) \]  

(44)

Figure 2.10: One-bit differential detector for GMSK [36].

If the carrier frequency is chosen so that \( \omega_c T_b = 2\pi k \), where \( k \) is an integer, then (43) can be reduced to [36]

\[ y(t) = \frac{R(t)R(t-T_b)}{2} \sin[\Delta \phi(T_b)] \]  

(44)

The receiver then resolves the transmitted symbol as a ‘1’ if the \( y(t) > 0 \) and ‘0’ otherwise.

In the case of 1-bit differential GMSK, no closed form mathematical expression describing the BER has been published. Consequently, the modem model must be validated empirically by comparing results to those in [36]. In order to observe the effects of the Gaussian filter, different BT product values (0.25, 0.32, 0.40, 0.5, 1) are used. Moreover, the same data rate is assumed for all cases for fair comparison. Table 2.3 summarizes the simulation scenarios for 1-bit differential GMSK based on different BT products.

The performances of GMSK under AWGN for the specified parameters are shown in Figure 2.11. The trends therein match reasonably well with those published in
[36] for BERs of $10^{-3}$ with a margin of less than 0.5 dB, for BT products of 1 and 0.5. For lower value BT products, the differences in BER trends increase slightly towards 1 dB. The difference in BER performances could be attributed to differences in simulation specific implementations (mapping versus sampling based, differences in channel filters specifications, etc.) [36]; hence, the simulation model is justifiably correct.

Table 2.3: Simulation Parameters for 1-bit Differential GMSK Under AWGN.

<table>
<thead>
<tr>
<th>Channel Mode</th>
<th>BT product</th>
<th>Source Data Rate (kbps)</th>
<th>Channel Date (kbps)</th>
<th>Symbol Rate (symbols/s)</th>
<th>Samples per Symbol</th>
</tr>
</thead>
<tbody>
<tr>
<td>GMSK One-bit differential detection</td>
<td>BT=0.25</td>
<td>64</td>
<td>64</td>
<td>64</td>
<td>16</td>
</tr>
<tr>
<td></td>
<td>BT=0.32</td>
<td>64</td>
<td>64</td>
<td>64</td>
<td>16</td>
</tr>
<tr>
<td></td>
<td>BT=0.4</td>
<td>64</td>
<td>64</td>
<td>64</td>
<td>16</td>
</tr>
<tr>
<td></td>
<td>BT=0.5</td>
<td>64</td>
<td>64</td>
<td>64</td>
<td>16</td>
</tr>
<tr>
<td></td>
<td>BT=1</td>
<td>64</td>
<td>64</td>
<td>64</td>
<td>16</td>
</tr>
</tbody>
</table>

Figure 2.11: BER performance of GMSK (various BT products) under AWGN.
2.2.3 Binary Frequency Shift Keying

Although binary frequency shift keying (BFSK) modulation is not considered as a candidate waveform for satellite-based tactical mobile communication in this study, understanding its operation is beneficial for later discussions on group demultiplexing, spread spectrum and tone jamming techniques. BFSK uses two signals with different frequencies to represent binary ‘1’ and ‘0’. The signal can be represented as [31]

\[ S_i(t) = A \cos(2\pi f_i t + \varphi_i), \quad 0 \leq t \leq T_b, \quad i = 1,2 \]  

where \( f_1 \) and \( f_2 \) are two distinct carrier frequencies, \( \varphi_1 \) and \( \varphi_2 \) are the corresponding initial phases at time \( t = 0 \), and \( T_b \) is the bit duration of the binary data. If \( \varphi_1 \neq \varphi_2 \), the two signals are non-coherent and the waveform is not continuous at bit transitions [33]. Note that non-coherent BFSK can only be demodulated through non-coherent means, whereas coherent BFSK can be demodulated by coherent and non-coherent techniques.

To effectively implement BFSK, it is necessary to establish orthogonality between tone carriers. Assuming that the initial phases of \( S_1(t) \) and \( S_2(t) \) are different, the correlation between them must be zero, as defined in (46).

\[
\int_{kT_b}^{(k+1)T_b} S_1(t)S_2(t) dt = \int_{kT_b}^{(k+1)T_b} A \cos(2\pi f_1 t)A \cos(2\pi f_2 t + \varphi) = 0, \quad \text{for} \quad \varphi_1 = 0, \varphi_2 = \varphi \quad (46)
\]

In order for the correlation between tone carriers to be zero, the following must hold true [33]:

\[ 2\pi(f_1 + f_2)T_b = 2n\pi \]  

(47)

and

\[ 2\pi(f_1 - f_2)T_b = m\pi \]  

(48)

where \( n \) and \( m \) are integers. Solving (47) and (48) for \( f_1 \) and \( f_2 \) yields [33]

\[ f_1 = \frac{n + m}{2T_b}, \quad f_2 = \frac{n - m}{2T_b} \]  

(49)
respectively. The difference between tone frequencies for non-coherent BFSK can then be taken as:

\[ f_{\Delta} = \frac{n}{2T_h} \]  

Note that the tone spacing requirements for non-coherent BFSK are twice that of coherent BFSK. A non-coherent BFSK modulator can be implemented by multiplexing the outputs of two different oscillators as dictated by the binary data stream; the corresponding block diagram is shown in Figure 2.12.

![Figure 2.12: Block diagram of non-coherent modulator for BFSK [31].](image)

Herein, non-coherent detection is implemented by using correlators, integrate-and-dump filters and square law detectors [31]. The topology of the corresponding non-coherent detection scheme is shown in Figure 2.13. The decision device resolves the data symbol as a ‘0’ if \( l_1^2 > l_2^2 \), and ‘1’ if \( l_1^2 < l_2^2 \).
Figure 2.13: Block diagram of non-coherent demodulator for BFSK [31].

2.3 Forward Error Correction Coding

Satellite communications is largely unidirectional. Consequently, if a corrupted message is received, the receiver does not have the luxury of requesting a re-transmit from the sender. Rather, the capability to detect and correct bit errors must be contained entirely within the message being transmitted. Forward error correction (FEC) encoding is the process in which an encoded message containing $n$ bits is constructed based on an input sequence containing $k$ bits (where $n > k$) in a given generator scheme. By increasing the length of the bit sequence beyond what is required for representation of the data, error detection and correction functionality can be introduced. The idea is to minimize the effects of noise during transmission by distributing the noise energy over a greater number of bits. This translates into a mechanism for minimizing the likelihood of receiving corrupted bits. The measured difference between the signal-to-noise ratios of the coded and uncoded sequences is referred to as coding gain. There exists many
different types of codes and deciding which to use is largely dependent on the nature of
the end-application.

Based on the literature survey and the implications of mobile communications,
this research study will consider convolutional coding. Convolutional codes are described
by three main parameters, \((n,m,k)\), where \(n\) and \(m\) are as described above and \(m\)
represents the number of memory bits in the encoder. FEC coding schemes can have
various code rates, or the ratio of the number of information bits, \(k\), to the number of
information and redundancy bits in \(n\). Coding rates of 1/2, and 1/3 are considered herein
so that the effect of lower and higher code rates can be ascertained. The constraint length,
\(K\), is defined as \(m+1\) and represents the number of bits in the encoder’s memory which
affect generation of the \(n\) output bits. For this study, the constraint length is chosen as \(K = 9\).

Generation of the code rates 1/2 and 1/3 are achieved using the encoders shown in
Figure 2.14 and Figure 2.15, respectively. These figures are derived from the optimum
convolutional codes for the chosen code rates for a constraint length \(K = 9\) which are
given in Table 2.4 [37]. Thus, for the code rate 1/2, the generator polynomials for each of
the respective outputs are given as:

\[
g^{(1)}(D) = 1+D^2+D^3+D^4+D^8 \tag{51}
\]

\[
g^{(2)}(D) = 1+D+D^2+D^3+D^5+D^7+D^8 \tag{52}
\]

which are then arranged into the subsequent generator matrix:

\[
G(D) = [g^{(1)}(D) ; g^{(2)}(D)] \tag{53}
\]
Similarly, for the convolutional code with rate 1/3, the generator polynomials for each of the respective outputs are given as:

\[ g^{(1)}(D) = 1 + D^2 + D^3 + D^4 + D^5 + D^6 + D^8 \]  
(54)

\[ g^{(2)}(D) = 1 + D + D^4 + D^7 + D^8 \]  
(55)

\[ g^{(3)}(D) = 1 + D + D^2 + D^4 + D^6 + D^7 + D^8 \]  
(56)

Arranging \( g^{(1)}(D) \), \( g^{(1)}(D) \) and \( g^{(1)}(D) \) into the corresponding generator matrix yields:

\[ G(D) = [g^{(1)}(D) ; g^{(2)}(D) ; g^{(1)}(D)] \]  
(57)

Table 2.4: Optimum Rate Convolutional Codes.

<table>
<thead>
<tr>
<th>( g^{(1)} )</th>
<th>Code Rate 1/2</th>
<th>Code Rate 1/3</th>
</tr>
</thead>
<tbody>
<tr>
<td>561_8 = 101 110 001_2</td>
<td>575_8 = 101 111 101_2</td>
<td></td>
</tr>
<tr>
<td>753_8 = 111 101 011_2</td>
<td>623_8 = 110 010 011_2</td>
<td></td>
</tr>
<tr>
<td>727_8 = 111 010 111_2</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
To decode the convolutional encoded binary sequence the Viterbi decoding algorithm is employed. The process begins with construction of a trellis diagram which represents all possible transitions from one state to the next. Assuming an information sequence of length $h$ the trellis diagram contains $h+m+1$ time units. The Viterbi decoder assumes a terminated code which starts in an all zero state and ends in an all zero state. Consequently, not all states can be reached within the first or last $m$ time units; however, in the center of the diagram, all states transitions are possible. There exists $2^k$ branches leaving and entering each state. The decoder calculates (starting from the final state back to the initial state) the path through which the trellis which has the lowest metric in order to decode the received codeword using hard-decision. At each time unit, it adds $2^k$ branch metrics to each previously stored path metric, compares the metrics of all $2^k$ paths entering each state, and selects the path with the lowest metric. The path yielding the lowest metric is known as the survivor. The survivor at each state is then stored along with the metric. The Viterbi algorithm is summarized in [37] as follows:

**Step 1.** Beginning at time unit $t = m$, compute the partial metric for the single path entering each state. Store the path (the survivor) and its metric for each state.

**Step 2.** Increase $t$ by 1. Compute the partial metric for all $2^k$ paths entering a state by adding the branch metric entering that state to the metric of the connecting survivor at the previous time unit. For each state, compare the metric of all $2^k$ paths entering that state, select the path with the lowest metric (the survivor), store it along with its metric, and eliminate all other paths.

**Step 3.** If $t < h+m$, repeat step 2; otherwise, stop.

Note that all decoding is performed using hard decision metric calculation wherein the Hamming distance is taken as the metric.
2.3.1 Performance of Convolutional Coded SDPSK Under AWGN

To illustrate the effect of convolutional coding on the SDPSK modulated waveform under AWGN, a preliminary suite of simulations was conducted using coding rates 1/2, 1/3, 1/4, and 2/3 as well as a data rate of 64 kbps. The parameters for simulation are summarized in Table 2.5. The resulting BER trends are shown in Figure 2.16.

Table 2.5: Simulation parameters for Convolutional Coded SDPSK Under AWGN.

<table>
<thead>
<tr>
<th>Modulation</th>
<th>Code Rate</th>
<th>Source Data Rate (kbps)</th>
<th>Channel Date (kbps)</th>
<th>Symbol Rate (symbols/s)</th>
<th>Sampling frequency (samples/symbol duration)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SDPSK</td>
<td>1/2</td>
<td>64</td>
<td>128</td>
<td>128</td>
<td>16</td>
</tr>
<tr>
<td></td>
<td>1/3</td>
<td>64</td>
<td>192</td>
<td>192</td>
<td>16</td>
</tr>
<tr>
<td></td>
<td>1/4</td>
<td>64</td>
<td>256</td>
<td>256</td>
<td>16</td>
</tr>
<tr>
<td></td>
<td>2/3</td>
<td>64</td>
<td>96</td>
<td>96</td>
<td>16</td>
</tr>
</tbody>
</table>

Figure 2.16: Performance of convolutional coded SDPSK under AWGN.
The simulation results indicate that the performance of coding rate 1/2 is better than that of coding rate 1/3, the performance of coding rate 1/3 is better than that of coding rate 1/4, and the performance of coding rate 2/3 is between coding rate 1/2 and 1/3; however, it should be noted that the margins between the various coding rate performances is reduced as $E_b/N_0$ gets larger. Table 2.6 shows the coding gain of SDPSK at each coding rate wherein a coding gain of approximately 0.5 dB is realized for a BER of $10^{-3}$.

Table 2.6: Performance of the convolutional coded SDPSK under AWGN.

<table>
<thead>
<tr>
<th>Code Rate</th>
<th>Uncoded $E_b/N_0 @ 10^{-3}$ (dB)</th>
<th>Conv. Coded $E_b/N_0 @ 10^{-3}$ (dB)</th>
<th>Coding gain (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1/2</td>
<td>8</td>
<td>7</td>
<td>1</td>
</tr>
<tr>
<td>1/3</td>
<td>8</td>
<td>7.5</td>
<td>0.5</td>
</tr>
<tr>
<td>1/4</td>
<td>8</td>
<td>7.8</td>
<td>0.2</td>
</tr>
<tr>
<td>2/3</td>
<td>8</td>
<td>7.4*</td>
<td>0.6</td>
</tr>
</tbody>
</table>

*: Estimated Value

2.3.2 Performance of Convolutional Coded GMSK Under AWGN

The effects of convolutional coding on 1-bit differentially detected GMSK waveform under AWGN was evaluated for coding rates 1/2, 1/3, 1/4, and 2/3 and BT products of 0.25, 0.32, 0.5 and 1. The simulation suite used the same data rates and number of samples in the case of SDPSK. The parameters for simulation are summarized in Table 2.5. The resulting BER trends are shown in Figure 2.17, Figure 2.18, Figure 2.19 and Figure 2.20, respectively.
Table 2.7: Simulation parameters for Convolutional Coded 1-bit Differential GMSK Under AWGN.

<table>
<thead>
<tr>
<th>Modulation</th>
<th>BT product</th>
<th>Coding Rate</th>
<th>Source Data Rate (kbps)</th>
<th>Channel Date rate (kbps)</th>
</tr>
</thead>
<tbody>
<tr>
<td>GMSK One-bit differential detection</td>
<td>BT=0.25</td>
<td>1/2</td>
<td>64</td>
<td>128</td>
</tr>
<tr>
<td></td>
<td></td>
<td>1/3</td>
<td>64</td>
<td>192</td>
</tr>
<tr>
<td></td>
<td></td>
<td>1/4</td>
<td>64</td>
<td>256</td>
</tr>
<tr>
<td></td>
<td></td>
<td>2/3</td>
<td>64</td>
<td>96</td>
</tr>
<tr>
<td></td>
<td>BT=0.32</td>
<td>1/2</td>
<td>64</td>
<td>128</td>
</tr>
<tr>
<td></td>
<td></td>
<td>1/3</td>
<td>64</td>
<td>192</td>
</tr>
<tr>
<td></td>
<td></td>
<td>1/4</td>
<td>64</td>
<td>256</td>
</tr>
<tr>
<td></td>
<td></td>
<td>2/3</td>
<td>64</td>
<td>96</td>
</tr>
<tr>
<td></td>
<td>BT=0.5</td>
<td>1/2</td>
<td>64</td>
<td>128</td>
</tr>
<tr>
<td></td>
<td></td>
<td>1/3</td>
<td>64</td>
<td>192</td>
</tr>
<tr>
<td></td>
<td></td>
<td>1/4</td>
<td>64</td>
<td>256</td>
</tr>
<tr>
<td></td>
<td></td>
<td>2/3</td>
<td>64</td>
<td>96</td>
</tr>
<tr>
<td></td>
<td>BT=1</td>
<td>1/2</td>
<td>64</td>
<td>128</td>
</tr>
<tr>
<td></td>
<td></td>
<td>1/3</td>
<td>64</td>
<td>192</td>
</tr>
<tr>
<td></td>
<td></td>
<td>1/4</td>
<td>64</td>
<td>256</td>
</tr>
<tr>
<td></td>
<td></td>
<td>2/3</td>
<td>64</td>
<td>96</td>
</tr>
</tbody>
</table>

Figure 2.17: Performance of convolutional coded GMSK (BT=0.25) under AWGN.
Figure 2.18: Performance of convolutional coded GMSK (BT=0.32) under AWGN.

Figure 2.19: Performance of convolutional coded GMSK (BT=0.5) under AWGN.
In Figure 2.17, coding gains are realized in the range of 10 - 15 dB for BT=0.25 (with lower coding rates corresponding to the most coding gains). Similarly, in Figure 2.18 the range of coding gains is 8 - 12 dB for BT=0.32. The spread between coding gains continues to decrease as the BT product is increased. It should be noted that Figure 2.17 and Figure 2.18 exhibit some oscillation in trends at high SNR values. Simulations were conducted using 10,000 data symbols can produce reliable BER accuracy to 10^{-3}. In order to achieve better resolution beyond this threshold the number of data symbols processed would have to increase by an order of magnitude, incurring excessive computational time and resources.

Figure 2.19 shows a coding gain spread of 1 dB (8.5 - 9.5 dB), while the coding gains of Figure 2.20 are overlapped at 8.5 dB for BT=1. For GMSK, the lower the convolutional coding rate, the better the BER performance. As a summary, Table 2.8 -
Table 2.11 show the coding gain of GMSK for all considered coding rates and respective BT products.

Convolutional coding shows substantial coding gain for one-bit differentially detected GMSK with lower BT product values. The lower the coding rate, the better the BER performance is; however, with the BT product increased, the coding rate shows no significant effect on the BER.

Table 2.8: Performance of the convolutional coded GMSK (BT=0.25) under AWGN.

<table>
<thead>
<tr>
<th>Code Rate</th>
<th>Uncoded $E_b/N_0$ @ $10^3$ (dB)</th>
<th>Conv. Coded $E_b/N_0$ @ $10^3$ (dB)</th>
<th>Coding gain (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1/2</td>
<td>21*</td>
<td>15*</td>
<td>6</td>
</tr>
<tr>
<td>1/3</td>
<td>21*</td>
<td>13*</td>
<td>8</td>
</tr>
<tr>
<td>1/4</td>
<td>21*</td>
<td>12</td>
<td>9</td>
</tr>
<tr>
<td>2/3</td>
<td>21*</td>
<td>17*</td>
<td>4</td>
</tr>
</tbody>
</table>

*: Estimated Value

Table 2.9: Performance of the convolutional coded GMSK (BT=0.32) under AWGN.

<table>
<thead>
<tr>
<th>Code Rate</th>
<th>Uncoded $E_b/N_0$ @ $10^3$ (dB)</th>
<th>Conv. Coded $E_b/N_0$ @ $10^3$ (dB)</th>
<th>Coding gain (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1/2</td>
<td>16</td>
<td>12</td>
<td>4</td>
</tr>
<tr>
<td>1/3</td>
<td>16</td>
<td>11.5*</td>
<td>4.5</td>
</tr>
<tr>
<td>1/4</td>
<td>16</td>
<td>11</td>
<td>5</td>
</tr>
<tr>
<td>2/3</td>
<td>16</td>
<td>13</td>
<td>3</td>
</tr>
</tbody>
</table>

*: Estimated Value

Table 2.10: Performance of the convolutional coded GMSK (BT=0.5) under AWGN.

<table>
<thead>
<tr>
<th>Code Rate</th>
<th>Uncoded $E_b/N_0$ @ $10^3$ (dB)</th>
<th>Conv. Coded $E_b/N_0$ @ $10^3$ (dB)</th>
<th>Coding gain (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1/2</td>
<td>12</td>
<td>9</td>
<td>3</td>
</tr>
<tr>
<td>1/3</td>
<td>12</td>
<td>9*</td>
<td>3</td>
</tr>
<tr>
<td>1/4</td>
<td>12</td>
<td>9*</td>
<td>3</td>
</tr>
<tr>
<td>2/3</td>
<td>12</td>
<td>10 (Estimated)</td>
<td>2</td>
</tr>
</tbody>
</table>

*: Estimated Value
Table 2.11: Performance of the convolutional coded GMSK (BT=1) under AWGN.

<table>
<thead>
<tr>
<th>Code Rate</th>
<th>Uncoded $E_b/N_0 \times 10^{-3}$ (dB)</th>
<th>Conv. Coded $E_b/N_0 \times 10^{-3}$ (dB)</th>
<th>Coding gain (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1/2</td>
<td>10.5</td>
<td>9*</td>
<td>1.5</td>
</tr>
<tr>
<td>1/3</td>
<td>10.5</td>
<td>9*</td>
<td>1.5</td>
</tr>
<tr>
<td>1/4</td>
<td>10.5</td>
<td>9</td>
<td>1.5</td>
</tr>
<tr>
<td>2/3</td>
<td>10.5</td>
<td>9</td>
<td>1.5</td>
</tr>
</tbody>
</table>

*: Estimated Value
Chapter 3

Satellite Payload Architecture Modeling

Fully regenerative On-Board Processing (OBP) systems are endorsed in literature as being the most effective architecture for maintaining signal quality under jamming circumstances [1], [3], [38]. Dehop-Rehop Transponder (DRT) systems have been proposed as economical alternatives, bridging the gap between passive repeater and full OBP architectures [39]; however, nothing has been published quantifying and comparing the performances of both DRT and OBP payloads through simulation or closed form analysis.

The objective of this study is to provide analysis and simulation to determine the performance of a frequency hopped (FH), frequency division multiplexed (FDM) waveform through both DRT and OBP satellite payloads under channel impairments towards application in tactical satellite communication systems. Several modems, channel models and the source-to-sink system hardware model (comprised of transmitting earth-terminals, DRT and OBP satellite payloads and receiving earth-terminals) are considered herein. This chapter describes the models of both architectures as well as their corresponding block components.
3.1 Dehop Rehop Transponder and On-Board Processing Architectures

The generalized block diagram of the complete source-to-sink system model is shown in Figure 3.1. A random binary data sequence is passed to the FEC encoder, followed by one of the modulators under study. The resulting modulated signals are then FDM multiplexed onto a single carrier frequency. A group hopping block is then applied to spread the bandwidth of the FDM formatted waveform using slow frequency hopping (i.e. multiple symbols are transmitted per hop). Although typically outperformed by fast frequency hopping (FFH), slow frequency hopping (SFH) can help mitigate the effects of jamming without incurring the expense of fast frequency synthesizing hardware [40]. The hopped waveform is then subjected to jamming and/or AWGN corruption. The satellite payload can be simulated using OBP or DRT architectures whose detailed block diagrams are shown in Figure 3.2 and Figure 3.3, respectively. The downlink transmission signals are subjected to the desired impairments (i.e. downlink jamming and/or AWGN) and recovered at the receiving earth terminals. Note that frequency translation operations are omitted from the proposed simulation model, because of the tremendous overhead incurred in computational resources they incur.
Using OBP processing, the signal is first group dehopped and digitized for further processing. The error resulting from digitization is introduced through a 14-bit quantizer block. Errors resulting from the sample-and-hold (S/H) process are not considered. The digitized signal is then channelized down to baseband using a multi-stage multicarrier
demultiplexing block [41]. After demodulation and decoding, the signal is re-encoded, re-modulated, re-multiplexed and re-hopped. The signal is also passed through a nonlinear TWTA, implemented herein using a frequency independent Saleh’s model [42].

As with OBP processing, the DRT processing payload first performs group dehopping followed by digitization. Digitizing the received signal at intermediate frequency (IF) allows for digital channelization techniques (not performed herein) to be used, giving the advantage of dynamic reconfiguration from TTC. After digitization, the FDM formatted signal is then rehopped and amplified through the nonlinear TWTA.

The subsequent sections of this chapter present detailed descriptions of select block components.

### 3.2 Quantization

The process of digitization consists of a sample and hold circuit (S/H) followed by a quantizer. The block diagram of this process is shown in Figure 3.4. The mathematical details quantitatively presenting this quantizer error term will be presented for both uniform and non-uniform cases. Sample and hold circuits are designed to instantaneously capture the input signal and hold that sample value constant until the next sample is taken. The output resulting from a random input signal \( r_0(t) \) and the ideal sample-and-hold is given as:

\[
r_0(t) = \sum_{n=-\infty}^{\infty} r(nT_s) h_0(t-nT_s)
\]

where \( T_s \) represents the sampling period, \( r(nT_s) = r[n] \) (discrete sequence notation) and represents the ideal samples of \( r(t) \) at \( t = T_s \), and \( h_0(t) \) is the impulse response of the zero-order-hold system, defined as
\[ h_0(t) = \begin{cases} 1 & 0 < t < 1/T_s \\ 0 & \text{otherwise} \end{cases} \] (59)

Equation (58) can be rewritten as:

\[ r_0(t) = h_0(t) \ast \sum_{n=-\infty}^{\infty} r(nT_s)\delta(t - nT_s) \] (60)

Note that equations (58) - (60) have been added here for reference and completeness. Simulations in this research work using Matlab are inherently based on discretized sequences and consequently, no effort will be made to model and compensate for the effects of the sample and hold module. After discretization, the sampled signal is then fed to the quantizer module.

Amplitude quantization maps samples of the continuous waveform into finite set of amplitudes which may or may not be uniform. Quantizers may be midrisers or midtreads, indicating whether the quantizer employs a vertical or horizontal component of the staircase function. Because we have a finite number of mappings, the digitized signal will contain quantization noise (i.e. the difference between the signal and the corresponding mapping level), defined as \( q_e(t) \) in Figure 3.4.

![Diagram of quantization process of signals](image)

Figure 3.4: Model of quantization process of signals.

### 3.2.1 Uniform Quantization with Rounding-Off

Uniform quantizers, as the name suggests, have steps of equal size as shown in Figure 3.5. This particular quantizer is a midriser type. A quantizer input value lies in the range of \( i\Delta - \Delta/2 \) to \( i\Delta + \Delta/2 \), where \( i \) is an integer (including zero) and \( \Delta \) refers to the quantum
(or step size). For any random input signal, error lies in the range of\(-\Delta/2 \leq q_e \leq \Delta/2\).

For the uniform quantizer (assuming uniform distribution of quantization noise), the quantizer variance, or quantizer error, can be simply modeled as [43]:

\[
\sigma_q^2 = \frac{\Delta^2}{12}
\]

where \(q_e\) is the error value taken as the difference between the quantized signal sample and the original sample and \(p(q_e) = 1/\Delta\) is the probability density function of the uniform quantization error [43],[44], assuming uniform density. Thus, after uniform quantization, the \(N_c\) channel FDM formatted digitized signal sequence \(\hat{s}_{RF}(nT_s)\) can be represented as:

\[
\hat{s}_{RF}(nT_s) = \sqrt{\frac{2E_s}{T_s}} \sum_{k=-\frac{N_c}{2}}^{\frac{N_c}{2}} \sum_{n=-\infty}^{\infty} \cos[2\pi(f_{LO} + f_k) + \theta]nT_s + q_e(nT_s)
\]

where \(k\) represents the index of a given channel sub-carrier and \(n\) is the index of the sampled sequence.

Assuming that the number of quantization levels provides a fine enough resolution, the error resulting from this quantization behaves as if it were an additive independent source of noise with zero mean and mean-square value governed by the step size \(\Delta\) [44].
3.2.2 Non-Uniform Quantization

Uniform quantizers are the most robust means of quantizing signals, as they are relatively insensitive to small changes and more responsive to larger changes; however, when considering the dynamics involved in communication signals, the robust behavior of the uniform quantizer is quite undesirable. That is, ideally it is best to use a quantizing scheme, wherein small changes in the signal dynamics are captured using small step sizes and large changes in signal dynamics are captured using large step sizes. It is for this reason that non-uniform quantizers are preferred in communication systems. In order to achieve this behavior, it is necessary to first pass the signal through a compressor, feed the compressed signal as an input to a uniform quantizer, and finally expand the signal using an expander (which is precisely the inverse function of the compressor). A device encompassing both compression and expansion modules is known as a compander. Figure 3.6 illustrates the signal companding process. Using a logarithmic compressor, for
example, implements this type of behavior and because of the nature of the logarithmic
scale, equal distances (or errors) represent equal ratios. That is to say, logarithmic
compression results in the signal having a constant signal-to-noise ratio. Thus, the signal
can be reconstructed with no degradation other than that introduced by the quantizer
itself.

Figure 3.6: Non-uniform quantization (companding) process [45].

The two most commonly used non-uniform quantization schemes are A-law (used in
Europe) and μ-law (used in North America).

The A-law compressor has is defined by the following characteristic [45]:

\[
y = \begin{cases} 
    \frac{Ax}{1 + \ln A} & 0 \leq x \leq \frac{1}{A} \\
    \frac{1 + \ln A}{1 + \ln A} & \frac{1}{A} \leq x \leq 1 
\end{cases}
\]

and has regions reflecting both linear (central region) and logarithmic (outer regions)
characteristics. Its quantum intervals are defined as [45]:

\[
\frac{dx}{dy} = \begin{cases} 
    \frac{k_1}{A} & 0 \leq x \leq \frac{1}{A} \\
    k_1x & \frac{1}{A} \leq x \leq 1 
\end{cases}
\]
Equation (64) indicates that the quanta used for small signals are reduced by a factor of $A/k_1$ (typically 25 dB as compared to uniform quantization). Typical values for $A$ are approximately 100 and the constant $k_1$ is defined by [45]:

$$k_1 = 1 + \ln A$$

(65)

The noise bounds are derived in [45] and are found to be:

$$\frac{k_1^2}{3N^2} \leq \frac{E^2}{S^2} \leq \frac{k_1^2}{3N^2} \left(1 + \frac{1}{A^2 S^2}\right)$$

(66)

where $A$ and $k_1$ are as described before, $N$ represents the total number of quantization levels, and $S$ is the signal amplitude distribution. For $S$ such that all amplitudes are less than $1/A$, the quantizing is uniform and the noise is a constant given as [45]:

$$E^2 \geq \frac{k_1^2}{3N^2 A^2}$$

(67)

For the industry standard value of $A = 87.56$, the average SNR is 38.0 dB [45].

The $\mu$-law compander was developed by Bell Systems and has the following characteristic [45]:

$$y = \frac{\ln(1 + \mu x)}{\ln(1 + \mu)}$$

(68)

where $\mu$ is a positive constant, and $x$ is the maximum signal amplitude. The logarithmic step size is proportional to [45]:

$$\frac{1}{x} \frac{dx}{dy} = \ln(1 + \mu) \left(\frac{1 + \mu x}{\mu x}\right)$$

(69)

which, at high levels tends asymptotically towards [45]:

$$k_2 = \ln(1 + \mu)$$

(70)

The error bounds are derived in [45] and are found to be:

$$\frac{k_2^2}{3N^2} \left(1 + \frac{1}{\mu^2 S^2}\right) \leq \frac{E^2}{S^2} \leq \frac{k_2^2}{3N^2} \left(1 + \frac{1}{\mu S}\right)^2$$

(71)
It is noted in [45] that A-law compression always results in a more uniform SNR over the range of levels than μ-law compression. Thus, non-uniform quantization with A-law compression is selected for this study and the number of quantization bits is 14.

### 3.3 Multi-stage Multicarrier Demultiplexing Group Demodulator

When processing received signals that are formatted using FDMA or MF-TDMA, a group demultiplexer is needed to filter each sub-band and bring it to baseband for demodulation and decoding. This “multirate signal processing” is achieved largely through use of filter banks and decimators. Filter banks that decompose a signal into \(N\) signal sub-bands are called analysis filter banks and those that combine \(N\) signal sub-bands are referred to as synthesis filter banks. The two mainstream techniques for implementing a group demultiplexing scheme are polyphase DFT (PDFT) which are derived from uniform filter banks, and Multi-stage Multicarrier Demultiplexer (MMCD). Both PDFT and MMCD methods exhibit comparably low computational complexity; however MMCD exhibits a higher degree of flexibility over its PDFT counterpart. Hence, this study will consider only MMCD group demultiplexing which can be developed from a two channel quadrature mirror filter (QMF), whose topology is shown in Figure 3.7.

![Figure 3.7: Two channel QMF with analysis and synthesis filter bank [46].](image-url)
Although Figure 3.7 shows both the analysis and synthesis sections (left-half and right-half, respectively) of the QMF, our focus will be limited to that of the analysis portion only. The output of the analysis filters can be expressed in the z-domain as [46]

\[ X_k(z) = H_k(z)X(z) \quad k = 0, 1 \]  

(72)

and the output of the decimated sequences are given as [46]:

\[ C_k(z) = \frac{1}{2} \left[ X(z^{1/2}) + X(z^{-1/2}) \right] \quad k = 0, 1 \]  

(73)

The second term in equation (73) reflects the possible aliasing factor that can arise from the input signal and the frequency response of the applied filters. The output of the upsamplers can be described as [46]:

\[ Y_k(z) = C_k(z^2) = \frac{1}{2} \left[ H_k(z)X(z) + H_k(-z)X(-z) \right] \]  

(74)

Finally, the reconstructed signal is given by [46], [47]:

\[ \hat{X}(z) = G_0(z)Y_0(z) + G_1(z)Y_1(z) \]  

\[ = \frac{1}{2} \left[ H_0(z)G_0(z) + H_1(z)G_1(z) \right]X(z) + \frac{1}{2} \left[ H_0(-z)G_0(z) + H_1(-z)G_1(z) \right]X(-z) \]  

(75)

Note that the second term in (75) contains aliasing from downsampling and imaging from upsampling. The effects of this aliasing and term can be seen in Figure 3.8.
In order for us to reconstruct the signal using the first term in (75), the second term must equate to zero. That is,

\[ [H_0(-z)G_0(z) + H_1(-z)G_1(z)]X(-z) = 0 \]  

(76)

which is satisfied under the following conditions \( G_0(z) = H_1(-z) \) and \( G_1(z) = H_0(-z) \).

Implementation of the sub-band filters can be simplified when they respect the mirror filter relationship given by [47]:

\[ H_1(z) = H_0(-z) \]  

(77)

In the frequency domain, the corresponding mirror relationship is given by [47]:

\[ H_1(e^{j\omega}) = H_0(e^{j(\omega-\pi)}) \]  

(78)

Let \( H(z) \) represent our prototype filter having the following z-transform [47]:

\[ H(z) = \sum_{i=-L}^{L} h_i z^{-i} \]  

(79)

The corresponding mirror filter \( H_{mirror}(z) \) can be implemented as follows [47]:

Figure 3.8: Spectral illustration of aliasing in QMF bank [47].
\[ H_{\text{mirror}}(z) = H(-z) = \sum_{i=-L}^{L} h_i (-z)^{-i} = \sum_{i=-L}^{L} (-1)^i h_i z^{-i} \]  

(80)

Therefore, a mirror filter can be generated by simply executing a sign change operation of \((-1)^i\) on the parent filter coefficients.

A special circumstance arises when FIR half-band filters are employed in QMF structures. Digital (FIR) half-band filters have the property that their transition bands are symmetric about the quarter sampling frequency, \(f_{\text{transition}} = f_s/4 = f_{\text{Nyquist}}/2\). That is, the passband frequency \(\omega_p\) and the stopband frequency \(\omega_s\) should sum to \(\pi\). The transfer function of a half-band filter can be derived from the expression for a factor of \(M=2\) polyphase decomposed lowpass filter as [48]:

\[ H(z) = \alpha + z^{-1} E_1(z^2) \]  

(81)

The impulse response of the transfer function \(H(z)\) satisfies the following condition [48]:

\[ h[2n] = \begin{cases} 
\alpha & n = 0 \\
0 & \text{otherwise} 
\end{cases} \]  

(82)

where \(\alpha\) assumes the value of 0.5. Based on equation (81), a special property of these filters becomes evident; with the exception of the center tap coefficient, all even-numbered coefficients (i.e. every other coefficient) are zero. For a value of \(\alpha = 0.5\), equation (81) can be rewritten as [48]:

\[ H(z) + H(-z) = 1 \]  

(83)

and consequently, if \(H(z)\) has real coefficients, then equation (83) implies [48]:

\[ H(e^{j\omega}) + H(e^{j(\pi-\omega)}) = 1 \]  

(84)

Thus, when compared to an arbitrary FIR filter, the half-band filter requires only half the number of coefficient multiplies per filter cycle. This constitutes a particularly
attractive characteristic for decomposing multiplexed signals. The frequency response of a typical half-band filter is shown in Figure 3.9. As noted previously, \( \omega_p = \pi/2 \) and \( \omega_s = 3\pi/4 \) sum to \( \pi \). It should also be noted that if the filter is implemented with length \( S \), then \( S+1 \) must be divisible by four, in order to maintain non-zero first and last coefficients [49].

![Figure 3.9: Frequency response of the FIR low and high pass halfband filters.](image)

The QMF analysis structure using half-band filters can be used to implement the MMCD. The basic topology of MMCD structure (4 channels in this case) is shown in Figure 3.10, where each processing block \( M \) represents the process shown in Figure 3.11.

The MMCD technique demultiplexes the carrier signal by recursively dividing the spectrum into high and low pass components centered at the carrier frequency. This process is repeated the multiple stages of filtering and decimation until, at last, all channels have been separated. The nature of the algorithm is such that maximum efficiency is obtained when the number of channels, \( N \), to be processed is a power of two; thus the channels are related to the number of stages by \( N_c = 2^L \), where \( L \) is the number of stages required in the MMCD. \( N - 1 \) modules are needed in all to demultiplex \( N \) channels. The number of stages \( L \) is governed by:
\[ L = \log_2 N \] (85)

It should be noted that after any demultiplexing stage, the signal must be bandlimited to \( \pi \).

![Block diagram of 4-channel MMCD](image)

Figure 3.10: Block diagram of 4-channel MMCD [50].

![Block diagram of MMCD processing module](image)

Figure 3.11: Block diagram of MMCD processing module.

The computational complexity of this approach has been given in [51] as:

\[
M_{MS} = \left[ \left( \frac{N_F}{2} + 1 \right) \left( \log_2 N_c - \frac{1}{2} \right) + N_G \right] 2^w
\] (86)

where \( N_F \) is the number of coefficients of the half-band filters, \( N_G \) is the number of coefficients of the last filter, \( N_c \) is the number of channels being demultiplexed and \( w \)
denoted the bandwidth of any single demultiplexed channel. The MMCD approach has the advantages of high modularity, design feasibility and structural compactness, while providing moderate flexibility and computational efficiency. Note that the system topology (i.e. binary tree structure) prohibits use of fast computational algorithms, such the discrete Fourier transform (DFT). Moreover, the efficiency of this technique decreases as the number of channels increases.

A preliminary simulation has been conducted based on the parameters shown in Table 3.1. The simulation takes place after the signal has been downconverted and does not include any of the signal preprocessing elements. Furthermore, this validation simulation will not include the effects of AWGN and other interference factors, so that correct operation of the MMCD filter bank can be verified. The input to the MMCD filter bank will be a four channel composite NC-QFSK signal (for illustrative purposes only) shown in Figure 3.12.

Table 3.1: Parameters for MMCD group demultiplexer simulation.

<table>
<thead>
<tr>
<th>Modulation Scheme</th>
<th>NCMFSK</th>
</tr>
</thead>
<tbody>
<tr>
<td>$M$</td>
<td>4</td>
</tr>
<tr>
<td>Number of Channels ($N_c$)</td>
<td>4</td>
</tr>
<tr>
<td>Symbol rate ($S_R$)</td>
<td>1000 Hz</td>
</tr>
<tr>
<td>Carrier Frequency ($f_c$)</td>
<td>20 kHz</td>
</tr>
<tr>
<td>Sampling Frequency ($f_s$)</td>
<td>$4f_c = 80$ kHz</td>
</tr>
<tr>
<td>Channel Guard Band ($GB$)</td>
<td>$2S_R = 2000$ Hz</td>
</tr>
<tr>
<td>Tone Spacing ($T_s$)</td>
<td>$2S_R = 2000$ Hz</td>
</tr>
<tr>
<td>Number of symbols</td>
<td>250</td>
</tr>
</tbody>
</table>
Figure 3.12: Composite 4-channel NCQFSK signal used in validation simulation of MMCD group demultiplexer.

Figure 3.13: Low (left) and high (right) quadrature mirror halfband filters used in MMCD group demultiplexing.
Per the discussion on QMFs and MMCDs, the basic processing block will be composed of low and high digital half-band filters. Their respective frequency response spectra are shown in Figure 3.13. After passing the multiplexed signal of Figure 3.12 through the first stage processing block of the MMCD filter bank shown in Figure 3.14 the following (1-sided) spectra results:

![Lower Branch After 1st Stage MMCD](image1)

![Upper Branch After 1st Stage MMCD](image2)

Figure 3.14: Spectral plot of first stage output of 4-channel MMCD group demultiplexer.

The spectra of channels 1 and 2 are shown in the top portion of Figure 3.14 (from left to right) and channels 3 and 4 are shown in the bottom portion (from left to right). When compared to Figure 3.12 we can see that the respective channel spectra match. Figure 3.15 shows the resulting (one-sided) spectra after the low and high branch are processed through the second stage blocks. It is evident that all share a common carrier frequency
that can then be demodulated using a common hardware block. The results of this simulation verify that the originally transmitted data sequence can be recovered without error using tree structured filter banks. It is now possible to perform initial studies on the BER of the OBP system under various coding and modulation schemes.

![Spectral plot of demultiplexed channels using 4-channel MMCD group demultiplexer.](image)

Figure 3.15: Spectral plot of demultiplexed channels using 4-channel MMCD group demultiplexer.
3.4 Frequency Hopped Spectrum Spreading

Spectrum spreading is used to combat the detrimental effects of jamming or interference. There are two basic types of spectrum spreading techniques used in digital communications: direct sequence (DS) and frequency (FH) or time hopping (TH). In general, DS spread spectrum is appropriate for applications using coherent detection. On the contrary, non-coherently detected schemes are often associated with FH spread spectrum. Since this study is concerned only with non-coherently detected schemes, the focus is restricted to FH spread spectrum. Figure 3.16 shows the typical block diagram of FH/SS communication system at the modem level.

![Block diagram of spread spectrum communication system.](image)

In FH spectrum spreading, the available channel bandwidth $W$ is subdivided into a large number of non-overlapped frequency slots. In any transmitted signal slot, the signal hops to one or more of the available frequency slots, which is controlled by a pseudorandom sequence in principle. The frequency hopping rate, $R_h$, is chosen to be higher or lower than the symbol rate $R_s$. If $R_h \leq R_s$, the FH system is said to be slow frequency hopped (SFH). If $R_h \geq R_s$, the FH system is said to be fast frequency hopped (FFH). Figure 3.17 shows an example of SFH and FFH using NC-BFSK for comparison.
3.5 Nonlinear Traveling Wave Tube Amplifier

Traveling wave tube amplifiers (TWTAs) are often employed for commercial satellite communications applications. A TWTA consists of the TWT itself and an associated electronic power conditioner (EPC), which converts the satellite bus voltage (typically 50V to 100V dc) to the necessary electrode voltages (kV dc) for the TWT. And while they do possess desirable efficiency characteristics, it is well known that TWTAs are highly non-linear, imparting amplitude and phase distortions upon the signal of interest. Figure 3.18 shows an example amplitude and phase characteristic for a TWTA. It is evident that the TWTA exhibits reduced non-linearity when operating with a reduced input power, i.e. with notable input (or output) back-off; however, operation of the amplifier with too much back-off will result in a prohibitively low efficiency. Therefore, to maintain reasonable efficiency, the TWTA must be operated close to the region of saturation. Nevertheless, degradation of the signal of interest is manifested as harmonics.
which result in ACI as well as inter-modulation. These effects are illustrated in Figure 3.19.

Figure 3.18: TWTA amplitude (left) and phase (right) characteristics.

Figure 3.19: Widening of a rectangular signal spectrum by non-linearity in the presence of noise [52].
The Saleh model is a widely accepted TWTA memoryless nonlinear amplifier model which accounts for disturbances in amplitude, $A(r)$, and phase, $\Phi(r)$, as a function of input signal, $x(t)$. Thus, it will be used throughout this study. In order to apply this model, the input signal is assumed to take the form [42]:

$$x(t) = r(t) \cos[\omega_0 t + \phi(t)]$$  \hspace{1cm} (87)

where $\omega_0$ is the carrier frequency, and $r(t)$ and $\Phi(t)$ are the envelope and phase of the input signal, respectively. Four parameters, namely, $\alpha_a$, $\beta_a$, $\alpha_\Phi$, and $\beta_\Phi$, determine the transfer characteristic for amplitude and phase of the particular TWTA through equations (88) and (89), respectively.

$$A(r) = \alpha_a r / (1 + \beta_a r^2)$$  \hspace{1cm} (88)

$$\Phi(r) = \alpha_\Phi r^2 / (1 + \beta_\Phi r^2)$$  \hspace{1cm} (89)

The output of the TWTA is given as [42]:

$$y(t) = A[r(t)] \cos[\omega_0 t + \psi(t) + \Phi[r(t)]]$$  \hspace{1cm} (90)

The specific values for $\alpha_a$, $\beta_a$, $\alpha_\Phi$, and $\beta_\Phi$ describing the TWTA model used in this study are taken from [42] and are summarized Table 3.2.

Table 3.2: Parameters of the Saleh model TWTA.

<table>
<thead>
<tr>
<th>TWTA Model Characteristics</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\alpha_a$</td>
<td>1.9638</td>
</tr>
<tr>
<td>$\beta_a$</td>
<td>0.9945</td>
</tr>
<tr>
<td>$\alpha_\Phi$</td>
<td>2.5293</td>
</tr>
<tr>
<td>$\beta_\Phi$</td>
<td>2.8168</td>
</tr>
<tr>
<td>Input Saturation Point</td>
<td>1.022 V</td>
</tr>
<tr>
<td>Output Saturation Point</td>
<td>0.9846 V</td>
</tr>
<tr>
<td>Output Phase Value $@\Phi$ Saturation</td>
<td>0.6606 rad.</td>
</tr>
</tbody>
</table>
In order to counter the non-linear effects of TWTA, the amplifier is backed-off from the input or output saturation point, depending on whether input back-off (IBOF) or output back-off (OBOF) is sought, down to the linear portion of the characteristic curve. Calculation of IBOF can be achieved by defining the new operating point as [53]:

\[
r(t)_{IBOF} = 10^{-x/20}
\]  

(91)

Then the input backed-off \(A[r(t)]\) and \(\Phi[r(t)]\) can be obtained by substituting the argument \(r(t)\) with \(r(t)_{IBOF}\). To apply \(x\) dB of OBOF, the corresponding input level of the new operating point can be computed using the following equations [53],

\[
r(t)_{OBOF} = \frac{\alpha_a - \sqrt{\alpha_a^2 - 4A[r]^2 \beta_a}}{2A[r] \beta_a}
\]  

(92)

where

\[
A[r]_{OBOF} = 10^{-x/20}
\]  

(93)

Preliminary simulations illustrating the degrading effect of TWTA on the BER of SDPSK, GMSK, and NCBFSK schemes under AWGN were conducted using the described TWTA model; the corresponding simulation results are shown from Figure 3.20 - Figure 3.25.

The difference in performance between the various IBOF and OBOF values when considering SDPSK (Figure 3.20 and Figure 3.21) and GMSK (Figure 3.22 and Figure 3.23) is marginal. In general, the larger IBOF / OBOF, yields better BER performance. It should be noted that an aberration is seen for SDPSK with OBOF (Figure 3.21). These results are reasonable, when considering the amplitude and phase characteristic curves presented and that SDPSK and GMSK are binary signaling schemes. Figure 3.24 and Figure 3.25 present results of the same simulation using NCBFSK wherein the trends
remain tightly grouped. In Figure 3.24, we see that the larger the IBOF, the worse the performance; in contrast, Figure 3.25 shows the opposite effect for OBOF.

In general for these modem level, single channel simulations, the effect of the TWTA and its nonlinearities is not significant; however, when coupled with multiple channel and additional impairments (i.e. jamming, ACI, etc.) the effects can become more pronounced. The TWTA in this study will be operated with 3 dB OBOF, to ensure a reasonable degree of operating efficiency while mitigating some of the nonlinearity.

Figure 3.20: SDPSK BER under AWGN and TWTA with various IBOF.
Figure 3.21: SDPSK BER under AWGN and TWTA with various OBOF.

Figure 3.22: 1-Bit DD GMSK BER under AWGN and TWTA with various IBOF.
Figure 3.23: 1-Bit DD GMSK BER under AWGN and TWTA with various OBOF.

Figure 3.24: NCBFSK BER under AWGN and TWTA with various IBOF.
Figure 3.25: NCBFSK BER under AWGN and TWTA with various OBOF.
Chapter 4

Impairment

When considering a source-to-sink satellite communication system, signal integrity can be compromised from any number of sources related to the transmission channel or system hardware, including the background thermal noise (i.e. additive white Gaussian noise (AWGN)), signal jamming, quantization, non-linear amplification, etc. In the previous chapter, impairment resulting from inherent hardware limitations, such as nonlinear TWTA and quantization noise was discussed in detail. This chapter focuses on impairments presented in the transmission channel. The critical channel impairments for the context of this study are partial-band noise jamming (PBNJ), and band-multitone jamming (BMTJ). A detailed discussion on the effects the aforementioned channel impairments, their corresponding modeling techniques and effects on the waveforms of interest is presented.

4.1 Modem Performance Under Partial Band Noise Jamming

A partial-band noise jammer (PBNJ) evenly distributes noise power $J$ over some frequency bandwidth $W_J$, which is a subset of the total spread bandwidth $W_{SS}$. It can be regarded as a concentrated Gaussian noise jammer in that it restricts its total power, $J$, to
a fraction $\rho$ ($0 \leq \rho \leq 1$) of the full spreading bandwidth, $W_{SS}$, as depicted in Figure 4.1. Ratio of the jamming bandwidth to the entire spreading bandwidth, $\rho$, is defined as [54]:

$$\rho = \frac{W_J}{W_{SS}}$$

(94)

Hence, the jamming noise power is spread uniformly over $W_J = \rho W_{SS}$, resulting in an increased jammer power spectral density [54],

$$\frac{J}{\rho W_{SS}} = \frac{N_J}{\rho}$$

(95)

where $N_J$ represents the jammer power spectral density of broad band noise jammer expressed as $N_J = J/W_{SS}$. This results in the degraded signal to jammer ratio (SJR) level as defined by [54],

$$\frac{E_b}{N_J / \rho} = \frac{\rho E_b}{N_J}$$

(96)

Figure 4.1: Partial Band Noise Jammer Frequency Distribution [54].

### 4.2 Performance of Uncoded, SFH Modems Under PBNJ

Because partial-band noise jamming is applied only towards a fraction of the total operating frequency band, a particular transmitted data symbol is successfully jammed with probability $\rho$ or unsuccessfully jammed with a probability $1 - \rho$. The error probability can then be expressed as [54]

$$P_b = (1 - \rho) \cdot P_1 + \rho \cdot P_2 \leq \rho \cdot P_2$$

(97)
where $P_1$ represents the bit error probability when jamming is absent from the transmission channel (i.e. $P_1 = 0$), and $P_2$ represents the bit error probability when jamming is present. The error probability for SDPSK under PBNJ can be derived from the theoretical error performance equation for SDPSK under AWGN by simply replacing the noise spectral density term $N_0$ with the concentrated jamming power spectral density divided by $\rho$, $N_J/\rho$. Therefore, the bit error rate for SDPSK under PBNJ only is given as [55]:

$$P_b = \frac{P}{2} e^{-\rho(2E_b/N_J)}$$

(98)

Figure 4.2 shows the simulation BER performance of SFH/SDPSK under PBNJ for parameters specified in Table 4.1. In order to practically simulate PBNJ environments, a probabilistic approach is taken. Thus, $\rho$ represents the probability that the signal transmission lies in the portion of the band which is experiencing jamming. Further note that the simulation was conducted at the modem level. It is evident from the trends in Figure 4.2 that for small values of $E_b/N_J$, $\rho = 1$ is the most effective jamming strategy – this corresponds to AWGN. As the value of $E_b/N_J$ is increased, use of smaller values of $\rho$ yields the most effective jamming strategy because of the increased jamming intensity.

Table 4.1: Simulation parameters for SFH/SDPSK under PBNJ.

<table>
<thead>
<tr>
<th>Date rate</th>
<th>SFH</th>
<th>Hopping rate</th>
<th>$\rho$</th>
</tr>
</thead>
<tbody>
<tr>
<td>256 kbps</td>
<td>16 symbols/hop</td>
<td>16 khops/sec</td>
<td>0.0001, 0.001, 0.01, 0.1, 0.5, 1</td>
</tr>
</tbody>
</table>
The worst-case performance of SFH/SDPSK can be found by differentiating (98) with respect to $\rho$ [55]

$$
\frac{dP_b}{d\rho} = \left[ -\rho \frac{E_b}{2N_j} + \frac{1}{2} \right] \exp\left[ -\rho \frac{E_b}{N_j} \right] = 0
$$

Solving for optimum $\rho$ gives [55]

$$
P_0 = \frac{1}{E_b / N_j}
$$

Substituting $P_0$ from (100) into (98) results in [55]

$$
P_{\rho_{\text{max}}} = \frac{1}{2E_b / N_j} \exp(-1)
$$

which is valid only for $P_0 \leq 1$. If $P_0 > 1$, then its value is taken as unity. Hence, the worst-case performance of SDPSK under PBNJ is given by [55]
\[
P_{\text{wc}} = \begin{cases} 
\frac{\exp(-1)}{2(E_b / N_J)} & Eb / NJ \geq 1.0 \\
\frac{1}{2} \exp(-E_b / N_J) & Eb / NJ < 1.0 
\end{cases}
\] (102)

Figure 4.3 shows the worst case performance for SFH/SDPSK. The trends therein where generated empirically by subjecting the waveform to various values of \( \rho \) and selecting the result that yielded highest BER for the specific \( E_b/N_J \) value being applied. As can be seen, the simulation of SFH/SDPSK under worst case PBNJ follows the theoretical bound of (102) well.

\[\text{Figure 4.3: Simulated BER performance of uncoded, SFH/SDPSK WCPBNJ.}\]

For the case of 1-bit differentially detected GMSK, no mathematical formulas describing the BER performance under AWGN are available. Consequently, its performance must be evaluated empirically using the specified simulation parameters in
Table 4.2 for PBNJ impairment. Figure 4.4 - Figure 4.6 show the comparisons of uncoded, SFH/GMSK BER performances for various BT product values under PBNJ for BT = 0.25, 0.5, and 1, respectively. It can be seen that PBNJ degrades the BER performances of uncoded, SFH/GMSK signals more than AWGN, when $E_b/N_J$ exceeds a certain threshold because of the increased jammer intensity.

Table 4.2: Simulation parameters for SFH/GMSK under PBNJ.

<table>
<thead>
<tr>
<th>Date rate</th>
<th>SFH</th>
<th>Hopping rate</th>
<th>BT</th>
<th>ρ</th>
</tr>
</thead>
<tbody>
<tr>
<td>256 kbps</td>
<td>16 symbols/hop</td>
<td>16 khops/sec</td>
<td>0.25, 0.5, 1</td>
<td>0.0001, 0.001, 0.01, 0.1, 0.5, 1</td>
</tr>
</tbody>
</table>

Figure 4.4: Simulated BER performances of uncoded SFH/GMSK under PBNJ for BT = 0.25.
Figure 4.5: Simulated BER performances of uncoded SFH/GMSK under PBNJ for $BT = 0.5$.

Figure 4.6: Simulated BER performances of uncoded SFH/GMSK under PBNJ for $BT=1$. 
As can be seen from the simulation results, the greater the value of $\rho$, the worse the BER performance for smaller values of $E_b/N_J$; however, beyond a certain threshold, PBNJ becomes the dominant degrading factor and smaller values of $\rho$ most severely degrade the BER. In addition, larger BT product values yield better BER performances because of the inclusion of more signal energy.

In the case of SFH/GMSK under worst case PBNJ, it is not possible to directly obtain the maximization of $P_b$ by mathematical derivation, as there is no mathematical expression for the BER of GMSK; however, it can be approximated through empirical means. The performance of SFH/GMSK under different values of $\rho$ is simulated to get the BERs for fixed increments of $E_b/N_J$. The worst case PBNJ can then identified by choosing the value of $\rho \in \{0.0001, 0.001, 0.01, 0.1, 0.5, 1\}$ that corresponds to the maximum value of $P_b$.

The results of the empirically determined worst case PBNJ for SFH/GMSK are shown in Figure 4.7. Therein, a nearly linear BER trend describes the relationship between $E_b/N_J$ and the corresponding BER for all BT values considered. As expected, the lower the BT product value (and in turn the amount of signal energy contained in the waveform) the worse the BER performance.
Figure 4.7: BER performances of uncoded SFH/GMSK Under Worst Case PBNJ.

4.3 Band Multitone Jamming (BMTJ)

A multitone jammer (MTJ) divides its total jamming power, J, into Q distinct equal powered random phase continuous wave (CW) tones. As a jamming strategy, it is typically more effective than PBNJ, because it is easy for a jammer to insert additional energy into the non-coherent detector. MTJs can be classified into two categories: band multitone jamming (BMTJ) and independent multitone jamming (IMTJ). The BMTJ strategy places n tones into each “jammed” M-ary band, where $n \in \{1, M\}$, for M-ary FSK and $n = 1$ for PSK classes, which include SDPSK and GMSK. IMTJ distributes jamming tones pseudo-randomly over all the available frequency hopped (FH) slots.

Figure 4.8 shows an example of both band multitone and independent multitone jamming strategies. For clarity of understanding, the jamming strategies are applied
towards an FH/NC-QFSK transmission. Note that for SFH systems, the minimum tone spacing, $R_c$, for FH/MFSK and FH/SDPSK is equivalent to the data transmission symbol rate [54], [55]. The power for each tone jammer can be calculated as,

$$J_t = \frac{J}{Q}$$

(103)

As indicated in Figure 4.8, given the tone spacing $R_c$, the total number of available FH slots is given as [54],

$$N_t = \frac{W_{SS}}{2R_c}, \text{for SDPSK and GMSK}$$

(104)

and

$$N_t = \frac{W_{SS}}{MR_c}, \text{for M-ary FSK}$$

(105)

Suppose $\alpha$ is the ratio between the signal power and the jamming tone power, i.e.

$$\alpha = \frac{S}{J_t}$$

(106)

The successful jamming of the signal tone is possible only where $\alpha$ is less than unity; note that for FH/SDPSK and FH/GMSK, there is only one signal frequency. It is assumed that the jammer has no forehand knowledge of the hopping sequence; as a result, it is advantageous for the jammer to place its tones in as many $M$-ary hopping bands as possible, even if most of these jammed bands contain a single tone. As with PBNJ, in order to practically simulate a BMTJ environment, a probabilistic approach must be taken. Herein, a parameter $\mu$ is used to represent probability that any symbol in a given M-ary band is jammed [54]. This study is concerned solely with the BMTJ jamming strategy.
4.4 Performance of Uncoded, Modems Under BMTJ

In order to analyze the effect of BMTJ impairment on SDPSK and GMSK, the channel is initially assumed to be dominated by tone jamming only (i.e. AWGN is not considered). Moreover, it is assumed that each of the jamming tone carrier frequencies coincides precisely with one of the $N$ available FH slots, with at most one tone per slot. When considering SFH/SDPSK, the total jammer power, $J$, is partitioned into $Q$ tones, and each FH band has bandwidth (or minimum tone spacing) $R_c = R_s$. Recall that $R_s$ is the transmitted symbol rate. The total number of available FH slots is defined in equation (104) and the probability that a particular FH band is jammed is given by $Q/N$. It is assumed that there is no inter-symbol interference present; therefore, the bit error performance of the SFH/SDPSK modulated waveform is precisely the same as that of SFH/DPSK. From [55], the corresponding error probability is given as

$$P_b = \begin{cases} 
\frac{1}{2}, & E_b / N_j < 1 \\
1, & E_b / N_j > 1 \\
2(E_b / N_j), & \frac{E_b}{N_j} < 1 \end{cases} \quad (107)$$
Simulated results must lie within the theoretical error bounds defined by equation (101). As with PBNJ impairment, in order to practically simulate the BMTJ impairment a parameter $\mu$ is introduced to represent the probability that the signal in a given M-ary band has been successfully jammed.

Preliminary simulations for uncoded SFH/SDPSK under BMTJ were conducted according to Table 4.3. The corresponding results are shown in Figure 4.8. Note that the empirically generated results do not exceed the upper bound defined by equation (101) validating the simulated SFH/SDPSK model.

It is evident from Figure 4.9, that when $E_b/N_j$ is smaller than the bound established by the worst case performance BER, the bit error probability is constant; however, once $E_b/N_j$ exceeds the worst case threshold value, the bit error probability abruptly tends to zero. Moreover, large values of $\mu$ show the worst performances in the region of small $E_b/N_j$, and smaller values of $\mu$ generate the worst performances under moderate to large values of $E_b/N_j$.

Table 4.3: Simulation parameters for SFH/SDPSK under BMTJ.

<table>
<thead>
<tr>
<th>Date rate</th>
<th>SFH</th>
<th>Hopping rate</th>
<th>$\mu$</th>
</tr>
</thead>
<tbody>
<tr>
<td>256 kbps</td>
<td>16 symbols/hop</td>
<td>16 khops/sec</td>
<td>0.001, 0.01, 0.1, 0.5, 1</td>
</tr>
</tbody>
</table>
Figure 4.9: BER performances of uncoded SFH/SDPSK under BMTJ.

Figure 4.10 shows the simulated BER performance of uncoded SFH/SDPSK under worst case BMTJ. Note that if a sufficiently large number of symbols could be simulated for \( \mu \), the simulated curve will match with the theoretical one. Comparing the curves in Figure 4.10, the simulated performance of SFH/SDPSK does not exceed the theoretical error bounds. Moreover, according to Figure 4.10 and Figure 4.3, the performance of SFH/SDPSK under WBMTJ is about 5 dB worse than that under WPBJ at a BER of \( 10^{-3} \).
Finally, the performance of uncoded SFH/GMSK under BMTJ (n=1) for various values of μ and BT products are assessed. Again, no mathematical models are available describing the BER performance of GMSK under jamming conditions. Therefore, an empirical approach is again taken. The simulation parameters are given in Table 4.4.

Table 4.4: Simulation parameters for SFH/GMSK under BMTJ.

<table>
<thead>
<tr>
<th>Date rate</th>
<th>SFH</th>
<th>BT</th>
<th>Hopping rate</th>
<th>μ</th>
</tr>
</thead>
<tbody>
<tr>
<td>256 kbps</td>
<td>16 symbols/hop</td>
<td>0.25, 0.5, 1</td>
<td>16 k hops/sec</td>
<td>0.001, 0.01, 0.1, 0.5, 1</td>
</tr>
</tbody>
</table>

Figure 4.11 - Figure 4.13 show the simulated BER performances of uncoded SFH/GMSK with BT = 0.25, 0.5 and 1, respectively. In all three Figures, it is evident that lower values of jamming probability most severely degrade BERs as the bit energy to...
jammer noise spectral density \((E_b/N_j)\) is increased. On the contrary, when \(E_b/N_j\) is larger, the smaller values of \(\mu\) more severely degrade BER performance due to the increased jamming energy. It is further evident that the error bound for SFH/GMSK is nearly linear based on the values of \(\mu\) assessed.

Figure 4.11: BER performances of uncoded SFH/GMSK under BMTJ (BT=0.25).
Figure 4.12: BER performances of uncoded SFH/GMSK under BMTJ (BT=0.5).

Figure 4.13: BER performances of uncoded SFH/GMSK under BMTJ (BT=1).
Using the same empirical approach described in section 4.2, the worst case effect of BMTJ on SFH/GMSK can be ascertained. The corresponding BERs are shown in Figure 4.14 wherein smaller the BT product values correspond to the worse performances, albeit a very small variation in results between the different BT products. Thus, increasing the BT product cannot significantly improve the performance of SFH/GMSK using 1-bit differential detection in the presence of BMTJ. Moreover, comparing Figure 4.14 and Figure 4.7, the performance of SFH/GMSK under WBMTJ is very close to that under WPBJ for small BT products. However, when BT = 1, the performance of SFH/GMSK under worst case BMTJ is about 1 dB worse than that of under WPBJ.

Figure 4.14: BER performances of uncoded SFH/GMSK under WBMTJ.
Chapter 5

Performances of DRT and OBP Payloads Under Critical Channel Impairment

Leveraging the detailed discussions on waveform design (Chapter 2), architectural design (Chapter 3) and channel modeling (Chapter 4), satellite system level simulations were conducted towards ascertaining the BER performances of SFH/SDPSK and SFH/GMSK, frequency division multiplexed (FDM) waveforms through both DRT and OBP satellite architectures under jamming interference. For all simulation trials, 10,000 data symbols were processed at a data rate of 64 kbps. Processing 10,000 data symbols yields BER resolutions to $10^{-4}$; however, the statistics provide accurate results to only $10^{-3}$. This is sufficient to gain BER performance insight for applications such as secure voice transmissions. The limited number of symbols can sometimes produce oscillations in the BER trends. Increasing the number of symbols becomes computationally prohibitive, and further resolution must be sought through hardware implementation in FPGAs or DSPs.

The FEC coding schemes used herein are convolutional codes of $(n, k, m) = (2, 1, 9)$ and $(3, 1, 9)$ with generator polynomials $[561; 753]_s$ and $[575; 623; 727]_s$, respectively. Application of the specified FEC coding results in coded transmission rates
of 128 kbps and 192 kbps, for coding rates 1/2 and 1/3, respectively. The corresponding chip rates, assuming a hopping rate of 4 symbols/hop, are 32 khops/sec and 48 khops/sec.

Downlink channels impairments are modeled as having strong \((E_b/N_0 = 10 \text{ dB})\) or moderate \((E_b/N_0 = 15 \text{ dB})\) AWGN interference. For SFH/GMSK, bandwidth-time (BT) product values of 0.5 and 1 were applied. Finally, PBNJ and BMTJ jamming strategies were assumed to have successful jamming probabilities \((\mu, \rho)\) of 0.25 and 0.5. The simulation environment parameters are summarized in Table 5.1.

Table 5.1: Simulation Parameters for DRT vs. OBP Under Critical Impairment.

<table>
<thead>
<tr>
<th>Modulation Schemes</th>
<th>GMSK (BT=0.5), GMSK (BT=1), SDPSK</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bit Rate</td>
<td>64 kbps</td>
</tr>
<tr>
<td>No. of Symbols</td>
<td>10,000</td>
</tr>
<tr>
<td>FEC Coding</td>
<td>Convolutional coding rate 1/2, 1/3</td>
</tr>
<tr>
<td>Coded Bit Rate</td>
<td>128 kbps (code rate 1/2)</td>
</tr>
<tr>
<td></td>
<td>192 kbps (code rate 1/3)</td>
</tr>
<tr>
<td>Hopping Rate</td>
<td>4 symbols/hop</td>
</tr>
<tr>
<td>Chip Rate</td>
<td>32 khops/sec (code rate 1/2)</td>
</tr>
<tr>
<td></td>
<td>48 khops/sec (code rate 1/3)</td>
</tr>
<tr>
<td>Uplink Impairment</td>
<td>PBNJ, (\rho = 0.25, 0.5)</td>
</tr>
<tr>
<td></td>
<td>BMTJ, (\mu = 0.25, 0.5)</td>
</tr>
<tr>
<td>(E_b/N_0) Uplink (dB)</td>
<td>0 - 30</td>
</tr>
<tr>
<td>(E_b/N_0) Uplink (dB)</td>
<td>-</td>
</tr>
<tr>
<td>Downlink Impairment</td>
<td>AWGN</td>
</tr>
<tr>
<td>(E_b/N_0) Downlink (dB)</td>
<td>10, 15</td>
</tr>
</tbody>
</table>

Sections 5.1 and 5.2 present detailed discussions on DRT and OBP system performances under uplink PBNJ and downlink AWGN and uplink BMTJ with AWGN and downlink AWGN, respectively.
5.1 System Performance Under Uplink PBNJ and Downlink AWGN

The BER curves resulting from convolutional coded SFH/GMSK modems (BT=0.5) in a DRT payload under uplink PBNJ and downlink AWGN are shown in Figure 5.1. Curves 1 and 3 indicate a nearly flat BER results for GMSK using code rate 1/2 in the presence of strong downlink AWGN ($E_b/N_0 = 10$ dB) with $\rho = 0.25$ and 0.5. Decreasing the code rate to 1/3 results in marginal improvement, as shown in curves 2 ($\rho = 0.25$) and 4 ($\rho = 0.5$). Curves 5 (rate 1/2) and 7 (rate 1/3) show that performance continues to improve (marginally) under moderate downlink AWGN ($E_b/N_0 = 15$ dB); however these BERs are not acceptable for any practical application. Using SFH/GMSK modems with a DRT payload under PBNJ becomes achieves practical BERs only when low rate coding is used and downlink AWGN is moderate ($E_b/N_0 = 15$ dB), as shown in curves 6 ($\rho = 0.25$) and 8 ($\rho = 0.5$). Note that curves 6 and 8 have BERs of $10^{-3}$ at 26.5 dB and 25.5 dB, respectively.

Figure 5.2 shows the BER performances of SFH/GMSK (BT=0.5) modems in an OBP payload under uplink PBNJ and downlink AWGN. The decoupling of uplink and downlink impairments offered by the OBP payload results in improved BER trends for all scenarios tested; however, the improvement in most curves is still insufficient for practical application. It is shown that SFH/GMSK with BT=0.5, convolutional coding rate 1/2 and $\rho=0.25$, 0.5 (curves 1 and 3, respectively) quickly saturates and cannot be reduced below a BER of $10^{-1}$, regardless of the $E_b/N_j$ applied. This indicates that the downlink AWGN is the dominant effect beyond SJR = 10 dB. Decreasing the coding rate to 1/3 allows the system to attain a BER of $10^{-2}$ at $E_b/N_j = 14$ dB for $\rho = 0.25$ and approximately $E_b/N_j = 10$ dB for $\rho = 0.5$ (curves 2 and 4). These results indicate that
using low rate convolutional codes with OBP processing significantly improves the performance for SFH/GMSK under uplink PBNJ and downlink AWGN. Note that the BER performance is further improved when the downlink AWGN intensity is reduced by 5dB, i.e. increasing $E_b/N_0$ to 15 dB, (curves 5-8). This is true for both coding rates of 1/2 and 1/3. Furthermore, as $E_b/N_J$ is increased, the BER performance is more adversely affected for $\rho = 0.25$ than when $\rho = 0.5$. This implies that highly concentrated jamming (i.e. small values of $\rho$) is more degrading to OBP when SJR is low.

Practical BER performances for SFH/GMSK with OBP processing are achieved only when low rate coding is used and downlink AWGN is moderate ($E_b/N_0 = 15$ dB), as shown in curves 6 ($\rho = 0.25$) and 8 ($\rho = 0.5$) in Figure 5.2. Therein, BERs of $10^{-3}$ are achieved for $E_b/N_J = 11.5$ dB (curve 6) and 14 dB (curve 8), resulting in gains of 15.5 dB and 11.5 dB (respectively) over their DRT counterparts in Figure 5.1. Note that the BER trend of curve 6 is slightly lower than curve 8 until $E_b/N_J$ exceeds 15 dB; thereafter, curve 8 produces no errors, while curve 6 is limited to $10^{-4}$. This is due to the reduced probability of successful jamming from $\rho = 0.25$, which in turn results in increased jamming intensity.
Figure 5.1: BER performances of SFH/GMSK modems (BT=0.5) in DRT payload under uplink PBNJ and downlink AWGN.

Figure 5.2: BER performances of SFH/GMSK modems (BT=0.5) in OBP payload under uplink PBNJ and downlink AWGN.
Figure 5.3 shows the BER performances of SFH/GMSK under uplink PBNJ and downlink AWGN when BT = 1 for DRT and OBP payloads, respectively. Note that curves 6 (p = 0.25) and 8 (p = 0.25) attain BERs of $10^{-3}$ at $E_b/N_j = 23$ dB and 18 dB, respectively, resulting in gains of 3.5 dB and 7.5 dB when compared to corresponding curves wherein BT = 0.5 (i.e. Figure 5.1). Hence, relaxing the BT constraint improves GMSK performance substantially in DRT systems.

According to Figure 5.4, SFH/GMSK with BT = 1 and OBP processing shows better performance than that of BT=0.5 in general. For example, SFH/GMSK with code rate 1/3 under uplink PBNJ (p = 0.5) and downlink AWGN ($E_b/N_0 = 15$ dB) attains a BER of $10^{-3}$ at nearly 10 dB (curve 8), an approximate 2 dB gain over the corresponding waveform in Figure 5.1, wherein the BT = 0.5. Similarly, curve 6 shows a BER of $10^{-3}$ at approximately 12 dB when p is reduced to 0.25; however, for most curves (1-5, 7), the influence of the considered BT products between corresponding trials is marginal, indicating that bandwidth can be saved with minimal BER performance cost. Hence, for OBP systems the effect of relaxing the BT product to unity is, in general, not as pronounced as it is for DRT systems. The impact of the BT product value is secondary to the processing gain which decouples uplink and downlink channel noises.
Figure 5.3: BER performances of SFH/GMSK modems (BT=1) in DRT payload under uplink PBNJ and downlink AWGN.

Figure 5.4: BER performances of SFH/GMSK modems (BT=1) in OBP payload under uplink PBNJ and downlink AWGN.
Figure 5.5 shows the BER trends produced for SDPSK modems in DRT payloads under uplink PBNJ and downlink AWGN. These trends group into one of three distinct regions. Curve 1 represents the theoretical bound for worst-case BER performance under PBNJ. For code rate 1/2 and strong downlink AWGN, curves 2 (\( \rho = 0.25 \)) and 4 (\( \rho = 0.5 \)) show similar performances and have BERs of \( 10^{-3} \) when \( E_b/N_J \) is approximately 12 dB and 14 dB, respectively. Curves 3, 5, 6, and 8 have BERs of \( 10^{-3} \) for \( E_b/N_J \) values of 10.5 dB, 11 dB, 11 dB, and 10 dB. Hence, similar power efficiencies are observed for SDPSK with low code rates, moderate uplink jamming intensity (\( \rho = 0.5 \)) and strong downlink AWGN (\( E_b/N_0 = 10 \) dB), and SDPSK with higher code rates, strong uplink jamming intensity (\( \rho = 0.25 \)) and moderate downlink AWGN (\( E_b/N_0 = 15 \) dB). The best performances are seen when downlink AWGN is reduced to 15 dB and code rate 1/3 is applied, as in curves 7 and 9, wherein a \( 10^{-3} \) BER is obtained for \( E_b/N_J = 8 \) dB and 7 dB, respectively.

Simulated performances of SFH/SDPSK modems with an OBP architecture are shown in Figure 5.6. The additional processing gain offered by OBP (which decouples up and down link channel interferences) results in trends which gravitate towards one of two distinct clusters and are distinguished based solely on coding rate. Curves 4, 6 and 8 all have code rate 1/2 and reach the \( 10^{-3} \) BER threshold for \( E_b/N_J \) of 7 - 7.5 dB. Curves 3, 5, 7, and 9 use code rate 1/3; for curves 3 and 9, a BER of \( 10^{-3} \) is observed for \( E_b/N_J = 5 \) dB and 6 dB, respectively. It is difficult to ascertain precisely where curves 5 and 7 produce error rate of \( 10^{-3} \) because no errors were resulted in the simulation for these curves beyond \( E_b/N_J = 4 \) dB. Thus, OBP processed SFH/SDPSK with convolutional code rate 1/2 requires an additional 2 - 3 dB of \( E_b/N_J \) to attain a BER of \( 10^{-3} \) than if code rate 1/3
was used. Use of low rate convolutional codes yields notable improvement for OBP processing under uplink PBNJ and downlink AWGN; however, the improvement is not as pronounced for SFH/SDPSK as it is for SFH/GMSK.

![Figure 5.5: BER performances of SFH/SDPSK modems in DRT payload under uplink PBNJ and downlink AWGN.](image-url)
5.2 System Performance Under Uplink BMTJ with AWGN and Downlink AWGN

Figure 5.7 illustrates the BER performances of SFH/GMSK (BT=0.5) under uplink BMTJ with AWGN and downlink AWGN in a DRT payload. Curves 1-5 and 7 of Figure 5.7 show that under strong uplink AWGN, communications are completely impaired. When using code rate 1/3, as in curves 6 (μ = 0.25) and 8 (μ = 0.5) performance is limited to less than $10^{-2}$. This indicates that the presence of moderate uplink AWGN is the dominant degrading factor and no practical performance can be realized for SFH/GMSK (BT=0.5) when used with DRT architecture.

Figure 5.8 shows the corresponding BERs resulting when SFH/GMSK (BT=0.5) in an OBP payload is subjected to the same impairments. The best performance is attained when $\mu = 0.5$ and the influence of the jammer power $E_b/N_J$ is diminished. It is
likely that the highest jamming power concentration (resulting from the lower values of \( \mu \)) corresponds to the worst performance when considering \( E_b/N_J > 15 \) dB. While use of OBP processing produces notable improvement (over DRT systems) for all configurations, only curves 6 (\( \mu = 0.25 \)) and 8 (\( \mu = 0.5 \)) can be considered for practical application, underscoring the need for powerful coding or improved receiver design to maintain GMSK signal integrity. BERs of \( 10^{-3} \) are seen in curves 6 and 8 for \( E_b/N_J \) just over 16 dB and 15 dB, respectively.

Figure 5.7: BER performances of SFH/GMSK (BT=0.5) modems in DRT payload under uplink BMTJ with AWGN and downlink AWGN.
Figure 5.8: BER performances of SFH/GMSK (BT=0.5) modems in OBP payload under uplink BMTJ with AWGN and downlink AWGN.

Figure 5.9 and Figure 5.10 show the improved BER performances of SFH/GMSK under uplink BMTJ with AWGN and downlink AWGN resulting from BT = 1 for DRT and OBP payloads, respectively. Note that curves 6 (ρ = 0.25) and 8 (ρ = 0.25) of Figure 5.9 are limited to BERs near $10^{-2}$, whereas the corresponding curves of Figure 5.10 require just under 14 dB to attain a $10^{-3}$ BER. Hence, substantial improvement is obtained for curves 6 and 8 by using OBP instead of DRT; when compared to SFH/GMSK BT = 0.5 with OBP (Figure 5.8), an approximate 2 dB gain results by relaxing BT to unity.
Figure 5.9: BER performances of SFH/GMSK (BT=1) modems in DRT payload under uplink BMTJ with AWGN and downlink AWGN.

Figure 5.10: BER performances of SFH/GMSK (BT=1) modems in OBP payload under uplink BMTJ with AWGN and downlink AWGN.
Figure 5.11 shows the BER trends for SDPSK modems in DRT payloads under uplink BMTJ and downlink AWGN. Curve 1 represents the theoretical bound for worst-case BER performance under BMTJ. The lowest BERs are obtained in the presence of moderate uplink AWGN ($E_b/N_0 = 15$ dB) and BMTJ by curves 7 (rate 1/3, $\mu = 0.25$) and 9 (rate 1/3 and $\mu = 0.5$). These curves require SJR values of approximately 7 dB and 7.5 dB for a $10^{-3}$ BER, respectively.

Figure 5.12 indicates that when OBP processing payloads using coded (rate 1/2) SFH/SDPSK modems are subjected to the specified channel interferences, BER trends are very similar, regardless of the value of $\mu$, or uplink AWGN intensity. Hence, for curves 2, 4, 6, and 8, an $E_b/N_J$ of approximately 10 dB is required to for a $10^{-3}$ BER. Curve 8 shows the lowest BER trend of the cluster due to the decreased uplink jamming intensity ($\mu = 0.5$) and reduced downlink AWGN intensity ($E_b/N_0 = 15$ dB). Decreasing the code rate to 1/3 results in a 2 - 3 dB gain for curves 3, 5, 7 and 9.
Figure 5.11: BER performances of SF/SDPSK modems in DRT payload under uplink BMTJ with AWGN and downlink AWGN.

Figure 5.12: BER performances of SF/SDPSK modems in OBP payload under uplink BMTJ with AWGN and downlink AWGN.
5.3 Discussion and Summary of Results

Herein, a comparative assessment of the performances of DRT and OBP payloads using convolutional coded SFH/GMSK and SFH/SDPSK waveforms under jamming has been presented. The simulated results provide quantifiable measures of the end-to-end BER for the described system and waveforms of interest and serve as a preliminary reference for system designers.

Results show that OBP architectures with SFH/GMSK (BT = 0.5) waveforms and code rate 1/3 under uplink PBNJ can realize power efficiency gains ranging from 11.5 dB - 15.5 dB when compared to corresponding DRT systems. Increasing the BT product to unity can provide gains of 3.5 dB - 7.5 dB for DRT architectures (over BT = 0.5). While the increased BT product also results in improved performances for OBP architectures, the gains are not as substantial as those seen in DRT systems.

OBP processed SFH/SDPSK exhibits remarkable immunity towards PBNJ and BMTJ with AWGN channel impairments and can be further improved by using low code rate convolutional codes. OBP processing of SFH/SDPSK offers gains ranging from 2 - 6.5 dB over DRT systems, depending on jamming intensity and coding rate used; while the performance of SFH/SDPSK is superior to that of SFH/GMSK (due to reduced ISI influence and use of optimum receiver design), it can only be considered for practical application if sufficient bandwidth and supporting terminal resources (i.e. power, physical size, cost, etc.) are available.

SFH/GMSK is more susceptible than SFH/SDPSK under the same impairment scenarios, irrespective of the BT product values. Under uplink BMTJ and AWGN, DRT processed SFH/GMSK cannot realize sufficient performance to be considered for
practical application, regardless of BT product value or satellite architecture used. Although the ability to control spectral sideband content allows support for a greater number of users in the available operating bandwidth, maintaining practical BER performances (particularly in DRT architectures) requires more powerful concatenated coding or iterative decoding schemes, or an enhanced receiver structure for SFH/GMSK to be considered a viable option for tactical communications.
Chapter 6

Analysis of Differential GMSK Under AWGN

Based on the results obtained in Chapter 5, the practicality of using convolutional coded SFH/GMSK in satellite-based tactical mobile communications is limited to OBP satellite payload architectures coupled with low coding rates and weak channel impairments. In order to improve practical BER performance and power efficiency of SFH/GMSK under severe channel impairment, an improvement must be made to the coding scheme or the receiver design. Using an improved maximum likelihood receiver structure would incur increased complexity (resulting from use of local oscillators and/or matched filters [33]), physical size and cost, and cannot be considered for application in dismounted mobile tactical applications.

The alternative is to improve the channel coding scheme which can be supported signal processing hardware (i.e. ASIC, DSP or FPGA) and will incur far less complexity, weight, and power consumption on the mobile terminal device. Turbo coded GMSK has been proposed for use in tactical communication systems, such as MUOS [19] and prior studies have proposed the use of various demodulator structures for GMSK waveforms, including max-Log-MAP demodulators for turbo coded GMSK [56], and iterative based GMSK demodulation and decoding [57], [58] with good results, particularly for low BT
product values; however, the much of the improvement in performance is due use of more complex receiver designs than that considered herein.

The remaining chapters of this dissertation are devoted to investigating use of turbo coded differential GMSK (using the simplified the differential receiver structure described previously [36]), which allows for reduced hardware implementation complexity and marginal signal processing delay and helps to offset signal processing delay and computational overhead incurred by the turbo decoder. The performance of turbo coded differential GMSK will be investigated first for AWGN and PBNJ channel interferences preliminarily.

In order to develop a turbo decoding scheme for differential GMSK, conditional likelihood functions describing the probability of a misresolved symbol given a specific demodulator output value must be developed so that accurate channel reliability estimates can be determined. Previous works have given preliminary insight into closed form analysis of the simplified differential GMSK receiver of interest under AWGN [36], [59]; however, the results did not provide a means for assessing the error probabilities through closed form or numerical methods. Consequently, the remainder of this chapter is devoted towards developing an analytical expression for the performance of 1-bit differential GMSK under AWGN.

6.1 Noise Characterization of Differential GMSK Under AWGN

Consider the input signal of a differential modulating detector in complex form

\[ s_i(t) = R\cos\phi(t) + jR\sin\phi(t) \]  \hspace{1cm} (108)
wherein \( R \) is the signal envelope and \( \phi(t) \) is the phase encoded message. Under noise, the input signal in complex form is

\[
o(t) = s_i(t) + n(t)
\]

(109)

where \( n(t) \) represents an independent Gaussian noise that can be rewritten as

\[
n(t) = n_c(t) + jn_s(t)
\]

(110)

Note that \( n_c \) and \( n_s \) are quadrature noise components of white, Gaussian, zero mean, complex wide sense stationary (WSS) noise with standard deviation \( \sigma_n \). The input SNR is defined as

\[
SNR = 10 \log_{10} \left( \frac{P}{\sigma_n^2} \right)
\]

(111)

where \( P \) is the modulated signal power and \( \sigma_n \) is the noise standard deviation. As indicated in Figure 2.10, the complex signal output of the 1-bit differential demodulator can be written as

\[
\gamma(t) = \text{Re}[j o(t) \phi^*(t-T)]
\]

(112)

\[
= \text{Re}[j(R \cos \phi(t) + jR \sin \phi(t) + n_c(t) + jn_s(t))(R \cos \phi(t-T) - jR \sin \phi(t-T) + n_c(t-T) + jn_s(t-T))] \\
= R^2[\cos \phi(t) \sin \phi(t-T) - \sin \phi(t) \cos \phi(t-T)] + R \cos \phi(t)n_c(t-T) - R \sin \phi(t)n_s(t-T) + R \sin \phi(t-T)n_c(t) - R \cos \phi(t-T)n_s(t) + n_c(t) + n_s(t-T) - n_c(t-T)n_s(t)
\]

where \( o(t) \) is as defined in (109), \( T \) is the symbol duration and * denotes complex multiplication. Having simplified (112), the output signal component of interest is given by

\[
S(t) = R^2[\cos \phi(t) \sin \phi(t-T) - \sin \phi(t) \cos \phi(t-T)]
\]

(113)

\[
= R^2[\sin(\phi(t-T) - \phi(t))] \\
= R^2[\sin(\Delta \phi(t))]
\]
wherein $\Delta \phi(t)$ corresponds to the difference in phase between two successive symbols.

Note that the output noisy component of the signal in (112) is

$$z(t) = R \sin \phi(t) x_1 + R \sin \phi(t-T) x_2 + R \cos \phi(t) y_1 + R \cos \phi(t-T) y_2 + x_2 y_1 - x_1 y_2$$  \hspace{1cm} (114)

where

$$x_1 = -n_s(t-T), \quad y_1 = n_s(t-T)$$  \hspace{1cm} (115)

$$x_2 = n_s(t), \quad y_2 = -n_s(t)$$

Recalling our assumption of the noise disturbance $n(t)$, $x_1, x_2, y_1, y_2$ are independent and Gaussian. Consequently, their linear combination is also Gaussian. The expectation of the noisy component of the signal is

$$E[R \sin \phi(t) x_1 + R \sin \phi(t-T) x_2 + R \cos \phi(t) y_1 + R \cos \phi(t-T) y_2 + x_2 y_1 + x_1 y_2]$$

$$= R \sin \phi(t) E[x_1] + R \sin \phi(t-T) E[x_2] + R \cos \phi(t) E[y_1] + R \cos \phi(t-T) E[y_2] + E[x_2 y_1] + E[x_1 y_2]$$  \hspace{1cm} (116)

The corresponding variance of (116) is then

$$R^2 \sin^2 \phi(t) E[x_1]^2 + R^2 \sin^2 \phi(t-T) E[x_2]^2 + R^2 \cos^2 \phi(t) E[y_1]^2 + R^2 \cos^2 \phi(t-T) E[y_2]^2 + E[x_2 y_1] + E[x_1 y_2]$$  \hspace{1cm} (117)

Taking $E[x_1]^2 = E[x_2]^2 = E[y_1]^2 = E[y_2]^2 = \sigma_n^2$, the $\sin$ and $\cos$ terms of equation (117) can be rewritten as

$$R^2 \left[ \cos^2 \phi(t) + \sin^2 \phi(t) \right] \sigma_n^2 + R^2 \left[ \cos^2 \phi(t-T) + \sin^2 \phi(t-T) \right] \sigma_n^2$$

$$= R^2 \sigma_n^2 + R^2 \sigma_n^2 = 2R^2 \sigma_n^2$$  \hspace{1cm} (118)
Note that \( x_2y_2 \) and \( x_2y_1 \) are not Gaussian; however, if \( E[x_1y_2]^2 + E[x_2y_1]^2 \) is small relative to the square of the expectation of \( x \), \( E[x]^2 \), where

\[
x = R \sin \varphi(t)x_1 + R \cos \varphi(t)y_1 + R \sin \varphi(t-T)x_2 + R \cos \varphi(t-T)y_2
\]

(119) then, it can be shown using the central limit theorem for dependent variables [60] that the noise from (114)

\[
z = x + x_2y_1 - x_1y_2
\]

(120) is nearly Gaussian with

\[
E(z) = E[x] + E[x_2y_1] - E[x_1y_2]
\]

(121) Leveraging independence of the variables

\[
E(z) = 0 + E[x_2]E[y_1] - E[x_1]E[y_2] = 0
\]

(122) Furthermore,

\[
E(z)^2 = E((x + x_1y_2 - x_2y_1)^2) = E[(x + x_1y_2 - x_2y_1)(x + x_1y_2 - x_2y_1)]
\]

(123)

\[
= E[x^2 + 2x_1y_2 - 2xx_1y_2 + x_1^2y_2^2 - xx_1y_2y_2 - xx_1y_2y_2 + x_1^2y_2^2]
\]

\[
= E[x^2 + 2E[xx_1y_2] - 2E[xx_1y_2] + E[x_1^2y_2^2] + E[x_1^2y_2^2]]
\]

Recalling the definition of \( x \) in (119) and expanding the middle terms of the simplified equation in (123) we see that

\[
xx_1y_2 = R \sin \varphi(t)x_1^2y_2 + R \cos \varphi(t)x_1y_1y_2 + R \sin \varphi(t-T)x_1x_2y_2 + R \cos \varphi(t-T)x_1y_2^2
\]

(124) Again, leveraging independence of the variables

\[
E[xx_1y_2] = R \sin \varphi(t)E[x_1^2]E[y_2^2] + R \cos \varphi(t)E[x_1]E[y_1]E[y_2] + R \sin \varphi(t-T)E[x_1]E[x_2]E[y_2] + R \cos \varphi(t-T)E[x_1]E[y_2^2]
\]

(125)
Since \(E[x_1] = E[x_2] = E[y_1] = E[y_2] = 0\),

\[
E[x_{1}y_{2}] = 0
\]  \hspace{1cm} (126)

and, similarly,

\[
E[x_{2}y_{1}] = 0
\]  \hspace{1cm} (127)

\[
E[x_1x_2y_1y_2] = 0
\]  \hspace{1cm} (128)

Therefore, recalling (118), (123) can be reduced to

\[
E[z]^2 = Var[z] = E[x^2] + E[x_1]^2 E[y_2]^2 + E[x_2]^2 E[y_1]^2
\]  \hspace{1cm} (129)

\[
= 2R^2 \sigma^2_n + \left(\sigma^2_n \right) \left(\sigma^2_n \right) + \left(\sigma^2_n \right) \left(\sigma^2_n \right) = 2\sigma^2_n \left(R^2 + \sigma^2_n \right)
\]

Hence, \(z\) (defined in (120)) is a Gaussian variable with zero mean and standard deviation

\[
\Delta = \sqrt{2} \sigma_n \sqrt{R^2 + \sigma^2_n}
\]  \hspace{1cm} (130)

Leveraging the uniqueness theorem of the Fourier transform, and representing \(z\) as a Taylor series expansion,

\[
E(\exp(itZt)) = 1 + \sum_{n=1}^{\infty} E[Z^n] \left(it\right)^n = 1 + \sum_{n=1}^{\infty} E[x + (x_1y_2 - x_2y_1)]^n \frac{it^n}{n!}
\]  \hspace{1cm} (131)

When \(n=0\), \((it)^n = 1\), and \(E(1)=1\). Using the independence properties of \(x_1, x_2, y_1, y_2\), it can be shown that [61]

\[
E[x + (x_1y_2 + x_2y_1)]^n \approx \left(\frac{\sqrt{R^2 + \sigma^2_n}}{\rho \sigma_n} \right)^n E[x]^n
\]  \hspace{1cm} (132)

Therefore, \(z\) is nearly Gaussian [61].

To verify this claim, a Matlab script was written to compare the density function of an actual Gaussian variable to that of the empirically generated density function defined by equation (114). The densities were generated for various standard deviations;
the empirical and theoretical densities for $\sigma = 0.1$ (arbitrarily chosen) are shown in Figure 6.1 and substantiate the above claim that the noise, $z$, is nearly Gaussian.

![Density plots: Empirical (blue) vs. Theoretical (red=zero mean Gaussian)](image)

Figure 6.1: Comparison of empirical (blue) and theoretical (red) Gaussian densities for $\sigma=0.1$.

### 6.2 Derivation of Error Probabilities for Differential GMSK

Given the analysis in the previous section, the detected signal just prior to thresholding can be modeled as

$$S(t) = R^2 \sin[\varphi(t) - \varphi(t - T)] + z$$  \hspace{1cm} (133)

where

$$z \sim N(0, \Delta)$$  \hspace{1cm} (134)

$$\Delta = \sqrt{2} \sigma_n \sqrt{R^2 + \sigma_n^2}$$  \hspace{1cm} (135)

$$\varphi(t) - \varphi(t - T) = \Delta \varphi = \sum_{i=-\frac{L}{2}}^{\frac{L}{2}} M_i \frac{T}{2 - iT} g(\tau) d\tau$$  \hspace{1cm} (136)
From Section 2.2.2, recall that $M_i$ is the NRZ formatted binary data sequence (prior to GMSK modulation), the modulating index, $h = 0.5$, and the shaped pulse train $g(t)$ in (136) is generated by creating an NRZ square wave and applying a Gaussian shaping response, $h(t)$, which carries ISI over $L$ symbol durations, as

$$g(t) = h(t) * rect\left(\frac{t}{T}\right)$$

(137)

We approximate integration of the shaping function, of (137) as

$$\Gamma(i) = \int_{\frac{T}{2} - iT}^{\frac{T}{2} + iT} g(\tau) d\tau = \frac{1}{\sqrt{2\pi\sigma_G T^2}} \int_{\frac{T}{2} - iT}^{\frac{T}{2} + iT} \int_{\frac{T}{2} - iT}^{\frac{T}{2} + iT} \exp\left(- \frac{(t - \tau)^2}{2\sigma_G^2 T^2}\right) dt d\tau$$

(138)

where

$$\sigma_G = \sqrt{\ln 2} \frac{2\pi B}{T}$$

(139)

Rearranging (141) as

$$\Gamma(i) = \frac{1}{\sqrt{2\pi\sigma_G T}} \int_{\frac{T}{2} - iT}^{\frac{T}{2} + iT} \frac{1}{T} \int_{\frac{T}{2} - iT}^{\frac{T}{2} + iT} \exp\left(- \frac{(t - \tau)^2}{2\sigma_G^2 T^2}\right) dt d\tau$$

(140)

and considering that

$$\lim_{q \to 0} \int_{\frac{q}{2}}^{\frac{q}{2}} g(t) dt = g(t)$$

(141)

if the symbol duration, $T$, is small, then (140) can be simplified as

$$\Gamma(i) = \frac{1}{\sqrt{2\pi\sigma_G T}} \int_{\frac{T}{2} - iT}^{\frac{T}{2} + iT} \exp\left(- \frac{(t)^2}{2\sigma_G^2 T^2}\right) dt \approx \frac{1}{\sqrt{2\pi\sigma_G}} \exp\left(- \frac{i^2}{2\sigma_G^2}\right)$$

(142)

For a received bit sequence that carries ISI over $L$ symbol durations, let
\[ \Delta \varphi = \sum_{i=-L/2}^{L/2} M_i \, d(i) = \sum_{i=-L/2}^{L/2} M_i d(i) \]  
\hspace{1cm} (143) 

where

\[ d(i) = \frac{\pi h}{\sqrt{2 \pi \sigma_G}} \exp \left( -\frac{i^2}{2\sigma_G^2} \right) \]  
\hspace{1cm} (144) 

\( i \) is the summation index (143), \( h = 0.5 \), and \( \sigma_G \) is as defined in (139). The demodulated differential GMSK signal is expressed as

\[ S = R^2 \sin(\Delta \varphi) + z \]  
\hspace{1cm} (145) 

Recall that the binary data symbols are resolved by taking the \( \text{sign}(S) \). Hence, if \( S \geq 0 \) the \( M^\text{th} \) data symbol corresponds to a ‘1’ data symbol and \( S < 0 \) corresponds to a ‘0’. Since GMSK is a binary signaling scheme, misresolved symbol errors can occur as

- \( M_0 = -1 \), when \( S \geq 0 \) or
- \( M_0 = +1 \), when \( S < 0 \)

Assuming that the probability of error occurs equally for both bit values, i.e.

\[ P(M_0 = 1) = P(M_0 = -1) \]  
\hspace{1cm} (146) 

then the probability of a bit being in error is

\[ P_b = \frac{1}{2} \left[ P(S > 0|M_0 = -1) + P(S < 0|M_0 = 1) \right] \]  
\hspace{1cm} (147) 

The average bit error probability for a transmitted data sequence of length \( L \), is

\[ \frac{1}{2^{L-1}} \sum_{l=1}^{2^{L-1}} P(S > 0|M_0 = -1), \text{for } V = \{M_{L/2}, M_{L/2+1}, \ldots, M_{L/2-1}, M_{L/2+1}\} = t(l) \]  
\hspace{1cm} (148)
where \( t(l) \) is one of the possible bit sequence combinations in the space \( V = \{M_{L/2}, M_{L/2 + 1}, \ldots, M_{L/2 - 1}, M_{L/2 + 1}\} \). The error probability of a given data symbol \( M_0 \) can then be found by substituting (142) into (148) and using the complementary error function integration as in (149)

\[
P(S > 0 | M_0 = -1) = \frac{1}{2^{L-1}} \sum_{l=1}^{2^{L-1}} \frac{1}{\sqrt{2\pi\Delta}} \int_{-\infty}^{\infty} \exp\left(-\frac{\theta^2}{2\Delta^2}\right) d\theta
\]

wherein \( \theta \) is the variable of integration (recall \( i \) in (142)). The conditional likelihood function that \( S = x \) given \( M_0 = -1 \) is

\[
f(S = x | M_0 = -1) = \frac{1}{2^{L-1}} \sum_{l=1}^{2^{L-1}} \frac{\exp\left(-\frac{(x - R^2 \sin(\varphi(l)))^2}{2\Delta^2}\right)}{\sqrt{2\pi\Delta}}
\]

where \( \Delta \) is as defined in (130) and \( \Delta \varphi(t) \) is as defined in (143). Similarly, for the case of a misresolved binary ‘1’

\[
P(S < 0 | M_L = 1) = \frac{1}{2^{L-1}} \sum_{l=1}^{2^{L-1}} \frac{1}{\sqrt{2\pi\Delta}} \int_{-\infty}^{\infty} \exp\left(-\frac{\theta^2}{2\Delta^2}\right) d\theta
\]

for \( V = \{M_{L/2}, M_{L/2 + 1}, \ldots, M_{L/2 - 1}, M_{L/2 + 1}\} = t(l) \) and the conditional likelihood function is given by

\[
f(S = x | M_L = 1) = \frac{1}{2^{L-1}} \sum_{l=1}^{2^{L-1}} \frac{\exp\left(-\frac{(x - R^2 \sin(\varphi(l)))^2}{2\Delta^2}\right)}{\sqrt{2\pi\Delta}}
\]

Thus, the theoretical bit error probability for one-bit differential GMSK is given by
This expression cannot be evaluated in closed form and must be implemented using numerical approximation.

### 6.3 Validation of Derived Error Probability for Differential GMSK

In order to validate the expression modeling bit error probability for 1-bit differential GMSK under AWGN, the simulated GMSK performance was compared against its theoretical counterpart for several BT products. The simulation parameters are outlined in Table 6.1. Figure 6.3 - Figure 6.4 show the comparative simulated and theoretical trends for GMSK under AWGN with BT = 0.4, 0.5 and 1, respectively.

<table>
<thead>
<tr>
<th>BT products</th>
<th>Number of Symbols</th>
<th>Data Rate (kbps)</th>
<th>E_b/N_0 (dB)</th>
<th>Channel Impairment</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.4, 0.5, 1</td>
<td>10,000</td>
<td>10</td>
<td>0 - 20</td>
<td>AWGN</td>
</tr>
</tbody>
</table>
Figure 6.2: Theoretical vs. simulated differential GMSK under AWGN with BT=0.4.

Figure 6.3: Theoretical vs. simulated differential GMSK under AWGN with BT=0.5.
Figure 6.4: Theoretical vs. simulated differential GMSK under AWGN with BT=1.

It is evident from the figures that the simulated results and theoretical bounds correspond well for the considered BT products over the considered SNR range. Note that as the BT product is decreased from 0.5 to 0.4, a marginal deviation of less than 1 dB between the theoretical bound and simulated BER performance is shown in Figure 6.2 at high SNR. This deviation results from the Gaussian approximation of the noise distribution in the above mathematical analysis and is clearly more pronounced for smaller BT product values. The results obtained for the BT products of interest sufficiently demonstrate the validity of (153) in the above mathematical analysis and the conditional likelihood probability functions given in (150) and (152) will enable development of a turbo decoding scheme that can support differential GMSK.
Chapter 7

Turbo-coded Differential GMSK

This chapter is devoted to the development and performance assessment of turbo coded 1-bit differential GMSK under AWGN and PBNJ interferences. The performances of convolutional coded differential GMSK (in Chapter 5) were shown to be insufficient for supporting tactical mobile communications with respect to power efficiency. Improved performances could be realized by using convolutional code rates lower than rate 1/3; however, this approach would work contrary to our objective of improving overall bandwidth efficiency.

Turbo coding supports superior error correcting capabilities by using random-like codes and a structured, efficient iterative decoding process that, using soft demodulated output values, achieves near Shannon limit performances [37]. Section 7.1 provides a brief treatment on turbo encoding, section 7.2 presents turbo decoding of differential GMSK under AWGN and finally, section 7.3 presents the BER performances of turbo coded GMSK under AWGN. A comparative assessment between turbo coded GMSK and convolutional coded GMSK for code rates 1/3 and 2/3 is also included. Section 7.4 demonstrates the performance of the designated scheme under PBNJ channel interference.
7.1 Turbo Encoding

Turbo encoding leverages the parallel concatenation of two systematic recursive convolutional codes (RSCCs) and a pseudorandom interleaver block as shown in Figure 7.1, wherein $d_k$ is the systematic bit, $P_1$ is the first parity check bit, and $P_2$ is second parity check bit. Hence, the output of this structure consists of one systematic bit and two parity check bits that are serialized. Note that the output $P_2$ is first randomized using a pseudorandom interleaver prior to passing through the second RSCC encoder. In addition, the incoming sequence of bits must be properly terminated to ensure that the encoder ends in an all zero state. Consequently, $m$ tail bits must be added to the bit sequence block length prior to encoding. The specific number of tail bits is determined based on the number of memory cells in the RSCC structures.

![Figure 7.1: Basic turbo encoding structure [37].](image)

The structure of the RSCC encoder Figure 7.1 is shown in Figure 7.2 and its corresponding generator matrix is given by [62]

$$G(D) = \begin{bmatrix} 1 & g_2(D) \\ g_1(D) & \end{bmatrix}$$

(154)
where \( g_1(D) \) represents the generator feedback polynomial and \( g_2(D) \) represents the generator feedforward polynomial, which define the RSCC topology. For this study, the generator polynomials are defined as [63]

\[
g_1(D) = [1 \ 0 \ 0 \ 1 \ 1]
\]

and

\[
g_2(D) = [1 \ 0 \ 0 \ 1 \ 1]
\]

Figure 7.2: G(23,31) RSCC encoder structure.

After the parity sequences \( p_1 \) and \( p_2 \) are generated by the RSCCs, they are punctured. Puncturing deletes selected bits from the encoded bit stream to reduce coding overhead and is required to realize higher code rates. The considered RSCC encoders have a code rate of 1/2 and each produce \( k \) information bits and \( k \) parity bits. Consequently, \( 2k \) parity bits are transmitted through the channel along with \( k \) information bits (as shown in Figure 7.1). The overall system code rate can be found according to

\[
\frac{1}{R} = \frac{1}{R_1} + \frac{1}{R_2} - 1
\]

Hence, the overall system code rate is 1/3.
In order to construct higher rate turbo codes (i.e. code rates greater than rate 1/3), puncturing must be applied. The higher rate codes typically assume rates of \( \frac{k}{k+1} \) where \( k \) is the number of information bits used [64]. The turbo code rates under consideration herein will be limited to rate 1/3 initially, for comparison with our previously used convolutional codes. In addition, rate 2/3 is used to demonstrate the performance of a higher rate code. The turbo codes were generated according to the parameters outlined in [64].

<table>
<thead>
<tr>
<th>Rate</th>
<th>Selected Polynomial</th>
<th>Selected P(p,q)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2/3</td>
<td>(23,31)</td>
<td>P(3,4)</td>
</tr>
</tbody>
</table>

Note that in Table 7.1, the puncturing pattern \( P(p,q) \) indicates that every \( p^{th} \) parity bit (from RSCC1) is retained from RSCC1 and every \( q^{th} \) parity bit from RSCC2 is retained; all other parity bits are deleted. Systematic bits are not punctured.

7.2 Turbo Decoding of Differential GMSK Under AWGN

MAP decoding takes the a priori probabilities of the input symbols and makes a soft output decision of what the received data symbol should be. Turbo decoding is an iterative process wherein the constituent decoders pass this soft-information (resulting from MAP decoding) from one to the other. The soft output of one decoder serves as the a priori information of the other decoder.

Figure 7.3 shows the block diagram of a typical turbo decoder structure which was originally designed to support BPSK modulation, but has been modified to support
GMSK. The structure receives $y_k^s$ (the systematic bit) and $y_k^{p1,p2}$ (the parity bits) in sequence from the GMSK demodulator. The iterative turbo decoder uses the Log-MAP decoding algorithm (so that computations consist of additions instead of multiplications) and it is assumed that an information bit assumes values of 1 and 0 with equal probability. The a priori information, $A_{1a}(u_k)$, of DEC1 is initialized to zero for all $k$ in the information word $u$, in preparation for the first decoding iteration. The extrinsic information, $A_{1e}(u_k)$, produced by DEC1 is interleaved and then passed as a priori information to DEC2: $A_{2a}(u_k) = A_{1e}(u_{int}(k))$, where $int(k)$ indicates interleaver mapping. Note that in simulations, a random interleaver with a size equal to the number of data bits is used. A single iteration is complete when DEC2 generates its output. The extrinsic information $A_{2e}(u_k)$ from DEC2 is then deinterleaved and fed back as a priori information to DEC1 for the next iteration: $A_{1a}(u_{int}(k)) = A_{2e}(u_k)$. This process is repeated for an arbitrary number of iterations [63].

Figure 7.3: Turbo decoder structure with conventional Log-MAP [63].
The key metrics involved in using Log-Map decoding are the branch metric, $\gamma$, the forward path metric, $\alpha$, the backward path metric, $\beta$ and the extrinsic information, $\Lambda_e$, which are defined as follows [63]:

$$
\gamma_k(e) = \ln P(s^e_k(e)|s_k^e(e)) + \ln P(y_k|X_k) \quad (158)
$$

$$
\alpha_k(s) = \max_{e \in e^k(e)_s} \left[ \alpha_{k-1}(s^e_k(e)) + \gamma_k(e) \right], \quad k = 1, \ldots, N-1 \quad (159)
$$

$$
\beta_k(s) = \max_{e \in e^k(e)_s} \left[ \beta_{k-1}(s^e_k(e)) + \gamma_k(e) \right], \quad k = 1, \ldots, N-1 \quad (160)
$$

$$
\Lambda_e(u_k) = \lambda^A_k(u|O) - \Lambda_a(u_k) - \lambda_k(c^{(i)}|I) \quad (161)
$$

where $u$ is the information word, $c$ is the codeword, $I$ is the soft output from the demodulator, and $O$ is the corresponding output estimate. The complete a posteriori (denoted by superscript A) log-likelihood ratio (LLR) of the information bit, $\lambda^A_k(u|O)$, is

$$
\lambda^A_k(u|O) = \max_{e \in e^k(e)_s} \left[ \alpha_{k-1}(s^e_k(e)) + \gamma_k(e) + \beta_k(s^e_k(e)) \right]
$$

$$
- \max_{e \in e^k(e)_s} \left[ \alpha_{k-1}(s^e_k(e)) + \gamma_k(e) + \beta_k(s^e_k(e)) \right] \quad (162)
$$

where $\max^*(\cdot)$ is defined as

$$
\max^*(z) = \ln \sum_e \exp(z) \quad (163)
$$

For equation (158), $\ln P(s^e_k(e)|s_k^e(e))$ is given as, [63]

$$
\ln P(s^e_k(e)|s_k^e(e)) = \begin{cases} 
\Lambda_a(u_k) - \ln (1 + e^{\Lambda_e(u_k)}), & u_k = 1 \\
- \ln (1 + e^{\Lambda_e(u_k)}), & u_k = 0 
\end{cases} \quad (164)
$$

The value in equation (164) is only based on the information word sequence $u_k$ and $\lambda_k(c^{(j)}|I)$ is given by [63]

$$
\lambda_k(c^{(j)}|I) = \ln \frac{P_k(c^{(j)} = 1|I)}{P_k(c^{(j)} = 0|I)} \quad (165)
$$
For 1-bit differential GMSK, the conditional likelihoods were found in Chapter 6 to be

\[
f(x | M = \pm 1) = \frac{1}{2^{L-1}} \sum_{j=1}^{2^{L-1}} \exp \left( \frac{x + R^2 \sin(\Delta \phi)}{2\Delta} \right) \]

(166)

where \( M \) is the intended transmitted data symbol, \( x \) is the soft demodulator output, \( \sin(\Delta \phi) \) is the idealized received value (bounded by unity), \( L \) is the number of bits influenced by the shaping function, \( \Delta \) is the Gaussian noise standard deviation, and \( R \) is the signal envelope. Substituting (166) into (165) and simplifying, the LLR can be expressed as

\[
\lambda_k (c^{(j)} | I) = \frac{2R^2}{\Delta^2} \sum_{j=1}^{2^{L-1}} x \sin(\Delta \phi)
\]

(167)

The expression in (167) is then substituted in the following to obtain the conditional probability of the received sequence given the intended transmitted sequence [63]

\[
\ln P(Y_k | X_k) = 0.5\lambda_k (c^{(1)} | I)(2c_k^{(1)} - 1) + 0.5\lambda_k (c^{(2)} | I)(2c_k^{(2)} - 1)
\]

(168)

The expressions in (162) and (168) are then used to compute the branch metric \( \gamma \). For further details on the MAP decoding algorithm, extensive treatments on turbo decoding can be found in [37], [62], [63], and [65].

### 7.3 Performance of Turbo Coded Differential GMSK Under AWGN

The performance of turbo-coded GMSK waveform is evaluated through sampling-based computer simulations using Matlab for code rates 1/3 and 2/3. For comparative purposes, simulations were also conducted for convolutional coded GMSK using the same coding rates 1/3 and 2/3. Note that convolutional coding rate 2/3 is obtained by puncturing the rate 1/2 code. Table 7.2 summarizes the simulation parameters for this study. Note that in
Table 7.2, $G_1(D) = [575;623;727]_8$, $G_2(D) = [561;753]_8$, $g_1(D) = [1 0 0 1 1]$ and $g_2(D) = [1 1 0 0 1]$, respectively.

Table 7.2: Simulation Parameters for turbo coded GMSK waveform under AWGN.

<table>
<thead>
<tr>
<th>Modulation</th>
<th>1-bit Differential GMSK</th>
</tr>
</thead>
<tbody>
<tr>
<td>BT</td>
<td>0.5, 1</td>
</tr>
<tr>
<td>Bit Rate (kbps)</td>
<td>10</td>
</tr>
<tr>
<td>No. of Symbols</td>
<td>80,000</td>
</tr>
<tr>
<td>Coding Scheme</td>
<td>Turbo-coding</td>
</tr>
<tr>
<td>Code Rates</td>
<td>1/3, 2/3</td>
</tr>
<tr>
<td>Generator</td>
<td>$[1 g_2(D)/g_1(D)]$</td>
</tr>
<tr>
<td>Coded Bit Rate (kbps)</td>
<td>30, 15</td>
</tr>
<tr>
<td>Channel Impairment</td>
<td>AWGN</td>
</tr>
<tr>
<td>$E_b/N_0$ (dB)</td>
<td>0 - 15</td>
</tr>
</tbody>
</table>

Figure 7.4 shows the BER performance of turbo coded GMSK with coding rate 1/3 and BT product of 0.5 for various decoding iterations as well as the BER trends corresponding to uncoded GMSK and convolutional coded GMSK cases. The simulated uncoded performance matches well with theoretical bound in the region of small $E_b/N_0$ values, however, a slight deviation is perceived for large $E_b/N_0$, which may occur due to the optimum approximation taken for the BER calculation.

The first iteration of turbo decoding shows approximately 2 dB degradation as compared to the convolutional coded case at BER of $10^{-3}$. The second iteration shows comparable 1 dB gain by turbo coding and convolutional coding at $10^{-3}$ BER. Note that no error is produced from $10^0$ to $10^{-5}$ BER in the turbo coded trend when $E_b/N_0$ is increased beyond 10 dB. For the third decoding iteration, turbo coded GMSK offers slight improvement over the second iteration. After the fourth decoding iteration, the next point beyond 7 dB in the BER curve is not visible due to the limited number of symbols processed in simulations. While it is possible to obtain further resolution by increasing
the overall number of symbols processed, the additional time and computational expense is prohibitive and hardware implementation should be pursued for further study. The dashed line delineates the expected BER values after the fourth iteration, which corresponds to an approximate 3 dB improvement in performance over convolutional coding at BER of $10^{-3}$.

Figure 7.5 shows the BER performance of turbo coded GMSK with coding rate 1/3 and BT product of 1. The trends shown therein are similar to those in Figure 7.4. Comparing the turbo coded GMSK with BT products of 0.5 and 1, turbo coded GMSK with BT product of 1 always shows better performance than that of BT product 0.5.

![Figure 7.4: Performance of turbo-coded (rate 1/3) GMSK (BT = 0.5) and convolutional coded (rate 1/3) GMSK (BT = 0.5) under AWGN.](image)
Figure 7.5: Performance of turbo-coded (rate 1/3) GMSK (BT = 1) and convolutional coded (rate 1/3) GMSK (BT = 1) under AWGN.

Figure 7.6 and Figure 7.7 show the BER performance of turbo coded GMSK under AWGN with coding rate 2/3 for BT products of 0.5 and 1, respectively. Coding rate 2/3 is obtained for turbo codes by puncturing the rate 1/3 code. The puncturing pattern used to generate rate 2/3 is shown in Table 7.1, where the puncturing pattern $P(p,q)$ indicates that the $p^{th}$ parity bit of every sequence of $q$ bits is retained and all other parity bits are deleted. The systematic bits are not punctured.

On comparing the performance of the two coding rates for turbo coded GMSK, we see that rate 1/3 performs better than rate 2/3, especially for BT = 0.5. The trend shows that use of low rate FEC coding can yield improvement for turbo coded GMSK under AWGN channel and its use in the modem can be justified. Moreover, the bandwidth efficiency gained by using turbo code rate 2/3 is also substantial and its use is well justified given the marginal degradation in performance that occurs when increasing
the code rate from 1/3 to 2/3. Table 7.3 summarizes the results for turbo coded GMSK under AWGN for the third iteration, compared to uncoded and convolutional coded GMSK at a BER of $10^{-3}$.

Figure 7.6: Performance of turbo-coded (rate 2/3) GMSK (BT = 0.5) and convolutional coded (rate 2/3) GMSK (BT = 0.5) under AWGN.
Figure 7.7: Performance of turbo-coded (rate 2/3) GMSK (BT = 1) and convolutional coded (rate 2/3) GMSK (BT = 1) under AWGN.

Table 7.3: BER performance summary of turbo coded GMSK under AWGN.

<table>
<thead>
<tr>
<th>Code Rate</th>
<th>BT</th>
<th>Uncoded $E_b/N_0$ @ 10^{-3}$ (dB)</th>
<th>Conv. Coded $E_b/N_0$ @ 10^{-3}$ (dB)</th>
<th>Turbo Coded $E_b/N_0$ @ 10^{-3}$ (dB)</th>
<th>Turbo vs. conv. coding gain</th>
</tr>
</thead>
<tbody>
<tr>
<td>1/3</td>
<td>0.5</td>
<td>12</td>
<td>10</td>
<td>8.2*</td>
<td>1.8</td>
</tr>
<tr>
<td></td>
<td>1</td>
<td>11.7</td>
<td>8.5*</td>
<td>7.5*</td>
<td>1</td>
</tr>
<tr>
<td>2/3</td>
<td>0.5</td>
<td>12</td>
<td>11.2*</td>
<td>8.5*</td>
<td>2.7</td>
</tr>
<tr>
<td></td>
<td>1</td>
<td>11.7</td>
<td>9.5*</td>
<td>7*</td>
<td>2.5</td>
</tr>
</tbody>
</table>

*: Estimated value.

The empirical results show that as few as two turbo decoding iterations can provide up to 2 dB gain over convolutional coded counterparts for a BER of $10^{-3}$ with further gains realized for additional decoding iterations. Using turbo code rate 2/3 offers significant bandwidth efficiency, over code rate 1/3, in exchange for marginal BER performance degradation (in the case of BT=0.5). Turbo coding with three decoding iterations, offers a 3.5 - 4.7 dB gain over uncoded GMSK, and a 1.8 - 2.7 dB gain over convolutional coded GMSK is observed (depending on the BT value). These results show
that use of turbo coded differential GMSK in handheld terminals under AWGN is indeed feasible and warrants further investigation into assessing their feasibility for use under targeted jamming conditions.

7.4 Performance of Turbo Coded Differential GMSK Under PBNJ

The mathematical analysis of differential GMSK under AWGN can be used to model its behavior under PBNJ, as it also has a Gaussian characteristic. The bit energy to noise spectral density, $E_b/N_0$, is replaced by the scaled bit energy to jamming noise spectral density, $\rho E_b/N_j$ (to reflect the increased jamming intensity) and the expression for overall bit error probability, equation (157) must be multiplied by $\rho$, where $\rho$ is defined as probability that a successful jam occurs. This section presents empirical BER performances for turbo coded GMSK under PBNJ; the simulation parameters are summarized in Table 7.4. Figure 7.8 - Figure 7.15 show the resulting BER performances.

Table 7.4: Simulation Parameters for turbo coded GMSK waveform under PBNJ.

<table>
<thead>
<tr>
<th>Modulation</th>
<th>GMSK</th>
</tr>
</thead>
<tbody>
<tr>
<td>BT</td>
<td>0.5, 1</td>
</tr>
<tr>
<td>Bit Rate (kbps)</td>
<td>10</td>
</tr>
<tr>
<td>No. of Symbols</td>
<td>20,000</td>
</tr>
<tr>
<td>Coding Scheme</td>
<td>Turbo-coding, Convolutional</td>
</tr>
<tr>
<td>Coding Rates</td>
<td>1/3, 2/3</td>
</tr>
<tr>
<td>Generator</td>
<td>$[1 g_2(D)/g_1(D)]$, $G_1(D)$, $G_2(D)$</td>
</tr>
<tr>
<td>Coded Bit Rate (kbps)</td>
<td>30, 15</td>
</tr>
<tr>
<td>Channel Impairment</td>
<td>PBNJ</td>
</tr>
<tr>
<td>$E_b/N_j$ (dB)</td>
<td>0 - 15</td>
</tr>
</tbody>
</table>
Figure 7.8: Performance of turbo-coded (rate 1/3) GMSK (BT = 0.5) and conv. coded (rate 1/3) GMSK (BT = 0.5) under PBNJ (p=0.25).

Figure 7.9: Performance of turbo-coded (rate 1/3) GMSK (BT = 0.5) and conv. coded (rate 1/3) GMSK (BT = 0.5) under PBNJ (p=0.5).
Figure 7.10: Performance of turbo-coded (rate 2/3) GMSK (BT = 0.5) and conv. coded (rate 2/3) GMSK (BT = 0.5) under PBNJ (ρ=0.25).

Figure 7.11: Performance of turbo-coded (rate 2/3) GMSK (BT = 0.5) and conv. coded (rate 2/3) GMSK (BT = 0.5) under PBNJ (ρ=0.5).
Figure 7.12: Performance of turbo-coded (rate 1/3) GMSK (BT = 1) and convolutional coded (rate 1/3) GMSK (BT = 1) under PBNJ (p=0.25).

Figure 7.13: Performance of turbo-coded (rate 1/3) GMSK (BT = 1) and convolutional coded (rate 1/3) GMSK (BT = 1) under PBNJ (p=0.5).
Figure 7.14: Performance of turbo-coded (rate 2/3) GMSK (BT = 1) and convolutional coded (rate 2/3) GMSK (BT = 1) under PBNJ (p=0.25).

Figure 7.15: Performance of turbo-coded (rate 2/3) GMSK (BT = 1) and conv. coded (rate 2/3) GMSK (BT = 1) under PBNJ (p=0.25).
As in section 7.3, the analysis for turbo coding results is compared to the corresponding convolutional coded waveform and is taken with respect to the third decoding iteration, and a BER threshold of $10^{-3}$. The results, summarized in Table 7.5, show a remarkable improvement in overall transmission power efficiency. The gains resulting from the use of turbo codes instead of convolutional codes range from 1.4 - 3.5 dB. It is interesting to note that for code rate 1/3, the largest gains are seen for $BT=0.5$, regardless of the jamming intensity $\rho$. When considering code rate 2/3, three of the four cases tested show gains of 3.5 dB or greater, indicating that BT product value and the jamming intensity $\rho$ do not much influence the BER performance. As with the results in section 7.3, comparable or slightly degraded performance is observed when using turbo code rate 2/3; however, the considerable gain in bandwidth efficiency completely justifies its use. Hence, these results also show that use of turbo coded differential GMSK in handheld terminals under PBNJ is indeed feasible.

Table 7.5: BER performance summary of turbo coded GMSK under PBNJ.

<table>
<thead>
<tr>
<th>Code Rate</th>
<th>BT</th>
<th>$\rho$</th>
<th>Uncoded $E_b/N_J @ 10^{-3}$ (dB)</th>
<th>Conv. Coded $E_b/N_J @ 10^{-3}$ (dB)</th>
<th>Turbo Coded $E_b/N_J @ 10^{-3}$ (dB)</th>
<th>Turbo vs. conv. coding gain</th>
</tr>
</thead>
<tbody>
<tr>
<td>1/3</td>
<td>0.5</td>
<td>0.25</td>
<td>16.2</td>
<td>7.2*</td>
<td>4</td>
<td>3.2</td>
</tr>
<tr>
<td></td>
<td>0.5</td>
<td>0.5</td>
<td>13.5</td>
<td>7.7</td>
<td>4.5*</td>
<td>3.2</td>
</tr>
<tr>
<td></td>
<td>1</td>
<td>0.25</td>
<td>15.2</td>
<td>6.6</td>
<td>5.2*</td>
<td>1.4</td>
</tr>
<tr>
<td></td>
<td>1</td>
<td>0.55</td>
<td>14</td>
<td>6.7</td>
<td>5.2*</td>
<td>1.5</td>
</tr>
<tr>
<td>2/3</td>
<td>0.5</td>
<td>0.25</td>
<td>16.2</td>
<td>10.5</td>
<td>6.5</td>
<td>4</td>
</tr>
<tr>
<td></td>
<td>0.5</td>
<td>0.5</td>
<td>13.3</td>
<td>7.5</td>
<td>6.1*</td>
<td>1.4</td>
</tr>
<tr>
<td></td>
<td>1</td>
<td>0.25</td>
<td>15.3</td>
<td>9.7</td>
<td>6.2*</td>
<td>3.5</td>
</tr>
<tr>
<td></td>
<td>1</td>
<td>0.5</td>
<td>14</td>
<td>9</td>
<td>5.5*</td>
<td>3.5</td>
</tr>
</tbody>
</table>

*: Estimated value.
Chapter 8

Conclusion

8.1 Summary

In this dissertation, advanced FH-FDMA DRT and OBP system architectures were modeled and simulated towards application in satellite-based tactical mobile communications system experiencing inherent degrading factors and targeted channel interferences. An extensive suite of sampling-based simulations was conducted using both architecture types and two waveform variants: convolutional coded non-coherent 1-bit differentially detected GMSK and convolutional coded SDPSK. Simulations provided a quantifiable measure of the differences in end-to-end BER performances for the architectures and waveforms of interest under the influences of uplink PBNJ, or BMTJ with AWGN, and downlink AWGN.

Although the performance of SFH/SDPSK was found to be superior to that of SFH/GMSK (due to reduced ISI influence and use of optimum receiver design), it can only be considered for practical application if sufficient bandwidth and supporting terminal resources (i.e. power, physical size, cost, etc.) are available. SFH/GMSK is more susceptible than SFH/SDPSK under the same impairment scenarios, regardless of the BT
product values and satellite architecture used; in order to substantially improve BER performance, a more effective channel coding scheme was needed.

Consequently, a detailed mathematical analysis of 1-bit differential GMSK under AWGN was developed wherein an expression for evaluating the theoretical BER performance bound through numerical methods was undertaken. This analysis enabled use of more powerful turbo coding in the waveform design towards making differential GMSK a viable option for satellite based tactical mobile applications. Simulations show that as few as two turbo decoding iterations can provide up to 2 dB gain over convolutional coded differential GMSK for a BER of $10^{-3}$ with further gains realized for additional decoding iterations. Remarkable improvements in power efficiencies were also realized when subjecting turbo coded GMSK to the effects of PBNJ interference, particularly for code rate 2/3. Both empirical investigations into differential GMSK BER performance under AWGN and PBNJ interferences effectively demonstrate that greater bandwidth efficiencies can be realized by using high code rates turbo codes with marginal BER performance degradation. These results strongly support use of turbo coding with differential GMSK under AWGN and PBNJ interferences, and warrant further investigation into their feasibility of use under tone jamming conditions.

### 8.2 Future Work

The work detailed herein has provided a basis for modeling and performance assessment of satellite communication systems wherein the nature of the waveform and system architecture has been taken into account. The results obtained for turbo coded GMSK have shown much promise in operating in mobile tactical environments. Further effort
should be applied towards system level simulations using the turbo coded GMSK waveforms. In addition, a theoretical analysis of the performance of differential GMSK under tone jamming interference should be undertaken so that turbo coding can be effectively applied therein.

The computer-based simulations have yielded a preliminary insight into the BERs achievable by the systems and waveforms under consideration; however, in order to obtain further accuracy and resolution into the system performance, hardware implementation must be sought. Towards this end, the convolutional coded SDPSK waveform model developed and simulated in this dissertation using Matlab have been ported to C and executed in hardware using Atmel’s Diopsis D940, a dual-core heterogenous processing platform shown in Figure 8.1 [66]. The processor consists of an ARM926 RISC microcontroller and the mAgicV very long instruction word (VLIW) 40-bit floating point DSP. This platform was chosen for implementation because its native support for C, and extensive supporting DSP library of function calls. Having multiple processing elements allows assignment of the more computationally intensive tasks (using the hArtes toolchain [67]) to the mAgicV DSP, leaving the ARM processor with the task of managing data flow. This approach dramatically reduces overall execution time.
Figure 8.1: Atmel Diopsis D940HF (ARM+DSP) development board.

As an initial example, the screenshot in Figure 8.2 illustrates the difference in execution times observed when performing modulation/demodulation on the ARM and mAgicV DSP processors for two transmission channels. Ten data symbols were generated for each channel and were then convolutional coded with rate 1/2 (resulting in a coded sequence of 36 bits). The bit sequence was 8x sampled resulting in 288 samples to process. No noise was introduced so that correct operation of the modem could be validated. For the ARM processor, modulation and demodulation requires 21.47 ms and 41.08 ms, respectively. In contrast, the mAgicV DSP requires only 1.204 ms for modulation and 7.73 ms for demodulation. This translates into a nearly 18x improvement in modulation execution time, and over a 5x improvement in demodulation execution time [68]. Note also that the implementation was performed offline, i.e. all data was generated and processed within the D940 single board computer.
In continuing this research work, the convolutional coded GMSK waveform model, turbo coded SDPSK and GMSK, as well as satellite architectural components must be ported from Matlab into C for implementation on the D940 platform. Also, a model must be developed (in C) for the satellite’s on-board packet switching module, to process information from various types of applications (voice, geospatial and navigation, images, etc.). An extensive hardware implementation will enable in-the-field real-time testing of a system that is more accurate (and faster) than those provided by computer simulation.
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