Overhead Channel Power Control in a Low Earth Orbit Satellite Communication System

by

Hussein M. Waljee

Submitted to the Department of Electrical Engineering and Computer Science in Partial Fulfillment of the Requirements for the Degree of Master of Engineering in Electrical Engineering and Computer Science at the Massachusetts Institute of Technology

May 22, 1998

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ABSTRACT

In the context of the Globalstar Low Earth Orbit satellite system, one of the critical bottlenecks affecting system capacity is the power available to each satellite. In order to better manage the use of this resource, four distinct strategies for power control of overhead channels are investigated. The main issues limiting the controllability of satellite power are identified as rain attenuation and uncertain system gains on the uplink path. The simplest strategy to successfully address these concerns is that of the “golden phone.” Within this context, an appropriate discrete-time controller is designed to provide acceptable power control in light of the inherent system uncertainties. Since the Globalstar system is not yet completely operational, the design process is made flexible to the incorporation of modified performance requirements as more experience is gathered. The designed controller is tested in simulation, and performs well in the face of uncertainty in system parameters and inputs. The ultimate result is the potential for an increased ability to manage satellite power, and thus increase the traffic carrying capacity of the Globalstar system.
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It’s over, baby, it’s over.

*hoy es hoy, con el peso de todo el tiempo ido,
con las alas de todo lo que sera manana.*

-Pablo Neruda
Chapter 1: Introduction

Wireless communication technologies are rapidly becoming absolute essentials for quick and reliable access to information sources and sinks. No longer is it acceptable to be out of touch when not sitting at a stationary computer or next to a land-line telephone handset. Similarly, it is becoming less excusable, and more frustrating, to be out of touch when outside the geographic wireless coverage area, or if the system capacity is exhausted.

To fill this need, private enterprise and government are rushing to develop systems that will provide global wireless communications coverage that is more accessible and reliable. However, wireless technologies by definition require the use of frequency allocations in the air, which are a scarce commodity. Even if these new systems can make a call possible and robust, accessibility and reliability would both suffer from limited system capacity as a result of limited bandwidth. This shortage is largely responsible for the latest developments in digital wireless communications standards which will enable more efficient use of frequency space in terrestrial cellular systems. From the microeconomic perspective of the suppliers, the motivation is obvious -- each additional user of a system provides added revenue, so the more users one can service, the more money one stands to make.

Globalstar is one system under development that uses a constellation of low earth orbit (LEO) satellites to provide ubiquitous global coverage for voice and data communications. After the system is reliably establishing and maintaining phone calls, the most critical technical goal is to maximize the system capacity. In the case of satellite communications, however, user capacity is influenced both by available bandwidth and by a new bottleneck -- available satellite power. Power efficiency is not a new goal for consumer electronics and personal communications equipment, because it directly influences battery size and talk-time, which are very marketable specifications. In terrestrial cellular systems, economics primarily require power efficiency in the phone, so the power control problem is near optimally solved in the case of traffic channels, which carry the actual phone call. By contrast, the overhead channels, which are only needed to establish and maintain the call, have come under less scrutiny for dynamic power efficiency. In cellular systems, these are
mainly generated by the cell-site, where power is cheap and unlimited, so efficiency is not critical. In Globalstar, however, these overhead channels use a large portion of the available satellite power, so controlling their resource usage is of critical importance. The setup of the Globalstar system, including the difference in role between the traffic and overhead channels is further explained in Chapter 2.

Noticing the need, this thesis addresses the design of a control strategy for overhead channel power in Globalstar. The result should be in accordance with the overall goals of minimizing satellite power wastage, and thus, maximizing user capacity of the overall system.

In order to establish the context in which this control system is developed, Chapter 2 presents a brief overview of the Globalstar system. It explains the major system segments, the basics of code division multiple access (CDMA) transmission, and the role of the aforementioned overhead channels. The chapter is intentionally very general and non-technical, in order to avoid unnecessary detail. Its purpose is to bring forth issues that eventually motivate and explain the power control method and design. If the system is already well-known to the reader, this chapter can be safely skipped, and the tables in Section 2.5 used instead as a reference.

In Chapter 3, the goal of overhead channel power control is revisited at a systems level. The power-limited nature of the forward link is introduced, and the motivation for power control better explained. The limitations in the current state of the system are identified, and they implicitly provide the design objectives of this proposed addition. Also, the major issues and obstacles affecting the ability of Globalstar to meet its goals are described. One of the factors identified is rain attenuation, which has been the subject of much on-going research in satellite communications. The relevant parts of this research are briefly presented by Chapter 4, in the specific context of Globalstar. By the end of Chapter 4, the background information necessary to directly address the power control strategy is established.

There are a number of strategies that can be utilized for estimating rain attenuation, and for directly controlling the satellite output power. Four different schemes of increasing complexity are described briefly in Chapter 5 in order to ascertain the relative advantages, disadvantages and costs of each. Chapter 6 is best
described as a feasibility study for the schemes involving the satellite beacon, and the new “golden phone”. The latter strategy emerges as a reasonable choice in terms of potential performance, and it is the basis for the subsequent design effort.

The design of a closed loop controller is developed using frequency-domain techniques in Chapter 7, and then finalized using a discrete-time domain analysis in Chapter 8. The final system undergoes simulation using SIMULINK to confirm performance and robustness in Chapter 9. The simulation also enables the testing of various contingency situations, and can show the effects of any unmodeled nonlinear dynamics that are unmodeled at the design stage.

The result of this development is a specific discrete-time controller for overhead channel power on the forward link of Globalstar. Perhaps a more important outcome is the establishment of a controller form that can be tuned quickly for different performance advantages. The eventual implementation of this controller should markedly increase the ability of Globalstar to intelligently manage the power available to it, and thereby increase the system capacity.
Chapter 2: The Globalstar System

The simplest way to conceive of Globalstar is as a regular land-based cellular system, but one in which the phone signals are bounced off a network of satellites in order to provide better geographic coverage. On the ground, the cell site and the user terminal communicate directly over the air, so the coverage area is inherently very limited. In Globalstar, the analog of a cell site is the Gateway (GW), which connects users to the terrestrial telecommunications infrastructure by relaying CDMA signals through one or more satellites that have line-of-sight contact with the user. The satellite itself, however, is simply a repeater. Its only function is to receive signals from the Gateway or the users, and then transmit that signal back to earth at a different frequency.

![Globalstar System Block Diagram](image)

**Figure 2.0-1: Globalstar System Block Diagram**

The block diagram in Figure 2.0-1 shows the structure of the Globalstar system. To the right, the Gateway relays calls to and from the generalized telecom network. Calls received from this network are then forwarded to the respective users via satellite over the Forward Link. Similarly, information originating from
Globalstar users is transmitted via satellite over the Reverse Link to the Gateway. As shown, the users each have a dedicated traffic channel for information transfer, but also share a set of overhead channels. These channels include the Pilot, Sync and Paging channels on the Forward Link as well as the Access channel on the Reverse Link. Their roles are further explained in Section 2.3.1.

2.1 System Segments

The Gateway consists of a central processing building, and up to four large antennas in the vicinity where each has a drive mechanism to track a satellite. The building houses the electronics that interface with the terrestrial phone switch infrastructure, generate the radio frequency (RF) carriers for the antennas, and perform the signal processing required to transmit and receive CDMA signals. This last task is by far the most complicated, and is briefly explained in Section 2.2. The Gateway is truly the workhorse of Globalstar, and is designed to operate without human intervention.

Intelligently coordinating the large network of Gateways, satellites and user terminals that comprise Globalstar is the role of the Ground Operations Control Center (GOCC), and the Satellite Operations Control Center (SOCC). The SOCC is the only system to deal with the satellites as objects rather than tools, and is responsible for making sure each stays on its prescribed trajectory, or ephemeris. The forty-eight satellites in the constellation travel on eight distinct low earth orbits at 1414 km altitude, and at a tangential velocity of 25757 km/hr. The global coverage pattern can be seen in Figure 2.1-1, and is noticeably geared for maximum overlapping coverage in the temperate latitudes. For an observer on the ground, a satellite sweeps from horizon to horizon in a speedy 17 minutes, and the time delay in ground-to-ground communication ranges from 12 ms at 10 degrees elevation to 9.4 ms overhead. In the power control context, the most important function of the SOCC is to validate the assumption that satellite position is known to the system very accurately.

The GOCC is considered by some to be the brain of Globalstar. It is mainly concerned with calculating system resource usage in conjunction with the SOCC, assigning available resources to the Gateways, and making critical system parameters available. The GOCC generates long-range plans based on projected traffic requirements and resource constraints such as available frequencies, available satellite power, Gateway
capacity usage and geographic service areas. These activities can be included under the umbrella of load leveling to use satellite and Gateway resources with optimal efficiency. In the context of overhead channel power control, the GOCC plays a critical role because it issues the resource allocation instruction (RAI) which specifies to each Gateway the amount of satellite power it is permitted to use during a particular satellite sweep.

2.2 Introduction to CDMA

Moving to a more basic level, the “physical layer” of the system is concerned with the actual transmission of the call through the air. It is a variation of the original ground-based Code Division Multiple Access (CDMA) technology, which provides a standard for channelized spread spectrum transmission, encryption, and power control. The details of this standard address the specifics of convolutional encoding, bit interleaving, signal constellations, and modulation that are applied to a digital stream in order to make its transmission efficient, robust and secure. Much of this processing is beyond the level of detail required to perform the power control task, so it is not addressed here. The relevant part of CDMA is the means by which many individual signals are summed together, sent over the same set of frequencies, and then successfully separated and identified on the receiving end.
2.2.1 Orthogonalization and Encryption

The first goal of CDMA is to create a set of orthogonal channels, so that the separate signals of different users do not mix together irrecoverably. In more traditional wireless systems, this is done by assigning a different frequency to each user (FDMA), or a different time slot to each user (TDMA), because that allows the signals to be separated easily on the receiving end. In CDMA, each user is assigned a “Walsh” code that is orthogonal to all the others, and can be likened to a specific language that is unique to the user. Every bit of information is sent as a series of 64 “Walsh chips”, and the result is summed and transmitted with the chips of all the other users in the system. Thus the signal of any one user is very small in power when compared to the total.

On the receiving end, one is metaphorically standing in a room full of people speaking different languages. If the receiver only knows one specific language, the voice using that language will stand out, and the rest will sound like noise. Similarly, if a unique code is reapplied at the receiver, the corresponding signal will increase dramatically in power. If this coding gain makes an individual signal powerful enough relative to the noise caused by all the rest, it can be accurately retrieved, or separated. Thus, with a sufficient signal to noise ratio, any one signal can be “pulled out of the noise.”

Once the Walsh code is applied, each signal occupies 1.2288 MHz of bandwidth -- that is, 1,228,800 chips are generated every second. The sum of a number of such orthogonal signals is still 1.2288 MHz wide, and this comprises a single CDMA channel. For one beam, each Gateway generates up to sixteen CDMA channels -- each around different carriers, which are unique combinations of eight frequencies and two polarizations. Thus, the CDMA channels generated by a Gateway are mutually orthogonal in the traditional FDMA sense.

Since the Walsh codes only separate the different signals within each CDMA channel, the same set of codes is reused in every CDMA channel. However, because satellite coverage is redundant as shown in Figure 2.1-1, it is possible to receive signals from two different Gateways using the same carrier frequency and polarization. There needs to be a way to identify the point of origin of each signal, so, a pseudo-random noise (PN) sequence is generated and multiplied onto the signal before transmission. The details of this process
are not addressed here, but by assigning unique phase offsets of this PN sequence to each satellite and Gateway combination, the frequency-coincident CDMA channels become orthogonal. Because of the one-to-one mapping, the identification of a PN offset also identifies the Gateway, the satellite, and the beam number on the satellite that is transmitting the signal received by the phone.

Another, perhaps more important function of PN sequences is that of encryption. By assigning the user a unique noise seed for PN generation, the sequence generated is uniquely known only to the system and the user. Thus, decoding the signal without knowledge of the seed is nearly impossible, and communication is secure.

2.2.2 Accurate Timing

One of the most critical issues in the implementation of the above coding and encryption techniques is that of timing. In the case of coding, when the procedure describes "reapplying" the Walsh code at the receiving end, the largest gain is achieved if the code is coherently reapplied so that it aligns well with the code originally applied at the transmitting end. In the case of encryption, the PN offset is related to phase; therefore, its identification requires some sense of absolute time. As timing becomes less certain, the decoding becomes non-coherent, coding gain decreases, the necessary transmitted power increases, and ultimately system capacity decreases. The coding process described above can be modified to better serve the non-coherent decoding case, but the resulting link will still carry less information than the coherent link.

Establishing a reference for coherent decoding can be quite complicated. One major factor is that satellite motion is fast, and therefore implies large time and frequency shifts as a result of the Doppler effect. Similar to a fire truck passing down the street, the time and frequency shift of the received signal changes over the satellite sweep. This issue of timing will eventually explain the role of overhead channels in Section 2.3.

2.2.3 System Capacity

Given the Walsh channelizing method described in Section 2.2.1, how is the capacity of a CDMA channel determined? In more traditional systems, capacity is easy to gauge because there are only a fixed number of frequency assignments or time slot assignments available. In CDMA, capacity is not deterministic, but is
rather interference-limited, or noise-limited. Moving back to analogy, in the room of conversations found at the receiving end, capacity is reached when the noise caused by others exceeds the ability of the target user to be heard. Each new user adds some noise to the noise floor, which in turn requires all other users to increase their respective power levels, or “raise their voice”. This subsequently increases the noise floor, and the cycle continues. Normally, this iterative process reaches an asymptotic limit which defines the new power level of each user in steady-state. However, at full capacity, the last user will cause this iterative process to go unbounded, and exhaust the power available to each user. This interference-limited channel capacity is a “soft” limit because it depends on the level of system noise that affects user signal-to-noise ratios. Since this noise level is time-varying, and to some degree controllable, the capacity also varies over time, and is not fixed by the system design per unit of bandwidth.

In Globalstar, the above analysis is in general true for each CDMA channel. However, when the system is considered in its entirety, a new factor begins to appear as a capacity bottleneck. As mentioned in the introduction, this factor is available satellite power, and its effect is discussed in Chapter 3.

2.2.4 Traffic Channel Power Control

Noticing that self-generated noise and interference is the dominant limit to CDMA system capacity, power control automatically becomes a very important issue. If any user terminal is amplifying its signal to unnecessarily high power levels, it is not only draining its own battery, but also creating more noise for other users. Therefore, it is decreasing the capacity of the entire CDMA channel in which it resides.

In order to prevent such wastage, a traffic channel power control system is included in basic CDMA. Although quite simple in concept, this power control loop is given the necessary resources to act very quickly, and its integrity is protected under most circumstances. In short, the system periodically monitors the quality of decoding on the receiving end, and based on a quality metric, issues one power control bit which is conveyed back to the transmitting end. The bit indicates either “raise the power” or “decrease the power”. Since, by default, the system favors the latter option, the decrease command is issued unless the decoding quality is less than some strict threshold. This ensures that the power being used at the transmitting end is kept to the minimum required for good quality reception.
2.3 Forward Link

The forward link (FL) is simply the path for information flowing to a Globalstar user terminal from a Gateway. As shown in Figure 2.3-1, the link further divides into the uplink and downlink segments.

![Figure 2.3-1: Globalstar Link Description and Frequency Plan.](image)

The antenna represents the Gateway, and the phone is the user terminal. The circular patterns show the footprint of the S-Band and L-Band antenna array sub-beams.

First of all, an example signal is traced through the forward link sequentially to reveal the stages involved. A given telephone signal is received by the Gateway and converted into a digital CDMA signal by a dedicated Globalstar Modulator Card (GMOD) which implements Walsh channelizing and error prevention coding. A number of these signals are then summed to compose a CDMA channel, and amplified under the auspices of the Gain Controller Unit (GCU). The result is upconverted to the specified carrier frequency, and finally transmitted from a highly directional dish antenna at about 5 GHz in the C-Band. The satellite receives this uplink signal using a rotationally symmetric antenna that has a gain which varies with satellite elevation. The on-board transponder amplifies and downconverts the received CDMA channel, then retransmits the result over one element of the S-Band downlink antenna array at about 2.5 GHz. The forward downlink is received by the Globalstar User Modem (GUM) which is inside the phone, or user terminal. This application
specific integrated circuit (ASIC) is responsible for multi-path diversity combining, and for undoing and decoding everything done previously by the GMOD. The result is, of course, the original voice signal.

Looking down from the satellite, the S-Band downlink has a coverage footprint as shown in Figure 2.3-1. The coverage is divided into sixteen separate areas, or sub-beams, which each correspond to a specific antenna array element aboard the satellite. The gain profile over any chord of this antenna footprint varies, because the pattern is asymmetric. As a result, the antenna pattern of each satellite is measured before being launched into space, and recorded as a matrix of gain values over a grid of points in the coverage area. The signal transmitted by each downlink sub-beam corresponds to a particular CDMA channel on the uplink. Since all the downlink sub-beams transmit at the same frequency and polarization, sixteen different uplink carriers (eight different frequencies) translate to one carrier frequency on the downlink. This explains the larger bandwidth required on the C-Band uplink (0.239 GHz) compared to the S-Band downlink (0.017 GHz).

2.3.1 Overhead Channels

The forward link of both Globalstar and terrestrial CDMA are obvious examples of one-to-many transmission schemes, because the central Gateway is transmitting to many users simultaneously. This observation, combined with the advantages derived from coherent demodulation of CDMA signals, make the creation of overhead channels on the forward link a natural evolution. In most systems, some resources (bits or bandwidth) are used to set up protocol, or coordinate the establishment of communications over other channels. Since it does not carry actual voice information, the bandwidth used for this coordination is a fixed or “overhead” cost to the system. In CDMA systems, overhead channels carry the added responsibility of providing a timing reference for use in coherent demodulation of Walsh codes, in identification of PN offsets, and in time and frequency tracking used to combat Doppler-related shifts. There are three types of overhead channels on the Globalstar forward link -- Pilot, Sync, and Paging.

The pilot channel plays the role of lighthouse within each CDMA channel by announcing the presence of a Gateway to all users in the corresponding coverage area. The pilot is a very simple signal which has “all ones” for bits, and which is encoded using the “Walsh 0” (all zeros) code. Since zero chips are modulated
into one unit of positive voltage, the result is essentially a constant positive voltage. This is then multiplied by a PN sequence which is generated by a noise seed that is common over all Gateways, and a PN offset that is unique to each Gateway-satellite combination. The resulting pilot signal is transmitted at a much greater power level than an average traffic signal, so it is easy to detect. When a user terminal is turned on, it searches the space of possible frequencies, PN offsets and Doppler shifts in order to find a pilot. Once the PN offset of this pilot is known, the identity of the transmitting Gateway, satellite, and satellite beam becomes uniquely clear.

Using this newly acquired PN offset, the sync channel of the same Gateway can then be decoded to retrieve system time, which helps to make coherent demodulation reliable. The sync also makes demodulation of the paging channel possible by providing the current data rate being employed. The paging channel contains a veritable wealth of information about channel assignments, system parameters, and neighborhood lists which make call maintenance more reliable and smooth.

The transmitted power of the sync and paging channels is linearly related to that of the pilot channel. Therefore, the control of radiated pilot power becomes the major concern addressed in this thesis.

2.3.2 Handoffs and the PSMM

Any two pilot signals from different Gateways together play another role that is important to the system -- they provide a relative power scale by which to judge the goodness of any Gateway connection. As mentioned earlier, a user terminal can often see multiple Gateways through different beams or satellites, but the one which provides the greatest pilot power is probably the best connection. For this reason, the phone continuously seeks out and tracks the three strongest pilots that it can see. For each of these, the GUM ASIC generates a Received Signal Strength Indicator (RSSI), which reports the estimated signal-to-noise ratio of the pilot based on measured power and accumulated energy. These reports are relayed back to the Gateway in the form of a Pilot Strength Measurement Messages (PSMM) over the reverse link, which is explained in Section 2.4.
This functionality is particularly good for handoffs between different satellites, beams and Gateways. If the RSSI for one pilot falls while another rises, the user should switch to the second Gateway. In Globalstar, dealing with handoffs efficiently has a large positive impact on system capacity, because satellites that travel from horizon to horizon so quickly make handoffs a common occurrence. In the context of this thesis, however, the PSMM is not introduced because of handoffs, but rather in the goal of providing feedback information to the power control system.

2.4 Reverse Link

The reverse link (RL) is the complement of the forward link for transferring information from the user terminal back to the Gateway and the connected ground network. Once again, it can be divided into an uplink and downlink portion as shown in Figure 2.3-1.

As in Section 2.3, an example signal is traced through the reverse link sequentially to reveal the stages involved. The user speaks into the phone, and the voice is encoded by the Globalstar vocoder, just as in the Gateway. This digital signal is then converted into a CDMA signal by the GUM ASIC, which does both the demodulation on the forward link, and the modulation on the reverse link. The Walsh encoded CDMA signal is then amplified and upconverted to the specified carrier frequency for the assigned CDMA channel, and finally transmitted from a vertical antenna at about 1.6 GHz in L-Band. The satellite receives this uplink signal using an array of seventeen somewhat directional antennas that have a coverage footprint as shown in Figure 2.3-1. The satellite transponder amplifies and upconverts the received signal, then simply retransmits the result of each receiver array element as one CDMA channel over the C-Band downlink antenna at about 7 GHz. The reverse downlink is received by the Globalstar Receiver Card (GREC) and then decoded by the Globalstar Demodulator (GDM) which resides in the Gateway. The result is, of course, the original voice signal generated by the phone.

2.4.1 Overhead Channels

The reverse link of Globalstar is a many-to-one communication scheme because many individual users transmit to a common receiving center in the Gateway. As a result, it is expensive in terms of bandwidth resources to provide the receiver with timing references from every transmitter. On the reverse link then,
non-coherent demodulation is the method used, and there is no need for a "pilot" channel. For call processing, however, the receiver needs to send messages back to the Gateway, like the PSMM from Section 2.3.2. These messages justify the creation of the Access Channel, which is the only overhead on the reverse link.

As alluded to in the section concerning CDMA timing, because non-coherent demodulation is a reality, the reverse link modulation scheme is slightly different from that of the forward link. This is true of both Globalstar, and terrestrial CDMA systems.

Since overhead channels play a minimal role in the reverse link, the entire link becomes peripheral to the aim of this thesis. For that reason, Section 2.4 is kept brief.

### 2.5 Summary and Reference

The above sections are intended to introduce the reader to the overall structure of Globalstar in order to provide a basis for meaningful discussion about overhead channel power control in Chapter 3. The most important concept presented is the motivation for including overhead channels like the pilot on the forward link -- that is, to establish coherent demodulation in the phone, and assist in general call processing.

In the process of explaining the relevant parts of Globalstar, a fairly substantial vocabulary has been introduced. This, of course, aids in future discussion, if remembered. In order to assist with the function of this chapter as a reference, the following tables identify the important acronyms and other vocabulary items that are now considered familiar.
### Table 2.5-1: Summary of Globalstar Terms

<table>
<thead>
<tr>
<th>Term</th>
<th>Explanation</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gateway</td>
<td>Equivalent of the cell-site in terrestrial CDMA. Supports Globalstar calls in a given geographic area.</td>
</tr>
<tr>
<td>GOCC</td>
<td>Ground Operations Control Center -- coordinates Gateways, and plans resource usage</td>
</tr>
<tr>
<td>SOCC</td>
<td>Satellite Operations Control Center -- manages the satellites, ensures ephemeris</td>
</tr>
<tr>
<td>GMOD</td>
<td>Gateway Modulator -- forward link CDMA transmission</td>
</tr>
<tr>
<td>GCU</td>
<td>Gain Controller Unit -- manages gain of Gateway</td>
</tr>
<tr>
<td>GUM</td>
<td>Globalstar User Modem -- CDMA reception and transmission in the phone</td>
</tr>
<tr>
<td>CDMA</td>
<td>Code Division Multiple Access -- the standard used for digital signal transmission</td>
</tr>
<tr>
<td>RAI</td>
<td>Resource Allocation Instruction -- message from GOCC to each Gateway</td>
</tr>
<tr>
<td>ASIC</td>
<td>Application Specific Integrated Circuit</td>
</tr>
<tr>
<td>RSSI</td>
<td>Received Signal Strength Indicator -- calculated for the pilots seen by phone</td>
</tr>
<tr>
<td>PSMM</td>
<td>Pilot Strength Measurement Message -- periodically reports RSSI to Gateway</td>
</tr>
<tr>
<td>FL</td>
<td>Forward Link -- for signals from Gateway to satellite to user</td>
</tr>
<tr>
<td>RL</td>
<td>Reverse Link -- for signals from user to satellite to Gateway</td>
</tr>
<tr>
<td>Pilot</td>
<td>FL Overhead Channel (unmodulated) which acts like a beacon for the Gateway</td>
</tr>
<tr>
<td>Sync</td>
<td>FL Overhead Channel that conveys system time</td>
</tr>
<tr>
<td>Paging</td>
<td>FL Overhead Channels that provide lots of system parameters</td>
</tr>
<tr>
<td>Traffic</td>
<td>Traffic Channels carry the actual voice/data communications for a given user</td>
</tr>
<tr>
<td>Access</td>
<td>RL Overhead Channel used for call processing</td>
</tr>
<tr>
<td>Walsh code</td>
<td>One of a set of orthogonal binary codes</td>
</tr>
<tr>
<td>Walsh chips</td>
<td>Each code consists of 64 binary chips</td>
</tr>
<tr>
<td>Chip Rate</td>
<td>Bandwidth of CDMA channel is 1.2288 MHz</td>
</tr>
<tr>
<td>PN sequence</td>
<td>Pseudo-Random noise -- looks random, but is known to system and reproducible</td>
</tr>
<tr>
<td>PN offsets</td>
<td>A mask applied on PN generator which causes a change in phase of PN sequence</td>
</tr>
</tbody>
</table>

Uniquely assigned to each Gateway-satellite-beam combination.
<table>
<thead>
<tr>
<th>Concept</th>
<th>Explanation</th>
</tr>
</thead>
<tbody>
<tr>
<td>noise floor</td>
<td>for any particular user, this is the sum of:</td>
</tr>
<tr>
<td></td>
<td>- power from sources external to the system;</td>
</tr>
<tr>
<td></td>
<td>- power from users of orthogonal Walsh codes in the same channel;</td>
</tr>
<tr>
<td></td>
<td>- power from users of neighboring satellite beams;</td>
</tr>
<tr>
<td></td>
<td>- power from users of Gateways with overlapping coverage areas.</td>
</tr>
<tr>
<td>interference-limited</td>
<td>This is a soft capacity limit for CDMA systems which occurs when all phones exhaust their ability to provide the necessary signal power to adequately exceed the noise floor on the receiving end once the Walsh code is reapplied and the inherent coding gain derived.</td>
</tr>
<tr>
<td>capacity</td>
<td></td>
</tr>
<tr>
<td>coding gain</td>
<td>The dramatic increase in signal to noise ratio that a signal experiences when the Walsh code that was used to encode it is reapplied at the receiver.</td>
</tr>
<tr>
<td></td>
<td>The constituents of the noise floor for a user are for the most part orthogonal to, or uncorrelated with, this code.</td>
</tr>
<tr>
<td>coherent decoding</td>
<td>Occurs in the presence of a timing reference which allows the Walsh code to be properly aligned when reapplied at the receiver. Provides superior coding gain than non-coherent decoding, and is used on the forward link of CDMA systems in a one-to-many environment.</td>
</tr>
<tr>
<td>non-coherent decoding</td>
<td>Correlated-energy based decoding that becomes necessary when the timing reference of the transmitter can not be reproduced efficiently at the receiver. Provides reduced coding gain, and is utilized on the reverse link of CDMA systems in a many-to-one environment.</td>
</tr>
</tbody>
</table>
Chapter 3: Overhead Channel Power Control

Once the overall Globalstar system is understood, the specific niche in which overhead channel power control resides can be explained in greater detail. First, the system conditions that bring about the need for additional power control are identified and fully explained. Also, since there is currently a mechanism for power management in place, the goals and limitations of the current open-loop system are identified. The major factors limiting current performance are rain attenuation and an uncertain satellite transponder gain. One management method for directing the impact of these uncertainties is evaluated in Section 3.4. Understanding these factors aids in evaluating the relative merits of the power control strategies presented in Chapter 5. For that reason, more insight is provided in Chapter 4 on current research regarding rain attenuation.

3.1 Emergence of Power Limitation

In ground-based CDMA, it is known that the channel capacity is essentially interference-limited. Also, the reverse link carries less capacity because it is at the disadvantage of using non-coherent demodulation. In order to maintain a two-way conversation, there needs to exist a reverse link traffic channel for each forward link channel, so it is the reverse link which becomes the bottleneck and dominates the capacity of the overall system. On the forward link, therefore, the system is a little less concerned with power control, because there is some spare margin. In the case of the overhead channels, pilot power is set above the minimum level to guarantee that intended users can acquire it quickly and accurately. The main reason to limit the pilot power is to demarcate the boundary of a particular cell, so that users in neighboring cells do not mistakenly attempt to contact the local cell when they are in reality too far away. There is also no need to dynamically set this power limit, because there are no significant gain or loss uncertainties that can universally change the received pilot power of all users over time. Finally, starting at the cell site, there is no lack of power availability-- it is rather cheap and plentiful.

In Globalstar, much of the above situation changes with the addition of a satellite that repeats and amplifies every signal on both the forward link and reverse link. Unfortunately, power on the satellite is not plentiful, and is not cheap. Since the satellite is driven mostly by solar energy, it follows that in steady-state operation,
the energy available to the communications payload over an orbit is equal to the integrated solar power
derived over that orbit. Solar exposure is not constant, so there is a rechargeable battery on-board which
stores energy for use when in the shadow of the earth, or for periods of high activity when instantaneous
solar power is not adequate. The battery also has a reserve, so, if more energy is needed in an orbit than is
collected, the traffic can still be supported. Utilizing this reserve, however, can severely impact the lifetime
of the battery and consequently, the entire satellite. This is therefore an extremely expensive way to sup-
port a larger communications load; in practice, it will not be done for routine operation.

How does this power limitation affect system capacity? In low traffic periods, each Globalstar CDMA chan-
nel is still interference-limited, as on the ground. However, it so happens that if every beam and sub-beam on
the satellite -- that is every available CDMA channel -- is loaded to its full interference-limited capacity, the
steady-state power availability of the satellite is exceeded. Since steady-state power is never to be exceeded,
the system is deemed power-limited. This limit is not enforced by the satellite, but rather by the GOCC and
the SOCC, which monitor the energy drained on each satellite by requesting periodic reports from the Gate-
ways that are using it. The GOCC then uses this information to allocate the amount of satellite energy each
Gateway is allowed to use over a given satellite pass. This command comes in the form of a Resource Allo-
cation Instruction (RAI), which also provides the ephemeris of the satellite, the maximum instantaneous
radiated power allowed by regulation in each downlink sub-beam, and finally, various estimated system
gains so that the Gateway is better enabled to meet the power and energy limits. The job of limiting power
usage is thus delegated by the GOCC to the Gateways.

In this new power-limited paradigm, the efficiency of both the forward and reverse links affects the capacity
of the system, because both use satellite power. Unlike in terrestrial cellular, the Globalstar forward link
does not have the luxury of slack capacity, so power control becomes more important. For that reason, a sim-
ple closed power control loop similar to that of the reverse link is implemented on the Globalstar forward
link traffic channels. This is an improvement, but the forward link also includes some powerful overhead
channels for which efficiency is left unchecked. There is no point in saving milliwatts in each traffic channel
if the pilot channel is routinely wasting watts -- hence the need for overhead channel power control.
3.2 Forward Link Power Management -- Open Loop

In the case of traffic channels, there is only one receiver, so power control is straightforward. The receiving user terminal generates a power control bit based on received power, and sends it over the already established reverse link to complete closed-loop control. In the case of overhead channels, every user is a receiver, so the feedback strategy is not as obvious.

On the forward downlink, each user in a given beam of the satellite is subject to widely and quickly varying reception because of changing topographic environments, atmospheric conditions, specular reflection, and even the angle in which the phone antenna is pointing. It would be undesirable to have these individual situations affect the pilot power for the entire beam.

In lieu of direct user feedback, a nominal operating point is established for the amount of overhead power to be radiated by each beam of the S-Band downlink antenna, $P_{SPEC}$. The various factors affecting the individual downlink paths are estimated in order to set an operating point by ultimately making trade-offs between capacity and system availability. In the case of terrestrial CDMA, knowing this operating point is usually enough to set the desired pilot power for a particular cell deterministically. In Globalstar, however, at least an open-loop controller is needed, because there are time-varying system gains along the path from the Gateway to the satellite output. Thus, the overhead channel power that is released from the Gateway must change over time in order to pre-compensate for the varying path gain and render the output of the satellite constant.

Most of these system gains vary as a function of satellite elevation, which in turn varies with time as in Figure 3.2-1a. The relevant factors that cause the path gain to vary are described below, and are also useful in understanding the systems described in later chapters.

One of the time-varying factors is the uplink path loss of the pilot signal, for which the profile is seen in Figure 3.2-2a. Radiated power decays as an inverse square function of distance, so path loss decreases as the satellite approaches overhead. Distance to the satellite is also responsible for a time-varying delay in round trip communication over the satellite, as mentioned in Section 2.1.
Figure 3.2-1: Elevation and Distance over a Satellite Sweep

The orbital path of the satellite specifies a plane passing through the center of the Earth. The line formed by the Gateway position and the center of the Earth makes an angle $\phi$ with above specified orbital plane.

a) (left) Distance from satellite to Gateway for various values of $\phi$.
b) (right) Elevation angle of satellite from Gateway for various values of $\phi$.

The thick horizontal line represents the horizon, and shows the maximum distance to the satellite. The thin horizontal line represents 10 degrees elevation, which is the minimum angle for Gateway contact.

Another of the time-varying uplink gains is that of the C-Band receiving antenna aboard the satellite. As implied by Figure 3.2-2b, this gain is small at 10 degrees elevation, then rises to reach a peak gain at 50 degrees elevation before falling once more by the time the satellite is overhead at 90 degrees. The antenna pattern is assumed to be rotationally symmetric, so the angular difference ($\phi$) between the satellite orbit plane and the plane in which the Gateway resides is of no concern. The relationship that is plotted in Figure 3.2-2b shows the combined antenna gain and path loss as a function of only the satellite elevation. The gain of the C-Band antenna also depends on the frequency used, so, in practice, there is a look-up table for each sub-beam frequency over elevation. The average curve shown in Figure 3.2-2b minimizes the peak-to-peak variation of the antenna gain over the uplink frequency band of 5.011 GHz to 5.250 GHz. Applying this average curve to the profile of elevation over a pass, Figure 3.2-2c shows the derived path gain profile over time during a satellite pass.

Lastly, the gain of the transponder aboard the satellite varies over time because of changes in aspects of its environment, like temperature or load level. This gain is not necessarily dependent on elevation; therefore,
Figure 3.2-2: Various Time-Varying Uplink Path Gains.

a) (top)  Path Loss because of distance travelled to satellite (dB)
b) (middle-left)  Uplink Path Gain (path loss and C-Band antenna gain) versus Elevation (dB)
c) (middle-right)  Profile of Uplink Path Gain versus time over satellite sweep (dB)
d) (bottom-left)  Example profile of predicted transponder gain over satellite sweep (dB)
e) (bottom-right)  Profile of total predicted GW gain required over time for constant satellite output.
the predicted transponder gain profile over a satellite sweep is supplied by the GOCC in the RAI. The prediction is made by the SOCC, which continuously monitors the vital statistics of the on-board environment using telemetry data, then applies a model to predict transponder gain. An example profile of transponder gain is given in Figure 3.2-2d.

The pre-correction to compensate for the above time-varying gains is carried out by the Forward Link Power Management feature, which resides in the Gateway. Prediction of the above uplink path gain profile is done in the Pre-Contact Gain Calculation (PGC) phase, and the appropriate compensation is implemented by the Transmit Power Tracking Loop (TPTL). This pre-correction activity is updated with a period of one second, and an example of a gain profile that the TPTL might attempt to follow over a satellite pass is given in Figure 3.2-2e.

3.2.1 Forward Link Energy Accountant -- FLEA

The TPTL is obviously an important addition in achieving the nominal operating point for output satellite power, $P_{SPEC}$. However, it does not fully ensure that the power usage mandate set out in the RAI is met by the Gateway. For that purpose, the Forward Link Energy Accountant (FLEA) is introduced. The FLEA measures the amount of power being radiated by the Gateway, applies to it the expected uplink path gain as found by the PGC, and integrates the result over time in order to estimate the amount of satellite energy drained in a particular sweep. If the Gateway seems in danger of exceeding the limit set in the RAI, it begins to limit the capacity of the system in various ways. Also, as mentioned in Section 3.1, this estimated power usage statistic is forwarded to the GOCC in order to enable more informed resource allocation in the future.

3.2.2 Limitations of TPTL and FLEA

Because this power management feature is currently operating in an open-loop manner, the main limitation in its accuracy is the unpredictability of uplink gain factors. Fortunately, both the gain of the satellite antenna and the path loss are known quite well because the ephemeris of the satellite is reliable. However, the transponder gain turns out to be less predictable, as does the path loss caused by atmospheric effects such as rain attenuation. The errors caused by these factors limit the ability of the TPTL to meet the satellite power out-
put specification, $P_{SPEC}$, as well as the ability of the FLEA to accurately estimate energy usage aboard the satellite.

The alternatives for dealing with unpredictability in the current system are described in Section 3.3, along with the respective consequences. However, the aim of this thesis is to design the additional control which should prevent the unpredictable factors from affecting the satellite output power. Possible design strategies for this addition are explored in Chapter 5, and include ideas like estimating the rain attenuation based on the received strength of the satellite beacon, or directly estimating the satellite power output based on measurements included in the telemetry stream carried by this beacon. If the latter is not available, a control loop could also be closed by using the periodic PSMM reports sent by the user terminals, or by building a forward link receiver at the Gateway. Before these strategies are discussed, an analysis of rain attenuation is presented in Chapter 4 in order to provide insight about one of these unpredictable factors, and the impact on the current system is evaluated in the following sections.

### 3.3 Consequences of Unpredictability

Given that there is an unpredictable factor in the uplink path gain, there are two ways in which it can impact the current system, and two clear management methods by which the system can direct this impact.

For the following discussion, suppose the open-loop TPTL system leads the Gateway to radiate a certain amount of pilot power, $P_{up}$, on the uplink, with the expectation that a corresponding output power, $P_{out}$, will result from the satellite. The actual power radiated at the output is $P_{real}$.

#### 3.3.1 Modes of Impact

If the unpredictable factor is a large unknown loss on the uplink path, less output power is generated than expected ($P_{real} < P_{out}$). The consequence is a reduced system availability from the point of view of the user, because it might take the phone longer to acquire the weaker pilot, or it might not acquire the pilot at all. Also, with reduced pilot power, the quality of the voice signal received might worsen due to time and frequency tracking errors. Ultimately, a call may be dropped altogether. A more subtle effect of the error is related to the activities of the FLEA. Believing that the power used is $P_{out}$ instead of $P_{real}$, the FLEA may
prematurely limit the user capacity of the system and render the satellite under utilized for the present pass. As a second order effect in the long term, the GOCC also believes that more energy has been drained from the satellite than in reality, and therefore, it may reduce the satellite power available for other Gateways to use -- once again reducing the system capacity utilization.

If, on the other hand, the unpredictable factor is an unknown gain, more than the expected output power is generated by the satellite ($P_{\text{real}} > P_{\text{out}}$). There are few negative consequences seen by the user in the short term, except a higher propensity of users in neighboring beams to prematurely handoff into this region. The long term consequences for the system are more detrimental, because the system is unknowingly violating the power limit set in the RAI. Through both first- and second-order effects as described above, the over-loaded satellite may be forced to tap reserve power and thereby reduce its battery lifetime.

3.3.2 Impact Management Methods

Faced with an unquantifiable uplink loss, it may be desirable to raise the power of the pilot for all time in order to avoid a reduction in system availability. Knowing that the pilot power is being increased, the FLEA can manage the traffic load in order to make sure that satellite power and energy usage do not violate the RAI limits. However, there are obvious consequences involving steady-state system capacity under this strategy, and these are discussed in Section 3.4 within the specific context of rain attenuation.

Similarly, faced with an unquantifiable uplink gain, the pilot power can be reduced in order to protect the satellite battery lifetime. Of course, this has an effect on system availability when the unknown gain is absent. Alternatively, the pilot power can remain the same, but the FLEA can assume that the worst-case (largest) uplink gain is applied, and then decrease traffic load accordingly. In this way, system capacity is again traded for an increase in battery lifetime without sacrificing system availability.

In general, for factors that can be either gains or losses, there is some combination of the above strategies that will trade off the relative benefits in system availability, system capacity, capacity utilization, and satellite battery lifetime. A systems designer who is knowledgeable about the relative cost of reductions in any of the above system parameters will be best equipped to decide the appropriate trade-off level amongst them.
For future discussion and comparison, however, the impact of uncertainty is always translated to, and quantified as, a reduction in system capacity.

### 3.4 Effect of Overdriving Overhead Channels on System Capacity

As mentioned in Section 3.2.2, one of the largest unpredictable factors affecting the total gain of the uplink path is rain attenuation. As will be discussed in Chapter 4, at Globalstar frequencies this loss could vary anywhere from 0 dB to 12 dB, as predicted by using certain rain rate estimation models and attenuation prediction models. For the most part, however, it can be assumed that the rain fade rarely exceeds 2 dB. In the spirit of Section 3.3.2, the current system developers have decided to minimize the impact of this rain attenuation on system availability by permanently increasing the gain of the Gateway by 1 dB. This is done primarily for the benefit of the overhead channels, which are needed for coherent demodulation. Of course, this improvement in system availability is not free, because system capacity suffers as a result.

Essentially, any satellite power that is used for the overhead channels is not made available for use by traffic channels, so the traffic capacity of the system decreases. If, for example, the overhead channels normally consume forty percent of the satellite power, a 1 dB increase raises this to fifty percent. Assuming that the rest of the power is used for traffic channels, the power available to them decreases from sixty to fifty percent, which is a reduction of one-sixth. That is a large capacity hit for some gain in robustness. Of course, moving from an “available power” metric to a “system capacity” metric requires some knowledge of the average power per user, so this is assumed to remain constant despite the 1 dB increase in pilot power. In the case where rain attenuation is present, this ratio might increase, and further reduce capacity, but this reduction is independent of that caused by extra overhead power. Even in a controlled situation, rain attenuation will always reduce system capacity, but this effect should be kept separate from the capacity wastage that is cause by the overhead overdriving policy.

In the above example, continuously overdriving the overhead channels by 1 dB leads to the wastage of ten percent of the satellite power when there are clear skies. The portion of satellite power that is devoted to traffic channels is reduced from sixty percent to fifty percent; therefore, one sixth of the possible traffic capacity is wasted. Simple statistics like these can be calculated for each combination of overdrive ratios (in this case...
Figure 3.4-1: Effect of Overdriving the Pilot

1 dB), and the percent of satellite power originally targeted for overhead channels (in this case forty percent). Figure 3.4-1 provides some idea of these statistical trends as these values vary. The figures assume that the only overhead channel in use is the pilot, but this causes no loss of generality because the power of the paging and sync channels are linearly proportional to that of the pilot.

Globalstar user channels on both the forward and reverse links have dynamic closed-loop power control systems, so they adapt to rain attenuation automatically, and optimally. The overhead channels, on the other hand, are continuously overdriven by 1 dB to make sure the system is robust to atmospheric attenuation of that magnitude. Under this solution, it is hard to disregard the large capacity reduction the system suffers for
95+% of the time, in order to maintain calls in periods of high rain which occur less than 5% of the time. The goal of this project is to create a dynamic control system for the pilot so that extra power is used only when it is needed. The recovered traffic capacity is the reward for a robust control strategy that provides good disturbance rejection in the case of rain attenuation.
Chapter 4: Rain Attenuation Analysis

One of the key hurdles in the establishment of efficient satellite communications is uncertainty in the channel, and more specifically, the atmospheric effects on radiated signals. The attenuation and depolarization caused by rain have been the subject of much research over the years. Perhaps one the first and most important conclusions made is that for the same rain rate, signal attenuation increases dramatically with carrier frequency. Since it has generally been possible to use lower frequencies for satellite systems, rain has not been a grave problem. However, presently, bandwidth assignments are using much higher carriers, so rain is becoming a critical concern. In fact, at the now popular Ka-band (20GHz), the attenuation can routinely exceed 15 dB. This increased criticality has led to a resurgence in rain related research [11].

Fortunately, Globalstar is blessed with a relatively low transmitting frequency of 5 GHz on the forward uplink. Statistics and attenuation models show that we can typically expect no more than 2 dB of attenuation even under the harshest circumstances, as is explained in the later sections of this chapter. It is possible, although unlikely, that such a fade would be fatal to a Globalstar call. However, if rain attenuation, or the threat thereof, is left dynamically uncompensated, it can potentially decrease the capacity of the entire Globalstar system, as explained in Section 3.4. Strategies for recovering this capacity by minimizing the effect of rain and other uncertainties are introduced in Chapter 5.

A number of studies have been conducted by various organizations to estimate signal attenuation at different frequencies, given a particular rainfall rate and satellite elevation. This is useful only if one has knowledge of the rainfall rate, so other studies have concentrated on historical rainfall statistics, and temporal rain patterns in different climates around the world. Brief overviews of these rain rate models and attenuation prediction models are presented in Section 4.1 and 4.2 respectively.

4.1 Rain Rate Statistics

Rain is itself an essentially statistical process, and most often characterized by “exceedance curves”, which specify the rain rate, in mm/hr, that is exceeded for a certain percent of all time. Such exceedance curves can be generated for any point location, given enough empirical data, but this is often tiresome and usually not
justified -- after all, the rain rate statistics in Buffalo are not likely to vary much from those of Toronto, for example. Various research groups have analyzed global rain rate data, and have created “climate regions” within which rain patterns can be considered statistically the same. Two such regional rain statistics models are the Global model, which is presented independently [5], and a slightly different model developed by the Consultative Committee on International Radio (CCIR) [14]. A comparative analysis of these two models has also been conducted, and tends to favor the Global model as statistically more reliable [5]. Maps showing both sets of climate regions, and the associated rain rate exceedance curves can be found in the references listed at the end of this thesis [3, 15]. The same rain climate models are also summarized in the satellite communications survey textbook by Pratt and Bostian, [20, pp. 334-340].

4.2 Attenuation Prediction Models

The rain rates observed are only part of the problem at hand, because the impact depends on how the rain rate affects signals that are travelling through it. Based on both theoretical and empirical results, a number of attenuation prediction techniques have been developed for use when the rain rate is known and assumed to be uniform across an entire cloud. Two of these attenuation models are the Simple Attenuation Model (SAM) that was developed with NASA support, and the CCIR attenuation model [15], which is not to be confused with the CCIR climate model for rain statistics that was mentioned in Section 4.1. These models take the value of the rain rate, $R$, in mm/h, the carrier frequency to be used, in GHz, and satellite elevation in degrees in order to predict the rain attenuation on the specified communication link.

Both models use a simple formula relating specific attenuation, $\alpha$ (dB/km), to frequency (GHz) and rain rate, $R$ (mm/h). The constants $a$ and $b$ in the equations below are dependent on the carrier frequency, and have been compiled into empirical formulas that are presented below for Globalstar frequencies [2, pp. 406; 3]:

$$\alpha = aR^b$$

$$f = 5.15\text{GHz} \quad a = 4.21 \times 10^{-5} f^{2.42} = 2.22 \times 10^{-3} \quad 2.9\text{GHz} \leq f \leq 54\text{GHz}$$

$$b = 0.851 f^{0.158} = 1.102 \quad f \leq 8.5\text{GHz}$$

Each of the two attenuation models then has its own method of calculating the effective length, $L_{\text{eff}}$ in km, over which the specific attenuation, $\alpha$, is applicable, and thereby obtains attenuation, $A$ in dB, as $\alpha L_{\text{eff}}$. The
equations and variable definitions required to make the prediction are given in Table 4.2-1. It is assumed, in the worst case, that the rain rate is greater than 10 mm/h and that the latitude of the observer is less than 30°. It is also assumed that the satellite elevation is greater than 10°, because this is the minimum satellite contact elevation for a Globalstar Gateway. If these assumptions need to be relaxed for any reason, a more complete set of equations is available in the references [20, pp. 334-340].

### Table 4.2-1: Equations Used in Attenuation Prediction Models

<table>
<thead>
<tr>
<th>Simple Attenuation Model</th>
<th>( L = \frac{H_e}{\sin(EL)} ), ( H_o = 0 )</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>( H_e = 4.8 + \log \left( \frac{R}{10} \right) ), (</td>
</tr>
<tr>
<td></td>
<td>( A = aR^b \left( 1 - e^{b\gamma \ln(R/10)L\cos(EL)} \right) ), ( R \geq 10\text{mm/h} )</td>
</tr>
</tbody>
</table>

| CCIR Model | \( h_R = 5.1 - 2.15\log\{1 + 10^{((|\Lambda_o| - 27)/25)}\} \) |
|------------|----------------------------------------------------------|
|            | \( L_s = h_R/(\sin(EL)), EL \geq 10° \) |
|            | \( r_p = \frac{90}{90 + 4L_s \cos(EL)} \) |
|            | \( A = aR_p^bL_{sr}r_p\left( \frac{P}{0.01} \right)^{-\rho} \) \( \rho = 0.33, \ 0.001 \leq P \leq 0.01 \) |
|            | \( \rho = 0.41, \ 0.01 \leq P \leq 0.1 \) |

- **A**: attenuation (dB)
- **EL**: satellite elevation with respect to observer (degrees)
- **\( \gamma \)**: empirical factor of 1/22
- **\( H_e \)**: SAM effective rain height above sea level (km)
- **\( H_o \)**: height of observer above sea level (km)
- **\( h_R \)**: CCIR effective rain height above sea level (km)
- **\( \Lambda_o \)**: latitude of observer (degrees)
- **\( L_s \)**: length of horizontal projection of the slant path to satellite. (km)
- **\( P \)**: percent of time the desired estimate of attenuation can be exceeded. (%)
- **R**: rain rate (mm/h)
- **\( r_p \)**: CCIR path reduction factor
- **\( R_p \)**: rain rate exceeded for 0.01% of the time (mm/h)
4.3 Results of Applying the Models

In order to evaluate the extent of rain fade that might affect Globalstar, a number of MATLAB scripts have been composed to carry out the calculations described by the different models, at the specific Globalstar uplink frequency of 5.15 GHz, and a worst case satellite elevation of 10 degrees.

One script, RainExc.m, finds the rain rate exceedance curve for every climate, by referencing the user-specified rain statistics model -- Global or CCIR. It then applies both the SAM and CCIR attenuation prediction models to each of the points on the rain rate exceedance curve. The result is a new set of exceedance curves showing the expected attenuation in dB that is exceeded for a certain percentage of the time.

Since there are too many climate regions to consider at once on the same plot, the number of rain attenuation exceedance curves presented in Figure 4.3-1 is reduced to the four most commonly found in the set of planned Globalstar Gateways.

Reading these curves, it is found that although one of the models predicts over 12 dB of rain attenuation for some very small amount of time, for 99.9% of the time the attenuation is less than 3 dB, using the SAM prediction model with the Global climate model. The other model combinations do not differ greatly at this exceedance percentage (0.01% of the time). In all but the rainiest climates, the attenuation is less than 2 dB most of the time. Thus, it is safely assumable, for the purposes of overhead channel power control, that the uplink rain attenuation ranges from 0 to 2 dB.

The MATLAB scripts used to carry out the model calculations, and which carry the rain rate exceedance information, are also found in the Appendix on page 116.

4.4 Frequency Scaling Techniques

One useful conclusion derived from the research is that rain attenuation frequency scaling empirically follows an approximately square relationship. In other words, knowing the attenuation at one frequency, the attenuation at another can be estimated using the following relationships:
The simple square relation is suggested by a satellite communications textbook [20], and the more complicated version by the Consultative Committee on International Radio (CCIR) [15].
The rain analysis mentioned above concerns only the magnitude of attenuation to be expected. Rain can also cause depolarization of signals, and research in this area is available. However, depolarization is not addressed further here because it is quite minimal at Globalstar frequencies.

4.5 Rain Dynamics

A select few of the references have analyzed statistics such as rain rate duration [13], frequency scaling methods [11], and even mitigation methods for these rain fades [6]. Most of the analyses, however, have been directed to higher frequency bands like the Ka-band, because they suffer greater attenuation than the C-band or S-band.

In general, rainfall studies have also been very temporally oriented, rather than spatially-oriented -- that is, they monitor rainfall rate for a particular location over time, rather than over a large area at the same time. This difference becomes important when analyzing LEO satellite systems, because the speed of the satellite over a cloud is conceivably faster than the rate of change in rainfall intensity, or in the movement of clouds. Nevertheless, spatial rain patterns can be inferred in order to find the worst case for rain fade dynamics over a Globalstar satellite sweep.

In order to determine the potential rate of increase in rain attenuation, it is assumed that the line-of-sight path to the satellite first encounters rain at its highest predicted extent. It is also known that the rain attenuation will increase as more of this communication path becomes affected by the rain. Using some simple geometric calculations, it is found that 2 dB of rain fade can be encountered within 15 seconds of first contact with the rain. The maximum rate of increase, then, should be no larger than 0.2 dB per second. This figure is of assistance in determining the speed with which the power control strategies of Chapter 5 must react to disturbances in the system.
Chapter 5: Survey of Possible Strategies

Globalstar capacity depends on the ability of the system to optimize and accurately monitor the downlink output power spent by the satellite over time. The TPTL system for overhead channel power control operates by predicting the uplink path gain, and then pre-compensating for it by setting the output power of the Gateway accordingly. However, as explained in Section 3.4, the factors of rain attenuation and other uncertain system gains make the task of prediction difficult and inaccurate. In order to compensate for these uncertainties, more information must be gathered and then used efficiently to quantify or combat these factors on a real-time basis. The following sections introduce four strategies that gather pertinent measurements from the system in order to predict the uplink gain more accurately, or control the satellite output power directly. These strategies are briefly described in this chapter, along with their relative advantages and disadvantages, both in performance and ease of implementation. Chapter 6 takes this analysis one step deeper by examining the feasibility of two strategies in detail.

5.1 C-Band Satellite Beacon

Each of the Globalstar satellites is equipped with a beacon that constantly transmits in a narrow frequency band at the lower end of the C-Band allocation on the reverse link. The information encoded in the beacon is telemetry data from the satellite, which will be discussed later. For the application of the particular strategy to be discussed in this section, the encoded information is irrelevant because only the received power of the beacon is needed.

In the Gateway Receiver card (GREC), the beacon can be treated like simply another reverse link user channel. Since each channel needs to be at a pre-determined power level for the demodulation process to function properly, every signal is fed to an Automated Gain Controller (AGC) which fixes the signal power at its output. Knowing the instantaneous gain value of the AGC, and knowing the fixed output signal power, an estimate of the input power can be calculated. This estimate takes the form of a Received Signal Strength Indicator (RSSI), and is important because the beacon travels through the same atmospheric path as the
uplink signal. If the power of the beacon under clear skies is predictable, then rain attenuation can be inferred by comparing this expected beacon power with the measured RSSI.

The beacon strategy described above is summarized in Figure 5.1-1. The received beacon power is measured at the Gateway, and then compared to the expected power based on knowledge of the system. The result enables the estimation of rain fade that might also affect the uplink signal, and thus provides a means to pre-compensate for this fade.

The power of the satellite beacon is defined in satellite specifications, but the accuracy with which it is controlled is of key importance. If the satellite steadily transmits at a specified absolute power that is common to all satellites, the rain fade estimate can be quite accurate with a simple calculation. The result of transmitted beacon power minus the received power and minus the expected downlink path gain indicates the atmospheric effect, or estimated rain fade, at the beacon frequency. Using the frequency scaling formulas mentioned in Chapter 4, the system can then estimate the uplink rain fade and increase the transmitted power accordingly.

If the absolute beacon power transmitted is not known or not consistent among satellites, there will be a bootstrapping problem when the first contact is made with each satellite at ten degrees elevation. For exam-
pie, if Satellite A always transmits a beacon at 0.5 watts (-3 dB) while Satellite B transmits at 1 watt (0 dB), the Gateway may incorrectly assume that 3 dB of rain fade is affecting the link from Satellite A, even though both links may have clear skies.

Once the bootstrapping problem is addressed, rain attenuation can still be detected by the trend in received power compared to the profile of the expected beacon power over time. If, however, the downlink path gain profile is not known over the sweep of a satellite, this entire strategy becomes useless.

Since the beacon power is independent of the uplink power, this is an open-loop system, and is structured more as a method to estimate rain attenuation. It may not be the most accurate of the four strategies, but it is very simple and attractive because all the necessary equipment is already in place at the Gateway and in the satellite. Only a few minor modifications are needed to implement this scheme.

One of the other motivating factors behind the beacon scheme is that a similar strategy has been previously tested using Advanced Communications Technology Satellites and higher carrier frequencies [6]. This documented uplink power control system estimates the rain fade acting on a pilot signal at 27 GHz, based on measurements of a 20 GHz beacon. Both signals suffer much greater atmospheric attenuation than Globalstar signals because of their higher frequencies. In the system presented, measurements are made to validate the accuracy of the frequency scaling techniques and to determine the control error of an open-loop system like the one proposed above. The results are satisfactory for the purposes of their system, because up to 30 dB of potential rain fade is estimated to within 2.5 dB. This error is larger than the acceptable bounds in Globalstar, because the rain fade itself is only up to 2 dB in most cases. However, the Globalstar system might provide some tools to obtain better accuracy. The control error may also be frequency dependent, so if the earlier scaling techniques are applied to the error, one could expect errors as low as +/- 0.3 dB.

The advantages of this scheme are that it is simple, cheap and quick to implement. It also promises to have a fast reaction time to fades, and has been documented as effective in another similar situation at higher frequencies. The largest drawbacks to this strategy are the detailed knowledge of the system required for accu-
racy, the dependence upon a frequency scaling factor, and the open-loop format. The last of these three drawbacks limits the use of this strategy to the compensation of only rain fade uncertainties.

5.2 **Golden Phone**

Answering for the key deficiency of the beacon scheme, this strategy proposes a closed-loop method of controlling the overhead channel power output of the satellite. The loop is closed using a “golden phone”, which is simply a Globalstar phone that has a potentially more accurate RSSI, and a better, more directional antenna that alleviates multi-path and specular interference. The phone is planted on top of the transmitting antenna at the Gateway, and it measures the received pilot power on the S-Band downlink. For each user terminal, recall that this same RSSI measurement is sent to the Gateway periodically on the reverse link as a Pilot Strength Measurement Message (PSMM).

![Golden Phone Closed Loop Scheme](image)

**Figure 5.2-1: Golden Phone Closed Loop Scheme**

Since the uplink and downlink portions of the forward link share the same atmospheric path, and because the rain fades at different frequencies are related using the scaling techniques, the rain attenuation can again be estimated. This, however, requires the knowledge of all other system gain on the forward link, and limits the use of the strategy to rain attenuation.
Remembering the ultimate goal of controlling the satellite output, however, the system can be re-structured to use the RSSI and knowledge of downlink gains to estimate the power transmitted by the satellite. The comparison of this estimate with the $P_{\text{SPEC}}$ identified in the RAI provides an error signal which is fed to an appropriate controller and then used to change the transmitted uplink power of the Gateway. The key in this strategy is to find the appropriate controller that provides the desired performance in light of inherent system delays. Any error in the measurement devices like the golden phone, or any uncertainty in the downlink system parameters can be treated as noise sources in the loop.

The golden phone strategy is depicted in Figure 5.2-1. Two advantages of this strategy are that it does not depend on an unreliable frequency scaling factor, and that its closed-loop nature makes it more versatile and accurate. More importantly, this system does not limit its usefulness to rain attenuation or fades that affect both the downlink and uplink in some related way. The error in the transponder gain pre-correction activity is one example of an error combatted by this system, but left untouched by the beacon scheme.

The main problem with this strategy is that the downlink signal being measured is not always available. Globalstar frequency allocations for each satellite and geographic area do not guarantee that each Gateway is assigned a CDMA channel in its resident S-Band beam. Another hurdle for this strategy is the sheer number of different system parameters that need to be known. These values might not be in an accurate or easily available form, so the look-up function may introduce significant errors and delays in the system, and thus constrain the loop performance.

Finally, with regard to implementation, the “golden phone” does not currently exist physically, but performance requirements for the receiver can be determined by this design, and incorporated into the golden phone design. Other than the phone itself, there is little extra capital involved in implementation, because the actual controller is digital, and expressed as part of the GCU software.
5.3 Telemetry Data Stream

For a closed loop control system, the best possible resource would be a direct measurement of the output variable, which is, in this case, satellite output power. Such a measurement is tantalizingly proposed by the telemetry data that is encoded by the Globalstar satellite beacon.

If this data were made available to every Gateway, the system built around it may be that of the golden phone closed-loop strategy, without a lot of the uncertainty. The measurement of output power is direct, so no inference is required, and hence downlink path gain estimates need not be made. The system compares the measured output power with the desired value, $P_{\text{SPEC}}$, in order to generate an error signal. This error signal is processed by an appropriate controller to determine a gain value for the overhead channels that are leaving the Gateway.

The advantages of this system include simplicity, and a lower time delay in the feedback loop, which allows greater stability and speed. The telemetry resource also has some rather important disadvantages, however. First of all, it is not clear exactly where and how this power measurement is taken in the satellite. Investigation of this issue should provide critical information concerning the form, reliability and accuracy of the telemetry data. Secondly, it is unclear how often the power output value is made available, because it cannot be requested – it is sent periodically among a list of many other pieces of telemetry data. Finally, the equipment used to decode and retrieve this data, the Telemetry and Command Unit (TCU), is not expected to reside in every Gateway. The TCU is primarily a utility for the SOCC, so it is installed at only four locations in the world. If chosen as part of the power control strategy, the inclusion of a TCU at every Gateway might be somewhat expensive.

5.4 User PSMM Reports

The fourth and final strategy addressed in this chapter utilizes equipment at locations other than those of the Gateway. Just as the golden phone produces an RSSI, each user terminal periodically and regularly produces a set of RSSI values for three different pilots, and these are relayed to the primary Gateway in the form of a PSMM. These measurements are inherently less accurate and more susceptible to unusual individual circumstances than those of the golden phone, but if collected for some time and analyzed, they may constitute
the basis of a pilot power control strategy. Before the information can be used, however, there needs to be 
some normalization for the type of user terminal generating the report.

This strategy forgets the specification of $P_{\text{SPEC}}$ watts at the output of the satellite, and deals with the more 
basic problem of providing each user with the necessary pilot flux density for proper coherent demodulation. 
One proposed method is to normalize the individual PSMMs and find the maximum, which supposedly re-
resents the maximum pilot power available to a user in a beam. This pilot power should meet some thresh-
old, but not exceed it extravagantly, so the threshold becomes the basis of uplink power control.

![Multiple PSMM Scheme](image)

**Figure 5.4-1: Multiple PSMM Scheme**

One problem -- and advantage -- of this strategy is essentially statistical in nature. By using information that 
is generated after the downlink on a number of unknown paths, there is some statistical variation resulting 
from the particular orientation of each individual user. The effect of a random downlink situation needs to be 
filtered out to make any conclusion about the uplink. By the same token, the statistical advantage of a large 
number of users may filter out the measurement errors that are experienced by a single golden phone.

Another disadvantage of this scheme is found in the details of its implementation. It turns out that the part of 
the Gateway which receives the PSMM is largely concerned with hand-off procedure, and is not physically 
or structurally near the GCU. The frequent shuttling of PSMM information is a drain on the Gateway data
infrastructure, so the added accuracy might not be worth the performance cost. This problem is compounded if the PSMMs are to be assigned geographic tags, which are useful in detecting a distinct area of rain that affects a small number of downlink beams.

One statistical issue that may at first seem to be a problem is the lack of enough actual users in a beam to provide the aforementioned statistical advantage in reducing the variance of the measurement error. Fortunately, since this power control system is attempting to squeeze the last ounces of capacity out of the system, one can assume that the scheme might be utilized only when that capacity is required. When there are fewer users, the pilot power can be increased a little more freely. When there are many users, a scheme based on user PSMM reports will have enough data points to set the pilot power efficiently.

The system diagram in Figure 5.4-1 provides a summary of this strategy. One of its advantages is statistical in nature, and the other is the continuous availability of user PSMM reports. For that reason, it is often considered together with the golden phone scheme. When the frequency allocations are favorable -- that is, when a Gateway has a geographic service area that includes itself -- the golden phone can provide a good calibration measurement by which to judge the user PSMMs. When the golden phone is shadowed, the ever-available PSMM scheme can take over control without the help of a calibrated golden phone. The biggest disadvantages of this strategy concern implementation, and can probably be overcome if the performance gain is shown to be significant.
Chapter 6: Feasibility Studies

The strategies presented in Chapter 5 provide many ways in which to build an overhead channel power control system; however, only one can be the fundamental strategy. Once in place, distinct aspects of the other strategies can be added to augment overall performance. In order to choose the fundamental scheme, feasibility studies are applied to each strategy in order to determine if adequate performance is possible. There is no reason to use a complicated strategy when a simpler one is adequate, so they are analyzed in order of complexity. The satellite beacon scheme is analyzed first in Section 6.1, but is determined inadequate for the control task. The “golden phone” strategy is subsequently visited in Section 6.2, and it has the potential for good performance, so it is further analyzed in Chapter 7.

6.1 Satellite Beacon Strategy

The satellite beacon is a constantly available resource from the point of view of the gateway. Also, the number of calculations, the amount of system knowledge, and the extent of redesign required to estimate rain attenuation are relatively minimal compared to the other strategies. As shown by the example presented in Section 5.1, many satellite system designers consider the beacon scheme as the default method for dealing with rain attenuation, and it has largely been assumed that such a scheme would suffice in Globalstar as well.

6.1.1 Description

This strategy estimates the uplink rain fade by using the attenuation of a known downlink signal that travels the same atmospheric path. The satellite beacon is the signal used, and its received power is measured on the ground using a slight modified Gateway Receiver Card (GREC) that essentially treats the beacon as if it were a regular traffic channel. External research has developed frequency scaling techniques which can be used to estimate the 5.15 GHz uplink fade based on the 6.88 GHz beacon fade, as shown in Section 4.4.

6.1.2 System Model

The block diagram describing this strategy is presented in Figure 6.1-1, and the variables used are defined under the picture. The working unit of the signal arcs is power in decibels or dBW. For this reason, the mul-
tiplicative noise sources and uncertainties can be expressed as additive influences on the system, and their conglomeration effect can be analyzed using superposition.

The figure shows two signal paths which branch from the beacon signal of constant expected power $P_{BTX}$. While the upper path tracks the actual beacon signal power, the lower path resides entirely within the proposed controller, and demonstrates the effort to track the expected beacon power as it travels the C-Band downlink. The expected multiplicative factors that affect the beacon include the gain of the satellite C-Band antenna, $G_{BTx}$, the path loss incurred over the distance travelled, $L_{PB}$, and the gain of the receiving antenna at the gateway, $G_{BRx}$. There is also an expected time-varying propagation delay, $D_{B1}$, and another time delay, $D_{B2}$, which is explained in the description of the upper path. The above gains and delays are predicted given all the information available to the open-loop TPTL control system a priori to satellite contact. Therefore,
the result of the lower path in the diagram is the time profile of the expected received beacon power, \( P_{BRx} \), during a satellite sweep.

The upper path in Figure 6.1-1 represents the actual beacon power as it travels the C-Band downlink. The multiplicative factors that affect this signal correspond well to those that affect the expected beacon power. However, in reality, the values of these factors are random variables with some probability distribution centered about the expected value. The ratio, or logarithmic difference, between the actual value, \( G_{SUB} \), and the expected value of each factor, \( G_{SUB} \), is labelled as the error, \( E_{SUB} \). Because the real value of these random gain factors can be expressed as the product of the expected value and this error ratio, the actual beacon power is affected by both of these components. Therefore, in the upper path, Figure 6.1-1 clearly shows the addition of the same expected gains as for the lower path, plus the corresponding error ratios that make each physical factor a true random variable. In addition to these obvious system gains, there are a number of other errors that unpredictably affect the actual beacon power. These include the loss due to rain attenuation, \( L_{RB} \), as well as the error inherent in the system which controls the transmitted beacon power, \( E_{PMB} \), and the error inherent in the measurement of the beacon power, \( E_{AGC} \). The final result of the upper path, which models the actual physical processes, is the real-time profile of beacon power measured by the automatic gain control (AGC) in the GREC.

The AGC requires some time to generate the power measurement, so this explains the time delay, \( D_{B2} \). The delays experienced by the actual beacon signal (\( D_{B1} \) and \( D_{B2} \)) may differ from the expected delays (\( D_{B1} \) and \( D_{B2} \)), but this discrepancy is not efficiently expressed by decoupling the random portion. The possible inconsistency between the actual delay and the expected delay is indicated by italicization.

Once the expected beacon power and the actual measured power are known for the same point in time, they can be compared in order to estimate the rain attenuation. The difference between these signals can be caused by any of the error ratios added to the upper path, but the entire difference is assigned to the rain factor. Therefore, as these error sources become larger, the accuracy of the rain attenuation estimate, \( L_{RB} \), decreases.
The key to this scheme is its ability to predict the rain fade at the uplink frequency, $L_{RU1}$, based on measurements of attenuation on the downlink frequency, $L_{RB}$. It is the constant $K_{BU}$ that performs the conversion from beacon frequencies to the uplink frequency, as shown on the right side of Figure 6.1-1. According to most references, an adequate relation between the two rain fades (in dB) is just the square of the frequencies involved (in GHz). A more complicated relation is suggested by the Consultative Committee on International Radio (CCIR), as mentioned in Chapter 4. In the case of Globalstar, the 5.15 GHz rain fade is being estimated based on measurements of the 6.88 GHz beacon data, so this leads to a scaling factor, $K_{BU} = 0.56$ or 0.608, depending on the method chosen. The accuracy and reliability of these relations at Globalstar frequencies has not been fully investigated.

### 6.1.3 Error Analysis

As determined in the description of the system model, the accuracy of the uplink rain fade estimate, $L_{RU1}$, is largely dependent on the size of the other system errors, and on the reliability of $K_{BU}$. The best way to analyze the potential benefit of this control strategy is then to quantify the other error sources.

For the purpose of simplification, it is assumed that the estimated delays, $D_{B1}$ and $D_{B2}$, are well known, and equal to the real delays $D_{B1}$ and $D_{B2}$. It is also assumed that the scaling factor, $K_{BU}$, calculated above is a reliable estimate of reality. In relaxing these assumptions, the rain fade estimate $L_{RU1}$ becomes less reliable, so the error bound which follows is not a maximum bound. The error sources mentioned in the above system description are found in Table 6.1-1 on page 55 along with current estimates of their ranges.

Adding the error figures in the table, the rain downlink fade estimate, $L_{RB}$, will include an error interval of +/- 1.67 dB. After applying the scaling factor $K_{BU} = 0.608$, the uplink rain fade has an accuracy of +/- 0.93 dB. Also, it should be noted that this estimate becomes available after a delay of $D_{B1} + D_{B2}$ seconds.
Table 6.1-1: Error Analysis of Beacon Strategy

<table>
<thead>
<tr>
<th>Error</th>
<th>Range</th>
<th>Brief Explanation</th>
</tr>
</thead>
<tbody>
<tr>
<td>$E_{PMB}$</td>
<td>+/- 0.5 dB</td>
<td>The power control loop which keeps the absolute beacon power relatively constant is only as accurate as the power meter it uses. Meters used for forward link power management in the gateway tend to have a +/- 0.5 dB slow moving offset. The power meter on the satellite is probably not any better.</td>
</tr>
<tr>
<td>$E_{BTx}$</td>
<td>+/- 0.7 dB</td>
<td>The antenna pattern of the satellite’s C-Band antenna is estimated based on measurements of a sample set. Unit to unit variation in these antennas can cause +/- 0.2 dB error. The antenna is assumed to be rotationally symmetric about the yaw axis; however, in practice, this is not always the case. Scattering of the antenna pattern due to other physical entities aboard the satellite can cause +/- 0.5 dB variation from the ideal constant circular pattern. The satellite’s yaw position will also vary over time, and perhaps even over a single satellite sweep.</td>
</tr>
<tr>
<td>$E_{BRx}$</td>
<td>+/- 0.3 dB</td>
<td>Unit to unit variation in the gateway antennas can cause +/- 0.1 dB. Although the actual satellite position is known, there is a pointing loss associated with the inaccuracy of the gateway antenna position. The maximum pointing error is specified as x degrees, and at x degrees from the center of the gateway antenna pattern, the gain drops to -0.4 dB. The error in the antenna gain estimate is then +/- 0.2 dB.</td>
</tr>
<tr>
<td>$E_{AGC}$</td>
<td>+/- 0.27 dB or +/- 0.17 dB</td>
<td>The Automatic Gain Control compensates for input powers of 40dBm to 100dBm with a resolution of 8 bits. This translates to 0.23 dB per LSB, so the quantization error is +/- 0.12 dB. Other errors in this method of estimating input power can be attributed +/- 0.15 dB. Since the beacon power envelope does not vary as much as that of a traffic channel, the dynamic range of the AGC can be modified to 9dBm instead of 60dBm, thus decreasing quantization error to +/- 0.02 dB.</td>
</tr>
<tr>
<td>$E_{LPB}$</td>
<td>none</td>
<td>In general, this error is small because the distance to the satellite is well-known and well-controlled. For that reason, this error can be lumped in as part of the disturbance to be estimated – the rain fade itself. If the estimate, $L_{RB}$, of downlink rain attenuation is off because of $E_{LPB}$, the estimate, $L_{RU}$, of the uplink fade, may be off in a correlated way, so the errors cancel.</td>
</tr>
</tbody>
</table>

1. The error estimates in this table were derived from personal communication with Steven Mollenkopf of the Power Management Group at QUALCOMM Incorporated.

6.1.4 Strategy Evaluation

It has been shown that by this scheme, a potential rain fade can be estimated to an accuracy of only +/- 0.93 dB with some delay. After relaxing the assumptions made about estimated delays and the frequency scaling factor, even more uncertainty invades the rain fade estimate.
The rain attenuation itself is between 0dB and 2dB most of the time, as shown in Chapter 4, so the above estimation accuracy does not provide much performance gain for the system. Also, this strategy attempts to compensate for a very specific error caused by rain, so it cannot help to reduce the effect of other uncertainties in the system, such as the largely varying transponder gain. Referring to Table 6.2-1, the total open-loop uplink error range is +/- 4 dB, which is the sum of $E_{TR}$, $E_{UTx}$, $E_{Lpu}$ and $E_{URx}$ in addition to the uplink rain fade, $L_{RU}$. Using the beacon strategy, only the last error is marginally alleviated, so the remaining uncertainty is +/- 3.86 dB. Clearly, this strategy is not adequate in the current context for controlling the satellite power output devoted to Globalstar overhead channels.

The concept of using the beacon signal to control uplink power can be very useful in slightly different contexts, however. One of the required conditions is the threat of much larger rain fades. Usually, large fades can be fatal to the communications task at hand. If these fades are also quickly changing, the beacon strategy has the advantage of providing fast compensation, because the delay $D_{B1} + D_{B2}$ is not very large, and the latter term is somewhat controllable. The inherent error in the strategy is also dwarfed by the benefit to system reliability.

For the beacon scheme to be useful, the rain attenuation should also be very significant when compared to other uncertainties in the path, because otherwise, as in this case, another strategy might be suited to compensate for both.

If the beacon scheme is chosen as the fundamental strategy, there are a number of ways to improve and extend its performance. As an extension of the rain fade estimation idea, if power measurements are available at points along the forward link such as on the satellite or at the receiving phone, each can spawn another estimate of the rain fade. Some combination of all these estimates should be better than any one alone, especially given the uncertain nature of the frequency scaling factor. However, if all the possible power measurements are utilized to derive rain fade estimates, the strategy becomes quite complicated.
6.1.5 Conditions for Reconsideration

As a final statement about this strategy, certain conditions should be satisfied before the beacon strategy is reconsidered for Globalstar overhead channel power control. The first possible condition is if the rain attenuation experienced once Globalstar is operating turns out to be much greater than predicted by the attenuation models (> 5 dB). The same attenuation increase can result from an increase in operating frequency of the uplink. The second condition is if the transponder gain and other forward link uncertainties turn out to be much more predictable, so that the rain fade is proportionally more important. In addition to these conditions, the reliability of the frequency scaling constant needs to be verified in the Globalstar context before it is used.

6.2 Golden Phone Strategy

Since the beacon strategy has been found inadequate, the next candidate needs to be chosen. The telemetry scheme is straightforward, but depends critically on the accuracy of the measuring equipment aboard the satellite -- something that is unknown and unchangeable in the scope of this project. Also, the current patent disclosure which introduces the multiple PSMM scheme requires the existence of a well-calibrated and well-positioned "golden phone". For the above reasons, and because of its relative ease of implementation, the golden phone scheme is chosen as the next candidate.

6.2.1 Description

This strategy places a special user terminal dubbed the "golden phone" at the Gateway in question. The phone continuously measures the downlink pilot power received from the satellite in the S-Band. Knowing the received power as well as the various system gains along the downlink path, the system then estimates the transmitted pilot power at the output of the solid-state power amplifier (SSPA) on-board the satellite. Since the power of this signal is to be controlled, it is compared to the desired pilot power specified by the GOCC. This comparison generates an error signal which is used as the basis of a closed-loop control system. The stability and performance of this loop depend on the controller design, the system delays, and the dynamics, if any, inherent in the current system. In contrast to the beacon scheme, the essential goal of this feedback loop is to control the satellite output power directly, rather than first estimate the uplink rain atten-
Because of its closed loop nature, this system should be more robust in the face of disturbances, and more flexible in its design options.

### 6.2.2 System Model

Figure 6.2-1 shows a simple system diagram for this scheme, which includes possible noise sources and uncertainties. The working unit of the signal arcs is power in decibels, or dBW. For this reason, the multiplicative noise sources and uncertainties can be expressed as additive influences on the system, and their conglomerate effect can be analyzed using superposition. Continuing the convention presented in the Beacon scheme Section 6.1.2, each possible disturbance to the system is separated into the expected value, which is deterministically known to the system, and the error ratio, which captures the real random variable nature of the input. By the labeling convention defined under the picture, the estimated value of any gain is italicized, while the real gain is in plain-face. The error ratio associated with the discrepancy between the two is labeled by an "E" with the same subscript as the gain in question.

Starting at the top-left corner of Figure 6.2-1, the main input, $P_{\text{pilot}}$, represents the power of the pilot signal which is released by the Gateway gain controller unit in the absence of this new “golden phone” system. This power level is set by the Gateway Forward-Link Power Management feature, which also aims to achieve a constant signal power at the output of the satellite, and which uses an open-loop method to pre-compensate for the expected uplink gains. The pre-compensated factors include the gain of the gateway transmitting antenna, $G_{\text{UTx}}$, the path loss resulting from radiation over the distance to the satellite, $L_{PU}$, the gain of the receiving antenna on the satellite, $G_{\text{URx}}$, and finally, the estimate of the satellite transponder gain, $G_{TR}$, which is provided in the RAI. Because these factors are pre-compensated, the only part of them that can potentially disturb the output are the various error ratio terms that are shown on the forward path of the diagram.

The pilot is first radiated by the transmitting antenna of the gateway, which has a gain, $G_{\text{UTx}}$, that is applied to the signal. The signal then normally degrades in power on the path to the satellite ($L_{PU}$) and is potentially faded by rain ($L_{RU}$). The propagation time to the satellite is represented by $D_{UI}$, which is, at most, 6.2 ms.
Figure 6.2-1: Golden Phone Simple System Diagram (working unit: dBW)

Labeling Convention:
- \( G_{\text{sub}} = G_{\text{sub}} + E_{\text{sub}} \): real gain = estimated gain + error in estimated gain
- \( L_{\text{sub}} = L_{\text{sub}} + E_{\text{sub}} \): real loss = estimated loss + error in estimated loss
- \( P_{\text{sub}}, P_{\text{sub}a}, E_{\text{Am}} \): actual, estimated / expected / predicted, measured, required power (dB)

Variables:
- \( D_{\text{UI}} \): delay of uplink power at satellite due to transmission time
- \( D_{\text{Di}} \): delay of received downlink power due to transmission time
- \( D_{\text{D2}} \): delay of received downlink power measurement due to measurement time
- \( D_{\text{DI}}, D_{\text{D2}} \): estimated total downlink delay
- \( E_{\text{DTX}} \): error in estimated gain of the S-Band downlink transmitting antenna on satellite
- \( E_{\text{LPD}} \): error in estimated path loss of the downlink signal (includes downlink rain fade)
- \( E_{\text{PMP}} \): error of the golden phone's power meter in estimating downlink reception
- \( G_{\text{UTX}} \): gain of the uplink transmitting antenna at the gateway
- \( G_{\text{URX}} \): gain of the uplink receiving antenna at the satellite
- \( G_{\text{DTRx}} \): gain of the downlink transmitting antenna on the satellite
- \( G_{\text{DRx}} \): gain of the downlink receiving antenna at the gateway
- \( G_{\text{OFFSET}} \): gain adjustment factor as a result of this control loop
- \( L_{\text{PD}} \): path loss of the downlink signal
- \( L_{\text{PU}} \): path loss of the uplink signal
- \( L_{\text{RU}} \): rain attenuation on the uplink
- \( L_{\text{RD}} \): rain attenuation on the downlink
- \( P_{\text{pilot}} \): unaltered pilot power that would result after Gain Control Unit (GCU)
- \( P_{\text{DTX}} \): downlink pilot power transmitted from satellite (before antenna)
- \( P_{\text{DRx}} \): downlink pilot power received at gateway
- \( P_{\text{SPEC}} \): desired downlink pilot power transmitted from satellite (command signal)
- \( \text{ERROR} \): difference between estimated and desired values of \( P_{\text{DTx}} \)
The receiving antenna at the satellite provides some gain, $G_{URx}$, then passes the signal to the satellite transponder which downconverts from C-band (5 GHz) to S-band (2.5 GHz) and applies a gain of $G_{TR}$. The resulting signal is the pilot power transmitted by the satellite, $P_{DTx}$, which is to be controlled.

Moving to the downlink, the pilot is radiated by one of the sixteen antennas, and amplified by a gain, $G_{DTx}$, which is elevation dependent, as implied by the honeycomb beam pattern shown in Figure 2.3-1. The pilot travels the same atmospheric path as the uplink, but at a lower frequency, so the rain attenuation, $L_{RD}$, and the path loss, $L_{PD}$, are possibly related to their uplink counterparts. The pilot then finds the gain of the receiving antenna, $G_{DRx}$, which is the antenna on the golden phone itself. The propagation time back to the ground, $D_{D1}$, is about the same as $D_{U1}$, but the pilot power measurement time, $D_{D2}$, is much larger.

Once the received downlink power measurement, $P_{DRx}$, has been calculated by the gateway equipment, the system enters the digital domain, and all calculations are done in the gain controller unit (GCU). The estimated value of the downlink transmit power, $P_{DTx}$, is inferred from the measurement $P_{DRx}$ by subtracting out the expected factors encountered on the downlink path. The delay factors are shown in the post-correction activity to represent the efforts of properly aligning the estimated gains profiles in time with those of reality.

The measured and estimated downlink transmit power value, $P_{DTx}$, is then compared to the required power output of the satellite, $P_{SPEC}$, in order to generate an error signal. This error is fed to a digital controller, $H$, which determines the appropriate value of $G_{OFFSET}$, and consequently changes the pilot power transmitted by the Gateway in the next time step. The treatment of this controller as linear and time-invariant (LTI) is appropriate even with the logarithmic working unit because the input and output use consistent units, and because the controller is implemented as a set of difference equations in a microprocessor rather than some physical analog hardware that might otherwise inherently work only on a linear scale.

The application of the control effort to the uplink signal completes the closed loop of this system. From the structure of this loop, it is likely that the controller transfer function $H(z)$ dominates the dynamics of the
response, and the delay factors will limit the performance of this response. The command to be followed by the output is shown in the diagram as \( P_{\text{SPEC}} \) in the feedback path. The major disturbances -- rain attenuation and transponder gain -- are in the forward path. Since all these signals are of relatively low frequency, the controller should, in general, act like an integrator so that the command signal has good transmission at low frequencies while the disturbances suffer major attenuation.

### 6.2.3 Error Analysis

In a procedure similar to the one employed with the beacon strategy, the performance of this control mechanism is largely judged by the error that is expected in pilot power at the output of the satellite transponder. The error analysis is performed with the assumption that \(|GH| > 1\), so that the controller chosen passes all the error signals in the feedback path but adequately attenuates all error signals in the forward path, as shown in Figure 6.2-2.

![Figure 6.2-2: Golden Phone Error Analysis Diagram](image)

Taking values from Table 6.2-1, the total expected error at the output is taken as the sum of the error intervals on the feedback path, which includes \( E_{\text{DTX}}, E_{\text{LPD}}, L_{\text{RD}}, E_{\text{DR}}, \) and \( E_{\text{PMP}} \). This analysis shows an error range of \(+/- 1.3\) dB at the output. Also, the time to compensate for a step input is going to be at least \( D_{U1} + D_{D1} + D_{D2} \), which is much longer than the expected delay for the beacon strategy. The total error figure, and specifically \( E_{\text{PMP}} \), may be reduced by changing the gain of the RSSI filter in the golden phone, \( G \), with some performance impact that is addressed in Chapter 7.
### Table 6.2-1: Error Analysis of Golden Phone Strategy

<table>
<thead>
<tr>
<th>Error</th>
<th>Range</th>
<th>Brief Explanation</th>
</tr>
</thead>
<tbody>
<tr>
<td>ET_R</td>
<td>+/- 2 dB</td>
<td>The gain of the transponder changes over time, and is less predictable than originally thought. Responding to environmental changes such as temperature and traffic load level, the gain can drift away from the expected value by up to 3 dB per minute, and reach +/- 2 dB.</td>
</tr>
<tr>
<td>L_RU</td>
<td>+/- 1 dB</td>
<td>The rain attenuation at 5 GHz can range from 0 dB to 2 dB as shown in Chapter 4.</td>
</tr>
<tr>
<td>E_U_Tx</td>
<td>+/- 0.3 dB</td>
<td>The transmitting antenna at the gateway is subject to pointing loss due to inaccuracy in the drivers that track the satellite. The specified limit of (x) degrees of error places the uplink path at the -0.4 dB point of the antenna gain pattern compared to the peak at 0 dB. The antenna gain thus varies by +/- 0.2 dB. There is also a +/- 0.1 dB unit-to-unit variation among gateway antennas.</td>
</tr>
<tr>
<td>E_L_PU</td>
<td>negligible</td>
<td>The ephemeris of the satellite is well known, and the random altitude of a gateway over sea level has little effect on the distance to the satellite.</td>
</tr>
<tr>
<td>E_U_Rx</td>
<td>+/- 0.7 dB</td>
<td>The antenna pattern of the satellite’s C-Band antenna is estimated based on measurements of a sample set. Unit to unit variation in these antennas can cause +/- 0.2 dB error. This antenna is assumed to be rotationally symmetric about the yaw axis; however, in practice, this is not always the case. Scattering of the antenna pattern due to other physical entities aboard the satellite can cause +/- 0.5 dB variation from the ideal constant circular pattern.</td>
</tr>
<tr>
<td>E_D_Tx</td>
<td>+/- 0.25 dB</td>
<td>Because of its complexity (16 beams) and non-symmetric nature, the S-Band antenna pattern for each satellite is recorded before launch, and the resulting data provided to the Gateway before satellite contact by the GOCC (in the RAI). The record shows the gain at a grid of points in the far field. Unit-to-unit and scattering errors are caught by the pre-launch record, but aging effects and interpolation cause +/- 0.25 dB of error.</td>
</tr>
<tr>
<td>E_L_PD</td>
<td>negligible</td>
<td>The ephemeris of the satellite is well known, and the random altitude of a gateway over sea level has little effect on the distance to the satellite. The downlink rain attenuation can be considered a component of this error.</td>
</tr>
<tr>
<td>L_R_D</td>
<td>+/- 0.25 dB</td>
<td>Using the frequency scaling formula, at 2.49 GHz, a +/- 1 dB fade at 5.15 GHz becomes slightly less than +/- 0.233 dB or +/- 0.287 dB depending on the method chosen</td>
</tr>
<tr>
<td>E_D_RX</td>
<td>+/- 0.3 dB</td>
<td>Same argument as for E_U_Tx and E_B_RX.</td>
</tr>
<tr>
<td>E_P_MP</td>
<td>+/- 0.5 dB</td>
<td>The error in the power meter of the phone is variable, and described in the RSSI Filter design. The error is actually +/- 1 dB, at the current gain setting, but that is reducible. Also, the variation is largely high frequency (uncorrelated every sample), so much of the error is attenuated by the closed loop control.</td>
</tr>
</tbody>
</table>

1. The error estimates in this table were derived from personal communication with Steven Mollenkopf of the Power Management Group at QUALCOMM Incorporated.
6.2.4 Strategy Evaluation

At first sight, these performance metrics are not too flattering of the golden phone strategy. However, upon a more complete survey, it becomes apparent that the open-loop uplink path error is +/- 4dB, so a total reduction of 5.4 dB in uncertainty has been achieved by the system. If uncertainty is mapped strictly to a reduction in system capacity using the methods of Section 3.4, this translates to a savings of up to 70% of the system user capacity.

Another attractive feature of this strategy compared to the beacon scheme is the lack of an influential and questionable frequency scaling factor. For completeness, there is still a potential role in this scheme for the scaling factor, because the downlink rain fade could be related to that of the uplink. However, since the purpose of this system is no longer to estimate the rain fade, it is of little concern. Fortunately, the Globalstar downlink frequency (2.5 GHz) is relatively low, so the resulting rain attenuation, $L_{RD}$, is very small.

Based on the above analysis, the golden phone strategy has the potential to eliminate much of the uncertainty that affects the pilot channel between the gateway and the output of the satellite. It may also compensate for unknown fades that are not yet identified. The only major disadvantage to this strategy is the possibility that the gateway is not included in its own coverage area, so that the downlink measurement becomes unavailable. Eventually, this condition may be eliminated by a higher level system, or compensated for by a backup system like the PSMM strategy that works to cover “shadow” periods. In the worst case, the current TPTL open-loop system can be employed alone. Whatever the decision with regard to this shortcoming, the golden phone is chosen as the fundamental strategy for overhead channel power control, and is further investigated in the next three chapters.
Chapter 7: Initial Controller Design

The “golden phone” strategy has the potential to control the output power on the satellite downlink more accurately than the open loop system currently in place. In order to make this feasibility a reality, however, the appropriate controller, H, must be found to provide an adequate response time, a small enough overshoot, zero steady-state error, and very good disturbance rejection. Finding this controller is the purpose of the next two chapters.

In order to make use of popular analog design techniques, the system is initially assumed to contain solely continuous-time signals. This assumption introduces the task of conveniently modeling pure time delays, which are present in this system. The commonly used Pade approximation, as described in Section 7.2, provides an adequate delay representation in the frequency domain, but the resulting time response is not ideal. For this reason, the frequency-domain analysis is done in continuous time, but time-domain performance is analyzed in discrete-time. The initial design of the frequency response is presented in this chapter without addressing the details of discrete-time conversion. The result is the emergence of two parameters that describe the sample space of reasonable controllers, as will become evident by the end of Chapter 7.

Following this development the issue of converting a controller to discrete-time is addressed in Chapter 8. The final design selection is then made by collecting time-domain performance statistics over the sample space, and considering the requirements presented in Section 7.3.

7.1 System Model

The basic golden phone system is quite well described by the diagram in Figure 7.1-1, which is reprinted from Section 6.2. Recall that signal arcs in the main loop represent the power of the overhead channels at different stages of the forward link. For simplicity and without loss of generality, the only overhead channel considered in operation is the pilot channel.

As shown in the diagram, there are a number of multiplicative factors that change the pilot power as it travels from the GMOD to the golden phone. Since many of these are naturally specified on a decibel scale, it
becomes convenient to do the same with pilot power. The units are then logarithmic (dBW), so factors that are linearly multiplicative become additive, as shown in the diagram.

By superposition, the output, $P_{DTx}$, is the sum of all the input signals multiplied by their respective transfer functions. In preparation for presenting the design, it is convenient to enumerate these transfer functions for the significant system inputs. The system shown in Figure 7.1-1 includes all the factors that may affect the output pilot power, and the point in the loop at which they arrive, so the transfer functions are derived therefrom, and listed in Table 7.1-1. The expected uplink gains -- $G_{UTx}$, $G_{TR}$, $L_{PU}$ and $G_{URx}$ -- are pre-compensated by the gateway; therefore, each input enters at two different locations, and there are two terms which differ by a delay in the transfer function $Y(s)/C(s)$. Similarly, the expected values of the downlink factors --
If the delays of downlink propagation and power measurement are well-known and predictable -- that is, $D_{D1} + D_{D2} = D_{D1} + D_{D2}$ -- this transfer function goes to zero, and the post-correction is successful. The transfer function becomes more significant to the output as the estimated delays become more incorrect.

Figure 7.1-2 greatly simplifies matters by assuming that the pre-correction and post-correction activities are synchronized with reality, and perfectly compensate for the expected path gains. There are only two significantly different positions at which an input can enter the simplified loop -- location A, the forward control path (uplink), and location B, the feedback control path (downlink). The corresponding transfer functions are listed in Table 7.1-1, and all those from other positions differ by at most a delay term.

<table>
<thead>
<tr>
<th>Table 7.1-1: Relevant Transfer Functions</th>
</tr>
</thead>
<tbody>
<tr>
<td>Loop Transfer Function:</td>
</tr>
<tr>
<td>$L(s) = e^{-sD_{U1}} e^{-sD_{D1}} e^{-sD_{D2}} H(s)G(s) = e^{-sD_{L}} H(s)G(s)$</td>
</tr>
<tr>
<td>Forward Path: $\frac{Y(s)}{A(s)} = \frac{1}{1 + L(s)}$ $Y(s) = P_{DTx}(s)$</td>
</tr>
<tr>
<td>Feedback Path: $\frac{Y(s)}{B(s)} = e^{-sD_{D1}} \left( \frac{H(s)G(s)}{1 + L(s)} \right)$</td>
</tr>
<tr>
<td>Pre-Corrected Factors: $\frac{Y(s)}{C(s)} = (e^{-sD_{D}} - e^{-sD_{U}})\left( \frac{1}{1 + L(s)} \right)$</td>
</tr>
<tr>
<td>Post-Corrected Factors: $\frac{Y(s)}{D(s)} = (e^{-sD_{D}} - e^{-sD_{D}})\left( \frac{H(s)G(s)}{1 + L(s)} \right)$</td>
</tr>
</tbody>
</table>
The forward path transfer function applies to the varying power of the pilot signal which emerges from the GMOD ASIC. It also applies to the uplink rain attenuation factor, L_{RU}, and the error in the estimated transponder gain, E_{GTR}, which are the two major disturbances in this system.

The feedback path transfer function applies to the random variable parts of the S-Band antenna gain, E_{DTX}; the downlink path loss, E_{LPD}; the downlink rain attenuation, L_{RD}; and the gateway receiving antenna, E_{DRX}. It also applies to the command input, P_{SPEC}, which is the desired output power.

### 7.2 Plant Model

The first step in designing the closed-loop controller is to understand the system it is attempting to control -- the plant. Looking at the loop transfer function in Table 7.1-1, the plant dynamics consist of a number of time delays, and the response of the power meter, G(s), at the receiving end of the downlink. The uplink and downlink propagation delays, D_{U1} and D_{D1}, are essentially equal and reach a maximum of 6.2 ms at the minimum contact elevation of 10 degrees.

#### 7.2.1 Power Meter Response

The power meter shown is the same as the one that calculates the Received Signal Strength Indicator (RSSI) for use in the PSMM. The RSSI filter provides a decibel measure of the received pilot channel signal-to-noise ratio, and it has a response similar to a single-pole filter. The time constant is nominally set at 26.45 ms for a regular phone, but is changeable by specifying a different filter gain. As shown in Table 7.2-1 and Figure 7.2-1, increasing the filter gain, G, results in a faster time constant, but a higher noise figure in the power measurement.

In the error analysis of Section 6.2.3, it is given that \( E_{PMP} \) is approximately +/- 0.5dB, even though the filter gain of \( G = 8 \) corresponding to \( \tau = 26.45 \) ms ordinarily maps to an error of +/- 1 dB. Basically, the RSSI filter is assumed to have a smaller gain, \( G = 4 \), in order to support the lower noise figure quoted. This new operating point has a slower time constant (\( t = 50 \) ms) but a better noise figure of +/- 0.75 dB. The lower number of 0.5 dB is justified because the golden phone is also expected to perform better than a standard phone by pro-
\[
P_{\text{DRx}} \rightarrow \frac{G(s)}{\tau s + 1} \rightarrow y = P_{\text{DRx}}
\]

RSSI Filter

\[
\tau = \frac{-T_s}{\ln(1 - G/2^9)} \quad E[y] = P_{\text{DRx}} \quad \sigma_y^2 = \frac{2(P_{\text{DRx}} + 1)}{(2^{10}/G - 1)}
\]

\(G = \text{RSSI Filter Gain}, \quad T_s = 512/(1.23 \, \text{MHz})\)

Table 7.2-1: RSSI Filter Statistics

<table>
<thead>
<tr>
<th>Gain</th>
<th>Time Constant (s)</th>
<th>Output Variation (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>-19.5</td>
<td></td>
<td></td>
</tr>
<tr>
<td>-20</td>
<td></td>
<td></td>
</tr>
<tr>
<td>-20.5</td>
<td></td>
<td></td>
</tr>
<tr>
<td>-21</td>
<td></td>
<td></td>
</tr>
<tr>
<td>-21.5</td>
<td></td>
<td></td>
</tr>
<tr>
<td>-22</td>
<td></td>
<td></td>
</tr>
<tr>
<td>-22.5</td>
<td></td>
<td></td>
</tr>
<tr>
<td>-23</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Figure 7.2-1: RSSI Filter Performance

a) Time constant, \(\tau\), versus filter gain, \(G\).
b) Noise Figure of output given input of -21 dB Signal to Noise

Providing lower ambient noise at reception. The implicit trade-off caused by the RSSI filter is between the closed-loop bandwidth and the output noise figure.

Another significant piece of the power meter response is the delay in communicating this RSSI measurement to the gain control unit (GCU) where the control strategy is implemented. This delay is not fixed, so if necessary, a maximum communication delay can be specified, and the gateway software designers will make sure that the specification is met. For now, it is safely assumed that the power measurement, once settled after about 150 ms, is not available instantaneously, but rather at the next sampling instant of the discrete controller, \(H(z)\). The actual choice of sampling rate is addressed in Chapter 8.
Finally, once the control effort, $G_{\text{OFFSET}}$, is calculated by the compensator, $H(z)$, there may be some small delay before this gain is applied to the outgoing pilot. This delay is also changeable, but is on the order of nanoseconds, which is essentially instantaneous for this system.

The delay structure of the system is summarized by the timing diagram of Figure 7.2-1.

![Timing Diagram](image)

**Figure 7.2-1: Timing Diagram**

"x" is the propagation delay (12ms)

### 7.2.2 Simplifying Power Meter Response

The transfer function of the power meter filter is basically a single-pole low pass filter with a nominal time constant of about 50 ms, and therefore a bandwidth of approximately 20 rad/s. It is assumed that this bandwidth is large compared to the overall control loop to be designed, so the power meter frequency response can be considered equal to unity. However, the controller may wait for the power measurement to settle before the output value can be trusted, and this takes approximately three time constants to happen. For the purposes of design, the RSSI filter is modeled as straight delay term of about 150 ms. This is a conservative estimate because it implies that no feedback is available until 150 ms have passed. The modeling assumption is eventually relaxed during the simulation phase.

With this simplification, the entire plant can be considered a pure delay. Referring to the timing diagram in Figure 7.2-1, and assuming a convenient sampling rate of 8 Hz, the total loop delay, $D_L$, is taken to be 250 ms. The design that follows assumes this delay value, but the MATLAB scripts that are provided in the
Appendix on page 116 can conveniently re-apply the design process for a different delay time or sampling rate.

### 7.2.3 Modeling the Delay Factors

As mentioned towards the beginning of this chapter, one of the hurdles in establishing a continuous-time model of this system is finding a good representation for the pure delay term. Many of the design techniques for analog controllers, especially in MATLAB, work with transfer functions that are rationals in $s$, the complex frequency-domain parameter. However, the Laplace transform of a delay, $T$, is an exponential, $e^{sT}$. In order to express the delay as a polynomial, one can use the Pade approximation.

\[
Pade(s) = \frac{\left[ 2 - Ts + \frac{(-Ts)^2}{2!} + \frac{(-Ts)^3}{3!} + \ldots + \frac{(-Ts)^n}{n!} \right]}{\left[ 2 + Ts + \frac{(Ts)^2}{2!} + \frac{(Ts)^3}{3!} + \ldots + \frac{(Ts)^n}{n!} \right]} \quad \text{for order } n
\]

![Figure 7.2-1: Pade Approximations of Varying Order](image)

**Figure 7.2-1: Pade Approximations of Varying Order**

a) (left) Delay of 1 second  
b) (right) Delay of 25 ms

The plots also include dashed lines showing the phase of the exponential being approximated and lines of constant phase at -90 degrees and -180 degrees.

The higher the order of the Pade approximation, the more accurately the phase response of the resulting expression matches that of an exponential. The phase curves for Pade polynomials of varying order are shown in Figure 7.2-1, along with dashed lines showing the true exponential being approximated, and
dashed lines showing the constant phase at -90 and -180 degrees. Two plots are provided to show that accuracy does not depend on the value of the delay. It should be noted that the phase of the exponential is really linear, but it looks exponential on a logarithmic frequency scale.

The phase response of the loop transfer function, \( L(s) \), is most important near the crossover frequency. For stability reasons, the phase response at crossover is usually no less than -180 degrees, so it is particularly important that the Pade function represent the phase of the delay accurately up to at least -180 degrees. A magnified version of the Pade response is shown in Figure 7.2-2a, and the approximations of orders 1, 2 and 3 are visible. An order of two might be adequate up to -180 degrees, but a order three is preferred. Further investigation reveals that if the Pade order is increased indiscriminately, the complexity of the system increases until the numerous zeros and poles added by the Pade actually make the entire system unstable.

The phase response emulation of the delay is satisfactory with a Pade approximation of order three, but in terms of the time-domain step response, the Pade does not perform very cleanly, as shown in Figure 7.2-2b. In fact, as the order is increased, the initial response becomes more oscillatory, and there is some concern that the approximation may distort the perceived time-dynamics of the overall system. The conclusion is that Pade works for frequency-domain design, but when time-domain performance is analyzed, the Pade approx-

![Figure 7.2-2: Pade Approximation](image)

a) (left) Close-up of phase response near -180 degrees.
b) (right) Step response of various order Pade approximations.
imation should be avoided. For time-domain analysis, the system is tested in discrete-time, because unit sample delays have a very clean representation in the z-domain.

7.3 Performance Requirements

The desired performance of this controller is expressed in terms of the regular metrics of goodness which follow. Since the main goal of the controller is to set the output power of the satellite at $P_{SPEC}$, there should be no steady-state error in this activity. This requires the output to follow changes in the command signal, $P_{SPEC}$, but also completely reject changes in the disturbance inputs such as rain and transponder gain error.

The disturbance signals above have the potential to grow as ramps in time. As shown in Chapter 4, the rain attenuation can change by as much as 0.2 dB per second. The transponder gain, on the other hand, is largely dependent on satellite temperature, and the system load level, which both change slowly. Therefore, the change in transponder gain has been estimated not to exceed 3 dB per minute. In addition to these two factors, if any other system gains -- expected or otherwise -- change significantly within a second, they will not be pre-compensated adequately by the open-loop controller operating at 1 Hz. The closed-loop control effort is responsible for filling in the gaps between TPTL sample points.

The TPTL also implicitly introduces small step inputs to the closed-loop system every second. If an unexpected external step input disturbs the system, however, it should be corrected quickly by the closed-loop to prevent any impact on system availability, or large drains on satellite power to which the FLEA is otherwise blind. Correspondingly, the peak overshoot of the controlled response should not be very large, otherwise the FLEA may act to reduce the user capacity based on a violation of the instantaneous power limit.

After consideration of various performance trade-offs, and discussion with the Power Management group at QUALCOMM Incorporated, the tentative design goal requires zero steady-state error in the tracking of a ramp input, and in the rejection of a ramp disturbance. For a unit step input, the output should suffer no more than a 50% overshoot, and should provide a rise time of less than one second to achieve 0.1 dB of the final value. Finally, the settling time of both the ramp response and step response should be less than five seconds.
Given these minimum time response requirements, the amount of in-band noise or error should be minimized.

<table>
<thead>
<tr>
<th>Type</th>
<th>Description of Requirement</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Goal</td>
<td>Error or noise in the output signal, $P_{D_Tx}$</td>
<td>minimized</td>
</tr>
<tr>
<td></td>
<td>Associated: Maximize Bandwidth of System</td>
<td></td>
</tr>
<tr>
<td>Constraint</td>
<td>Steady-state Error to Ramp Input</td>
<td>0</td>
</tr>
<tr>
<td>Constraint</td>
<td>Peak overshoot of Time Response</td>
<td>&lt; 50%</td>
</tr>
<tr>
<td>Constraint</td>
<td>Rise Time to within 0.1 dB of Final Value</td>
<td>&lt; 1 sec</td>
</tr>
<tr>
<td>Constraint</td>
<td>Settling Time to within 0.1 dB of Final Value</td>
<td>&lt; 5 sec</td>
</tr>
<tr>
<td>Goal</td>
<td>Substantial Delay Robustness</td>
<td>&gt; 400 ms</td>
</tr>
</tbody>
</table>

In addition to these quantifiable specs, the design should be robust to changes in the loop delay time. If one recalls the inverse relationship between the speed of response and the noise rejection of the RSSI filter, noise rejection can be increased at the expense of extra measurement delay, so it is desirable to minimize the impact on performance and stability if such a trade-off is made. Also, if the gateway infrastructure is, for some reason, unable to deliver power measurements in one sample delay, the system performance should not become unstable.

7.4 **Steady-State Error**

Using conventional design techniques, the performance of a closed-loop system is largely explained by the frequency response of the loop transfer function, $L(s)$. This transfer function is the cascade combination of, $H(s)$, the controller to be designed, and plant transfer function. From Section 7.2, the latter includes air propagation delays, the internal communication delay, and the response of the power meter, $G(s)$, which is also simplified to a delay.

7.4.1 **Double Integral**

Since the plant transfer function is modeled as just a delay, $e^{-sT}$, it has unit magnitude at all frequencies. The magnitude response of $L(s)$ is then determined solely by $H(s)$. The goal of the controller is essentially to make the closed-loop transfer function from $P_{SPE}$ to the output equal one in the frequency range where
PSPEC is significant, and the transfer function for disturbances practically equal to zero in the frequency range where the disturbance power is significant. Since all of these signals are slowly changing, it is the low frequency response that is most important. Revisiting the appropriate closed-loop transfer functions in Table 7.1-1, it becomes clear that at low frequencies, $|H(s)|$ should be very large in order to achieve the above goals. Control theory further suggests that double integral control is necessary to reject ramp disturbances with zero steady-state error, as constrained by the performance requirements of Section 7.3. The calculations in Table 7.4-1 confirm this notion.

### Table 7.4-1: Steady-state Error Analysis

<table>
<thead>
<tr>
<th>Controller Type</th>
<th>Closed-loop Transfer Function of Disturbance to Output</th>
<th>Steady-state Error for unit Step Input</th>
<th>Steady-state Error for unit Ramp Input</th>
</tr>
</thead>
<tbody>
<tr>
<td>Proportional</td>
<td>$H(s) \frac{1}{1 + Ke^{-sD_L}}$</td>
<td>$\lim_{s \to 0} \left( \frac{1}{s} \frac{Y(s)}{A(s)} \right)$</td>
<td>$\lim_{s \to 0} \left( \frac{1}{s^2} \frac{Y(s)}{A(s)} \right)$</td>
</tr>
<tr>
<td>Integral</td>
<td>$\frac{K}{s} \frac{1}{1 + Ke^{-sD_L}/s} = \frac{s}{s + Ke^{-sD_L}}$</td>
<td>0</td>
<td>$\frac{1}{K}$</td>
</tr>
<tr>
<td>Double Integral</td>
<td>$\frac{K}{s^2} \frac{1}{1 + Ke^{-sD_L}/s^2} = \frac{s^2}{s^2 + Ke^{-sD_L}}$</td>
<td>0</td>
<td>0</td>
</tr>
</tbody>
</table>

The final value theorem is used to calculate the steady-state error of the closed-loop response to both step and ramp inputs given a proportional controller, an integral controller, or a double integral controller. Indeed, at low frequencies, only the double integral response is sufficient to meet the zero steady-state error condition.

#### 7.4.2 Lead Zero

Recognizing the monotonically decreasing magnitude response of the proposed double integral controller, the concept of phase margin (PM) can be used to analyze the stability of the closed-loop system. Using this tool, it is clearly established that pure double integral control is unstable since the phase response of the loop transfer function, $L(s)$, is always less than -180 degrees at all frequencies.
\[ H(s) = \frac{K}{s^2} \quad \text{(Double Integral)} \]
\[ G(s) = 1 \quad \text{(Power Meter)} \]
\[ L(s) = e^{-sD_{L}} H(s) G(s) \]
\[ \angle L(s) = \angle H(s) + \angle e^{-sD_{L}} = -\pi - \omega D_{L} \leq -180^\circ \]
\[ \angle L(s) + 180^\circ \leq 0^\circ \quad \therefore \text{no PM available} \]

One simple solution for stability is to add a Proportional Derivative (PD) term to the controller, or equivalently add a zero to the transfer function \( H(s) \) at some positive frequency, \( a \) radians, as shown below. The new singularity provides up to 90 degrees of additional phase at crossover frequencies greater than the zero frequency, \( a \), as shown below. It is also confirmed that with the zero, the steady-state response to a ramp disturbance is still exactly zero.

\[ H(s) = \frac{K(s + a)}{s^2} \]

\[ \angle L(s) = \text{atan} \left( \frac{\omega}{a} - \pi - \omega D_{L} \right) = -\frac{\pi}{2} - \omega D_{L} \leq -90^\circ \quad \text{for } \omega \gg a \]

This simple controller design of two poles at the origin of the s-plane, and one real left-half-plane zero, is now the assumed structure of the loop controller, as shown by \( H(s) \) above. Therefore, the gain, \( K \), and the location of this "lead zero", \( a \), are the two remaining degrees of freedom in the design process.

### 7.5 Zero Placement

Expressed in the frequency domain, the requirements at the end of Section 7.3 become constraints on the crossover frequency (bandwidth), and the corresponding phase margin of the loop transfer function, \( L(s) \). These can then be translated into requirements on the gain, \( K \), and the zero location, \( a \). Phase margin provides a measure for the degree of stability of the closed-loop system, and it can be used to loosely indicate the shape of a unit step response. When the crossover frequency is fixed, various phase margins lead to the
step responses shown in Figure 7.5-1, where the more oscillatory responses correspond to the lower phase margins. For the purpose of developing a design technique, a phase margin of 55 degrees is chosen because it provides a reasonably quick and non-oscillatory step response. This choice is temporary, however, and the phase margin is more thoughtfully chosen in Chapter 8.

7.5.1 Max Zero Design

An initial exercise is to determine how the assumed phase margin, 55 degrees, can be achieved in L(s) by correctly placing the lead zero and the gain. The maximum positive phase available from the controller H(s) is 90 degrees, as shown in Section 7.4.2. However, if the zero location, $a$, is too large, this maximum phase is not realized until a frequency at which the delay factor in L(s) has already made 55 degrees of phase margin impossible. The highest frequency zero location is determined theoretically in the simultaneous equations of Table 7.5-1, which essentially set the maximum value of the L(s) phase response to provide the given phase margin, $PM$. Alternatively, the MATLAB script MaxZero.m in the Appendix, iteratively decreases the lead zero location, and looks for the first $a$ that provides the desired phase margin in L(s). Specifically, this function returns the controller gain, $K$, and zero location, $a$, required in the face of a particular plant delay, $mdelay$, and for a desired phase margin, $despm$. The frequency response of the MaxZero controller corresponding to $despm=55$ and $mdelay=250$ms is shown in Figure 7.5-2, and the 55 degrees of phase margin occurs at the peak of the phase response, as expected. The corresponding gain for this design is 1.15, and
Set peak of phase response to provide the desired PM:
\[
\begin{align*}
\text{atan} \frac{\omega}{a} - \omega D_L &= PM \\
\frac{d}{d\omega} PM &= \frac{1}{a} \left[ \frac{1}{1 + (\omega/a)^2} \right] - D_L = 0
\end{align*}
\]

Crossover Frequency
\[
\omega = \sqrt{a(a - 1/D_L)}
\]

Table 7.5-1: MaxZero Design

The zero location is 0.3467 rad/sec. Henceforth, the zero location of a MaxZero design for any given phase margin will be referenced as \(a_o\).

7.5.2 MinZero Design

With a little analysis, it is clear that the MaxZero design provides the lowest possible crossover frequency for the specified phase margin and plant delay. If the zero location, \(a\), is brought closer to the s-plane origin than \(a_o\), then more of the associated phase boost from the controller is available at frequencies, \(\omega\), greater than \(a\). In other words, at the previous MaxZero crossover frequency, reducing the zero location makes more phase margin available, and consequently, permits a higher crossover frequency for the same desired phase margin. The frequency response of the loop transfer function, \(L(s)\), is shown in Figure 7.5-3a as the zero location decreases. In the figure, it should be noted that the peak of the phase response gets higher as the zero gets lower, but the phase margin remains constant at 55 degrees. Also, the bandwidth of the system, which is essentially represented by the crossover frequency, increases as the zero decreases. The relative position of the new zero, \(a\), is recorded as a fraction \(q\) of the original zero location, \(a_o\), from the MaxZero design. The relationship between this fraction and the resulting crossover frequency, \(f_{CP}\), is depicted in Figure 7.5-3b.
The limiting case as the fraction $q$ approaches zero is called the MinZero design. The bandwidth gained as the fraction goes to zero is finite, and this convergence is graphically implied by Figure 7.5-3a. In the MinZero design, the double-integral controller degenerates to a single integral controller with a constant -90 degrees of phase. Thus, in the limiting case, the bandwidth limit is determined by the desired phase margin, and the phase response of the plant, which is governed by the loop delay, $D_L$. The absolute maximum bandwidth for stability occurs when the phase of the plant is exactly -90 degrees, while the maximum bandwidth for 55 degrees of phase margin is somewhat lower. Both cases are indicated in Figure 7.5-4, along with the bandwidth of the MaxZero design for 55 degrees. Of course, no performance gain is free in nature, so as the bandwidth increases, it turns out that time-domain performance and delay robustness are sacrificed. These trade-offs are explored in the following sections.

### 7.6 Delay Robustness

Using phase margin as a measure of stability, the major issue here is whether delay robustness is affected by the choice of zero location, $a=qa_o$, in the spectrum between the MinZero and MaxZero designs.

In order to analyze robustness, the MATLAB script TradeOff.m in the Appendix iteratively assumes some plant delay, and finds the corresponding MaxZero and MinZero designs for the desired phase margin,
This controller design is then used to control a plant of variable delay. As this new plant delay is increased from the assumed value, the phase margin of $L(s)$ is monitored, and the plant delay that first violates some minimum allowable phase margin, $minpm$, is recorded. This analysis determines the maximum plant delay for which acceptable stability can be sustained by each combination of designed plant delay and lead zero placement strategy.

Figure 7.6-1a shows the result of this analysis for $despm = 55$ degrees, and $minpm = 20$ degrees. This graph is useful for determining the delay robustness, and also for showing the effect of a changing plant delay on the loop crossover frequency. If the plant delay assumption in Section 7.2 is ever changed, this graph can quickly indicate the relative performance of another controller designed under the new plant delay assumption.

The ratio of maximum delay to designed delay for the two zero placement strategies is shown in Figure 7.6-1b as the designed phase margin, $despm$, takes various values, while $minpm$ stays constant at 20 degrees. For
Figure 7.6-1: Delay Robustness Analysis

a) Assumed delay and corresponding maximum delay versus crossover frequency

b) Ratio of maximum delay (for 20 deg PM) over designed delay (0.25 s) versus the desired phase margin.

a planned phase margin of 55 degrees, the MaxZero strategy shows a ratio of close to 3.2, and the MinZero a ratio of 2. Therefore, the former is considered more robust to plant delay variations. For the assumed plant delay of 250ms, if a controller is designed for 55 degrees of phase margin using the MaxZero strategy, it is acceptably stable unless the plant delay exceeds 800ms, whereas with the MinZero strategy, it becomes unacceptable with plant delays of more than 500 ms.

In general then, as the fraction \( q \) tends to zero, the bandwidth increases, but delay robustness decreases. A pictorial explanation for this may be seen in the Nichols plot of Figure 7.6-2. For the MaxZero design, the crossover of unit magnitude occurs at a phase peak, so its phase margin is affected minimally under the effect of increased delay. The effect is essentially to shrink the plot vertically.

7.6.1 Higher-Order Designs

The above idea is extended to address briefly the possibility of higher-order controller designs. In assembling a controller with more poles and zeros, the goal is probably to provide the maximum phase boost in the area of crossover. However, as shown in Figure 7.6-1a, crossover frequency strongly depends on the assumed plant delay. If this assumption is reliable, the crossover frequency can be safely raised using a higher-order controller. However, the speed with which the resulting phase response decreases after the
crossover frequency is going to be more steep on a logarithmic scale. If this steepness is translated to the Nichols plot of Figure 7.6-2, the new system would experience a very quick decrease in phase margin for the same change in plant delay as for the MaxZero and MinZero. Thus, the conclusion is that higher-order controllers provide inferior delay robustness to even the MinZero design, and the option is discarded unless it is determined that the added time-domain performance is worth the sacrifice in delay robustness.

7.7 Time-Domain Performance

The time domain performance of a particular feedback system is generally judged by a simulated step response based on the closed-loop transfer function. In the specific case of a second-order system, however, the step response is commonly analyzed directly from the closed-loop transfer function by using the concepts of the damping ratio, $\zeta$, and natural frequency, $\omega_n$. Many systems with order greater than two can also be analyzed by considering only the dominant poles of the system, and then using a second-order approxi-
In the cases where this can be done, the phase margin of the loop transfer function, \( L(s) \), is strongly related to the damping ratio, \( \zeta = \text{PM}/100 \). Consequently, the phase margin can carry much information about the step response of the closed-loop system.

### 7.7.1 Negation of Second-Order Approximation

In the case of the overhead channel power control loop, it would be particularly convenient to use a second-order approximation if possible, since the Padé approximation is not acceptably behaved in the time-domain, as concluded in Section 7.2.3. The controller \( H(s) \) is of second order, but \( L(s) \) is of higher order because the delay term is modeled as a third-order Padé approximation. However, if the closed-loop system, \( Y(s)/B(s) \), has only two dominant poles, an approximation may still be made. A portion of the root locus of \( L(s) \) using the nominal MaxZero controller is shown in Figure 7.7-1a, and magnified near the origin in Figure 7.7-1b. The larger radius poles and zeros in this plot are the result of the plant delay approximation. Initially, the two complex poles near the imaginary axis of the \( s \)-plane appear dominant. However, the pole on the negative real axis near the lead zero is too close to the origin, so it is not overshadowed, and a second-order approximation to the system cannot be made.

This conclusion is somewhat intuitive when the difference between the MaxZero and MinZero design is considered. As noted during its development, the MinZero design is essentially a single integral, or a domi-
nant pole compensator. In that case, it would be expected that the MinZero response to a ramp input should not be the same as that of the MaxZero, even though the achieved phase margin is the same. Thus, if the zero location, $a$, of the MaxZero design is lowered, it follows that the shape of the time-domain response may also be affected.

### 7.7.2 Performance Analysis

As a result of the above conclusion, it is necessary to conduct a full time-domain analysis of the various controller designs. The purpose of this analysis is to decide on the best controller for the performance requirements of Table 7.3-1. In section 7.4, it became clear that the two control parameters left to be decided are the gain, $K$, and the zero location, $a$, of $H(s)$. The range of possible values assumed by these parameters comprises one sample space of possible controllers. In Section 7.5, this sample space was transformed so that it can be alternately described by the designed phase margin, $despm$, and the same zero location. The latter is more conveniently described as a fraction $q$ of the zero location, $a_w$, found by the MaxZero strategy given $despm$. The time-domain analysis in Chapter 8 explores this sample space of controllers, and monitors the speed and shape of the step and ramp responses that are achieved.
Chapter 8: Controller Evaluation and Selection

In order to select the appropriate controller for the golden phone strategy, the various trade-offs between stability, robustness, and speed of response must be well understood. Through Chapter 7, the form of the proposed controller, $H(s)$, has been established, but the specific values of the gain $K$ and the zero location $a$ are yet to be determined. The ranges of these parameters are explored in this chapter in order to ascertain their effect on the performance of a corresponding controller.

Although the pilot signal in a Globalstar channel is inherently continuous-time when sent through the air, the power of that signal is itself seen as a discrete-time signal by the controller. This follows from the fact that the controller is implemented as a difference equation in the microprocessor of the GCU. For this reason, and because of the inaccuracy of the Pade representation of the delay in the time-domain, the continuous-time controller is converted in Section 8.1 to a discrete-time equivalent with transfer function denoted by $H(z)$. Once converted, the time-domain performance can be accurately analyzed, as in Section 8.2. Given the performance constraints of Table 7.3-1, an appropriate controller is selected in Section 8.3, and this design is then simulated in Chapter 9.

8.1 Conversion to a Discrete-Time Controller

In a time when the versatility of digital computing equipment is hard to overlook, many system controllers are being transformed to discrete-time equivalents, so there is good documentation of the various transformation methods. The two main issues involved are the choice of an adequate sampling frequency, and correctly changing the original input-output differential equations associated with $H(s)$ to difference equations associated with $H(z)$.

8.1.1 Sampling Rate Selection

The sampling rate used to discretize the power of the continuous-time signal coming to the gateway on the downlink largely determines the quality of control that can be applied to the system. With a high sampling rate, the resulting signal appears more like the original continuous-time version, and more of the dynamic
behavior is captured. However, there is a danger of ill-conditioning if the sampling rate is too high, because there will then be very little change from sample to sample.

For the purposes of control, the Nyquist Criterion suggests that the sampling frequency should be at least twice the closed-loop bandwidth of the system -- that is, approximately two times the crossover frequency, \( \omega_c \), of the loop transfer function, \( L(s) \). In practice, however, this minimum rate does not provide adequately smooth time-domain performance, so signals are usually oversampled by a factor of 6 to 40 [9, p.485].

Following a discussion of sampling rate selection in the Franklin and Powell textbook [9] on digital control systems, slower sampling rates can add a significant delay to the control system because continuous information is not available until the next sampling instant. In the case of the golden phone system, this delay is already addressed in the timing diagram of Figure 7.2-1. In general, however, it is suggested that the sampling rate be greater than 20 times the closed-loop bandwidth (\( \omega_c/\omega_c > 20 \)) in order to realize a delay that is less then 10% of the rise time to a step response.

Another observation is that, if there is noise in the system, then sampling at higher rates spreads this noise power over a higher range of frequencies, even though the working bandwidth of the control system may stay the same. This means that a greater amount of this noise power is automatically filtered out by the low-pass nature of the control system. In the golden phone case, this is particularly true of the error in the power measurement, \( E_{PMP} \), which is essentially uncorrelated white noise that is filtered by the system on its way to the output.

Finally, in an analysis of the system response to white noise disturbances, it is determined that a sampling rate of greater than 20 times the bandwidth provides minimal degradation of disturbance rejection compared to the original continuous time controller. [9, p.491]

In the control system at hand, the maximum closed-loop bandwidth for a plant delay of 250ms is 1.0 Hz, as seen in Figure 7.5-4 on page 79; however, it is more realistically in the range of the MinZero bandwidth of 0.3 Hz. From the above analysis, a 25 times oversampling seems to be large enough to create an accurate
discrete-time emulation of the continuous controller, so this suggests a required sampling rate of 7.5 Hz. In the timing diagram of Figure 7.2-1 on page 69, 8.0 Hz seems to give the gateway infrastructure a comfortable amount of time to communicate the RSSI measurement to the GCU. Consequently, 8.0 Hz is chosen as the nominal sampling rate.

### 8.1.2 Transformation Techniques

Once the sampling rate is chosen, the second issue of conversion is transforming the actual controller. There are at least three popular transform methods: the zero-order hold equivalent (ZOH); the zero and pole mapping method (ZP); and the bilinear transformation, with the option of “pre-warping”.

The ZOH method approaches the problem in the time-domain, and creates a controller that, given a sampled version of the continuous input, generates an output that is equivalent to the sampled version of the output from the continuous controller. The ZP method, on the other hand, directly maps the zeros and poles from the continuous-time controller to their equivalent locations in the z-plane using the map \( z = e^{Ts} \), where \( T \) is the sampling period -- in this case 0.125 seconds.

Lastly, the bilinear transformation is derived from the trapezoid rule of numerical integration. The mapping for this transform is given below, and has the particular advantage of pre-warping. This feature lets the designer choose a frequency at which the discrete equivalent of the controller perfectly matches the frequency response of the original continuous-time version.

All of the above transform methods have been tested on the MaxZero controller of 55 degrees phase margin to observe the relative merits of each. Since the major design emphasis until now has been on the crossover frequency, \( \omega_c \), it is chosen as the pre-warping set point in the bilinear transform. As a result of this specific customization, the bilinear transform with prewarping provides the best frequency response emulation of the continuous-time controller. The frequency response of the MaxZero controller and the result of the bilinear transformation of \( H(s) \) is shown in Figure 8.1-1.
8.1.3 *Direct Discrete-Time Design*

In the process of emulating the controller developed in Chapter 7, one might wonder whether a design that is formulated directly in discrete-time would perform much better than the continuous-time equivalent. This possibility has been investigated, and the important result is that the two design techniques essentially follow an equivalent pattern of development, so the results should not be largely different.

### 8.2 Time-Domain Performance

The purpose of this analysis is to determine the time-domain performance of different controllers in the realm of possibilities such that one can be chosen in accordance with the performance goals of Section 7.3.
This realm of possibilities, or sample space, of controllers is determined by the range of values assumable by the gain, \( K \), and the zero location, \( a \), in the transfer function, \( H(s) \). At the end of Chapter 7, however, this same sample space is also described by each unique combination of desired phase margin, \( \text{despm} \), along with the zero location, \( a \). The latter is not described in absolute terms, but rather by the fraction \( q \) of the zero location, \( a_0 \), that results from the MaxZero strategy. The desired phase margin is a good measure of stability, and the fraction maps to the choice of design strategy between MinZero and MaxZero, so this description of the sample space is easier to grasp in terms of its expected performance.

The following analysis is conducted using the MATLAB script PeakFrac.m in the Appendix, which iteratively explores the sample space and records the measures of interest defined in Table 7.3-1. The continuous-time controller corresponding to each combination of \( \text{despm} \) and \( q \) is specified, and converted to discrete-time using the bilinear transform method with prewarping matched to the continuous crossover frequency. The step and ramp responses of the discrete design are then calculated, and the figures of crossover frequency, achieved phase margin, peak overshoot, rise time, and settling time are recorded.

These results are presented in the following surface plots, which are generated by the script PlotPeakFrac.m, and which give a good sense of the general pattern of different measures over the sample space. A short explanation of why these particular trends exist over the sample space is also given. Once the results are recorded, a design decision is made in Section 8.3 using all of these measures simultaneously.

The two graphs in Figure 8.2-1 show the pattern of achieved phase margin and crossover frequency as the designed phase margin, \( \text{despm} \), and the fraction \( q \) are varied. The results for achieved phase margin are intuitive, because once the \( \text{despm} \) is chosen, the choice of fraction is irrelevant. As the fraction decreases from one to zero, the peak phase margin of \( L(s) \) increases, but the gain is readjusted to keep the phase margin, \( \text{despm} \), constant (see Figure 7.5-3). The second plot above shows that the benefit of decreasing the fraction \( q \) is an increase in crossover frequency, and a corresponding increase in gain. As the designed phase margin is increased, extra stability is achieved at the sacrifice of bandwidth and gain.
The first requirement of Table 7.3-1 to be analyzed is the peak overshoot, which is shown in Figure 8.2-2. If the second order approximation of the $L(s)$ had held, this would be closer to a true plane because achieved phase margin would closely correlate to the damping ratio, which solely determines the peak overshoot. As it stands, there is still a good correlation between the surfaces for phase margin and peak overshoot, but as the fraction decreases, the peak overshoot also seems to decrease, thus tilting its plane over the sample
space. At higher phase margins, the bump in the plane is probably caused by a breakdown of the correlation between phase margin and the damping ratio.

![Figure 8.2-3: Rise Time to -0.1 dB (unit step input)](image)

The second metric shown in Figure 8.2-3 is the initial rise time to -0.1 dB (0.9772) in response to a unit step input. The time seems to increase both with despm, and the fraction used. This is a reasonable result because both of these trends also lead to a lower crossover frequency, which corresponds to a reduction in speed of response.

The third metric shown is the settling time of a step response to within 0.1 dB of the final value. The result seen in Figure 8.2-4 is a little more complicated than the previous ones, so it is broken into regions for analysis. At high phase margins and high fractions, the settling time mimics the behaviour of the rise time. This makes sense since the low crossover frequencies in this region make the system slow both in rise time and settling time. In the area of low phase margin and high fraction, the step response also takes a long time to settle, but the rise time is quick. An explanation of this is the oscillatory behavior that is characteristic of lower phase margins, which causes the response to settle slowly. Finally, the area of high phase margin but very low fraction seems to have surprisingly good settling time. This behaviour is explained by a combination of good stability (less oscillation) resulting from the phase margin, along with a controller that looks
increasingly like a first-order compensator as $q$ approaches zero. The step response in this region is therefore simply a smooth exponential that monotonically increases to a final value.

In general, the settling time does not decrease with lower fractions as much as rise time does. This is probably because the zero location is approaching the $s$-plane origin as $q$ goes to zero. As the controller gain increases to keep the phase margin constant, one of the closed-loop poles depicted in the root locus of Figure 7.7-1 is approaching this low frequency zero; therefore, it causes very slow dynamics that give rise to long-tailed transients. In terms of settling time, this slow behaviour offsets the gain in speed that is achieved with higher crossover frequencies, as clearly demonstrated by the drop in risetime over the same area of the sample space.

The last of the requirements to be investigated is the settling time in response to a unit ramp input, which is shown in Figure 8.2-5. The behaviour of this surface is dominated by the large increase in settling time as the absolute zero location, $a = qa_0$, decreases in frequency. Both an increased $despm$ and a decreased fraction $q$ lead to a lower $a$, and describe a controller that is more first-order in nature than before. In the high phase margin and low fraction area of Figure 8.2-5, the measure of settling time becomes saturated at 30 seconds because the controllers of that region are essentially first-order and thus demonstrate a steady state error in
response to a ramp (see Table 7.4-1). In other words, they never settle down to within 0.1 units of the desired output.

8.3 Controller Selection

Now that time-domain behaviour of the controllers in the sample space is better understood, an informed decision can be made about an appropriate controller for the current system. In order to make a selection, all the performance metrics need to be considered simultaneously, so overlaid contour plots are generated to aid in this process. The base of these plots is of filled contour type, showing the various levels of the settling time exceeded in response to a step input. The surface plot of the same data is shown in Figure 8.2-4, as described above. The contours shown correspond to the settling time values of \([0, 1, 2, 3, 3.5, 4, 5, 6, 8, 10]\) seconds. This can be changed in the script PlotPeakFrac.m, but they are chosen to provide maximum information without clutter.

In Figure 8.3-1, the settling time plot is overlaid with a contour plot of rise time in response to a step response, and also the peak overshoot fraction, which is shown by dashed lines. Figure 8.3-2 shows the overlaid contour plot of the crossover frequency, which is a good measure for the amount of disturbance rejection.
Figure 8.3-1: Overlaid Contour Plots of Peak Overshoot and Rise Time for Step Input

achieved by the system, because faster disturbances can be rejected. Finally, in Figure 8.3-3, the contour plot of the settling time in response to a ramp input is once again overlaid on the plot of settling time to a step. For convenience, this basic settling time plot is essentially used like a map by which to compare the three plots. The choice of any one spot in this sample space leads to a unique combination of all the performance metrics being considered. After some analysis, an operating point is chosen at $\text{despm} = 43$ degrees, and $q = 0.7$.

The real advantage of this analysis, however, is not necessarily to choose a definite operating point right away. If it is ever determined that the performance requirements in Table 7.3-1 need to be modified, a new design awaits quick identification in the plots above. If the assumed plant delay is changed, the MATLAB
Figure 8.3-2: Overlaid Contours of Achieved Bandwidth (Hz)

Figure 8.3-3: Overlaid Contours of Settling Time to Ramp Input
scripts that generated the plots can be modified to make a new set with updated numbers. As long as the plant delay is always translated to a two sample delay, the actual shape of the above trends should not change.

The controller designed at an operating point of 43 degrees phase margin with \( q = 0.7 \) is chosen because it provides good performance without sacrificing too much stability or delay robustness. The expected overshoot is between 20 and 30 percent, the settling time to both the step and ramp inputs is less than 5 seconds, the rise time to a step is less than 0.8 seconds, and the crossover frequency is between 0.3 and 0.4 Hz. From Figure 8.2-4 on page 91, the delay robustness ratio is probably just slightly less than 2 so that plant delays less than 500 ms can be absorbed by the system with some degradation of response, but continued stability.

The operating point is further supported when the contour plots of that region are analyzed. If the despm value were decreased, the delay robustness ratio and the peak overshoot percentage are sacrificed for better settling times and a slightly higher bandwidth. If, on the other hand, the fraction is decreased, the settling times and robustness are sacrificed for better bandwidth and shorter rise times. Other such trade-off relationships can also be gleaned from the contour plots but they are not addressed here specifically.

### 8.4 Summary of Controller Performance

The final choice of a controller provides good delay robustness and time-domain performance in the face of a 250 ms plant delay. The controller can accurately follow a ramp-like command, and equivalently reject linearly increasing disturbances. The rise time and percentage overshoot requirements are both met. To summarize this nominal controller, \( H_p(z) \), and its performance, the following equations and figures can be used as a reference.

\[
H_p(s) = \frac{K(s + a)}{s^2} = \frac{K(s + qa_0)}{s^2}
\]

\[
H_p(z) = \frac{b_0 + b_1 z + b_2 z^2}{(z - 1)^2} = \frac{K_d(z + 1)(z - a_d)}{(z - 1)^2}
\]
Figure 8.4-1: Various Plots for Description of Controller Performance
Table 8.4-1: Properties, Parameters and Values

<p>| | | |</p>
<table>
<thead>
<tr>
<th></th>
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<th></th>
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</thead>
<tbody>
<tr>
<td>$q$</td>
<td>0.7</td>
<td>$b_2$</td>
</tr>
<tr>
<td>$a_o$</td>
<td>0.707946</td>
<td>$b_1$</td>
</tr>
<tr>
<td>$a$</td>
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<td>$b_0$</td>
</tr>
<tr>
<td>$f_c$</td>
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<td>$T_s$</td>
</tr>
<tr>
<td>$K$</td>
<td>2.11922</td>
<td>$a_d$</td>
</tr>
<tr>
<td>$K_d$</td>
<td>0.1374</td>
<td>$D_L$</td>
</tr>
<tr>
<td>PM</td>
<td>46.13 deg</td>
<td>$D_{max}$</td>
</tr>
</tbody>
</table>

Table 8.4-2: Performance of Chosen Controller

<table>
<thead>
<tr>
<th>Requirement</th>
<th>Constraint</th>
<th>Achieved</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rise Time to 0.1 dB of final value</td>
<td>&lt; 1 sec</td>
<td>0.75 sec</td>
</tr>
<tr>
<td>Settling Time to 0.1 dB of final value (unit step input)</td>
<td>&lt; 5 sec</td>
<td>4.625 sec</td>
</tr>
<tr>
<td>Settling Time to 0.1 units of error (unit ramp error input)</td>
<td>&lt; 5 sec</td>
<td>3.1 sec</td>
</tr>
<tr>
<td>Settling Time to 0.1 dB of error (unit ramp error input)</td>
<td>***</td>
<td>5.5 sec</td>
</tr>
<tr>
<td>Percentage Overshoot</td>
<td>&lt; 50%</td>
<td>29.35%</td>
</tr>
</tbody>
</table>
Chapter 9: Simulation

The design process described in Chapter 7 and the selection method described in Chapter 8 have resulted in the controller, $H_p(z)$, as described in Section 8.4. During this process of design, however, a number of assumptions have been made in order to simplify various aspects of the system and to enable the use of linear time-invariant analysis methods. The main purpose of the simulation phase is to relax these assumptions and confirm that the system performance is not adversely affected. By implementing the controller in a model that more closely resembles the real system, one can also see how performance is affected by changes in certain system parameters, or by various combinations of inputs.

The scope of this chapter on simulation is mainly to establish the model itself, and identify the previous assumptions that can be relaxed as a result. Once the model is defined, the nominal controller presented in Chapter 8 is inserted and tested. These tests confirm expected performance and enable one to analyze the effect of new variables that are introduced by a more complete description of the golden phone system.

9.1 The SIMULINK Model

The simulation is built in a SIMULINK environment because much of the design has been done using MATLAB scripts, and SIMULINK is a natural progression from that workspace. The advantages found in this environment are a very attractive user interface as well as a faster running time and an elegant treatment of sampling issues. The main model diagram is presented in Figure 9.1-1, and this includes both the controller itself and most of the signals and blocks that are critical to the dynamics of the system. There are, however, a number of subsystems that have been condensed into single blocks in order to reduce clutter on the main page, so these are presented in Figure 9.1-2 and Figure 9.1-3.

The main page of the SIMULINK model very much resembles the system diagram presented in Chapters 6 and 7. The signal in the top left hand corner is the power of the pilot as it emerges from the GMOD card. It is then pre-compensated by the “Uplink Estimate” subsystem and amplified by the control effort that is calculated by the GCU. The resulting signal is sent over the “Uplink”, and the result is the satellite output power which needs to be controlled. This power then travels the “Downlink” and is measured by the RSSI.
filter in the golden phone, G(s), with an implied communication delay of \(D_{D2}\). As previously explained, this power measurement on the downlink can be noisy, so a uniformly distributed random noise source is chosen to represent this uncertainty. The noise strength is defined by \(+/-\) \(\text{DownNoise}\), and it is applied with a sampling interval of \(\text{ErrStep}\). This noisy measurement is then passed through a single-pole anti-aliasing filter with cutoff of 3 Hz, which prepares the signal for sampling by the zero-order-hold block with a period of \(\text{ConStep}\), which is nominally set at 8 Hz. This represents the interface between the physical system and the domain of the GCU microprocessor in which the nominal controller \(H_p(z)\) resides.

The “Downlink Estimate” subsystem emulates the expected gains of the downlink path with the help of information received in the RAI, and the result is subtracted from the sampled power measurement entering the GCU. The result of this post-correction activity is recorded as \(P_{DTx\_inf}\), or the inferred satellite output power. This inferred value is then compared to the specified power, \(P_{SPEC}\), which is shown as a step input in the diagram. Finally, the derived error signal from this comparison is fed to the discrete-time controller, \(H_p(z)\). The \(G\)-offset generated by the controller encounters a gain block and a saturation block before being applied to the pilot power, but these exist more for testing purposes than to model reality. Most
of the time, ExtraGain is set to 1, and the cutoffs SaturationUp and SaturationDown are set to 10 and -10 respectively. In reality, the overhead power control system has been assigned 5 dB of the link budget, but there is nothing that restricts the controller from exceeding this assignment.

Almost all of the blocks in the model have been set up to take their values from variables defined in the MATLAB workspace rather than in the model. Also, every distinct factor that is shown in the system diagram of Figure 7.1-1 is included in the simulation model; therefore, each one has a block as well as a variable assignment that must exist for the simulation to run. As noted before, however, there are only two distinct positions at which a factor can enter this loop. At each of these locations, one factor has been chosen to implement user-specified signal patterns, and another has been chosen as a random noise-like input. On the uplink, the user-specified signals are input via the rain attenuation factor, L_RU, and on the downlink by P_SPEC. The random noise input for the uplink portion is provided by G_URx, and the downlink portion is cov-
Before moving on to the results, it should be noted that the default values for parameters in this model are provided by the script ThsLoad.m which is provided in the Appendix on page 116.

9.2 Performance Verification

The first set of tests conducted on the above model are to verify the expected performance of the controller presented in Section 8.4. Initially, performance of the controller is verified in a hybrid system consisting of both continuous-time signals (at 200 Hz simulation rate) and variously sampled discrete-time signals. In order to make the hybrid nature of the plant an isolated change, the RSSI filter \( G(s) \) is set to unity and the power meter is represented only by the delay \( D_{D2} \). The simulation driver does not deal well with continuous-time delays that are of the same duration as \( \text{SysStep} \), so the uplink and downlink propagation delays, \( D_{U1} \) and \( D_{D1} \), are each assigned 10ms (two time steps) and the remaining 230ms of delay are assigned to \( D_{D2} \).
The response of this system to multiple step inputs is shown in Figure 9.2-1. The three inputs are the events of $P_{\text{pilot}}$ taking a unit step at $t = 1$ second, $P_{\text{SPEC}}$ taking a step at $t = 5$ seconds, and $L_{\text{RU}}$ taking a negative step at $t = 10$ seconds. As expected, the system rejects the step inputs of $P_{\text{pilot}}$ and the increased rain attenuation by showing zero steady state error, while it follows the command given by $P_{\text{SPEC}}$. Also, the shape of the step response closely resembles that of the expected closed-loop step response and error response predicted in Section 8.4.

The next two graphs in Figure 9.2-2 show the response of the system to a ramp in $L_{\text{RU}}$ that drops from 0 to -10 dB in 5 seconds. This represents the maximum rate of increase for rain attenuation at C-Band, as specified in Chapter 4. The responses shown are again in accordance with the expected shape from Chapter 8, however, there is seemingly some noise at the output while the rain fade is ramping. This noise-like pattern is the first visible indication that the disturbance is continuous while the controller is discrete. Basically, the rain fade is changing between sampling instants, so the variation is hidden from the controller and proceeds directly to the output. In general, however, the response of the system is quite acceptable because what would have led to -1 dB of output power now leads to almost no change.
Now that the basic system has been modeled, it is time to relax the assumption that the power meter can be considered a pure delay. As mentioned in Section 7.2, the RSSI filter really acts like a single-pole filter with a time constant of 53.1 ms. Inserting this response as the transfer function $G(s)$, this leads to 159.3 ms of delay if one waits for the system to settle before polling the meter. The remainder of the assumed 250 ms delay is kept in $D_{D2}$ which becomes 70.9 ms. Testing this new system, one would expect slightly better performance, because more information is being fed back sooner. The results of the step response are shown in Figure 9.2-3, and it is almost exactly like the one of Figure 9.2-1. If one looks closely, however, there is a slight change in the shape of the response, and there is a lower peak overshoot. Both of these indicate an increase in phase margin, which is expected from a decrease in feedback delay. From now onwards, the system is tested in the configuration described above, in which $D_{D2}$ is 70.7 ms, and the power meter is a single-pole filter with a time constant of 53.1 ms.

### 9.3 Robustness

Another performance metric that needs to be verified is the robustness of the controller under varying system circumstances. The most critical parameter in this respect is the communication delay, so in Figure 9.3-1a a single step response is shown as the value of $D_{D2}$ is increased in integer multiples. As expected, the response becomes less stable as delay increases. The last value to be tested is six times the original 70.7 ms, so the
total communication delay experienced is in excess of 420 ms. The design requirements call for at least 400 ms of plant delay robustness, so this is achieved by the nominal controller.
The second robustness test presented concerns the gain of the loop. If for some implementation reason the gain of the controller is changed, the system should not easily become unstable. The results of this test are provided in Figure 9.3-2, where an ExtraGain of 3 brings the system close to the minimum acceptable performance. This is an interesting result, because the gain margin predicted for the loop by theoretical means is 2.92. This margin is thus exceeded by the simulated system, and the discrepancy is attributed to the relaxation the power meter delay assumption.

![Simulated Time Response](image)

**Figure 9.3-2: Gain Robustness Test**

The gain values tested were ExtraGain = [0.25, 0.5, 1, 1.5, 3]

The last robustness test involves changing the sampling rate at which the system operates. At first, the discrete-time controller is kept constant while the operating rate is varied in the simulation. For a reduction in the operating rate, it is expected that the system should react more slowly to inputs, whereas when the rate is increased, the response should be faster. Another option is to change the sampling rate with which the controller is converted to discrete-time.

The results of such sampling changes are presented in Figure 9.3-3. In the first graph, the operating rate is increased while the controller is kept constant. The result is an increase in response speed, as well as a decrease in stability, because the same continuous-time delay translates into a greater number of sample
Figure 9.3-3: Changing the Sampling Rate

a) (top left) The result of keeping the nominal controller while operating at 16 Hz and 32 Hz
b) (top right) The result of designing and operating the controller at 4 Hz.
c) (bottom left) The result of designing and operating the controller at 40 Hz
d) (bottom right) The result of designing the controller at 40 Hz, but operating at 8 Hz.

delays. If, however, the same nominal $H(s)$ is converted to $H_p(z)$ at a higher rate of 40 Hz, and then used in a control system operating at 40 Hz, the results in Figure 9.3-3c show that the response is essentially the same as the nominal controller operating at 8 Hz, with a smaller response lag. In Figure 9.3-3d, the performance of this new $H_p(z)$ is degraded by operating at the regular 8 Hz.


## 9.4 Measurement Noise Transmission

One of the main errors that the closed-loop system will experience at the output is the noise that is inherently included in the RSSI power measurement. The assumed noise figure for the golden phone with an RSSI gain of $G=4$ is $\pm 0.5$ dB. The noise is modeled as uniform between these two extremes and entered into the loop as $E_{\text{PM}}$. The time step used to enter this noise is the same as the controller sampling frequency of 8 Hz. As shown in Figure 9.4-1, the noise that results at the output of the satellite has been lowered in frequency. This is expected because that input position sees a closed-loop system that looks like a low-pass filter with a bandwidth of 0.35 Hz. The variance of the input noise is 0.0846 (1/12), while that of the output signal, $P_{\text{DTx}}$, is 0.0178. Obviously, there is some reduction in the noise variance at the output, but for the most part, noise moves through to the output quite freely.

When the sampling rate of the input noise is increased to 40 Hz, it is expected that the noise power is spread across more of the frequency spectrum, and therefore, the system should be able to filter more of it out. However, about the same result is seen at the output in both Figure 9.4-2a and Figure 9.4-1a. The explanation of this departure from the expected is in the modeling of noise as a uniform random number, which does not necessarily provide a constant level of total noise power. The two diagrams show that the basic in-band noise power at the output stays the same in the two cases. In Figure 9.4-2, the variance of the noise is 0.0795, while that of the output is 0.0221.
Since the sampling rate of the noise is not a strong parameter, maybe the expected drop in noise transmission can be achieved by changing the sampling rate of the controller. Testing the version of the nominal controller discretized at 40 Hz, the noise response is given in Figure 9.4-3. In this case, the variance of the input noise is still 0.0795, but that of the output is much lower at 0.0035. The sampling rate of the controller is thus established as one way to decrease noise transmission from the feedback path to the output. The other way, as previously discussed, is to change the gain of the RSSI filter, since this determines the uncertainty in the
actual power measurement. An implicit trade-off between sampling rate, RSSI filter gain, loop delay, and noise transmission emerges from the results presented above. For now, however, the sampling rate is kept at 8.0 Hz because this clock is readily available in the GCU. It should be noted that if measurement noise is ever determined to be excessive, and time performance can not be sacrificed, an increased sampling rate can reduce the transmission of measurement noise to the output.

One other parameter that can have a significant effect on noise transmission is the ExtraGain, because a higher loop gain results in a higher crossover frequency, and more of the measurement noise may pass through to the output. If ExtraGain is set at 2, the resulting noise at the output is depicted in Figure 9.4-4. Visibly, there is an increase in the noise power at the output, and this is confirmed by a rise in variance from 0.0178 to 0.0509.

To summarize the findings of the noise transmission analysis, as shown in Figure 9.4-1, the +/- 0.5 dB noise from the RSSI filter translates to +/- 0.23 dB of uncertainty at the output if the noise is modeled as uniformly distributed, or +/- 0.26 dB if modeled as Gaussian noise (95% confidence intervals).

### 9.5 Uplink Errors

Just as the noise on the downlink sees a low-pass filter on its way to the output, the uplink uncertainties face a high-pass filter in their transmission to the output. This is good because the two most important uplink dis-
turbances -- rain fade and transponder gain uncertainty -- have already been addressed as changing quite slowly in comparison to the bandwidth of the closed-loop control system. The successful rejection of a ramp input in Section 9.2 confirms that these errors are adequately eliminated by the golden phone strategy.

There are, however, other error sources on the uplink, and they have the potential to change somewhat more rapidly. In order to model these inputs, a white noise generator is shaped by a low-pass filter with a time constant of 10 seconds. This creates a signal that is random, but somewhat correlated, as shown by the autocorrelation of upnoise in Figure 9.5-1c. This noise shaping allows random walks to be taken by the signal, as shown in the results of Figure 9.5-1a,b. The resulting upnoise is added in the position of the uplink receiving antenna, \( G_{URx} \) and monitored as it carries through to the satellite output. Based on results presented in the figures, the major benefit of the control effort is in compensating for the random walk aspect of the disturbance. The higher frequency portions, however, pass through to the output before anything can be done by the controller.

The uplink noise variation is measured as a root-mean-square value of 0.7285, which translates to an error of \( \pm 1.5 \) dB if noise can be considered Gaussian. The error that passes to the satellite output has an RMS of 0.1898 at a designed sampling rate of 8.0 Hz, and is found to be 0.1766 if the sampling rate is moved to 40 Hz. By these results, one can expect an error of \( \pm 0.38 \) dB from the uplink path, and there is little advantage in a higher sampling frequency for disturbance rejection.

Following the example of the measurement noise transmission that increases with bandwidth, there is some possibility that a similar decrease in uplink error transmission might result from a higher loop gain. When ExtraGain is set to 2, there is no visible improvement in performance, and the resulting RMS value at the output is again 0.1872. It is obvious that the small increase in bandwidth is not particularly effective in reducing the uncertainty at the satellite output.

### 9.6 Simulation Summary

In the preceding simulation scenarios, it has been confirmed that the performance promised by the controller in Chapter 8 is indeed achieved. The shape of the predicted step response is confirmed, as well as the ability
of the system to reject ramp-like disturbances with zero steady-state error. In addition to this verification procedure, the system has been subjected to a number of robustness tests in response to varying loop gain, communication delay, and sampling rates. The result is an assurance that the system is reliable in the face of uncertainty in system parameters. Also, an interesting trade-off relationship between the downlink noise transmission and the loop sampling rate has been found, and this development may assist in future design.
Given the nominal controller design, and a sampling rate of 8.0 Hz, a partial error analysis has also been conducted based on simulation results. In Section 6.2.3, all of the uplink gains are taken to be perfectly rejected by the control system. However, in the error analysis, a correlated noise signal is added to the uplink path, and the high frequency content is shown to affect the output with an error of +/- 0.38 dB.

On a similar note, it is shown that the high frequency components of downlink errors are significantly attenuated on the path to the satellite output. For the particular case of noise added at the RSSI filter, the uncertainty decreases from +/- 0.5 dB at the source to +/- 0.23 dB at the output. Therefore, the corresponding gain in certainty offsets the extra error that is added by the uplink path. In the end, the error figure of the golden phone strategy should be increased by +/- 0.11 dB to +/- 1.41 dB. This estimate may be further reduced because the other downlink errors may experience an attenuation in the same way as $E_{PMP}$. Such an analysis would be premature for these factors because they have not been empirically characterized, but by including their entire range of error, the error analysis for the system is conservative.

It is obvious that a lot of interesting analysis can be conducted with the help of the simulation model established in Section 9.1. However, as stated at the beginning of the chapter, the results presented here are purposely contained in their scope in order to simply confirm the performance of the nominal controller design. Some directions for future analysis and potential improvements in the nominal design are suggested in Section 10.2.
Chapter 10: Discussion

The discrete-time controller that has been developed and simulated in the course of this thesis has been shown to provide the desired performance in controlling the output pilot power of the Globalstar satellite.

10.1 Design Summary

The performance constraints established for this system include time-domain measures as well as a minimum level of delay robustness in order to absorb unexpected conditions in the gateway infrastructure. Under these constraints, the goal of the system was to provide as much disturbance rejection as possible, because factors such as rain attenuation and uncertain transponder gains can interfere with the power management ability of the system.

In response to these goals, the strategy of the “golden phone” was chosen because it provided the potential to overcome both of the major disturbances, and at the same time offered a more tunable closed-loop design that can be easily implemented at the individual gateways. The design process then characterized the golden phone system, and developed a discrete-time controller to provide the required and desired performance. Finally, this nominal controller was tested in simulation, and the promised performance verified.

Throughout the design process, a conscious effort has been made to keep the design procedure flexible and easily repeatable, because of current uncertainty in model parameters. Also, the exact consequences of failing to meet the performance requirements given in Section 7.3 are unclear because the Globalstar system is not yet fully operational. Once the system is running, it may become clear that, for example, a 20% peak overshoot is more detrimental to the satellite lifetime than a slow response is to call stability. This would lead to a change in the design requirements of Section 7.3, and another iteration of design process. In this case, the only design step that needs re-investigation is the final choice of controller from the contour plots of Chapter 8. Once that choice is made, the MATLAB scripts provided as well as the SIMULINK model can be quickly modified.
In order to implement the current nominal design, the system requires the addition of a golden phone on top of each transmitting antenna at the gateway. This is simply a modified Globalstar phone with a more directional antenna pattern in order to reduce the multi-path interference that normally degrades reception quality. The controller portion of the design requires a sampling rate of 8 Hz inside the GCU, and expects a measurement to arrive from the RSSI filter with a time delay of less than 80 ms. It is also expected that the coefficients of the controller have an accurate floating point representation, and that any calculated control effort changes the power of the transmitted pilot within a matter of nanoseconds.

Given the current understanding of system parameters operating in the Globalstar forward link, the golden phone strategy and the corresponding controller, \( H_p(z) \), are a very good combination to accurately control the overhead channel power at the satellite output. This increased accuracy improves the reliability of intelligent power management, and thus increases the capacity of the Globalstar system.

### 10.2 Future Directions

As a result of the uncertainty in system parameters, and in the understanding of the consequences of poor performance, there are many directions in which the design of a power control system can be extended into the future.

The first analysis that should be conducted involves better characterizing the inputs affecting the system. Once the basic Globalstar system is made operational, a prototypical golden phone should be placed at the gateway, and the S-band downlink signals measured over time. As the system encounters rain, data can be accumulated to record the extent and the dynamics of the rain attenuation that is experienced. This information will be specific to Globalstar frequencies, and to communication with non-stationary satellites, and should greatly improve the estimates of the extent and speed of rain attenuation. The data received during clear sky conditions should also provide some idea of the other unexpected gains in the system like that of the transponder; the communication delay of the RSSI measurement can also be better evaluated. Eventually, as all of this knowledge is accumulated, the controller design can be tuned to operate in the better defined system. For now, the nominal controller is simply made more robust in order to compensate for the present uncertainty.
A second category of analysis is to develop a cost function for use in trading off the different specifications of Globalstar performance such as system capacity, capacity utilization, satellite battery lifetime, and system availability. For the analysis in this thesis, it was assumed that only the system capacity would be hit by the presence of uncertainty and the resulting inefficiency in power management. This assumption is good for comparison of different strategies, but needs to be relaxed in order to set the nominal bias points of overhead channel power, for example. With the development of a cost function, it can be decided that other aspects of performance can be sacrificed in order to regain a minimum system capacity. Such a decision might affect the performance requirements that are placed on the designed controller, as shown in the example of the last paragraph.

Finally, the most interesting of future analyses might be to combine portions of all four strategies in order to make up for relative weaknesses in each, and cumulatively provide more functionality. One such combination was already suggested in Section 5.4, whereby the PSMM strategy compensates for the potential “shadow” periods in the golden phone scheme. Another such example is to use the received beacon power as a signal to indicate the presence or absence of rain, rather than an accurate measure of the extent of rain. For example, if the presence of rain is known, the golden phone system might be customized to compensate for downlink rain attenuation by adjusting its post-correction activity. Such dynamic control scenarios can be tested by extending the simulation model provided in Chapter 9.

In the event that the above studies cannot be completed, the current design can be trusted to robustly provide good performance in combatting uplink disturbances. The result of this implementation is that less uncertainty drives the output power of the satellite, so more intelligent power management methods can be implemented. Without a sacrifice of either system availability or lifetime, the ultimate benefit is an increase in the Globalstar system capacity, which was the ultimate goal of the overhead channel power control system.
11 Appendix: MATLAB Scripts

The following MATLAB scripts have been generated during the development of this thesis. Some of these have been referred to in the text, while others are mentioned here for the first time. The three distinct sets of code are described in separate sections for ease of reference.

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The scripts archived here should be run in the Matlab directory with the subdirectories data and graphs available. Many of the scripts here also rely on some functions specific to the MIT class 6.302. These are related to the ramp response, and any supporting scripts. If in the MIT Athena environment, only the appropriate path (/mit/6.302/matlab) need be added. Many of the scripts also assume the existence of previously generated data files, but these can be regenerated by uncommenting the code in the respective scripts.
11.1 Rain Attenuation Studies

**RainExc.m**

```matlab
function RainExc(data, meth, El);
% RAINEXC custom function
% Generates Rain Attenuation Exceedance Curves for
% Specific Climates relative to Globalstar given the
% climate data and attenuation model to be used.
% RainExc(data, meth, El)
% data: 0=CCIR, l=Global (default=l)
% meth: 0=CCIR, l=SAM (default=l)
% El: elevation in degrees
% Output to GlSAM.ps, GlCC.ps, CCSAM.ps or CCCC.ps
% depending on the data and method chosen
% See also CLIMATE
%
if nargin==0
    data = 1;
    meth = 1;
    El = 10;
end

mark = 7;
if data==1
    w=1:10;
    Res = climate(l, meth, El, 0);
    hold off
    semilogx(Res(l,w), Res(8, w), '>-', 'markersize', mark)
    hold on
    semilogx(Res(l,w), Res(11, w), 'd-', 'markersize', mark)
    semilogx(Res(l,w), Res(12, w), 's-', 'markersize', mark)
    semilogx(Res(l,w), Res(13, w), 'x-', 'markersize', mark)
    grid on
    xlabel('Percentage of Time')
    ylabel('Attenuation in dB')
    legend('D: Chicago, USA', 'F: San Diego, USA',...
    'G: Puebla, Mexico', 'H: Suphanburi, Thailand')
    if meth==0
        title('CCIR Exceedance Curves for Various Global Climates')
        print -deps -epsi graphs/GlCC.ps
    else
        title('SAM Exceedance Curves for Various Global Climates')
        print -deps -epsi graphs/GlSAM.ps
    end
else
    w=1:7;
    Res = climate(0, meth, El, 0);
    hold off
    semilogx(Res(l,w), Res(6, w), '>-', 'markersize', mark)
    hold on
```

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semilogx(Res(l,w), Res(11, w), 'd-', 'markersize', mark)
semilogx(Res(l,w), Res(14, w), 's-', 'markersize', mark)
semilogx(Res(l,w), Res(15, w), 'x-', 'markersize', mark)
grid on
xlabel('Percentage of Time')
ylabel('Attenuation in dB')
legend('E: San Diego, USA', 'K: Chicago, USA', ...
 'N: Puebla, Mexico', 'P: Suphanburi, Thailand')
if meth==0
    title('CCIR Exceedance Curves for Various CCIR Climates')
    print -deps -epsi graphs/CCCC.ps
else
    title('SAM Exceedance Curves for Various CCIR Climates')
    print -deps -epsi graphs/CCSAM.ps
end
end

RainLoad.m

% Sets up the Rain Data.
% The matrices are entered in the same form as the reference table
% They are then transposed so that each CLIMATE is a ROW
% The corresponding percentages are defined in the P vectors...

Global = [ 29 45 58 70 78 90 108 126 165 66 185 253;  
            21 34 44 54 62 72 89 106 144 51 157 220.5;  
            13.5 22 28.5 35 41 50 64.5 80.5 118 34 120.5 178;  
            10 15.5 19.5 23.5 28 35.5 49 63 98 23 94 147;  
            7 11 13.5 16 18 24 35 48 78 15 72 119;  
            4 6.4 8 9.5 11 14.5 22 32 52 8.3 47 86.5;  
            2.5 4.2 5.2 6.1 7.2 9.8 14.5 22 35 5.2 32 64;  
            1.5 2.8 3.4 4 4.8 6.4 9.5 14.5 21 3.1 21.8 43.5;  
            .7 1.5 1.9 2.3 2.7 3.6 5.2 7.8 10.6 1.4 12.2 22.5;  
            .4 1 1.3 1.5 1.8 2.2 3 4.7 6 0.7 8 12]';

Pgl = [0.001 0.002 0.005 0.01 0.02 0.05 0.1 0.2 0.5 1];

CCIR = [ 0 1 0 3 1 2 0 0 0 2 0 4 5 12;  
            1 2 3 5 3 4 7 4 13 6 7 11 15 34;  
            2 3 5 8 6 8 12 10 20 12 15 22 35 65;  
            5 6 9 13 12 15 20 18 28 23 33 40 65 105;  
            8 12 15 19 22 28 30 32 35 42 60 63 95 145;  
            14 21 26 29 41 54 45 55 45 70 105 95 140 200;  
            22 32 42 42 70 78 65 83 55 100 150 120 180 250]';

Pcc = [1 0.3 0.1 0.03 0.01 0.003 0.001];
rainrate.m

function M = rainrate(Rsam, Rcc, P, El, graphtoggle)
% M = rainrate(Rsam, El, graphtoggle)
% M = rainrate(Rsam, El, methsel)
% M = rainrate(Rsam, Rcc, P, El, graphtoggle)
% This function uses the SAM and CCIR methods of rain fade prediction.
%
% Mode Zero:
% The function takes five arguments, and plots the exceedance curves for
% rain data and percentages given, using both methods. (Provide one line
% from the rain rate exceedance tables... then specify if you want a
% graph)
%
% Mode One:
% The function takes three arguments where both Rsam and El are vectors.
% In this case, methsel selects the method (CCIR=0, SAM=1) to be used.
% Returned is a matrix of attenuation at each (rain, el) pair, and a
% surface plot showing this result. (Provide a range of elevations and
% rain rates to be tested, and choose a calc method)
%
% Mode Two/Three:
% The function takes three arguments in which either Rsam or El is a vec-
% tor
% It returns a graph and corresponding matrix of attenuation using both
% methods and sweeping over the vector variable. (Provide a rain rate and
% a sweep of elevation, or an elevation and a range of rain rates... Then
% specify if you want a graph)
%
% Rsam -- mm/h rain rate(s) to be tested -- corresponding to P, if given
% Rcc -- mm/h rain rate exceeded 0.01 percent of the time
% P -- percentages of time Rsam rates are exceeded
% El -- satellite elevation angles of interest
% graphtoggle -- 0 to suppress graphs, 1 to allow graphical output
% methsel -- 0 is CCIR method, 1 is SAM method

Asam is the attenuation for the given Rsam
Acc is the attenuation exceeded for 0.01% of the time
Accp is the attenuation exceeded for P% of the time given Rcc
R(P) is the mm/h rain rate exceeded P% of the time given Rcc

**********
% *** Checking for correct arguments modifying for 3 args
**********

if nargin==3
    graphtoggle = P;
    El = Rcc;
    Rcc = Rsam;
    P = 0.01*ones(size(Rsam));
elseif (nargin==5)&&(length(Rsam)==length(P))
    error('Rsam and P must have same length')
elseif nargin==5
    Rcc = Rcc*ones(size(Rsam));
elseif (nargin==3)&&(nargin==5)
    error('this function requires 3 or 5 arguments')
end
Identifying the Mode of Operation

```matlab
if nargin==5
    mode = 0;
elseif (length(El)>1)&(length(Rsam)>1)
    mode = 1;
    methsel = graphtoggle; 
    graphtoggle = 1;
elseif length(El)>1
    mode = 2;
elseif length(Rsam)>1
    mode = 3;
else
    graphtoggle = 0;
end
```

Setup and Loop Begin

```matlab
clear M;
f = 5.15; % GHz 
lat = 20; % degrees
Ho = 0; % km

sweep = 1; % default
rain = 1; % default

for rain = 1:length(Rsam) 
    for sweep = 1:length(El)

        Method SAM 
        see Pratt/Bostian for calc ref.

        a = 4.21e-5 * f^2.42; % good for 2.9 <= f <= 54 GHz
        b = 0.851 * f^0.158; % good for f <= 8.5 GHz
        alphal = a * Rsam(rain)^b; % uses whatever R(P)

        if lat <= 30
            Hi = 4.8; % km
        else
            Hi = 7.8 - 0.1*lat; % km
        end

        if Rsam(rain) <= 10
            He = Hi; % km
        else
            He = Hi+log10(Rsam(rain)/10); % km
        end

        L = (He-Ho)/sin(El(sweep)*pi/180); % km

        if Rsam(rain) <= 10
            Asam = a*Rsam(rain)^b*L; % dB
        else
```

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\[ d = \frac{b}{22} \log\left(\frac{R_{sam}(rain)/10}{10}\right) \cos(El(sweep) \cdot \pi/180) \]
\[ L_{eff} = \frac{1 - \exp(-dL)}{d} \]
\[ Asam = \alpha_{eff} \cdot L_{eff} \text{ dB} \]

%%% Method CCIR
%%% see Pratt/Bostian for calc. ref.

\[ \alpha_2 = a \cdot R_{cc}(rain)^b \]
\[ h_R = 5.1 - 2.15 \log_{10}(1 + 10^{((lat-27)/25)}) \]
\[ L_s = (h_R - H_0) / \sin(El(sweep) \cdot \pi/180) \% \text{ good for El > 10} \]
\[ r_p = 90/(90+4*L_s*\cos(El(sweep) \cdot \pi/180)) \]
\[ Acc = \alpha_2 \cdot L_s \cdot r_p \% \text{ dB} \]

\[ \text{if } P(rain) \leq 0.01 \% \text{ good for 0.001} \leq P \leq 0.1 \]
\[ g = 0.33; \]
\[ \text{else} \]
\[ g = 0.41; \]
\[ \text{end} \]
\[ Accp = Acc \cdot (P(rain) / 0.01)^{-g} \% \text{ dB} \]

%%% Recording the results

\[ \text{if } mode = 0 \]
\[ M(rain, 1:3) = [P(rain) \ Asam \ Accp]; \]
\[ \text{elseif } mode = 2 \]
\[ M(sweep, 1:3) = [El(sweep) \ Asam \ Accp]; \]
\[ \text{elseif } mode = 3 \]
\[ M(rain, 1:3) = [R_{sam}(rain) \ Asam \ Accp]; \]
\[ \text{elseif } mode = 1 \]
\[ M(rain, sweep) = \text{methsel} \cdot Asam + (1-\text{methsel}) \cdot Accp; \]
\[ \text{else} \]
\[ M = 0; \]
\[ \text{end} \]
\[ \text{end for sweep} \]
\[ \text{end for rain} \]

%%% Graphing

\[ \text{if } graphtoggle = 1 \]
\[ \text{if } mode = 0 \]
\[ \text{hold off} \]
\[ \text{semilogx}(M(1:length(R_{sam}), 1), M(1:length(R_{sam}), 2)) \]
\[ \text{hold on} \]
\[ \text{semilogx}(M(1:length(R_{sam}), 1), M(1:length(R_{sam}), 3), '--) \]
\[ \text{grid on} \]
\[ \text{title('Attenuation exceedance curves using SAM and CCIR methods')} \]
\[ \text{xlabel('Percent of Time (mm/h)')} \]
\[ \text{ylabel('Attenuation in dB')} \]
\[ \text{elseif } mode = 3 \]
\[ \text{hold off} \]
climate.m

function Res = climate(data, method, El, graph)
% Res = climate(data, method, El, graph)
% Generates the climate specific exceedance curves
% Asks for the desired rain data set -- CCIR(0) or Global(1)
% Asks for the desired calculation method -- CCIR(0) or SAM(1)
% El is the elevation this attenuation will occur be calc for.
% graph toggles graphical output on(1) and off(0)
% Defaults are 40 degrees, SAM and Global, and graph on

if nargin ==0
    data = 1;
    method = 1;
    El = 40;
    graph = 1;
end

RainLoad;% initialize the climactic rain rate figures
clear Res;

if graph
    clf;
    semilogx(0.1,0) % to prevent “hold on” from establishing linear axes
    hold on
end

%%%%% Setup the Data/Options

if data
[a, b] = size(Global);
RainM = Global;
Res(1, 1:b) = Pgl;
else
    [a, b] = size(CCIR);
    RainM = CCIR;
    Res(1, 1:b) = Pcc;
end

if method% tailored to output of rainrate
    column=2;
else
    column=3;
end

%%%%% Calcs, Plotting
for iter = 1:a
    Temp = rainrate(RainM(iter,:), El, 0);
    Res(iter+1, :) = transpose(Temp(:, column));
    if graph
        semilogx(Res(1,:), Res(iter+1,:))
    end
end

if graph
    grid on
    xlabel('Percentage of Time')
    ylabel('Attenuation in dB')
    title('Attenuation Exceedance Curves for different Climates')
end
11.2 Golden Phone Controller Design

MaxZero.m

function [K, zero, maxpm] = MaxZero(mdelay, despm)

% MAXZERO custom function
% Finds the Primary Design with the highest lag zero frequency.
% This is the response with the slowest response.
% Max Phase Margin should be equal to Phase Margin
% when K and zero are used.
% [K, zero, maxpm] = MaxZero(mdelay, despm)
% GoldenParameters required: maxfreq, minfreq
% see also LOOPGAIN, DELFIND, DELMARGIN,

GoldenParameters

minrange= minfreq;
maxrange= maxfreq;

for base = [0 -1 -2]
    for index = flipr([log10(minrange):10^base:log10(maxrange)
                      log10(maxrange)]),
        zero = 10^index;
        [a,b,c,d,e, fh, maxpm, K] = loopgain([1 zero], [1 0 0], mdelay, 0);
        w = sqrt(zero*(1/mdelay-zero));
        maxpm = 180+degrees(-pi - w*mdelay + atan(w/zero));
        if maxpm > despm
            minrange=zero;
            break;
        else maxrange=zero;
        end
    end
end

PeakFrac.m

matchPM = 1;
Ts = 0.125;
mdelay = 0.125;
[plantnum, plantden] = pade(mdelay, 3);
Plant = tf(plantnum, plantden);
PlantD = tf(1, [1 zeros([1 ceil(mdelay/Ts)])], Ts);
fracvec = [1 0.9 0.8 0.7 0.6 0.5 0.4 0.3 0.2 0.1 0.01 0.0001];
despmvec = [65 60 55 50 45 40 35 30 25 20 15 10];

clear PM Peak WCP Settle Rise2 Risel K zero

for j = 1:length(despmvec)
    despm = despmvec(j);
disp(['Next PM -- ' num2str(despm)]);
[K(j, zero(j), maxpm(j)]) = MaxZero2(mdelay, despm);

for i = 1:length(fracvec)
    disp(['Next Frac -- ' num2str(fracvec(i))])
    if i=1
        K2(j,1) = K(j);
    else
        K2(j,i) = fraczero(K(j), zero(j), despm, fracvec(i), mdelay, Wcp, Wcg);
        % the Wcp and Wcg are to delimit the fraczero search range
        % they come from the i=1 pass...
        AnaCont = tf(K2(j,i)*[1 zero(j)*fracvec(i)], [1 0 0]);
        LoopTrans = AnaCont*Plant;
        [Gm, Pm, Wgc, Wcp] = margin(LoopTrans);
    end
end

if digital
    DigCont = c2d(AnaCont, Ts, 'prewarp', Wcp);
    DigLoop = DigCont*PlantD;
    LoopTrans = DigLoop;
    [Gm, Pm, Wgc, Wcg] = margin(LoopTrans);
    if i=1
        [dzero(:,j), dpoles(:,j), dK(:,j)] = zpkdata(LoopTrans, 'v');
    end
end

[y, t] = step(AnError, 30);
risel = 0; rise2 = 0; settle = 0;
riselfound = 0; rise2found = 0; inband = 0;
for k = 1:length(y)
    if (y(k)<0.1) & not(riselfound)
        risel = t(k);
        riselfound = 1; end
    if (y(k)<0.0233) & not(rise2found)
        rise2 = t(k);
        rise2found = 1;
        inband = 1; end
    if (abs(y(k))>0.0233) & inband
        inband = 0;
        settle = 0;
    elseif (abs(y(k))<0.0233) & not(inband)
        inband = 1;
        settle = t(k); end
end
[rnum, rden] = tfdata(feedback(1, AnaCont*Plant), 'v');
[r, x, t] = ramp(rnum, rden, 0:0.1:30);

rsettle = 0; rsetfound = 0; inband = 0;
for k = 1:length(r)
    if (abs(r(k))>0.1) & inband
        inband = 0;
        rsettle = 30;
    elseif (abs(r(k))<0.1) & not(inband)
        inband = 1;
        rsettle = t(k); end
end
Risel(j,i) = risel;
Rise2(j,i) = rise2;
Settle(j,i) = settle;
RSettle(j,i) = rsettle;
Peak(j, i) = -min(y);
PM(j, i) = Pm;
WCP(j, i) = Wcp;
end

if digital
    save data/TheDigOne2 PM Peak WCP Settle Rise2 Risel ...
    K zero dzero dK despvec fracvec RSettle K2
else
    save data/TheBigOne2 PM Peak WCP Settle Rise2 Risel ...
    K zero despvec fracvec RSettle
end

PlotPeakFrac.m

% The variables digital and matchPM must be defined in the MATLAB
% workspace before running this script.
% Digital indicates whether the step and ramp responses were done in
% discrete-time (1) or continuous-time (0)
% MatchPM is vestigial, and should be set to 1.
if matchPM
    if digital
        load data/TheDigOne2
        method = ' *digital PM*
    else
        load data/TheBigOne2
        method = ' *analog PM*
    end
else
    if digital
        load data/TheDigOne
        method = ' *digital Wcp*
    else
        load data/TheBigOne
        method = ' *analog Wcp*
    end
end

i = gcf;
method = ' ;

figure(i+1)
surf(fracvec, despvec, Peak)
view(-130, 30)
H = title(['Peak Overshoot Fraction' method]);
set(H, 'FontSize', 12)
xlabel('Fraction of MaxZero Location')
ylabel('Designed MaxZero Phase Margin')
zlabel('Overshoot Fraction')
colormap('bone')
brighten(0.5)
brighten(0.5)
print -depsc -epsi graphs/ThesisGraphs/TimePerf/CrossFreq.eps

%%%%% Contour Plots Begin Here

figure(i+7)
clf;
contourf(fracvec, despmvec, Settle, [0 1 2 3.5 4 5 6 8 10])
colormap('bone')
colorbar
brighten(0.3)
H=text(0.15,73,{' Filled Contour Plot of Settling Time (seconds)',...
  'Solid: RiseTime to -0.1 dB, Dashed: Peak Overshoot'});
set(H, 'FontSize', 12)
xlabel('Fraction of MaxZero Location')
ylabel('Designed MaxZero Phase Margin')
hold on;
[CRs2, HRs2] = contour(fracvec, despmvec, Rise2, [0.5 0.6 0.8 1 1.25 1.5 2], 'k-');
set(HRs2, 'LineWidth', 2);
clabel(CRs2, HRs2)

[CPk, HPk] = contour(fracvec, despmvec, Peak, 0.1:0.1:0.9, 'k--');
set(HPk, 'LineWidth', 1);
clabel(CPk, HPk)
print -depsc -epsi graphs/ThesisGraphs/TimePerf/Contour.eps

%%%%% Contour Plot Number Two

figure(i+8)
clf;
contourf(fracvec, despmvec, Settle, [0 1 2 3.5 4 5 6 8 10])
colormap('bone')
colorbar
brighten(0.3)
H=text(0.15,73,{' Filled Contour Plot of Settling Time (seconds)',...
  'Solid: Crossover Frequency or BandWidth (Hertz')});
set(H, 'FontSize', 12)
xlabel('Fraction of MaxZero Location')
ylabel('Designed MaxZero Phase Margin')
hold on;
[CW, HW] = contour(fracvec, despmvec, WCP/2/pi, 'k-');
set(HW, 'LineWidth', 2);
clabel(CW, HW)
print -depsc -epsi graphs/ThesisGraphs/TimePerf/Contour2.eps

%%%%% Contour Plot Number Three

figure(i+9)
clf;
contourf(fracvec, despmvec, Settle, [0 1 2 3.5 4 5 6 8 10])
colormap('bone')
colorbar
brighten(0.3)
H=text(0.15,73,{' Filled Contour Plot of Settling Time (seconds)',...
  'Dashed: Settling Time for Unit Ramp (seconds')});
set(H, 'FontSize', 12)
xlabel('Fraction of MaxZero Location')
ylabel('Designed MaxZero Phase Margin')

hold on;
[CRS, HRS] = contour(fracvec, despmvec, RSettle, [0 1 1.5 2 3 4 6 10 20 30], 'k--');
set(HRS, 'LineWidth', 2);
clabel(CRS, HRS)
print -depsc -epsi graphs/ThesisGraphs/TimePerf/Contour3.eps

zeromargin.m

function [fcvec, pmvec, fhvec, mpmvec, Kvec, zero] = ... 
    zeromargin(mdelay, graph, despm, fracvec)

% ZEROMARGIN custom function
% Golden Phone Scheme. Takes an assumed loop measurement delay and the
% desired phase margin for that delay. Finds the appropriate controller
% using the "max zero" method. Then tests lower frequency zero locations
% which are specified as an input vector of fractions (fracvec).
% Statistics for these new systems are gathered and returned. Suggestion
% to add 1.0 as one of the elements in the fracvec input.
% [fcvec, pmvec, fhvec, mpmvec, Kvec, zero] = ...
%    zeromargin(mdelay, graph, despm, fracvec)
%
% GoldenParameters required:
%  maxtime, maxfreq, minfreq, freqres
% %
% See also LOOPGAIN, DELFIND, DELMARGIN
%

GoldenParameters;
minrange= minfreq;
maxrange= maxfreq;
wexp = log10(minfreq*2*pi):freqres:log10(maxfreq*2*pi);
wvec = 10.^wexp;

disp('Initialized')

for base = [0 -1 -2]
    for index = fliplr([log10(minrange):10^base:log10(maxrange)]
        log10(maxrange)]
        zero = 10.^index;
        [a,b,c,d,e,f, maxpm, K] = loopgain([1 zero], [1 0 0], mdelay, 0);
        if maxpm > despm
            minrange=zero;
            break;
        else maxrange=zero;
        end
    end
end

disp('Found Primary Max Zero Design.')

% At this point, we have the highest frequency zero possible,
% And the lowest frequency crossover. Graphing this result below
if maxpm > despm
    timevec = 0:1e-3:maxtime;
    if graph close all; end
%
% We start testing lower frequency zeros now...

for i=1:length(fracvec)
    [a,b,c,d,e,fhvec(i),mpmvec(i),h] = ...
    loopgain([1 fracvec(i)*zero], [1 0 0], mdelay, 0);
    if mpmvec(i)<despm
        warning(['num2str(fracvec(i)) is not feasible -- margin not
achieved'])
    else
        H(i,1:length(wvec)) = freqs(a, b, wvec);
%
% Finding the crossover frequency with 55 degrees margin
% Assumes a monotonically decreasing phase after the max pm peak
%
    for j = 1:length(wvec)
        if (wvec(j)>=fhvec(i)*2*pi)&(angle(H(i,j))<((despm-180)*pi/180))
            pmvec(i) = angle(H(i,j-1))*180/pi + 180;
            Kvec(i) = 1/abs(H(i,j-1));
            fcvec(i) = wvec(j-1)/2/pi;
            break;
        end
    end
%
% Graph the result of the new 55 degree system
%
    if graph
        [a,b,c,d,e,f,g,h] = ...
        loopgain(Kvec(i)*[1 fracvec(i)*zero], [1 0 0], mdelay, graph);
        disp(num2str(fracvec(i)))
        figure(5); step(a, c, timevec)
        title(['Step Response. ' num2str(despm) ' margin; frac. input ' ...
num2str(fracvec) ' rad/s, and mdelay ' num2str(mdelay) ' s'])
        grid on
        axis([0 maxtime -0.5 1.5])
        if graph-1
            print -deps -epsi graphs/zero.ps
        end
    commonlabel(1:5, 'hold on');
    end
    end
else
    warning(['no acceptable zero in range to achieve ' num2str(despm) ' degrees p.m.'])
end

loopgain.m

function [aL, bL, sum, fc, pm, fh, maxpm, gain] = ...
    loopgain(num, den, mdelay, graph)
% LOOPGAIN custom function
% [aL, bL, sum, fc, pm, fh, maxpm, gain] = loopgain(num, den, mdelay, graph)
% num = numerator of controller TF (desc. poly form)
% den = denominator of controller TF (desc. poly form)
% mdelay = measurement delay on downlink (in seconds)
% graph = level of graphical output (0=none; 1=screen; 2=screen&print(file))
% aL = numerator of command TF
% bL = numerator of error TF
% sum = denominator of both TFs
% fc = crossover frequency (unit magnitude response)
% pm = current phase margin (for given num/den)
% fh = crossover frequency corresponding to maxpm
% maxpm = maximum possible phase margin
% gain = additional gain required to achieve maxpm

% See GoldenParameters.m for settings:
% This function requires:
% maxfreq, minfreq, freqres
% maxtime
% tdelay
% order
% delayerror
% See also LOOPHOLD, FINDSMOOTH, DELMARGIN

%%%%
%%% Control Parameters
%%%%

GoldenParameters

maxpm = -pi;% radians (initializing)
pm = -pi;% radians (initializing)
factor = 1;% placekeeper (initializing)
wh = 0;% radians (initializing)
wC = 0;% radians (initializing)
delay = 2*tdelay + mdelay; % seconds (DU1+DD1+DD2), or loop delay
wexp = log10(minfreq*2*pi):freqres:log10(maxfreq*2*pi);
% create frequency vector for freqs
timevec = 0:delay/10:maxtime; % create time vector for step, ramp

%%%%
%%% Transfer Functions
%%%%

%%%% Loop Gain
[a, b] = pade(delay, order);
aL = conv(num, a);
bL = conv(den, b);

%%%% Closed Loop Denominator
df = length(bL) - length(aL);
if df == 0
    sum = bL + aL;
elseif df > 0
    sum = bl + [zeros([1,df]) aL];
else
    sum = [zeros([1,-df]) bl] + aL;
end

%%%%% Downlink Factor of (1-delay)
[c, d] = pade(delayerror, order);
df = length(d) - length(c);
if df == 0
    diff = d - c;
elseif df > 0
    diff = d - [zeros([1,df]) c];
else
    diff = [zeros([1,-df]) d] - c;
end
dnum = conv(aL, diff);
dden = conv(sum, d);

%%%%
%%% ESTIMATE OF PHASE MARGIN STATS
%%%%

H = freqs(aL,bL, 10.^(wexp));
posmag = 1;
if angle(H(1))>0
    scrolled = 1;
else scrolled = 0;
end
for iter = 2:length(H)
    if (abs(H(iter)))<=1&posmag% detecting when gain < 0dB
        phasemargin = angle(H(iter))-2*pi*scrolled; % records the phase margin there
        wc = 10^(wexp(iter)); % crossover frequency
        posmag = 0;
    end
    diff = (angle(H(iter-1))-angle(H(iter)));% detecting angle function falling
    scrolled = scrolled + 1;% under pi, thus beginning the state where the arg is misleadingly positive.
if diff < -1.5*pi
    scrolled = scrolled - 1; % arg is misleading -- reset the maxpm.
endif
if not(scrolled)
    maxpm = max(angle(H(iter)), maxpm);
    if angle(H(iter))==maxpm
        factor = abs(H(iter));
        wh = 10^(wexp(iter)); % max margin crossover frequency
    end
end
pm = phasemargin*180/pi + 180;
maxpm = maxpm*180/pi + 180;
gain = 1/factor;
fh = wh/2/pi;
fc = wc/2/pi;

%%%%
%%% GRAPHING
%%%%

if graph

%%% LOOP FREQUENCY RESPONSE

figure(1)
freqs(aL,bL, 10.^wexp);
title('Loop Gain Frequency Response')
if graph-1
print -deps -epsi graphs/margin.ps
end

%%% CLOSED LOOP FREQUENCY RESPONSE

figure(2)
freqs(aL, sum, 10.^wexp)
hold on
subplot(212); hold on;
freqs(bL, sum, 10.^wexp)
freqs(dnum, dden, 10.^wexp)
title('Command, Error, and Downlink Frequency Responses')
hold off
subplot(212); hold off;
grid on
if graph-1
print -deps -epsi graphs/closed.ps
end

%%% STEP RESPONSES

%%% Command Response
figure(3)
subplot(311)
step(aL, sum, timevec)
title('Closed Loop Control Step Response')
xlabel(' ')
grid on

%%% Error Response
subplot(312)
step(bL, sum, timevec)
title('Closed Loop Error Step Response')
xlabel(' ')
grid on

%%% Downlink Response
subplot(313)
step(dnum, dden, timevec)
title('Downlink Step Response')
grid on
if graph-1
    print -deps -epsi graphs/step.ps
end

%%% RAMP RESPONSE

%%% Command Response
figure(4)
subplot(311)
[y,x] = ramp(1,1, timevec);
plot(timevec, y, '-.') % show ramp input
hold on
ramp(aL, sum, timevec)
title('Closed Loop Control Ramp Response')
xlabel(' ') % xlabel(' ') is commented out
grid on

%%% Error Response
subplot(312)
ramp(bL, sum, timevec)
title('Closed Loop Error Ramp Response')
xlabel(' ') % xlabel(' ') is commented out
grid on

%%% Downlink Response
subplot(313)
ramp(dnum, dden, timevec)
title('Downlink Ramp Response')
grid on
if graph-1
    print -deps -epsi graphs/ramp.ps
end
**fullphase.m**

```matlab
function [mag, phase, wvec] = fullphase(num, den)
% FULLPHASE custom function
% Unwraps the output of the ANGLE function.
% Only for functions monotonically decreasing in phase
% starting at phase = [0 -359]
% [mag, phase, wvec] = fullphase(num, den)
% GoldenParameters required: maxfreq, minfreq

GoldenParameters
wexp = log10(minfreq*2*pi):freqres:log10(maxfreq*2*pi);
wvec = 10.^wexp;
H = freqs(num,den,wvec);
if angle(H(1))>0
    scrolled = 1;
else scrolled = 0;
end
mag(1) = abs(H(1));
phase(1)=180/pi*angle(H(1))-360*scrolled;
for iter = 2:length(H)
    diff = (angle(H(iter-1))-angle(H(iter)));
    if diff < -1.5*pi % detecting angle function falling
        scrolled = scrolled + 1;% under pi, thus beginning the state
        % where the arg is misleadingly positive.
    end
    if diff > 1.5*pi% detecting angle function exceeding
        scrolled = scrolled - 1; % arg is misleading -- reset the maxpm.
    end
    mag(iter)=abs(H(iter));
    phase(iter)=180/pi*angle(H(iter))-360*scrolled;
end
```

**GoldenParameters.m**

```matlab
maxfreq = 60/2/pi;% Hz
minfreq = 0.01;% Hz
freqres = 0.01; % rad/s log scale (usually 0.01)
maxtime = 8;% seconds
timestep = le-3; % seconds
tdelay =12e-3; % seconds (DU1=DD1)
order = 3;% for pade function... if not doing the downlink, 6 can be used
T = 0.25; % seconds delay for padeapprox
range = 1:6;% pade order vector for padeapprox
delayerror = T/10;
```
fraczero.m

function Kvec = fraczero(K, zero, despm, fracvec, mdelay, Wcp, Wcg)

disp('fraczero')
GoldenParameters;
minrange= Wcp/2/pi;
maxrange= Wcg*2/2/pi;
wexp = log10(minfreq*2*pi):freqres:log10(maxfreq*2*pi);
wvec = 10.^wexp;

% We start testing lower frequency zeros now...
for i= 1:length(fracvec)
    [a,b,c,d,e,fhvec(i),mpmvec(i),h] = ...
        loopgain([1 fracvec(i)*zero], [1 0 0], mdelay, 0);
    if mpmvec(i)<despm
        warning([num2str(fracvec(i)) ' is not feasible -- margin not
achieved'])
    else
        H(i,1:length(wvec)) = freqs(a, b, wvec);
    end
end

% Finding the crossover frequency with despm degrees margin
% Assumes a monotonically decreasing phase after the max pm peak
for j = 1:length(wvec)
    if (wvec(j)>=fhvec(i)*2*pi)&(angle(H(i,j))<((despm-180)*pi/180))
        pmvec(i) = angle(H(i,j-1))*180/pi + 180;
        Kvec(i) = 1/abs(H(i,j-1));
        fcvec(i) = wvec(j-1)/2/pi;
        break;
    end
end
end
11.3 Ephemeris Data

**UpPath.m**

```matlab
load data/Ephemeris
load data/GPoly
[i j] = size(el);

%%% following code useful if the data is not saved in above files.
% path1 = polyval(G1, vec1);
% path2 = polyval(G2, vec2);
% path1a = polyval(G1, points1);
% path2a = polyval(G2, points2);

% for i = 1:i
% for j = 1:j
% t = el(i,j);
% pathloss(i,j) = -20*log10(dist(i,j)*1000);
% if t < 49.65
%    pathgain(i,j) = polyval(G1, t);
% else
%    pathgain(i,j) = polyval(G2, t);
% end
% end
% end

% save data/PathLG pathgain pathloss
load data/PathLG
A = asin(6366/7780);
Dhorizon = 7780*cos(A);
PGhorizon = polyval(G1, 0);
PLhorizon = -20*log10(Dhorizon*1000);

A = asin(6366/7780*sin(radians(100)));
elevationl0 = sqrt((6366)^2 + (7780)^2 - 2*6366*7780*cos(radians(80)-A));
PG10 = polyval(G1, 10);
PL10 = -20*log10(elevationl0*1000);

figure(1)
clea;
hold on
plot(points1, path1a,'o');
plot(points2, path2a,'s');
plot(vec1, path1,'-');
plot(vec2, path2,'-');
hold off
grid on
legend('polynomial 1','polynomial 2',0)
title('Polynomial Fit of Path Gain versus Satellite Elevation')
xlabel('Satellite Elevation (degrees)')
ylabel('Path Gain (dB)')
print -depsc -epsi graphs/PathGainEl.eps
```

figure(3)
clea;

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plot(time_m, pathgain(1,:), '-','...
  time_m, pathgain(2,:), '-','...
  time_m, pathgain(3,:), '-','...
  time_m, pathgain(4,:), '-','...
  time_m, pathgain(5,:), '-','...
  time_m, pathgain(6,:), '-'
grid on
hold on
plot(time_m, PGhorizon*ones([1 length(time_m)]), 'kx')
plot(time_m, PG10*ones([1 length(time_m)]), 'k-')
title('Total Path Gain over the Course of Satellite Sweep')
ylabel('Path Gain (dB)')
xlabel('Time (min)')
axis([0 20 floor(PGhorizon)-1 ceil(max(max(pathgain)))+1])
legend( [num2str(phi_d(1)) ' deg'],..., [num2str(phi_d(2)) ' deg'],..., [num2str(phi_d(3)) ' deg'],..., [num2str(phi_d(4)) ' deg'],..., [num2str(phi_d(5)) ' deg'],..., [num2str(phi_d(6)) ' deg'])
print -depsc -epsi graphs/PathGain.eps

figure(4)
clf;
plot(time_m, pathloss(1,:), '-','...
  time_m, pathloss(2,:), '-','...
  time_m, pathloss(3,:), '-','...
  time_m, pathloss(4,:), '-','...
  time_m, pathloss(5,:), '-','...
  time_m, pathloss(6,:), '-'
grid on
hold on
plot(time_m, PLhorizon*ones([1 length(time_m)]), 'kx')
plot(time_m, PL10*ones([1 length(time_m)]), 'k-')
title('Path Loss resulting from Distance Travelled')
ylabel('Path Loss (dB)')
xlabel('Time (min)')
axis([0 20 floor(PLhorizon)-1 ceil(max(max(pathloss)))+1])
legend( [num2str(phi_d(1)) ' deg'],..., [num2str(phi_d(2)) ' deg'],..., [num2str(phi_d(3)) ' deg'],..., [num2str(phi_d(4)) ' deg'],..., [num2str(phi_d(5)) ' deg'],..., [num2str(phi_d(6)) ' deg'])
print -depsc -epsi graphs/PathLoss.eps

load data/Transponder
rand('state', J)
clear TrF Tr T

AvgTime = 4;
Tspeed = 4; % samples per minute of Transponder
J = rand('state');
T = round(rand(1, Tspeed*time_m(length(time_m))+AvgTime).*2.-1);
Tr(1) = 0;
for i = 1:length(T);
    Tr(i+1) = T(i) + Tr(i);
end

TrL = filter(1/AvgTime*[ones([1 AvgTime]), 1, Tr), 1, Tr);
TrF = 123.*TrL(AvgTime+1:length(TrL));
TrF2 = [TrF(1) interp(TrF(2:length(TrF)), 15)];

PhiIndex = 2;
for i = 1:length(time_s)
    UpGain(i) = TrF2(i) + pathgain(PhiIndex,i);
end

figure(5)
clear
plot((0:length(T)-AvgTime)/Tspeed, TrF)
grid on
title('Example Transponder Profile Over Satellite Sweep (Random)')
axis([0 20 floor(min(TrF))-1 ceil(max(TrF))+1])
ylabel('Path Loss (dB)')
xlabel('Time (min)')
print -depse -epsi graphs/Transponder.eps

figure(6)
clear
hold on
plot(time_m, -UpGain)
plot(time_m, -123.-pathgain(PhiIndex,:), '--')
grid on
title('Predicted Gain Required to make Satellite Output Constant')
axis([0 20 floor(min(-UpGain))-1 ceil(max(-UpGain))+1])
ylabel('Gateway Gain Required (dB)')
xlabel('Time (min)')
legend('With Transponder Gain Profile', 'With Constant Transponder of 123 dB', 0)
print -depse -epsi graphs/UpGain.eps

Ephemeris.m

%%% the following code is useful if the Ephemeris data set has not yet
%%% been saved in the file below

%time_m =0:1/60:20;
%time_s =60*time_m;
%time_over =10*60;
%phi_d = [0 5 10 15 20 25];

%for i = 1:length(phi_d)
%    phi = radians(phi_d(i));
%    [dist(i,:), el(i,:), theta] = ephemeris(time_s, time_over, phi);
%end
%save data/Ephemeris dist el time_m time_s time_over phi_d

load data/Ephemeris

figure(1)
cif;
plot(time_m, el(l,:), 'b-', time_m, el(2,:), 'g-', time_m, el(3,:), 'r-', time_m, el(4,:), 'c-', time_m, el(5,:), 'k-', time_m, el(6,:), 'm-')
axis([0 20 0 90])
grid on
hold on
plot(time_m, 10*ones([1 length(time_m)]), 'k-')
title(['Satellite Elevation over Sweep' ...
   ' for various Orbital Plane Differences (phi)'])
ylabel('Satellite Elevation (degrees)')
xlabel('Time (min)')
legend( [num2str(phi_d(l)) ' deg'], ...
   [num2str(phi_d(2)) ' deg'], ...
   [num2str(phi_d(3)) ' deg'], ...
   [num2str(phi_d(4)) ' deg'], ...
   [num2str(phi_d(5)) ' deg'], ...
   [num2str(phi_d(6)) ' deg'], ...
   'contact @ 10 deg')
hold off

figure(2)
cif;
hold on;
plot(time_m, dist(l,:), 'r-', time_m, dist(2,:), 'g--', time_m, dist(3,:), 'b-.', time_m, dist(4,:), 'c-', time_m, dist(5,:), 'k-', time_m, dist(6,:), 'm--')
grid on
A = asin(6366/7780);
Dhorizon = 7780*cos(A);
A = asin(6366/7780*sin(radians(100)));
elevation10 = sqrt((6366)^2 + (7780)^2 - 2*6366*7780*cos(radians(80)-A));
axis([0 20 floor(min(min(dist))/1000)*1000 (ceil(Dhorizon/250)+1)*250])
plot(time_m, Dhorizon*ones([1 length(time_m)]), 'kx')
plot(time_m, elevation10*ones([1 length(time_m)]), 'k-')
title(['Distance to Satellite over Sweep' ...
   ' for various Orbital Plane Differences (phi)'])
ylabel('Distance to Satellite (km)')
xlabel('Time (min)')
legend( [num2str(phi_d(l)) ' deg'], ...)
ephemeris.m

function [dist, el, theta] = ephemeris(t_vec, t_satover, phi)

h_obs = 0; % km
R_earth = 6438; % km
R_sat = 1414 + R_earth; % km
R_obs = h_obs + R_earth; % km
M_earth = 5.98e24; % kg
G = 6.67e-11; % Nm^2/kg^2
v = sqrt(G*M_earth/(R_earth*1000));
w = v/(R_sat*1000);

for t = 1:length(t_vec)
    theta(t) = w*(t - t_satover);
    dist(t) = sqrt(R_obs^2 + R_sat^2 - 2*R_obs*R_sat*cos(theta(t))*cos(phi));
    el(t) = degrees(acos((R_obs^2 + dist(t)^2 - R_sat^2)/(2*dist(t)*R_obs))) - 90;
end
11.4 Loading Simulation

**ThsLoad.m**

```matlab
endtime = 20;% simulation end time
SysStep = 0.005;% continuous time signals time step

ConFreq = 8;% Controller Sampling Frequency Hz
ConStep = 1/ConFreq;% Controller Time Step s
OpenStep= 0.1;% Open Loop Sample Period
ErrStep = ConStep;

% Designing System in Analog Domain
%load data/TF-A-Frac0375% QuickLoad frac=0.375
%load data/TF-A-Frac1000% QuickLoad frac=1
load data/NominalCont% QuickLoad final design

mdelay = 0.25; % The Long way
frac = 1;
despm = 55;
%[fcvec, pmvec, fhvec, mpmvec, Kvec, zero] = ...
zeromargin(mdelay, 0, despm, frac)

%%% requires Kvec, frac, zero, fcvec to be defined after this point

% Converting to Digital Domain
ConStep = 1/ConFreq;
[numd,dend] = bilinear(Kvec*[1 frac*zero], [1 0 0], ConFreq, fcvec);

% Direct Digital Design
%numd = [1 -0.9 0];% Pole Cancellation method
dend = [1 -2 1];

ExtraGain = 1;

GPnum = [1];
GPden = [0.0531 1];
%GPden = 1;

DownNoise = 0.5;
UpNoise = 4;
UpToggle = 1;
DownToggle = 1;

aanum = 1;
aaden = 1;
%aaden = [1/(2*pi*3) 1];

SaturationUp = 10;
SaturationDown = -10;
```
off = [0; 0];
endtime = 0;

stepf = [0; 0; 10-SysStep0; 10 -1; -1];

stdy = [1; 1];

PropDel= ceil(0.006/SysStep)*SysStep;

PilotTime = 1;
PilotStep = -1;
SpecTime = 5;
SpecStep = 0;

G_GW = off;
G_UTx = off;
L_PU = off;
L_RU = [0 0 5 0 10 -1 20 -1];

D_Ul = stdy(:,1) * PropDel.*stdy(:,2);

% the reason for above is to make prop. delay greater than SysStep, 
% and an integer multiple of SysStep

G_URx = off;
G_TR = off;
P_DTx = off;
L_PD = off;
L_RD = [L_RU(:,1) 0.2*L_RU(:,2)];
D_D1 = D_Ul;
G_DRx = off;
% D_D2 = mdelay;
% D_D2 = mdelay - 2*PropDel;
D_D2 = mdelay - 2*PropDel - 3*GPden(1);

L_PD_est = off;
G_DTx_est = off;
D_D1_est = D_D1;
G_URx_est = off;
G_TR_est = off;
D_Ul_est = D_Ul;
L_PU_est = off;
G_DRx_est = off;
D_D2_est = D_D2;

P_user = off;
P_userReq = off;
12 References


29 Various authors. RSSI Filter Description. August 1997. QUALCOMM Proprietary.