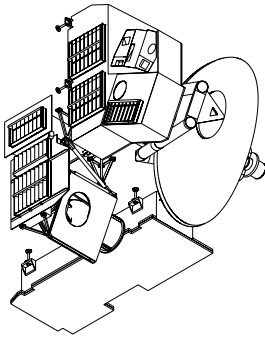


JPL D-15210
EOS MLS DRL 601 (part 2)
ATBD-MLS-02

Earth Observing System (EOS)

Microwave Limb Sounder (MLS)

EOS MLS Level 1 Data Processing Algorithm Theoretical Basis



$$\dot{\mathbf{P}}_i^L = \frac{1}{\eta_r^{ML}} \left(\frac{C_i^L - \hat{C}_i^S(L)}{\hat{g}_i(L)} + \eta_r^{MS} \dot{\mathbf{P}}_r^S - (1 - \eta_r^{ML}) \dot{\mathbf{P}}_r^{BL} + (1 - \eta_r^{MS}) \dot{\mathbf{P}}_r^{BS} \right)$$

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15 October 1999

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Release Record

Version	Date	Comments
1.0	15 Jan 1999	Initial version
1.1	15 Oct 1999	<p>Released following formal review of Version 1.0 by NASA board reviewing the EOS CHEM Algorithm Theoretical Basis Documents. This document received the top grade of ‘A’ from the review board. The board recommendations which apply to this document are italicized below, followed by the (non-italicized) responses of the MLS team.</p> <ul style="list-style-type: none"> • <i>Investigate image enhancement methods for possible implementation in Level 1 processing ...</i> This was investigated during the design of EOS MLS, and our early investigations indicated that the appropriate place in data processing to implement resolution enhancement is Level 2 (retrieval), not Level 1. • <i>A more complete (compared to that described in the ATBD’s) validation plan should be developed for the experiment. Particular emphasis should be on self-consistency plans and in-orbit calibration and performance monitoring.</i> This is planned, and will be documented in the MLS Calibration Plan. See [4] for examples of this kind of activity on UARS MLS. • <i>The performance of the electronics unique to EOS MLS should be assessed.</i> This is planned as part of instrument development, and will be documented in the MLS Functional Verification Plan. <p>No changes to the document were needed as a result of these recommendations. Changes to Version 1.0 are described below, and reflect expected progress and correction of a few minor errors:</p> <ul style="list-style-type: none"> • Units for rotation matrices in Table 4.2 corrected. • Equations (4.13) and (4.15) giving atmospheric limb radiance corrected (radiance offset components were not being divided by antenna reflectivity, ρ). • Clarification added preceding Equations (4.16) as to why Level 1 processing treats radiometric calibration as a double sideband calculation, even though earlier formulations (e.g., Equation (4.5)) are given single sideband. • Minor changes to correct grammatical and spelling errors, or clarify the meaning of some statements.

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Chapter 1

Introduction

EOS MLS measures thermal limb emission from molecules and radicals of special interest in the stratosphere and upper troposphere, with some measurements extending into the mesosphere. By measuring emission spectra as a function of limb altitude it is possible during ground processing to infer concentration profiles of various species including ClO, BrO, HNO₃, N₂O, CO, H₂O, H₂O₂, HO₂, O₃, OH, HCl, HCN and SO₂. Measurement of O₂ emission provides pressure and temperature data necessary for accurate constituent retrievals, and the spacecraft inertial reference unit (IRU) provides the pointing reference necessary for measurement of geopotential height. An overview of the EOS MLS experiment is given in [1].

The EOS project has defined several ‘Levels’ of data, the definitions below taken from the MTPE EOS REFERENCE HANDBOOK (1995 EDITION)[2]:

Level 0: *Reconstructed, unprocessed instrument/payload data at full resolution; any and all communications artifacts, e.g., synchronization frames, communications headers, duplicate data removed.*

Level 1A: *Reconstructed, unprocessed instrument data at full resolution, time-referenced, and annotated with ancillary information, including radiometric and geometric calibration coefficients and georeferencing parameters, e.g., platform ephemeris, computed and appended but not applied to the Level 0 data.*

Level 1B: *Level 1A data that have been processed to sensor units.* For MLS these include calibrated instrument radiances and related (e.g., instrument engineering) data.

Level 2: *Derived geophysical variables at the same resolution and location as the Level 1 source data.*

Level 3: *Variables mapped on uniform space-time grid scales, usually with some completeness and consistency.*

Level 4: *Model output or results from analyses of lower level data, e.g., variables derived from multiple measurements.*

This document gives the theoretical basis for algorithms used in the EOS MLS Level 1 processing software (i.e., the software used to generate daily Level 1B data files from Level 0 and ancillary data input). The primary tasks of this software are to:

1. Qualify each data quantity using instrument configuration and checksum data as well as data transmission quality flags. Apply statistical tests for data quality and “reasonableness.”
2. Calibrate the instrument engineering data (e.g., voltages, currents, temperatures and encoder angles).
3. Interpolate filter channel reference measurements onto the times of each data integration and difference the interpolates from the measurements.¹
4. Interpolate each filter channel radiometric gain to the times of limb measurements.
5. Determine the total signal power analyzed by each Digital Autocorrelator Spectrometer (DACS) during each data integration.
6. Convert each DACS data integration from an autocorrelation measurement in the time domain into a spectral measurement in the frequency domain.
7. Estimate separately the spectrally smoothly-varying and spectrally-averaged components of the limb port signal arising from antenna emission and scattering effects.
8. Calibrate the limb radiances, including corrections for antenna transmission and radiation offsets. This calculation also provides an estimate of the random component of uncertainty (noise) on each limb radiance.
9. Combine spacecraft inertial pointing and star tracker data with spacecraft and GHz antenna structural/thermal data and scan mechanism encoder data to estimate the boresight angles for each radiometer.
10. Collect and generate ancillary data (e.g., tangent point location, local solar time, local solar zenith angle, flags for bright objects in field of view) which are needed in Level 2 processing.
11. Produce a Log file summarizing instrument performance and outputs.

This software processes the Level 0 product into the Level 1B data set used as the primary input to retrieval (Level 2) software. Level 1 software also processes and/or generates additional ancillary data not included in the instrument Level 0 stream, but which are needed at Level 2. These data include tangent point locations and solar zenith angles at the tangent points.

The list below indicates the data products for which the algorithms described here will be used:

1. Calibrated limb radiances (and uncertainties) for all channels.
2. Estimates of instrument spectral baseline.

¹The reference view for limb measurements is nominally the space view, but the processing software also allows the ambient calibration target views to be used as the primary reference. For clarity, use of the space reference view is assumed in the rest of this document.

3. Calibrated engineering data (e.g., voltages, currents and temperatures).
4. System noise temperatures.
5. Reference χ^2 .
6. Interpolated channel radiometric gains.

Radiances, including spectral baseline, are written to the daily Radiance File, and the engineering and related diagnostics data (such as the noise temperatures, reference χ^2 and radiometric gains) to the Engineering File.

1.1 The ATBD review

A Peer Review for the main ATBDs of the EOS Chem I experiments was held on 18 May 1999 at the Goddard Space Flight Center (GSFC). All EOS MLS ATBDs were well received (receiving the top grade of 'A'), and the board recommendations for the Level 1 ATBD (this document), together with responses, are given in the release record at the beginning of this document.

Chapter 2

Overview and heritage

2.1 Overview of EOS MLS Level 1 data processing

Level 1 software provides the initial steps in the processing of EOS MLS instrument and ancillary data. The inputs and outputs of this process are illustrated in Figure 2.1. Input data to Level 1 processing software are the “raw” uncalibrated instrument spectrometer and engineering data, spacecraft ancillary data, startup information, and command and parameter files. The command files provide the mechanism for informing the processing software of the sources of input data and any other details of the processing to be performed, while the user inputs represent additional command line (runstream) parameters. The parameter files contain calibration data and conversion parameters necessary to convert the raw data into calibrated output quantities. The startup file (if available) contains data structures created at the end of processing of the previous contiguous data set which serve to provide continuity at the startup of a new processing run.

The primary output file used as input to Level 2 processing contains the calibrated limb radiances and uncertainties together with any additional data (such as pointing and baseline information) needed by that software. Additional output files contain calibrated engineering and performance diagnostic data, logging data which provides a “quick look” of instrument and software operation, and a termination file which serves as the startup file for a subsequent data processing run.

2.2 Heritage

The EOS MLS instrument is a direct descendent of UARS MLS, with increased and enhanced spectral coverage by virtue of:

1. a greater number of radiometers (5 unique center frequencies versus 3),
2. a larger number (19 versus 6) standard filter banks with broader spectral coverage ($\sim 1,300$ MHz versus ~ 500 MHz),
3. the addition of 5 mid-band filter spectrometers with ~ 200 MHz bandwidth,
4. 12 additional broad individual filter channels in the IF passbands of the 118 and 240 GHz radiometers, and

5. the addition of 4 Digital Autocorrelator Spectrometers (DACS) with ~ 10 MHz bandwidth and 0.2 MHz resolution to complement selected standard filter banks.

In addition to the increased complement of radiometers, their IF bandwidths range from 7 to 13 GHz, substantially larger in both absolute and relative terms than the 0.5 to 3 GHz bandwidths of the UARS radiometers. The data sampling rate of 6 Hz combined with continuous scanning and narrower fields of view at 240 and 640 GHz for EOS MLS provides substantially improved vertical resolution compared to the step-scanned implementation with 0.5 Hz sampling on UARS.

The front-end mixers are all Schottky diode implementations as before, but a significant enhancement in terms of reliability comes from the use of monolithic diodes in all front-end mixers, and a significant reduction in signal/LO diplexing complexity, together with the substantially larger IF bandwidths, arises from the use of dual-diode subharmonically pumped mixer implementations. The antenna/telescope for the GHz measurements is based on the design flown on UARS and is very similar in external dimensions.

In-flight radiometric calibration of the GHz radiometers is performed using a Switching Mirror which directs the radiometer FOVs sequentially to “cold space” and ambient target references, as on UARS. For the THz radiometers a single scanning reflector serves for both limb scanning and radiometric calibration. Scanning of the both GHz and THz FOVs through the atmospheric limb is continuous, but the data digitization implementation in the filter spectrometers is derived directly from the UARS design. This allows the radiance calibration algorithms for EOS MLS to be directly descended from, and very similar to, those used in UARS MLS.

Engineering data are acquired using circuitry based on the design concepts used in UARS MLS with minor enhancements to increase data resolution, allowing slightly modified ver-

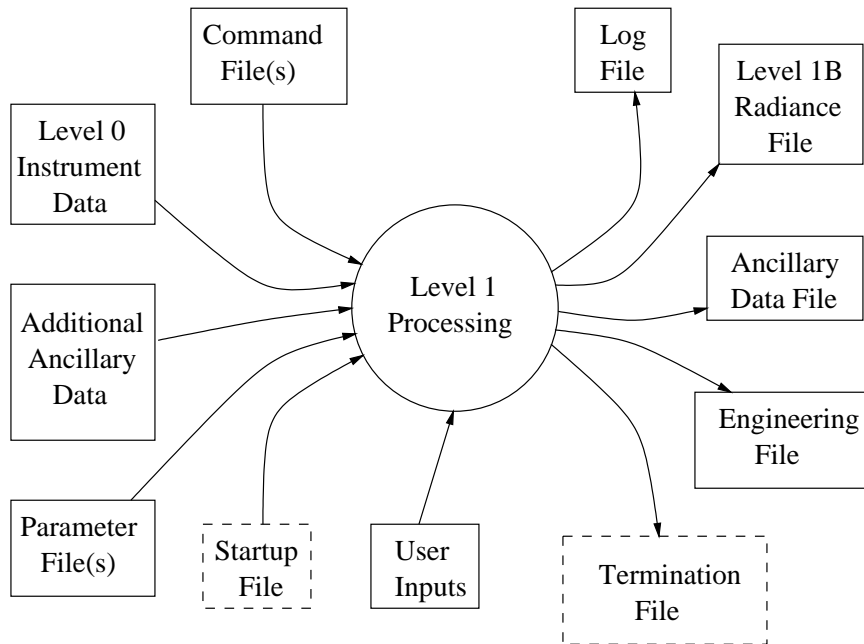


Figure 2.1: High level dataflow diagram for Level 1 data processing.

sions of the UARS processing algorithms to be used for calibrating these data, and identical algorithms to be used for converting the calibrated data into engineering units.

The considerable experience obtained with the processing of data from, and operation of, UARS MLS over a 7 year period has had a considerable influence on the design of the processing software for this instrument. Key concepts developed for UARS MLS Level 1 processing, such as the generation of Calibration Windows (described in Appendix D), and the use of startup and termination files, are retained in EOS Level 1 software. Significant differences between the UARS and EOS MLS instruments are described in Appendix A.

Chapter 3

The EOS MLS instrument

EOS MLS views the atmospheric limb in the orbital plane with a vertical scan centered $\sim 25.4^\circ$ below the spacecraft velocity vector. The nominal atmospheric scan ranges from tangent heights of ~ 2.5 km (~ 15 km for the THz measurements) to ~ 60 km, but is fully programmable to accommodate alternate measurement strategies. Measurements are made in 5 bands with radiometers centered near 118, 190, 240, 640 GHz and 2.5 THz, called R1 through R5 respectively. The 2.5 THz radiometer resides in its own assembly referred to as the “THz module,” the other radiometers being grouped together in the larger “GHz module.” Both radiometer modules share a common Spectrometer module which houses 19 25-channel filter bank spectrometers, 12 additional broad filter channels (~ 0.5 GHz bandwidth), 5 11-channel mid-band spectrometers (~ 200 MHz bandwidth) and 4 128-channel digital autocorrelator spectrometers (DACS, 10 MHz bandwidth) reside in the GHz module. The GHz and THz modules contain separate antennæ/telescopes, scan systems and radiometric calibration targets, but their operation is synchronized via the common Command and Data Handling assembly (C&DH). A simplified signal flow block diagram of the instrument is shown in Figure 3.1.

3.1 Spectral bands

EOS MLS measures thermal emission from the Earth’s atmospheric limb in 21 spectral bands, each subdivided into 25 channels. The 21 bands (B1 to B21) are covered by seven radiometers operating at frequencies defined in Table 3.1 which also identifies the principal species affecting each band. R1 is implemented with 2 radiometers (R1A and R1B) to provide redundancy for the crucial pressure/temperature (P/T) measurements. R5 is implemented with 2 radiometers to provide additional signal-to-noise for the OH measurements, with a redundant pressure measurement band implemented on the second radiometer to provide measurement resilience. The bands enclosed in parentheses in Table 3.1 are turned off in the nominal instrument operating modes.

Additional bands are analyzed by 5 mid-band spectrometers, 12 broad (~ 500 MHz) filter channels and 4 digital autocorrelator spectrometers (DACS). The mid-band spectrometers and DACS provide additional resolution for sub-bands within spectral regions analyzed by selected 25-channel filter bank spectrometers. The placement of all spectrometers, together with representative atmospheric radiance spectra, are illustrated in Figure 3.2.

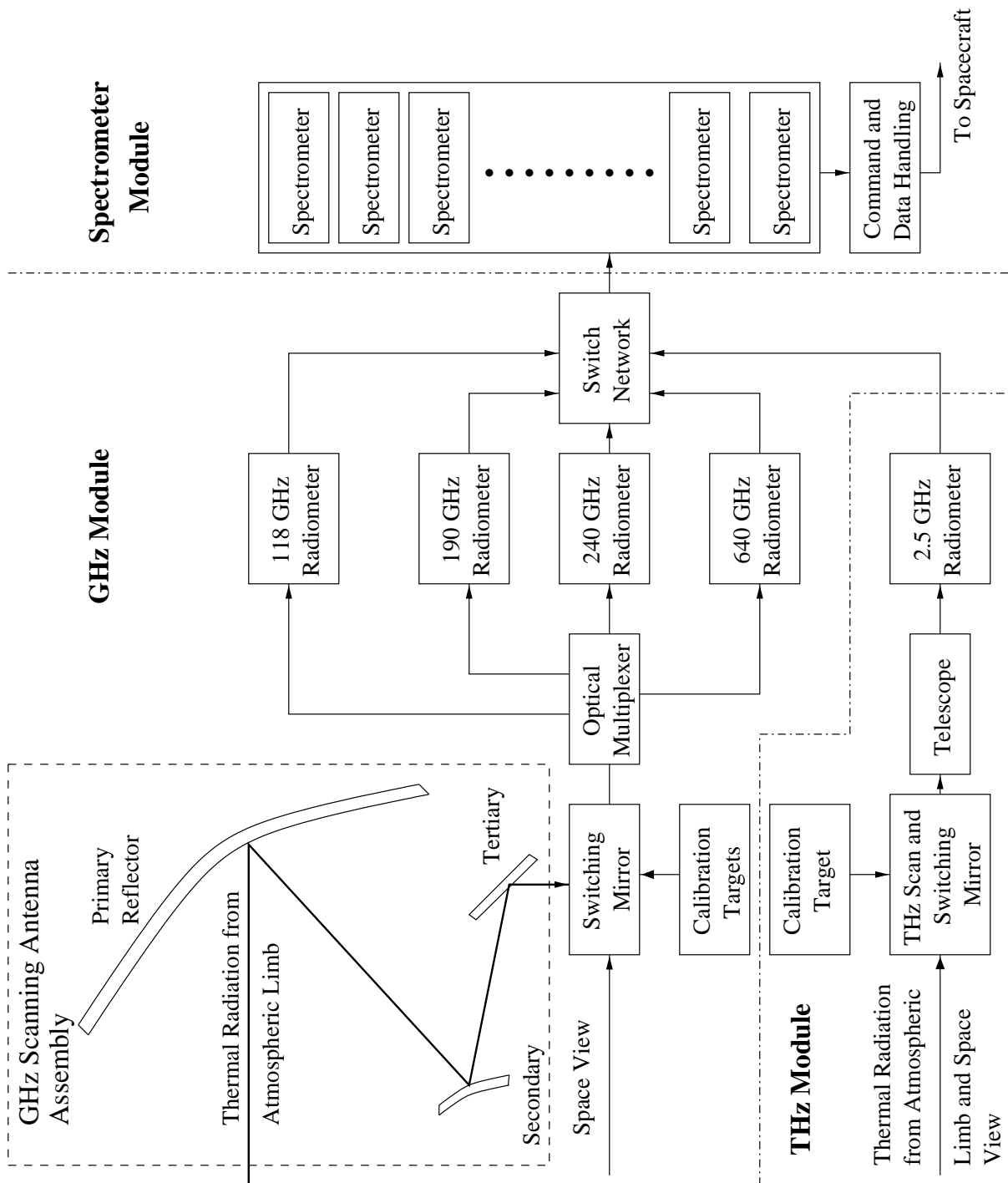


Figure 3.1: Simplified signal flow block diagram for EOS MLS. Some spectrometer channels (broad filters) are implemented within the radiometer IF downconversion chains. Redundant 118 GHz and dual 2.5 THz radiometers are not shown.

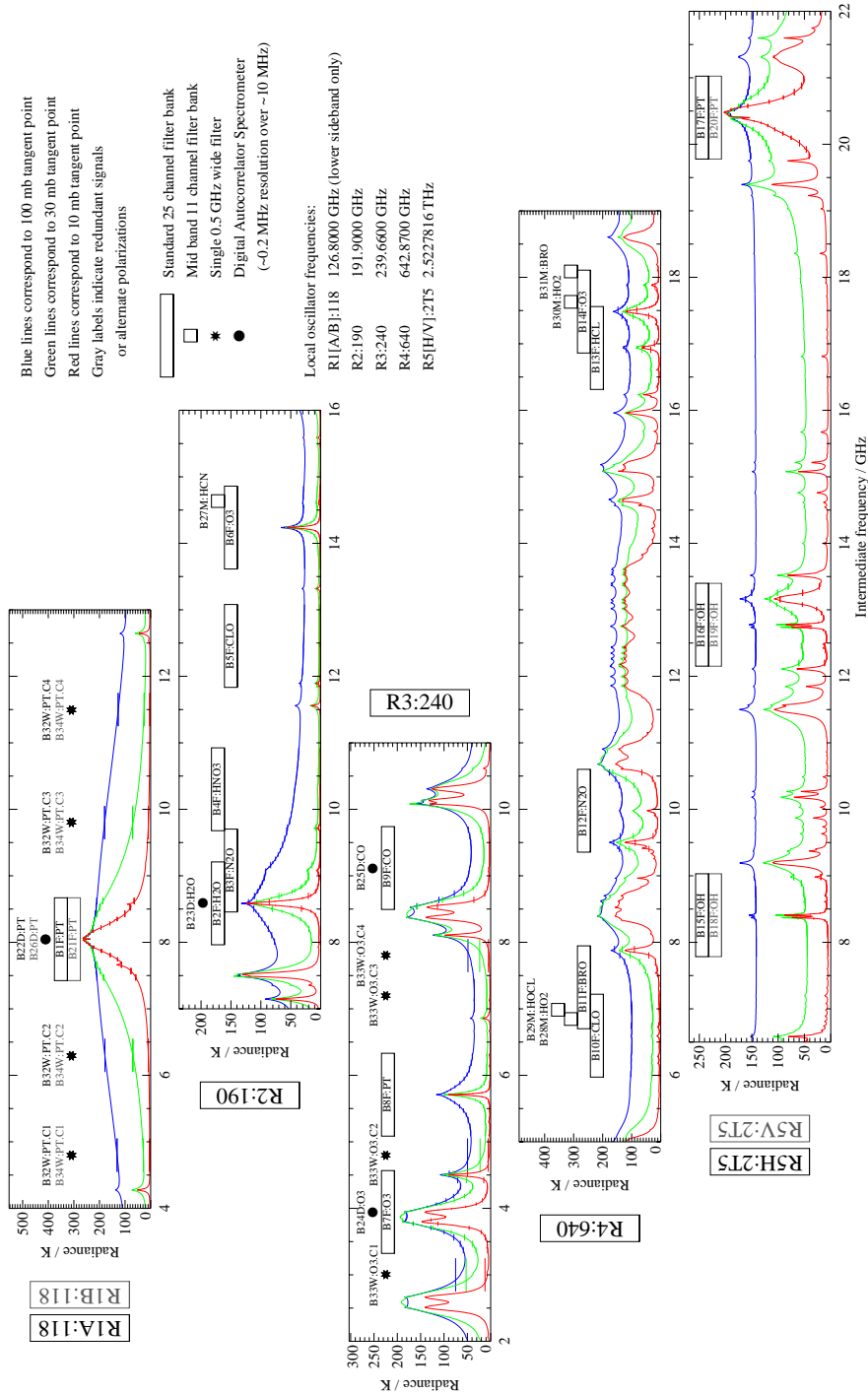


Figure 3.2: This figure shows the bands measured by EOS MLS, and the spectral coverage of all spectrometers. The abscissa of each panel indicates the IF frequency in GHz. The three spectra in each panel correspond to nominal atmospheric radiances for tangent pressures of 10, 30 and 100 mb and assume single sideband response for R1 and equal relative sideband responses all other radiometers. This figure is from N.J. Livesey.

Table 3.1: MLS Spectral Bands analyzed with 25-channel Filter Bank Spectrometers. Bands in parentheses are redundant (for measurement resiliency) and are not normally powered on. Negative IF frequencies indicate the the primary signals come from the lower sideband of the radiometer. See text for additional details.

Radiometer 1st. LO frequency	Band	IF Center Frequency (GHz)	Primary Measurement
R1 126.8000 GHz	B1 (B21)	-8.0470	O ₂
R2 191.9000 GHz	B2	-8.5858	H ₂ O
	B3	9.0798	N ₂ O
	B4	-10.3013	HNO ₃
	B5	12.4566	ClO
	B6	-14.2367	O ₃
R3 239.6600 GHz	B7	-9.1168	CO
	B8	-5.7085	P/T
	B9	-3.9449	O ₃
R4 642.8700 GHz	B10	-17.4844	O ₃
	B11	-16.9373	HCl
	B12	6.5959	ClO
	B13	7.3237	BrO
R5 2522.7816 GHz	B14	9.9785	N ₂ O
	B15	-8.4081	OH
	B18	-8.4081	OH
	B16	-12.7759	OH
	B19	-12.7759	OH
	B17 (B20)	-20.4012 -20.4012	P P

3.2 Spectrometers

EOS MLS implements 4 different “types” of spectrometer:

1. A set of 19 25-channel “standard” filterbank spectrometers,
2. A set of 5 11-channel “mid-band” filterbank spectrometers spectrometers,
3. 12 discrete, non-adjacent, non-overlapping “broad filter” channels, and
4. 4 128-channel digital autocorrelator spectrometers (DACs).

3.2.1 Standard spectrometers

The standard filter spectrometer design is based on UARS MLS, but increases the number of channels per spectrometer from 15 to 25, and the analyzed bandwidth from ~ 500 to $\sim 1,300$ MHz, while reducing mass, power consumption and volume. Details of channel relative positions and widths for both these and the mid-band filterbank spectrometers are given in Table 3.2.

Table 3.2: Positions and widths of filters in the 25-channel and 11-channel (mid-band) filter bank spectrometers. Channel positions are with respect to spectrometer center frequency which is 1,300 MHz for the standard (25 channel) spectrometers, and 200 MHz for the mid-band spectrometers. Widths are the nominal 3 dB channel widths.

channel #	position (MHz)	width (MHz)	channel #	position (MHz)	width (MHz)	channel #	position (MHz)	width (MHz)
1	-575	96	8	-79	32	19	119	48
2	-479	96	9	-51	24	20	175	64
3	-383	96	10	-31	16	21	239	64
4	-303	64	11	-17	12	22	303	64
5	-239	64	12	-7	8	23	383	96
6	-175	64	13	0	6	24	479	96
7	-119	48	14	7	8	25	575	96
			15	17	12			
			16	31	16			
			17	51	24			
			18	79	32			
			← mid-band filter bank →					
← 25-channel filter bank →								

3.2.2 Mid-band spectrometers

The mid-band spectrometers replicate the center 11 channels of a standard 25-channel filter bank spectrometer. These spectrometers are fabricated from a subset of the building blocks used to manufacture the more complex, modular 25-channel filterbank spectrometers.

3.2.3 Broad Filter channels

These are discrete single channel filters which use the detector/digitizer building block employed in the 11 and 25-channel spectrometers.

The common implementation of the back end electronics for all filter spectrometer channels allows the use of common algorithms within Level 1 software processing of all 500+ filter spectrometer channels implemented in the instrument.

3.2.4 Digital Autocorrelator Spectrometers (DACS)

The DACS implement high resolution (~ 0.2 MHz) spectrometers with uniform spectral coverage over a measurement bandwidth of 10 MHz. Their implementation and operation is described in detail in Appendix G.

The operation of a DACS is very similar conceptually to that of a Fourier Transform Interferometer (FTI). Consider Figure 3.3 which shows a simplified FTI in the upper half, and the analogous DACS below. In the FTI the band-limited input signal is split into 2 paths using a beam splitter, and recombined at the detector. As the path length is varied by altering the position of the moving reflector, the detector records the interferogram generated by the optical signals traversing the 2 paths. The path difference varies from 0 (to capture the zero

order signal when the 2 path lengths are identical, in order to determine the total power in the input signal), to a maximum path difference which defines the longest wavelength (lowest frequency) distinguishable by the measurement system. The path length differences between successive readouts of the detector define the bandwidth of the measurement system; the maximum path length difference sets the resolution. Measurement bandwidth and resolution are limited by the Nyquist criterion to wavelengths corresponding to twice the path length increment and twice the total path length difference respectively.

The DACS utilize a shift register to provide the delayed (path length altered) form of the digitized input signal, and a chain of simple multipliers generate the equivalent of the interferogram. Instead of a single “detector,” the output from each multiplier is accumulated simultaneously. This is a major advantage of the DACS compared to the FTI, since all “path length differences” are measured simultaneously with subsequent signal-to-noise benefits.

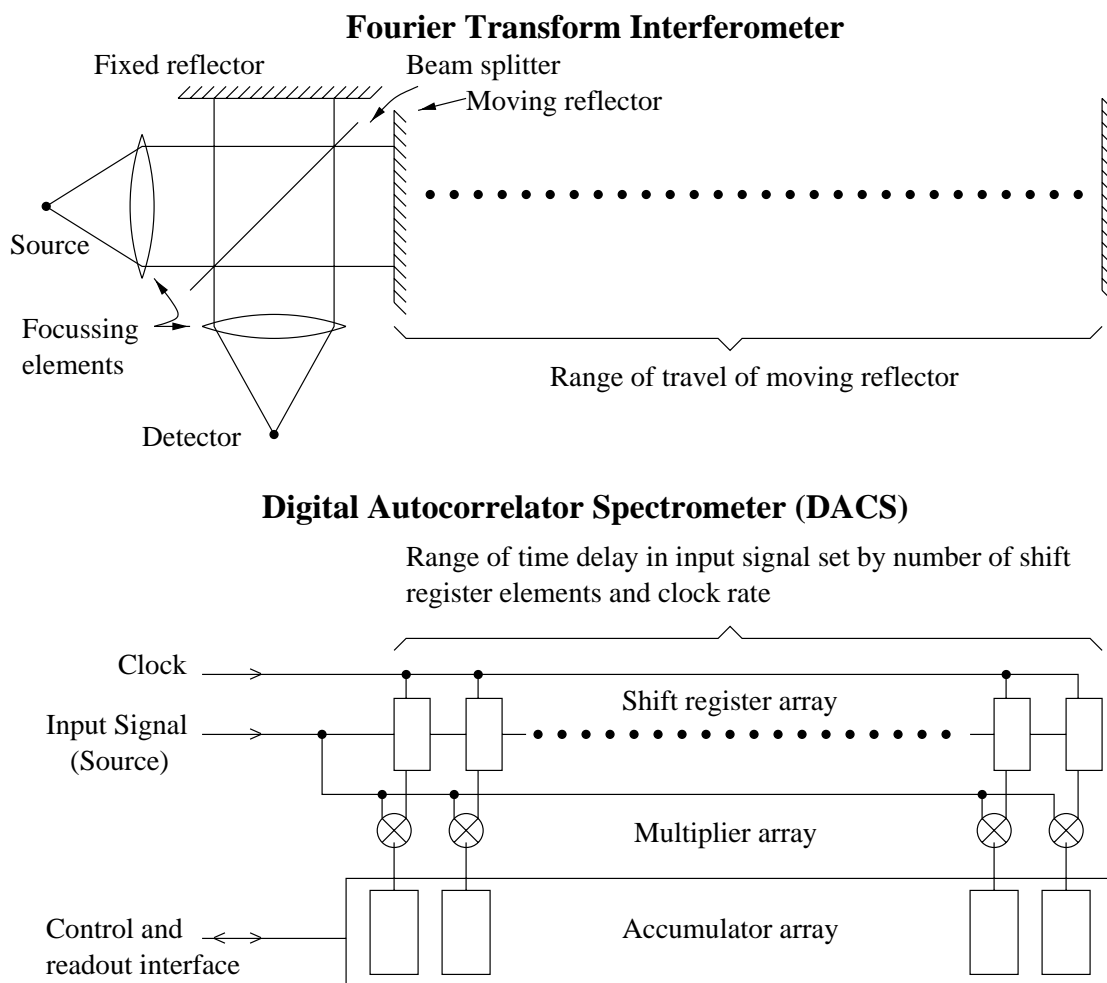


Figure 3.3: Simplified block diagram showing signal flow in a Fourier Transform Interferometer (upper portion) and a digital autocorrelator spectrometer (lower portion). Additional circuitry in the DACS (not shown) is used to measure input signal power during each data integration.

Other obvious advantages include a completely electronic implementation with no precision optical or moving parts, and high accuracy in the frequency domain for the transformed signal since the sampling clock is derived from the extremely accurate instrument Master Oscillator. At the end of a data integration period the accumulators contain the measured autocorrelation function of the input signal, the same quantity measured by the FTI as its path length is swept. The coarse quantization of the input signals by the DACS results in a distortion in the measured autocorrelation function, and a transformation (described in Appendix G) is required to obtain an estimate of the true autocorrelation function before Fourier transforming into the frequency domain.

In both measurement systems the input signal is typically band-limited to less than the bandwidth implied by the Nyquist criterion in order to enhance signal-to-noise ratio (by virtue of oversampling). In the case of the EOS MLS DACS this is also done to ease the implementation of the input band defining filter. The mathematical conversion of the measured autocorrelation function/interferogram into the frequency domain is the same for both systems, the DACS having the advantage that no additional processing and interpolation is necessary to correct for imprecision in the relative delays of the 2 signal paths, including mirror placement for the critical zero path length measurement.

The DACS are the only new element in EOS MLS requiring Level 1 processing algorithms substantially different from UARS MLS. Similar DACS with larger input bandwidths (~ 450 MHz *vs* ~ 10 MHz) but identical digitization and multiplication implementations have been procured for future use on the Submillimeter Limb Sounder (SLS) program. Every opportunity will be used to take advantage of all that is learned during the development of software for these earlier implementations.

3.3 Timing and FOV scanning

The basic internal timing events of EOS MLS are the minor and major frames (MIF and MAF) which define the integration repeat cycle and FOV atmospheric scan cycle respectively. Instrument operation over timescales of one MIF to an orbit are sketched in Figure 3.4. Spectrometer data are integrated simultaneously in all active channels during every MIF, each of which has a nominal duration of $\frac{1}{6}$ s (but is programmable over the range $\frac{1}{6}$ to $\frac{1}{3}$ s). Each MIF starts with a 5 ms dead time during which housekeeping activities (e.g. resetting counters) take place. Each MAF consists of an integer number of MIFs, and there are 240 MAFs (nominal) per orbit. The number of MAFs per orbit is programmable, but we choose an integer (and even) number of limb scans per orbit. We also require the sampled latitude bands to be the same for ascending and descending observations which forces the integer number of limb scans per orbit to be a multiple of 4. Since the orbital period will vary slightly with orbit decay and lunar gravitational influences, the synchronization of limb scans to the orbit is a software function of the Command and Data System (C&DH), with synchronization maintained by the addition or deletion of one MIF as needed at the end of each limb scan.

An important consequence of the factors (1) that limb scans are synchronized to the (variable length) orbit, and (2) that MAFs are composed of an integer number of (nominally fixed length) MIFs, is that MAFs are not of constant length, but will typically vary in length by 1 MIF. The main impact of this in Level 1 processing is that calibration measurements (i.e., Space and Target dwells) will have a slightly non-uniform distribution in time.

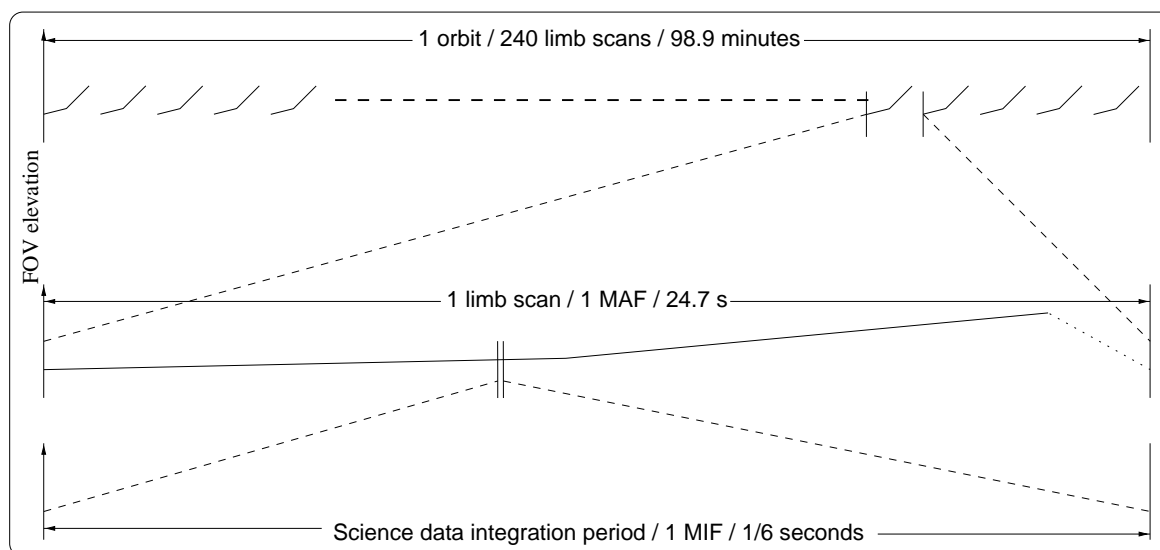


Figure 3.4: Sketch of MLS operation over timescales of an orbit, a major frame (MAF) and a minor frame (MIF). Major and minor frame boundaries are coincident, requiring an integer number of MIFs per MAF. The patterns replicated sequentially along an orbit represent limb scans (antenna angular position) and are not shown to scale. The gaps between limb scans shown in the orbit duration segment, and the dotted portion of the MAF length segment, are used to perform radiometric calibration and retrace prior to the start of the subsequent limb scan.

In-flight radiometric calibration is accomplished in the GHz module by sequencing the switching mirror through three positions as illustrated in the lower portion of Figure 3.5. This provides simultaneous radiometric calibration of all GHz bands and channels. Motion of both the antenna and switching mirror is smoother than shown in the figure to limit disturbances to the spacecraft which could adversely effect the accuracy of atmospheric measurements of MLS and other instruments on the platform. The nominal observing mode devotes ~ 1 s of each MAF to viewing the ambient calibration target, and ~ 2 s to viewing cold space. The THz module uses a single scanning mirror which serves for both atmospheric scanning and viewing the radiometric calibration references. The cold space view for this module is obtained by raising the limb FOV $\sim 1^\circ$ above the highest atmospheric view, and the target is viewed by rotating the FOV by $\sim 186^\circ$ from the nominal limb observing position, as shown in the upper portion of Figure 3.5. All mechanism activities are triggered by software command from the instrument C&DH, and the limb scanning portions of both GHz and THz radiometers are synchronized to start on the same MIF. Although GHz and THz limb scans start simultaneously, it is necessary to stagger their ends slightly due to operational constraints on peak power, ripple current and disturbance torques imposed by the spacecraft. The nominal scan profile ends the THz limb scan approximately 1 s before that of the GHz.

The GHz module contains two ambient calibration targets, one at approximately the temperature of the surrounding structure, the other designed to float to a temperature ~ 20 K lower than the structure, using a radiator for passive cooling. The floating target also contains heating elements which can be used to elevate the target temperature by ~ 20 K. This target

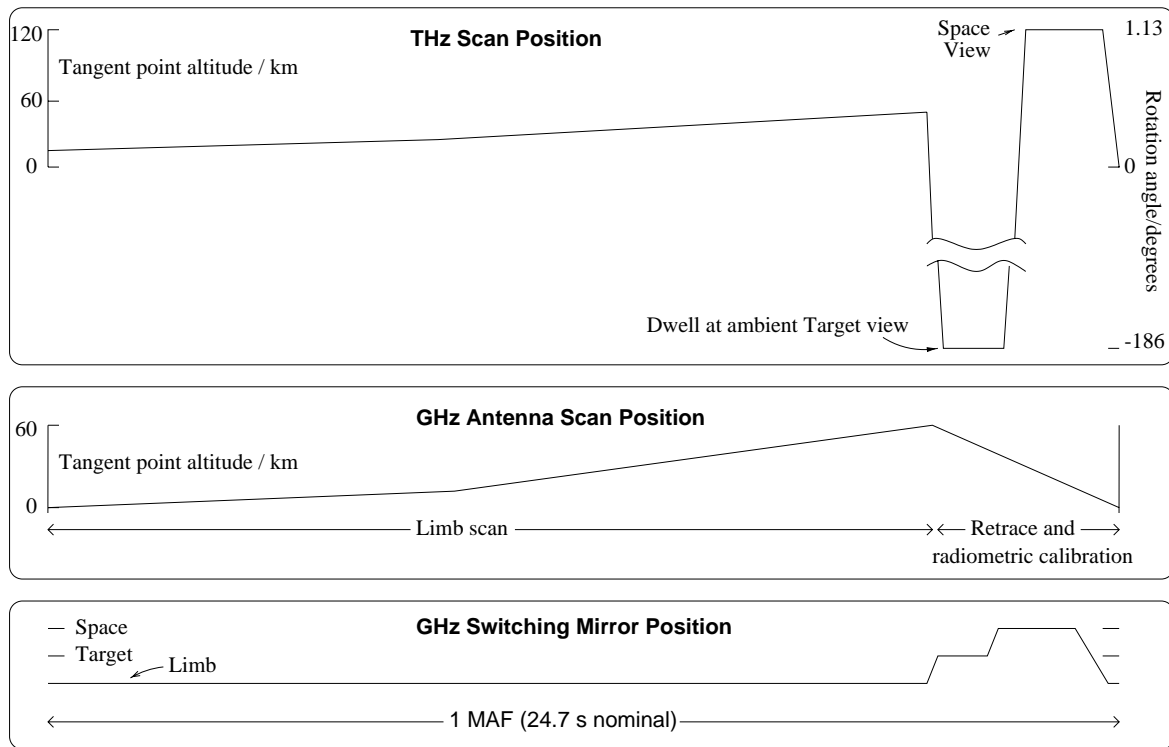


Figure 3.5: The top panel shows a nominal limb scan and calibration cycle for the THz scan mirror which also serves as the switching mirror for in-flight radiometric calibration. The center panel shows a nominal limb scan and retrace for the GHz antenna. The lower panel illustrates the use of the GHz switching mirror to provide in-flight radiometric calibration for all GHz radiometers by simultaneously directing their fields of view to an internal ambient calibration target, followed by a view to cold space.

serves to provide an “aliveness” check before launch without the need for an external target with an umbilical connection to the ground support equipment (GSE) system, removing a limitation present during portions of integration of the UARS instrument with the spacecraft. Due to packaging constraints, there is only a single calibration target in the THz module.

Chapter 4

EOS MLS Level 1 data processing algorithms

4.1 Radiances

The MLS radiometers are heterodyne systems which receive power $h\nu/\{\exp(h\nu/kT) - 1\}$ per unit frequency range when viewing a blackbody source at temperature T which completely fills their FOV, where ν is frequency, h is Planck's constant, and k is Boltzmann's constant (see Appendix B). The MLS signals originate thermally, and it is convenient to measure radiant power per unit bandwidth, \dot{P}_ν , in units of temperature so that the measure converges to the absolute temperature, T , in the long wave (Rayleigh-Jeans) limit where $h\nu \ll kT$ and classical statistical mechanics applies. The dot above a symbol indicates a measured quantity. For the blackbody this is

$$\dot{P}_\nu^{BB} = \frac{h\nu}{k\{\exp(h\nu/kT) - 1\}} \quad (4.1)$$

The long wave expansion of this expression and values (in Kelvin) of the individual terms for various temperatures are given below for $\nu = 115$ GHz, approximately the lowest frequency received by EOS MLS:

$$\begin{aligned} \dot{P}_\nu^{BB} &= T - \frac{h\nu}{2k} + \left(\frac{h\nu}{2k}\right)^2 \frac{1}{3T} \dots \\ 297.249 &= 300 - 2.760 + 0.008 \dots \\ 97.266 &= 100 - 2.760 + 0.025 \dots \\ 0.821 &= 2.7 - 2.760 + 0.940 \dots \end{aligned} \quad (4.2)$$

Temperatures of the atmospheric regions measured by the MLS, and of its internal calibration target, are within the range 150 – 300 K, for which $\dot{P}_\nu^{BB} \approx T - \delta T$, where δT is a nearly temperature independent offset, different for each radiometer, increasing in magnitude as the frequency of observation increases. This behavior is readily apparent from Figure 4.1 which plots radiance versus blackbody temperature over the range 0 to 300 K for frequencies corresponding to the nominal centers of the bandpasses of all 5 MLS radiometers. Values of δT for temperatures of 2.7, 150 and 300 K at the center frequencies of all EOS MLS radiometers are given in Table 4.1. The full formula (4.1) is used throughout processing of MLS data.

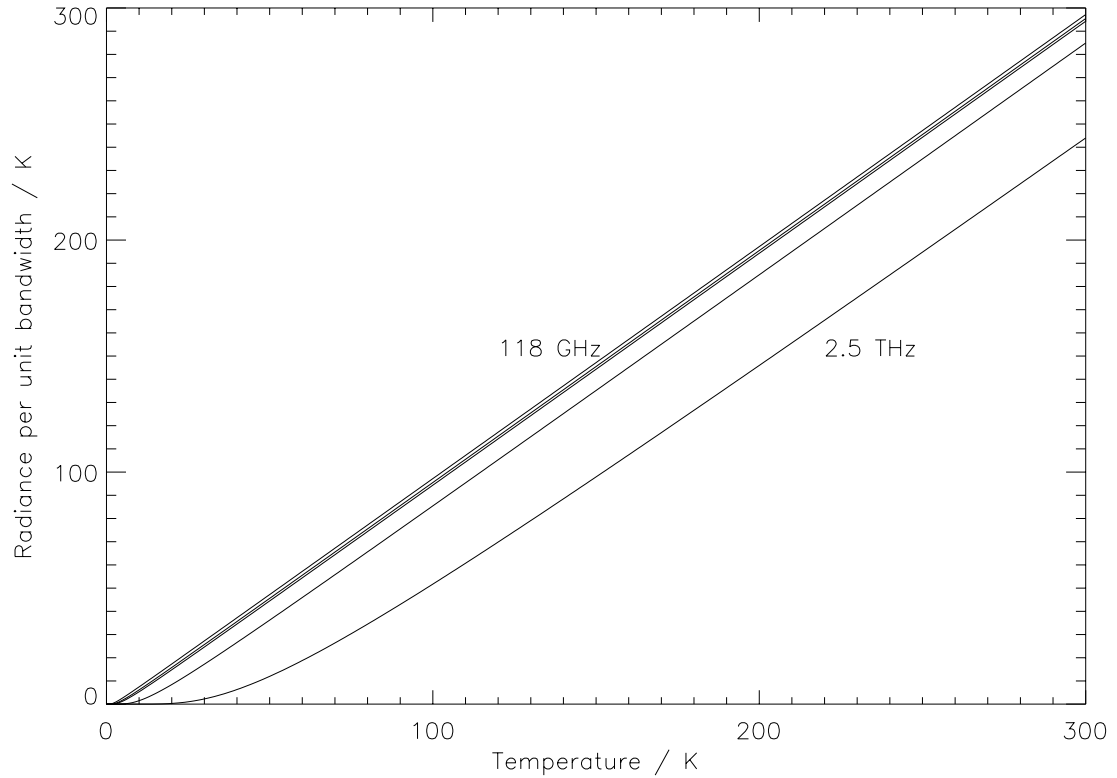


Figure 4.1: Plots of received blackbody radiance per unit bandwidth versus blackbody temperature over the temperature range 0 to 300 K at frequencies corresponding to the center frequencies of the 5 EOS MLS radiometers (118, 190, 240 and 640 GHz and 2.5 THz). The approximately linear relationship between physical temperature and radiance over the atmospheric temperature range (~ 150 to 300 K) is readily evident, as is the frequency dependent offset.

Table 4.1: This table lists the differences, δT , between physical temperature, T , and received blackbody radiance per unit bandwidth (Kelvin). δT is computed at the nominal center frequencies of all EOS MLS radiometers, and at physical temperatures of 2.7, 150 and 300 K corresponding to the physical temperatures of “cold space” and approximately the full range of observed temperatures.

ν / GHz	δT		
	$T = 2.7$ K	$T = 150$ K	$T = 300$ K
118	1.907	2.814	2.823
190	2.378	4.513	4.536
240	2.536	5.685	5.722
640	2.700	14.834	15.096
2,500	2.700	52.076	56.043

4.1.1 Radiometric signals and the calibration process

MLS measures atmospheric limb radiances captured by the GHz and THz antennæ. For both subsystems regular periodic radiometric calibration is performed in flight using a switching mirror. The operation of the 2 switching mirrors is independent, but synchronized at both the MIF and MAF level by the C&DH so that the start of limb scanning is coincident for both modules, and with slightly staggered (but overlapped) radiometric calibration sequences to meet spacecraft peak power, ripple current and disturbance torque requirements.

For both GHz and THz modules the use of a switching mirror for radiometric calibrations ensures that all radiometers within a module are calibrated concurrently, but in each case the measured radiances at the limb ports of the switching mirrors must be corrected for the FOV transformations and losses in the telescope optics. For the GHz module a 3 element offset Cassegrain telescope scans the atmosphere, directing this radiation into the limb port of the switching mirror. For the THz module the switching and atmospheric scanning mirrors are one and the same. In both cases the FOV response scanning the atmosphere will be characterized over a $\sim\pm 6^\circ$ angle from pre-launch FOV calibrations, and the much larger solid angle outside of this region will be characterized using scattering models. For the GHz FOV the $\pm 6^\circ$ region is a cone centered on the main response, while for the THz system this region is one dimensional (i.e., collapsed into a response in the vertical) because of the method used for its measurement.

Below we discuss radiometric response, effects introduced by the antenna, radiometric calibration using the switching mirror, and the impacts of noise on radiance measurements. Radiometric calibration of the DACS is discussed separately at the end of this chapter.

4.1.2 Radiometric Response

The response of radiometer filter channel i is proportional to received power \dot{P}_i^{MX} obtained by integrating the power per unit frequency and per unit solid angle, $\dot{I}_\nu^{MX}(\theta, \phi)$, incident on the switching mirror (M) from view X , where X is one of L , T or S , representing the limb, target and space view ports respectively. The integrals are evaluated over angle and frequency with weighting functions $G^M(\nu, \theta, \phi)$ and $F_i(\nu)$ which describe the angular and frequency response of the antenna and radiometer respectively:

$$\dot{P}_i^{MX} = \frac{1}{4\pi} \int_\nu \int_\Omega \dot{I}_\nu^{MX}(\theta, \phi) F_i(\nu) G^M(\nu, \theta, \phi) d\Omega d\nu \quad (4.3)$$

where $F_i(\nu)$ is normalized to unit area ($\int_\nu F_i(\nu) d\nu = 1$), $\int_\Omega G d\Omega = 4\pi$ (Appendix B), and the integrals are evaluated over the full range of frequencies and solid angles over which the instrument has a response. For the THz radiometers the switching and antenna scanning mirrors are the same device. Since EOS MLS scans continuously through the limb while integrating spectrometer outputs, the forward model used in Level 2 processing must include the effects of this smearing in its computations.

The following considerations are important in relating the radiation incident upon the switching mirror to that from the antenna, target, and space ports:

1. In the case of the GHz module, the views from the switching mirror are restricted by baffles, coated with absorbing material, which define solid angles Ω_{MX} for view X . We

define quantities \dot{P}_i^X for $X = L, T$, or S from (4.3) by restricting the integral to the solid angle Ω_{MX} :

$$\begin{aligned} \dot{P}_i^X &= r_l \times \frac{\int_0^{\nu_{LO}} \int_{\Omega_{MX}} \dot{I}_\nu^X(\theta, \phi) F_i(\nu) G_{i,l}^M(\theta, \phi) d\Omega d\nu}{\int_{\Omega_{MX}} G_{i,l}^M(\theta, \phi) d\Omega} + \\ & r_u \times \frac{\int_{\nu_{LO}}^\infty \int_{\Omega_{MX}} \dot{I}_\nu^X(\theta, \phi) F_i(\nu) G_{i,u}^M(\theta, \phi) d\Omega d\nu}{\int_{\Omega_{MX}} G_{i,u}^M(\theta, \phi) d\Omega} \end{aligned} \quad (4.4)$$

$$(\dot{I}_\nu^{MX} \equiv \dot{I}_\nu^X \text{ for directions within solid angle } \Omega_{MX})$$

where r_l and r_u are the radiometer relative responses for channel i in the lower and upper sidebands (lower sideband only for the single sideband 118 GHz radiometer), and use has been made of the fact that relative sideband response does not vary significantly over the spectral region measured by a spectrometer channel in one sideband. ν_{LO} is the frequency of the radiometer local oscillator. To reduce the complexity of Equation 4.4, and many of those which follow in this chapter, channel dependence of r_l and r_u is not called out explicitly.

2. The fractional ranges of ν over which the above integrands have significant magnitude are very small, allowing the dependence on ν to be removed from the expression for G^M when performing the integrals (i.e., constant values for ν corresponding to the center frequencies of the channel in each sideband are used when evaluating the integrals for each sideband).
3. Radiation \dot{I}_ν^X from the calibration target and from space is isotropic and has a black-body spectrum.

The restriction of the beam pattern by the baffles of the GHz switching mirror cavity is illustrated in Figure 4.2. The baffles are approximately elliptical in shape and coated with microwave absorber. They are designed to intercept the gain pattern at the same level for all ports. The most severe truncation is for the 118 GHz radiometers where the gain pattern is intercepted at the ~ 30 dB contour. This indicates that ~ 0.999 of the power in the nominal beam passes through the baffle, represented by the parameter η . With the above definitions, for the limb view,

$$\dot{P}_i^{ML} = r_l \times \left(\eta_{i,l}^{ML} \dot{P}_{i,l}^L + (1 - \eta_{i,l}^{ML}) \dot{P}_{i,l}^{BL} \right) + r_u \times \left(\eta_{i,u}^{ML} \dot{P}_{i,u}^L + (1 - \eta_{i,u}^{ML}) \dot{P}_{i,u}^{BL} \right) \quad (4.5)$$

where $\eta_{i,s}^{ML} = \frac{1}{4\pi} \int_{\Omega_{ML}} G_i^M(\theta, \phi) d\Omega$ is the fraction of the “volume” under the gain pattern which is inside Ω^{ML} , and $\dot{P}_{i,s}^{BL}$ is the radiation from the baffle in the limb view, averaged over the solid angle outside Ω_{ML} , for sideband s . $\dot{P}_{i,s}^L$ is the appropriate sideband component from the right hand side of Equation 4.4, thus accounting for the relative sideband response of the radiometer. Similarly, for the target ($X = T$) and space ($X = S$) views,

$$\dot{P}_i^{MX} = r_l \times \left(\eta_{i,l}^{MX} \dot{P}_{i,l}^X + (1 - \eta_{i,l}^{MX}) \dot{P}_{i,l}^{BX} \right) + r_u \times \left(\eta_{i,u}^{MX} \dot{P}_{i,u}^X + (1 - \eta_{i,u}^{MX}) \dot{P}_{i,u}^{BX} \right) \quad (4.6)$$

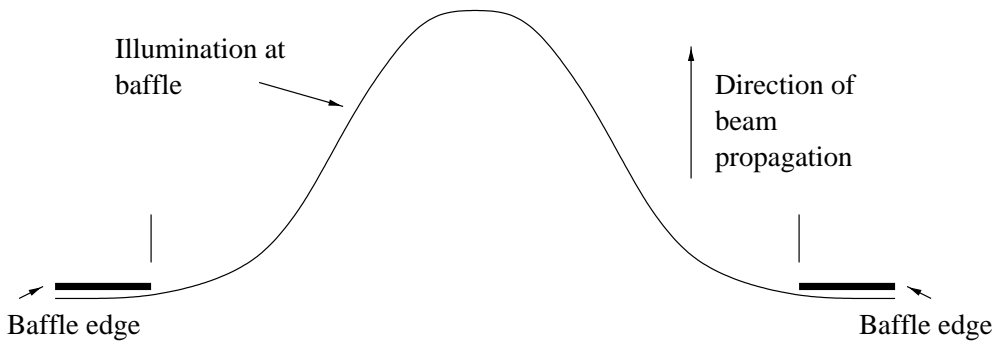


Figure 4.2: Truncation of beam pattern by the baffles of the GHz switching mirror cavity. A cross section through the center of the approximately elliptical cavity port is shown.

where $\eta_{i,s}^{MX}$ and $\dot{P}_{i,s}^{BX}$ are the corresponding values for these two views, and $\dot{P}_{i,s}^X = \dot{P}_{\nu_{i,s}}^X$.

4.1.3 Antenna Effects

The radiance intercepted by the switching mirror when viewing the limb consists of not just the desired atmospheric limb radiance, but also small components arising from antenna emission, diffraction and scattering. In addition, the limb radiance itself is attenuated due to losses in the antenna system. For the THz system, which uses the same mirror for both switching (calibration) and limb scanning, the situation is slightly simpler. We discuss the more complex GHz case in detail below.

The GHz antenna, interposed between the switching mirror and the atmospheric limb, transforms the FOV of the radiometer so that when calculating the radiance collected by the antenna, \dot{P}_i^A , from the limb spectral intensity incident upon the antenna, $\dot{I}_\nu^A(\theta, \phi)$, $G_i^M(\theta, \phi)$ must be replaced by $G_i^A(\theta, \phi)$, the measured antenna gain. Imperfections in the antenna and practical limitations on its characterization result in the introduction of further terms. One limitation is that $G_i^A(\theta, \phi)$ is measured only over directions included in a solid angle Ω_A , about $\pm 6^\circ$ from the boresight axis. The calculated response for directions outside this range, and the calculated ohmic loss in the antenna, are used to estimate an effective transmission loss of the radiation received within Ω_A and an additive radiation offset from the antenna.

Antenna transmission is more easily described by considering the antenna as a transmitter with illumination function $G_i^A(\theta, \phi)$ (restricted to solid angle Ω_A). Two processes are involved:

1. Ohmic loss: the antenna transmission due to ohmic loss is ρ_r^A where a fraction $(1 - \rho_r^A)$ of the incident radiation is absorbed, and
2. diffraction and scattering: of the remainder, in any narrow frequency band a fraction $(1 - \eta^A)$ is transmitted into directions outside the solid angle Ω_A over which the antenna pattern is characterized in detail.

Antenna ohmic loss, ρ^A , varies sufficiently slowly with frequency that a single value suffices for each radiometer r . Antenna efficiency, η^A , however, has sufficient frequency dependence

that we must provide values for each channel i and sideband s , where

$$\eta_{i,s}^A = \frac{1}{4\pi} \int_{\Omega_A} G_{i,s}^A(\theta, \phi) d\Omega \quad (4.7)$$

Radiation offsets arise from the same two processes: (1) ohmic loss: radiation $(1 - \rho_r^A) \dot{P}_{i,s}^{OA}$ is emitted by the antenna, where $\dot{P}_{i,s}^{OA}$ is an effective brightness for the antenna thermal emission; (2) diffraction and scattering: radiation $(1 - \eta_{i,s}^A) \rho_r^A \dot{P}_{i,s}^{SA}$, characterized by an effective brightness $\dot{P}_{i,s}^{SA}$, is scattered outside solid angle Ω_A .

Accounting for loss and scattering as described above relates \dot{P}_i^L , the radiance at the limb port of the switching mirror, to $\dot{P}_{i,l}^{OA}$ and $\dot{P}_{i,u}^A$, the radiance collected by the antenna within solid angle Ω_A for the two sidebands:

$$\begin{aligned} \dot{P}_i^L &= \rho_r^A (r'_l \eta_{i,l}^A \dot{P}_{i,l}^A + r'_u \eta_{i,u}^A \dot{P}_{i,u}^A) + \\ & r'_l (1 - \rho_r^A) \dot{P}_{i,l}^{OA} + r'_u (1 - \rho_r^A) \dot{P}_{i,u}^{OA} + r'_l (1 - \eta_{i,l}^A) \rho_r^A \dot{P}_{i,l}^{SA} + r'_u (1 - \eta_{i,u}^A) \rho_r^A \dot{P}_{i,u}^{SA} \end{aligned} \quad (4.8)$$

where

$$\begin{aligned} r_l &= r'_l \eta_{i,l}^A \\ \text{and } r_u &= r'_u \eta_{i,u}^A \end{aligned} \quad (4.9)$$

are the lower and upper sideband responses through the entire signal chain including antenna effects, whereas r'_l and r'_u are the sideband responses for the radiometers and optics up to the radiometer port of the switching mirror. The quantities r'_l and r'_u , normalized according to

$$r'_l + r'_u = 1 \quad (4.10)$$

are measured during spectral calibration of the instrument, while $\eta_{i,s}^A$ are determined during FOV calibration.

Calculated values of \dot{P}_i^A are produced during Level 2 processing in a Forward Model (where they are designated \dot{P}_i^A) by integrating $\dot{I}_\nu^A(\theta, \phi)$ over the angular response of the antenna within solid angle Ω_A and over the frequency response of a radiometer channel, where

$$\dot{P}_i^A = \frac{\int_\nu \int_{\Omega_A} \dot{I}_\nu^A(\theta, \phi) F_i(\nu) G_i^A(\nu, \theta, \phi) d\Omega d\nu}{\int_{\Omega_A} G_i^A(\theta, \phi) d\Omega} \quad (4.11)$$

It is important to note that \dot{P}_i^A is computed separately for both sidebands of each radiometer (except for the single sideband 118 GHz radiometers), which allows the frequency dependence of G to be removed from the integrals by substituting the value for ν in the center of channel i in each measurement sideband. These components are then combined appropriately using the radiometer sideband responses to produce the Forward Model quantity \dot{P}_i^A given by

$$\begin{aligned} \dot{P}_i^A &= r_l \times \frac{\int_0^{\nu_{LO}} \int_{\Omega_A} \dot{I}_\nu^A(\theta, \phi) F_i(\nu) G_{i,l}^A(\theta, \phi) d\Omega d\nu}{\int_{\Omega_A} G_{i,l}^A(\theta, \phi) d\Omega} + \\ & r_u \times \frac{\int_{\nu_{LO}}^\infty \int_{\Omega_A} \dot{I}_\nu^A(\theta, \phi) F_i(\nu) G_{i,u}^A(\theta, \phi) d\Omega d\nu}{\int_{\Omega_A} G_{i,u}^A(\theta, \phi) d\Omega} \end{aligned} \quad (4.12)$$

The Forward Model calculations of \dot{P}_i^A include the effects of vertical smearing due to motion of the FOV during the limb data integration by including the time dependence of G and including the motion of the FOV during each data integration. This does not affect any of the derivations below and, to improve clarity in discussion of the key points, we do not include this time dependence here or later in this chapter.

The calibrated radiance estimates produced by Level 1 processing must correspond to the quantity \dot{P}_i^A in Equation 4.12 calculated by the Forward Model during Level 2 processing. The Level 1 estimate of \dot{P}_i^A is given by $\langle \dot{P}_i^A \rangle$ where

$$\begin{aligned} \langle \dot{P}_i^A \rangle &= r_l \dot{P}_{i,l}^A + r_u \dot{P}_{i,u}^A \\ &= \frac{\dot{P}_i^L}{\rho_r^A} - \left(\frac{1}{\rho_r^A} - 1 \right) (r_l' \dot{P}_{i,l}^{OA} - r_u' \dot{P}_{i,u}^{OA}) - r_l' (1 - \eta_{i,l}^A) \dot{P}_{i,l}^{SA} - r_u' (1 - \eta_{i,u}^A) \dot{P}_{i,u}^{SA} \end{aligned} \quad (4.13)$$

We now expand Equation 4.13 to explicitly include separate terms for the three reflectors in the GHz antenna system, as these will be explicitly accounted for in Level 1 processing to improve accuracy in the absolute values of calibrated MLS radiances. This expansion leads to

$$\langle \dot{P}_i^A \rangle = \frac{\dot{P}_i^L}{\rho_r^1 \rho_r^2 \rho_r^3} - \alpha_r \quad (4.14)$$

where

$$\begin{aligned} \alpha_r &= \left\{ r_l' \times \left(\rho_r^1 \rho_r^2 \rho_r^3 (1 - \eta_{i,l}^{AA}) \eta_{i,l}^1 \dot{P}_{i,l}^{SA} + (1 - \rho_r^1) \rho_r^2 \rho_r^3 \eta_{i,l}^1 \dot{P}_{i,l}^{O1} + \right. \right. \\ & \rho_r^2 \rho_r^3 (\eta_{i,l}^2 - \eta_{i,l}^1) \dot{P}_{i,l}^{S1} + (1 - \rho_r^2) \rho_r^3 \eta_{i,l}^2 \dot{P}_{i,l}^{O2} + \rho_r^3 (\eta_{i,l}^3 - \eta_{i,l}^2) \dot{P}_{i,l}^{S2} + \\ & \left. (1 - \rho_r^3) \eta_{i,l}^3 \dot{P}_{i,l}^{O3} + (1 - \eta_{i,l}^3) \dot{P}_{i,l}^{S3} \right) + \\ & r_u' \times \left(\rho_r^1 \rho_r^2 \rho_r^3 (1 - \eta_{i,u}^{AA}) \eta_{i,u}^1 \dot{P}_{i,u}^{SA} + (1 - \rho_r^1) \rho_r^2 \rho_r^3 \eta_{i,u}^1 \dot{P}_{i,u}^{O1} + \right. \\ & \left. \rho_r^2 \rho_r^3 (\eta_{i,u}^2 - \eta_{i,u}^1) \dot{P}_{i,u}^{S1} + (1 - \rho_r^2) \rho_r^3 \eta_{i,u}^2 \dot{P}_{i,u}^{O2} + \rho_r^3 (\eta_{i,u}^3 - \eta_{i,u}^2) \dot{P}_{i,u}^{S2} + \right. \\ & \left. (1 - \rho_r^3) \eta_{i,u}^3 \dot{P}_{i,u}^{O3} + (1 - \eta_{i,u}^3) \dot{P}_{i,u}^{S3} \right) \} / (\rho_r^1 \rho_r^2 \rho_r^3), \end{aligned} \quad (4.15)$$

where i,s indicates channel i , sideband s (l or u), r indicates radiometer r , and where

$$\begin{aligned} \dot{P}_{i,s}^L &= \text{limb port radiance power,} \\ \rho_r^k &= \text{reflectivity of reflector } k = 1, 2, 3 \text{ (primary, secondary, tertiary),} \\ \eta_{i,s}^{AA} &= \text{optical transmission of the antenna system: the product of scattering } (\eta_{i,s}^{AS}) \end{aligned}$$

$$\begin{aligned}
& \text{and diffraction } (\eta_{i,s}^{AD}) \text{ from the primary plane,} \\
\eta_{i,s}^k &= \text{optical transmission of reflector } k, \\
\dot{P}_{i,s}^{SA} &= \text{radiance power in the limb hemisphere but outside FOV measurement angle} \\
& \quad \Omega_A, \\
\dot{P}_{i,s}^{Sk} &= \text{radiance power illuminating the spillover solid angle for reflector } k, \text{ and} \\
\dot{P}_{i,s}^{Ok} &= \text{power thermally emitted by reflector } k.
\end{aligned}$$

The next section of this document describes how \dot{P}_i^L appearing in Equation 4.14 is obtained during in-flight radiometric calibration of the instrument. The quantities ρ_r^k , $\eta_{i,s}^{AA}$ and $\eta_{i,s}^k$ appearing in Equations 4.14 and 4.15 are produced during instrument FOV calibration. Precalculated values are used for $\dot{P}_{i,s}^{SA}$ and $\dot{P}_{i,s}^{Sk}$. $\dot{P}_{i,s}^{Ok}$ is calculated from temperatures measured by sensors on the appropriate reflectors.

4.1.4 The radiance calibration process

In-flight radiometric calibration is performed during every limb scan using the switching mirror (scanning mirror in the case of the THz module) to direct the FOVs of all radiometers to the space and internal calibration target ports. The filter bank detectors are operated at sufficiently low signal power to provide a linear relationship between input radiance and channel output, illustrated in Figure 4.3, where the abscissa indicates signal power collected by the switching mirror, and the ordinate indicates the output counts from the voltage-to-frequency (V/F) converter digitizer of an individual channel i . The discussion below applies to all filter spectrometer channels, including the single filter channel implemented in each DACS as a signal power monitor.

In the formulation below, the sideband dependence of the receivers and of the baffle transmissions (shown in Equation 4.5) is omitted, since the error introduced by this approximation is negligible.

The output of channel i for each of the three switching mirror positions is given by

$$\begin{aligned}
C_i^L &= g_i \left(\eta_r^{ML} \dot{P}_r^L + (1 - \eta_r^{ML}) \dot{P}_r^{BL} \right) + C_i^O + C_i^Z, \\
C_i^T &= g_i \left(\eta_r^{MT} \dot{P}_r^T + (1 - \eta_r^{MT}) \dot{P}_r^{BT} \right) + C_i^O + C_i^Z, \\
C_i^S &= g_i \left(\eta_r^{MS} \dot{P}_r^S + (1 - \eta_r^{MS}) \dot{P}_r^{BS} \right) + C_i^O + C_i^Z.
\end{aligned} \tag{4.16}$$

where C_i^X are the outputs of channel i for the three switching mirror positions X . C_i^O are the counts that would be generated if observed radiances were zero (i.e., generated by radiometer and IF noise alone), and g_i is the overall radiometric gain of channel i (expressed in counts per Kelvin of signal brightness). C_i^Z are the offsets built into the spectrometer digitization system, measured nominally once per orbit to facilitate determination of system temperature (described later) as a diagnostic of radiometer performance. The measurement of C_i^Z is performed by commanding the IF gains of all radiometers to such a low value that the spectrometers see essentially no RF signal at their inputs. Although counts C_i^O are shown the same for all views in Equation 4.16, it must be noted that this offset is time dependent due to thermal drifts and noise, and that its power spectrum invariably displays

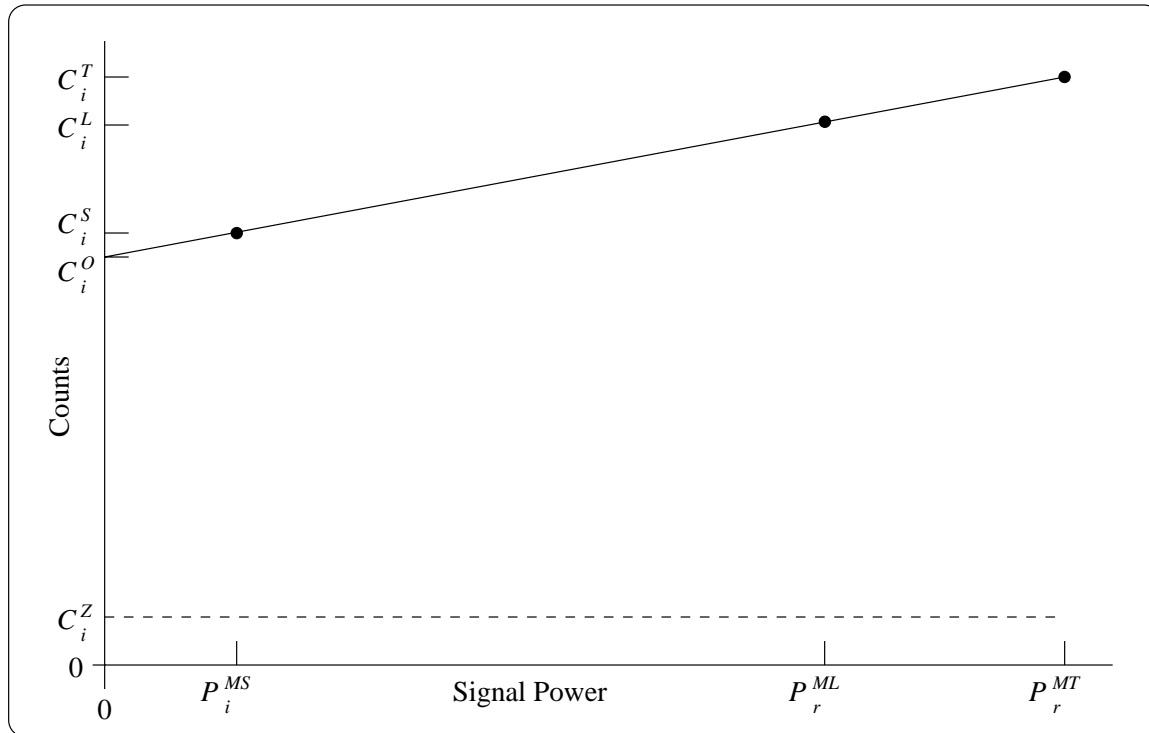


Figure 4.3: Plot illustrating the linear relationship between filter bank channel digitized output (counts) and power collected by the switching mirror. S , A and T correspond to the space, limb (antenna) and target views of the switching mirror respectively. C_i^O is the offset generated by instrument noise, and C_i^Z is the offset built into the digitizing system.

some form of $\frac{1}{f}$ dependence, usually over the timescales of radiometric calibrations for this instrument. Another important fact to note is that C_i^O dominates the digitized signal because system temperatures in all radiometers exceed the temperature of any radiometric source. One of the most crucial aspects of the radiometric calibration data processing is thus the interpolation of reference counts (C_i^S and C_i^T) necessary for estimating channel reference counts and gains at the times of the limb views, since this must be done in a manner which reduces the variance (noise) on the interpolated quantities without the introduction of bias. Of particular concern are the low frequency variations in signal path gain which tend to have a $\frac{1}{f}$ characteristic. These issues are discussed later.

4.2 Radiance calibration algorithms

Radiance calibration of each filter channel is a multi-step process: channel gain is first estimated at the time of each limb radiance measurement; this gain is then applied to the difference between observed limb and estimated reference signal (interpolated to the time of each limb measurement). The random component of uncertainty is then computed, and after a complete MAF of data has been processed as just described, an estimate of the spectrally flat component for each radiometer is generated. The above steps are now discussed further, followed by a description of the radiance calibration algorithms used with the DACS data.

UARS experience indicates that channel gain varies slowly and smoothly between calibration measurement cycles, and EOS MLS is designed similarly to retain this important characteristic. This allows us to determine the gain of an individual channel at the times of ambient calibration target views, and then interpolate this sequence of gain measurements onto the times of limb measurements.

Channel gain at the time of a target view, $\hat{g}_i(T)$, is estimated by

$$\hat{g}_i(T) = \frac{C_i^T - \hat{C}_i^S(T)}{\left(\eta_r^{MT} \epsilon_r \dot{P}_r^T - \eta_r^{MS} \dot{P}_r^S - (1 - \eta_r^{MT}) \dot{P}_r^{BT} + (1 - \eta_r^{MS}) \dot{P}_r^{BS}\right)} \quad (4.17)$$

where the ‘‘radiation offset’’ terms in the denominator involving \dot{P}_r^{BT} and \dot{P}_r^{BS} are provided from FOV calibrations. Temperature sensors on the ambient calibration targets, switching mirror baffle structures, and antenna reflectors allow all quantities \dot{P} in the above expression to be determined from instrument engineering data. The estimated space reference counts at the time of the target view, $\hat{C}_i^S(T)$, are provided by interpolation using a weighted quadratic fit over a ~ 5 MAF window of measurements (with the MAF at the center of the sequence containing the limb data being processed). The estimate of gain at the time of each limb view, $\hat{g}_i(L)$, is then obtained by a similar interpolation (i.e., using the same weighting function, window length and quadratic fit). Solving the first and third of equations (4.16) for \dot{P}_i^L , and using $\hat{g}_i(L)$ for g_i , gives

$$\dot{P}_i^L = \frac{1}{\eta_r^{ML}} \left(\frac{C_i^L - \hat{C}_i^S(L)}{\hat{g}_i(L)} + \eta_r^{MS} \dot{P}_r^S - (1 - \eta_r^{ML}) \dot{P}_r^{BL} + (1 - \eta_r^{MS}) \dot{P}_r^{BS} \right) \quad (4.18)$$

where $\hat{C}_i^S(L)$ is the interpolated value of the space counts at the time of the limb view, obtained using a ~ 5 MAF window and quadratic fit as for the gain interpolation.

4.3 Interpolation of Space references

The performance of the radiance calibration algorithms described earlier is highly dependent upon the manner in which the interpolated space (reference) counts, $\hat{C}_i^S(L)$, are determined. Further discussion of this topic is given in Appendix E. The chosen method is to perform a quadratic least squares fit (as a function of MIF number) to a sequence of reference measurement groups of duration ~ 5 MAFs disposed as symmetrically as possible about the MIF for which the interpolate is being calculated. The interpolations are performed on raw data counts from the filter spectrometer channels. The noise, ΔC_i , on each individual measurement is given by the radiometer equation:

$$\Delta C_i = \frac{(C_i^S - C_i^Z)}{\sqrt{B_i \tau}} \quad (4.19)$$

where:

i is the channel index,

C_i^S is the value, in counts, of the raw reference measurement,

C_i^Z is the value, in counts, of the offset built into the digitizer system (see Section 4.1.4),

B_i is the predetection noise bandwidth in Hz, and

τ is the postdetection integration time in seconds.

Values of B_i are provided for each channel from prelaunch calibrations, and τ is determined directly from telemetry. The digitizer offsets, C_i^Z , are known from prior (UARS) experience to be a stable quantity, and the effects of small errors in their values have negligible impact upon the calibrated limb radiances and associated uncertainty estimates. Since they do not provide essential science data, their values will only be measured periodically for diagnostic purposes (once per orbit), and the last measured valid values of these quantities are adequate for Level 1 computational purposes.

Equation 4.19 provides a reliable estimate of the rms noise on an individual reference measurement. In practice there will be a small additional component of noise arising from gain variations common to all channels of a given radiometer. The most important manifestation of this additional source of noise arises when computing the rms of a time series of data, where it is seen that the ratio of measured to expected rms rises above unity as the duration of the time series extends. It is also seen when fitting a quadratic to a time series of measurement as described earlier for space reference interpolation, where the χ^2 for the fit increases from unity as the calibration window duration is increased. The effects of this noise are diminished by providing less weight to reference measurements more distant from the interpolation point. These topics are discussed in more detail in Appendix C and D where it is pointed out:

- (1) that a common weighting (apodizing) function must be used for all channels whose calibrated radiance data is to be used together in higher level analyses, and
- (2) for the range of planned EOS MLS in-orbit operating modes there is only a very weak dependence on the actual form of the weighting function for the calculated radiances and uncertainties.

4.3.1 Space reference interpolation details

To the space reference count sequence from each channel we fit the function

$$\hat{C}(j) = a + b \times C^S(j) + c \times (C^S(j))^2 \quad (4.20)$$

where the channel index i has been dropped for clarity, and j is an abbreviation for $t(j)$, the time corresponding to measurement j . In practice it is sufficient to replace $t(j)$ with a discrete time index (MIF number) corresponding to each reference measurement. The best-fit coefficients are given by the standard relationships:

$$a = \frac{1}{\Delta} \begin{vmatrix} \Sigma \frac{C(j)}{\sigma^2(j)} & \Sigma \frac{j}{\sigma^2(j)} & \Sigma \frac{j^2}{\sigma^2(j)} \\ \Sigma \frac{jC(j)}{\sigma^2(j)} & \Sigma \frac{j^2}{\sigma^2(j)} & \Sigma \frac{j^3}{\sigma^2(j)} \\ \Sigma \frac{j^2C(j)}{\sigma^2(j)} & \Sigma \frac{j^3}{\sigma^2(j)} & \Sigma \frac{j^4}{\sigma^2(j)} \end{vmatrix} \quad (4.21)$$

$$b = \frac{1}{\Delta} \begin{vmatrix} \Sigma \frac{1}{\sigma^2(j)} & \Sigma \frac{C(j)}{\sigma^2(j)} & \Sigma \frac{j^2}{\sigma^2(j)} \\ \Sigma \frac{j}{\sigma^2(j)} & \Sigma \frac{jC(j)}{\sigma^2(j)} & \Sigma \frac{j^3}{\sigma^2(j)} \\ \Sigma \frac{j^2}{\sigma^2(j)} & \Sigma \frac{j^2C(j)}{\sigma^2(j)} & \Sigma \frac{j^4}{\sigma^2(j)} \end{vmatrix} \quad (4.22)$$

$$c = \frac{1}{\Delta} \begin{vmatrix} \Sigma \frac{1}{\sigma^2(j)} & \Sigma \frac{j}{\sigma^2(j)} & \Sigma \frac{C(j)}{\sigma^2(j)} \\ \Sigma \frac{j}{\sigma^2(j)} & \Sigma \frac{j^2}{\sigma^2(j)} & \Sigma \frac{jC(j)}{\sigma^2(j)} \\ \Sigma \frac{j^2}{\sigma^2(j)} & \Sigma \frac{j^3}{\sigma^2(j)} & \Sigma \frac{j^2 C(j)}{\sigma^2(j)} \end{vmatrix} \quad (4.23)$$

and

$$\Delta = \begin{vmatrix} \Sigma \frac{1}{\sigma^2(j)} & \Sigma \frac{j}{\sigma^2(j)} & \Sigma \frac{j^2}{\sigma^2(j)} \\ \Sigma \frac{j}{\sigma^2(j)} & \Sigma \frac{j^2}{\sigma^2(j)} & \Sigma \frac{j^3}{\sigma^2(j)} \\ \Sigma \frac{j^2}{\sigma^2(j)} & \Sigma \frac{j^3}{\sigma^2(j)} & \Sigma \frac{j^4}{\sigma^2(j)} \end{vmatrix} \quad (4.24)$$

where the summations are over j , the time index of the measurements. In these relationships the quantity $\sigma(j)$ is the rms noise on the measured reference value $C(j)$ at time j . In the absence of any apodization or weighting σ is the same as the expected radiometer noise on individual reference integrations, given by Equation 4.19. To attenuate the effects of signal drifts due to temperature and gain variations we choose to weight each measurement with an exponential roll-off $w_r(j)$ given by

$$w_r(j) = e^{-\left(\frac{|\delta j|}{\lambda_r}\right)} \quad (4.25)$$

where δj is the time difference between interpolation point and reference measurement j .

The parameter λ_r sets the rate at which the relative weighting given to a particular measurement drops off with its distance (time) from the interpolation point. Optimal values of λ_r will be determined from ground tests and in-flight data, and read by Level 1 software as user inputs. A nominal value of 150 MIFs has been selected initially which provides an order of magnitude reduction in weighting for the most distant reference measurements compared to the closest ones for the nominal scan/switching profile. We combine these weightings with the noise levels given by Equation 4.19 to obtain a modified estimate of the noise contribution from each measurement given by σ' where:

$$\sigma'(j) = w_r(j)\sigma(j) = w_r(j) \frac{(C_i^S - C_i^Z)}{\sqrt{B_i\tau}} \quad (4.26)$$

This modified value of $\sigma'(j)$ is used in place of $\sigma(j)$ when performed the quadratic fit via Equations 4.21 through 4.24

4.3.2 Gain interpolation details

Filter channel gains are evaluated at the times of the ambient calibration target views, with the space reference counts quadratically interpolated onto the times of the target measurements as described above. All of the evaluated gains within the calibration window for a given channel are then interpolated onto the times of the limb measurements using the quadratic interpolation scheme described in Section 4.3.1, except that for gain interpolation we do not apply any apodization (weightings) to the gain estimates before the fit is performed. This is because in the estimation of gain given by Equation 4.17 the common impact of gain variations on target and reference counts is largely removed by the differencing of these two quantities in the numerator.

4.3.3 Limitations of the quadratic interpolator

The quadratic fit used for reference and gain interpolation functions extremely well under nominal operating conditions, as demonstrated on UARS MLS, but can exhibit well-known minor shortcomings under certain conditions:

1. The quality of the fit is strongly influenced by bad data.
2. When used as an extrapolator, the quadratic fit exhibits large uncertainties.

The first of these deficiencies is handled by trapping bad data points and excluding them from the fit. This is done by initially excluding data which are outside prescribed limits (set by user inputs), and then rejecting data points which lie more than 6σ from a preliminary quadratic fit on the remaining data. This technique was used successfully on UARS MLS. The second issue is not actively addressed in this software for the following reason: the only condition under which the quadratic fit is used to generate an interpolate at or near one extreme of its temporal extent is when the edge of a calibration window approaches the MAF currently being processed. This will occur typically only when a gain change is commanded or a radiometer is turned on or off. Commanded gain changes have only occurred twice in the 7 years of UARS MLS operation to date, and are not expected more frequently on EOS MLS. Mode changes which turn radiometers on or off are planned to take place on time intervals of at least several weeks, and will be followed by a warm up period lasting several hours before science data is stable enough to be of value to Level 2, which makes the brief increase in uncertainty irrelevant. We thus anticipate operating the quadratic interpolator in a mode in which it exhibits relatively large errors so rarely, and for such brief intervals, that the loss in data quality will be negligible. It should be noted that when the quadratic interpolator is being operated in a mode that exhibits larger than typical uncertainties, these will be reported correctly to Level 2.

4.4 Radiance uncertainties

Appendix C provides a detailed description of noise and the calibration process as it pertains to EOS MLS, and Appendix E provides expressions for measurement covariances when the noise power spectrum contains both white and $\frac{1}{f}$ components. Radiance information in the calibrated bands output by Level 1 software is used in two distinct ways during Level 2 processing, and it is necessary to determine and catalog radiance uncertainties accordingly.

Most retrievals at Level 2 depend mainly upon the spectral contrast in one or more bands, and these results are only weakly affected by errors or uncertainties in the average values (offsets) of these radiances. For these retrievals the most important uncertainties are those which introduce channel-to-channel noise. This noise component arises from radiometer noise present on both limb and reference measurements, and the effects of gain variations are minor in comparison. We refer to this class of error as relative radiance uncertainty.

Several important measurements, particularly water vapor in the upper troposphere, are retrieved from broad spectral regions covered by a combination of one or more standard filterbanks and several broad filter channels. There are no strong spectral features in these measurements, and the dominant source of information arises from the absolute radiances in the measured spectrum. In the calculation of the relative radiance uncertainties discussed above we take care not to include the absolute uncertainty component arising from gain

variations, making it necessary to compute them separately. The correlated uncertainties arise from gain variations in the front-end receivers and HEMT IF amplifiers common to all channels of each radiometer.

The methods of calculating relative and absolute radiance uncertainties are given below. Note that these uncertainties reflect only the contributions from noise on the data integrations, not systematic uncertainties due to errors in, for example, baffle transmissions or other optical properties. The algorithms for calculating the relative uncertainties reflect the method used for reference interpolation, and are modified from those used in UARS data processing only to reflect the minor differences between the two instruments (i.e., variable length MAF, different calibration sequence details). The uncertainties generated by these algorithms, as well as the radiances themselves, were validated as part of UARS Level 1 processing validation using the Optimal Calibrator (Appendix E) for comparison. Results from both methods are virtually indistinguishable, the algorithms given below producing a slightly lower estimate of uncertainty than the Optimal Calibrator, as is to be expected. Of more importance are the radiances themselves, where results from binning of zonal mean CIO data have produced undistorted spectra with channel-to-channel radiance differences of ~ 0.001 K, a dynamic range of 1 part in 10^6 . This result strongly validates the use of the quadratic interpolator.

4.4.1 Relative radiance uncertainties

It is shown in Appendix C that the uncertainty in relative radiance, ΔT_{rel} , is given by

$$\Delta T_{rel} = \sqrt{\frac{(T_{sys} + T_{sig})^2}{B\tau} + (\Delta R)^2 + \left(T_{sig} \times \frac{\Delta g}{g}\right)^2} \quad (4.27)$$

which combines in quadrature three separate noise terms, discussed further below. It is important to note that the right hand term inside the square root includes the noise from just the spectral contrast (limb radiance) and gain uncertainty. The uncertainty derived for the total signal (limb plus system) and gain uncertainty is determined separately and book kept under absolute radiance uncertainty, discussed later in this chapter.

Radiometer noise contribution

The first term inside the square root of Equation 4.27 is simply the noise on an individual limb radiance, given by the radiometer equation:

$$\Delta T_{rel} = \frac{(T_{sys} + T_{sig})}{\sqrt{B\tau}} \quad (\text{radiometer noise contribution}) \quad (4.28)$$

T_{sys} is the current estimate of system temperature, determined once per orbit as a performance monitoring diagnostic, and T_{sig} is the calibrated limb radiance. Based on UARS experience and the similarity of the EOS MLS design, T_{sys} is expected to vary by $<0.5\%$ orbitally which introduces negligible error in this component of estimated radiance uncertainty.

Interpolated reference noise contribution

The second term inside the square root of Equation 4.27 represents the noise in the interpolated space reference:

$$\Delta T_{rel} = \Delta R \quad (\text{interpolated reference uncertainty contribution}) \quad (4.29)$$

This uncertainty is inferred from the estimates of uncertainty in the quadratic fit coefficients of Equation 4.20, and is given by the expression:

$$(\Delta R)^2 = \frac{1}{g^2} [1 \quad j \quad j^2] S^{-1} \begin{bmatrix} 1 \\ j \\ j^2 \end{bmatrix} \quad (4.30)$$

where g is the channel gain in units of Counts per Kelvin and the matrix S is given by:

$$S = s^2 \begin{bmatrix} \Sigma \frac{1}{w^2(j)} & \Sigma \frac{j}{w^2(j)} & \Sigma \frac{j^2}{w^2(j)} \\ \Sigma \frac{j}{w^2(j)} & \Sigma \frac{j^2}{w^2(j)} & \Sigma \frac{j^3}{w^2(j)} \\ \Sigma \frac{j^2}{w^2(j)} & \Sigma \frac{j^3}{w^2(j)} & \Sigma \frac{j^4}{w^2(j)} \end{bmatrix} \quad (4.31)$$

s^2 is the sample variance, given by:

$$s^2 = \frac{1}{(N-3)} \Sigma [C - C(j)]^2 \quad (4.32)$$

and N is the number of data points included in the fit. From UARS experience we expect the sequence of space measurements to be well behaved, and replace the determination of the sample variance with a reliable estimate given by the radiometer equation:

$$s^2 = \frac{(C^S - C^Z)^2}{B\tau} \quad (4.33)$$

where $(C^S - C^Z)$ is the count contribution from the reference radiance with the digitizer offset removed.

Interpolated gain noise contribution

The third and final term inside the square root of Equation 4.27 adds the uncertainty due to noise on gain measurements. Channel gains are evaluated at the times of the ambient target views and then interpolated onto the times of the limb measurements using the same quadratic fit and weightings discussed earlier for interpolation of the space reference counts. The expression used to determine channel gain at the times of the ambient target views, Equation 4.17, differences the actual ambient target counts with interpolated space reference counts. This differencing serves as a high pass filter which attenuates the effects of common drifts in space and ambient target measurements, and results in the radiometer noise on the ambient target measurements being the dominant source of uncertainty in interpolated gain (UARS MLS experience).

The dominant contributor to random uncertainty in gain determination thus arises from noise on the calibration target measurement, C^T , allowing us to write:

$$\left(\frac{\Delta g}{g}\right)^2 = \left(\frac{\Delta C^T}{C^T}\right)^2 = \frac{1}{B\tau} \quad (4.34)$$

directly from Equation 4.17 and the radiometer equation.

This uncertainty applies to a single estimate of gain, but several estimates are combined by the quadratic fit used to interpolate these estimates onto the times of the limb measurements.

The weightings w applied to these estimates as part of the interpolation serve to reduce the noise-induced gain uncertainty to:

$$\left(\frac{\Delta g}{g}\right)^2 = \frac{1}{B\tau} \times \frac{1}{(\sum w(j))^2} \sum w^2(j) \quad (4.35)$$

and giving for this component of relative radiance uncertainty:

$$\Delta T_{rel} = T_{sig} \sqrt{\frac{1}{B\tau} \times \frac{1}{(\sum w(j))^2} \sum w^2(j)} \quad (\text{gain noise uncertainty contribution}) \quad (4.36)$$

4.4.2 Absolute radiance uncertainties

The atmospheric radiance signals of interest appear as noise power in the radiometer IF outputs, typically much smaller than the noise contribution from the radiometers themselves. Gain variations modulate the total IF noise power, causing the overall signal counts from each filter channel to exhibit low frequency noise characteristics even when viewing a scene of fixed radiance. The typical nature of these variations is that they have a $\frac{1}{f}$, or similar, dependence so that their magnitude increases with increasing time scale. Care is being taken with subsystem design and choice of components (HEMT IF amplifiers in particular) to ensure that the breakpoint frequencies of such gain variations are no higher than necessary.

The largest effect of these gain variations is on limb radiances measured most distant in time from the encompassing space reference views. For the absolute radiance uncertainty determination we require knowledge of the shape and magnitude of the post detector noise power spectrum in the instrument's orbital operating environment, which will be obtained by Fourier transforming sufficiently long (a small even integer number of orbits) time series of ambient target data (see [11] for details). These data will be acquired on occasional days (estimated at $\sim 2-3$ per year during the first year of operation and approximately annually thereafter) devoted to in-orbit instrument characterization, since the data analyses require a sequence of target view data equally separated in time, a situation not achieved during nominal instrument operation. The non-spectrally flat component in these power spectra, $S_g(f)$, will be determined during off-line processing of these special data sets, not during routine Level 1 processing.

The term of interest here in the expression for limb port radiance (Equation 4.18) is the one which differences actual limb counts with interpolated space reference counts. The quadratic interpolation used to estimate the reference signal at the time of the limb measurement can be formulated as a sequence of weights applied to each reference measurement included in the fit. The (signal - reference) estimate may then be formulated as a linear combination of weighted signal and reference measurements. This sequence in the time domain is then Fourier transformed to provide the equivalent response of the interpolation and differencing process in the post-detector frequency domain, $H(f)$. This process is described in Appendix D. Using these values for S and H we can then evaluate the absolute radiance uncertainty, ΔT_{abs} , from:

$$\Delta T_{abs} = (T_{sys} + T_{sig}) \sqrt{\int_0^\infty S_g(f) H(f) df} \quad (4.37)$$

This differs from the expressions in Appendix C only in normalization.

The integral inside the square root of Equation 4.37 will be evaluated for each radiometer for the nominal calibration window, measurement sequence and timings, for each MIF within the central MAF. This will be done numerically during off-line processing, resulting in a vector of multipliers for each radiometer, one for each MIF in the central MAF. These vectors will be provided to Level 1 processing as user inputs, and absolute radiance uncertainties will be determined during Level 1 processing by multiplying the sum of system and signal radiances by the component of the vector corresponding to the radiometer and MIF of the measurement.

An approximation being made in evaluating the vector of uncertainty multipliers is that a constant length MAF and fixed observing sequence is being assumed. In practice the length of the MAF will vary by $\pm\frac{1}{6}$ s to maintain synchronization between the scanning pattern and the orbit. The additional error introduced by ignoring this minor variation in timing is negligible compared to the errors themselves. It should also be noted that the error multiplier vector will be symmetric about the center of the limb observing portion of the MIF, and so only the first half of the vector will be supplied to Level 1, the last half being just a mirror image of the first half, as shown in Figure C.4. The vector supplied to Level 1 will accommodate the longest MAF sequence expected during nominal instrument operation, and for shorter MAFs the vector will be truncated to fit the actual length of the atmospheric measurement sequence.

4.5 Additional algorithms for DACS

The DACS, described in detail in Appendix G, process signals in an entirely different manner from the filter spectrometers, resulting in quite different processing algorithms. Some details of the DACS data processing depend upon the mode chosen for their operation (e.g., constant input signal power for all measurements or constant signal chain gain). Below we describe the mode in which these spectrometers will be operated, followed by the resulting data processing algorithms.

4.5.1 Operating mode

The mode and environment in which the DACS will be operated on EOS MLS is:

1. The gains of the IFs feeding the DACS will be kept constant during limb scans – i.e., no “automatic gain control” will be implemented to keep the input RF power to the DACS constant, a fairly common practice with DACS used to analyze radio telescope data.
2. Digitizer thresholds will be changed (if necessary) while the radiometers are viewing the cold Space reference, and during no other part of the viewing/measurement cycle. These threshold levels will be included in the telemetry generated by the instrument.
3. Signal chain stability will be sufficient that any changes to digitizer thresholds during space reference measurements will be considerably smaller than the rms levels of the space reference input signals.

The digitizer thresholds will be set to ± 0.9 of the rms of the input signal level when viewing the Space reference, which when combined with the simplified multiplier arrangement of this implementation results in a S/N ratio 0.87 times that of a continuous correlator.

4.5.2 DACS data processing

The DACS data are processed via the following steps:

1. uncompress data,
2. normalize,
3. convert measured 2 bit autocorrelation to best estimate of the continuous autocorrelation function,
4. Fourier transform into the frequency domain, and
5. convert from relative to absolute signal intensity (i.e., perform radiometric gain calibration).

Uncompressing of the DACS data will depend upon the final implementation of the lossless compression scheme implemented in the C&DH. The need for compression of these data arises because the DACS generate approximately 50% of the internal instrument data traffic, and take the instrument data rate $\sim 30\%$ above the 100 kbps data rate allocation to the spacecraft if no compression is invoked. Fortunately the majority of the DACS channels exhibit extremely low correlations allowing simple lossless truncation by simply reducing the word size of these channels by truncation of several msbs (which will be all zeroes for these channels). Precise details of the compression/decompression will be included in a future version of this document when the compression algorithm details are finalized.

The second step in the processing of these data is normalization, which consists of dividing the counts in each channel by the counts from the zero delay channel:

$$R(j) = R'(j)/R(0) \quad (4.38)$$

This equation and those below are taken from Appendix G which should be consulted for further details. Next we convert the measured 2 bit correlations in each channel to estimates of the continuous (multi-bit) correlation coefficients via Equations G.13, repeated here for convenience¹:

$$\begin{aligned} R_{cont} &= 1.146R_2 - 0.049R_2^2 \quad \text{for } 0.0 \leq R_2 < 0.9 \\ &= 1.340R_2 - 0.340R_2^2 \quad \text{for } 0.9 \leq R_2 \leq 1.0 \end{aligned} \quad (4.39)$$

and transformation into the frequency domain via a real Fourier transform:

$$P\left(\frac{j}{2M\Delta t}\right) = \frac{1}{M} \left[R(0) + 2 \sum_{m=0}^M R(m\Delta t) \times \cos(\pi m j / M) \right] \quad (4.40)$$

In practice a real Fast Fourier Transform (FFT), rather than the Discrete Fourier Transform (DFT) shown above, will be implemented to minimize computational overhead. This is aided by the implementation of 129, rather than 128, lags in the DACS. This allows the creation of a 256 (2^8) element real data vector by appending to the 129 element measurement vector the 127 element vector consisting of the original 129 element vector, reversed in time order, with

¹These equations are appropriate for signal powers small compared to T_{sys} , the case for EOS MLS. They will likely be modified in the future to correct for the effects of finite signal power.

the first and last elements deleted. The data vector being transformed thus has a number of elements which is a power of 2, an optimal number of elements for the FFT.

The final step is to scale the measured power spectrum using the radiometric calibration measurement and power measurement channel data to obtain calibrated radiance spectra in the same units (Kelvin) as the other spectrometer data. It is shown in Appendix H how the counters implemented in the DACS to monitor the digitizer thresholds will be used to measure the power in the input signal during each data integration, resulting in:

$$\sum \dot{P}_i^L = (T_{sys} + \dot{P}^S) \times \frac{301.55 - 760.07 \times f + 448.23 \times f^2}{100.0} \quad (4.41)$$

where:

- $\sum \dot{P}_i^L$ is the sum of the Limb radiances in all DACS channels,
- T_{sys} is the mean system temperature across the DACS IF band,
- \dot{P}^S is the mean Space view radiance for the measurement band, and
- f is the fraction of digitizer states during the data integration that occurred within ± 0.9 of the Space view rms signal amplitude.

4.6 Spectral baseline

The finite apertures of the MLS optical systems introduce small truncations at the edges of the beams viewed by the radiometers. These truncations create diffraction lobes which vary with signal frequency (i.e., channel) resulting in small channel-dependent antenna patterns and losses through the optical paths from antenna to receivers. Although these effects are small, less than ~ 0.1 K across the UARS MLS bands, they result in a channel-dependent signature in the radiances measured through the GHz antenna which can be measured. This signature, called spectral baseline, is observed when differencing signals measured through the antenna (when observing the atmosphere with a sufficiently large tangent height that atmospheric contributions to the spectrally varying component of observed radiance are negligible) from those measured through the space port. The THz optical design uses a common scanning mirror for observing both signal and reference, and is thus less susceptible to the generation of baseline artifacts. Spectral baseline will be determined by Level 1 processing for both GHz and THz radiometers.

UARS MLS experience indicates that these small baseline signatures can be latitude dependent, and can be different for ascending and descending portions of the orbit. For EOS MLS Level 1 software we choose latitude bins similar to those used for UARS, $\sim 22.5^\circ$, resulting in a total of 16 bins (4 north of the equator for ascending and descending portions of the orbit, and the same number south of the equator). Limb observations with tangent point altitudes above selected thresholds will be used to update the appropriate bin (selected by tangent point latitude and ascending/descending flags) by adding the limb/space port difference (in calibrated radiance units) to the data in that bin and incrementing counters which indicate the number of radiances accumulated in each bin. At the end of a processing run these information are converted into average radiances by simply dividing the accumulated radiance by the total number of accumulated radiances in that bin. This results in a spectrum for each band and bin which will be written to the radiance file for use by Level 2 processing.

The tangent point altitudes above which each channel will be included in the baseline computation will be indicated by user input, and will be initially defined by those responsible for the Forward Model and retrieval algorithms. These altitudes will likely be modified after launch based on analyses of in-flight data, as may be the selection of latitude bins. Spectral bands with strong narrow atmospheric signals at the highest altitudes scanned need only baseline data from the wing channels. The maximum nominal tangent height of the GHz radiometer observations is ~ 60 km, and occasional periodic scans to higher altitudes may be included to improve baseline data quality. This decision will also depend of analysis of in-flight data.

4.7 Diagnostics

Several diagnostic quantities will be routinely computed during Level 1 processing and written to the Engineering data file. These quantities are intended to be plotted and examined daily, to provide a clear picture of instrument performance, and hence must be limited in quantity and simple to interpret. To provide end-to-end visibility into system performance these quantities are all derived from spectrometer outputs, and are:

1. System temperatures,
2. Reference χ^2 ,
3. Interpolated channel gains, and
4. Channel zeros.

This selection is based on UARS MLS experience, and may change as further insight is gained into the nuances of EOS MLS behavior with analyses of EM and FM data.

4.7.1 System temperature

System temperature measurements provide an instant view of radiometer performance at a high level. This diagnostic quantity will be measured relatively infrequently (once per orbit initially) since it will remain stable in normal operation, and is expected (UARS experience) to vary by less than 1% orbitally. The measurement is required at Level 1 for the computation of radiometric uncertainty, and is obtained by reducing the IF gains of all bands to their minima so that the filter spectrometer outputs are just the offsets of the digitizers in each channel. System temperature for each filter channel i is computed from the expression:

$$T_{sys}(i) = \frac{(\hat{C}_i^S(Z) - C_i^Z)}{\hat{g}_i(Z)} - \dot{P}_i(2.7) \quad (4.42)$$

where $\hat{C}_i^S(Z)$ are the interpolated space view counts at the time of the zero measurement, C_i^Z are the zero counts, and $\hat{g}_i(Z)$ is the interpolated gain at the time of the zero measurement for channel i . $\dot{P}_i(2.7)$ is the radiative “temperature” of cold space, which, as per Equation 4.1, ranges from ~ 0 to ~ 0.8 K for the full complement of EOS MLS radiometers. The equations given earlier in this chapter for these quantities are used in generating this diagnostic. Interpolated quantities are determined as described earlier in this chapter, and the temperature of cold space is assumed to be 2.7 K.

If multiple consecutive zero measurements are performed, the expression above will still be used to compute a single value of system temperature for each channel by substituting the mean value of the sequence of zero measurements, and interpolating onto the time of the center of the measurement sequence.

4.7.2 Reference χ^2

System temperature measurements provide a valuable, but incomplete, indication of instrument performance in regards to sensitivity. This is because the noise level on a channel is extremely sensitive to signal chain stability over timescales from a single data integration up to a calibration window (~ 3 minutes). A broad standard spectrometer channel (96 MHz) is subject to radiometer noise which has an rms of only 1 part in $\sim 4,000$ of the digitized channel output in a single $\frac{1}{6}$ s data integration. This means that variations in channel gain of less than 0.0004 dB impart significant additional uncertainty into calibrated limb radiances.

The effects of smooth gain variations, such as those caused by orbital temperature variations, are largely removed by the Level 1 radiance calibration algorithms, but variations due to interference, gain variations over the timescales of a MAF, or degradations in signal chain electronics, are harmful. A suitable diagnostic for revealing degradations of this nature is a plot of reference χ^2 for each filter channel. The space reference is observed for more MIFs than the ambient target, so it is chosen for this diagnostic, and determined by interpolating the reference counts onto the times of each reference measurement and computing the mean square differences between measurements and interpolates for each channel. The quantity written to the engineering file is the ratio of the observed mean square deviation to that predicted simply by the radiometer equation. This diagnostic has proven invaluable on UARS MLS during in-flight operation where it has provided immediate indication of such problems as vibrationally induced noise from the switching mirror, allowing rapid correction of the problem by minor changes to operating modes. This diagnostic is also currently in use for analyzing the performance of some EOS MLS breadboard radiometers where the variation of χ^2 with channel bandwidth, and its behavior as the timing of the reference measurements is changed, has led to the rapid discovery of signal instabilities due to ground loops and other test setup issues, greatly facilitating the correction of such problems.

4.7.3 Interpolated gains

The Level 1 radiometric calibration process effectively conceals potential performance changes such as substantial drifts in channels gains which are indicative of failures or degradations in electronic subsystems. We thus write the interpolated channel gains (Equation 4.17) at the times of the ambient target views to the engineering data file for daily review.

4.7.4 Channel zeros

Performance of the filter channel post-detector electronics is monitored by reviewing the offsets of the channel digitizers, measured in orbit whenever IF gains are reduced to their minima to monitor T_{sys} . This parameter reveals any problems due to DC amplifier offset drifts, and on UARS MLS was used to detect a small number (~ 3) of filter channel post detector amplifiers which degraded during ground testing, and were replaced prior to launch (no further channels have degraded in this manner after more than 7 years in orbit). No

algorithms are necessary to determine this quantity – all channel ‘zero’ data are simply written to the Level 1 Engineering file.

4.8 Calibration and ancillary inputs to Level 1 processing

Table 4.2 lists parameters needed by Level 1 software to implement the algorithms described in this document. The parameters listed below ϵ_r are discussed further in the next chapter.

Table 4.2: Summary of parameters required by Level 1 software. PRD is an abbreviation for Platinum Resistance Device, the type of temperature sensor used to monitor the ambient calibration targets. IRU is the abbreviation for Inertial Reference Unit, the spacecraft attitude determination system.

Symbol	Units	Description	Purpose
B_i	Hz	Noise bandwidth of each filter channel	Radiometric noise determination
τ	s	Integration time	Radiometric noise determination
$S_g(f)$	$\text{W Hz}^{-\frac{1}{2}}$	Post detector noise power spectral density for each radiometer	Absolute radiance uncertainties
r'_l, r'_u	–	Relative sideband responses from mixer to switching mirror	Radiometric calibration
ρ_r^k	–	Reflectivity of antenna element k	Antenna emission (radiance) determination
$\eta_{i,s}^{AA}$	–	Antenna beam efficiencies	Limb port to limb radiance conversion
$\eta_{i,s}^{MX}$	–	Baffle transmissions	Radiometric calibration
$\eta_{i,s}^k$	–	Optical transmission of antenna reflector k	Radiometric calibration
ϵ_r	–	Calibration target emissivities	Target radiance determination
$d\alpha$	K^{-1}	Pointing thermal coefficients	Absolute pointing. These coefficients are combined with measured MLS structural temperatures to determine thermal distortions
α, ϵ	$^\circ \text{Count}^{-1}$	Encoder coefficients	Conversion of encoder counts to pointing angles
E	–	Rotation matrices from instrument to spacecraft reference cubes	Absolute pointing determination
D	–	Rotation matrix from spacecraft reference cube to IRU reference frame	Absolute pointing determination
E^T, D^T	K^{-1}	Thermal coefficients of E and D	Absolute pointing. These coefficients are combined with spacecraft thermal data to correct for thermal distortions in spacecraft structure between MLS and the IRU.
R_0	Ω	0 C resistances of individual PRD temperature sensors	Determination of cal target temperatures
C_l, C_h	Ω, V	Engineering Data Hybrid internal calibration values	Conversion of engineering data “counts” to engineering units
a, b	–	PRD conversion coefficients	Conversion of PRD resistance into inferred temperature
c, d, e, f	–	Thermistor conversion coefficients	Conversion of thermistor resistance into inferred temperature

Chapter 5

Engineering data

The MLS instrument data system comprises a redundant pair of central Command and Data Handling (C&DH) subsystems linked via serial data and synchronization busses to a network of ~ 30 Remote Interface Units (RIU). Each RIU contains an Engineering Data Hybrid (EDH) which is used to acquire analog engineering data. A partial list of the capabilities provided by an EDH includes:

1. multiplexing (selection) of input signals from one of 16 external or 6 calibration (internal or external) sources,
2. selection of input polarity,
3. selection of signal gain,
4. selection of source excitation current (none, 0.1 mA for thermistors or 1.0 mA for Platinum Resistance Device temperature sensors),
5. selection of signal offset, and
6. digitization via a V/F converter, with the output from the V/F recorded by logic in the RIU.

Not all combinations of signal chain attribute listed above are selectable without restriction, and the EDH is targeted towards the measurements of voltage and resistance (to support both PRDs and thermistors) as shown in Table 5.1. Resolution of the digitized result is limited by the external counter gate time and by the inherent noise of the V/F converters, but is designed to meet or exceed 16 bits for all measurements. The nonlinearity of the signal chain is dominated by the V/F converter, and is less than 0.05% of full scale signal for all measurement types. This level of nonlinearity is such that no corrections are needed in the processing software.

Figure 5.1 shows a block diagram of the EDH. One of 16 input sources is selected via an analog multiplexer which allows any input source to be either a voltage or resistance (for temperature measurements). Resistive sources are excited by a current source of either 0.1 or 1.0 mA, and the sensors may be connected in 2, 3 or 4 wire configuration. In addition, it is possible to reverse the polarity of the excitation current supplied to PRDs in order to repeat the resistance measurement. By averaging the inferred resistances measured using both polarities of excitation current, the effects of thermocouple-induced voltage errors in these

resistance measurements are eliminated. This is necessary to achieve the desired accuracies for PRD measurements of internal calibration target temperature.

Input signal source and polarity, excitation current, and channel gain and offset are all selected simultaneously via an externally provided 20 bit control word which is generated under software control within the host RIU, allowing for considerable flexibility in measurement sequencing. The output from the EDH consists of a (nominally) square wave signal from the V/F converter which is of the same type used in all of the filter spectrometers. This V/F converter is operated over a substantially restricted portion of its potential operating range (~ 40 kHz to ~ 80 kHz) in order to provide improved linearity compared to a system which operates with a low frequency limit much closer to zero; it should be noted that the operating ranges of the V/F converters in the filter spectrometers are restricted inherently by the system temperatures of the radiometers which tend to be large compared to the atmospheric signals.

An input channel is selected towards the beginning of a MIF, and the rest of the MIF is used to allow the analog circuitry to settle. In the case of resistance (temperature) measurements, some of this settling time may be required to charge up any feedthrough filters in the signal path from the EDH to the external sensor. The measurement system is designed to allow complete settling in the ~ 1 MIF allocation. The subsequent MIF is used to measure the average frequency of the V/F converter output.

This timing scheme is illustrated in the upper half of Figure 5.2. Measurement sequences will be repeated starting on MAF boundaries, which implies that all engineering measurements made by any RIU can and will be completed in the duration of a MAF. In order not to “waste” measurement time, some measurements may be selected more than once during a MAF, and Level 1 processing will separately process (and write to the Engineering file) measurements which are repeated during a MAF. The only exception to this is for the calibration targets PRD data; for these data Level 1 processing will additionally combine (by averaging) the 2 measurements taken of each PRD resistance in order to eliminate offsets due to thermocouple junctions in the wiring between EDH and sensor. The temperature deduced from this average will be used in radiometric calibration, but the individual temperatures

Table 5.1: Input signal types and attributes supported by the Engineering Data Hybrid. Measurements are of voltages or resistance. PRD1 and PRD2 refer to Platinum Resistance Device temperature measurements over a relatively narrow temperature range (PRD1) for the internal calibration targets, and a wider range (PRD2) for the sensors monitoring the structure external to the main modules.

Measurement Type	Minimum	Maximum	Units
Voltage	-1	+7	Volts
PRD1	460 (-20)	640 (+70)	Ω (C)
PRD2	300 (-100)	700 (+100)	Ω (C)
Thermistor	5.0 (-80)	0.318 (+100)	k Ω (C)

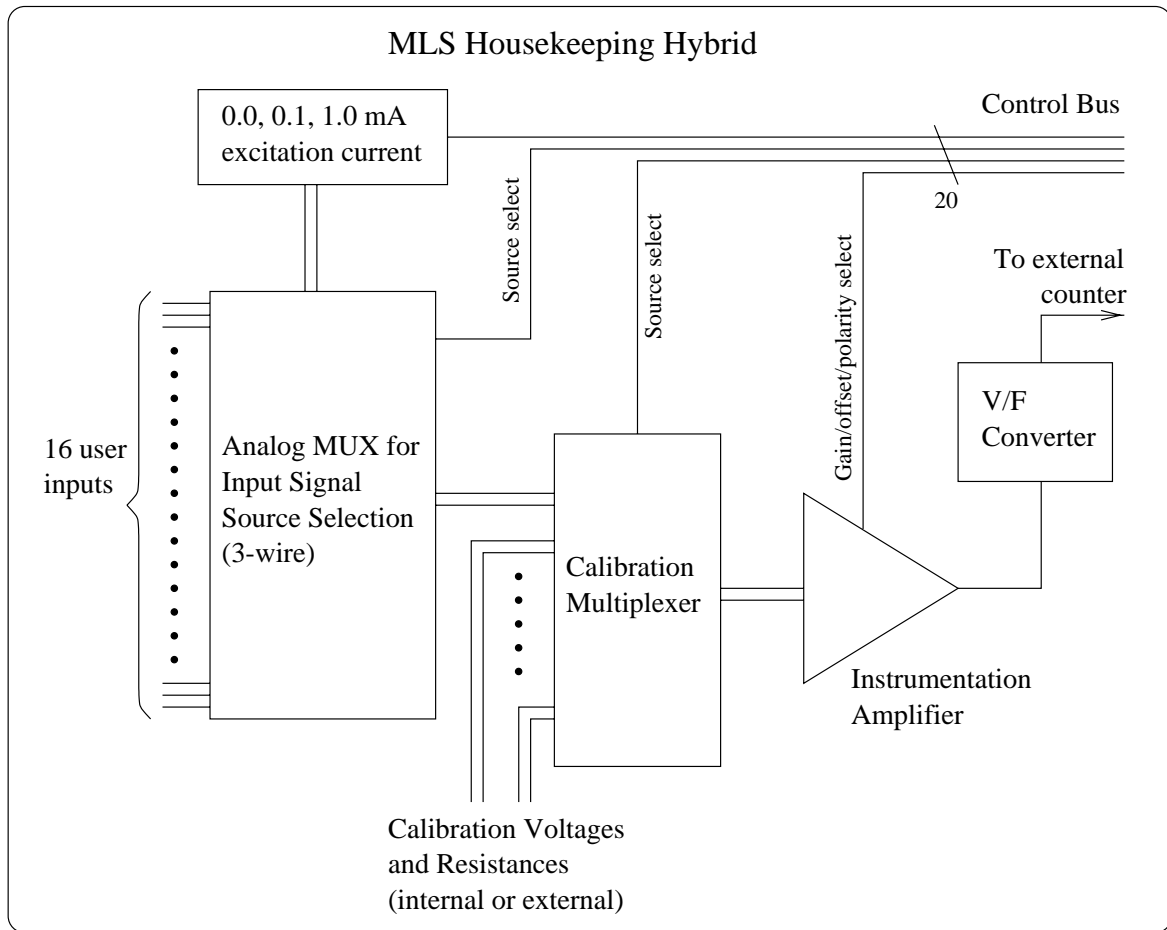


Figure 5.1: Block diagram of the Engineering Data Hybrid. The upper portion of the figure shows timing events at the MIF level; the lower illustrates the operation of the external counters used as the digitizer. See text for further details.

deduced from each excitation current polarity will also be written to the engineering file for diagnostic purposes.

Figure 5.2 shows additional details of the V/F frequency measurement scheme. The RIU generates a Gate signal to indicate when the V/F output is to be recorded. The rising and falling edges of this signal indicate to the control logic that the subsequent rising edges of the V/F output denote the actual period during which the number of cycles of both the V/F output (N_v) and of the higher frequency timebase (N_t) are recorded. This results in a variable measurement interval which is slightly delayed from the commanded one, but since the minimum V/F frequency is ~ 40 kHz, the variation and delays are small and inconsequential. The control logic is designed so that a V/F converter which is stopped or running at a speed so low that the measurement cycle has not terminated in time for the next one to commence (due for example to a large input signal of the opposite polarity to the one assumed for that channel), the measurement cycle is forced to terminate, an error is recorded, and the next measurement cycle is started correctly and on time. The measurements of N_v and N_t are used

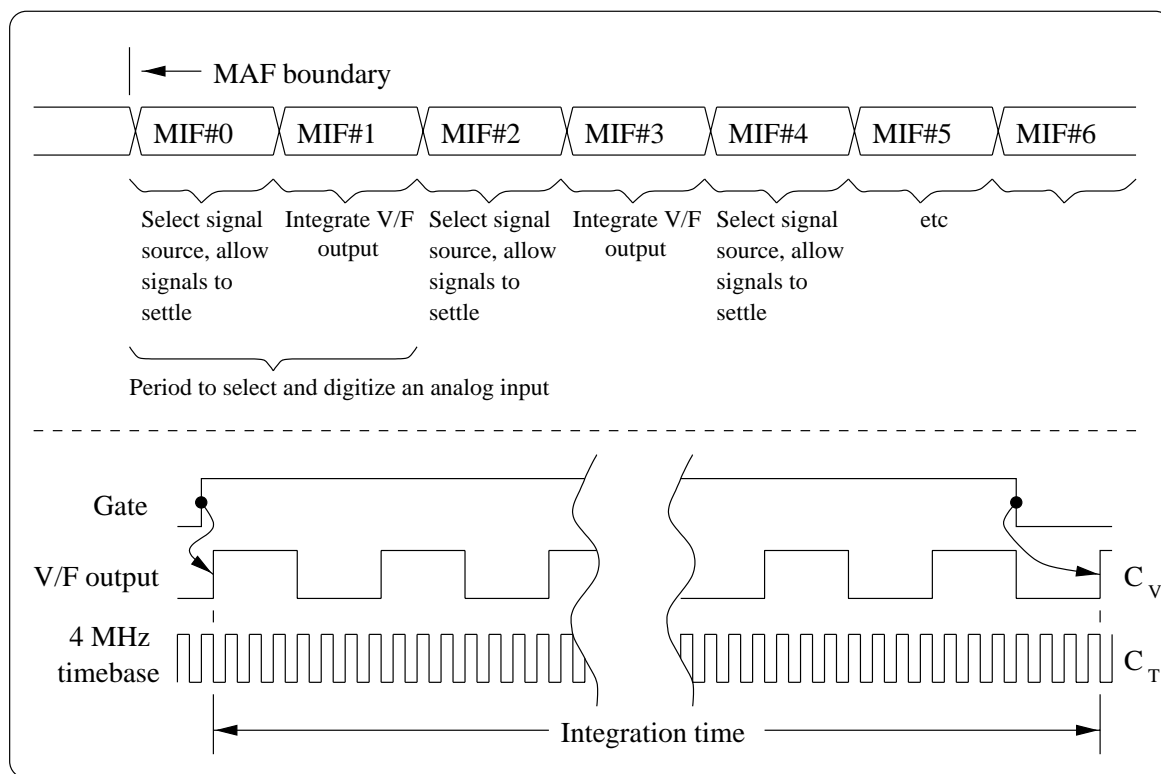


Figure 5.2: High level illustration of EDH timing. The upper portion of the figure shows timing at the MIF level, and the lower portion illustrates the relationship between commanded measurement interval (defined by the Gate signal) and that set by the V/F output. See text for further details.

to deduce the average frequency of the V/F converter output, f_v , from the straightforward relationship:

$$f_v = 4.0 \times 10^6 \times \frac{N_v}{N_t} \quad (5.1)$$

where 4.0×10^6 is the frequency of the timebase in Hz.

Advantages of this technique over the one used on UARS MLS (in which the V/F output is counted without additional information from a second counter monitoring a timebase) include:

1. the resolution of the measurement is limited by the frequency of the timebase, not the frequency of the V/F converter,
2. high resolution may be obtained even with the V/F converter run at a relatively low frequency and over a limited portion of its dynamic range (which improves linearity significantly),
3. by choosing a lower operating frequency range for the V/F converter than would be the case if the dual counter scheme were not employed, overall power consumption can be reduced, and

4. the inferred result (a frequency) is independent of the signal integration time.

Penalties for this measurement approach include:

1. larger counters are needed (40 bits total *vs* 15 for this implementation),
2. the control logic is more complex, and
3. slightly more onboard processing is required, since we wish to telemeter the derived quantity f_v to the ground instead of the directly measured quantity N_v .

The resolution of this measurement system is ultimately limited by phase noise (jitter) in the V/F converter output signal which tends to have a nominally $\frac{1}{f}$ frequency characteristic. This means that for a substantial range of signal integration durations the S/N of the measured frequency remains essentially constant. The EDH signal conditioning electronics is designed to introduce much less noise into the measurement cycle than the V/F jitter, so that the dominant source of noise in a digitized signal measurement should be limited typically by the V/F jitter or by thermal drifts. When recording internal calibration points, the V/F frequency jitter is approximately 1 part in 500,000 peak-to-peak. To allow all MLS engineering data to be included in the 512 bits s^{-1} engineering data stream broadcast by the spacecraft we truncate the telemetry words for f_x to 20 bits in length. So that resolution is not lost unnecessarily, we choose a modified form of Equation 5.1 for evaluating f_x :

$$f_x = F_{\text{norm}} \times \frac{N_v}{N_t} \quad (5.2)$$

where F_{norm} is an EDH-specific normalization factor used within the C&DH to ensure that dynamic range is not wasted. This 20 bit word will be extracted from an IEEE-754 compliant floating point number by the C&DH as a result of evaluating the expression given in Equation 5.2, and will consist of the 20 msbs of the mantissa. Since this floating point format has an implied leading “1,” the Level 1 software will set this bit to obtain a 21 bit integer prior to further processing.

All analog quantities measured by an EDH are converted to engineering units in a similar manner. Measurements are classified according to “type” by the source of the input signal:

1. a voltage (or another signal type, transformed to a voltage for digitization by the EDH),
2. a resistance consisting of a YSI 44906 thermistor in parallel with a precision 4.990 k Ω resistance, or
3. a resistance consisting of a Rosemount 118AKT2F PRD.

The PRD category is further subdivided, since PRD resistance may be read over one of two possible ranges, as shown earlier in Table 5.1. All input signals are converted into frequencies as described earlier. The frequency measurements of all non-calibration inputs are then converted into engineering units via linear interpolation using EDH measurements of calibration references. Each measurement frame (MAF) from an EDH contains at least one set of calibration pairs for each data type.

Analog data inputs are first calibrated using high and low calibration frequencies from the EDHs in each RIU:

$$x = \left(\frac{f_x - f_l}{f_h - f_l} \right) (C_h - C_l) + C_l, \quad (5.3)$$

where:

x is the desired input signal in engineering units, f_x is the frequency measured from the corresponding subchannel,
 f_h is the “high” calibration frequency for that channel type,
 f_l is the corresponding “low” calibration frequency, and
 C_h and C_l are the high and low calibration parameters for this channel, which are measured during test of the EDH.

If more than one measurement of f_h and/or f_l is made during a given MAF, the values of f_h and f_l for that MAF are averaged before being applied in Equation 5.3. Multiple measurements of f_x for a given subchannel are processed independently however, and recorded in the Engineering output file as separately time tagged quantities. If no valid measurements of f_h and/or f_l are present in the data record for a given MAF, the most recent prior values are used. If no prior values exist (e.g., at the start of a daily processing run) then the values from the default file are used.

5.1 Conversion to engineering units

5.1.1 Voltages and currents

In the case of input voltages, Equation 5.3 gives the input signal to the EDH in units of Volts. For scaled voltages, and currents converted into voltages for recording by the EDH, the appropriate additional conversions need to be applied.

5.1.2 Temperature – PRDs

For resistive input sources, Equation 5.3 gives the resistance of the source in Ohms. For PRD sensors the measured resistance, R , is converted to a temperature, T , using the relationship:

$$T = \frac{a \times (R_{tlm} \times 500.0/R_0 - 500.0)}{(1.0 - b \times R_{tlm} \times 500.0/R_0)} \quad (5.4)$$

where:

T is the inferred temperature in Celsius,
 $a = 0.48945548411$,
 $b = 7.20107099888 \times 10^{-5}$, and
 R_0 is the resistance of the sensor at 0 C.

This expression is accurate to ~ 0.15 C over the temperature range -50 C to +150 C. By customizing the coefficients a and b (not proposed here) the error is reduced to ~ 0.08 C. The expression above is the one used in UARS MLS data processing.

Note that the calibration target PRDs are measured with both polarities of excitation current during each MAF. The inferred temperatures will likely be slightly different for each polarity, and the average of these temperatures for each sensor is the quantity to be used for radiometric gain determination and to be written to the Engineering file.

5.1.3 Temperature – thermistors

As for the PRDs, Equation 5.3 is used to determine the resistance presented by the sensor at the input to the EDH. Each thermistor is paralleled with a 4.990 k Ω precision resistor, and Equation 5.5 is used to convert the measured resistance, R_{in} , into the resistance of the thermistor sensor, R_{th} , in Ohms, as follows:

$$R_{th} = \frac{(4990.0 \times R_{in})}{(4990.0 - R_{in})} \quad (5.5)$$

The thermistor resistance is then converted into temperature (Celsius) using the relationship:

$$T = \frac{1.0}{(c + \log(R_{th}) \times (d + \log(R_{th}) \times (e + \log(R_{th}) \times f)))} - 273.16 \quad (5.6)$$

where:

$$\begin{aligned} c &= 1.286212 \times 10^{-3}, \\ d &= 2.355213 \times 10^{-4}, \\ e &= 9.826046 \times 10^{-8}, \text{ and} \\ f &= 8.835732 \times 10^{-8} \end{aligned}$$

This expression was derived for UARS MLS data processing using vendor supplied data.

5.2 Data quality and reasonableness

As for all telemetry processed by Level 1 software, engineering telemetry will be checked for reasonableness using available flags, CRC and checksum information. In addition, measurements of EDH calibration values will be limit checked, and appropriate estimates used in the absence of current data, together with the generation of diagnostic flags. Non-calibration inputs cannot be checked for reasonableness in most cases.

Chapter 6

Ancillary data

There are two primary classes of ancillary data in the context of level 1 processing: those generated by the spacecraft or Flight Dynamics team, some of which are sent directly to MLS, and some generated on the ground by ground processing and made available to Level 1 software together with the instrument Level 0 data; and additional data needed by Level 2 which involves processing at Level 1, such as estimated tangent point location in an Earth referenced coordinate system and local solar time at the tangent point(s) of the observations.

The spacecraft generated/derived ancillary data listed in the Tables below are taken from the Instrument Reference Report [7].

6.1 Spacecraft provided/related ancillary data

The ancillary data in Table 6.1 will be transferred to MLS via the 1553 spacecraft interface in “real time,” while the data set in Table 6.2 will be available to Level 1 during ground processing of data. All of these data will be written to Level 1 output files and hence be available to higher levels of MLS processing. X, Y and Z in the tables are the primary spacecraft axes about which all pointing information is reported. MLS observes in the XZ plane, with X being the nominal direction of flight of the spacecraft, and Z the nominal nadir direction. X, Y and Z form a conventional right-handed coordinate system.

The quaternion used to indicate spacecraft attitude and solar/lunar positions is a four element vector. Three of the elements specify a direction. The fourth element is the rotation angle (in a right handed coordinate sense) about that vector. The first three elements are dimensionless, the angle is reported as the cosine of half the rotation angle.

6.2 Level 1 processed ancillary data

Table 6.3 lists those ancillary data required by Level 2 which is produced by Level 1 processing. These data will be written to an Ancillary Data file.

Each data record of the ancillary file will be tagged with the time of the start of the MIF to which the data applies. In addition, the records will be tagged with MIF, MAF and orbit counters. The MIF counter indicates the MIF within the current MAF. The MAF counter is reset to zero for the MAF in which the tangent points cross the equator heading south to north. The orbit counter is generated by Level 1 software, and marks each orbit sequentially

within a daily file, starting with orbit number zero. The orbit counter is incremented when the MAF counter is reset.

The master horizontal coordinate, ϕ , referenced in Table 6.3 is the angle between the normal to the geoid, normal to the nominal limb ray path (boresight), and the equator illustrated in Figure 3.7 of the Level 2 Theoretical Basis document [13]. This coordinate ranges between 0 and 360° (2π radians) at the start of processing, but in the ancillary data file will not be reset to 0 as it ranges beyond 360°, but allowed to increase throughout the day. Level 2 processing will convert it back to modulo 360 (or 2π) as necessary. This representation is at the request of Level 2 designers.

Table 6.1: Data sent from the spacecraft directly to MLS via the 1553 bus for local processing. These data are sent in two different data packets, each sent at a 1 Hz rate, delineated by the entries labeled Time Stamp. Attitude angles are reported as a 4 element quaternion. ECI denotes Earth Centered Inertial. IRU indicates Inertial Reference Unit, the spacecraft gyro subsystem. SCE indicates Spacecraft Converter Electronics, the MLS interfaces to the spacecraft power supply. See text for definition of the X, Y and Z axes, and of the quaternion. Packet headers have not been shown.

Description	Units
Time stamp	μs
Spacecraft Position – ECI X	m
Spacecraft Position – ECI Y	m
Spacecraft Position – ECI Z	$m s^{-1}$
Spacecraft Velocity – ECI X	$m s^{-1}$
Spacecraft Velocity – ECI Y	$m s^{-1}$
Spacecraft Velocity – ECI Z	$m s^{-1}$
Attitude Angle – Q1	quaternion
Attitude Angle – Q2	quaternion
Attitude Angle – Q3	quaternion
Attitude Angle – Q4	quaternion
Roll	radians
Pitch	radians
Yaw	radians
Roll Rate	$arcsec s^{-1}$
Pitch Rate	$arcsec s^{-1}$
Yaw Rate	$arcsec s^{-1}$
Time stamp	μs
Solar Position – X	quaternion
Solar Position – Y	quaternion
Solar Position – Z	quaternion
Lunar Position – X	quaternion
Lunar Position – Y	quaternion
Lunar Position – Z	quaternion
Descending Node Crossing	ms
Oblateness Offset – X	arcsec
Oblateness Offset – Y	arcsec
Oblateness Offset – Z	arcsec
Oblateness Offset Rate – X	$arcsec s^{-1}$
Oblateness Offset Rate – Y	$arcsec s^{-1}$
Oblateness Offset Rate – Z	$arcsec s^{-1}$
SCE number 2 Voltage	V
SCE number 3 Voltage	V
IRU Fault Status	N/A

Table 6.2: Spacecraft-related data made available to Level 1 software during ground processing. Where these data are the same as those shown in the previous table, the accuracy and resolution of data provided on the ground are much higher.

Spacecraft Parameter	Description	Knowledge; Resolution	Time Tag Accuracy	Temporal Resolution
Position	ECI XYZ	500 m; 1 m	50 ms	1 Hz
Velocity	ECI XYZ rates	0.1 m s^{-1} ; 1 mm s^{-1}	50 ms	1 Hz
Attitude (roll/pitch/yaw)	Euler angle to orbit frame	65 arcsec; <0.2 arcsec	50 ms	8 Hz
Attitude Rate (roll/pitch/yaw)	Euler angle rates	$0.1 \text{ arcsec s}^{-1}$; $0.01 \text{ arcsec s}^{-1}$	50 ms	8 Hz
Sun Vector	Spacecraft to Sun XYZ	0.075° ; 0.01°	50 ms	1 Hz
Moon Vector	Spacecraft to Moon XYZ	0.75° ; 0.01°	50 ms	1 Hz
IRU Status and Gyro Status	IRU Configuration bits, Gyro Status Bits	Bits and gyro ID	50 ms	1 Hz
Altitude	Distance along Geocentric or Geodetic Vertical to Ellipsoid	15 m; <0.1 m	50 ms	1 Hz
IRU Gyro Angle	Accumulated angle in Gyro Coordinates	<1 arcsec; <0.1 arcsec	50 ms	8 Hz
Spacecraft Converter Voltage	SCE Voltage	<1 V; <0.1 V	50 ms	1 Hz

Table 6.3: Spacecraft and pointing-related data to be written to the Level 1 Ancillary Data output file. These data are primarily for use by Level 2 processing, and this table will be updated at the request of Level 2 designers. These data will be tagged with time, MIF, MAF and orbit counters. The definition of “North” is currently TBD.

Item	Description	Rate
Tangent points	Tangent point locations for R1A and R5H, the primary pointing radiometers for the GHz and THz modules. Reported as geocentric altitude, latitude and longitude.	MIF
Solar zenith angle	Solar zenith angle at the tangent points reported above.	MIF
Radiometer offsets	Radiometer to radiometer boresights determined from spacecraft and instrument thermal data.	MAF
Spacecraft location	Spacecraft geocentric altitude, latitude and longitude.	MIF
Elevation rate	Estimated elevation rate of the GHz and THz boresights derived from spacecraft attitude rate and actuator encoder data.	MIF
Solar position	Angle of the sun <i>wrt</i> R1A and R5H boresights.	MIF
Lunar position	Angle of the moon <i>wrt</i> R1A and R5H boresights.	MIF
Bright body positions	Angles of other “bright bodies” (e.g., Venus, Jupiter, etc., which may impart noticeable signals if in FOV) <i>wrt</i> R1A and R5H boresights.	MIF
Boresight direction	Angle between R1A and R5H boresights and “North.”	MIF
ϕ spacecraft	Spacecraft master horizontal coordinate. See text.	MIF
ϕ tangent point	Tangent point master horizontal coordinate. See text.	MIF

Chapter 7

Resource estimates

In this chapter we provide estimates for the key resource requirements of EOS MLS Level 1 processing, including file and daily I/O volumes, main memory requirements, and processing capability. Memory and processing capability are machine dependent, and only approximate estimates can be made at this time. The intent here is to bound the requirements of this software and show that they are reasonable. Since this software inherits much from its UARS MLS predecessor, many of the estimates can be checked for reasonableness by comparison. We also discuss how I/O is minimized in this processing, and how use will be made of parallel processing to reduce wall clock time of daily processing runs. To simplify the discussion we assume an operating mode which maximizes data rate while meeting the current instrument power allocation, realizing that other modes may be invoked which reduce all requirements discussed below.

7.1 Input data volume

The input to Level 1 processing is dominated by the instrument data stream, sent to the spacecraft at an average rate of 10^5 bits s^{-1} , a daily volume of 1 GB (GB \equiv gigabyte, 2^{30} bytes). Ancillary data and overhead for the HDF file format headers and links is unlikely to raise the overall daily input data volume above ~ 1.2 GB.

7.2 Output data volume

The instrument data input to Level 1 processing is marginally compressed ($\sim 30\%$ compression) to meet the 10^5 bits s^{-1} data rate allocation using a simple lossless algorithm for the filterbank and DACS science data. The corresponding output data products are not compressed, primarily because they are in single precision floating point format for both radiances and uncertainties. If data storage ever becomes an issue, these formats can be changed to scaled integers (as for UARS) with corresponding small loss of dynamic range and increase in overhead at Level 2 to convert back to scientific format for processing.

The largest source of output data is the calibrated radiances for the filter and DACS channels, including the relative uncertainty calculated with each filter channel radiance. The daily volumes for these are computed in the following tables.

Number of filter channels: ¹	$19 \times 25 + 5 \times 11 + 8 = 538$
Readout rate:	120 times per MAF
MAFs per orbit:	240
Orbits per day:	14.6
Bytes per datum:	8 (includes radiance and uncertainty estimate)
Daily total:	1.8 GB
Number of DACS channels:	$128 \times 4 = 512$
Readout rate:	120 times per MAF
MAFs per orbit:	240
Orbits per day:	14.6
Bytes per datum:	4 (we assume a common uncertainty for all channels)
Daily total:	0.9 GB
Number of engineering channels:	500 (upper limit for analog monitors)
Readout rate:	Once per MAF
MAFs per orbit:	240
Orbits per day:	14.6
Bytes per datum:	4
Daily total:	0.007 GB

The daily totals are the products of all of the entries in each table, and the analog engineering data is seen to be insignificant in volume compared to the science data. Details such as the time tags for engineering data have been omitted and are likely to approximately double the daily volume for such data, but its volume still remains insignificant. The major missing category of data from the above tabulations is the diagnostic data required to monitor instrument radiometric performance. These data include the reference χ^2 , interpolated channel gains and reference counts, and digital data such as phase lock status, and are likely to amount to about a quarter of the size of the radiance data volume, ~ 0.7 GB (derived from comparison with UARS). The daily total data volume in the above categories is thus estimated to be ~ 4.0 GB.

7.3 I/O minimization

All filter channel data are gathered in a 5 MAF calibration window for radiometric calibration, with smaller windows for DACS and engineering data. Data are logically appended to the end of this window and deleted from its beginning as each MAF is processed. This results in sequential access to the input data files, and the same for the output files, reducing file I/O to the minimum possible.

In the first versions of Level 1 software it is likely that baseline determination will be performed after all of a day's data has been processed, resulting in a second read of the output radiance file. This data can readily be determined as the radiance file is written for

¹In the primary (nominal) MLS operating mode only 8 of the 12 broad filter channels are turned on.

the first time, and this optimization will be incorporated after the algorithms for baseline determination are tested and verified.

The instrument data (Level 0) files input by Level 1 processing retain the packet structure generated by the C&DH, documented in [12]. Output files will be in Hierarchical Data Format (HDF) format, and as data are written, links will be updated and records written. A major advantage of this scheme is that files will only be of the size necessary to store actual output products, and the file records will not contain space for TBDs, spares, or data that might exist at some future time. This is a major benefit of an intelligent file structure, and also serves to minimize files sizes, and hence I/O volumes. This should be particularly evident for the EM which implements only 5 25-channel spectrometers, and none of the 11-channel, broad filter or DACS subsystems. This will result in a reduction in the radiance output file size to just over 10% its size for the full instrument in its primary operating mode.

7.4 Main memory requirements

There are three main components to the memory required by this software – the code itself, static data (such as calibration information), and workspace for the data in the calibration window and the results of computations on these data. UARS experience indicates that code size will be insignificant compared to available memory. The calibration window data structures contain 5 MAFs of instrument data and the derived output products. Instrument data for this window size amounts to 1.5 MB, and the output data (calibrated radiances, uncertainties, etc.) will increase memory requirements to ~ 10 MB. Parallel processing (discussed below) requires duplication of certain working data structures since multiple threads will require some private working variables. Assuming that calibration and related data amount to a similar data volume indicates that the memory footprint of Level 1 software will be well under 100 MB. This is to be contrasted to the 8,192 MB available in the current version of the EOS MLS Science Computing Facility (SCF), indicating that main memory requirements are not an issue

7.5 Processing capability

To obtain an initial estimate of processing capability we have chosen to scale using UARS MLS Level 1 software performance as the benchmark. This is felt to be a reliable starting point because of the similarity of the algorithms for both instruments.

UARS Level 1 data processing is currently performed on a desktop Alphastation 4000 (300 MHz Alpha 21064 processor) and is I/O limited, taking ~ 24 minutes of wall clock time, but only 65 seconds of CPU time, to process an entire days worth of instrument and ancillary data. The current SCF (an SGI Origin-2000 with 16 processors) benchmarks at ~ 40 times the speed of the Alpha (this estimate comes from published SPECfp_rate95 figures for both machines). The I/O is also considerably faster than for the Alpha due to the direct connection to a fast, efficient, local RAID array for file storage. The number of channels to be processed is ~ 10 times larger for EOS, and these channels are output at an ~ 12 times higher rate.

Simply scaling the CPU time up by the increase in data volume and down by the estimated performance increase of the SGI compared to the Alpha indicates that about 3 minutes of CPU time is required to process a day's worth of EOS MLS data. We justify this simple

scaling as follows. For the filter spectrometer channels the nominal UARS and EOS Level 1 calibration windows encompass a similar number (~ 60) of space reference views for quadratic interpolation during radiometric calibration processing. Since all other aspects of processing for each channel are also very similar, scaling is felt to provide a highly reliable metric for estimating these processing requirements. The DACS represent the only completely new processing of significance compared to UARS. For the filter spectrometers each channel is examined individually by the Level 1 software to look for unexpected gain changes or abnormally high reference χ^2 , since each channel has its own failure mechanisms. This is not the case for the DACS because of their digital implementation, and eliminates much of the processing that takes place for the filter channels before radiometric calibration is performed. Gain determination for each DACS is from a single filter channel which monitors its full RF input power. The unique processing steps performed on the DACs data consist of converting each measured autocorrelation into an estimate of the true (multi-bit) autocorrelation, and the subsequent Fourier transform into the frequency domain. The conversion is relatively straightforward, and care has been taken in the design of the DACS to ensure that the transform is efficient. This has been accomplished by providing 129 lags in each DACS which results in a 256 element vector² being transformed, very efficient using an FFT. We estimate that the combined overhead of the various DACS processing steps is less than that required for processing an equivalent number of analog filter channels, but for this work assume that the cost of processing a single filter and DACS channel is the same. This will be verified as soon as DACS data are available, either from future EOS MLS brassboard models or SLS prototypes currently in house.

We estimate I/O time by assuming a sustainable transfer rate to and from the local disk storage of 10 MB s^{-1} (a single 100 Mbit Ethernet channel to a local RAID system, as on the current EM SCF system). If we assume that input data are read once and output files written and read once (the read being to derive diagnostic information such as baseline spectra), then $\sim 10^{10}$ bytes total are transferred to and from the disk storage in the processing of one day's data. At a 10 MB s^{-1} transfer rate this implies ~ 17 s of I/O time. We thus estimate ~ 20 minutes of wall clock time to process a single day of EOS MLS data at Level 1, very similar to the current figure for UARS MLS.

The above arguments do not take into account improvements in processing power in the time between now and production processing, nor the increase in processing speed attainable through the addition of more processors. Similarly, disk I/O speed can, and will, be substantially increased by increasing the number of connections from the RAID array to the processor system from the current single 100 Mbit non-shared Ethernet path to several (up to 4 possible in the current system), or to even faster IEEE-1394 interconnects expected to be readily available in the timescale of the final SCF. We thus conclude that processing time, including waits for external I/O transactions, will not present any issues or problems.

²The 129 element data vector from each DACS is converted into a 256 element vector for Fourier transforming by appending all but the first and last element of the DACS data in reverse order to the original vector. The returned transform consists of just the real part of the returned data, the imaginary part being all-zero.

7.6 Speedup through parallel processing

Many areas exist in this software to achieve speedup through parallelism, but it only makes sense to parallelize the portions of the code which produce the largest return for the least effort, and to do this in a portable manner so that the software is not tied to a single processor vendor and family. All appropriate computers for this processing task in the timescale of EOS are likely to be symmetric multi-processor (SMP) systems with substantial shared memory and a small number (8 to 128) of processors. Such systems are available from SGI, HP, IBM, Sun and DEC (now Compaq), and the following discussion directly relates to the SGI Origin 2000, the version of the SCF used for development of this software. Our research indicates that changing to an alternative machine/vendor in the future would be a minor (but undesirable) perturbation, not a major hurdle.

The obvious areas for parallelization are:

1. filter channel radiance calibration and
2. DACS processing.

There are just over 500 filter channels requiring identical processing from raw data numbers into calibrated radiances and uncertainties. This can be thought of as a loop by channel in the source code, and we propose allowing several loop iterations to be performed as separate threads in parallel. There are no dependencies between threads, so this is easy to implement, and the speedup should be largely proportional to the number of available processors.

The DACS can be processed as 4 separate threads (one per DACS), or, more likely, will be processed sequentially with the parallelism at a lower level. The latter approach is likely to be more efficient since one of the main tasks of DACS processing is the FFT for which efficient parallel libraries are likely to exist for all potential computer architectures that could be used for MLS production processing.

The above methods of parallelism will be built into the software from its initial release used with the EM, and further refinements will only be included in the unlikely event that further reduction in processing time is required.

7.7 Summary

EOS MLS Level 1 processing time is highly dependent on the chosen computer system, but unlikely to exceed 20 minutes of “wall clock time” to process a day’s worth of data. Main memory footprint is likely to be well under 100 MB, daily I/O will be of order 10 GB, and daily output data volume of order 4 GB. It has been shown that I/O traffic can be readily minimized if necessary, and processing algorithms will be implemented to run in parallel where appropriate to make efficient use of available CPU resources. It has also been shown that I/O time dominates, reducing the value of optimization of CPU resources compared to the more significant gains derived from steps taken to eliminate unnecessary I/O and minimize overall I/O. It should be pointed out that the current SCF with 8 GB of RAM is capable of keeping all input and output files memory resident if necessary.

We conclude that the resource requirements of EOS MLS Level 1 software will be reasonable and acceptable, especially since the estimates generated above do not assume any increase in capability over the current version of the SCF.

Chapter 8

Additional topics relevant to Level 1 data processing

8.1 Quality control and exception handling

It is essential that Level 1 processing software be robust when it encounters unexpected or absent data. Such events will arise during the instrument turn on sequence, from single event upsets which impart errors in telemetered data, and possibly from failures or degradations in instrument hardware. User inputs (from command files or the runstream command string) allow the processing of any engineering or science data to be inhibited. This provides a mechanism for eliminating the processing of data points which display characteristics of such a nature that it is impossible (or not worth) devising software workarounds to deal with them.

This is best explained by an example. Consider the case of an intermittent ambient calibration target temperature sensor which returns occasional bad data. The errors in such data could be so small that the conversion from counts to temperature units presents no problems within Level 1 processing, but since these data are used in conjunction with the data from the other temperature sensors on the target, the radiometric gain calibration algorithms could output incorrect results at a level which impacts science data quality. The approach taken on UARS to handle data hits in calibration target temperature telemetry, and which will be used in this software, is to compare the temperatures reported by all related temperature sensors and reject any readings which lie outside of a predetermined scatter range. This action is also reported in the Level 1 log file so that the situation can be examined off-line to determine whether the sensor should be declared bad. To date no engineering sensors have failed or even noticeably degraded on UARS, but occasional data hits, attributed to single event upsets, have been observed in such data as that from the calibration target temperature sensors and the antenna shaft angle encoder.

Engineering data is generally processed by a limited set of software procedures (functions) which perform conversions of calibration and monitor point data sets into engineering units such as Volts, Amps and Celsius. For each specific monitor point there will exist a database entry which defines the upper and lower limits for both the calibration readings and the datum being converted. If any of these limits are violated, the data point will be marked as “bad” and a log file entry generated.

For limb radiance measurements Level 1 processing makes no sophisticated attempts to

determine data validity other than declaring radiances outside the range 0 to ~ 350 K to be “bad,” and rejecting baseline radiances outside of ranges set by user inputs. It is possible, and indeed necessary, to perform quality checks on reference measurements before they are used for calibration of the associated limb data. Space view reference measurements are processed to determine their χ^2 , a valuable indicator of system stability. This is done by calculating the rms scatter of these measurements against the quadratic fit performed during radiance calibrations. These data have proven extremely valuable on UARS MLS. Space reference measurements which occur more than 6 standard deviations from the interpolating quadratic are rejected, and the fitting process repeated without these data. Ambient target measurement sequences are filtered for data data points similarly.

Other instrument performance parameters which will be determined routinely include system temperatures and time series of interpolated channel space counts and gains. These, together with the full complement of engineering data, serve as valuable quality control information which will be plotted and examined on a daily basis once the instrument is in orbit.

8.2 In-orbit “tuning” of algorithms

In addition to use of in-orbit data for validation of instrument operation and validation of Level 1 processing discussed further below, in-orbit data provides vital information necessary for fine tuning of some parameters obtained from prelaunch estimates. Important examples of data/algorithms which will be tuned using in-orbit data include:

1. Antenna loss and scattering parameters (ρ and η),
2. gain variation ($\frac{1}{f}$ noise) parameters (breakpoints and slopes),
3. radiometer-to-radiometer relative boresights,
4. instrument frame of reference with respect that of the spacecraft, and
5. baseline characteristics.

Experience with UARS indicates that in most cases simple refinements to prelaunch parameters will arise from analysis of in-orbit data. Relative boresights of the radiometers will be enhanced from study of both routine atmospheric radiances, and signals obtained from those occasions when the moon traverses the antenna FOV. Special scan sequences will be employed to maximize the data return from the moon scans, which only happen a few times a year, and this refinement is likely to improve incrementally as new data are acquired. Refining the algorithms which predict instrument pointing in the Earth’s reference frame is also likely to be an incremental process. Baseline characterization is an example where we will assume that EOS MLS behaves like its UARS predecessor, and careful study of in-orbit data will be required to characterize any idiosyncrasies in the observed data, and relevant algorithms updated based on observations of true instrument behavior.

Another major area in which algorithm tuning may be required concerns unanticipated behavior in instrument data. The approach taken with this software is to build in resilience against expected or likely events such as

1. a non-standard scan sequence which does not produce enough (or any) radiometric calibration sequences within the calibration window for the algorithms described in Chapter 4,
2. data “hits” (e.g., from cosmic ray induced single event upsets in electronic components) which create temporary data inconsistencies, and
3. data out of prescribed limits, as will happen during turn-on sequences.

On UARS MLS data characteristics were found to change after more than a year of in-orbit operation due to wear in the bearings of the switching mirror and antenna scanning mechanism. The changes to science data were not handled in an optimal manner by the pre-launch Level 1 software, and changes were eventually made to both on-board instrument software and the ground processing software to minimize, and in some cases eliminate, any impact to science data quality. Such quirks cannot be predicted in advance, and the approach taken here is not to try to anticipate and program against the unlikely, but design and build robust software which can be cleanly and safely enhanced when such behavioral anomalies have been analyzed and characterized.

8.3 Use of Level 1 algorithms in instrument testing

The primary motivations for using Level 1 software with the EM and FM versions of the instrument during instrument integration and testing are to test both the software and the instrument. The EM does not implement the full radiometer or spectrometer complement, but generates every data type to be processed by this software with the exception of that from the DACS. This allows the filter spectrometer radiometric calibration algorithms and all engineering data processing algorithms to be extensively exercised relatively early in the program. Level 1 software will be used to routinely calibrate and catalog the instrument engineering data for both versions of the instrument. Functional verification tests of system noise levels and sensitivity will also exercise the radiometric calibration algorithms for the filter channels, but the “science” data from many tests will require off line processing unavailable in this software. Such tests include end-to-end spectral sweeps and relative sideband measurements.

We anticipate that the instrument ground test environment will be far more volatile than that seen routinely in orbit, and thus provide an excellent robustness test for this software.

8.4 Validation

Instrument testing is likely to uncover most errors in algorithms or coding of this software, but in-flight data will provide a rich environment for validating many aspects of the instrument, and also many output data products from Level 1, as was the case for UARS MLS. By implication, the validation of data products is also a validation of the algorithms and implementation used to generate them, and the calibration parameters used in the processing algorithms.

Certain features of atmospheric radiances are highly predictable and dependable, providing ideal data for validating both instrument operation and the associated software. When observing the limb at tangent heights of 100 km or more, most filter channels should report

radiances close to that of cold space. Conversely, when observing with tangent heights close to, or slightly below, the Earth's surface, most channels will return saturated radiances.

Appendix A

Significant differences from UARS MLS

One of the more significant differences between UARS and EOS MLS is the choice of a continuous scan. This was implemented on EOS MLS to provide radiance measurements more densely spaced in the vertical than for UARS MLS, and thus allow better vertical resolution in retrieved geophysical parameters. The EOS MLS nominal integration time is chosen to provide negligible or acceptable “smearing” of the FOV during individual integrations. The limb scan is slower in the upper troposphere and lower stratosphere to provide more observation on these regions of the atmosphere which are currently of greater scientific interest than in the middle and upper stratosphere. The $\frac{1}{6}$ s integration time corresponds to FOV vertical movement at the tangent point of ~ 0.3 km in the upper troposphere and lower stratosphere, and ~ 1 km in the middle and lower stratosphere.

The continuous scan has no impact on Level 1 processing algorithms, but the following instrumental differences result in changes or additions to the software:

1. Two radiometric calibration targets are included in the GHz module instead of a single target as on UARS MLS. One of these targets is thermally well-coupled to the GHz structure and serves as the primary warm reference during in-flight operation. The other target, which is similar in construction (i.e., is not intended to have degraded emissivity), is thermally decoupled from the structure and attached to a radiator patch so that during in-orbit operation it floats to a temperature ~ 20 K below the primary target. This target also contains resistive heaters which allow its temperature to be raised 20–25 K above the temperature of the primary target during ground testing. The power to heat this target is provided internally from the instrument, but this capability will be removed prior to launch. The secondary target thus provides a valuable aliveness test during all phases of ground testing which was not available on the prior instrument. In addition, it serves as a in-flight backup should there be any problem with either the primary target or the Switching Mirror system which prevents use of the primary target.
2. A frequency synthesizer is included in the instrument to allow in-flight spectral calibration of the 11 and 25-channel filter bank spectrometers. This addition is included because of the long design life of EOS MLS (5 years in orbit) compared to UARS MLS (18 months).

3. Four high resolution digital autocorrelator spectrometers (DACS) are included on EOS MLS for accurate mesospheric and upper stratospheric measurements. Although minor in terms of instrument construction, these devices provide approximately half of the data generated by the instrument, and are a significant workload for the Level 1 processing software and hardware.
4. The data transferred from the instrument to the spacecraft for processing on the ground is packet oriented (conforming to CCSDS packetization conventions¹) as opposed to the fixed timing, fixed record structure generated by UARS MLS. This means that the fixed length record (with fixed location contents) file structure used for the prior instrument at Level 1 is inappropriate, and an HDF structure² conforming as closely as possible to EOS Project guidelines will be used for EOS MLS.

From the viewpoint of Level 1 processing the differences between UARS and EOS MLS are small. The most significant differences arise from the introduction of the DACS with their completely new processing requirements, and the slightly variable length MAF which eliminates some algorithm performance optimizations possible with the fixed length UARS MAF.

¹Consultative Committee for Space Data Systems.

²Hierarchical Data Format.

Appendix B

Heterodyne radiometers and black body radiation

This chapter is taken from Appendix B from [4], modified slightly for inclusion in this document.

The MLS heterodyne radiometers receive power $h\nu/\{\exp(h\nu/kt) - 1\}$ per unit frequency range when viewing a black body source which completely fills their FOV, where ν is frequency, h is Planck's constant and k is Boltzmann's constant. Our objective here is to relate the power received by a coherent (heterodyne) radiometer to the temperature of a black body which completely fills its FOV. By 'coherent' we mean that electromagnetic radiation is coupled to the radiometer in a manner which preserves its phase — this places constraints on the modes which are received and influences the effective area of the aperture which 'collects' the radiation. Let $I_\nu(\theta, \phi)$ be the intensity (Watts Hz⁻¹ m⁻² ster⁻¹) of unpolarized radiation incident upon a collecting aperture from direction (θ, ϕ) . The power in frequency interval $d\nu$ delivered through the aperture to a single-polarization radiometer can then be written

$$dP_\nu = \frac{1}{2} d\nu \int_{\Omega} I_\nu(\theta, \phi) A_e(\theta, \phi) d\Omega , \quad (\text{B.1})$$

where $A_e(\theta, \phi)$ is the effective collecting area, the factor of $\frac{1}{2}$ is due the radiometer accepting only one polarization, and the integral is over solid angle Ω .

A general expression for $\int A_e(\theta, \phi) d\Omega$ is derived from considerations of a thermodynamic equilibrium situation. Let the collecting aperture be immersed in a cavity of black body radiation, and let thermal equilibrium be established at temperature T between the black body and the radiation in the transmission line which matches the aperture to the radiometer. If V is volume of the cavity, then the black body radiation intensity is given by

$$I_\nu^{BB} = \left\{ \frac{\mathcal{N}_3(\nu) \langle E \rangle_T}{V} \right\} \left\{ \frac{c}{4\pi} \right\} . \quad (\text{B.2})$$

The first factor in braces in (B.2) is the average energy per unit volume; the second is the conversion to isotropic radiation intensity with c being the speed of light. $\mathcal{N}_3(\nu)$ is the number of modes per unit frequency interval in the 3-dimensional cavity, and $\langle E \rangle_T = h\nu / \{\exp(h\nu/kT) - 1\}$ is the average energy in a single mode at temperature T and frequency

ν [e.g., Vol. 3, chap 4 of *Feynman et al.*, 1963]. The thermal equilibrium power in the transmission line within the frequency interval $d\nu$ which is moving towards the aperture is

$$dP_\nu^T = \left\{ \frac{\mathcal{N}_1 \langle E \rangle_T}{L} \right\} \left\{ \frac{v}{2} \right\} d\nu , \quad (\text{B.3})$$

where L is the line length. The first factor in braces in (B.3) is the average energy per unit length in the line, and the second converts it to power moving towards the aperture where v is propagation speed in the line (in equilibrium, half the power moves towards the aperture and half moves away from it). \mathcal{N}_1 is the number of modes per unit frequency, and $\langle E \rangle_T$ is the same as in (B.2) since the average energy per mode at thermal equilibrium depends only on temperature and frequency. In thermal equilibrium the transmission line power moving towards the aperture will equal that collected from the radiation field by the aperture and delivered to the line. Setting (B.3) equal to (B.1), and using (B.2) for $I_\nu(\theta, \phi)$ which can be taken outside the integral since the black body radiation is isotropic, leads to

$$\int_{\Omega} A_e(\theta, \phi) d\Omega = 4\pi \frac{v}{c} \frac{\mathcal{N}_1/L}{\mathcal{N}_3(\nu)/V} . \quad (\text{B.4})$$

Using (B.2) and (B.4) in (B.1) gives, for a black body source,

$$\frac{dP_\nu^{BB}}{d\nu} = \langle E \rangle_T \frac{\mathcal{N}_1}{L} \frac{v}{2} . \quad (\text{B.5})$$

The significant difference between (B.5) and (B.3) is that (B.3) requires the transmission line to be in thermal equilibrium with the black body, whereas (B.5) does not. Heterodyne radiometers, such as in MLS, use a ‘single mode’ transmission line for which $\mathcal{N}_1 = 2L/v$, so (B.5) becomes

$$\frac{dP_\nu^{BB}}{d\nu} = \langle E \rangle_T , \quad (\text{B.6})$$

$$= h\nu / \{ \exp(h\nu/kT) - 1 \} , \quad (\text{B.7})$$

which is the relation between the black body temperature and the power per unit frequency received by a heterodyne radiometer whose FOV is completely filled by the black body.

Although it is not needed for the above derivation, the number of modes per unit frequency for a 3-dimensional blackbody cavity of volume V is $\mathcal{N}_3(\nu) = 8\pi\nu^2 V/c^3$. When this and the expression given above for \mathcal{N}_1 are used in (B.4), we obtain

$$\int_{\Omega} A_e(\theta, \phi) d\Omega = c^2/\nu^2 = \lambda^2 , \quad (\text{B.8})$$

where λ is wavelength of the radiation. By invoking detailed balancing at thermal equilibrium (the principle that equilibrium must apply to *each* frequency, direction and polarization [e.g., section 9-15 of *Reif*, 1965]), and using the preceding arguments leading to equation (B.4), the effective aperture area for collecting radiation from direction (θ, ϕ) is shown to be given by

$$A_e(\theta, \phi) = \frac{\lambda^2}{4\pi} G(\theta, \phi) , \quad (\text{B.9})$$

where $G(\theta, \phi)$ is the aperture's angular 'gain'. (If unit power is delivered to the aperture by the line, the amount radiated within solid angle $d\Omega$ in direction (θ, ϕ) is $G(\theta, \phi) d\Omega/4\pi$). Note that $\int G(\theta, \phi) d\Omega = 4\pi$, as follows from using (B.9) in (B.8). Combining (B.1) and (B.9) gives

$$dP_\nu = d\nu \frac{1}{4\pi} \int_{\Omega} \dot{I}_\nu(\theta, \phi) G(\theta, \phi) d\Omega , \quad (\text{B.10})$$

where

$$\dot{I}_\nu = \frac{1}{2} \lambda^2 I_\nu . \quad (\text{B.11})$$

Equation (B.10) is applied several places in this document. Equations (B.8), (B.9) and (B.10) describe general properties of coherent radiometer systems. The derivations given here are based on notes from classes taught by A.H. Barrett and D.H. Staelin at the Massachusetts Institute of Technology.

References

- Feynman, R. P., R. B. Leighton, and M. Sands, *The Feynman Lectures on Physics*, Addison-Wesley, New York, 1963.
- Reif, F., *Fundamentals of Statistical and Thermal Physics*, McGraw-Hill, New York, 1965.

Appendix C

Calibration and noise

In this appendix we discuss the effects of noise on a total power measurement system such as EOS MLS. The familiar expressions for the sensitivity of an individual measurement are derived, and extended to allow for the effects on calibration measurements of gain variations which increase in spectral density with decreasing frequency, usually referred to as “ $\frac{1}{f}$ ” noise. The following discussion builds largely on a prior work describing the sensitivity of total power radiometers with periodic calibration [5], and retains similar terminology. It should be noted that the description which follows does not relate exclusively to total power measurements systems, but is applicable to *any* measurement system which combines discrete calibration/reference and scene measurements.

In the interests of keeping the mathematical expressions compact and comprehensible, the assumptions are made that gain variations between calibration measurements are small compared to the random noise on an individual measurement, and that the noise on a single reference measurement is of the same magnitude as the noise on a single Limb measurement. The first assumption is reasonable for an instrument with the characteristics expected for EOS MLS, and the second is removed in the algorithms implemented for Level 1 processing.

In the latter part of this appendix we discuss the partitioning of the uncertainties in calibrated radiance into two components: a random component which appears as noise on the spectral contrast in a measured spectral feature; and a correlated component which appears as an overall gain variation which primary serves to create uncertainty in the absolute value (offset) of an observed spectrum. These two components are determined separately during Level 1 processing to suit the needs of Level 2 retrieval algorithms.

C.1 Description of the measurement system

MLS integrates and digitizes observed radiances with constant integration times, equally spaced in time. A nominal measurement sequence consists typically of a set of Limb (L) integrations lasting 20 s followed by measurements viewing the Space port (S) and internal ambient Calibration Target (C). Space measurements are of duration ~ 2 s, and Target measurements ~ 1 s. All measurements are broken into short, regularly spaced and fixed duration intervals called Minor Frames (MIF) of nominal duration $\frac{1}{6}$ s, and the entire measurement sequence is repeated once every Major Frame (MAF), of nominal duration 24.7 s. MAFs are of slightly varying duration (but always comprise an integer number of MIFs) to accommodate their synchronization to the spacecraft orbital period, and there is a delay consisting of

a small integer number of MIFs as scan and switching mechanisms transition between Limb, Space and Target views. Further details were provided in Chapter 1 of this document.

This timing is shown in Figure C.1 which indicates a single Limb observation together with the most recent preceding and succeeding calibration measurement pairs. Note that all data integrations are of constant duration, and that calibration measurements consist of a sequence of consecutive views. The period between successive data integrations is very small (~ 5 ms) compared to the data integration time ($\frac{1}{6}$ s nominal). The start of the first limb data integration following a calibration measurement sequence denotes the start of a MAF. During Level 1 processing we need to determine the estimated reference¹ and gain for each channel at the times of the Limb observations.

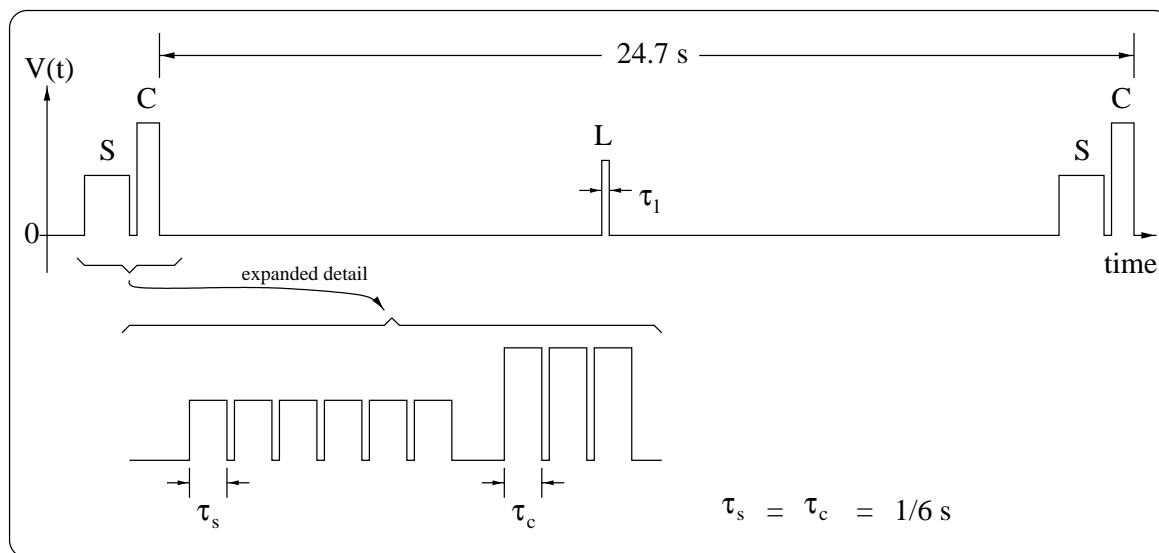


Figure C.1: Figure showing the relative output voltages and nominal timings for a single Limb observation (L) and adjacent sequences of Space (S) and ambient Calibration Target (C) reference observations. The vertical axis represents the output from a spectrometer channel, which in the case of EOS MLS is a digitized quantity. The lower portion of the figure expands the calibration views to show that they consist of sequences of data integrations, all of which are of the same duration (i.e., $\tau_l = \tau_s = \tau_c$). The interval between successive data integrations is greatly exaggerated.

It is common practice to express the sensitivity of a total power radiometer in terms of noise equivalent temperature difference, ΔT , given by:

$$\Delta T = T_{\text{sys}} \sqrt{\frac{1}{B\tau} + \left(\frac{\Delta G}{G}\right)^2} \quad (\text{C.1})$$

where:

T_{sys} is the system temperature, commonly provided by a Y-factor (hot/cold load) measurement,

¹For this discussion we assume that only the Space reference measurements are to be interpolated to the times of the Limb measurements, but in practice the software is structured to allow either Space or Target views to be selected as the primary reference to be differenced from the Limb measurements.

B is the predetection noise bandwidth of the measurement channel,
 τ is the postdetection integration time of the measurement, and
 $\frac{\Delta G}{G}$ represents the normalized rms fluctuation of radiometer power gain.

Although this equation indicates the noise on an individual signal integration, it ignores completely the effects of low frequency noise on the calibration measurements used in the determination of calibrated radiance. For a system with $\frac{1}{f}$ noise the full relationship between ΔT and T_{sys} clearly needs to include details of the timing of the calibration measurements with respect to the limb measurement, and also details of the noise power spectrum.

A simple illustration of this measurement system is shown in Figure C.2 in which the radiometer response function (H) and channel noise power (S_r) are shown as a function of frequency. Two points must be clearly understood in this figure; the horizontal axis indicates frequency in spectrometer post-detector *output* domain, not the radiometer input domain, and $H(f)$ is a power response function derived from the function which provides the estimate of the difference between a Limb measurement and its surrounding reference measurements (calibration filter).

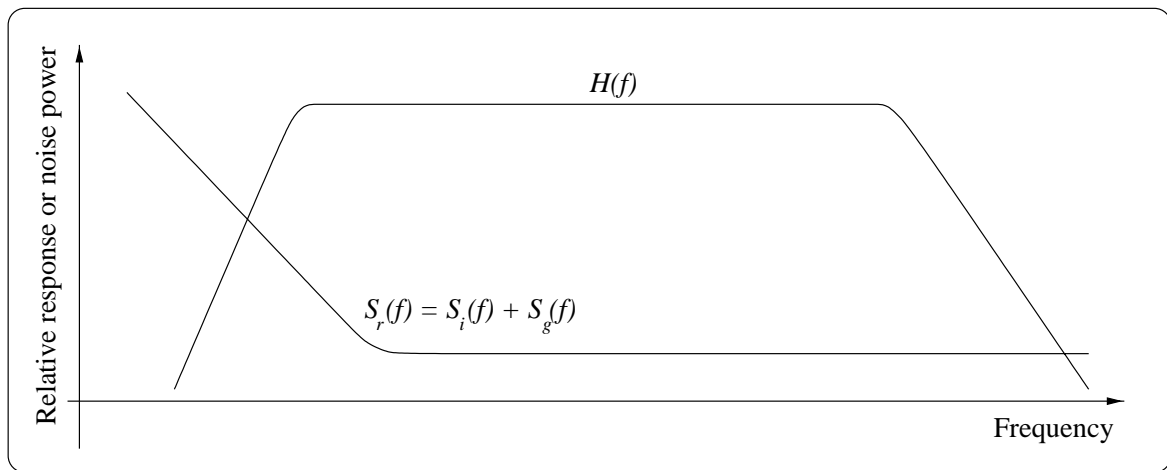


Figure C.2: Simplified representation of the noise power spectrum, $S(f)$, and the calibration filter power response function, $H(f)$, in the post-detector frequency domain. The noise power spectrum has 2 components; a frequency independent (white noise) component $S_i(f)$ and a “ $\frac{1}{f}$ ” component $S_g(f)$. The calibration filter response function, $H(f)$, is highly idealized, but displays the essential characteristics that its response falls to zero at DC and as frequency tends to infinity.

To illustrate this more clearly, consider the simple measurement sequence shown in Figure C.3 in which a single limb measurement is differenced from the next Space reference measurement. Both measurements are single MIF integrations (i.e., $\tau_l = \tau_s = \tau$) with the Limb measurement centered at $t = 0$ and the reference measurement at $t = T$. These measurements are combined to determine their difference by subtracting the reference signal from the Limb signal, shown by the weightings in the lower half of the figure. The estimate of Limb minus reference signal difference is given by the filter shown in the lower half of the figure. These weightings also represent the integration periods for both measurements, and

for an input signal $e^{2\pi i f t}$, the output of the calibration filter, $R(f)$, is given by

$$R(f) = \frac{1}{\tau} \left(\int_{-\frac{\tau}{2}}^{+\frac{\tau}{2}} e^{2\pi i f t} dt - \int_{T-\frac{\tau}{2}}^{T+\frac{\tau}{2}} e^{2\pi i f t} dt \right) \quad (\text{C.2})$$

The normalization factor $\frac{1}{\tau}$ is included to indicate that we require unity DC input to an integrator for unit time produces unity output. The right hand side of this equation evaluates to produce

$$R(f) = \frac{\sin(\pi f \tau)}{(\pi f \tau)} - \frac{\sin(\pi f \tau)}{(\pi f \tau)} e^{2\pi i f T} \quad (\text{C.3})$$

where the sinc functions arise from the integrator responses and the phase factor is a result of the temporal separation (by time T) of the two integrations.

The desired radiometer response function for noise power, $H(f)$, is simply the squared magnitude of $R(f)$, or $R(f)R^*(f)$.

To be useful the calibration filter must include more information than just that from the closest reference view. In practice we combine the data from several reference views enclosing the Limb measurement to obtain a lower noise estimate of the reference at the time of the Limb view.

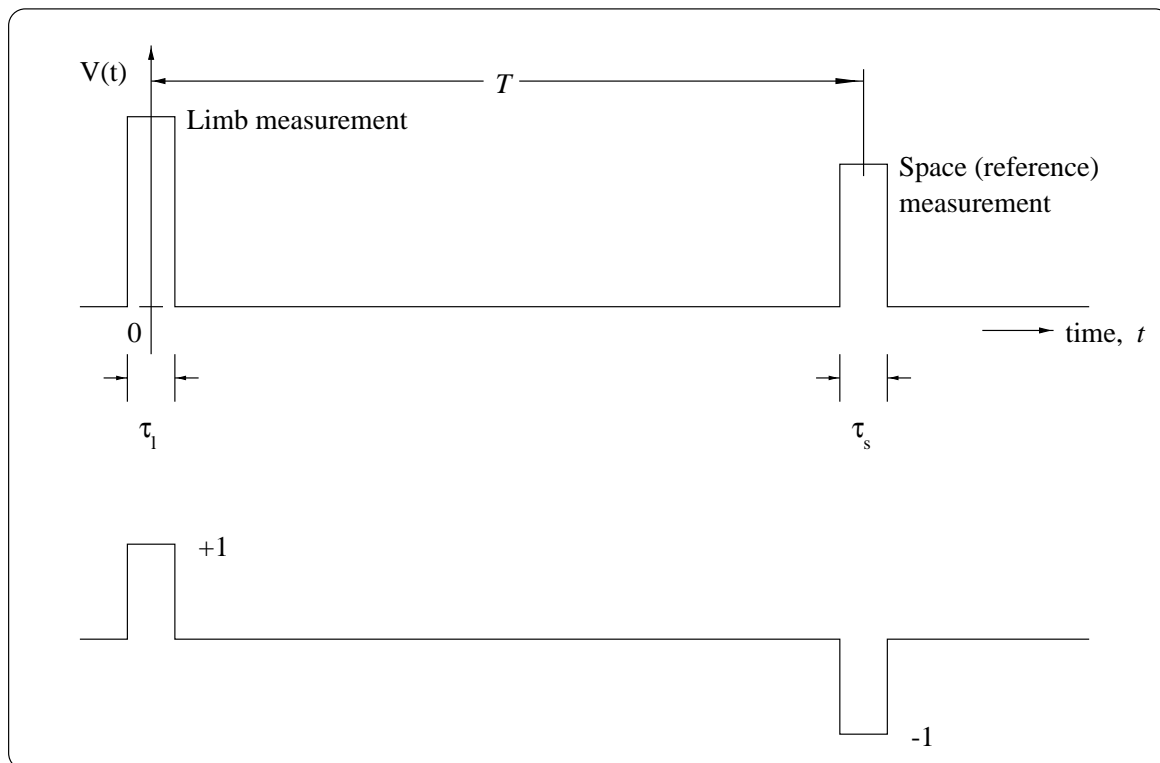


Figure C.3: The upper plot indicates the output of a channel while viewing the Limb (at $t = 0$) followed by a view to the Space reference (at $t = T$). The lower plot indicates the relative weightings applied to the two measurements by a simple calibration filter which merely differences the two measurements.

Equation C.2, and the corresponding expression for $H(f)$, are readily extended for the case in which multiple reference views are combined and differenced from a single Limb measurement:

$$H(f) = \left| \text{sinc}(\pi f \tau) - \text{sinc}(\pi f \tau) \sum_k w(t - kt_c) e^{2\pi i f(t - kt_c)} \right|^2 \quad (\text{C.4})$$

where the factors w represent the weighting applied to each reference measurement, and for convenience of representation we have assumed that all reference measurements are equally spaced in time (by time t_c) and symmetrically disposed about the single Limb measurement. The weights are subject to constraints which ensure that they produce the desired interpolate, discussed in the next chapter.

The noise power, $(\Delta T)^2$, in a single Limb measurement is given by the convolution of the post detector noise power spectrum, $S(f)$, and the power response of the post-detector system, $H(f)$:

$$(\Delta T)^2 = c^2 \int_0^\infty S_r(f) H(f) df \quad (\text{C.5})$$

where c is the channel gain, usually expressed in units of Kelvin per volt, or, in the case of EOS MLS, Kelvin per count. Minor assumptions which have been made, both of which are reasonable for EOS MLS, are:

1. the noise in c has been neglected, and
2. the noise power spectrum has been assumed to be the same for both signal and reference views.

Returning to more familiar representations, the “white noise” component of radiometer noise, $S_i(f)$, is given by:

$$S_i(f) = \frac{2T_{sys}^2}{c^2 B} \quad \text{for } 0 \leq f \ll B \quad (\text{C.6})$$

in units of V^2/Hz or $\text{Counts}^2/\text{Hz}$ where B is the measurement channel predetection noise bandwidth in Hz, and the previously stated assumptions have been retained. The condition on bandwidths given to the right of this equation is discussed further in Appendix F. Combining previous results gives:

$$\left(\frac{\Delta T}{T_{sys}} \right)^2 = \frac{1}{B\tau} + \frac{1}{B\tau} \sum_k w^2(t - kt_c) + \left(\frac{\Delta G}{G} \right)^2 \quad (\text{C.7})$$

where the left hand term on the right hand side of this equation gives the white noise contribution from the Limb measurement, the center term gives the white noise term on the combined reference measurements, and the right hand term gives the “ $\frac{1}{f}$ ” contribution to the Limb/reference difference. The gain variation term is given by:

$$\left(\frac{\Delta G}{G} \right)^2 = \left(\frac{c}{T_{sys}} \right)^2 \int_0^\infty S_g(f) H(f) df \quad (\text{C.8})$$

where $S_g(f)$ represents the noise component with non-white spectral density.

C.2 Correlated and uncorrelated noise

For Level 2 processing needs it is convenient to separate the uncertainty contributions generated by the spectrally flat and $\frac{1}{f}$ components of $S(f)$. Analyses of radiances which rely solely on spectral contrast are influenced almost entirely by the spectrally flat noise component, while analyses which depend upon accurate knowledge of absolute radiances need to include the uncertainty contribution arising from gain variations, given by the $\frac{1}{f}$ component.

We refer to the uncertainty generated by spectrally flat noise as “uncorrelated noise,” and compute this during Level 1 processing for each individual limb radiance. The “correlated noise” arising from gain variations is assumed to be identical for all channels of a given radiometer (this will be tested on the EM and FM versions of the instrument), and computed as an uncertainty vector for each active radiometer. The elements of this vector are single numbers for each MIF of the current MAF record, and the correlated noise for a given limb radiance is determined by multiplying the sum of the system temperature and calibrated limb radiance by the vector element corresponding to the MIF during which the limb measurement was made.

The form of correlated and uncorrelated uncertainties during a MAF is similar, shown in Figure C.4. The uncertainties are smallest for limb radiances measured close to reference measurements, and largest for measurements made approximately mid-way between reference measurements. For analyses which rely upon the absolute values of observed radiances it is necessary to combine both sources of error (in quadrature) to obtain the total estimated uncertainty. In order to determine these noise contributions separately we make some minor

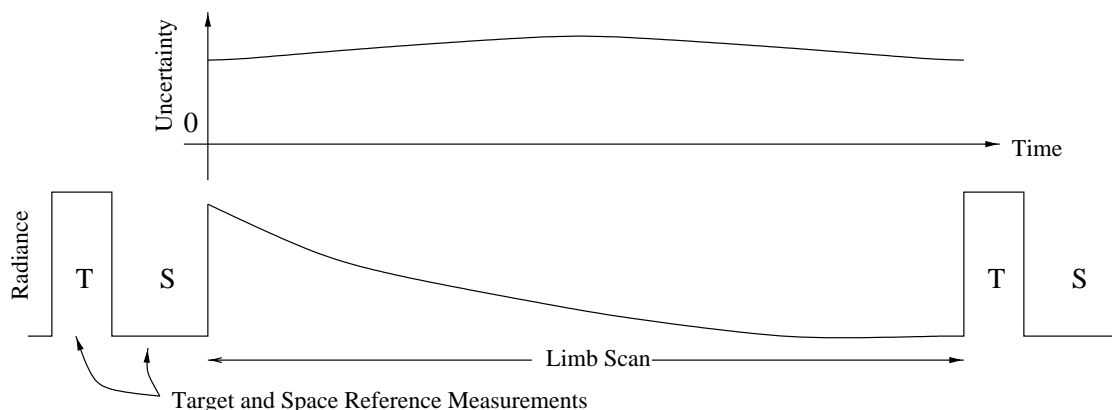


Figure C.4: Figure showing the behavior of radiance uncertainties arising from both correlated and uncorrelated sources. The uncertainties are seen to be smaller for limb radiance measurements made closest to the reference measurements, and largest for measurements mid-way between reference measurements. It is assumed for this figure that T_{sys} is much larger than any observed radiance, resulting in an uncertainty plot symmetric about the center of the limb scan. This approximation is not made in Level 1 processing.

approximations, discussed below.

C.2.1 Uncorrelated noise

The random noise, ΔT , on an individual radiance measurement is given by

$$\Delta T \approx \frac{(T_{sys} + T_{sig})}{\sqrt{B\tau}} \quad (\text{C.9})$$

The interpolation scheme used to calibrate each limb radiance also includes the noise on the reference measurements, and additional uncertainty due to the interpolation process. These additional noise contributions are much smaller than the noise on an individual limb observation, but are included in the final estimate of the random (spectral) component of uncertainty.

Including these additional sources of noise gives for the estimate of uncertainty:

$$\Delta T = \sqrt{\frac{(T_{sys} + T_{sig})^2}{B\tau} + (\Delta R)^2 + \left(T_{sig} \times \frac{\Delta G}{G}\right)^2} \quad (\text{C.10})$$

This noise estimate is calculated for all channels each measured limb radiance. ΔR is the noise on the interpolated space reference, derived from the error covariances in the quadratic fit used for interpolation, and ΔG is the noise on the interpolated channel gain. It is important not to confuse the right hand term inside the square root which contains the noise-induced uncertainty in channel gain with the similar looking gain variation term discussed earlier (see Section 4.4.1). It is similarly important to note that this expression for uncertainty scales the signal radiance by the noise on interpolated gain, not the sum of signal and system generated signals. The reason for this is that the interpolation of space references has essentially removed the gain variation effects from the data processing to determine spectral contrast. The gain variation component of uncertainty is not ignored, but evaluated and included as part of the determination of absolute radiance uncertainties, described later. The main assumptions made in expressing relative radiance uncertainty using Equation C.10 are

1. that the apodizing function applied to the measurement variances during the quadratic fit used for reference interpolation serves as a low enough bandwidth filter to allow radiometer noise to dominate any $\frac{1}{f}$ components, and
2. that gain variations over the timescale of a MAF introduce scaling errors small compared to ΔT on an individual data integration.

For EOS MLS to meet its sensitivity requirements both of the above conditions will also have to be met. Tests performed on brassboard radiometers and IF subsystems indicate that these requirements will be satisfied.

Each individual calibrated limb radiance produced by Level 1 software and written to the output radiance file includes an estimate of the random component of uncertainty arising from front end radiometer noise calculated using Equation C.10 above.

C.2.2 Correlated noise

The correlated noise determination depends upon knowledge of the $\frac{1}{f}$ characteristics (break-point frequency and slope) of the post-detector outputs of the channels of a given radiometer. UARS experience indicates that this can only be determined accurately in the final orbital

operating environment, and [11] describes a method for determining these parameters which separates the effects of orbital harmonics arising from periodic thermal variations from the underlying random gain variations. The expressions for covariance derived in Appendix E (Equations E.13 in particular) then allow the magnitude of this uncertainty source to be computed.

Appendix D

Quadratic interpolation of reference radiance signals

This appendix provides additional material to augment that presented in Section 4.3.1, and in particular clarifies the need for a common weighting function and calibration window for the quadratic reference interpolation used with all channels of a radiometer.

A significant computational task of the Level 1 software is the interpolation of space reference measurements onto the times of the limb and target reference measurements. We ideally require that this interpolation be performed in a manner which minimizes the uncertainty in the interpolated results for each measurement channel. In practice however we must not allow the interpolation to introduce any biases which may create spectral artifacts. This prevents use of an Optimal Calibrator such as the one developed for use with UARS MLS and described in Appendix E. An Optimal Calibrator is one which implements an interpolation scheme which minimizes the variance on the reference interpolates when the measurements are contaminated by $\frac{1}{f}$ noise (typical for gain variations). The description of the Optimal Calibrator is included in this document since it provides a precise description of how to determine the uncertainty (noise) on a calibrated measurement for data with noise characteristics expected for this instrument

The total power measurement scheme used by EOS MLS is described in Appendix C. Key points of relevance to the following discussion are (1) reference measurements are taken periodically in groups, and (2) the temporal separation of these reference measurement groups is of order half a minute.

In order to ensure that the noise on the interpolated reference is substantially lower than that on each individual limb radiance measurement we need to combine the information from several reference measurements during interpolation. The ideal situation is to have reference measurements disposed symmetrically about the limb measurement being calibrated, a situation not achievable in reality, but which is approximated closely enough by combining data from several groups of reference measurements on either side of the limb radiance being calibrated.

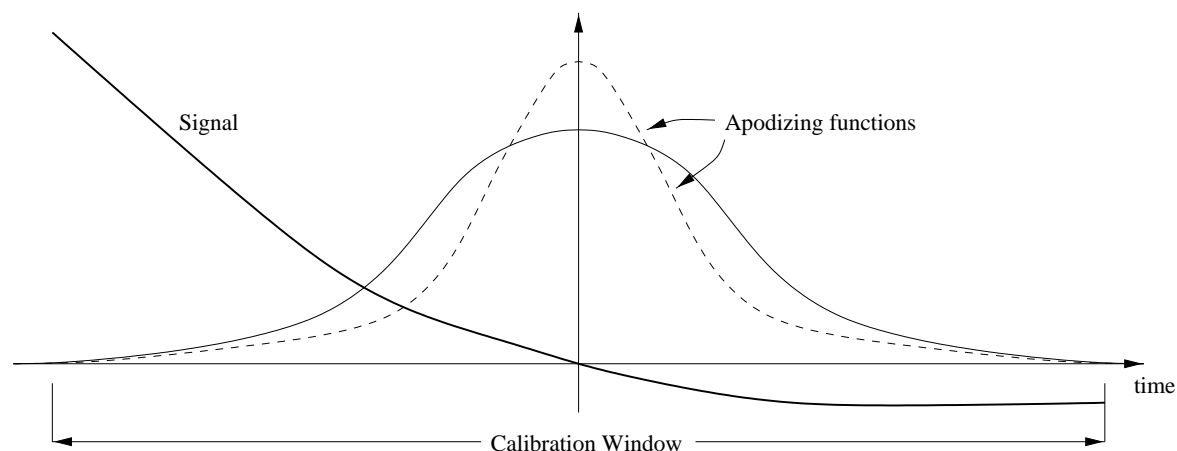


Figure D.1: Figure illustrating a signal with drift and two possible apodizing functions to be applied prior to fitting for interpolation. The vertical axis corresponds to either signal “counts” or apodizing magnitude. See text for details.

D.1 Interpolation and spectral bias

Consider Figure D.1 in which the bold solid curve represents the “true” reference signal as a function of time, and two different apodizing functions are shown. The two example apodizing functions shown differ in halfwidth, the difference in magnitude representing appropriate normalization. The magnitudes of the apodizing functions signify the relative weights applied to the reference measurements during the fitting of the interpolation function to the data. The references are being interpolated onto the time at the center of the plot (i.e., at the location of the y axis), and we size the Calibration Window so that it spans several groups of reference measurements¹.

If we assume that the bold curve labeled “signal” represents a contiguous sequence of reference measurements, it is clear by observation (and a little thought) that the mean value of the apodized signal is different for the two different apodizing functions. Both apodizing functions result in a mean signal above the horizontal axis of the example plot, the narrower function giving in a mean signal much closer to the axis than the broad one. The noise on each measurement comes predominantly from two sources, (1) a spectrally flat random component mainly from the radiometer front-ends, and (2) a component arising from gain variations in the signal chains. The gain variations arise both from thermal changes in the signal paths during the period of the Calibration Window, and from $\frac{1}{f}$ type gain variations in the HEMT amplifiers used as first IF amplifiers in all radiometers. We anticipate that the gain variations will be highly correlated in all channels of a given band, and well correlated even between bands of a given radiometer. This hypothesis is felt to be well founded based on discussions with experts in the field, and will be verified using EM, FM and in-flight data.

An optimal interpolator will select broad apodizing functions for the channels with narrowest predetection bandwidths and narrow ones for the wide channels. For the example

¹The Calibration Window is the maximum time interval over which reference measurements are included when performing the fit to these measurements for interpolation purposes.

illustrated in the figure above the optimal interpolator would bias the estimate of the signal at the time represented by the intersection of the two axes more for narrow than for broad channels. Since individual limb measurements are not interpolated, the limb/reference differences would thus look like an inverted emission spectrum for this example if both signals were spectrally flat (or even if they were both taken looking at a common reference and merely relabeled for analysis purposes). Since most MLS measurements are derived from the spectral contrast in the observed spectrum, this is clearly unacceptable. Note that these biases only arise if the order of the drift in the signal is higher than the order of the fitting polynomial (or constraint order in the case of the Optimal Interpolator). Such drift characteristics may arise for instance when the sun impinges on a previously eclipsed portion of the MLS containing sensitive signal chain electronics.

D.2 Apodizing functions and the length of the Calibration Window

The previous discussion makes it clear that measurements which are derived from spectral contrast of the observed radiances require the use of a common apodizing function for all measurements which are to be used as an ensemble in subsequent data processing (i.e., bands which are “standalone” require use of a single apodizing function for all channels within the band, and bands whose spectral data are combined in some sense require use of a common apodizing function for all channels of those bands).

We strongly desire that the random component of uncertainty on an individual limb radiance be dominated by radiometric noise, and not significantly degraded during radiometric calibration processing by noise on the associated reference measurements. Based on the discussions on noise and calibration in Appendix C, if gain variations do not contribute significantly to uncertainty, we have for the noise on an individual calibrated measurement:

$$\Delta T = T_{\text{sys}} \sqrt{\frac{1}{B\tau} + \frac{1}{B\tau} \sum_k w^2(t - kt_c)} \quad (\text{D.1})$$

Space reference measurements occur in contiguous groups of twelve meaning that 24 space reference measurements are used to generate the interpolated estimate of space signal at the time of a limb measurement in the simple case that a linear fit is used with data from adjacent calibration measurement groups. In the equation above this means that a common weighting, w , of $\frac{1}{24}$ is used, resulting in a signal to noise degradation due to noise on the reference measurements of only 2%.

Experience with UARS MLS indicates that we wish to use a larger calibration window than this, and to fit to the space references with a quadratic, in order to capture second order components in the signal drifts. The previous arguments also indicate that a short calibration window is appropriate since this minimizes any deleterious effects of $\frac{1}{f}$ noise, and a longer window makes scarcely any improvement in the noise of the calibrated limb radiance. An apodizing function strongly favors calibration measurements closest to the measurement being calibrated, and tapers to a low value for the most distant calibrations is appropriate. The chosen scheme is shown in Figure D.2 below. Further details of measurement timing are given in Appendix C.

As shown in Figure D.2, we choose a Calibration Window of 6 MAFs since this provides 3 reference measurement groups in either side of the central MAF of limb radiances being

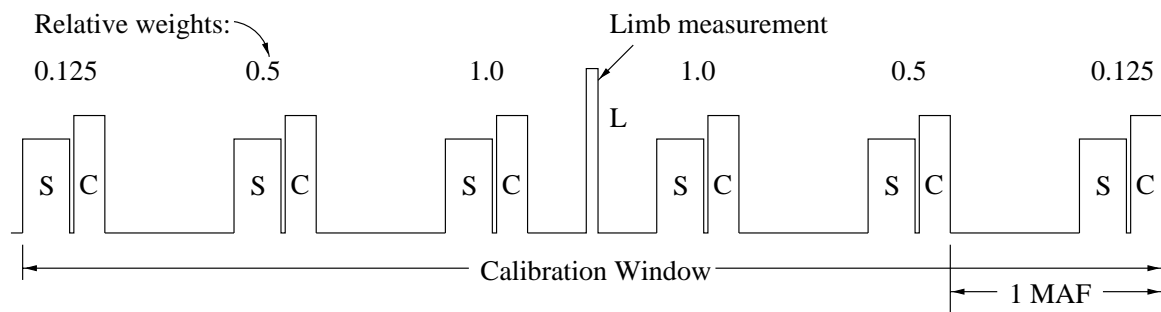


Figure D.2: Figure illustrating relative timings of calibrations used with a given limb radiance measurement. S, C and L represent Space, Calibration Target and Limb views respectively. The Central MAF is the one containing the Limb view being processed. Relative weights are the weightings applied to each calibration measurement when fitting to the calibration sequence for interpolation. Figure is not to scale. See text for additional details.

processed. The relative durations of all measurements in the figure are exaggerated for clarity. Relative weightings of 1 and 0.125 are applied to the calibration measurements closest and furthest from the limb measurement being processed, with a weighting of 0.5 for the intermediate calibrations. These weightings may change slightly after data from the EM and FM have been analyzed, but the algorithms given below are unlikely to change.

The short duration of the proposed Calibration Window relative to the one used for UARS MLS data processing (~ 1.5 minutes *vs* ~ 12 minutes) justifies the assumption above that the noise on individual calibrated radiances will be dominated by the spectrally flat component.

D.3 The quadratic interpolator formulated as a sequence of weights

In order to determine absolute radiance errors (Section 4.4.2) it is necessary to express the operation of generating the difference between limb and interpolated reference counts as a sequence of weights applied to each measurement. This calculation is performed off-line as part of the determination of the absolute radiance error multipliers supplied to Level 1 as user inputs, but described here so that the process is documented in a convenient place.

The quadratic least squares fit consists of minimizing the function f given by:

$$f(a, b, c) = \sum_j (a + bx_j + cx_j^2 - y_j)^2 \quad (\text{D.2})$$

where a , b and c are the quadratic coefficients, and x_j are the “times” at which the measurements y_j were made. To use this equation as an interpolator at “time” zero we need only determine the value of coefficient a . Note that any measurement sequence can be interpolated onto any time simply by offsetting the time coordinates to define the interpolation point to occur at time zero. The solution for a which minimizes the rms difference between measurements and model is given in section 4.3.1, and reproduced here in slightly modified

form:

$$a = \frac{1}{\Delta} \begin{vmatrix} \Sigma C(j)W(j) & \Sigma jW(j) & \Sigma j^2W(j) \\ \Sigma jC(j)W(j) & \Sigma j^2W(j) & \Sigma j^3W(j) \\ \Sigma j^2C(j)W(j) & \Sigma j^3W(j) & \Sigma j^4W(j) \end{vmatrix} \quad (\text{D.3})$$

where $C(j)$ are the counts and $W(j)$ are the relative weightings given to the measurement at time j , and Δ is a constant for any given measurement sequence. From this relationship we may directly write:

$$\begin{aligned} a'(j) = & \frac{C(j)}{\Delta} \left(W(j) \times \left(\Sigma j^2W(j) \times \Sigma j^4W(j) - \Sigma j^3W(j) \times \Sigma j^3W(j) \right) - \right. \\ & \Sigma jW(j) \times \left(jW(j) \times \Sigma j^4W(j) - j^2W(j) \times \Sigma j^3W(j) \right) + \\ & \left. \Sigma j^2W(j) \times \left(jW(j) \times \Sigma j^3W(j) + j^2W(j) \times \Sigma j^2W(j) \right) \right) \end{aligned} \quad (\text{D.4})$$

where

$$a = \Sigma a'(j) = \Sigma w(j)C(j) \quad (\text{D.5})$$

Note that we have now expressed a as a sequence of weights $w(j)$ which can be applied directly to the measurements $C(j)$ (as a dot product) to determine the interpolate of C at time zero (i.e., $j = 0$). A further simplification arises from the recognition that a simple normalization applies to the weights:

$$\Sigma w(j) = 1 \quad (\text{D.6})$$

since we require unit result when applying the weights to a uniform input. This removes the need to evaluate Δ , a significant computational saving.

An important point to note in the expression for w is that the weights do not depend upon the input data at all, only on their position with respect to the interpolation point, and on the relative weighting given to each measurement when performing the fit. Since all measurements in a module (GHz or THz) share the same reference timing, and the arguments given earlier show that a common relative weighting must be given to related measurements, once the vector of weights to be applied to the measurements has been determined, the same vector is then used with the data from all channels sharing the same input weighting to determine the interpolate at a given MIF. This is in sharp contrast with the more conventional approach in which the data dependent coefficients a , b and c are determined for each channel and then used to generate the desired interpolates at each MIF. The conventional approach requires a complex calculation initially, followed by a set of relatively trivial calculations to determine the interpolates. The ‘‘vector of weights’’ approach described above substitutes a simple calculation that must be repeated for each MIF onto which we are interpolating, but has the additional saving that many channels can share vector w in determining the interpolates.

D.3.1 Relative radiance uncertainty estimate

The noise variance on an individual limb integration, $(\Delta T_{limb})^2$, is given by:

$$(\Delta T_{limb})^2 = \frac{(T_{sys} + T_{limb})^2}{B\tau} \quad (\text{D.7})$$

The noise variance on the interpolated reference measurement, $(\Delta R)^2$, is given by:

$$(\Delta R)^2 = \frac{(T_{sys} + T_{ref})^2}{B\tau} \times \Sigma w^2 \quad (\text{D.8})$$

In Chapter 4 it is shown that the variance contribution from noise on the interpolated gains, $(\Delta T_g)^2$, is given by:

$$(\Delta T_g)^2 = \frac{T_{limb}^2}{B\tau} \times \Sigma w_g^2 \quad (\text{D.9})$$

where w_g are the normalized weights used to quadratically interpolate the gain measurements onto the times of the limb measurements.

We thus arrive at an expression for the relative uncertainty on an individual calibrated limb measurement:

$$\Delta T_{rel} = \frac{1}{B\tau} \sqrt{(T_{sys} + T_{limb})^2 + (T_{sys} + T_{ref})^2 \times \Sigma w^2 + T_{limb}^2 \times \Sigma w_g^2} \quad (\text{D.10})$$

The relative ease with which this expression can be evaluated should be compared to those given in Section 4.4.1.

D.4 Some examples

The operation of the weights-based interpolator is best clarified by some examples of the weight sequences arising for anticipated interpolation scenarios. Figure D.3 closely simulates the nominal measurement sequence planned for EOS MLS with consecutive groups of 12 MIFs devoted to observation of cold space. These reference groups repeat every 149 MIFs, and, as can be seen from the figure, 6 groups of reference measurements are included in the fit for the interpolation. The relative weighting given to the input data falls off exponentially with temporal distance from the MIF onto which the references are being interpolated with a $\frac{1}{e}$ length of 150 MIFs, corresponding to the description in Section 4.3.1.

The upper left panel shows the absolute weights applied to the reference measurements when interpolating onto the MIF in the precise center of the reference measurement sequence. The quantity Σw^2 in the panel is the sum of the squares of the interpolation weights, and indicates the square of the noise contribution from the interpolated reference (where unity would indicate that the reference noise contribution was of the same magnitude as the noise on a single low radiance limb measurement). The upper right panel shows the weights applied to the reference measurements to generate the interpolate for a MIF adjacent to one reference measurement group. Note that the noise on the interpolated reference is slightly larger than in the previous case.

The lower two panels give the weights for generating an interpolate just beyond and well beyond the rightmost reference measurement respectively. Note the huge increase in the uncertainty on the interpolate as the quadratic is used as an extrapolator instead of an interpolator.

D.5 Non-standard measurement sequences

It is necessary for Level 1 software to operate robustly even when the measurement sequence departs radically from nominal. Non standard sequences will occur during pre-launch instrument testing and characterization, and will occur in flight when scan programs are being

updated or special test are being performed. Experience with UARS MLS indicates that the primary conditions that must be accounted for are:

1. a continuous sequence of measurements of one type (e.g., as when staring at the Space or Calibration Target port for extended periods in order to derive stability characteristics from off-line analyses, and
2. a Calibration Window containing insufficient (or no) reference measurements for processing of Limb data in the Central MAF.

It is imperative that Level 1 software be robust under all instrument operating conditions, i.e., that it continue to run even when the instrument is operating in a mode such that limb data cannot be calibrated (e.g., when staring at the internal calibration target for an extended period such as when characterizing signal chain stabilities).

This is handled primarily by performing extensive sort and qualify operations on all data within the calibration window prior to processing the central MAF. Once this has been done radiometric calibration is only performed if adequate calibration data exists within the calibration window. If insufficient or no calibration data is present, previous calibration data (if present) are extrapolated and used. When the estimated uncertainty on the extrapolated data exceeds preset thresholds (set by default data files), default data are used instead, and the radiance data flagged as “bad” by being tagged with a negative radiance error.

The instrument has a range of operating conditions which represent a “standard operating mode,” and for which Level 1 processing will operate seamlessly. This range is set by the allowable range of MIF and MAF durations, and by the presence of a radiometric calibration sequence at least once per MAF. Non-standard measurement sequences may violate these conditions, typically by the absence of the full radiometric calibration sequence for an extended period of time. These sequences are intended to provide data which will be derived directly from the Level 0 data files using analysis tools developed expressly for that purpose.

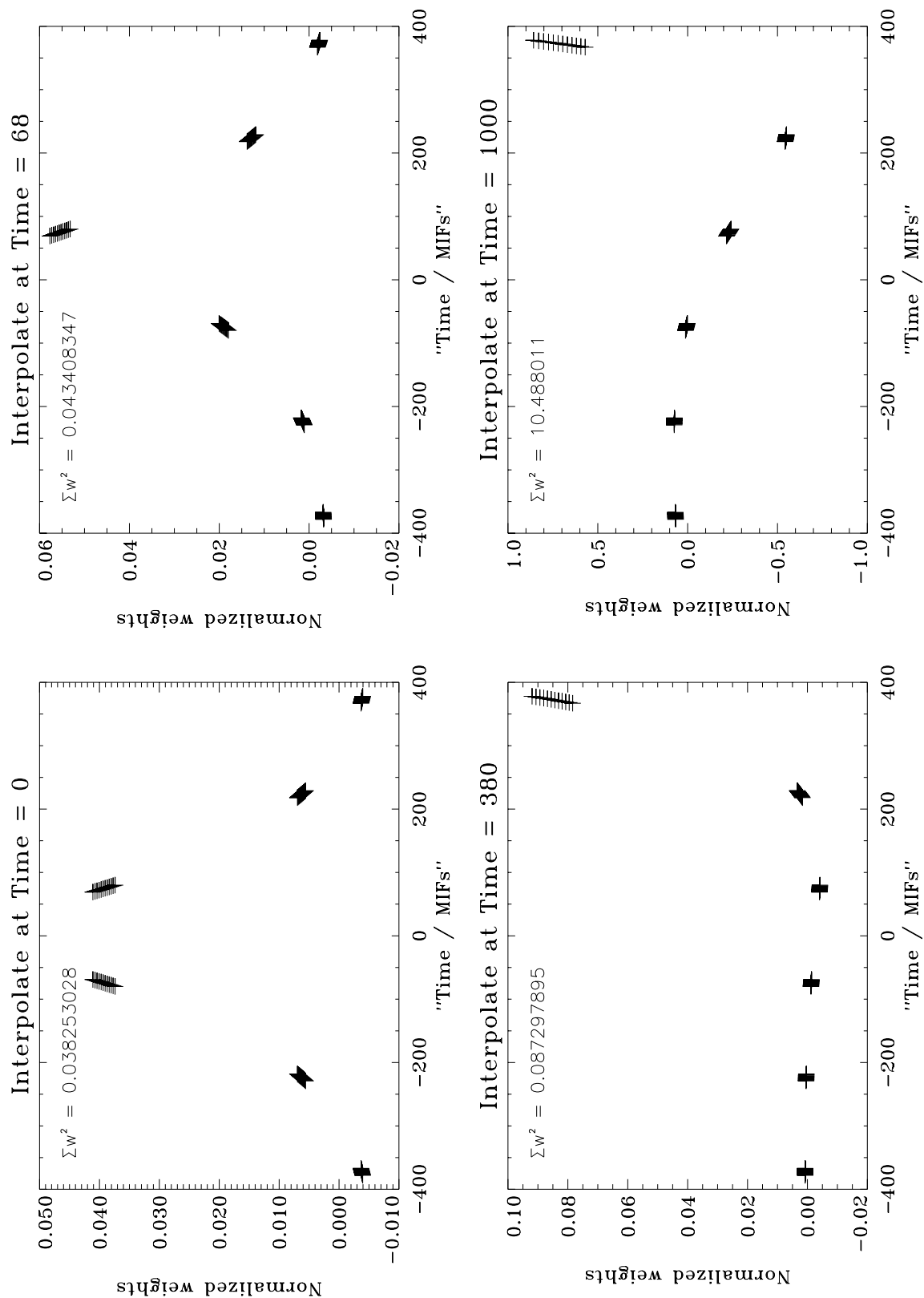


Figure D.3: These panels show the weights applied to reference measurements (as a dot product) to determine interpolates at “times” 0, 68, 380 and 1000. See text for additional details.

Appendix E

Optimal interpolation of reference radiance signals

The appendix describes the Optimal Calibrator developed by Peckham, described in [6], and adheres closely to the content and terminology used therein. This work led to the development of algorithms, given below, which evaluate calibrated limb radiances with minimum variance. It was shown in the previous appendix that such algorithms can lead to spectral bias, precluding their use with EOS MLS data during routine Level 1 data processing. The description below is included in this document because it provides the key algorithms necessary for determination of the coefficients to be used in Level 1 processing to determine the magnitude of the absolute radiance uncertainties. These coefficients will be determined with off-line (non-production) analyses of in-orbit data, and presented to Level 1 software as user inputs.

E.1 Filter weights and constraints

Combining Equations C.7 and C.8 gives

$$\left(\frac{\Delta T}{T_{\text{sys}}}\right)^2 = \underbrace{\frac{1}{B\tau}}_1 + \underbrace{\frac{1}{B\tau} \sum_k w^2(t - kt_c)}_2 + \underbrace{\left(\frac{c}{T_{\text{sys}}}\right)^2 \int_0^\infty S_g(f)H(f) df}_3 \quad (\text{E.1})$$

For a given channel noise bandwidth and integration time, the noise on a single uncalibrated Limb integration (term 1) is set by the system temperature, T_{sys} . The noise on the interpolated reference views due to system temperature (term 2) is dictated by the relative weights (w) applied to each reference measurement. In the absence of low frequency noise and gain fluctuations we would choose to interpolate using a large number of reference measurements, allowing the noise on the interpolated reference to be very small compared to the noise on the Limb integration. The power response of the calibration filter in term 3, $H(f)$, is of course a function directly related to the weights. As the temporal span covered by the calibration filter increases, the filter accepts more of the noise which increases in spectral density with decreasing frequency. The goal is to determine the set of weights, w , which minimize the right hand side of Equation E.1 while simultaneously providing the desired interpolates of the reference sequence.

E.2 Constraints

Before the set of weights which minimize the right hand side of Equation E.1 can be determined, it is necessary to determine the constraints to which they are subject. In the general case in which we wish to determine the difference, D_T , between a set of M Limb measurements (S_m) and K reference measurements (C_k), we define the following filter

$$D_T = \sum_{m=1}^M v_m S_{m+T} - \sum_{k=1}^K w_m C_{k+T} \quad (\text{E.2})$$

where D_T represents the smoothed output sequence (at time T) with any zero level (offset) removed. It should be noted that the reference sequence is typically chosen to be longer than the scene measurement sequence, and disposed as symmetrically as possible about it. In the case that all signals are time independent, and $S_m = C_k$, the output of the filter should be zero, requiring

$$\sum_{m=1}^M v_m = \sum_{k=1}^K w_m. \quad (\text{E.3})$$

Since we require drifts and low frequency gain variations to cause only slow variations in signal levels over timescales of several MAFs, it is appropriate to approximate drifts by a low order polynomial of order R in time, where $R = 2$ or 3 is likely to suffice. We may represent the signal S (or C) by

$$S = \sum_{i=0}^R a_i t^i \quad (\text{E.4})$$

(where t is the time of the center of the integration for measurement i) which combines with E.2 to give

$$\sum_{m=1}^M v_m \sum_{i=0}^R a_i t_m^i = \sum_{k=1}^K w_m \sum_{i=0}^R a_i t_k^i. \quad (\text{E.5})$$

The summations may be rearranged to give (my thanks to Dr. Nathaniel Livesey for this suggestion):

$$\sum_{i=0}^R a_i \sum_{m=1}^M v_m t_m^i = \sum_{i=0}^R a_i \sum_{k=1}^K w_m t_k^i \quad (\text{E.6})$$

which implies

$$\sum_{m=1}^M v_m t_m^r = \sum_{k=1}^K w_m t_k^r \quad \text{for } r = 0..R. \quad (\text{E.7})$$

This result is readily obtained for the case $R = 2$ by subtracting Equation E.3 from the appropriate expansion of Equation E.6. The proof for successively higher orders of R is obtained by incrementally increasing the value of R in the expansion of Equation E.6 and subtracting the constraints determined by the prior evaluations for lower orders.

For EOS MLS we have chosen to calibrate each Limb radiance individually, with no binning at Level 1. This leads to the reduced requirement that

$$\sum_{k=1}^K w_m t_k^r = t_m^r \quad \text{for } r = 0..R. \quad (\text{E.8})$$

In the derivations which follow we continue to derive the expressions for both the case of calibrating each Limb radiance individually (the baseline), and the case of calibrating multiple Limb radiances together, should the need for this ever be required at a later time. Binned radiances are typically generated during non-production analyses (e.g., for zonal mean data of species with small atmospheric signals), and it must be recognized that the noise levels estimated for the binned products will be incorrect if the covariances between adjacent limb measurements are not taken into account.

E.3 Variance and the noise spectrum

From Equation E.2 we can express the variance of the filter output, $E\{D^2\}$, where E is the expectation operator, as

$$E\{D^2\} = \sum_{mn} v_m v_n E\{S_m S_n\} + \sum_{kl} w_k w_l E\{C_k C_l\} - 2 \sum_{mk} v_m w_k E\{S_m C_k\} \quad (\text{E.9})$$

where m and n cover the range from 1 to M , and k and l cover the range from 1 to K . The reduced expression for the case of a single Limb measurement with surrounding reference measurements is given by

$$E\{D^2\} = E\{S^2\} + \sum_{kl} w_k w_l E\{C_k C_l\} - 2 \sum_k w_k E\{S C_k\}. \quad (\text{E.10})$$

For a unit amplitude input signal at frequency f the term $S_m C_k$ may be written

$$S_m C_k = \frac{1}{\tau_c \tau_l} \left(\int_{-\frac{\tau_l}{2}}^{+\frac{\tau_l}{2}} e^{2\pi i f t} df \int_{T-\frac{\tau_c}{2}}^{T+\frac{\tau_c}{2}} e^{2\pi i f t} df \right) \quad (\text{E.11})$$

where T is the interval between the centers of the two integrations (indexed by the subscripts m and n), with similar expressions for the other covariances. For UARS MLS the low frequency post-detector noise power spectrum varied as $\frac{1}{f^\alpha}$, with α ranging from ~ 1 to ~ 2.5 , with breakpoint frequencies¹ of $\sim \frac{1}{10}$ Hz and lower [11]. Similar characteristics are expected for EOS MLS. For such noise characteristics the covariance $E\{S_m C_k\}$ may thus be written

$$E\{S_m C_k\} = \frac{1}{\tau_l \tau_c} \int_0^\infty \left(1 + \left(\frac{f_c}{f} \right)^\alpha \right) \text{sinc}(\pi f \tau_l) \text{sinc}(\pi f \tau_c) \cos(2\pi f T_{mk}) df \quad (\text{E.12})$$

where f_c is the breakpoint frequency of the $\frac{1}{f}$ noise and T_{mk} is the time between the centers of the two data integrations.

Following the work of Peckham [6] we convert the expressions for the covariances into dimensionless forms by means of the substitutions $f' = \frac{f}{f_c}$, $\tau_l' = 2\pi f_c \tau_l$, $\tau_c' = 2\pi f_c \tau_c$, $T_k' = 2\pi f_c T_k'$, etc. Dropping the primes, the full expressions for the covariances become

$$\begin{aligned} E\{S_m S_n\} &= \frac{f_c}{\tau_l'^2} \int_0^\infty (1 + f'^{-\alpha}) \text{sinc}^2\left(\frac{f' \tau_l'}{2}\right) \cos(f' T_{mn}) df \\ E\{C_k C_l\} &= \frac{f_c}{\tau_c'^2} \int_0^\infty (1 + f'^{-\alpha}) \text{sinc}^2\left(\frac{f' \tau_c'}{2}\right) \cos(f' T_{kl}) df \\ E\{S_m C_k\} &= \frac{f_c}{\tau_l' \tau_c'} \int_0^\infty (1 + f'^{-\alpha}) \text{sinc}\left(\frac{f' \tau_l'}{2}\right) \text{sinc}\left(\frac{f' \tau_c'}{2}\right) \cos(f' T_{mk}) df \end{aligned} \quad (\text{E.13})$$

¹The breakpoint frequency, f_c , is the frequency at which the $\frac{1}{f}$ and spectrally flat noise power spectral densities are equal.

where T_{ab} is the time difference (in the appropriate units) between the centers of integrations a and b .

E.4 Minimum variance estimate

Equation E.9 may be rewritten

$$E\{D^2\} = c + \sum_{kl} A_{kl} w_k w_l - 2 \sum_k b_k w_k \quad (\text{E.14})$$

where

$$c = \sum_{mn} v_m v_n E\{S_m S_n\}, \quad A_{kl} = E\{C_k C_l\}$$

and

$$b_k = \sum_m v_m E\{S_m C_k\}$$

Introducing Lagrange multipliers, λ_r , the minimum of $E\{D^2\}$, subject to the constraints of Equation E.7, occurs when w_k satisfy

$$\sum_{l=1}^K A_{kl} w_l + \sum_{r=1}^{R+1} k^{r-1} \lambda_r = b_k \quad (\text{E.15})$$

for $k = 1$ to K . Note that the range of r has been changed from 0 to R (in Equation E.7) to 1 to $R + 1$ to conform to the usual matrix notation. These equations, together with the constraints themselves, may be summarized in matrix form as

$$\begin{pmatrix} \mathbf{A} & \mathbf{P} \\ \mathbf{P}^T & \mathbf{O} \end{pmatrix} \begin{pmatrix} \mathbf{w} \\ \lambda \end{pmatrix} = \begin{pmatrix} \mathbf{b} \\ \mathbf{q} \end{pmatrix} \quad (\text{E.16})$$

Here, \mathbf{A} is a square matrix of dimension K , \mathbf{O} is the null matrix of dimension $R+1$, $P_{kr} = k^{r-1}$ and

$$q_r = \sum_{m=1}^M (m + \beta)^{r-1} \nu_m$$

The minimum variance corresponding to these values is given by

$$E\{D^2\} = c - (\mathbf{b}^T \mathbf{q}^T) \begin{pmatrix} \mathbf{w} \\ \lambda \end{pmatrix} \quad (\text{E.17})$$

Efficient evaluation of the integrals in Equations E.13 is described in [6]. Since we will only evaluate these expressions during off-line processing to obtain the data needed for determination of absolute radiance offsets for EOS MLS, efficiency is not paramount. As discussed in Chapter 4, for determination of the absolute radiance uncertainties we only need to evaluate the right hand term of Equation E.1. This is readily accomplished using Simpson's rule.

E.5 Notation

The symbols used in this Appendix occasionally duplicate those used with different meanings elsewhere in this document. The table below lists the first occurrence of each symbol here, together with a brief description.

Notation used in this chapter

Symbol	units	Equation	description
f	Hz	E.12	post-detector signal frequency
f_c	Hz	E.12	breakpoint frequency of post-detector $\frac{1}{f}$ noise component
i		E.4	measurement index
i		E.12	the square root of minus one
k		E.2	target measurement index
m		E.2	reference measurement index
\mathbf{q}		E.16	vector – see text
t	s	E.12	time
t^i	s	E.4	the time corresponding to measurement i
v_m		E.2	sample weighting applied to Space reference measurement m
w_m		E.2	sample weighting applied to Target reference measurement m
\mathbf{A}		E.16	covariance matrix
C_k	Counts	E.11	Target counts at time index k
D_T	Counts	E.2	Smoothed time sequence of signal minus reference counts
E	Counts ²	E.9	variance of output interpolate
K		E.2	total number of Target reference measurements used in an interpolation
M		E.2	total number of Space reference measurements used in an interpolation
O		E.16	null matrix
P		E.16	matrix – see text
R		E.4	order of interpolating polynomial
S	Counts	E.4	a sequence of reference measurements
S_m	Counts	E.11	Space counts at time index m
T	s	E.2	the time onto which the interpolate is being determined
α		E.12	slope of $\frac{1}{f}$ component of noise power spectrum
λ_r		E.15	vector of Lagrange multipliers
τ_c	s	E.11	integration time of each target reference measurement
τ_l	s	E.11	integration time of each limb measurement

Appendix F

Integrator and detector noise relationships

In this appendix we discuss the noise bandwidth of an integrator and the noise characteristics at the output of a square law detector. The assumptions that are conventionally made in the derivation of the standard results are also elaborated.

F.1 Noise bandwidth of an integrator

The noise bandwidth, f_n , of an electronic filter with frequency response in the voltage domain of $A_v(f)$ is given by

$$f_n = \frac{1}{A_v(f_0)A_v^*(f_0)} \int_0^\infty A_v(f)A_v^*(f) df \quad (\text{F.1})$$

where $A_v(f_0)$ is the maximum gain of the filter (at frequency f_0). This noise bandwidth is the equivalent bandwidth of a hypothetical filter (i.e., a rectangular one with infinitely sharp cutoffs) which passes the same noise power as the “real” filter represented by A_v . For integrators such as are used in MLS as the post-detector filter-digitizers we have the situation that the voltage gain of the filter is a maximum at zero frequency, and is directly proportional to the integration time. Consider the basic integrator circuit shown in Figure F.1. For an

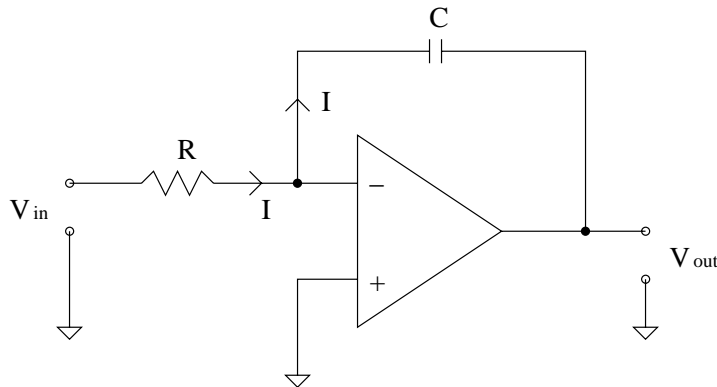


Figure F.1: Schematic of an electronic integrator.

integrator in which $RC = 1$ we may define the voltage gain (which is obtained with a DC input signal) as unity per unit integration time. Using the expression for integrator response derived in Appendix C, the noise bandwidth is then given by

$$f_n = \int_0^\infty \frac{\sin^2(\pi f \tau)}{(\pi f \tau)^2} df = \frac{1}{2\tau} \quad (\text{F.2})$$

where integral is evaluated directly using the relationship

$$\int_0^\infty \frac{\sin^2(px)}{x^2} dx = \frac{\pi p}{2} \quad (\text{F.3})$$

which may be obtained from any reasonable mathematical handbook with tabulations of integrals.

An alternative method of deriving this result is by use of Parseval's theorem which formalizes the equivalence of power expressed in the frequency and time domains:

$$w_n = \int_0^\infty |H(jw)|^2 dw = \pi \int_0^\infty h^2(t) dt = \pi E \quad (\text{F.4})$$

An input signal consisting of a unit impulse (1 V for 1 s) compressed to an infinitesimally small time leads to an output voltage $\frac{1}{RC}$ for the integrator shown in Figure F.1. If a 1Ω is presented as a load at the output of the integrator, the energy dissipated in this load is $\frac{1}{(RC)^2}$ W. Normalizing by the power gain of the integrator, $\frac{\tau}{(RC)^2}$, gives

$$E = \left(\frac{RC}{\tau}\right)^2 \times \frac{\tau}{(RC)^2} = \frac{1}{\tau} \quad (\text{F.5})$$

Thus

$$w_n = 2\pi f_n = \frac{\pi}{\tau} \quad (\text{F.6})$$

or

$$f_n = \frac{1}{2\tau} \quad (\text{F.7})$$

as before.

F.2 The detection process

The following sections are based on prior work by Robinson [8]. Most practical detectors fall into one of two categories: square law or linear detectors, characterized by the law relating their output current to their input voltage. For a square law detector we have:

$$I = \alpha V^2, \quad (\text{F.8})$$

and for a linear detector:

$$\begin{aligned} I &= \alpha V, & V &\geq 0 \\ I &= 0, & V &< 0 \end{aligned} \quad (\text{F.9})$$

At sufficiently low input levels, all detectors exhibit square law characteristics, and at high enough input levels, all behave approximately like linear detectors. The tunnel diode detectors used in EOS MLS filter channels are chosen to exhibit close to ideal square law behavior over the full range of input signals, and we shall concentrate on detectors with this characteristic below.

A typical system involving a detector is shown in Figure F.2 below. The input signal consists of $V(t)$ containing both noise and signals¹ in a band of width B centered at B_0 , and the output of interest is at low frequencies from DC to B . The bandwidth B need not be well defined, but we require that the signal power be negligible at frequencies outside the limits, say $B_0 \pm 2B$, and that B is less than approximately $B_0/4$ (which is true for all MLS filter channels). The low frequency output from the detector contains components from DC to

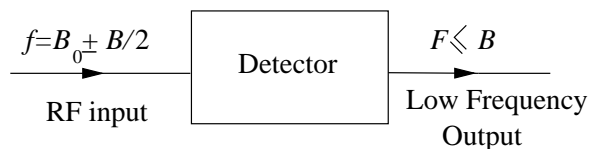


Figure F.2: Basic detector configuration. F is used to denote post detector frequencies, and f RF input signal frequencies.

B arising from difference beats between input components. There are also sum components near $2B$ which are rejected by MLS post detector electronics, which we shall neglect.

The properties of the low frequency output are discussed most conveniently in terms of the envelope of the input. Let the input be expanded as a Fourier series in an interval T so that, adopting an unit of time $T/2\pi$, we have

$$V(t) = \sum_{n=0}^{\infty} a_n \cos(nt + \phi_n), \quad (\text{F.10})$$

for which only those coefficients a_n for which

$$B_0 - \frac{B}{2} < \frac{n}{T} < B_0 + \frac{B}{2} \quad (\text{F.11})$$

have values appreciably different from zero. We may manipulate the previous expressions to obtain

$$V(t) = A \cos(n_0 t) - B \sin(n_0 t) \quad (\text{F.12})$$

where

$$\begin{aligned} A &= \sum a_n \cos((n - n_0)t + \phi_n), \\ B &= \sum a_n \sin((n - n_0)t + \phi_n), \end{aligned} \quad (\text{F.13})$$

and

$$n_0 = B_0 T \quad (\text{F.14})$$

¹For systems such as MLS measuring thermal emission, the input signal ideally comprises just thermal noise.

corresponds to the center frequency of the RF input signal band.

We can then write $V(t)$ as

$$V(t) = R(t) \cos(n_0 t + \theta), \quad (\text{F.15})$$

where

$$R(t) = \sqrt{(A^2 + B^2)} \quad (\text{F.16})$$

and

$$\tan(\theta) = A/B. \quad (\text{F.17})$$

Since n ranges only over the limited interval $(B_0 \pm B/2)T$, no component of A or B varies more rapidly with t than $B/2$, and no component of $R(t)$ more rapidly than B . Thus $R(t)$ is a slowly varying function of t compared with $\cos(n_0 t)$, and is the envelope of $V(t)$ shown in Figure F.3.

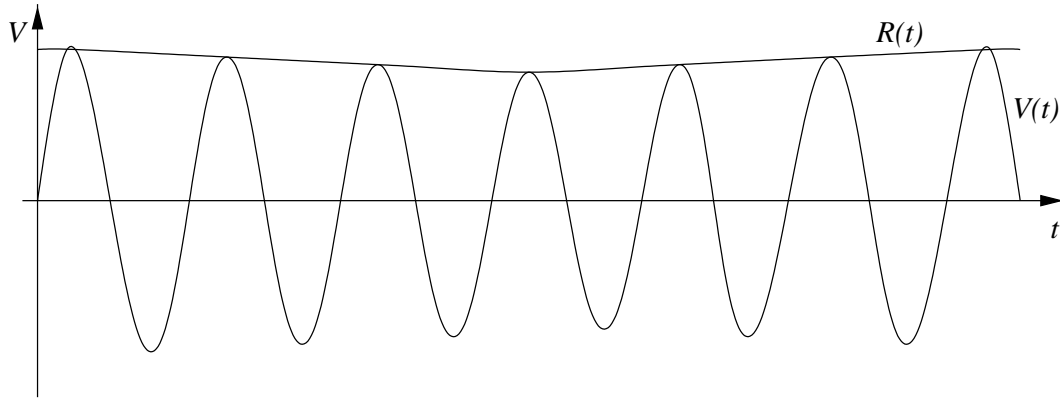


Figure F.3: The envelope $R(t)$ of an amplitude modulated signal $V(t)$.

F.3 Noise characteristics of a square law detector

A signal $V_0 \cos(n_0 t)$ applied to a square law detector results in an output current I where

$$I = \alpha V_0^2 \cos^2(n_0 t) = \frac{1}{2} \alpha V_0^2 (1 + \cos(2n_0 t)) \quad (\text{F.18})$$

and α is a constant representing the “efficiency” of the detector. We are normally just concerned with the DC term in this expression:

$$I_{dc} = \frac{\alpha}{2} V_0^2. \quad (\text{F.19})$$

If we apply an input signal $R(t) \cos(n_0 t + \theta)$, the low frequency output of the square law detector is clearly given by

$$I_{lf} = \frac{\alpha}{2} R^2(t) \quad (\text{F.20})$$

We may rewrite this equation as

$$I_{lf} = \frac{\alpha}{2}(A^2 + B^2) = \frac{\alpha}{2} \sum_n \sum_m a_n a_m \cos((n-m)t + \phi_n - \phi_m), \quad (\text{F.21})$$

or in the form of a Fourier series

$$I_{lf} = \sum_k C_k \cos(kt + \phi_k). \quad (\text{F.22})$$

recognizing that terms of frequency k arise only when $m = n \pm k$. In particular, the DC term is

$$I_{dc} = C_0 = \frac{\alpha}{2} \sum_n a_n^2. \quad (\text{F.23})$$

If the input is a noise process with a power spectrum $w(f)$ over a predetection bandwidth B centered at $F = 0$ we have

$$\sum \langle a_n^2 \rangle = 2 \int_{f_0-B/2}^{f_0+B/2} w(f) df, \quad (\text{F.24})$$

and so

$$I_{dc} = C_0 = \alpha \int_{f_0-B/2}^{f_0+B/2} w(f) df. \quad (\text{F.25})$$

The term of frequency k is given by

$$C_k \cos(kt + \theta_k) = \frac{\alpha}{2} \left(\sum_n a_n a_{n-k} \cos(kt + \phi_n - \phi_{n-k}) + \sum_n a_n a_{n+k} \cos(kt + \phi_{n+k} - \phi_n) \right). \quad (\text{F.26})$$

The two terms in this expression yield identical results, for they differ only for terms with $n < k$, i.e., $f < B$, where the input is in any case zero. We therefore have

$$C_k \cos(kt + \theta_k) = \alpha \sum_n a_n a_{n+k} \cos(kt + \phi_{n+k} - \phi_n). \quad (\text{F.27})$$

The phases ϕ_{n+k} and ϕ_n are independently and randomly distributed, and so the ensemble average of this expression is zero. Thus

$$\langle C_k \rangle = 0, \quad k \neq 0. \quad (\text{F.28})$$

If we square Equation F.27 we have

$$C_k^2 \cos^2(kt + \theta_k) = \alpha^2 \sum_n \sum_m a_n a_m a_{n+k} a_{m+k} \cos(kt + \phi_{n+k} - \phi_n) \cos(kt + \phi_{m+k} - \phi_m). \quad (\text{F.29})$$

In an ensemble average the cross-terms with $m \neq n$ drop out, again because of the independence of the phases, and we are left with

$$\langle C_k^2 \rangle = \alpha^2 \sum_n \langle a_n^2 a_{n+k}^2 \rangle. \quad (\text{F.30})$$

Since the a_n with different n are independent, this is equivalent to

$$\langle C_k^2 \rangle = \alpha^2 \sum_n \langle a_n^2 \rangle \langle a_{n+k}^2 \rangle. \quad (\text{F.31})$$

We now express $\langle C_k^2 \rangle$ in terms of a power spectrum $W(F)$, where $F = k/T$, giving

$$\begin{aligned}\langle C_k^2 \rangle &= (\text{limit } T \rightarrow \infty) 2W(F)/T, \\ \langle a_n^2 \rangle &= (\text{limit } T \rightarrow \infty) 2w(f)/T.\end{aligned}\tag{F.32}$$

Sums over n then become T times integrals over f , i.e.,

$$\sum_n \rightarrow T \int df,\tag{F.33}$$

so that

$$W(F) = 2\alpha^2 \int_{f_0 - \Delta f_0/2}^{f_0 + \Delta f_0/2} w(f)w(f + F) df.\tag{F.34}$$

We thus have now expressed the output spectrum in terms of input spectrum.

F.3.1 White noise input to a square law detector

It is normally an implicit assumption that the RF signal into a spectrometer power detector consists of white noise with constant power spectral density w within a well defined bandpass B , giving

$$I_{dc} = \alpha w B\tag{F.35}$$

and the power spectrum

$$W(F) = 2\alpha^2 w^2 B \left(1 - \frac{F}{B}\right).\tag{F.36}$$

The noise spectrum is thus seen to have its maximum intensity near DC, falling to zero at B .

The total low-frequency output noise is obtained from Equation F.36 by integrating from DC to $F \geq B$, and is given by

$$\Delta I^2 = \alpha^2 w^2 B.\tag{F.37}$$

If the bandwidth Δf of the measurement system is very much less than B , a measurement of the DC current can be made to a resolution of $I_{dc} \pm (w\Delta F)^{\frac{1}{2}}$. The signal to noise of the measurement is therefore $(2\Delta F/B)^{\frac{1}{2}}$. In terms of system temperature and ΔT we have:

$$\frac{\Delta T}{T_{\text{sys}}} = \left(\frac{2\Delta F}{B}\right)^{\frac{1}{2}}.\tag{F.38}$$

We showed earlier in this chapter that the noise bandwidth of an integrator is given by the reciprocal of twice the integration time τ , i.e.,

$$\Delta F = \frac{1}{2\tau}\tag{F.39}$$

leading to the more familiar expression

$$\frac{\Delta T}{T_{\text{sys}}} = \sqrt{\frac{1}{B\tau}}.\tag{F.40}$$

F.3.2 White noise with a CW signal

A situation likely to arise in several filter spectrometer channels on EOS MLS is that the atmospheric and electronic noise signal input to the power detector will be contaminated by an unmodulated continuous wave signal S via leakage from local oscillators used for 2nd and 3rd IF downconversion. These contaminating signals will be very stable in both frequency and amplitude. Proceeding as before, the resulting expression for the DC term at the output of the detector for a contaminating signal $S \cos(2\pi ft_0)$ at the center of the predetection passband may be shown to be

$$I_{dc} = \frac{\alpha}{2}(S^2 + 2wB), \quad (\text{F.41})$$

meaning that the signal and noise are detected independently.

The power spectrum is rather less simple than before since noise between DC and $F = B$ can arise either from beats between noise components or between noise and the CW signal, whereas noise between $B/2$ and B arises solely from noise beats alone. This results in

$$\begin{aligned} W(F) &= 2\alpha^2 w \left(S^2 + wB \left(1 - \frac{F}{B} \right) \right) \quad \text{where } 0 < F < \frac{B}{2} \\ W(F) &= 2\alpha^2 w^2 B \left(1 - \frac{F}{B} \right) \quad \text{where } \frac{B}{2} < F < B \end{aligned} \quad (\text{F.42})$$

This spectrum is illustrated in Figure F.4. The total low frequency noise obtained by integrating Equations F.42 from DC to $F \geq B$ is

$$\Delta I^2 = \alpha^2 wB(S^2 + wB). \quad (\text{F.43})$$

The presence of a steady signal within the predetection passband which significantly contributes to the detector DC output is thus seen to also increase the noise output power.

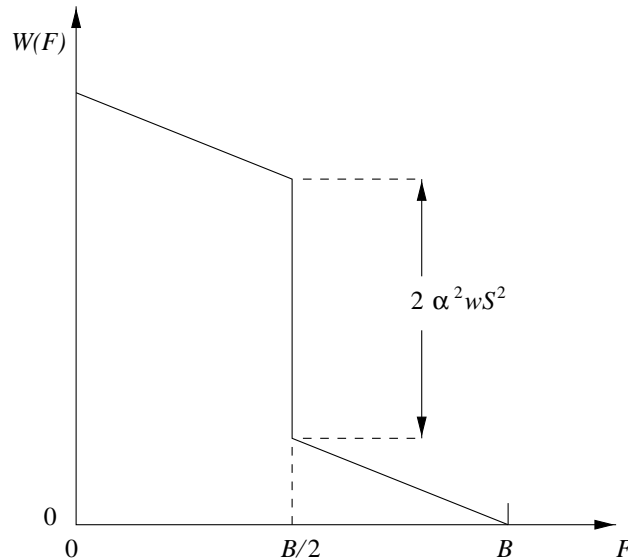


Figure F.4: Power spectrum, $W(F)$, of the low frequency output of a square law detector with white noise input in bandwidth B and a signal S at the center of the passband.

Appendix G

Digital Autocorrelator Spectrometers (DACS)

Digital autocorrelators similar to the ones implemented in EOS MLS have been used elsewhere, and their implementation and theory of operation described in the literature (e.g., [9] and [10]). For EOS MLS they serve to provide relatively narrow bandwidth (~ 10 MHz) and high resolution (~ 200 kHz) spectrometers with uniform channel spacing. Their main benefits compared to a conventional filter spectrometer arise from the largely digital implementation with resulting compactness, ease of fabrication, high stability, and relatively low power consumption.

The signal flow through an EOS MLS DACS is illustrated in Figure G.1. The signal band of interest is down-converted to baseband, then bandpass filtered to ~ 10 MHz bandwidth. This signal is then coarsely digitized to 2 bits and fed to the CMOS DACS chip where it passes through a 128 element shift register. Each of these delayed signals is multiplied by the incoming 2 bit signal, and the output of each multiplier is added to a 4 bit accumulator with the carry being fed to the main counter array. The accumulator effectively performs as a divide by 16 (i.e., 4 bits of equivalent prescale). An additional channel (not shown) accumulates the product of the incoming signal with itself (the “zero lag”). All digitization, shifting and accumulation is synchronous with the external 25 MHz clock, with asynchronous 24 bit counters to minimize logic area. Additional data buffers (not shown in the figure) are implemented to allow the 24 bit counter array contents to be saved, and the counters themselves cleared to start the next data integration before the data readout is complete. A control input allows the data accumulation to be suspended so that the asynchronous counters can settle prior to being loaded into the readout registers and cleared. The measurement of science data by the DACS and filter channels is kept coincident by using the 5 ms blanking pulse between MIFs to suspend data integration in both types of spectrometer. The data path to the RIU is byte wide, and successive bytes are addressed by an internal counter which autoincrements after each data transfer.

G.1 Digitizer and multiplier operation

We start the description of the 2 bit autocorrelator by discussing a “conventional” implementation, followed by minor modifications which result in a simplified unit with similar

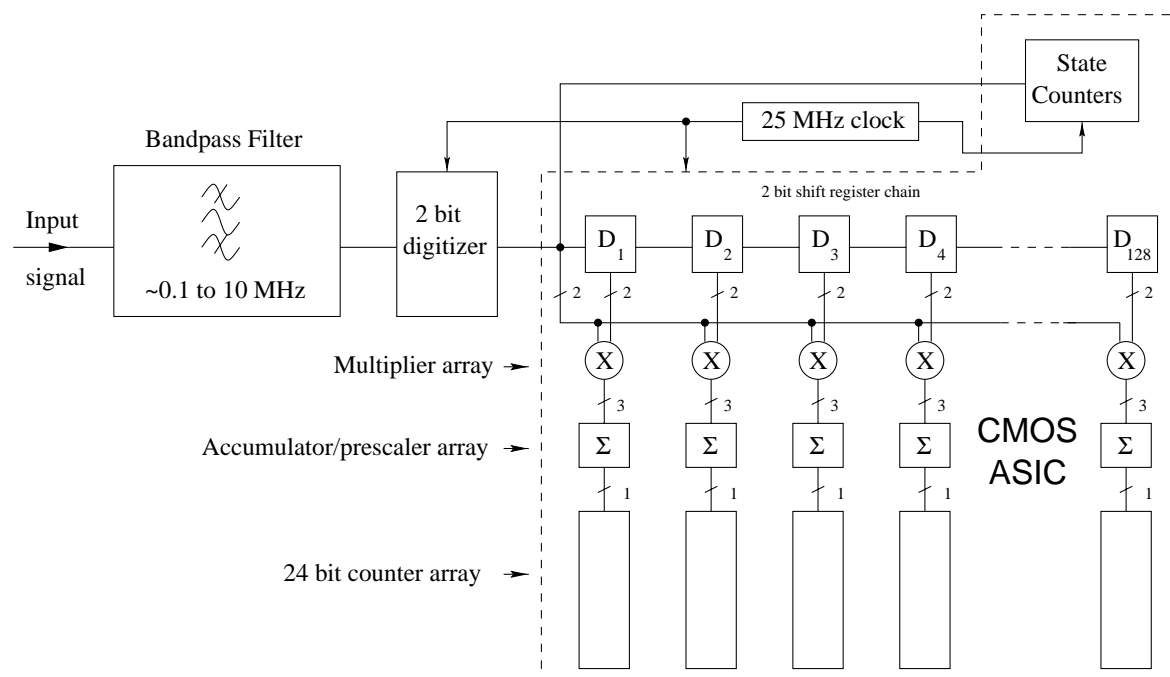


Figure G.1: Simplified schematic of the 2 bit DACS used on EOS MLS. The portion enclosed in the dashed box is fabricated on a single CMOS chip, and includes control logic for enabling/disabling and clearing the counter array, and a byte serial readout interface to read the 24 bit counter array into an accompanying RIU. The state counters record the total number of occurrences of each of the four possible digitizer output states during each data integration, and are used to determine total input RF signal power. See text for further details.

performance characteristics, as implemented on EOS MLS. The 2 bit digitizer outputs one of 4 possible states which are assigned weighting factors of $-n$, -1 , $+1$ and $+n$. When two voltages, quantized according to the scheme shown in Table G.1, are multiplied, the products shown in Table G.2 are generated.

If P_{11} , $P_{1\bar{1}}$, \dots , $P_{\bar{2}\bar{2}}$ are the probabilities of finding the two noise voltages jointly in the states designated by the associated P subscripts, and the sampling rate is N per second, the net rate of accumulation of counts in an add-subtract counter (accumulator) at the multiplier

Table G.1: The 2 bit quantization scheme implemented in an “ideal” 2 bit digital autocorrelator. The digitizer thresholds are set at voltages of $-V_0$, 0 and $+V_0$. Digitizer outputs are in 2’s complement notation.

Input Voltage Range:	$-\infty < V \leq -V_0$	$-V_0 < V \leq 0$	$0 < V \leq V_0$	$V_0 < V \leq \infty$
Sign bit	1	1	0	0
Magnitude bit	1	0	0	1
State designation	$\bar{2}$	$\bar{1}$	1	2
Weighting factor	$-n$	-1	$+1$	$+n$

output will be

$$\langle N_2 \rangle = \langle N_{22} - N_{2\bar{2}} \rangle + \langle N_{12} - N_{1\bar{2}} \rangle n^{-1} + \langle N_{11} - N_{1\bar{1}} \rangle n^{-2} \quad (\text{G.1})$$

where

$$\begin{aligned} \langle N_{11} \rangle &= 2P_{11}N, & \langle N_{1\bar{1}} \rangle &= 2P_{1\bar{1}}N, & \langle N_{12} \rangle &= 4P_{12}N \\ \langle N_{1\bar{2}} \rangle &= 4P_{12}N, & \langle N_{22} \rangle &= 2P_{22}N, & \langle N_{2\bar{2}} \rangle &= 2P_{2\bar{2}}N \end{aligned}$$

and probabilities have been equated wherever the symmetry of the system allows it.

Maximum count rate occurs with two fully correlated voltages ($\rho = 1$), under which condition the only nonzero probabilities are P_{11} and P_{22} . Hence a two-bit correlation coefficient, ρ_2 , can be defined by

$$\begin{aligned} \rho_2 &= N_2/N_{\max} \\ &= \frac{(P_{22} - P_{2\bar{2}}) + 2(P_{12} - P_{1\bar{2}})n^{-1} + (P_{11} - P_{1\bar{1}})n^{-2}}{(P_{22} + P_{11}n^{-2})_{\rho=1}} \end{aligned} \quad (\text{G.2})$$

Integrals defining the probabilities appearing in Equation G.2 take the form (assuming Gaussian statistics apply):

$$P_{11} = \frac{1}{2\pi(1-\rho^2)^{\frac{1}{2}}} \int_0^{v_0} \int_0^{v_0} \exp\left(-\frac{v_1^2 + v_2^2 - 2\rho v_1 v_2}{2(1-\rho^2)}\right) dv_1 dv_2 \quad (\text{G.3})$$

where the voltages v are normalized to the rms voltage at the digitizer input and ρ is the continuous correlation coefficient of the two signals. The other probabilities are given by the same equation with the integration limits set accordingly.

For $\rho \ll 1$ it is possible to expand the exponentials and arrive at the approximations

$$\begin{aligned} P_{11} &= \frac{1}{4}\phi^2 + \frac{\rho}{2\pi}(1-E)^2 \\ P_{1\bar{1}} &= \frac{1}{4}\phi^2 - \frac{\rho}{2\pi}(1-E)^2 \\ P_{12} &= \frac{1}{4}\phi(1-\phi) + \frac{\rho}{2\pi}E(1-E) \end{aligned}$$

Table G.2: Two bit multiplication table for the quantization scheme of Table G.1. Table entries are the outputs of each DACS multiplier for all possible combinations of delayed (V_1) and undelayed (V_2) digitized signals.

V_2 state	V_1 state			
	$\bar{2}$	$\bar{1}$	1	2
2	$-n^2$	$-n$	$+n$	$+n^2$
1	$-n$	-1	$+1$	$+n$
$\bar{1}$	$+n$	$+1$	-1	$-n$
$\bar{2}$	$+n^2$	$+n$	$-n$	$-n^2$

$$\begin{aligned}
P_{1\bar{2}} &= \frac{1}{4}\phi(1-\phi) - \frac{\rho}{2\pi}E(1-E) \\
P_{22} &= \frac{1}{4}(1-\phi)^2 + \frac{\rho}{2\pi}E^2 \\
P_{2\bar{2}} &= \frac{1}{4}(1-\phi)^2 - \frac{\rho}{2\pi}E^2
\end{aligned} \tag{G.4}$$

where $E = \exp(-\frac{1}{2}v_0^2)$, and ϕ is the probability integral $\phi(v_0)$, given by

$$\phi(v_0) = (2\pi)^{-\frac{1}{2}} \int_{-v_0}^{v_0} \exp(-\frac{1}{2}v^2)dv \tag{G.5}$$

Probabilities appearing in the denominator of Equation G.2 are the probabilities of finding either v_1 or v_2 in the range 0 to v_0 or v_0 to ∞ , that is,

$$P_{11} = \frac{1}{2}\phi(v_0), \quad P_{22} = \frac{1}{2}(1-\phi(v_0)), \quad \text{for } \rho = 1$$

Substituting these results in Equation G.2 leads to

$$\frac{\rho_2}{\rho} = \frac{2}{\pi} \left(\frac{(1-2n^{-1}+n^{-2})E^2 + 2(n^{-1}-n^{-2})E + n^{-2}}{1-\phi+\phi n^{-2}} \right) \quad \text{for } \rho \ll 1 \tag{G.6}$$

The variance on the output of the counter following each multiplier is determined by the total number of input counts. For Gaussian statistics this is given by

$$\langle N_2^2 \rangle - \langle N_2 \rangle^2 = \langle N_{22} \rangle + \langle N_{2\bar{2}} \rangle + \frac{\langle N_{12} \rangle + \langle N_{1\bar{2}} \rangle}{n^2} + \frac{\langle N_{11} \rangle + \langle N_{1\bar{1}} \rangle}{n^4} \tag{G.7}$$

In this equation the N_{12} and $N_{1\bar{2}}$ counts are smoothed by a factor of n^2 because their mean rate is reduced by the prescaling factor n and their variance is further reduced by the same factor n before they enter the main counter. Similar reasoning applies to the smoothing of the N_{11} and $N_{1\bar{1}}$ counts. Note that for this to be true we require the measurement bandwidth to be much larger than the signal bandwidth.

For low signal to noise ratio ($\rho \ll 1$), substituting from Equations G.4 gives

$$\begin{aligned}
\langle N_2^2 \rangle - \langle N_2 \rangle^2 &= N \left((1-\phi)^2 + 2\phi(1-\phi)n^{-2} + \phi^2 n^{-4} \right) \\
&= N(1-\phi+\phi n^{-2})^2 \\
&= N_{max}^2/N
\end{aligned} \tag{G.8}$$

The signal to noise of the two-bit correlator, R_2 , is given by

$$\begin{aligned}
R_2 &= \langle N_2 \rangle (\langle N_2^2 \rangle - \langle N_2 \rangle^2) \\
&= (N_2/N_{max}) N^{\frac{1}{2}} \\
&= \rho_2 N^{\frac{1}{2}}, \quad \rho_2 \ll 1
\end{aligned} \tag{G.9}$$

If the same number of samples were multiplied in a continuous correlator and averaged, the signal to noise ratio R_c would be given by

$$R_c = \rho N^{\frac{1}{2}} \tag{G.10}$$

and hence

$$R_2/R_c = \rho_2/\rho, \quad \rho \ll 1 \tag{G.11}$$

Thus the sensitivity of the two bit correlator relative to a continuous correlator is also given by the right hand side of Equation G.6. The maximum value of R_2/R_c is 0.88 for $n = 3$ and $v_0 = 1.0$.

G.2 A simplified two bit autocorrelator

A significant simplification in the implementation of the two bit correlator arises from simply omitting generation of the low level products (± 1) in Table G.2. Elimination of the appropriate terms from Equations G.2 and G.8 leads to the result

$$\frac{R_{2a}}{R_c} = \frac{2}{\pi} \frac{E^2 + 2E(1-E)n^{-1}}{((1-\phi)^2 + 2\phi(1-\phi)n^{-2})}, \quad \rho \ll 1 \quad (\text{G.12})$$

where R_{2a} is the signal to noise ratio of the simplified correlator. The ratio R_{2a}/R_c has a maximum of 0.87 for $n = 3$ and $v_0 = 0.9$. This simplified implementation is the one chosen for EOS MLS, with additional simplifications to the multiplier/accumulator which ease the design and implementation still further without any change in performance, discussed below.

G.3 The EOS MLS implementation

The multiplication table for the simplified 2 bit correlator is shown in Table G.3. With the value of n (3) chosen for the EOS MLS implementation an obvious additional simplification is to divide each multiplier output by 3 so that the quantities added into each accumulator are in the set $-3, -1, 0, +1$ and $+3$. Finally, it is possible to offset the output of the multipliers by $+3$ so that only the positive quantities 0, 2, 3, 4 or 6 are added during each accumulation. The inclusion of this offset significantly simplifies the implementation of the accumulators and counters in the DACS since only positive numbers are added to the accumulators, and only up-counters need to be implemented. The offset is removed in the instrument as part of the C&DH data compression activity. The final simplified multiplier algorithm as implemented on EOS MLS is given in the right hand side of Table G.4.

G.4 Estimate of true autocorrelation

Before the measured autocorrelation can be Fourier transformed into the frequency domain it is necessary to convert the measurements into an estimate of the continuous autocorrelation of the input signal. For the system described we estimate the continuous autocorrelation,

Table G.3: Simplified two bit multiplication table. Table entries are the outputs of each DACS multiplier for all possible combinations of delayed (V_1) and undelayed (V_2) digitized signals. This multiplication table is the same as the full one given earlier, but with the innermost products set to zero.

V_2 state	V_1 state			
	$\bar{2}$	$\bar{1}$	1	2
2	$-n^2$	$-n$	$+n$	$+n^2$
1	$-n$	0	0	$+n$
$\bar{1}$	$+n$	0	0	$-n$
$\bar{2}$	$+n^2$	$+n$	$-n$	$-n^2$

Table G.4: The unbiased multiplier algorithm for $n = 3$ is shown in the table at the left, and the biased implementation used in EOS MLS autocorrelators is shown at the right. Delayed and undelayed signals correspond to V_1 and V_2 in Table G.3.

Delayed signal state					Undelayed signal state	Delayed signal state				
2	1	1	2			2	1	1	2	
-3^2	-3	$+3$	$+3^2$	2	2	0	2	4	6	
-3	0	0	$+3$	1	1	2	3	3	4	
$+3$	0	0	-3	1	1	4	3	3	2	
$+3^2$	$+3$	-3	-3^2	2	2	6	4	2	0	

R_c , from the measured two-bit one, R_{2a} , using

$$\begin{aligned}
 R_c &= 1.146R_{2a} - 0.049R_{2a}^2 \quad \text{for } 0.0 \leq R_{2a} < 0.9 \\
 &= 1.340R_{2a} - 0.340R_{2a}^2 \quad \text{for } 0.9 \leq R_{2a} \leq 1.0
 \end{aligned} \tag{G.13}$$

where these expressions have been derived from fits to Equation G.12. These equations assume that the signal is small compared the system temperature contribution to the signal input power to the digitizers. This approximation will be removed in a future release of this document.

G.5 Gain calibration

Figure G.1 shows an analog power measurement channel (with data integrations synchronous with all other filter channel and DACS measurements) implemented to provide a measure of the total average RF signal power being analyzed during each MIF as described in Chapter 4. In Appendix H we describe the use of the counters implemented in the DACS to monitor the 2 bit A/D converter thresholds to monitor input signal power. In practice we expect this digital power monitor to closely match the performance of a dedicated analog channel, and may choose to delete the analog channel after the adequacy of the digital power monitor has been verified by tests on DACS breadboards.

G.6 Transformation to the frequency domain

The power spectrum, $P(f)$, of the DACS input signal is calculated by performing a discrete Fourier transform on the estimate of the continuous autocorrelation function, obtained from the measured 2 bit measurement of the autocorrelation function as discussed above. The M channels (corresponding the $M = 128$ delay values in the shift register) are transformed into M points in the frequency domain via the relationship:

$$P\left(\frac{j}{2M\Delta t}\right) = \frac{1}{M} \left[R(0) + 2 \sum_{m=0}^M R(m\Delta t) \times \cos(\pi m j / M) \right] \tag{G.14}$$

where

$P(\frac{j}{2M\Delta t})$ is the signal power at DACS input frequency $\frac{j}{2M\Delta t}$
 $R(0)$ is the correlation coefficient of the zero delay channel (after normalization), and
 $R(m\Delta t)$ is the normalized autocorrelation for delay $m\Delta t$.

G.7 Notation

In this chapter we have chosen to use the conventional symbols for most quantities. This results in a conflict with use of the same symbols elsewhere in this document (e.g., ρ is used here to signify correlation coefficient, and in Chapter 4 to signify reflector reflectivity). Rather than inventing a new alphabet, the table below lists the symbols used in this chapter as an aid to comprehension.

Notation used in this chapter

Symbol	units	1st. occurrence	description
Δt	–	G.14	time between DACS data samples (40 ns)
$\phi(v_0)$	–	G.5	probability integral of v_0
ρ	–	G.2	correlation coefficient
ρ_2	–	G.2	2-bit correlation coefficient
E	–	G.4	$\exp(-\frac{1}{2}v_0^2)$
j	–	G.14	frequency index (1..128)
m	–	G.14	delay channel index (1..128)
M	–	G.14	number of delay channels (128)
n	–	Table G.2	weighting given to highest order multiplier products
N	Hz	G.1	Sampling rate (25 MHz)
N_2	Hz	G.1	Rate of counts accumulation for 2 bit DACS
N_{lm}	Hz	G.1	Rate of counts accumulation for state lm
P_{lm}	–	G.1	Probability of undelayed and delayed signals being in states l and m respectively
R_c	–	G.10	Signal to noise of a continuous correlator
R_2	–	G.9	Signal to noise of the 2-bit correlator
R_{2a}	–	G.12	Signal to noise of the simplified 2-bit correlator
v_0	–	G.3	Normalized (<i>wrt</i> rms) non-zero digitizer thresholds
v_1, v_2	–	G.3	Normalized (<i>wrt</i> rms) signal voltages
V	V	Table G.1	Signal voltage or digitizer threshold
V_0	V	Table G.1	Modulus of non-zero digitizer thresholds
V_1, V_2	V	Table G.2	Digitized undelayed and delayed signals
$\bar{2}, \bar{1}, 1, 2$	–	Table G.1	The four possible digitized signal states

Appendix H

The EOS MLS DACS implementation

In the previous Appendix several variants of the 2 bit DACS were discussed in detail, ending with the implementation used on EOS MLS. Below we restrict the discussions to just the EOS implementation, and include representative radiometer performance estimates in some calculations where this helps to provide insight.

H.1 The digitizer as a power monitor

The output of the digitizer during each data integrations are monitored by 4 state counters (Figure G.1) which record the total number of occurrences of the data states $\bar{2}$, $\bar{1}$, 1 and 2. This is illustrated in Figure H.1 which shows an input signal with Gaussian characteristics in the time domain being digitized via a 2 bit A/D converter with transition thresholds at 0 and $\pm 0.9\sigma$ of the rms of the input signal. In this representation of the system the A/D converter outputs a pulse every clock cycle on one of 4 output lines which indicate the state ($\bar{2}$, $\bar{1}$, 1 or 2) of the input signal that was just digitized. The 2 bit digitizer thresholds of the EOS MLS DACS are adjusted to $\pm 0.9\sigma$ of the rms Space view signals to provide optimum signal to noise for the measured autocorrelations. In practice the implementation of the digitizer and counter electronics is substantially different from that shown in the figure, but the functionality is identical.

In addition to providing a mechanism for controlling the A/D converter thresholds, the state counters provide an indication of the signal power (relative to that during Space view) during each integration. This may be thought of in 2 ways: a power detector diode followed by an integrator in which the A/D converter acts as a crude detector, and the following counters serve as the integrator; or a statistics monitor in which the ratio of occurrences of the inner to outer states represents the signal power.

Both viewpoints must lead to the same results, so we choose the statistics monitor viewpoints to analyze further since it follows more closely the description in the previous chapter.

The contents of the 4 counters are used by the C&DH to adjust the threshold levels of the A/D converter so that while viewing the Space port reference we have:

$$\frac{(N_{\bar{1}} + N_1)}{(N_{\bar{2}} + N_{\bar{1}} + N_1 + N_2)} = 0.63189, \quad \text{and}$$

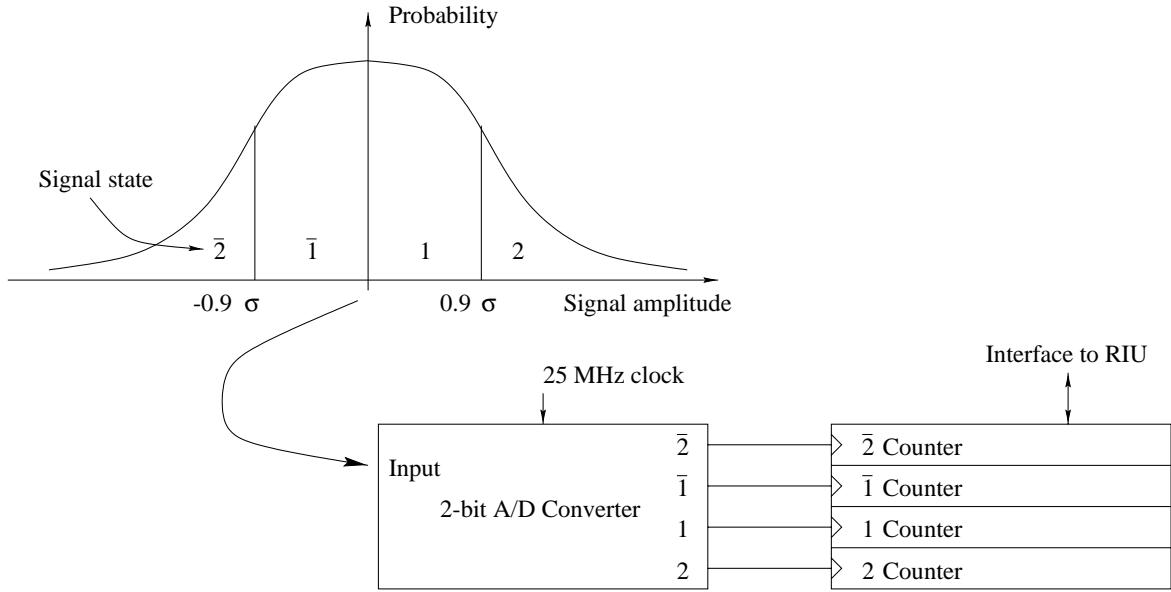


Figure H.1: This figure shows a Gaussian signal being digitized by a 2 bit A/D converter, with a set of counters recording the total number of occurrences of each digitizer output state during a complete data integration cycle.

$$N_{\bar{2}} + N_{\bar{1}} = N_1 + N_2 \quad (\text{H.1})$$

where N_m are the number of occurrences of state m during the data integration. This sets the digitizer thresholds at 0 and $\pm 0.9\sigma$.

H.2 Signal power estimation

When viewing a source of greater radiance than the Space reference the probability distribution shown in Figure H.1 is flattened (and broadened), causing the threshold levels of the digitizer be shifted. For example, if the radiometer has a T_{sys} of 1000 K, and the ambient calibration target is at a temperature of 300 K, when viewing the target the digitizer thresholds are moved to $0.9/1.3 = 0.69\sigma$ of the input signal. This is illustrated in a slightly different way in the top panel of Figure H.2 which plots the fraction of counts in states $\bar{1}$ and 1 (compared to the total number of samples in the integration) as a function of signal brightness temperature when the digitizer thresholds are maintained at the Space view settings of 0 and $\pm 0.9\sigma$. As can be seen from the figure, the relationship between signal radiance and fraction of counts in the inner 2 states is fairly linear. A quadratic will be fit to this relationship and used to convert the measured fraction into an equivalent brightness temperature in Level 1 processing. It is important to determine the signal to noise of the signal power estimate given from the quantities N_m , and compare it to the quality of measurement given by a dedicated analog signal power measurement channel at the input to the digitizer. We make this determination as follows. The variance on the quantity $N_{\bar{1}} + N_1$ is just the square root of the total number of number of counts in those states, and is readily obtained from the relationship $N_{\bar{2}} + N_{\bar{1}} + N_1 + N_2 = N\tau$ (where N is the sample rate of the DACS, nominally 25 MHz, and τ is the signal integration time, nominally 1/6 s) and the measured fraction of

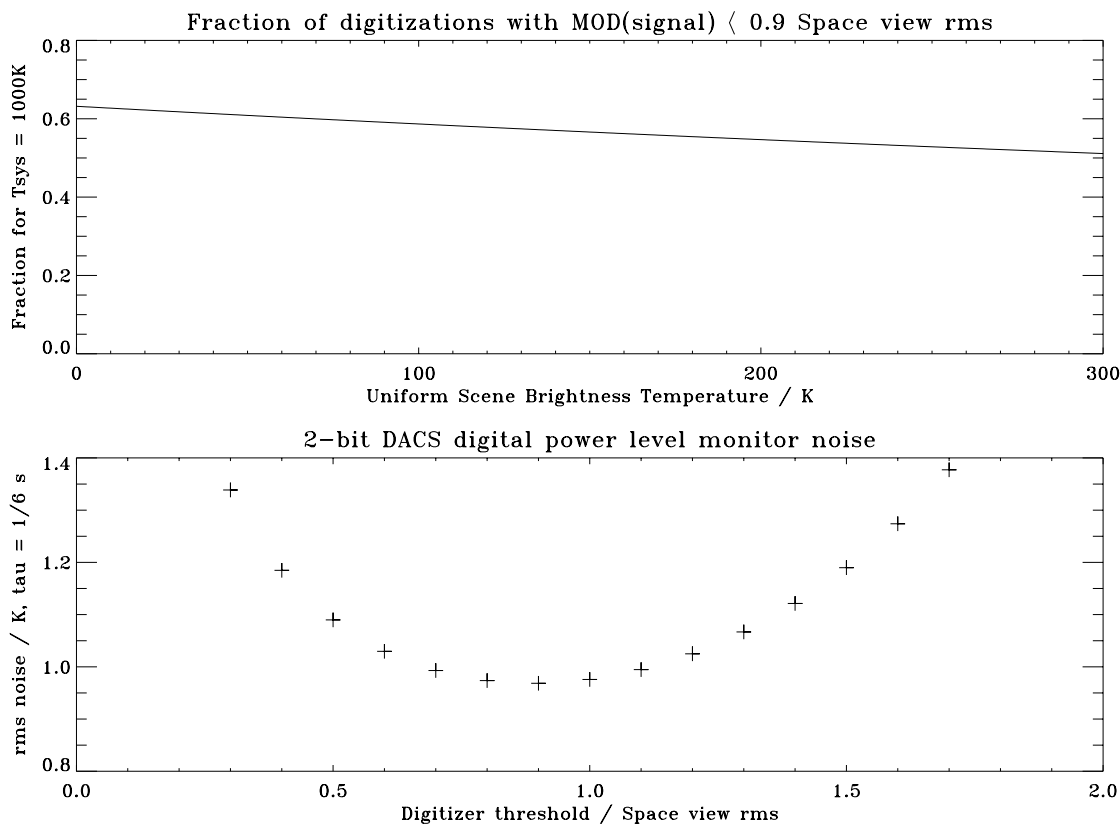


Figure H.2: This figure shows a Gaussian signal being digitized by a 2 bit A/D converter, with a set of counters recording the total number of occurrences of each digitizer output state during a complete data integration cycle.

counts in the inner 2 digitizer states (shown in the top plot of Figure H.2). The “gain” of the system is indicated by the slope of this figure, and the estimated noise on the signal power measurement is plotted in the lower panel of the figure for a range of digitizer threshold settings. Note that the optimum setting for the outer thresholds is $\pm 0.9\sigma$, the one which gives optimum signal to noise in the autocorrelation measurements for the reduced multiplier algorithm implemented in the EOS MLS DACS. The baseline plan is to maintain the digitizer thresholds at 0 and $\pm 0.9\sigma$ during Space reference views, but it has been suggested by H.M. Pickett that the thresholds be adjusted so that mid-range atmospheric radiances provide optimum signal to noise instead. Implementing this change involves just a minor change to the operating parameters of the C&DH digitizer threshold control software task, and no change to the DACS hardware, and will be investigated at a later time.

An analog power measurement channel would have rms noise at best of 0.77 K, given by the radiometer equation for $B = 10$ MHz and $\tau = 1/6$ s. In practice the analog filter would almost certainly have a noise bandwidth less than the ideal of 10 MHz, leading to a measured noise much closer to that given just from analysis of the digitizer count statistics. Certain other observations arise with use of an independent analog channel to measure signal power, in particular:

1. The analog channel would have high gain, precision, DC amplifiers with noise (both spectrally flat and with $\frac{1}{f}$ characteristics) and drifts independent of the behavior of the digital signal paths, and
2. the analog signal processing would involve an interpolations identical to those described in Chapter 4 to reduce the deleterious effects of noise and drifts on the reference measurements.

We thus conclude that there are probably no substantial benefits from including an analog signal power measurement channel, a fairly common practice with DACS used to analyze radio telescope signals, and possibly no real benefits at all in the EOS implementation. Note also that the filter spectrometers implement analog channels with 6 MHz nominal bandwidths at the centers of the DACS passbands which can serve as checks on the DACS performance and behavior. The difficulties with using these channels operationally are that their bandwidths are less than optimal (i.e., they provide noisy estimates of signal power) and, more important, their shapes do not match the bandpass shapes to the inputs of the DACS. This latter point is especially important since it means:

1. The shapes and positions of the 6 MHz channels need to be known better for this application than when they are used to provide their standard science data, and
2. the processing required to use signal power measured in these filter channels requires detailed knowledge of both the filter and DACS input passband shapes, and is somewhat cumbersome.

H.2.1 Signal power estimation details

Figure H.3 repeats the top panel of Figure H.2, modified to indicate signal power as a additional percentage of the signal presented to the DACS input when viewing the Space reference. A quadratic fit to these data provide the result that the signal power, t , input to the DACS when viewing a signal source is given by:

$$t = 301.555 - 760.07 \times f + 448.23 \times f^2 \quad (\text{H.2})$$

where:

- t is the percentage increase in input signal power compared to that measured when viewing the Space reference, and
- f is the fraction of digitizer counts in states N_1 and $N_{\bar{1}}$ combined.

For example, when the fraction of digitizer states N_1 and $N_{\bar{1}}$ is 0.55 of the total digitizations during a data integration, the total signal power input to the DACS is 19.106% higher than that measured during observation of the Space reference. If the Space reference signal, the sum of T_{sys} and T_{space} , was 1000 K, then the average signal in each DACS channel would have been 191.06 K. This total signal is then apportioned to each DACS channel (in the time frequency domain) according to the relative spectral signal, P , obtained from Equation G.14.

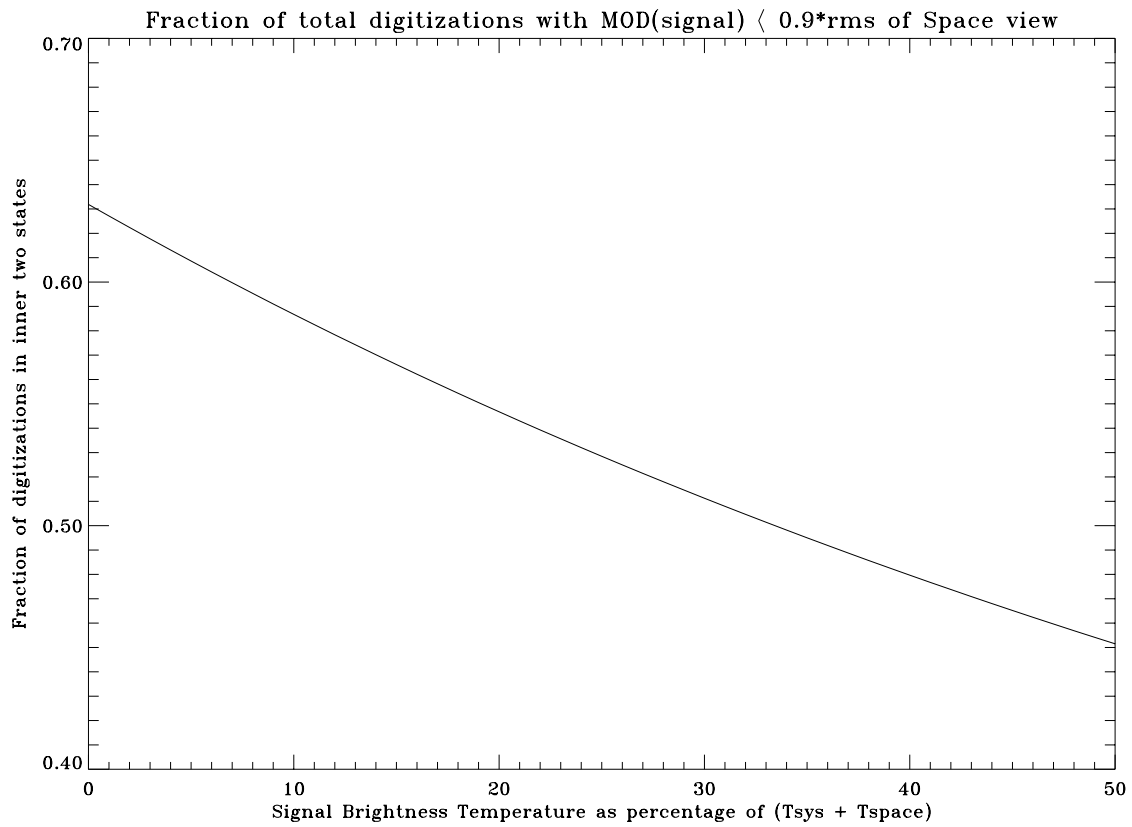


Figure H.3: Figure showing the relationship between total signal power and the fraction of 2-bit DACS signal digitizations in the inner two digitizer states. See text for additional details.

Appendix I

Notation

The first occurrence of each symbol in this document is indicated in the table below, together with a brief description. If a symbol has more than one use, the first occurrence with each usage is included. To minimize confusion where the same symbol may have more than two meanings in this document, separate Notation tables are provided at the ends of the Appendixes discussing Optimal Interpolation and the DACS. Appendix F which describes the noise characteristics of integrators and detectors, and Appendix B which discusses heterodyne radiometers, both use several symbols which have slightly different meanings than elsewhere in this document, and since both are largely self-standing and fairly brief, they are not covered by the following Notation table.

Symbol	Unit	Eqn.	Description
a		4.20	quadratic fit coefficient
a		5.4	conversion coefficient for PRD data processing
b		4.20	quadratic fit coefficient
b		5.4	conversion coefficient for PRD data processing
c	ms^{-1}	B.2	speed of light
c		4.20	quadratic fit coefficient
c		5.6	conversion coefficient for thermistor data processing
d		5.6	conversion coefficient for thermistor data processing
e		5.6	conversion coefficient for thermistor data processing
f		5.6	conversion coefficient for thermistor data processing
f_h	Hz	5.3	V/F converter output frequency during low calibration measurement
f_l	Hz	5.3	V/F converter output frequency during high calibration measurement
f_v	Hz	5.1	V/F converter output frequency
g_i	Counts K^{-1}	4.16	radiometric gain of channel i at the switching mirror
$\hat{g}_i(L)$	Counts K^{-1}	4.18	interpolated gain of channel i at time of limb view
h	Js	4.1	Planck's constant
i,s		4.5	channel i , sideband s (s is one of l or u)
j		4.20	MIF (time) index for quadratic fit
k	JK^{-1}	4.1	Boltzmann's constant
r		4.5	identifies one of seven radiometers

r_l, r_u		4.4	lower and upper sideband responses of entire signal chain (including antenna)
r'_l, r'_u		4.8	lower and upper sideband responses of signal chain from radiometers (mixers) to switching mirror
s	Counts	4.31	sample variance
w_j		4.25	weighting applied to measurement j in least squares fit
x	V/ Ω	5.3	calibrated multiplexed analog value
B_i	Hz	4.19	Predetection bandwidth of filter channel i
C_h		5.3	multiplexed analog 'high' calibration value
C_l		5.3	multiplexed analog 'low' calibration value
C_i^L	Counts	4.16	digitizer output viewing atmospheric limb
C_i^N	Counts	4.16	digitizer output random noise
C_i^O	Counts	4.16	digitizer output offset
$\hat{C}_i^S(T)$	Counts	4.18	interpolated space reference signal at time of target view for channel i
C_i^S	Counts	4.16	digitizer output when viewing space
C_i^T	Counts	4.16	digitizer output viewing target
C_h	Hz	5.3	multiplexed analog 'high' calibration input
C_l	Hz	5.3	multiplexed analog 'low' calibration input
$F_i(\nu)$		4.3	radiometer frequency response function
$G^M(\nu, \theta, \phi)$		4.3	gain function at switching mirror
$G_r^A(\theta, \phi)$		4.7	antenna gain function for radiometer r
$H(f)$		4.37	Frequency (power) response of post-detector radiometric calibration process
\dot{I}_i^L	K	4.12	Level 2 estimate of limb radiance reaching switching mirror
$\dot{I}_\nu(\theta, \phi)$	WSr ⁻¹ Hz ⁻¹	4.3	spectral radiance per unit frequency per unit solid angle
N		4.32	number of data points
N_t		5.1	Number of timebase cycles counted in an engineering data integration
N_v		5.1	Number of V/F converter cycles counted in an engineering data integration
\dot{P}	K	4.1	measured radiant power per unit bandwidth
\dot{P}^A	K	4.8	Limb radiance
$\langle \dot{P}_i^A \rangle$	K	4.13	Level 1 estimate of atmospheric limb radiance
\dot{P}_i^A	K	4.12	Forward Model estimate of limb radiance collected by the antenna
\dot{P}^L	K	4.8	radiance at switching mirror limb port
R_{in}	Ω	5.5	measured resistance of parallel thermistor/resistor pair
R_{th}	Ω	5.5	inferred thermistor resistance
R_{tlim}	Ω	5.4	measured PRD resistance
$S_g(f)$		4.37	non-spectrally flat component of post-detector noise power spectral density
T	K	4.1	temperature
T	C	5.4	inferred temperature (Celsius) of PRD

Δ		4.24	coefficient determinant in polynomial least squares fit
ϵ_r		4.17	calibration target emissivity for center frequency of radiometer r
$\eta_{i,s}^{MX}$		4.5	integrated gain over baffle aperture X
$\eta_{i,s}^A$		4.8	antenna transmission (scattering)
$\eta_{i,s}^k$		4.15	optical transmission of reflector k
λ_r	MIFs	4.25	$\frac{1}{e}$ rolloff distance for relative weightings in least squares fit
ν	Hz	4.1	radiation frequency
ν_{LO}	Hz	4.4	Local Oscillator frequency
ρ_r^A		4.8	antenna ohmic loss (cumulative)
$\rho_r^{1,2,3}$		4.15	ohmic loss for reflector 1,2,3
$\sigma(j)$	Counts	4.26	rms of measurement j
$\sigma'(j)$	Counts	4.26	weighted rms of measurement j
τ	s	4.19	data post-detector integration time
Ω_A		4.7	angle over which antenna pattern is defined
Ω_{MX}		4.4	solid angle defined by baffles
$d\Omega$		4.4	element of solid angle [$d\Omega = \sin(\theta) d\theta d\phi$]

Appendix J

Glossary

A/D	analog-to-digital
ADC	analog-to-digital converter
ASIC	Application Specific Integrated Circuit
C&DH	Control and Data Handling Assembly
CCSDS	Consultative Committee for Space Data Systems
CW	Continuous wave
D/A	digital-to-analog
DACS	Digital AutoCorrelator Spectrometer
DFT	Discrete Fourier Transform
EDH	Engineering Data Hybrid
EM	Engineering Model
EOS	Earth Observing System
FFT	Fast Fourier Transform
FM	Flight Model
FTI	Fourier Transform Interferometer
GB	Gigabyte, $\sim 10^9$ bytes
GSE	Ground Support Equipment
HDF	Hierarchical Data Format
HEMT	High Electron Mobility Transistor
HIRDLS	High-Resolution Dynamics Limb Sounder
IRU	Inertial Reference Unit
MAF	Major Frame
MB	Megabyte, $\sim 10^6$ bytes
MFLOPS	Millions of Floating Point Instructions per Second
MIF	Minor Frame
MIPS	Millions of Instructions per Second
MLS	Microwave Limb Sounder
msb	most significant bit
PRD	Platinum Resistance Device (temperature transducer)
RAID	Redundant Array of Inexpensive Disks
RIU	Remote Interface Unit
SCF	Science Computing Facility
SIDS	Simulated Instrument Data Set

SLS	Submillimeter Limb Sounder
SMP	Symmetric Multi-Processor, a form of parallel computer
S/N	Signal-to-noise
TBR	to be revised
TES	Tropospheric Emission Spectrometer
UARS	Upper Atmosphere Research Satellite
UTC	Coordinated Universal Time (GMT)
V/F	voltage-to-frequency
VFC	V/F converter

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