Design and performance evaluation of a modular imaging spectrometer instrument

Xiaofan Feng

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DESIGN AND PERFORMANCE EVALUATION OF A MODULAR IMAGING SPECTROMETER INSTRUMENT

by

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Doctor of Philosophy at the Center for Imaging Science

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December 1995

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December 12, 1995
DESIGN AND PERFORMANCE EVALUATION OF A MODULAR IMAGING SPECTROMETER INSTRUMENT

by Xiaofan Feng

Submitted to the Center for Imaging Science
on December 12, 1995, in partial fulfillment of the requirements for the degree of the Doctor of Philosophy

This thesis involved the design and testing of a modular imaging spectrometer instrument (MISI), which can provide hyperspectral image data at a very high ground resolution. The instrument can serve as an airborne laboratory for many remote sensing applications. The optical/mechanical/electrical system was designed using a system engineering approach with emphasis on system and sub-system modulation transfer function (MTF) analysis. Extensive modeling was used to predict the system performance and to aid the design trade process. Many sensor testing methodologies were developed to evaluate the performance of this electro-optical imaging system. The system engineering approach, the modeling tools developed, and the sensor performance testing methods can also be applied to other electro-optical (EO) imaging systems design and testing. Performance evaluation experiments verify that MISI has met its image quality goals. By using finite element analysis and optical image formation theory, the performance of a high-speed scan mirror was modeled. This model was used as a design tool to aid the development of the scan mirror assembly. A model-based algorithm was developed to derive the MTF of the imaging system from its edge spread function. By using prior information about the system, the MTF can be derived from very noisy edge traces. Finally, an alternative EO imaging system design criterion -- the information criterion, is investigated as a overall image quality measure. The results show that line-scan imaging system can be optimized informationally by shaping the electronic filter response and detector aperture. The analytical results also show that the traditional line-scan imaging system design does not maximize information, and the informationally optimized design with a diamond-shaped aperture can provide up to 1.5 bits more information than the traditional design over a broad range of radiance fields. Computer simulation confirms that by combining image acquisition and image processing, informationally optimized design tends to maximize image fidelity.
Acknowledgments

The author wishes to thank my team members: John Schott and Timothy Gallagher for more than five years of cooperation and hard working. Together, we made it happen.

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Many thanks to my managers and colleagues at Xerox Corporation for their continuous support and understanding.

And lastly, I'd like to thank my wife, my children, my parents and parents-in-law for their sacrifice in the last five years. Without their continuous support, this thesis would never be possible.
Dedications

To my teachers

A teacher affects eternity; he can never tell where his influence stops.

Henry B. Adams (1838–1918)
This is the first image collected by the modular imaging spectrometer instrument (MISI) on December 1, 1995. The instrument was mounted on a mechanical turn-table. The fast-scan motion was provided by the scan-mirror, and the slow scan motion (horizontal direction) was generated by rotating the turn-table with hands. Data were capture by a 12-bit A/D system and were later converted to 8-bit TIFF file.
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1. Introduction

Electro-optical (EO) imaging systems convert electromagnetic radiation at optical wavelengths to electrical signals for information extraction, source detection, and analog visual display. They improve the sensitivity of the human visual system and extend our vision to the infrared and ultraviolet spectrum. In the past thirty years, we have observed tremendous progresses in electro-optical imaging system development and application. Electro-optical imaging systems have been widely used in space exploration, earth remote sensing, military infrared imaging, machine vision, astronomy, document processing, and medical imaging.

![Figure 1.1](image)

**Figure 1.1** A functional flow diagram for a general electro-optical system

Electro-optical imaging systems use optical radiation as a carrier of information about the object being sensed or observed and convert the EM energy into electrical signals for data processing. The basic building blocks of a generalized EO imaging system are given in Figure 1.1. First there must be a source of optical radiation. After having been modified by passing through the intervening medium and selectively collected by the optical system, the optical energy is converted into electrical energy by a detector. The electrical signal is amplified and processed by the signal conditioning electronics. And finally, the output electrical signal is processed for information extraction or image display. The information to be sensed can be either in the source (e.g. the temperature or emissivity of an object in thermal imaging) or in the medium (e.g. the reflectance of an object or the transmittance of the medium to be sensed in the visible and near infrared imaging). Since optical radiation is used as an information carrier, EO imaging has at least five dimensions: intensity, wavelength, polarization, and spatial location \((x, y)\). The amplitude of the output electrical signal is proportional to the intensity of the optical radiation. Position information is coded by the location of the signals in time within the sequence. Spectral information can be extracted using imaging spectrometer instruments, and polarization is yet to be extensively exploited.

Due to the multi-disciplinary nature of electro-optical imaging systems, the design and performance evaluation of these systems requires not only a knowledge of several technical fields, including optics, detectors, noise theory, atmospheric effects, signal
processing and mechanical design, but also a system engineering approach to integrate the EO system. This thesis provides a system engineering approach to the design of an electro-optical line-scan imaging system. Central to the research is the application of system MTF and sub-system MTF analysis to the electro-optical-mechanical system. Computer modeling is extensively employed as a design trade tool in optical, mechanical, and electrical system design. It also provides the methodologies and procedures for establishing many sensor performance measures used in imaging sensors. The performance evaluation complements the image chain analysis by allowing us to evaluate the results of actual hardware testing on theoretical design criteria. Although this study is based on an EO remote sensing instrument, the system engineering approach, the modeling tools developed, and the testing methods can be applied to other EO systems as well.
2. **Background**

2.1 **History of Electro-Optical Imaging Systems**

The field of electro-optical engineering results from the merger of two related fields: the science of optics and electrical engineering. Whereas the history of optics goes back about 5,000 years, the history of electricity and electronics is only 200 years old.

2.1.1 **1870-1950: The Birth of Television**

The research on the first major electro-optical imaging system began in the late 1870s. Two events of that decade gave impetus to the search for the way of transmitting pictures and views or what was called “seeing by electricity”: the discovery of the light sensitivity of selenium in 1873, and the invention of Bell's telephone in 1876 (Abramson 1977).

By the 1880s, several “seeing-by-electricity” schemes were revealed (Shiers 1977). A mosaic of small separate squares of selenium was designed as the sensor, and magnetically operated aperture array was designed as the receiver. One of the most famous schemes for television was patented by Paul Nipkow on January 6, 1884 in Germany. This is the master television patent, for his solution was exceedingly simple and fundamentally sound - a spinning perforated disk. The scanning disk had 24 holes in a spiral near the outer rim. With the disk rotating, each hole reveals the field of view in consecutive strips; or line by line, as in reading the text. The transmitter consists of fixed lenses, a scanning disk, and a selenium cell. As the disk rotates, each hole scans the image sequentially. The selenium cell converts light intensity to electricity. At the receiver end, light from a local source, polarized in one plane by the first prism, is blocked by the second prism. Current through a coil is proportional to the conductivity of the selenium. A high level of light at the transmitter therefore increases the strength of the magnetic field and rotates the plane of polarization accordingly. The light passing through the analyzer then has an intensity proportional to the original picture element.

Nipkow’s scheme is simple yet elegant. Mechanical problems related to rapid scanning and synchronism were neatly solved by a spinning disk. This technology dominated the early years of television research but was ultimately abandoned as impractical. The basic elements of an all-electronic television system became available only after 1927, when the American engineer Philo T. Farnsworth developed his dissector tube.
The discovery in 1887 of photoelectric emission by H. Hertz led to the creation of inertialess photoelectric devices which played an important role in the development of television.

The discovery of radio wave by Hertz in 1888, the invention of radio communication by a Italian electrical engineer and inventor Guglielmo Marconi in the late 1890s[1][Zworykin et al. 1995], the discovery of the electron by Thomson in 1897, the invention of the cathode tube by K. F. Brown in 1897, and the electron tube by J. Fleming and Lee De Forest in 1904-1906. All of these laid the foundation for the subsequent development of electronics and radio engineering - optical electronic television.

In 1933, Zworykin of Radio Corporation of America (RCA), announced his iconoscope or television “electric eye”, thereby freeing the system from mechanical imaging limitations. In essence the iconoscope was a cathode-ray tube which could scan an optical image of the scene being televised and produce a video signal. The tube was able to store the electrical charge corresponding to any given scene elements throughout the scanning period, thereby accumulating a much higher signal than otherwise possible. This greatly enhanced sensitivity represented a significant advance. With the new iconoscope camera, RCA tested a virtually all-electronic system and the radio-relaying of television in the field during the first several months of 1933.

On July 1, 1941, fully commercial television was authorized in the United States. Soon after world war II, television was fully developed. It took almost a century to develop the television — the first electro-optical imaging system. But this imaging system has changed our twentieth-century world as much as the invention of gunpowder, the airplane, the telephone or the automobile. The technology and knowledge acquired in the development of television were readily applied to the other electro-optical imaging systems like thermal imagers and remote sensors. Television cameras placed aboard U.S. spacecraft successfully transmitted to earth formerly inaccessible information about outer space. The Mariner series of spacecraft, from 1965 to 1972, returned thousands of photographs of Mars. Spacecraft of the Ranger and Surveyor series relayed thousands of close-up pictures of the lunar surface for scientific analysis and preparation for lunar landings, before the successful U.S. landing on the moon in July 1969, during which live

---

1 For his pioneer work in the field of wireless telegraphy, Marconi shared the 1909 Nobel Prize in physics with the German physicist Karl Ferdinand Braun.
color broadcasts were made from the surface of the moon. Since 1960, television cameras have also been used extensively on orbiting weather satellites. Vidicon cameras trained on the earth record pictures of cloud cover and weather patterns during the day, and infrared cameras record nighttime pictures [Zworykin and Feldman, 1995].

2.1.2 1950-1970: Infrared Imaging -- Conquest of Darkness

According to the Planck's radiation law, the spectral radiation from a blackbody is given by

\[
M(\lambda) = \frac{C_1}{\lambda^5 \exp\left(\frac{C_2}{\lambda T}\right)} \quad (W \text{ m}^{-2} \mu\text{m}^{-1})
\]

where \( \lambda \) is the wavelength (\( \mu \text{m} \)), \( C_1 = 3.7418 \times 10^{12} \text{ W.cm}^2 \) is the first Planck constant, \( C_2 = 1.4387 \text{ cm K} \) is the second Planck constant, \( T \) is the absolute temperature in Kelvin. According to this formula, solid bodies at a temperature above absolute zero radiate energy, which is primarily in the infrared portion of the spectrum. Thus with the EO infrared system, we can still "see" the emitted IR radiation from any object even in the darkness.

Infrared radiation was first discovered by the astronomer William Herschel in the 17th century. Although the first infrared imaging system was the Evaporagraph (Lloyd 1975), which was developed in the 1930s, it could not perform most thermal imaging tasks because of inherent limitation in contrast, sensitivity, and response time. The first practical thermal imager was built by the Army in 1952. It was a single element, two-dimensional, slow framing scanner which recorded the imagery on photographic film and consequently was not a real-time device.

The development of fast-framing thermal imagers was not feasible until the late 1950’s. Two significant developments in photo-detectors were the indium antimonide (InSb) cell by Pieke in 1955, and the mercury cadmium telluride (HgCdTe) detector by Lawson in 1959. Both of these enhanced the application of EO sensors in the infrared. In 1960, Perkin-Elmer Corporation built a ground-based forward-looking infrared (FLIR) imager (Lloyd 1975). It used two rotating refractive prisms to generate a spiral scan for its single element InSb detector. It had a 5-degree circular field of view, a detector of angular resolution of 1 milliradian, a frame rate of 0.2 frame per second, a thermal sensitivity of about 1°C, and a long-persistence phosphor CRT display.
The airborne FLIRs were flight tested in 1965, and they were so successful that they 
spawned an amazing proliferation of airborne FLIRs and applications. From that point 
on, the FLIR business burgeoned, and between 1960 and 1974 at least sixty different 
FLIRs were developed and several hundred were produced (Lloyd 1975). FLIR 
technology reached maturity in the late 1970s with the development of HgCdTe detector 
arrays, which perform at near the theoretical thermal sensitivity limit while providing 
signal responses fast enough to sustain the wider bandwidth necessary for the high ratios 
of field of view to resolution (pixels per frame) requirements. FLIR systems have been 
extensively used in modern warfare, and more recently, there have been considerable 
civilian applications include law enforcement, fire fighting, and border patrol (Holst and 


Remote sensing is the process of acquiring information from the environment by the use 
of a sensor that is not in physical contact with the object under study. Man has been a 
remote-sensing creature since his very beginning. The ability of his eyes, ears, and nose 
to sense conditions in his surrounding environment often meant the difference between 
life and death. But it was very recently that EO sensors were implemented in aircraft and 
spacecraft. These sensors have produced good quality real-time pictures from other 
continents and other planets. Thus the new field of EO systems and satellite 
communication has made us aware of what exists beyond the range of human vision and 
in other spectral regions.

The electromagnetic radiation is a very important information carrier. The study of how 
light interacts with matter led to the theoretical development of the quantum theory in the 
twentieth century, where light acted as a probe to query the atomic world. For the same 
reason, optical radiation can be used to probe the earth environment. From a radiation 
point of view, remote sensing is the study of how electromagnetic radiation interacts 
with the terrestrial object to be sensed. For a sensor in the sky, the radiant energy 
collected by the sensor is given by (Schott 1989b)

\[ L(\lambda) = \frac{1}{\pi} E(\lambda) \tau_1(\lambda) r(\lambda) \tau_2(\lambda) + L_d(\lambda) r(\lambda) \tau_2(\lambda) + e(\lambda) \tau_2(\lambda) L_b(\lambda) + L_u(\lambda) \] (2.2)

where \( L \) is the spectral radiance seen by a detector above the atmosphere; \( E \) is the spectral 
irradiance of the sunlight as it strikes the top of the atmosphere; \( \tau_1 \) and \( \tau_2 \) are the 
transmission coefficients of the atmosphere along the Sun-Earth and Earth-sensor paths; \( r \)
is the spectral reflectance of the ground cover; $L_d$ is the “down-welled” spectral radiance, a combination of the sunlight scattered downward by the atmosphere and the downward thermal emission from the atmosphere; $L_u$ is the “up-welled” spectral radiance, a combination of the sunlight scattered upward by the atmosphere and the upward thermal emission from the atmosphere; $L_b$ is the spectral radiance from an ideal blackbody at the same temperature as the temperature of ground cover; and $\varepsilon$ is the emissivity of the ground cover. By measuring the radiance reaching the sensor, the material properties (reflectance, emissivity, and temperature) of ground cover can be inferred with proper radiometric calibration methods and digital image processing techniques (Schott 1989a). Since all terms in Eq. 3-2 are functions of wavelength, multispectral imaging is often used to extract more information. Figure 3.2 shows the spectral windows used in remote sensing.

![Figure 3.2 Spectral Windows Used in Remote Sensing of the Earth [Schott, 1996]](image)

The EO imaging sensors used in earth remote sensing share many hardware features with the EO sensors used in infrared imaging. All employ telescope type of optical systems to collect electromagnetic energy, scanners to provide two-dimensional images, and solid state detectors to convert optical radiation into electrical signals. The main differences are that the infrared imagers are fast-framing real-time imagers with only one spectral band, while most remote sensors are strip mappers with multispectral bands for better extraction of spectral features. Also digital image processing techniques are almost always used to process the remote sensing data (Schowengerdt 1983).

The placing of remote sensors in space by NASA for the purpose of making earth observations began in the early 1960s, and came about as an off-shoot of the decision to
land men on the moon. The testing of lunar-orbiting remote sensors on-board an aircraft over terrestrial sites and interpretations opened an opportunity for collecting data for the study of the earth. In 1960, NASA launched the first systematic space-based observation platform, the Television Infrared Observation Satellite (TIROS-I). TIROS-I was the first meteorological satellite and carried a low resolution imaging system. Since the launch of TIROS-I, NASA has launched more than 40 meteorological and environmental satellites with steadily improving sensor data-collection capabilities (Allison and Schnapf 1983). These satellites have provided uninterrupted images of the earth and its atmosphere in the visible and infrared spectrum.

The first satellite designed specifically to collect data of the earth’s surface and resources was the Earth Resources Technology Satellite (ERTS-1 later named Landsat 1) launched in 1972 (Simonett 1983). It carried a four-channel Multispectral Scanner (MSS), a three-camera return-beam vidicon (RBV), a data collection system, and two videotape recorders. The MSS operated in the following spectral intervals: band 4 (0.5 - 0.6 μm), band 5 (0.6 - 0.7 μm), band 6 (0.7 - 0.8 μm), and band 7 (0.8 - 1.1 μm). The three independent cameras of the RBV covered three spectral bands: blue-green (0.47-0.577 μ), yellow-red (0.58-0.68 μ), and near-infrared (0.69-0.83 μ). Both of these systems viewed a ground area of approximately 185 km square with a resolution of 80 meters. Landsat 2 was launched in 1975, and Landsat 3 was launched in 1978 with a fifth band added in the thermal infrared (10.4-12.6 μ).

The limitations of the Landsat series are its modest spatial resolution, spectral channels not precisely aligned to known absorption bands, and the inordinate time delay before the data become available to the user (Simonett 1983). Landsat-4 was the second generation of the Landsat mission launched in 1982. Landsat-4 also employ the same four bands of the Landsat 3 MSS but added an advanced MSS called the Thematic Mapper (TM). The TM sensor is designed to achieve better ground resolution (30 m), more separation between spectral bands, and better calibration and radiometric accuracy (SBRC 1984). The spectral band, detector, radiometric accuracy, and application of the TM sensor are shown in Table 2-1 (Chen 1985; Engel 1983).

The basic elements of the TM instrument are shown in Figure 3.2 (SBRC 1984). Scene energy from earth is first reflected by the scan mirror, and focused by the telescope onto the detectors of Bands 1, 2, 3, and 4 at the prime focal plane and onto the detectors of Bands 5, 6, and 7 at the cold focal plane via the relay optics.
Table 2-1 Landsat Thematic Mapper Spectral Band Selection and Applications

<table>
<thead>
<tr>
<th>Band</th>
<th>Spectral range (μm)</th>
<th>Detectors</th>
<th>Radiometric Resolution</th>
<th>Principal applications</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.45-0.52</td>
<td>SiPD</td>
<td>0.8% NEΔρ</td>
<td>Coastal water mapping; Soil/vegetation differentiation; Deciduous/coniferous differentiation</td>
</tr>
<tr>
<td>2</td>
<td>0.52-0.60</td>
<td>SiPD</td>
<td>0.5% NEΔρ</td>
<td>Green reflectance by health vegetation</td>
</tr>
<tr>
<td>3</td>
<td>0.63-0.69</td>
<td>SiPD</td>
<td>0.5% NEΔρ</td>
<td>Chlorophyll absorption for plant species differentiation</td>
</tr>
<tr>
<td>4</td>
<td>0.76-0.90</td>
<td>SiPD</td>
<td>0.5% NEΔρ</td>
<td>Biomass surveys; Water body delineation</td>
</tr>
<tr>
<td>5</td>
<td>1.55-1.75</td>
<td>InSb</td>
<td>1.0% NEΔρ</td>
<td>Vegetation moisture measurement; Snow/cloud differentiation</td>
</tr>
<tr>
<td>6</td>
<td>10.4-12.5</td>
<td>HgCdTe</td>
<td>0.5K NEDT</td>
<td>Plant heat stress measurement; Other thermal mapping</td>
</tr>
<tr>
<td>7</td>
<td>2.08-2.35</td>
<td>InSb</td>
<td>2.4% NEΔρ</td>
<td>Hydrothermal mapping; Geology mapping</td>
</tr>
</tbody>
</table>

One fundamental difference between the TM and MSS is that the MSS scans in only one direction, while the TM scans and obtains data in two directions, which enables the TM to scan at a lower speed than the MSS. The scan-line corrector is used to compensate for the forward motion of the satellite. Between the scan-line corrector and the prime focal plane is an oscillating shutter mechanism called the calibrator. The calibrator is synchronized with the scan mirror in such a way that it brings the calibration sources sequentially in view of the detectors during each scan mirror turnaround. For bands 1 through 5 and 7, there are 16 detectors per band (cf. Figure 2.2), providing a 30-meter ground instantaneous field of view (GIFOV). For band 6 four detectors, provide a 120-meter GIFOV. A single active scan of the object-plane scan mirror sweeps a ground swath of 185 km by 480 meters. The analog signal of each individual detector is sampled at the rate of 1 sample per IFOV, and each sample is quantized into an 8-bit binary word giving an encoding resolution of 256. The data from all 100 channels are combined into an 84.9 Mbps (megabits per second) data stream which is transmitted to the ground station through NASA’s Tracking and Data Relay Satellite System (TDRSS).
Landsat 6 was designed to provide 30-meter spatial resolution multispectral image data continuity via an Enhanced Thematic Mapper (ETM) sensor which includes a Panchromatic (PAN) band with 15-meter spatial resolution (Mowle and Dennehy 1991).
Landsat 6 was launched in 1994 but unfortunately failed to reach orbit; another launch (Landsat 7) has been planned in 1997.

Over the past twenty years Landsat satellites have been the principal operational earth observing and resource monitoring platforms. They have provided the repetitive global coverage necessary to support both domestic and international image data applications. The image data generated by these satellites has greatly enhanced the understanding of such varied earth sciences as geology, agriculture, forestry, hydrology, oceanography and bathymetry. It has been proven that Landsat satellites serve a vital role in environmental monitoring, mapping, disaster assessment, oil/mineral exploration, land use and regional planning.

Besides the Landsat program of the US, the French space agency launched its own earth observing satellite, SPOT (Système Probatoire d'Observation de la Terre), in 1984. SPOT employs high-resolution visible (HRV) imaging systems which can provide panchromatic images of 10-meter ground resolution. Another innovation in SPOT is the off-nadir viewing capability which provides considerable flexibility in scheduling times of image acquisition and stereo viewing capability.

Figure 2.3 shows the schematic of the SPOT imaging system. The HRV is a push-broom scanner which employs one-dimensional arrays of charge-coupled detectors (CCD) on its focal plane. Successive lines of ground coverage are obtained as the satellite moves over the earth. The HRV system operates in either a multispectral mode or a high-resolution panchromatic mode. The 6000 CCDs of the high-resolution mode have a 10 x 10 m ground resolution cell and cover the spectral range of 0.51 - 0.73 μm. The multispectral mode of HRV employs three arrays of 3000-element CCDs with a ground resolution cell of 20 x 20 m. The three spectral bands are: green (0.50 - 0.59 μ), red (0.61 - 0.68 μ), and near-infrared (0.79 - 0.89 μ).
The pushbroom scanner employed in SPOT has the advantages of no moving parts, high geometric fidelity, and longer life expectancy than the mechanical line scanners of the Landsat series. But it has two distinct disadvantages: (1) Each detector in a linear array must be individually calibrated to produce a uniform response, and (2) good quality large detector arrays (more than 1000 elements) in the infrared (> 1 μ) are presently very expensive.

In the interim, electro-optical imaging systems on-board aircraft are being widely used for remote sensing studies. Airborne remote sensing provides more flexibility in changing sensor configuration, and higher spatial and radiometric resolution. The first generation of airborne electro-optical imaging systems were cross-track multispectral scanners shown in Figure 2.4 [Sabins 1986]. The scan mirror sweeps across the terrain in a pattern of parallel scan lines oriented perpendicular to the flight direction. Radiation from the ground is reflected by the scan mirror and then focused onto a detector or entrance aperture of a spectrometer. Table 2-2 lists the characteristics of the Daedalus multispectral scanner (Sabins 1986)
Figure 2.4 Cross-track Multispectral Scanner.

Table 2-2 Characteristics of the Daedalus aircraft multispectral scanner

<table>
<thead>
<tr>
<th>Aircraft altitude</th>
<th>19.5 km</th>
</tr>
</thead>
<tbody>
<tr>
<td>Scanner IFOV</td>
<td>1.25 mrad</td>
</tr>
<tr>
<td>Ground resolution cell</td>
<td>24 by 24 m</td>
</tr>
<tr>
<td>Scan angle</td>
<td>42°</td>
</tr>
<tr>
<td>Image-swath width</td>
<td>14.7 km</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Band</th>
<th>Wavelength, μm</th>
<th>Spectral region</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0.38 to 0.42</td>
<td>UV and blue</td>
</tr>
<tr>
<td>2</td>
<td>0.42 to 0.45</td>
<td>Blue</td>
</tr>
<tr>
<td>3</td>
<td>0.45 to 0.50</td>
<td>Blue</td>
</tr>
<tr>
<td>4</td>
<td>0.50 to 0.55</td>
<td>Green</td>
</tr>
<tr>
<td>5</td>
<td>0.55 to 0.60</td>
<td>Green</td>
</tr>
<tr>
<td>6</td>
<td>0.60 to 0.65</td>
<td>Red</td>
</tr>
<tr>
<td>7</td>
<td>0.65 to 0.70</td>
<td>Red</td>
</tr>
<tr>
<td>8</td>
<td>0.70 to 0.80</td>
<td>Reflected IR</td>
</tr>
<tr>
<td>9</td>
<td>0.80 to 0.90</td>
<td>Reflected IR</td>
</tr>
<tr>
<td>10</td>
<td>0.90 to 1.10</td>
<td>Reflected IR</td>
</tr>
</tbody>
</table>

Figure 2.5 shows the optic schematic of the Environmental Research Institute of Michigan (ERIM) experimental M-7 multispectral scanner. It employs a rotary scanner and Dall-Kirkham type of telescopes [Swin and Davis, 1978]. After the incoming radiation is reflected off the scan-mirror, a portion of it (the part which would be lost by the obscuration of the secondary mirror) is reflected off a folding mirror into a dichroic beam splitter. The dichroic is ground into a spherical mirror so that the reflected radiation is images by the fold into detector 1B (0.33-1.0 μm). The radiation pass
through the dichroic mirror is imaged through a refracting system into a MCT detector (2–14 μm). The portion of radiation that was not reflected out of the main beam enters the Dall-Kirkham telescope (12.3 cm). The converging beam from the telescope is split by the second dichroic mirror into a short wave IR channel (0.9–2.6 μm) and a visible and near IR spectrometer (0.4–0.9 μm). Twelve fiber optic bundles were used to select spectral radiation and directed the radiation to photomultiplier tubes. Table 2-3 summarizes the specifications of the scanner [Swin and Davis, 1978].

Recent upgrades have resulted in the M-7 Mapper, which is the M-7 mated with a state-of-the-art high bandwidth, six-degrees-of-freedom, motion measurement subsystem [Carmer, 1994]. This makes the M-7 Mapper uniquely able to accomplish precision geometric image correction, orthorectification, geocoding, and multipass mosaics. New spectral bands provide more options in the 0.35 to 12.5 micrometer regions. The number of new bands has been increased by a factor of two, that is, from the original 15 bands to 30 bands plus an optional narrow or broadband visible channel that will replace the ultraviolet band when desired; thus, the new total is 31 bands. These new spectral bands will allow for emulation of current civilian remote sensing satellites, such as SPOT and Landsat TM and assessment of various potential strategies for spectrally upgrading those sensors. Particular emphasis has been placed on adding bands that will split existing bands (e.g., splitting TM bands 5 and 7 each into two narrow bands) as well as adding bands not currently available (e.g., two shortwave infrared [SWIR] bands in the 1.0 to 1.3
micrometer region and a 4.6 to 5.3 micrometer band in the medium wavelength infrared (MWIR). Likewise, multiple bands in the long-wave infrared (LWIR) region will expand on the one band offered by TM band 6. Performance for these new bands is similar to that of current bands.

Table 2-3 Nominal M-7 scanner-performance characteristics

<table>
<thead>
<tr>
<th>Characteristic</th>
</tr>
</thead>
<tbody>
<tr>
<td>15 spectral bands in ultraviolet, visible, infrared regions</td>
</tr>
<tr>
<td>90° external FOV (±45° from nadir)</td>
</tr>
<tr>
<td>2 mrad maximum spatial resolution, 3 mrad nominal</td>
</tr>
<tr>
<td>0.1°C nominal thermal resolution</td>
</tr>
<tr>
<td>1 percent nominal reflectance resolution</td>
</tr>
<tr>
<td>5 radiation reference ports</td>
</tr>
<tr>
<td>12.25-cm-diameter collector optics</td>
</tr>
<tr>
<td>60 or 100 scans/s</td>
</tr>
<tr>
<td>Direct current to 90 kHz electronic bandwidth</td>
</tr>
<tr>
<td>Roll-stabilized imagery</td>
</tr>
</tbody>
</table>

The advantage of M-7 is that all channels are inherently registered, thus make image processing a lot easier. But detector 1A and 1B receive much less energy because of the smaller aperture size of the secondary mirror. The use of photomultiply tubes and fiber optic bundles limits the number of channels available for the spectrometer. Also the one percent nominal reflectance resolution is too low for many applications.

The multispectral sensors described above sample the electromagnetic spectrum over bands approximately 100 nm wide. Sensors with this kind of spectral resolution allow isolation of only broad spectral features, severely limiting the degree of possible material or land cover classification (Schott 1989b). Imaging spectrometers are a new generation of imaging sensors which can acquire 100 to 200 contiguous spectral bands simultaneously, which makes possible the acquisition of reflectance spectra with higher spectral resolution for each pixel sufficient to identify directly those materials which have diagnostic spectral features in their reflectance spectra (Vane and Goetz 1985) (Goetz et al. 1985).

The first imaging spectrometer instrument for remote sensing study is the airborne imaging spectrometer (AIS) developed at Jet Propulsion Laboratory (JPL). The instrument uses a pushbroom along-track scanner similar to SPOT. As the aircraft moves forward, the instrument collects data over a ground swath width of 64 pixels with 128 spectral bands in the spectral region of from 1.2 to 2.4 μm. The instrument layout is shown in Figure 2.6 (Wellman 1987). Optical radiation from the ground is focused on a 4.23 x 0.142 mm slit, which defines the line image of the instrument in object space as
well as the aperture of the grating spectrometer. The heart of the instrument is a 64 x 32 element HgCdTe hybrid focal plane array. To provide 128 spectral bands, the grating has to step four positions.

![AIS Optical Schematic](image)

Figure 2.6 AIS Optical Schematic

NASA also developed the airborne visible and infrared imaging spectrometer (AVIRIS) which provides extended spatial and spectral coverage. The Airborne Visible Infrared Imaging Spectrometer is an instrument that flies onboard a NASA ER-2 airplane (a U2 spy plane modified for increased performance). The instrument is a unique optical sensor that images in 224 contiguous spectral channels, or bands, in wavelengths from the infrared (400 nm) to the ultraviolet (2500 nm). Its science objectives are to study Earth's ecology, geology, snow hydrology and oceanography to achieve an understanding of the environment and global climate change.

AVIRIS uses a scanning mirror to sweep back and forth ("whisk broom" fashion), producing 614 pixels for the 224 detectors each scan as shown in Figure 2.7. Fiber optics bundles were used to transmit optical radiation from the primary focal plane to the four spectrometers. Each pixel produced by the instrument covers an approximately 20 meter square area on the ground (with some overlap between pixels), thus yielding a ground
swath about 11 kilometers wide.

Figure 2.7 AVIRIS Optics Schematic

Besides airborne imaging spectrometers, space-borne imaging spectrometers are also being developed. The moderate resolution imaging spectrometer (MODIS) (Salomonson et al. 1988) will be launched late in the 1990s or early in the next century as part of the Earth Observing System (EOS) programs.

2.2 The Modular Imaging Spectrometer Instrument (MISI)

2.2.1 Background

The Digital Imaging and Remote Sensing (DIRS) group at RIT’s Center for Imaging Science has operated an airborne electro-optical line-scanner capable of operating in the visible, mid-wave infrared or long-wave infrared for many years. The instrument is a cross-track scanner as shown in Figure 2.4. This system has been a key component of much of the research the laboratory has conducted for a variety of sponsors and for numerous applications. The system has gone through several upgrades over the years with the current performance specifications listed in Table 2-5. This system is one of only a few such systems operating in the country. Its performance specifications still enable its use in support of many programs. However, it became increasingly apparent in the late 1980’s that anticipated imaging requirements could not be met by continued
modification of the existing system. As a result, the DIRS lab began a more structured effort to evaluate expected requirements. This was motivated by new R&D thrusts that required better radiometric resolution than was currently available with the existing system and a need for simultaneous acquisition in multiple spectral bands. In addition, the DIRS lab and its traditional supporting organizations were showing increasing interest in radiometric fidelity and multispectral/hyperspectral tradeoffs. In general, it became clear that there is a need for the capability to collect data other than that is currently available. This was further motivated by the perception that the value of such a collection system or systems would be enhanced if it was used in the university environment. In part, this was due to the educational value of having students involved in overall system development and ongoing upgrade and operation. However, another strong motivation was the need for such an instrument in the academic community where such concerns as radiometric fidelity can be carefully monitored and where new concepts can be easily implemented and evaluated at reasonable costs. Given this initial motivation, the next step involved an effort to identify expected uses of an airborne collection laboratory for the turn of the century and beyond.

Table 2-5 Partial Listing of Original and Current System Specifications

<table>
<thead>
<tr>
<th></th>
<th>Original Bendix LN3 (≈1970)</th>
<th>Current System</th>
</tr>
</thead>
<tbody>
<tr>
<td>Optics</td>
<td>3”, F4 Cassegrain 120°</td>
<td>3”, F4 Cassegrain 120°</td>
</tr>
<tr>
<td>Field-of-view</td>
<td>120° 2.5 Milliradians</td>
<td>120° 1 milliradian</td>
</tr>
<tr>
<td>IFOV</td>
<td>Digital delay line</td>
<td>Digital delay line</td>
</tr>
<tr>
<td>Roll compensation</td>
<td></td>
<td>Recorded gyro signal and post-processing to remote roll</td>
</tr>
<tr>
<td>Image recording</td>
<td>film</td>
<td>Wide band FM with ADC on ground</td>
</tr>
<tr>
<td>Bandpass</td>
<td>Interchangeable detectors (3-5, 8-14, and PMT)</td>
<td>Interchangeable detectors</td>
</tr>
<tr>
<td>Signal passband</td>
<td>138 K Hz</td>
<td>346 K Hz</td>
</tr>
<tr>
<td>NEDT (8-14 mm)</td>
<td>≈0.25 K</td>
<td>≈0.3 K</td>
</tr>
<tr>
<td>Calibration</td>
<td>Blackbody &amp; fixed gain</td>
<td>Blackbody &amp; fixed gain</td>
</tr>
<tr>
<td>Effective GIFOV at 1000 ft</td>
<td>2.5 ft x 2.5 ft</td>
<td>1 ft x 1.3 ft</td>
</tr>
<tr>
<td>Interchangeable spectral filters in flight</td>
<td>yes (4)</td>
<td>yes (4)</td>
</tr>
<tr>
<td># pixels per line</td>
<td>838</td>
<td>2,096</td>
</tr>
</tbody>
</table>

2.2.2 Expected Uses

In a research and development environment, it is always difficult to forecast expected requirements. Therefore, we chose to pursue two approaches in parallel. The first was to
look at uses that emphasized development of generalized methods rather than narrow applications. The second approach was to design as much adaptability into the system as practical to allow for future changes. We also considered continuation or growth of areas which had been long-term uses of the existing system. These expected uses and brief explanations of the expected requirements are listed below.

**Assessment of Radiometric Calibration Techniques**

Development and evaluation of radiometric calibration techniques have been an important application of the existing system, particular in the LWIR region [Schott, 1993]. Future use for this application requires increased radiometric precision, high radiometric accuracy, and for some applications, multiple spectral bands in a window. In addition, there is an expected requirement to develop and evaluate radiometric calibration techniques in a broader range of spectral bands, including the visible, SWIR, and MWIR. Because of the need for careful ground truth and repeated testing under various atmospheric conditions, this type of testing is often difficult without regular access to an airborne imaging system.

**Thermal Infrared Surveys and Temperature Mapping**

Temperature mapping has also been a long-term use for the current system that has different requirements. These studies have included mapping of power-plant cooling water discharges and heat-loss surveys for energy conservation. These studies require absolute radiometric calibration, large area coverage in most cases, and ground instantaneous field of view (GIFOV) values as small as one foot for heat loss studies. We would expect that while the application may change in the future, there will be an ongoing need for fairly large-area mapping of apparent or absolute surface temperatures with an increasing need for greater temperature fidelity.

**Engineering Assessment of High Altitude and Space-Based Systems**

A major use of the existing system has been to under-fly satellite systems to evaluate post-launch performance and improve understanding the satellite data of the lower spatial and/or radiometric resolution [Schott, 1993]. With the new generation of sensors planned for the polar orbiters and some of the high-altitude prototypes, we expect an ongoing need for well-calibrated high-resolution sensors to under-fly and evaluate satellite systems. We expect research in this area to involve increasing numbers of spectral bands with greater spectral resolution. This area is also likely to involve research into spatial-
spectral resolution or bandwidth tradeoffs. A system designed to support research in this area would be capable of sensing in many bands and, ideally, have the capability of high spatial and spectral resolution.

Source of Data for Algorithm Development Studies

The ever increasing number of applications of remote sensing and the increased amount of spatial and spectral information, coupled with advanced computers and software, is generating a new wave of studies aimed at improved image analysis algorithms. These will require reasonably large area coverage of targets of interest with a broad range of spatial, spectral, angular, and radiometric requirements. While all requirements cannot be forecast and no system will meet them all, we see this need as reinforcing the requirement for an adaptable system which can be operated from a variety of locations. Also, because algorithm studies are often not heavily mission-driven and are, therefore, often poorly funded, the collection platform often needs to be in an environment where low-cost operation is possible.

Validation of Synthetic Scene Models

A growing thrust in the remote sensing community is the development of modeling tools for synthetic image generation and the development of exploitation algorithms based on aspects of these models. Much of this work is aimed at reducing the need for expensive data collections. However, there is a need for limited collection for model validation. Ideally, these collections would utilize highly calibrated modest-cost platforms.

Environmental Assessment Studies

Remote sensing for environmental assessment has traditionally been done with a combination of aerial photography and earth resources satellites (e.g. Landsat, SPOT, AVHRR, CZCS ...). Future work will undoubtedly continue to require a mixture of airborne and space-based systems. However, the complexity of environmental signatures and the more subtle problems that we will seek to address will require a wider range of spectral information than photography alone can handle. In addition, there appears to be a growing requirement for active imaging systems to study certain environmental conditions. These requirements suggest that multispectral sensors with fairly wide fields of view to facilitate survey work will be required and also points to the need to consider active illumination for both reflection and luminescence studies.
Signature Assessment Studies

In the remote sensing community, methods for “seeing” have traditionally taken two courses. The first involves scientists identifying or speculating that based on some physical characteristics an object or the condition of an object should be observable. This leads to a requirement to collect the appropriate data using existing systems or to develop a system to collect the appropriate data to test the proposed approach. The second course, which has been the more common one, is that when new sensors become operational one can analyze scenes of interest to “look for” physical manifestations of objects or conditions of interest. While this may seem backwards, our knowledge-base of the optical phenomena associated with most natural surfaces is often inadequate for the more straightforward approach. We expect to use the MISI instrument to pursue both courses for signature assessment. The detailed spatial, spectral, and angular data available from the baseline system will offer the opportunity to look for many new signatures. At the same time, MISI’s modular design should allow for relatively easy adaptation to look at the earth in ways defined by studies of the physical characteristics of objects of interest (i.e. at a particular spectral band, view angle, or luminescence region). The use for signature assessment thus pushes us to many spectral bands, a wide range of resolutions and view angles, and requires a highly adaptable system for future signature assessment.

Based on these expected uses, a set of initial requirements were identified.

2.2.3 Initial Requirements

Based on the expected uses described above, a set of initial requirements and priorities were assembled as indicated in Table 2-6. These were prioritized as “required” if there was no sense in building a replacement system if the requirement was not met, “highly desirable” if a clear demand for the capability had been defined, and “desirable” if expected uses had been identified. The design objective was to meet all the required goals, as many of the highly desirable goals as possible, and be adaptable enough to meet the desirable goals in the future.
<table>
<thead>
<tr>
<th>Required</th>
<th>Very Desirable</th>
<th>Desirable</th>
</tr>
</thead>
<tbody>
<tr>
<td>Replace existing LWIR system option</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>Maintain 0.3 m GIFOV</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>Improve system NEDT from 0.25 K to 0.1 K</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Increase dynamic range from ≈ 7 bits to 10 bits</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Add split band for calibration</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Add several bands for signature assessment</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Improve radiometric calibration to 0.25 K</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Maintain a minimum of 90° FOV for survey work &amp; large area collections</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Maintain a capability for a several km ground swath (6 km minimum)</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Improve system noise with lower resolution imaging at higher altitudes</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>Replace existing MWIR system option</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>Increase GIFOV from 0.75 to 0.3 m</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Achieve system NETD of 0.1 K</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Add split band for calibration</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Provide the same record &amp; geometry specs as in LWIR</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>Multispectral requirements</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Be able to simulate Landsat TM spectrally</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Achieve a minimum GIFOV of 0.3 m in some bands with 0.6 in all bands</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Achieve an NEAρ of 0.2% with GIFOV of 1 m or better</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>Hyperspectral requirements</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- 15 nm or better resolution from 0.4 - 1 mm</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- GIFOV ≈0.6 m with S/N &gt;150</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- S/N &gt;250 for GIFOV ≈3 m</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- 15 nm or better spectral resolution from 1 - 2.5 mm</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- GIFOV 1 m with S/N &gt;100</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- S/N &gt;200 for GIFOV ≈3 m</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>Miscellaneous requirements</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Capable of upgrade for laser induced fluorescence</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Capable of upgrade for lunar induced fluorescence</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Capable of upgrade for polarization signature studies</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Capable of use from range lab for signature &amp; feasibility studies</td>
<td>X</td>
<td>X</td>
</tr>
<tr>
<td>- Capable of upgrade for ultra spectral imaging from range lab</td>
<td>X</td>
<td>X</td>
</tr>
</tbody>
</table>

### 2.3 Comparison of MISI with Other Line Scanners

Table 2-7 lists some of the key specs of four line scanners including MISI. Among the other three, the Landsat TM is the most important space base remote sensor, and AVIRIS is the hottest imaging sensor today. The ERIM M-7 was chosen because it shares many
same design features with MISI. Also ERIM is a private, nonprofit, high-technology organization similar to us, where the primary goal of the instrument is to collect radiometric calibrated data for remote sensing researches. These three systems represent the state-of-art multispectral line scanners.

Table 2-7  Comparison of MISI with other line scanners

<table>
<thead>
<tr>
<th>Spectral Bands</th>
<th>Spectral Resolution (µm)</th>
<th>Spatial Resolution (Meter)</th>
<th>NEAR or NEDT</th>
<th>Field of View (degree)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Landsat TM</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>4 VNIR (16)</td>
<td>≈0.05</td>
<td>30</td>
<td>0.5%</td>
<td>12</td>
</tr>
<tr>
<td>2 SWIR (16)</td>
<td>0.2</td>
<td>30</td>
<td>1%</td>
<td>12</td>
</tr>
<tr>
<td>1 LWIR (4)</td>
<td>2</td>
<td>120</td>
<td>0.5K</td>
<td></td>
</tr>
<tr>
<td>AVIRIS</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>0.4-0.7</td>
<td>0.01</td>
<td>20</td>
<td>0.2%</td>
<td>30</td>
</tr>
<tr>
<td>0.65-1.25</td>
<td>0.01</td>
<td>20</td>
<td>0.25%</td>
<td>30</td>
</tr>
<tr>
<td>1.2-1.82</td>
<td>0.01</td>
<td>20</td>
<td>0.25%</td>
<td>30</td>
</tr>
<tr>
<td>1.78-2.40</td>
<td>0.01</td>
<td>20</td>
<td>0.5%</td>
<td>30</td>
</tr>
<tr>
<td>M-7²</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>0.33-1</td>
<td>0.7</td>
<td>5</td>
<td>1%</td>
<td>90</td>
</tr>
<tr>
<td>0.4-0.9 (124)</td>
<td>≈0.04</td>
<td>5</td>
<td>1%</td>
<td>90</td>
</tr>
<tr>
<td>0.9-2.6</td>
<td>1.7</td>
<td>5</td>
<td>1%</td>
<td>90</td>
</tr>
<tr>
<td>2.0-14</td>
<td>12</td>
<td>5</td>
<td>1% or 0.1K</td>
<td>90</td>
</tr>
<tr>
<td>MISI⁵</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>0.4-1.0 VNIR</td>
<td>0.6</td>
<td>0.25</td>
<td>0.1%</td>
<td>90</td>
</tr>
<tr>
<td>2 SWIR</td>
<td>0.2</td>
<td>0.5</td>
<td>0.1%</td>
<td>90</td>
</tr>
<tr>
<td>3-5 MWIR (2)</td>
<td>2</td>
<td>0.25</td>
<td>0.4K</td>
<td>90</td>
</tr>
<tr>
<td>MWIR (2)</td>
<td>1</td>
<td>0.5</td>
<td>0.3K</td>
<td>90</td>
</tr>
<tr>
<td>8-14 LW (2)</td>
<td>6</td>
<td>0.25</td>
<td>0.6K</td>
<td>90</td>
</tr>
<tr>
<td>LWIR (3)</td>
<td>2</td>
<td>0.5</td>
<td>0.1K</td>
<td>90</td>
</tr>
<tr>
<td>0.4-0.7 (38)</td>
<td>0.01</td>
<td>0.5</td>
<td>&lt;0.5%</td>
<td>90</td>
</tr>
<tr>
<td>0.7-1.0 (38)</td>
<td>0.01</td>
<td>0.5</td>
<td>&lt;0.5%</td>
<td>90</td>
</tr>
</tbody>
</table>

² Estimation based on published SNR value. [Green et al., 1993].

³ The specs reported here are for the original M-7 system.

⁴ The number channels in this spectral region

⁵ The SNR can be improved by flying at a higher attitude with some loss of spatial resolution.
Figure 2-8  Comparison of Images from 6 Different Scanners
The specs in Table 2-7 show very favorable to Misi. Misi has the highest spatial resolution (more than one order of magnitude higher). And Misi's SNR numbers are also among the best. Keys to Misi's impressive performance compared to M-7 are: (1) large aperture optics (15.24 cm vs. 12.25 cm), (2) the use of hybrid scanning which enables Misi to scan at a lower rate, and (3) the use of pyramid mirror to separate the focal planes which results in a very low energy loss in spectral separation. Dichroic mirrors used in M-7 are most likely to have a very low throughput.

Figure 2-8 shows images from six different sensors. A digitized air-photo image was used to simulate Landsat, SPOT and Misi images. These images show how important spatial resolution is in image interpretation. The three real images are generated by the RIT scanner (Misi's predecessor), AVIRIS and the ERIM M-7 Mapper. Clearly, the RIT scanner image contains the most spatial details, and the AVIRIS image has the lowest spatial resolution, but has the best signal-to-noise ratio. The image from M-7 looks very noisy even on the computer screen. The image from Misi is expected have at least the same spatial resolution as its predecessor (RIT scanner) and similar or better SNR than AVIRIS.

2.4 Challenges and Contributions

Misi is a unique electro-optical instrument capable of collecting images at a very high ground resolution with a spectral coverage from 0.4 μm in the visible to 14 μm in the long-wave infrared. The development of Misi represents a new technical challenge to the design, modeling, and performance evaluation of the electro-optical systems. In meeting these challenges, several new techniques of modeling and testing of EO imaging system were developed, which represent the contributions of this thesis. Below are a summary of the four major contributions.

2.3.1 System Engineering Approach to EO Imaging System Design

Electro-optical system design involves seven basic disciplines [Waldman et al. 1992]:

1. Radiation theory: target size and target-to-background and intrinsic contrast.
2. Atmospheric spectral transmissions.
3. Optical design.
4. Detector.
5. Electronic signal processing.
6. Video display and digital image processing.
7. Human search processes (and, for FLIR systems, visual perception).
Due to the multi-disciplinary nature of EO systems, a system engineering approach must be used to integrate the above seven disciplines. Although there are many books describe the general rules of EO imaging system design [Seyrafi 1984; Spiro and Schlessinger 1989, Wyatt, 1991], there are very few published articles on comparison between system design and the final test results. This study is intended to apply the system engineering approach to the design of a multispectral imaging system. Central to this study is the application of system and sub-system MTF analysis of optics, detector, and electronic signal processing. Special emphasis is placed on the comparison of the design specifications with performance evaluation.

The key to the electro-optical system engineering process and a central theme of this thesis is the development of a series of appropriate models. The models can not only predict the system performance, but also be used as a design tradeoff tool in the design process.

2.3.2 Design and Modeling of an Opto-mechanical Scan Mirror

The scan mirror assembly is a critical component of the Modular Imaging Spectrometer Instrument (MISI). Scan mirrors have been used in many airborne and space borne imaging systems, but the image degradation caused by their deformation have not been studied extensively. This is because: (1) for some systems such as the Landsat Thematic Mapper™, although the scan mirror is quite large, it scans at a low speed (for TM, the scan frequency is only 8 Hz), and (2) for some real-time imaging systems like FLIR (forward looking infrared) systems, although the scan mirror scans at a very high speed, the dimension of the scan mirror is very small. It is also possible that for some systems, deformation induced image quality degradation exists, but it is not recognized. Fraedrich et al. (Fraedrich and Confer 1990) did sub-system MTF analysis of the Naval Research Laboratory imaging radiometer. They estimated the scanner contribution to the MTF degradation by dividing the measured MTF of the integrated sub-system by the product of the individually calculated MTFs. It was found that the scan mirror contribution was quite significant. The dynamic deformation of scan mirror under angular acceleration can be derived theoretically for some simple mirror configurations (Brosens 1972) (Conrad 1975), but for most practical scan mirrors, the derivation approach is too complex to pursue. In this study, a modeling method is developed to model the performance of scan mirrors. First, the mirror dynamic deformation is calculated using a finite element analysis (FEA) method, and second, the image quality of the scan mirror is calculated based on this deformation using optical image formation theory. This method can be
used as a design trade tool for scan mirror mechanical design. And finally, an experimental method is described to validate the model.

The biggest advantages of this method are: (1) it can be used for any scan mirror configuration as long as its geometrical properties can be defined in the computer, and (2) by combining dynamic modeling with optical modeling, the mirror designer can link the mechanical design with optical image quality.

2.3.3 A Model Based Algorithm to Derive MTF from Edge Tracing Data

The modulation transfer function (MTF) is one of the most significant image quality descriptors of any imaging system. Traditionally, the MTF of an imaging system can be measured with images of known input, such as sine-wave or square-wave targets, point sources or line sources, and edges. Knife-edge methods are extensively used for operational evaluation because edge is the only natural target available. But knife-edge methods are often inaccurate because of noise. Noise in the measurement not only causes the measured MTF to have random fluctuations, but also results in a bias at higher frequencies [Blackman, 1968].

In this thesis, we mathematically formulated the process of deriving MTF from edge trace data. Our analytical results confirm the conclusion of Swing and McCamy [Swing and McCamy, 1969] that knife-edge method is inaccurate because of noise. Our results also indicate that averaging over several edge traces may introduce an error in the measured MTF. A new algorithm was developed to derive MTF from edge trace data. The algorithm is based on the fact that the overall MTF of any electro-optical imaging system is the product of optical transfer function of the optics, detector aperture function, and the transfer function of electronic signal processing. The transfer function of each subsystem can be described by a model with a few parameters. By using prior knowledge of the imaging system, the MTF of an imaging system can be derived from very noisy edge tracing data.

2.3.4 Electro-optical System Design for Information channel capacity

The performance of EO imaging systems is inevitably constrained by (1) the spatial frequency response (resolution), (2) the signal-to-noise (SNR), and (3) the sampling passband and the aliasing associated with the sampling. The first two factors are well recognized as imaging system design criteria, but the third factor is often neglected.
Since the main objective of any imaging process is to gather information, it is natural that information should be used as a figure-of-merit for design tradeoff.

In this thesis, the information theory was applied to the design of an airborne line-scan imaging system with the emphasis on the differences between the traditional and the informationally optimized design. We used information channel capacity as a figure of merit to optimize the design tradeoffs involving the shape and size of the detector aperture, the electronic cutoff frequency and the order of the electronic filter. Our study indicates that traditional EO system design is not optimized for information, an information optimized design can provide 1.5 bit more information that the traditional design, but at the cost of expensive digital restoration and accurate characterization of instrument in the data collection process. Computer simulation shows that by combining image gathering and digital image processing (restoration), information optimized design can provide images with higher fidelity than that of the traditional design.
3 MISI Design and Modeling

This section describes the design and modeling of the MISI. Several modeling tools are developed to predict the system performance and to aid the design tradeoff process.

3.0 Design Considerations

3.0.1 Review of Competing Technologies

The primary methods available for electro/optical airborne imaging employ framing systems (video, two-dimensional arrays, high rate flying spot scanners), linear array cameras operated in a pushbroom fashion, or line scanners. These methods, or some hybrid, encompass most approaches to multispectral airborne imaging.

3.0.1.A Framing Systems

While we considered a variety of combinations of framing systems that might have covered the spectral range of interest, combining them into a common system with any spatial registration fidelity did not seem practical. Using common optics and off-the-shelf focal planes or flying spot scanners also seemed impractical, because of the large sizes of the focal planes and the spectral band segmentation problems.

3.0.1.B. Push Broom Linear Array Approaches

The use of many linear arrays (e.g. the SPOT approach) or two-dimensional arrays, one dimension spatial - one dimension spectral (e.g. the AIS approach)) appear to hold much more promise. The main limitation of this approach at the present time is the field of view issue. In order to maintain wide fields of view and good resolution, a large number of elements are required in the linear arrays. For a system with a 90° FOV to achieve a 0.3 meter spot size at nadir when flown at 0.3 km would require 2000 element arrays. While these are readily available in the 0.4-1 μm region, availability and cost for the SWIR, MWIR, and LWIR regions become prohibitive.

3.0.1.C. Line Scanner Configuration

The line scanner configurations eliminate most of the detector and optical problems associated with framing and push broom systems and can also be the simplest system capable of high calibration accuracy (Montagu and DeWeerd 1993). They have the
limitation of using a single detector to scan all of the input data in each band. This
generally means that the signal throughput must be very high for the detectors and pre-
amps and the dwell times can be very short, potentially resulting in poorer signal-to-noise
performance. On the plus side, because the optics and focal planes are comparatively
simple, line scanners are much easier to modify and generally have more unused room at
the focal plane to allow for noise reduction schemes such as time delay and integration
(TDI).

The requirements for high resolution, wide field of view, many spectral bands, high
calibration accuracy, and particularly the requirements for modularity, expandability, and
cost resulted in the selection of a hybrid line scanner approach.

3.0.2 Design Concepts and Background Calculations

Given a design based on line scanner approaches and the concern about signal-to-noise,
several initial calculations were made to refine requirements. Table 3.0-1 lists several
physical constraints that governed the following calculations.

We decided to design the system to fit in camera holes designed for standard mapping
cameras. This allows for a more extended use of the system and does not depend on the
rather rare 28-inch camera hole currently available in Calspan Corporation's Aztec C
aircraft. This smaller camera hole requires the scan optics to be located just above the
aircraft skin to avoid vignetting as shown in Figure 3.0-1. This requirement of having the
scan mirror in the bottom of an approximately 17-inch (43 cm) diameter hole led to a
design with an immediate fold after the scan mirror as shown in Figure 3.0-2. Space
constraints dictate that the diameter of the largest practical scan mirror (which controls
the limiting aperture on the telescope) is 6 inches (15.24 cm). This constraint, coupled
with the 0.3 km minimum flight altitude and the desire to not push the F# much below
3.5, led to the rest of the optical system design as detailed in Section 3.3.
Figure 3.0-1  Location of the scan mirror relative to aircraft skin

Table 3.0-1  Physical Limits for Design Considerations

<table>
<thead>
<tr>
<th>A. Fixed Parameters</th>
<th>B. Derived Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
</tr>
<tr>
<td>• minimum flying height</td>
<td>1 milliradian</td>
</tr>
<tr>
<td>• inside diameter of standard camera port (i.e. OD of scanner shell) (N.B. available Aztec C has a 71 cm camera hole)</td>
<td>0.3 km (1000 feet)</td>
</tr>
<tr>
<td></td>
<td>43 cm (17 inches)</td>
</tr>
<tr>
<td>• minimum ground speed of aircraft (N.B. this is for an Aztec C, most camera planes would be slower)</td>
<td>110 mph (49 m/sec)</td>
</tr>
<tr>
<td>• minimum ground instantaneous field-of-view</td>
<td>0.3 m (1 foot)</td>
</tr>
<tr>
<td></td>
<td></td>
</tr>
<tr>
<td>• optical aperture</td>
<td>15.2 cm (6 inches)</td>
</tr>
<tr>
<td>• scan rate at minimum altitude</td>
<td>82 revolutions/sec</td>
</tr>
<tr>
<td>• maximum signal bandwidth for 1 milliradian detector</td>
<td>257x10^3Hz</td>
</tr>
</tbody>
</table>
In order to achieve a one foot (0.3m) spot size (GIFOV) when operated at the lowest practical flight altitude (≈0.3 km) a minimum instantaneous field of view (IFOV) of one milliradian is required, i.e.

\[
IFOV = \frac{GIFOV}{H} = \frac{0.3m}{300m} = 0.001
\]  

(3.0.1)

The minimum aircraft ground speed (v) dictates the time per revolution (\(t_r\)) of the scan-mirror to provide continuous sampling of adjacent 0.3m wide lines of data according to

\[
t_r = \frac{GIFOV}{v} = \frac{0.3m}{49(m/s)} = 6.1 \times 10^{-3} \text{sec}
\]  

(3.0.2)

This leads to high sample rates (low dwell time) and high signal bandwidths computed as:

\[
t_s = \frac{t_r}{N_r} = \frac{6.1 \times 10^{-3} \text{sec}}{6283} = 9.7 \times 10^{-7} \text{[Seconds/sample]}
\]  

(3.0.3)
\[ f_s = 1 / t_s = 1.03 \times 10^6 \text{[samples/second]} \]  
\[ f = f_s / (2 \text{samples/cycle}) = 5.05 \times 10^5 \text{[cycles/sec]} \]  

where \n
- \( t_s \) = number of seconds between sample centers (dwell time if continuously sampling)  
- \( f_s \) = sample frequency  
- \( f \) = signal bandwidth (i.e. sample frequency adjusted for Nyquist limit of 2 samples per cycle).  
- \( N_r \) = number of IFOV's per revolution = \( 2\pi \text{radians/1.10}^{-3} \text{radians} = 6283 \).  

The required mirror scan rate of 163 Hz \((1/t_r)\) led to poor signal-to-noise ratio and mechanical problems due to the high spin rate for the large (6 inch) mirror. To overcome this problem, a hybrid approach was adapted such that the spectral bands with one milliradian IFOV would be sampled two lines at a time. Each revolution of the scan mirror will scan two lines, thus the forward advance of the aircraft before the next sweep can be twice as far and still achieve contiguous coverage (cf. Figure 3.0-3).  

Thus \n
\[ t_r = \frac{2 \times \text{IFOV's}}{N_r} = 12.25 \times 10^{-3} \text{sec} \]  

with a scan rate of \( 1/t_r = 82 \) revolutions/sec and \n
\[ t_s = t_r / N_r = 1.9 \times 10^{-6} \text{ sec} \]  
\[ f_s = 1 / t_s = 513 \times 10^3 \text{[samples/second]} \]  
\[ f = f_s / 2 = 257 \times 10^3 \text{[cycles/sec]} \]
These values generate more acceptable maximum scan rates, data rates, and S/N values. In addition, it was decided to use a variable scan rate mirror to slow the scan rate at higher altitudes. For example, at 0.6 km the GIFOV is only 0.6 m which just equals the ground advance for a single one milliradian IFOV detector at an 82 Hz scan rate. Since we have two detectors, the scan rate can be cut in half again and still maintain coverage. If the sample integration time is also doubled, the signal-to-noise can be increased by approximately $\sqrt{2}$ (cf. Section 3.4). Thus, at least four scan rates and sample integration times are planned.

### 3.0.3 Radiation Transfer Model

The radiance from the ground is given by Eq. 2.2. For the reflective channels (i.e. the visible-near-infrared channels), Eq. 2.2 can be reduced to

$$L(\lambda) = (1/\pi)E(\lambda)\tau_1(\lambda)r(\lambda)\tau_2(\lambda) + L_d(\lambda)r(\lambda)\tau_2(\lambda) + L_u(\lambda)$$  \hspace{1cm} (3.0.10)

and for the long-wave infrared channels and the mid-wave infrared channels operated in the dark, Eq. 2.2 can be reduced to

$$L(\lambda) = L_d(\lambda)r(\lambda)\tau_2(\lambda) + \varepsilon(\lambda)\tau_2(\lambda)L_b(\lambda) + L_u(\lambda)$$  \hspace{1cm} (3.0.11)
Figure 3.0.4 Radiation transfer from ground to the sensor

Figure 3.0.4 shows the radiation transfer from ground to the sensor. A portion of the ground radiance is collected by the sensor optics and focused to the detector elements in the focal plane. Assuming Lambertian behavior for the ground cover [Feng et al. 1993], the radiation collected by the optical aperture is given by

$$ \phi(\lambda) = \int_{\Omega} L(\lambda) \cos(\theta) A_g d\Omega \approx L(\lambda) A_g \frac{\pi D^2}{4a^2} = L(\lambda) A_d \frac{\pi D^2}{4f^2} = \frac{\pi L(\lambda) A_d}{4 F_\#^2} $$

(3.0.12)

where $\Omega$ is the solid angle of the optical aperture subtended the ground pixel, $\theta$ is the sensor view angle, $F_\# = \frac{f}{D}$ is the f-number of the optics, $A_g$ is the area on the ground and $A_d$ is the detector area. The energy reaches the detector is given by

$$ \phi_d = \int_{\Delta \lambda} \frac{\pi L(\lambda) A_d}{4 F_\#^2} \tau_0 d\lambda \approx \frac{\pi L(\lambda) A_d}{4 F_\#^2} \tau_0 \Delta \lambda $$

(3.0.13)

where $\tau_0$ is the throughput of the optical system, and $\Delta \lambda$ is the spectral bandwidth of the sensor. The noise equivalent power (NEP) is defined as

$$ NEP = \frac{\sqrt{A_d \Delta f}}{D^*} $$

(3.0.14)
where $\Delta f$ is the bandwidth of electronic signal determined by Eq. (3.0.9), and $D^*$ is the detectivity of the detector. The signal-to-noise ratio (SNR), noise equivalent spectral radiance (NER), noise equivalent reflectance (NE$\Delta \rho$), and noise equivalent temperature difference (NETD) are given by

$$SNR = \frac{\phi_d}{NEP}$$  \hspace{1cm} (3.0.15)

$$NER = \frac{4F_h^2NEP}{\pi A_d\tau_o \Delta \lambda}$$  \hspace{1cm} (3.0.16)

$$NE\Delta \rho = \frac{NER}{\partial L/\partial R}$$  \hspace{1cm} (3.0.17)

$$NETD = \frac{NER}{\partial L/\partial T}$$  \hspace{1cm} (3.0.18)

Where $\partial L/\partial R$ is the radiance change corresponds to unit reflectance change, and $\partial L/\partial T$ is the radiance change corresponds to one degree temperature change. These equations will be used to evaluate the radiometric performances of the MISI.

Based on these initial requirements, conditions, and preliminary calculations, a system was designed and more detailed subsystem designs initiated as described in the following sections. For design purposes, the instrument was divided into the following subsystems:

A. Mechanical
B. Optical
C. Detectors/Preamps
D. Signal Processing
E. Calibration

The design of each of these subsystems will be treated individually in these sections. In order to place these subsystems in a clearer context, we will first briefly describe the overall system.

### 3.1.1 System Overview

Figure 3.1-1 shows an overall view of the MISI system. The primary opto-mechanical package is designed to have the scan mirror and fore-optics drop into a standard camera hole. This results in a 17-inch diameter snout on the system with the main mounting plate extending beyond the camera hole for shock mounting of the overall assembly. The detectors and preamplifiers are mounted in the upper section of the primary opto-
mechanical assembly. The conditioning electronics and recording system are rack mounted in a separate package.

Figure 3.1-1 Overview of MISI System

The overall optical system with the various focal planes is shown in Figure 3.1-2. The 6-inch clear aperture scan mirror will spin at up to 82 revolutions per second and folds the image onto a second fold mirror which reflects the image into the F3.33 Dall-Kirkham Cassegrarian-style telescope. The converging image is split onto four slightly off-axis (less than 2°) focal planes by a four-sided pyramid mirror as shown in Figure 3.1-2. The on-axis rays pass through a hole in the center of the pyramid mirror and are used to sample the VNIR region. Of the four off-axis focal planes, two are in the along scan direction (one leading, one lagging the primary optical axis), and two are in the along track direction (one fore and one aft of the primary scan axis). For the initial design, only the on-axis and along scan focal planes are used. The along track focal planes are intended for additional modules.
Figure 3.1-2 Scanner Optical Schematic

Figure 3.1-3 shows the projection of the various detectors and spectrometer apertures onto the ground, as well as the field of view of the fore and aft focal planes. Table 3.1-1 shows the IFOV, operating temperature, and nominal spectral bandwidth of the detectors planned for initial development. The modular nature of the focal planes allow for easy addition of new focal planes or modification of an existing one.
Table 3.1-1 Summary of MASI Channels

<table>
<thead>
<tr>
<th>Channel</th>
<th>Wavelength (um)</th>
<th>Spatial Resolution (mrad)</th>
<th>Detector</th>
<th>Number of Detectors</th>
<th>Operating Temperature</th>
</tr>
</thead>
<tbody>
<tr>
<td>VNIR 1</td>
<td>0.5-0.9</td>
<td>1</td>
<td>Si</td>
<td>2</td>
<td>ambient</td>
</tr>
<tr>
<td>SWIR 1</td>
<td>1.5-1.75</td>
<td>2</td>
<td>InSb</td>
<td>1</td>
<td>77 K</td>
</tr>
<tr>
<td>SWIR 2</td>
<td>2.07-2.35</td>
<td>2</td>
<td>InSb</td>
<td>1</td>
<td>77 K</td>
</tr>
<tr>
<td>MW 1</td>
<td>3-5</td>
<td>1</td>
<td>InSb</td>
<td>2</td>
<td>77 K</td>
</tr>
<tr>
<td>MW 2</td>
<td>3.4-2</td>
<td>2</td>
<td>InSb</td>
<td>1</td>
<td>77 K</td>
</tr>
<tr>
<td>MW 3</td>
<td>4.5-5</td>
<td>2</td>
<td>InSb</td>
<td>1</td>
<td>77 K</td>
</tr>
<tr>
<td>LW</td>
<td>8-14</td>
<td>1</td>
<td>HgCdTe</td>
<td>2</td>
<td>77 K</td>
</tr>
<tr>
<td>LW 1</td>
<td>8-10</td>
<td>2</td>
<td>HgCdTe</td>
<td>1</td>
<td>77 K</td>
</tr>
<tr>
<td>LW 2</td>
<td>10-12</td>
<td>2</td>
<td>HgCdTe</td>
<td>1</td>
<td>77 K</td>
</tr>
<tr>
<td>LW 3</td>
<td>12-14</td>
<td>2</td>
<td>HgCdTe</td>
<td>1</td>
<td>77 K</td>
</tr>
<tr>
<td>Vis. Spect</td>
<td>0.4-0.7</td>
<td>2</td>
<td>Si Array</td>
<td>38</td>
<td>ambient</td>
</tr>
<tr>
<td>VNIR Spect</td>
<td>0.7-1.0</td>
<td>2</td>
<td>Si Array</td>
<td>38</td>
<td>ambient</td>
</tr>
</tbody>
</table>

Figure 3.1-3 Projection of Detectors from All Focal Planes onto the Ground

The scan mirror will operate at four user selectable speeds from 82 rps to approximately 10 rps with proportional increases in the dwell times (integration time constant of the signal conditioning circuits). An optical encoder, coupled to the mirror shaft, will provide synchronization timing pulses every milliradian to control line start, pixel sampling, and sampling of calibration standards.
The scanner's total field of view will be 90° (±45) with calibration standards viewed every rotation during the dead time. Figure 3.1-4 shows the scan sequencing for each rotation. The dual blackbodies will be sampled each line and the visible and near infrared (VNIR) calibrator will be imaged once each line with the intensity incremented every ten lines through eight levels.

![Diagram](image)

Figure 3.1-4 Time Sequences of MISH Signal Trains Generated Using Sync Signals and Signal Combinations

The system utilizes one sync disk with two tracks. One track has one single line used to generate the start of line signal. The start of line signal tells the data collection system to start capturing data, and the other track contains a circular grating, which provides pulses every milliradian (cf. Figure 3.1-5). In this system, the pixel clock is actually used to trigger the analog-to-digital (A/D) conversion.
For each thermal channel, the video data contains 90 degree of ground data, and two blackbody references for thermal calibration. Each visible and near infrared signal will consist 90 degree of ground data and a single calibration. There is only one calibration source for the VNIR channels. It will be stepped every 25 lines in approximately 10% increments in intensity, thereby generating a calibration step wedge for onboard calibration.

Each detector has an individual preamplifier, and the signals from the preamplifiers will be grouped into similar response groups for final conditioning (cf. Figure 3.1-6). The final data recording is 12 bits per sample.

The MISI instrument is designed to function not only as an airborne instrument, but will also be capable of panoramic imaging from RIT's North Range Lab. In this mode, the imager will be tilted $\approx 45^\circ$ and mounted on a slow speed turntable. By removing the stray light cowling and shifting the line start location, images of the horizon ($\pm 45^\circ$) will be acquired (cf. Figure 3.1-7). This panoramic mode will allow full up system testing, calibration, alignment, etc. in a laboratory environment, as well as provide a collection platform for signature studies and atmospheric calibration studies. It will also facilitate brass board testing of new modules or system upgrades.

The details of each subsystem are contained in the following sections. Plans for additional focal planes are not treated to any extent because they are not part of the base-
line system. However, the fore and aft focal plane areas will be available and the base line focal planes are modular and easily removed to allow installation of other focal planes. In fact the LWIR and MWIR/SWIR focal planes are mounted in standard 2-inch diameter side-looking dewars to facilitate interchange with other detector/filter combinations or arrays.

![Diagram of scanner and turntable](image)

**Figure 3.1-7** Atmospheric propagation and sky radiation phenomena study at the Roof Lab

Plans for active illumination with the MISI are also not treated to any extent since they are not part of the base-line system. However, an option for active illumination is available by placing a fold on the back of the secondary in the Cassegrain and taking advantage of the central obscuration of the telescope (cf. Figure 3.1-8). This option could potentially be used for active reflective imaging to measure emissivity values in the LWIR (i.e. to separate out temperature and emissivity effects) using CO₂ lasers or for laser-induced luminescence studies where the laser illumination and sensor are at different wavelengths.
3.2 Mechanical Subsystem

The mechanical design of the scanner was governed by the idea of limiting the operating costs. In order to do this, the instrument must fit into a common port for 9-inch format mapping cameras which are used in low and medium altitude aerial photography in light aircraft.

With this as a primary goal, the system is designed to fit in a 17-inch diameter camera port. The 6-inch collection aperture and a 90° scan angle demand that the scan mirror be in close proximity to the external skin of the aircraft. It is these criteria that drove the design to a scan mirror looking into a Cassegrain via a 45° fold mirror (cf. Figure 3.1-2).

The design is built around three “levels” in the vertical direction; first, the scan mirror and fold, second, the Cassegrain and calibration sources, and third, the focal planes (cf. Figure 3.2-1).
The scan mirror is a reinforced aluminum shell dynamically balanced and belt-driven by a DC motor. An optical encoder is mounted on the scan mirror assembly which provides 1 milliradian angular resolution. The fold mirror is an elliptical flat with a 6 1/8-inch clear aperture. Both the scan mirror assembly and the fold mirror are mounted to the main 17-inch diameter base plate shown in Figure 3.2-1.

The Cassegrain is located above the fold mirror looking downward into the fold mirror. Space is available on the back of the secondary mirror for a small 45° mirror (c.f. Figure 3.1-8). That mirror can be used to fold the laser beam from the active source into the scanner’s optical path. The secondary mirror has three point adjustment for fine alignment of the Cassegrain. The secondary mirror mount can also be adjusted along the optical axis with respect to the primary for proper focus of the Cassegrain. As a whole system, the Cassegrain is movable along the optical axis for proper focus at the focal plane level.

The focal plane is in the level above the primary mirror of the telescope. This level has an “inverted pyramid” mirror with a center bore through it. This gives five separate focal planes; one on-axis, two off-axis along the scan direction, and two off-axis along the flight line direction. To minimize distortions, only the three focal planes on the scan line axis are presently being used.
Figure 3.2-1 Opto-Mechanical Levels.
The two IR dewars are placed on the sides of the pyramid. These dewars are mounted in a collar-type clamp to facilitate vertical alignment. In addition, these collars may also be adjusted in the horizontal plane for alignment and focus.

The center bore in the pyramid is used by the high resolution (1 mrad) silicon detectors and the two imaging spectrometers. Two optical fibers (1.5 mm) are placed on the sides of the silicon detector to collect energy for the visible and near infrared spectrometers. The two fibers are 3 milliradians off-axis. The silicon detectors and fibers are mounted on a common fixture which is adjustable for focusing of the system.

Instability of the aircraft platform in flight necessitates a mean for monitoring roll, pitch, and yaw. Initially, a single axis roll-monitoring gyro system will be installed on the scan head mounting plate. The amount of roll per scan line is measured by the gyro system and is recorded for ground digital image processing. The amounts of pitch and yaw are not measured for the base-line system, but will be upgraded in the future if they are needed.

All electronic systems except the pre-amplifiers will be in a standard rack mount configuration with proper shock mounts. All the pre-amplifiers will be located very close the focal planes to reduce the noise and interferences.

3.3 Optical Subsystem

The optical layout of the MISI system is shown in Figure 3.1-2. The scene energy from the ground is first reflected off the scan mirror, and then folded into the telescope by the fold mirror. The scene radiation is then focused by the telescope to form an image on the primary focal plane of the telescope. The primary focal plane is sub-divided into five small focal planes via a pyramid mirror. The five small focal planes are long-wave (LW), short/mid-wave (SMW), visible and near infrared (VNIR), and two unused focal planes for future expansion. For the three focal planes being used now, each focal plane corresponds to a slightly different area in the scan direction. A given object in the scene will be sampled first by the SMW focal plane, then the VNIR focal plane, and last by the LW focal plane (cf. Figure 3.1-3).

3.3.1 Scan Mirror

The scan mirror assembly is a critical component of the MISI. The scan mirror is a cantilevered rotary mirror with an effective optical aperture of 6" (76.2 cm) (cf. Figure
In order to maintain the spatial resolution in the flight direction, the mirror must rotate at about 82 Hz (4800 rpm) (cf. Section 3.0.2). The centrifugal force created by the high rotation speed and the large size of the mirror will cause deformation of the mirror surface. This deformation will consequently affect the image quality of the scanner system. Also due to the asymmetrical configuration of the cantilevered mirror, the unbalanced force and torque, if not properly counter-balanced, will cause detrimental vibrations. The space limitation of the scan mirror assembly further complicates the problem. A modeling approach was used in the mirror design process to compromise between image quality and mechanical stability. The mirror dynamic deformation is calculated using a finite element analysis (FEA) method, and the image quality of the scan mirror is calculated based on this deformation using optical image formation theory.

\[ F = \rho \Delta v \omega^2 R \]  

(3.3.1)
where $\rho$ is the density of the material, $\Delta v$ is the volume of the solid body, $\omega$ is the angular velocity, and $R$ is the radius of the solid body to the rotation axis. According to the theory of elasticity, we have three sets of equations (Oden and Ripperger 1980):

(1) Equilibrium equations (which show the relationships between the first derivatives of stresses and the body forces):

$$
\frac{\partial \sigma_x}{\partial x} + \frac{\partial \tau_{yx}}{\partial y} + \frac{\partial \tau_{zx}}{\partial z} + F_x = 0
$$

$$
\frac{\partial \tau_{xy}}{\partial x} + \frac{\partial \sigma_y}{\partial y} + \frac{\partial \tau_{zy}}{\partial z} + F_y = 0
$$

$$
\frac{\partial \tau_{xz}}{\partial x} + \frac{\partial \tau_{yz}}{\partial y} + \frac{\partial \sigma_z}{\partial z} + F_z = 0
$$

(3.3.2)

(2) Stresses-strains-relationship (the generalized Hooke's law):
\[ \varepsilon_x = \frac{1}{E} [\sigma_x - \nu(\sigma_y + \sigma_z)] \quad \gamma_{xy} = \frac{\tau_{xy}}{G} \]

\[ \varepsilon_y = \frac{1}{E} [\sigma_y - \nu(\sigma_x + \sigma_z)] \quad \gamma_{yz} = \frac{\tau_{yz}}{G} \]

\[ \varepsilon_z = \frac{1}{E} [\sigma_z - \nu(\sigma_x + \sigma_y)] \quad \gamma_{xz} = \frac{\tau_{xz}}{G} \]  

(3.3.3)

(3) Strains-displacements-relationship:

\[ \varepsilon_x = \frac{\partial u_x}{\partial x} \quad \varepsilon_y = \frac{\partial u_y}{\partial y} \quad \varepsilon_z = \frac{\partial u_z}{\partial z} \]

\[ \gamma_{yz} = \frac{\partial u_z}{\partial y} + \frac{\partial u_y}{\partial z} \]

\[ \gamma_{zx} = \frac{\partial u_x}{\partial z} + \frac{\partial u_z}{\partial x} \]

\[ \gamma_{xy} = \frac{\partial u_y}{\partial x} + \frac{\partial u_x}{\partial y} \]  

(3.3.4)

where \( \sigma \) is the normal stress, \( \tau \) is the shearing stress, \( E \) is Young's modulus of elasticity and \( G \) is the modulus of rigidity of the material, \( \varepsilon \) is the strain, \( \gamma \) is the shear strain, \( u \) is the displacement, and \( F \) is the body force, for the mirror case, if we neglect the gravitational force, \( F \) is the centrifugal force. The subscript denotes the direction of stresses, strains, and forces.

Due to the complex structure of the scan mirror, it is impossible to solve Equation 1-4 to get an analytical solution. Thus a numerical approximation method must be used to solve for displacement at the mirror surface.

Finite element analysis (FEA) is a method which transforms an engineering system with an infinite number of unknowns to one that has a finite number of unknowns related to each other by elements of finite size as shown in Fig. 3.3-3 (Desai and Abel 1972). These elements are interconnected at the joints which are called nodes. For each element, the displacements vector \{u\}, which represent the responses to applied actions, and nodal force \{f\} are related by
\[ [K] \{u \} = \{f \} \tag{3.3.5} \]

where \([K]\) is the stiffness matrix. The stiffness matrix consists of the coefficients of the equilibrium equations derived from the material and geometric properties of an element and obtained by use of the principle of minimum potential energy. Once \([K]\) is determined for each element, all the individual \([K]\) matrices are assembled to form the set of simultaneous equations. The overall equilibrium relations between the total stiffness matrix \([K]\), the total load vector (centrifugal force) \([F]\), and the nodal displacement vector for the entire body \([r]\) will again be expressed as a set of simultaneous equations.

\[ [K] \{r\} = \{F\} \]

With proper geometric boundary conditions (for the mirror case, it is the mirror mount), this equation can be solved for the nodal displacements.

Figure 3.3-3 FEA Model of Scan Mirror

In this thesis, a general purpose finite element analysis program (ANSYS) was used to calculate the mirror deformation under the centrifugal force (Swanson 1990).

3.3.1.B Optical Modeling: Optical Image Formation

Fig. 3.3-4 shows a schematic of a deformed scan mirror. The wave-front aberration is defined as the path length difference of light between the wave-front reflected by the
deformed mirror with the wave-front reflected by the perfect plane mirror. According to Fig. 3.3-4, the wave front aberration is given by

$$w(\xi, \eta) = \frac{2\pi \delta(\xi, \eta)}{\lambda \cos(45^\circ)}$$

(3.3.6)

where $\xi$ and $\eta$ are pupil coordinates, $\delta(\xi, \eta)$ is the deformation in the normal direction of the scan mirror as shown in Fig. 3.3-4, and $\lambda$ is the wavelength of light.

Figure 3.3-4 Wave-front aberration and deformation

For the case of an object at infinity (a good approximation for most remote sensing applications), the point spread function (PSF) of the scan mirror is the squared modulus of the Fourier transformation of the generalized pupil function (Goodman 1968).

$$PSF(x, y) = \left| k \int_{-\infty}^{\infty} a(\xi, \eta) \cdot \exp(-i w(\xi, \eta)) \cdot \exp(-i \frac{2\pi}{\lambda f} (x\xi + y\eta)) \, d\xi \, d\eta \right|^2$$

(3.3.7)

where $x$ and $y$ are the coordinates in the image space, $k$ is a normalization constant, $a(\xi, \eta)$ is the aperture function which is the amplitude of the generalized pupil function, the phase term $w(\xi, \eta)$ in the generalized pupil function is the wave-front aberration given by
Eq. 3.3-6, and $f$ is the focal length of the telescope. The modulation transfer function (MTF) of the scan mirror is the modulus of the Fourier transformation of the point spread function

$$ \text{MTF}(f_x, f_y) = \left| \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \text{PSF}(x,y) \exp(-i2\pi(xf_x + yf_y)) \, dx \, dy \right| $$

(3.3.8)

where $f_x$ and $f_y$ are the spatial frequency in the x and y directions. Fig. 3.3-5 shows the PSF of a diffraction limited system (aberration free). The PSF has a symmetrical ring structure which is generally referred to as the Airy pattern. The PSF of a scan mirror with deformation will be much worse than that.

The diffraction limited blur circle, which is defined to be the radius of the first dark ring in the point spread function (cf. Fig. 3.3-5) is given by [Goodman, 1968]

$$ \sigma = 1.22 \frac{\lambda f}{D} $$

(3.3.9)

where $D$ is the diameter of the mirror. For this system, $f/D = 3.333$, the blur circle is 2.57 $\mu$m at $\lambda = 0.6328$ $\mu$m. But in the presence of deformation and aberration, the blur circle will be much larger.

Figure 3.3-5 The point spread function of a diffraction limited system
3.3.1. C Shell Structured Mirror

The image quality of the mirror were evaluated for three scan-mirror configurations (Feng et al. 1993): (1) aluminum flat mirror, (2) solid mirror, and (3) shell structured mirror. The flat mirror is not strong enough to produce reasonable image quality at high speed, while the solid mirror is strong enough, but too heavy to be balanced. The shell structured mirror is a compromise of the flat mirror and solid mirror. The mirror surface is supported by a rib structure as shown in Fig. 3.3-6 to keep the strength of the solid mirror and to reduce the weight. Also note that some ribs are removed to leave room to put in the counter-balance weight. Each rib is 5 mm thick, and the mirror surface is 10 mm thick. Again the whole assembly is made of aluminum material. The mirror was modeled using ANSYS. The element type used in the analysis is 4-node quadrilateral shell. Fig. 3.3-7 shows the deformation contour of the shell structured mirror at 80 Hz. The maximum deformation is about 27 μm.

![Figure 3.3-6 Shell structured mirror](image)

(left: without the mirror surface  
Right: with the mirror surface)
Figure 3.3-7 The deformation of shell structured mirror (80 Hz)

Fig. 3.3-8 shows the PSF of the shell structured mirror. The image quality is slightly worse than that of a solid mirror, but the PSF is still completely contained in the detector area, which satisfy our spatial resolution requirement.

Figure 3.3-8  The PSF of shell structured mirror at 80 Hz
Since this mirror is asymmetrical about the rotation axis, it will create a large unbalanced force when rotating at a high speed. Two counter-balance weights must be used to dynamically balance the scan mirror. Fig. 3.3-9 shows the mirror balancing scheme, where $F_0$ is the unbalanced force caused by the mirror. In order to have the scan mirror assembly dynamically balanced, the two weights ($m_1$ and $m_2$) and distance ($L_1$ and $L_2$) must satisfy the following equations to simultaneously balance the centrifugal and asymmetric torque effects (Equilibrium conditions).

The equilibrium conditions are:

$$
F_1 = F_0 + F_2 \\
F_2 L_2 + F_1 L_1 - M_1 = F_0 R + M_0
$$

(3.3.10)

where: $F_1 = \frac{m_1 \omega^2}{r_1}$ is the centrifugal force of $m_1$, and $F_2 = \frac{m_2 \omega^2}{r_2}$ is the centrifugal force of $m_2$, $M_0$ is the torque of the mirror surface, $M_1$ is the torque about its center of gravity due to the asymmetrical shape of counter-balance block $m_1$. For this mirror rotating at 80 Hz, $F_0 = 7660$ Newton, $M_0 = 255$ Newton\*m. The masses required to balance the mirror assembly are 1.8 kg for $m_1$ and 0.9 kg for $m_2$, which is only a quarter of the mass needed to balance the solid mirror [Feng et al. 1994]. Both masses can be easily fitted into the available space in the mirror assembly.
Figure 3.3-9 Dynamic balance of shell structured mirror

The fold mirror is a 9-inch by 6-inch flat glass mirror. A final coat of silicon dioxide (SiO2) was applied to protect the silver coating.

The scan-mirror assembly also generates timing pulses to provide synchronization for the electro-mechanical scanner.

3.3.2 Telescope System

The scene energy reflected by the fold mirror enters a telescope consisting of a primary and secondary mirror, a configuration referred to as a Dall-Kirkham design. The focal length of the telescope is 20 inches, and the f-number is 3.333. A trade off study found that the longer the distance between the primary and secondary, the better the off-axis image quality, and the larger the central obscuration due to the secondary mirror. A spacing of 5 inches was chosen as a compromise between the off-axis image quality and throughput. The primary mirror is a 6-inch ellipse with a curvature of 0.0575 inch⁻¹ and aspheric factor of 0.51. The secondary is a 2.55 inch sphere with a curvature of 0.07647 inch⁻¹. Figure 3.3-10 shows the actual ray tracing output generated by a commercial optical design software (SuperOslo).
The diffraction limited resolution is given by Equation 3.3-9. For D=6" and f=20", the diffraction limited resolution will be $\sigma = 4 \, \mu m$ at the wavelength of 1 \, \mu m.

The actual resolution is worse than that due to aberrations. For this system, the on-axis image quality is almost diffraction limited, but the off-axis image quality is much worse due to coma. The RMS blur at 20 IFOV off-axis is 0.14 mm, which is still only a third of an IFOV. Figure 3.3-11 shows the spot diagram at the center of the field of view and at 20 IFOV off-axis. Figure 3.3-12 and 3.3-13 show the MTF of the telescope. The cutoff frequency is about 25 lines/mm at 20 milliradian off-axis, which is still much higher than the cutoff frequency of detector aperture (1 line/mm for the high resolution channels).

The overall resolution of the optical system is also limited by the scan mirror when operated at high scan rate. The on-axis image quality is limited by the scan mirror dynamic deformation (cf. Fig. 3.3-8), the off-axis image quality is limited by the telescope. The finial image quality is limited by the detector aperture (cf. Fig. 4.2-6).
Figure 3.3-11 The spot diagrams of the telescope predicted by SuperOslo
Figure 3.3-12 On-axis MTF of the telescope based on optical ray-tracing (SuperOslo Output)

Figure 3.3-13 Off-axis MTF of the telescope predicted by optical ray-tracing at 20 milliradian off-axis (SuperOslo Output)
3.3.3 Pyramid Mirror

A pyramid mirror is used to split the telescope field of view into five small focal planes. Figure 3.3-14 shows the schematic of the pyramid mirror and Figure 3.3-15 shows the mirror in relation to the cryogenic focal planes and dewar assemblies.

![Pyramid Mirror Diagram]

Figure 3.3-14 Pyramid mirror showing the projection of the off-axis focal planes
The pyramid mirror is made of glass with Al coating. Again, a SiO2 coating is used to protect the Al coating.

### 3.3.4 Imaging Spectrometer

A silicon photodiode array imaging spectrometer was designed to sample the visible and near infrared spectrum with high spectral resolution (10 nm). Since each spectrometer covers only the spectral range of about 400 nm, two spectrometers are employed: one for the visible and the other for the near infrared. A silica fiber is used to relay the optical radiation from the VNIR focal plane to the entrance aperture of each spectrometer as shown in Figure 3.3-16. The Littrow configuration is used because of the simplicity. A lens is used for both the incidence beam and exit beam, which makes the spectrometer very compact. Figure 3.3-16 shows the optical layout of the imaging spectrometer.
Figure 3.3-16  Spectrometer and detector assembly

The spectrometer design parameters are as follows:

- Grating density: 1200 lines/mm
- Blaze angle: 17.5° for the visible and 26.7° for the near-infrared
- Focal length: 75 mm
- Spectral resolution: 10 nm
- F-number: 3.333 (matches the telescope)

Table 3.3-1 shows the diffraction angle (β) as a function of the wavelength. Figure 3.3-17 shows the ray tracing provided by SuperOslo with a doublet as the spectrometer lens. The blur spot is less than 0.5 mm across the full field of view, which is still smaller than the detector size (1 mm). Since a well-designed camera lens is used, the blur due to the spectrometer optics is negligible.
Table 3.3-1 The Relationship Between Wavelength and Diffraction Angle and Spectral Bandwidth

<table>
<thead>
<tr>
<th>Wavelength (nm)</th>
<th>sin(β)</th>
<th>β (diff. angle)</th>
<th>dβ/dλ</th>
<th>Δλ</th>
<th>x</th>
</tr>
</thead>
<tbody>
<tr>
<td>400</td>
<td>0.129796</td>
<td>-13.0421</td>
<td>0.00123</td>
<td>9.49075</td>
<td>-17.372</td>
</tr>
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<td>0.00120</td>
<td>9.72428</td>
<td>-12.629</td>
</tr>
<tr>
<td>500</td>
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<td>0.00120</td>
<td>9.88948</td>
<td>-7.9282</td>
</tr>
<tr>
<td>550</td>
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<td>0.00120</td>
<td>9.98161</td>
<td>-3.2127</td>
</tr>
<tr>
<td>600</td>
<td>0.36979</td>
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<td>0.00120</td>
<td>9.99559</td>
<td>1.57537</td>
</tr>
<tr>
<td>650</td>
<td>0.42979</td>
<td>4.95490</td>
<td>0.00120</td>
<td>9.92539</td>
<td>6.50210</td>
</tr>
<tr>
<td>700</td>
<td>0.48979</td>
<td>8.82750</td>
<td>0.00120</td>
<td>9.76450</td>
<td>11.6473</td>
</tr>
<tr>
<td>750</td>
<td>0.54979</td>
<td>12.8533</td>
<td>0.001230</td>
<td>9.50513</td>
<td>17.1129</td>
</tr>
<tr>
<td>750</td>
<td>0.33647</td>
<td>-14.637</td>
<td>0.001240</td>
<td>9.36144</td>
<td>-19.587</td>
</tr>
<tr>
<td>800</td>
<td>0.39647</td>
<td>-10.941</td>
<td>0.00122</td>
<td>9.63973</td>
<td>-14.498</td>
</tr>
<tr>
<td>850</td>
<td>0.45647</td>
<td>-7.1395</td>
<td>0.00120</td>
<td>9.84552</td>
<td>-9.3942</td>
</tr>
<tr>
<td>900</td>
<td>0.51647</td>
<td>-3.20327</td>
<td>0.00120</td>
<td>9.96877</td>
<td>-4.1974</td>
</tr>
<tr>
<td>950</td>
<td>0.57647</td>
<td>0.90367</td>
<td>0.00120</td>
<td>9.99750</td>
<td>1.18299</td>
</tr>
<tr>
<td>1000</td>
<td>0.63647</td>
<td>5.23021</td>
<td>0.001205</td>
<td>9.91690</td>
<td>6.86534</td>
</tr>
<tr>
<td>1050</td>
<td>0.69647</td>
<td>9.84569</td>
<td>0.00121</td>
<td>9.7076</td>
<td>13.0162</td>
</tr>
<tr>
<td>1100</td>
<td>0.75647</td>
<td>14.855</td>
<td>0.00124</td>
<td>9.34270</td>
<td>19.893</td>
</tr>
</tbody>
</table>

Figure 3.3-17 Optical ray-tracing of the imaging spectrometer

3.3.5 Focal Planes

Of the five possible focal planes, only three are used in the base-line design. The other two are for future expansions.

3.3.5.A Long-wave Focal Plane

There are five HgCdTe (Mercury Cadmium Telluride) detectors on the LW focal plane. Among them, two detectors are 1-mrad, broad-band detector (8-14 μm), and three are 2-mrad detectors; each covers a 2 μm spectral band from 8 to 14 μm. Since pixel registration of different scan lines is extremely difficult, all detectors must be aligned in the same scan line. The detectors are also arranged such that the spacing between
detectors is an integer number of IFOV so that images from different detectors can be registered simply by shifting an integer number of pixels. The detectors are placed close together to keep them near the telescope's optical axis where image quality is the best. Keeping detectors together can also reduce the time delay between data collection for each band, which minimizes registration errors that might result due to slight changes in scan-mirror velocity or aircraft motion. Figure 3.3-18 shows the LW focal plane arrangement.

![Diagram of LW focal plane arrangement]

Figure 3.3-18 Long-wave focal plane arrangement

In order to reduce the background radiation, the filters are mounted on the cold stop which is thermally connected to the cold finger, thus each detector only sees the radiation in the filter bandpass. Besides, each detector has its own cold stop aperture, which can effectively limit the acceptance angle. Figure 3.3-19 shows the cold stop geometry.
3.3.5.B. Short/Mid-Wave Focal Plane*

The SMW focal plane consists of 7 InSb detectors. Two 1-IFOV broad-band (3-5 µm) detectors and four 2-IFOV detectors which cover the spectral region of 1.5-1.75 µm, 2.07-2.35 µm, 3-4.2 µm, and 4.5-5 µm. Figure 3.3-20 shows the SMW focal plane arrangement. Like the LW focal plane, filters are mounted on the cold stop to minimize the background radiation.

* Note that while this focal plane was designed as specified, it has not yet been acquired due to budget limitations.
On the VNIR focal plane (cf. Figure 3.3-20), there are two 1-IFOV broad-band silicon detectors and two 1-mm diameter silica fibers to transmit the optical radiation to the entrance pupils of the spectrometers.

Each imaging spectrometer focal plane consists of a 38-element photodiode array having DIP-type ceramic case with quartz glass window. The array output is fed to the pre-amplifier circuit via a ribbon cable. The length of the cable must be as short as possible to reduce the stray capacitance. The spacing between detector elements is 0.1 mm, which gives a fill factor of 90%, while keeping the cross-talk between the adjacent elements very small. The disadvantage of this array is that the element size (4.4 x 0.9 mm) doesn’t match the aperture size (1.5 mm diameter). The large size of the detector causes a larger junction capacitance. Since custom design is possible, but expensive, we chose to use a stock array for initial design and validation. Figure 3.3-16 shows the imaging spectrometer with the diode-array.

The optical design presented in this section is based on the modular, cost and image quality requirements of the scanner. Except the scan mirror and telescope, everything else could be modified for specific applications. For example, by adding an infrared fiber
on the VNIR focal plane, we can add a short-wave infrared imaging spectrometer. The next section will deal with how to convert the optical energy collected by the optical system into electrical signal.

3.4. Detector/Preamplifier Design

The MISI employs three focal planes resulting in an imaging spectrometer with 13 discrete detectors and two 38-element linear arrays. Each channel has its own video chain which includes detector, preamplifier, and post-amplifier (signal processing). The nominal spectral coverage, spatial resolution, and critical design parameters are outlined in Table 3.4-1. Other design parameters include: cross-talk between adjacent channels <1%, detector/preamplifier rise time < 1 dwell time, and maximum signal overshoot <10%.

Table 3.4-1 MISI Spectral Coverage, Resolution, and Critical Design Parameters

<table>
<thead>
<tr>
<th>Channel</th>
<th>Wavelength (µm)</th>
<th>Spatial Resolution (mrad)</th>
<th>Detector</th>
<th>Number of Detector</th>
<th>Max. Signal Flux (µw)</th>
<th>NETD/NEΔP (1000 ft)</th>
<th>NETD/NEΔP (10,000 ft)</th>
</tr>
</thead>
<tbody>
<tr>
<td>VNIR</td>
<td>0.5-0.9</td>
<td>1</td>
<td>Si</td>
<td>2</td>
<td>1.1</td>
<td>0.1</td>
<td>&lt;0.1</td>
</tr>
<tr>
<td>SWIR 1</td>
<td>1.5-1.75</td>
<td>2</td>
<td>InSb</td>
<td>1</td>
<td>0.55</td>
<td>0.2</td>
<td>&lt;0.1</td>
</tr>
<tr>
<td>SWIR 2</td>
<td>2.07-2.35</td>
<td>2</td>
<td>InSb</td>
<td>1</td>
<td>0.18</td>
<td>0.31</td>
<td>0.1</td>
</tr>
<tr>
<td>MW 1</td>
<td>3-5</td>
<td>1</td>
<td>InSb</td>
<td>2</td>
<td>0.05/320 K</td>
<td>0.39 K</td>
<td>0.12 K</td>
</tr>
<tr>
<td>MW 2</td>
<td>3-4.2</td>
<td>1</td>
<td>InSb</td>
<td>1</td>
<td>0.13/320 K</td>
<td>0.13 K</td>
<td>0.08 K</td>
</tr>
<tr>
<td>MW 3</td>
<td>4.5-5</td>
<td>2</td>
<td>InSb</td>
<td>1</td>
<td>0.06/320 K</td>
<td>0.06 K</td>
<td>0.2 K</td>
</tr>
<tr>
<td>LW 1</td>
<td>8-14</td>
<td>1</td>
<td>HgCdTe</td>
<td>2</td>
<td>1.22/320 K</td>
<td>0.07 K</td>
<td>0.02 K</td>
</tr>
<tr>
<td>LW 2</td>
<td>10-12</td>
<td>2</td>
<td>HgCdTe</td>
<td>1</td>
<td>1.8/320 K</td>
<td>0.08 K</td>
<td>0.03 K</td>
</tr>
<tr>
<td>LW 3</td>
<td>12-14</td>
<td>2</td>
<td>HgCdTe</td>
<td>1</td>
<td>1.38/320 K</td>
<td>0.1 K</td>
<td>0.04 K</td>
</tr>
<tr>
<td>Vis. Spect</td>
<td>0.4-0.7</td>
<td>2</td>
<td>Si Array</td>
<td>38</td>
<td>0.01-0.07</td>
<td>0.3</td>
<td>0.1</td>
</tr>
<tr>
<td>VNIR Spect</td>
<td>0.7-1.0</td>
<td>2</td>
<td>Si Array</td>
<td>38</td>
<td>0.01-0.07</td>
<td>0.3</td>
<td>0.1</td>
</tr>
</tbody>
</table>

3.4.1 Detectors

VNIR Focal Plane

There are only two broad-band Si detectors on the VNIR focal plane. Since the signal level is quite high, any Si detector with 0.5 mm x 0.5 mm geometry can be used.

---

6 Maximum signal flux is based on tabulated data from Electro-optical Handbook [RCA, 1973]

7 NETD computed at 300 K, and NEDr expressed in terms of reflectance units (%). These are predicted values based on expected detector specifications.
SMW Focal Plane

For the short-wave infrared channels, either InGaAs or InSb detectors can be used. InGaAs detectors have a clear advantage of higher detectivity \((D^*=10^{13})\) than that of the InSb detectors \((D^*=10^{11})\). The disadvantage is that the InGaAs detector with extended wavelength range is a relatively new technology with very few tests done in the LN\(_2\) environment. On the other hand, the InSb detector is a very popular short/mid-wave infrared detector. It has been widely used in remote sensing instruments, including the Thematic Mapper. And even with the low detectivity of the InSb detector, the noise equivalent reflectance is on the order of 0.1 reflectance units. For the four mid-wave channels, InSb detector can provide background limited detectivity.

Long-wave Focal Plane

All the long-wave detectors are photoconductive HgCdTe detectors. Spectral characteristics are determined by the composition of this compound and peak response can be varied for various applications by choosing the proper alloy. This property can be used to simplify the spectral filtration of different channels. For the two broad-band, and 12-14 \(\mu m\) channels, the cutoff wavelength of the detector is 14 \(\mu m\). For the 8-10 \(\mu m\) channel, a cutoff wavelength of 10 \(\mu m\) can be chosen, thus eliminating the need for a bandpass filter. And for the 10-12 \(\mu m\) channel, the cutoff wavelength can be 12 \(\mu m\). A 10 \(\mu m\) cut-on low-pass filter can be used instead of a 10-12 \(\mu m\) bandpass. With this method, the three bandpass filters are no longer needed, instead we only need two long pass filters, which are much easier to find than the bandpass filters with the exact cuton and cutoff wavelengths. The detector specifications are listed in Appendix I.

Imaging Spectrometer

There are about 38 channels for each imaging spectrometer. The 130 KHz bandwidth requirement excludes the use of CCD or CID array. The ideal detector would be a 35-element \(1(H) \times 1(W)\) mm array with parallel output. The closest stock match identified are the HAMAMATSU S4111, S4112, S4113, and S4114 arrays developed for the detection of very low light such as in spectrometry. They cover a wide wavelength range with standard linear arrays of 35, 38, and 46 elements. The S4114 series is especially designed for low junction capacitance and high speed response. The disadvantage is that the light sensitive area \((4.4 \text{ mm} (H) \times 0.9 \text{ mm} (W))\) does not match the requirement. This will result in an increase in the junction capacitance which will inevitably increase the
contribution of boosted voltage noise. The S4114 series is used for the two imaging spectrometers. Electrical characteristics of S4114 are listed in Appendix II.

3.4.2 Preamplifier

Two types of photon detector are used in this instrument: photoconductive (PC) and photovoltaic (PV) detectors. They differ in their construction and in the electrical sensing circuits (preamplifiers) required. Photoconductors are poor conductors whose conductivity is improved by the presence of the photon-generated carriers. In order to detect a photon signal, a bias circuit is used to convert the conductance change into a voltage change. Photovoltaic detectors are diodes that generate a photo-current when photons are detected. A transimpedance amplifier (TIA) is used to convert the photocurrent into voltage. TIA offers lowest noise and best linearity under a wide range of conditions. The output voltage signal is the product of the photocurrent and the feedback resistance value.

3.4.2.A. Transimpedance Amplifier

All but the LW detectors are photovoltaic. Since the zero-biased photovoltaic detector behaves like a high-impedance current source, a low-impedance TIA can be used to convert photocurrent to voltage. Figure 3.4-1 shows a typical TIA circuit.

Figure 3.4-1 Transimpedance amplifier for PV detectors

The characteristics of this operational amplifier maintains the diode near 0 V bias. All photocurrent from the detector flows through the feedback resistor (R_f). The feedback
capacitor $C_f$ is added to control the gain peaking caused by the input junction capacitor. The value of $C_f$ depends on the detector junction capacitor. The frequency response of the photodiode/operational-amplifier is determined by the time constant $\tau = R_f \cdot C_f$. The 3 dB cutoff frequency is given by

$$f_c = \frac{1}{2\pi R_f C_f}$$

(3.4.1)

3.4.2. B Photodiode/OP-AMP Noise Spectrum

Like other types of light sensors, the lower limits of light detection for photodiodes are determined by the noise characteristics of the detector and preamplifier. Figure 3.4-2 shows the detector/OP-AMP noise model. Where $R_d$ is the detector shunt resistance, $C_d$ is the junction capacitance, $R_f$ is the feedback resistor, $C_f$ is the feedback capacitor, $e_n$ is the voltage noise of the amplifier, and $i_n$ is the current noise.

![Figure 3.4-2 Noise model of photodiode/op-amp](image)

The effects of various noise sources can be summarized below:

1. **Boosted voltage noise**

   $$e_n^{2\text{ boost}} = (1 + \frac{Z_f^2}{Z_d^2}) e_n^2 H^2$$

   (3.4.2)

   where: $H$ is the transfer function of the electronic filter, and $Z_f$ is the transimpedance gain, which is given by
\[ Z_f = \frac{R_f}{\sqrt{1 + (2\pi f R_f C_f)^2}} \]  
(3.4.3)

\( Z_d \) is the detector reactance, which is given by

\[ Z_d = \frac{R_d}{\sqrt{1 + (2\pi f R_d C_d)^2}} \]  
(3.4.4)

e\(_n\) is the preamplifier input voltage noise, and \( H \) is the transfer function of the post-amplifier.

2. Amplifier input current noise

\[ e_{n\text{ current}}^2 = i_n^2 Z_f^2 H^2 \]  
(3.4.5)

where \( i_n \) is the preamplifier input current noise.

3. Feedback resistor Johnson noise

\[ e_{n\text{ R}}^2 = \frac{4 k T Z_f^2 H^2}{R_f} \]  
(3.4.6)

where \( k \) is Boltzmann's constant and \( T \) is temperature in degrees Kelvin.

4. Detector Johnson noise

\[ e_{n\text{ R}}^2 = \frac{4 k T Z_f^2 H^2}{R_d} \]  
(3.4.7)

5. Background noise

\[ e_{n\text{ background}}^2 = 2 i_b q Z_f^2 H^2 \]  
(3.4.8)

where \( i_b \) is the background current and \( q \) is electron charge \((q=1.6 \times 10^{-19} \text{ Coul})\).
Because of the large bandwidth, 1/f noise is not included here. Also the preamplifier input current noise is very small as compared to the boosted voltage noise and background noise.

The total noise can be approximated by

\[
\text{Total } e_n(f) \approx \left[ \frac{Z_f^2}{Z_d} e_n^2 + i n Z_f^2 + \frac{4kT Z_f^2}{R_d} + \frac{4kT Z_f^2}{R_f} \right]^{1/2} H
\]

(3.4.9)

Figure 3.4-3 shows the spectral noise components for three background levels: (1) 300 K background with 60 degrees field of view (background current \(i_b=7 \times 10^{-6} \text{ A}\)), (2) ideal case (no background radiation), and (3) with cold spectral filter and 10 degree field stop \((i_b=4 \times 10^{-7} \text{ A})\). The background noise is calculated using Eq. 3.4.8. The parameters used here are as follows:

- \(R_f=300 \text{ K}\Omega\)
- \(C_f=5.3 \text{ pF}\)
- \(R_d=250 \text{ K}\Omega\)
- \(C_d=400 \text{ pF} \quad \text{(InSb detector)}\)
- \(T=300 \text{ K}\)
- \(T_{\text{det}}=77 \text{ K}\)
- \(e_n=4.5 \text{ nv/}\sqrt{\text{Hz}} \quad \text{(OPA-627)}\)

Figure 3.4-3(a) \(i_{bn}=1.5\text{pA/Hz}^{1/2}\)
The first graph shows that without the cold stop, the detector is background limited throughout the spectrum. With the cold stop, the noise is limited by the background noise and Johnson noise at low frequencies, and the boosted $e_n$ noise at high frequencies.

In a low signal condition (Figure 3.4-3b), the resistor noise limits the low-frequency noise. In either case, the wide band noise is limited by the boosted $e_n$ noise.
Since the signal-to-noise ratio is proportional to \( \sqrt{R_f} \) at low frequency, it would be tempting to select a large \( R_f \) value. But large resistance would demand a small feedback capacitance (\( C_f \)) which might cause the preamplifier to be unstable. Figure 3.4-4 shows the equivalent input noise current as a function of feedback resistance. As the feedback resistance increases from 5 K\( \Omega \) to 100 K\( \Omega \), the noise current decreases considerably. But further increase of \( R_f \) would cause only a slight decrease in the noise current, the detector is boosted voltage noise limited. So \( R_f \) should take the value between 100 - 500 K\( \Omega \).

![Graph showing noise current as a function of feedback resistance](image)

Figure 3.4-4 Noise current as a function of feedback resistance

The above analysis is valid for the four 2-mrad InSb detectors which have similar electrical characteristics. But for the two high resolution 1-mrad broad-band detectors, the electrical characteristics are different. Most notably, the shunt resistance is about four times larger and the junction capacitor is about a quarter of that of the 2-mrad detectors. The parameters used for the 1-mrad InSb detector are as follows:

\[
\begin{align*}
R_f &= 500 \text{ K}\Omega \\
R_d &= 1000 \text{ K}\Omega \\
R_{_{RF}} &= 300 \text{ K} \\
C_f &= 1.38 \text{ pF} \\
C_d &= 100 \text{ pF (InSb detector)} \\
T_{_{RF}} &= 300 \text{ K} \\
T_{det} &= 77 \text{ K} \\
e_n &= 4.5 \text{ nV/} \sqrt{\text{Hz (OPA-627)}} \\
i_b &= 1.85 \times 10^{-7} \text{ A (background current)}
\end{align*}
\]

Figure 3.4-5 shows the noise spectrum of the 1-mrad InSb detectors. The graphs show that at low frequencies the noise is limited by the background noise, at high frequencies
it is limited by the boosted $e_n$ noise. The wide-band noise is limited by both the boosted voltage noise and background noise.

![Graph of noise spectrum vs. frequency.](image)

Figure 3.4-5 Noise spectrum of 1-mrad InSb detector

The relationship between the noise current and feedback resistance is plotted in Figure 3.4-6. The upper curve is without the background noise, the lower curve is with the background noise. In either case, the noise current would not increase significantly if the feedback resistance is greater than 200 KΩ. The best feedback resistance value is between 200 KΩ and 1 MΩ.

![Graph of noise current vs. feedback resistance.](image)

Figure 3.4-6 Noise current vs. feedback resistance

The preamplifier for the two panchromatic channels (VNIR) is very simple due to the high signal level and high performance of the silicon detectors. Figure 3.4-7 shows the
noise spectrum of the panchromatic Si channels. The noise is completely signal-limited throughout the spectrum.

Figure 3.4-7 Noise Spectrum of Panchromatic Si Channel ($R_f=1$ MΩ, $C_f=0.5$ pF)

The pre-amplification for the Si diode array turns out to be the most difficult one because of low signal level (photocurrent of the order of $10^{-9}$ A). Figure 3.4-8 shows the noise spectrum for the Si diode array.

Figure 3.4-8 Noise spectrum of Si diode-array

The graph indicates that the feedback resistor noise ($R_f=5$ MΩ) dominates most of the spectrum except at the high frequency, which is dominated by the boosted voltage noise. The detector noise is very low so that it is definitely not detector limited performance.
The background noise is computed based on a nominal background flux of approximately 3x10^{-9}W on the detector which represents the flux from a 10% reflector. The wide band noise is limited by the feedback resistor noise. The boosted voltage noise will be higher than that is predicted here in actual circuit because the stray capacitance of the leads and the input capacitance of the amplifier will add to the junction capacitance.

Figure 3.4-9 Noise current vs. feedback resistance for Si diode array

Figure 3.4-9 shows the relationship between the equivalent input noise current and feedback resistor. The optimal resistor value would be 10 MΩ, but this resistor would require the feedback capacitor to be as small as 0.1 pF to keep the bandwidth of 115 KHz. This capacitor is too small to compensate for the gain peaking effect. So the design parameters for the Si-diode array are R_f=5 MΩ and C_f=0.25 pF.

Table 3.4-2 summarizes the parameters used for the four types of transimpedance amplifier.

---

8 The feedback resistance can be larger if the scanner is operated at lower scan rate.
Table 3.4-2 Critical Design Parameters for the TIA

<table>
<thead>
<tr>
<th>Band</th>
<th>Panchromatic Si (1-mrad)</th>
<th>InSb (2-mrad)</th>
<th>InSb (1-mrad)</th>
<th>Si-diode (2-mrad)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_d$</td>
<td>2 GΩ</td>
<td>250 KΩ</td>
<td>1 MΩ</td>
<td>2 GΩ</td>
</tr>
<tr>
<td>$C_d$ (PF)</td>
<td>15</td>
<td>400</td>
<td>100</td>
<td>35</td>
</tr>
<tr>
<td>$R_f$</td>
<td>1 MΩ</td>
<td>300 KΩ</td>
<td>500 KΩ</td>
<td>5 MΩ</td>
</tr>
<tr>
<td>$C_f$ (PF)</td>
<td>0.7</td>
<td>5</td>
<td>1.5</td>
<td>0.25</td>
</tr>
<tr>
<td>OP-AMP</td>
<td>OPA102</td>
<td>OPA627</td>
<td>OPA627</td>
<td>OPA102</td>
</tr>
<tr>
<td>Background rad. (W)</td>
<td>1.1E-06</td>
<td>2E-7</td>
<td>6E-8</td>
<td>6E-9</td>
</tr>
<tr>
<td>Background cur. (A)</td>
<td>5E-7</td>
<td>4E-7</td>
<td>1.8E-7</td>
<td>3E-9</td>
</tr>
<tr>
<td>$i_n$ (Boosted) (PA)</td>
<td>25</td>
<td>351</td>
<td>121</td>
<td>8</td>
</tr>
<tr>
<td>$i_n$ (Rf) (PA)</td>
<td>62</td>
<td>80</td>
<td>87</td>
<td>19</td>
</tr>
<tr>
<td>$i_n$ (Rd) (PA)</td>
<td>1.5</td>
<td>44</td>
<td>31</td>
<td>1</td>
</tr>
<tr>
<td>$i_n$ (background) (PA)</td>
<td>182</td>
<td>122</td>
<td>115</td>
<td>10</td>
</tr>
<tr>
<td>Total noise (PA)</td>
<td>194</td>
<td>383</td>
<td>190</td>
<td>21</td>
</tr>
<tr>
<td>Detector noise (PA)</td>
<td>2</td>
<td>508$^9$</td>
<td>384</td>
<td>2</td>
</tr>
<tr>
<td>Measured noise (PA)</td>
<td>&lt;100</td>
<td>NA$^{10}$</td>
<td>NA</td>
<td>15 ($\Delta f=60$kHz)</td>
</tr>
</tbody>
</table>

3.4.2.C. The Voltage Amplifier

Since the PC detectors are low-impedance detectors ($\approx 150 \, \Omega$), there is no need for front-end buffering. They can be modeled as photoresistors. Since the resistance value typically ranges from 20 to 150 $\Omega$, there are not many semiconductors with noise levels adequate for such low resistance.

For a typical HgCdTe detector, the noise equivalent power is given by

$$NEP = \frac{\sqrt{\Delta f}}{D^*} = \frac{\sqrt{0.05*0.05*230000}}{4E10} = 6E - 10(W)$$

(3.4.10)

For a biasing current of 3 mA, the responsivity ($R$) is about 3000 V/W. The noise equivalent voltage of the detector is

$$NEV=NEP*R=1.8E-6 \, V$$

(3.4.11)

$^9$ Note that for the InSb detector, the detector noise is greater than the detector/OP-AMP combination noise. This is because the InSb detector is background limited, the detector noise is determined with 60° FOV. In application, a cold stop is used to limit the FOV to close to 10°

$^{10}$ The InSb detector has not been acquired, thus it was not tested.
Assuming the detector noise is white noise, the noise density is about 3.75 nV/√Hz. Figure 3.4-9 shows a circuit schematic for the HgCdTe detector. A current source, which consists of a Zener diode and two bias resistors ($R_b$), provides current for both the detector leg of the amplifier and the reference leg. An ultra-low noise, high precision monolithic instrument amplifier (IN103) was used in the first stage. The values of $R_f$ and $R_g$ are very low to reduce the resistor noise. At room temperature, the resistance noise is given by

$$e_n = 1.3 \times 10^{-10} \sqrt{R} \quad (V/\text{Hz}^{1/2})$$  

(3.4.12)

![Circuit schematic for the HgCdTe detector](image)

Figure 3.4-10  Preamplifier for the HgCdTe detector

The IN103 has a fixed gain of 100. The output form IN103 is buffered to increase its ability to drive the signal conditioning circuit as well as to provide better protection against noise. The output signal from the pre-amplifier is given by

$$V_o = 100(V_a - V_b)$$  

(3.4.13)

where $V_a$ and $V_b$ are the voltage level at the input of INA103.

Biased by the current source, the detector can be modeled as a variable voltage source. The voltage $V_a$ is given by
\[ V_s = i_s R_d \]  \hspace{1cm} (3.4.14)

The photon-induced signal is given by

\[ \Delta V_s = i_s \Delta R_s \]  \hspace{1cm} (3.4.15)

In order to prevent preamplifier saturation due to the offset generated by the detector bias, detector temperature change, or even background change, a feedback loop (correlated sampling) is introduced into the circuit. When the scanner looks at a blackbody, the analog switch is closed. The sample-and-hold circuit samples the output of the amplifier and holds its value to provide 0.1% of the output signal to the reference leg of the amplifier. That is,

\[ \Delta V_s = 0.01V_o = 0.01 \times 100(V_a - V_b) = V_a - V_b \]  \hspace{1cm} (3.4.16)

rearranging yields

\[ V_b' = V_b + \Delta V_b = V_a \]  \hspace{1cm} (3.4.17)

and the output signal becomes

\[ V_0 = 100(V_a - V_b) = 0 \]  \hspace{1cm} (3.4.18)

The output signal was forced to zero for this blackbody. But for the other blackbody and the scene, the output signal is given by

\[ V_0 = 100(V_a - V_b) = 100(V_{in} - V_{blackbody}) \]  \hspace{1cm} (3.4.19)

Not only does this feedback loop provides DC restoration, but it also provides low-pass filtering of the signal. Since in the time domain, Eq. 3.4.19 is equivalent to

\[ V_0(t) = 100V_{in}(t) * [\delta(t) - \delta(t - \tau)] \]  \hspace{1cm} (3.4.20)

where * denotes convolution operation and \( \tau \) is the time interval between the blackbody and the scene. For this scanner, \( \tau \) varies from \( T/4 \) to \( T/2 \) (\( T \) is the period of the scan, for the flight height of 1000 ft, \( T = 0.0125 \) second). In the frequency domain, the transfer function has a sinusoid form

\[ T(f) = \sin(2\pi ft + \phi) \]  \hspace{1cm} (3.4.21)
Figure 3.4-11 shows the frequency response for $\tau = 0.005$ second. The frequency response is very low at the low frequency where the $1/f$ noise of the detector and amplifier is dominant. This method will certainly eliminate the effect of bias drift, detector temperature change, and moreover, will greatly reduce the $1/f$ noise.

![Transfer function of the clamping circuit (correlate sampling)](image)

Figure 3.4-11  Transfer function of the clamping circuit (correlate sampling)

The circuit in Figure 3.4-10 can be simplified by removing the two analog switches as shown in Appendix III (This is equivalent to closing both switches in Figure 3.4-10). The circuit becomes a closed loop integrator where the S/H circuit integrate the response of the reference blackbody and the output of the integrator is fed back to the reference leg of the bridge to cancel the drifting. Although this circuit cannot always clamp the output to zero when it sees the reference blackbody, the correlated sampling scheme described above can still be used in software.

The input noise density of this amplifier is about $3 \, \text{nv}/\sqrt{\text{Hz}}$, which is slightly smaller than the detector noise. It is expected that this detector/preamplifier will deliver detector limited performance.

### 3.5 Conditioning Signal Processing

The purpose of the signal-conditioning subsystem is to establish the noise/signal bandwidth, which determines the signal-to-noise ratio (SNR) of each. Since the spatial variation on the ground is converted into the temporal variation in the electrical signal, the temporal characteristics also determine the sensor's spatial response. There are at least two competing goals in determining the bandwidth: (1) to provide the best possible
estimate of the source radiometric flux in the presence of noise and (2) to preserve the spatial details. The first goal requires that the noise bandwidth be as narrow as possible to reduce the noise, while the second goal requires a wide bandwidth to preserve the spatial details, such as lines and edges.

The situation is further complicated by the sampling process. If the signal being sampled contains information with frequency content greater than half the sampling frequency, the resulting sampled signal carries spurious information (aliasing), which in effect acts as noise. The conditioning electronics must reject any frequency component greater than the Nyquist frequency.

The electronics must provide enough gain to yield useful output levels and operate over a dynamic range sufficient to provide linear response over the expected flux levels. The electronics must also provide various sync signals and human interfacing for scanner operation. Fig 3.5-1 shows a functional diagram of conditioning electronics.

3.5.1 Information Bandwidth

The integration time (or so called dwell time) is the time it takes for a detector to scan across an IFOV on the ground. The integration time $T_d$ is $2.17 \mu s$ for the 1 mrad channels and $4.34 \mu s$ for the 2 mrad channels (cf. Section 3.1). Conventionally (Seyrafi 1984), the electrical bandwidth is chosen to be

$$f = \frac{1}{2T_d} = 230 \text{ kHz} \quad (3.5.1)$$

The time constant of the electrical system is given by
The electrical system response to a step function is given by

\[ f(t) = 100[1 - \exp(-t/T_c)] \]  

\[ T_c = \frac{1}{2 \pi f} = 0.7 \mu s \]  

Besides electronics, the optics and detector also affect the spatial response of the sensor. Fig. 3.5-2 shows the sensor scan direction linear model. The line spread function of the system in the scan direction (prior to sampling) \( \text{LSF}_x(x) \) is described by the convolution of the line-spread functions of the three components:

\[ \text{lsf}_x(x) = \text{OSF}(x) \ast \text{DSF}(x) \ast \text{ESF}(x) \]  

where \( \text{OSF}(x) \) is the optics line-spread function, \( \text{DSF}(x) \) is the detector aperture function, \( \text{ESF}(x) \) is the electronic impulse response function, and \( \ast \) denotes the convolution operation. In the frequency domain, the scan-direction system-transfer function is given by

\[ \text{TF}_x(f_x) = \text{OTF}_x(f_x) \ast \text{DTF}_x(f_x) \ast \text{ETF}_x(f_x) \]  

where \( \text{OTF}_x(f_x) \), \( \text{DTF}_x(f_x) \), and \( \text{ETF}_x(f_x) \) are the optical transfer function, detector transfer function, and electronic frequency response function, and \( f_x \) is the spatial frequency expressed in the unit of cycles per milliradian (cycles/mrad).

The OTF of the telescope (on-axis) can be estimated using a regression method on the measured MTF of the telescope (cf. Section 4.2.2)

\[ \text{OTF}(f_x) = 1.00 - 0.0006 f_x - 0.00067 f_x^2 \]
The detector transfer function is given by a sinc function

\[ DTF = \frac{\sin(b \pi f)}{b \pi f} \]  

(3.5.7)

where \( b \) is the width of the detector. Suppose the electronic filter is a one-pole Butterworth filter (simple RC filter), the electronic transfer function is given by

\[ ETF(f) = \frac{1}{\sqrt{1+(f/f_n)^2}} \]  

(3.5.8)

where \( f_n \) is the electronic cutoff frequency.

These three transfer functions are plotted in Fig 3.5-3 for the high resolution channels. Since theIFOV’s are 1 mrad for the high resolution channels and 2 mrad for the low resolution channels, the Nyquist frequency is 0.5 cycle/mrad and 0.25 cycle/mrad respectively.

\[ \text{Figure 3.5-3 MISO scan direction MTF with one-pole low-pass filter} \]

It is evident that there is significant frequency response beyond the Nyquist frequency. If the signal is not band-limited at the Nyquist frequency, there will be aliasing which acts like noise. In order to reduce the aliasing noise, a high-order electronic filter must be employed to filter out the high frequency content. Fig. 3.5-4 shows the MTF of MISO using a 3-pole Butterworth low-pass filter. It is obvious that the high frequency content
is greatly reduced.

![Graph showing OTF, DTF, ETF', and TF(total) with Nyquist frequency](image)

**Figure 3.5-4 MISI scan direction MTF with three-pole low-pass filter**

### 3.5.4 Synchronization

Synchronization is very important for the scanner operation. It tells when the start of the scan line occurs and when the blackbodies are visible. The sync signal is also used as a trigger for the sample-and-hold circuit in the A/D converters. The sync signal is derived from the scan mirror assembly where an optical shaft encoder was installed on the scan motor shaft to provide 1 milliradian angular resolution sync pulse. The encoder also provides a line sync pulse which will be used as the start of line signal.

### 3.7. Radiometric Calibration

Absolute calibration is essential for a variety of scientific studies and image analysis applications. For example, the removal of atmospheric effects for building heat loss and water quality studies require absolute radiometric data. To extend signatures extracted from a particular scene to data collected in different scenes or at different times under different atmospheric conditions, or even to data collected by different sensors, requires both an absolute radiance and atmospheric correction.

Two types of calibration procedures are used to calibrate MISI absolutely: (1) onboard calibration to maintain the instrument stability, and (2) laboratory calibration to maintain the absolute accuracy and traceability to the NIST.
3.7.1 Onboard Calibration

Due to the detector gain and offset fluctuation with temperature and background radiation level, onboard calibration is necessary to provide reference level for every scan line. Onboard calibration also provides DC restoration for the thermal infrared channels (cf. 3.4).

3.7.1.A. Thermal Channels

Onboard calibration for the thermal channels is more important than for the reflective channels, because both the gain and offset change with the temperature of the focal plane, the temperature of optical elements, and the temperature of the instrument housing.

There are two methods of onboard calibration, one method is using an internal calibrator similar to the one used by TM. An oscillating flag passes the focal plane every scan line. A mirror is mounted on the head of the flag which reflects the radiation from a stationary blackbody to the detector. The second method is to place two blackbodies above the scan mirror assembly. For every scan line, the scan mirror, which is also the entrance aperture of the optics, is flooded by the radiation from the blackbodies. The second method represents a radiometric calibration of the complete optical train of the system. Since two blackbodies are used, both the gain and offset of each channel are calibrated for every scan line.

The second calibration method is used for MISI because of the high calibration accuracy. Figure 3.7-1 shows the schematic of scan mirror and two blackbodies. The blackbodies will be full aperture devices. Both will be thermally controlled, (one may be heat only) and have temperature measurement thermocouples in addition to the controller for accurate monitoring of the surface temperatures. The blackbodies are located well up into the scanner to minimize air flow across the surface which can introduce calibration problems due to high thermal gradients at the surface of the blackbody.

For each channel, the relationship between the radiance (L) and output digital count (DC) is given by

\[ DC = G\cdot L + O \]  \hspace{1cm} (3.7.1)

where G is the gain and O is the offset. Applying Equation 3.7.1 to the two blackbodies, we have
\[ DC_1 = G \cdot L_b(T_1) + O \]
\[ DC_2 = G \cdot L_b(T_2) + O \]  
(3.7.2)

Figure 3.7-1 Schematic of onboard blackbodies

where \( T_1 \) and \( T_2 \) are temperatures of the blackbodies, \( L_b(T_1) \) and \( L_b(T_2) \) are the radiances of from the blackbodies, which are given by the Planck's equation. Solving Equation 3.7.2 for \( G \) and \( O \) yields

\[ G = \frac{DC_2 - DC_1}{L_b(T_2) - L_b(T_1)} \]  
(3.7.3a)

\[ O = DC_1 - \frac{DC_2 - DC_1}{L_b(T_2) - L_b(T_1)} L_b(T_1) \]  
(3.7.3b)

With the calibrated gain and offset value, the absolute radiance reaching the entrance aperture can be determined using Equation 3.7.1

3.7.1.B Reflective Channels

Both the gain and offset of the reflective channels do not change as much as the thermal channels, but they still change with time during the operation. For example, the dark current of the photodiode and the offset voltage of the pre-amplifier change with temperature, the variation of the power voltage will certainly affect the output signal, and stray light will affect the offset value.

Since the changes in gain and offset are small and the fluctuation is slow, it is possible to use only one calibration source with variable output radiance. One way of changing the
light level is to vary the electrical current flow through the calibration lamp. But this method is susceptible to instability and introduces color temperature change which complicates the calibration process. The second method is to use a filter wheel with eight apertures as shown in Figure 3.7-2. Since the area of the aperture can be accurately measured, the output radiance can be stepped precisely through eight levels.

Figure 3.7-2 shows a configuration for reflective channel calibration. Light from a lamp passes through an aperture on a filter wheel and illuminates the integrated cylinder. The amount of light passes through the aperture is determined by the size of the aperture. With eight apertures on the filter wheel, the calibration system can provide eight intensity levels. The lamp power supply must be carefully regulated to ensure the stability. Optical filters can be inserted into the optical path to spectrally balance the energy to better simulate the reflected solar spectra. An opening slit on the bottom of the integrated cylinder illuminates the scan-mirror.

![Figure 3.7-2 Onboard calibration for the reflective channels](image)

The relationship between the radiance and the digital count is again given by Equation 3.7.1. Since more than two radiance levels are used, the gain and offset values can be better estimated using a linear regression method, where the independent variable is
radiance and the dependent variable is the digital count. The slope term is the gain and the intercept term is the offset.

3.7.2 Laboratory Calibration

Both the thermal and reflectance channel will require periodic calibration of the on board calibrators. Special calibrators and calibration procedures will be developed for this purpose. The thermal channel calibration will use instrumentation that is already in-house and expand on established calibration procedures used with RIT’s current IR line scanner. The instrumentation for reflective channel calibration was developed based on a 20 inch integration sphere [Ganiear, 1995].

3.7.2.A. Laboratory Standards and Procedures for the Thermal Channels

An 8”-square thermally controlled blackbody will be used for cross-calibration of the on board blackbodies. The CI system’s blackbody is controllable from -10° to 100° C and has absolute temperature calibration accuracy of .01 C. The surface has an emissivity of 0.97 ± .02. With the scanner at fixed gain and bias, the laboratory BB will be imaged at a series of temperature settings under ambient background conditions. The effective within-band radiance for each band will be computed from the Planck equation:

\[ L = \int_{0}^{\infty} \beta(\lambda) (\varepsilon L_{T\lambda} + r L_{B\lambda}) d\lambda \]  (3.7.4)

where \( L \) is the within band radiance, \( \beta(\lambda) \) is the normalized spectral response of the system for the channel of interest, \( \varepsilon \) and \( r \) are the emissivity and reflectivity of the blackbody and \( L_{T\lambda} \) and \( L_{B\lambda} \) are the spectral radiance values from the Planck equation for the BB temperature \( T \) and the background temperature \( B \).

A relation between observed digital count and radiance will then be established. The two on board calibrators will then be calibrated by setting their controllers, reading out a digital count for the setting value, and converting the digital count to radiance using the relationship for the laboratory standard. A relationship between control setting value and radiance can then be established for each on board blackbody. In practice, the on board setting values will generally be expressed as the nominal temperature values of the BB’s to facilitate intuitive use. This provides the nominal calibration of the onboard calibrators.
This entire process will be repeated with the ambient environment approximately 20K below the previous operation. This can be easily accomplished in the North Range Lab by moving the instrumentation outdoors during the winter or on a cool night. An environmental shroud will be used to minimize excessive air flow over the laboratory blackbody and to simplify characterization of the background temperature. If the results of the calibrations at the two temperatures show any significant difference, then a series of experiments to establish the degree of dependence on ambient temperature will be run and the internal temperature of the scanner will need to be monitored.

For the MWIR channels, the entire calibration will be conducted in a darkened environment. For the initial test, a full daylight environment will also be used to determine if any stray light problems exist.

Additional one-time tests will include effects of power supply voltage level on calibration and tests to ensure that temperature extremes in the onboard calibrators do not change background calibration effects.

3.7.2.B. Laboratory Standards and Procedures for the Reflectance Channels

The laboratory standard for the reflective channels will be a 20" integrating hemisphere with a 8" illumination port modeled after the NASA system described by Barker et al. [Ganiear, 1995]. The integrating sphere is illuminated internally by four tungsten-halogen lamps. The output radiance characteristics of the sphere were quantified with the use of a detector calibrated to a NIST standard lamp. The spectral characteristics of the hemisphere at each setting will be calibrated by illuminating a spectrometer calibrated using NIST traceable lamp standards using an approach similar to Markham et al. The absolute calibration of the hemisphere will use detector standards to monitor the absolute level and the stability of the radiance exiting the hemisphere. These radiance standard values will be transferred to the on-board calibrators using procedures similar to those used in the Thematic Mapper calibration as described by Barker et al. (1985b). The spectral sensitivity of the MISI will be evaluated using an 8" collimator and a monochromator. The polarization sensitivity will be evaluated for each channel by placing a polarizer over the collimator to maximize throughput from a tungsten halogen lamp and then rotating the entire collimator while illuminating the MISI. The internal calibrators will be calibrated by comparison with the hemisphere when it floods the scan mirror. In scan mode the MISI will image the hemisphere and the internal calibrator as both are stepped through their power levels. A digital count versus effective in-band
radiance relation will be developed for the hemisphere based on its known radiance and spectral character. From this relationship, the digital count for each level of the on-board calibrators can be converted to in-band radiance.

The stability of the calibration to variations in aircraft power levels and ambient flux levels will be evaluated by varying these levels over expected ranges. Calibration stability will be tested by regular repetitions of the calibration procedures at selected levels.
4. Performance Evaluation

4.1 Overview of MISI Evaluation

MISI was built based on the design in Section 3. The telescope was fabricated first. Test result shows very good agreement with model prediction. The mechanical fixture used to hold the two mirrors could not be fitted into MISI, thus the fixture was redesigned and fabricated. As expected the scan-mirror was the most costly and troublesome part of the instrument. Our design was compromised because of the difficulties in fabricating the shell structure (Figure 3.3-6). The casting process requires a 1.5 degree drafting angle which added mass to the tip of the mirror; this is the worst place to add mass. First-order analysis indicated that this compromise would not degrade the image quality very much, but it did require more accurate balance. After several months of work with vendors, the scan-mirror was balanced in November of 1994. The image quality of the mirror was found to be unacceptable due to a larger amount of aberration caused by stress in the casting process. It was refinished by Optical Technologies who fabricated and assembled the scan-mirror. Due to the budget constraints, the mid-wave focal plane has not been acquired. It will be added as a upgrade to the baseline system in the future. The high speed data acquisition and recording system was acquired, but it is still being tested stage. The entire MISI system was assembled in March of 1995. Figure 4.1-1 shows a photograph of the assembled MISI. Since then, extensive tests were performed to evaluate its performance against the design specifications. This section will briefly describe how each subsystem was evaluated to confirm that its performance meets specifications. This analysis complements the image chain analysis by allowing the system engineer to evaluate the results of actual hardware testing on theoretical design criteria.
4.2 Optical Subsystem Validation

4.2.1 Scan-mirror

Because the scan-mirror spins at a very high speed, the scan-mirror had to be tested both statically and dynamically. In the static test, the mirror surface profile was measured by a Flatmaster200XR at GCA Tropel. Figure 4.2-1 shows the measured surface contour. The static deformation ranges from -3 μm to +0.8 μm. The deformation data were converted to wavefront aberration, and the blur (PSF) was calculated according to Eq. 3.3.7. Figure 4.2-2 shows calculated point spread function. The PSF is not diffraction limited due to static
deformation, but the blur is well within one IFOV of the MISI thus it will not cause significant MTF degradation.

Figure 4.2-1 The static deformation as measured by GCA Tropel

Figure 4.2-2 The PSF of the Scan-mirror Calculated from Static Deformation

Ideally, we could measure the dynamic deformation of the mirror surface using interference methods. However, the scan-mirror rotates at such a high speed that it is impractical to record the interference pattern. Since the deformation causes image quality degradation, the line spread function caused by the dynamic deformation of the mirror surface can be measured. Figure 4.2-3 shows a schematic of the optical layout to measure the LSF of the scan-mirror. The light from the laser was collimated by a 6 inch collimator. The
Collimated light illuminated the scan-mirror. The reflected light from the scan-mirror was collected by the second collimator (MISI telescope), and focused onto a linear array CCD camera. A digital oscilloscope was used to record and display the CCD video output signal.

Figure 4.2-3 Schematic for measuring the LSF of the scan-mirror

When the scan-mirror rotates, the image of the pinhole scans across the CCD array. Since the integration time of the CCD is much longer than the time required to scan the pinhole image across the CCD, the signal on the CCD is given by

\[ s(x) = \int_{-\infty}^{\infty} h(x, y) \, dy \]  

(4.2.1)

where \( h(x, y) \) is the image of the pinhole. Since the pinhole is very small, \( h(x, y) \) is approximately the point spread function of the optical system, and \( s(x) \) is actually the line spread function. By measuring the LSF of the scan-mirror at low speed (without deformation) and high speed (with deformation), it is possible to calculate the blur due to scan-mirror deformation. This method was successfully used in the testing of the Bendix mirror (Feng et al. 1994). The LSF of the scan-mirror at low speed (10 Hz) and high speed (76 Hz) were measured. The difference between the two LSFs is very small as compared the LSF itself. This test was done before the mirror was repolished. The static LSF is
about one IFOV, thus the blur caused by mirror dynamic deformation must be less than one IFOV. Since the dynamic deformation is proportional to the square of scan rate, the blur caused by dynamic deformation at the scan rate of 40 Hz will be very small as compared to the detector size. Since the centrifugal force is always point outward from the rotation axis, the effects of temperature and vibration do not cause any degradation in image quality in normal operation.

4.2.2. Telescope

The telescope system was fabricated according to the design specification (cf. Section 3.3). The image quality of the telescope was tested using the configuration shown in Figure 4.2-4. Light from the laser was focused onto a 10 μm pinhole, which was located on the focal point of a 2.25 meter F/15 collimator. The collimated beam filled the aperture of the telescope and a point image was formed on the focal plane of the telescope. This point image was re-imaged onto a 384 x 490 CCD array using a 40X microscope objective. The magnified image was display on a video monitor.

![Figure 4.2-4 Configuration to measure the image quality of the telescope system](image)

Since the pin-hole size is relatively small (2.2 μm on the telescope focal plane) and the on-axis image qualities of the collimator and the microscope objective are better than that of the telescope, the image recorded on the CCD array is approximately the point spread function of the telescope. Figure 4.2-5 shows the on-axis PSF of the telescope.
The PSF is similar to the Airy disk of a diffraction-limited optical system. The diameter of the first dark ring is about 13 pixels horizontally and 15 pixels vertically. Since the size of the CCD pixel is 17.25 μm wide by 14.25 μm high, the diameter of the dark ring is estimated to be 5.6 μm horizontally and 5.3 μm vertically. The diameter of the Airy disk is given by Equation 3.1-9.

\[ \delta = 2 \frac{1.22D_0}{D} = 5.15 \mu m \]  

(4.2.2)

This result indicates that the telescope is nearly diffraction limited on-axis as designed.

The off-axis image quality was measured by tilting the telescope a small angle relative to the incident beam. The PSF of the telescope was measured at off-axis angles of 10 milliradian and 20 milliradian as shown in Figure 4.2-6 and 4.2-7. It can be seen that both the size and shape of the measured PSF at 20 milliradian off-axis (cf. Figure 4.2-7) is very similar to the spot diagram predicted by the optical design software (cf. Figure 3.3-11).
The MTF of the telescope was computed by the following procedures: (1) project the PSF onto the x-axis (horizontal axis) and y-axis (vertical axis) to calculate the line spread function (LSF) along the x and y axis respectively, (2) compute a discrete Fourier transform
of the LSFs, (3) compute the modulus of the transform and normalize to unity at zero frequency. Figure 4.2-8 shows the measured MTF for (a) on-axis, (c) 10 milliradians off-axis.

Figure 4.2-8(a)

Figure 4.2-8(b)
The measured MTFs are lower than the ones predicted by the optical design software. This difference is mainly caused by the non uniform background in the images captured by the CCD camera. Although digital image processing can improve MTF calculation, it is not necessary to do so since even the MTF at 30 mrad off axis is much better than the MTF of the detector. Figure 4.2-8(c) shows the MTF due just to sampling by the 1 milliradian detector. The system MTF is not telescope limited.

4.3 Detector and Conditioning Electronics Validation

The signal-to-noise ratio is the most important figure of merit in the analog imaging chain. The SNRs must be measured for every channel to determine whether the design specifications are met. Figure 4.3-1 shows a setup used to measure the SNR. The signal from the detector is amplified by the electronics and fed into the spectrum analyzer to determine the spectra of signal and noise. Since the signal from the broadband Si detector is very strong, effort was concentrated on testing the photodiode array used in the imaging spectrometer and the MCT detectors used in the long-wave focal plane.
4.3.1 Testing of the Photodiode Array

The photodiode array was tested using the pre-amplifier circuit shown in Figure 3.4-1. A Burr-Brown OPA102 operational amplifier was used because it has very low input capacitance and noise. Both the pre-amplifier and the detector were placed in a shielded metal box to prevent RF interference. Figure 4.3-2 shows a measured noise density and the predicted noise density based on Eq. 3.4.9 with $R_f = 3.9 \, \text{M} \Omega$, $C_f = 0.5 \, \text{pF}$, $C_d = 24 \, \text{pF}$, and filter cutoff frequency $f_c = 130 \, \text{kHz}$. The theoretical predicted equivalent input noise current is about 25 pA. The predicted noise number can be reduced by half to 13 pA if the instrument is operated at a scan rate of 40 Hz ($R_f = 10 \, \text{M} \Omega$, $C_f = 0.33$, $f_c = 50 \, \text{KHz}$). To achieve the predicted noise figure, extreme care must be taken to reduce noise sources from other equipment and a high-order lowpass filter must be used to reject the high frequency noise. For the experiment described here, the measured noise is about twice the predicted (50 pA). It is believed that most of the extra noise was due to inadequate filtering (only first-order low-pass filters was used in the test).
4.3.2 MISI Long-wave Detector (MCT) Testing

MISI design calls for the detectivity to exceed than $4 \times 10^{10}$ (W.cm/$\sqrt{Hz}$); which will yield less than 0.1K NEDT for most channels. There are two high-resolution (1 mrad) broadband channels and 3 low-resolution (2 mrad) narrow band channels cover approximately 8-10μm, 10-12μm and 12-14μm wavelength regions.

4.3.2.2 Manufacture's specifications

The long-wave (LW) focal plane was fabricated by Belov Technologies. Due to the limited selection of stock filters, the spectral regions for each detector were not exactly as designed. Nevertheless, they cover the spectral regions of greatest interest. The final specifications for the MCT detectors are listed in Appendix II. Table 4.3-1 is a summary of these detectors.

Table 4.3-1 MCT Detector Specifications Provided by Belov Technologies

<table>
<thead>
<tr>
<th>Spectral region</th>
<th>Detector 1</th>
<th>Detector 2</th>
<th>Detector 3</th>
<th>Detector 4</th>
<th>Detector 5</th>
</tr>
</thead>
<tbody>
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<td>9.7-11.3</td>
<td>8.0-13.9</td>
<td>9.0-13.7</td>
<td>8.3-9.9</td>
<td>11.7-14.4</td>
<td></td>
</tr>
<tr>
<td>8 E+10</td>
<td>6.2 E+10</td>
<td>6.2 E+10</td>
<td>7.0 E+10</td>
<td>5.2 E+10</td>
<td></td>
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<td>6000</td>
<td>21000</td>
<td>6400</td>
<td>1700</td>
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<td>126.4</td>
<td>114.8</td>
<td>160.7</td>
<td></td>
</tr>
</tbody>
</table>

Figure 4.3-2 The Noise Power Spectra of Photodiode Array
4.3.2.B. Testing of Detector Responsivities

A blackbody was placed in front of the dewar filling the detector field of view. A chopper is used to modulate the radiation and provide the voltage amplifier with a reference. For a five-degree temperature change around the background temperature of 300K, the voltage change (Δv) is recorded. The radiant flux difference between temperature T2 and T1 is given by

\[ \Delta \Phi = A_d \sin^2(\theta) \int_{\lambda_1}^{\lambda_2} (M(\lambda, T_2) - M(\lambda, T_1)) d\lambda \]  

(4.3-1)

Where θ is the half-acceptance angle (20 degree), A_d is the area of the detector, and M is given by the Planck’s Equation. The responsivity is given by

\[ R = \frac{\Delta v}{\Delta \Phi} \]  

(4.3-2)

The circuit used in the testing process was a low-impedance voltage amplifier very similar to that shown in Figure 3.4-10 with a gain of 1000. The bias current was set to be 5 mA. A feedback loop was added so that when the reference (one of the two blackbodies) was presented to the detector, the circuit output was clamped to zero. This clamping circuit effectively eliminates the low frequency noise from bias, detector temperature, and 1/f noise. These noises are very significant for MCT detectors. It also increases the dynamic range of the detection system. By changing the reference blackbody temperature, it is possible to cover a wide range of scene radiances. The noise of the circuit alone is about 2 mV rms (measured with the bandwidth of the voltage meter ≈ 300 kHz).

Table 4.3.2 lists the measured responsivities and estimated NEDT. The radiance flux (ΔΦ) is calculated from Eq. 4.3-1 with T2=301 and T1=300, and acceptance angle of θ = atan(0.5/f#)=8.5 (MISI’s collecting angle) and system throughput of 50%. The measured responsivities are still worse than the specifications. Most of the channels still satisfy our NEDT requirement, except channel 2 and 5 which have lower responsivity. It seems that the noise associated with these two detectors is due to amplifier, thus it is possible to reduce the NEDT by a factor of close to 2 by using 10 mA bias.

Table 4.3.2
4.3.2.C. Testing of Detector Frequency Response

A glow bar was placed on the input slit of a 2-meter collimator (F/# = 10) to produce a collimated infrared beam. This beam was deflected by the scan-mirror and focused onto the focal plane by the telescope (See Figure 4.3-3). By spinning the scan-mirror, the focused spot was scanned across the detector. The response from the detector was used to determine the detector frequency response. The rise time and width (μs) are listed in the Table 4.3-3 (note that the scan frequency is 10 Hz).

![Detector Frequency Response Measurement](image)

**Table 4.2-3** MCT Detector Frequency Response

<table>
<thead>
<tr>
<th></th>
<th>Detector 1</th>
<th>Detector 2</th>
<th>Detector 3</th>
<th>Detector 4</th>
<th>Detector 5</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rising time (μs)</td>
<td>17</td>
<td>8</td>
<td>8</td>
<td>12</td>
<td>10</td>
</tr>
<tr>
<td>Width (μs)</td>
<td>36</td>
<td>17</td>
<td>17</td>
<td>31</td>
<td>33</td>
</tr>
</tbody>
</table>

The theoretical width should be 16 μs for the 1 mrad detector and 32 μs for the 2 mrad detector. The measured width agrees with the predicted width. We believe that the rise time was caused by the finite size of the slit and optical blur instead of detector and preamplifier frequency response.
4.4 Overall System Performances Validation

MISI was assembled in March of 1995. Since then, it has undergone extensive testing to determine its overall image quality. The most important image quality factors are the system modulation transfer function and overall signal-to-noise ratio.

4.4.1 The Measurement of the Optical Point Spread Function (PSF)

The optical point spread function is the response of the optical system (including scan-mirror, fold mirror, and telescope) to a point source. In this experiment, the point source was simulated by placing a 10 \(\mu\)m pinhole at the front focal point of a lens with 2 meter focal length. The pinhole was illuminated by a He-Ne laser. The image of the point source was captured by a video camera. Since all the MISI focal planes are not accessible to the video camera, a coherent fiber bundle (d=1/8") was used to transfer the image from the focal plane to the CCD sensor of the video camera. Figure 4.4-1 shows the measured PSF. The image on the left is the PSF on the primary focal plane (center of the field of view). The image on the right is the optical PSF at the long-wave focal plane, which is about 30 IFOV off axis. The coma is clearly visible as expected from optical ray tracing. The rectangular box in Figure 4.4-1 represents the size of an one milliradian detector (high resolution). The optical blur is much less than the detector aperture.

Figure 4.4-1 PSF of the MISI Optical System Measured by a Video Camera

4.4.2 The Measurement of System LSF

The LSF is the system response to a line object. Since the Fourier transform of a LSF is one central line of the 2-D system MTF, the frequency response of the system can be easily derived from the LSF. The setup used to measure the LSF is very similar to that used to measure the optical PSF described in section 4.4.1, except that the pinhole was
replaced with a line source. In this experiment, a tungsten wire of diameter 0.25 mm was used. The use of a hot wire and a reflective collimator (Newtonian telescope) made it possible to measure the LSF of both the visible and infrared channels. The penalty of this setup is that the wire is not a theoretical line object, and the image quality of the Newtonian collimator is much worse than that of the 2-meter collimator used in the PSF measurement. Thus the MTF derived from this experiment is probably lower than the true MTF of the system. Figure 4.4-2 is the measured LSF of the high resolution Si detector. Figure 4.4-3 is the MTF of the Si detector derived from the LSF for the Si channels.

![Graph](image)

Figure 4.4-2 The Measured LSF of the Si Detector: The size of the detector is one milliradian
Figure 4.4-3 The Measured MTF of the Si Detector

Figure 4.4-4 shows the measured LSFs of the long-wave focal plane. The narrow curve is the LSF of the high resolution channel (one milliradian), and the wide one is the LSF of the low resolution channel. There is a "bump" in the measured LSF. This is most likely caused by stray reflections from the edge of the pyramid. Figure 4.4-5 plots the derived MTF of the two channels. Again the dimples in the MTF curves are caused by the bumps in the LSF.

Figure 4.4-4 LSF of Thermal Infrared Channels
For all channels, the MTF at the Nyquist frequency is around 0.4, which is very close to the model predicted MTF as shown in Figure 3.5-3.

4.4.3 The Measurement of the Edge Spread Function (ESF)

The PSF and LSF measurements were performed in the laboratory to determine the frequency response of an imaging system. Because a good approximation to a perfect point or line source is very difficult to find in the operational environment. The operational MTF of MISI had to be derived from the measured edge spread function. Since an edge is the only natural target available in the operational environment, ESF measurement is the only technique available to evaluate the frequency response of operational imaging systems in actual use. But MTF measured from the ESF is inherently inaccurate due to noise in the data (c.f. Section 4.5.1). The problem became even worse for MISI, because (1) MISI was still in the test stage and the electronics were not optimized, and (2) the edge targets available were of very low contrast (only a few percent) at the Roof Lab of the Carlson Building. This made it impossible to derive any useful information from the measured ESF using traditional knife-edge methods. A model-based algorithm was developed to derive MTF from ESF (section 4.5). By using prior information about the imaging system, the operational MTF was derived from very noisy edge scans.
Figure 4.4-6 shows the ESF of a high-resolution Si channel as recorded by a digital oscilloscope. The sync signal is the output from the optical encoder. Each clock cycle in the sync signal represents one sample for the high resolution channel or one milliradian in the angular resolution. The measured ESF was fitted to the MISI's spatial frequency model as shown in Figure 4.4-7. Figure 4.4-8 shows the MTF of the high resolution Si detector derived from the ESF. The MTF is again very similar to the MTF measured from LSF in Figure 4.4-3 with the MTF difference at the Nyquist frequency less than 0.05.

Figure 4.4-6 Measured Edge Spread Function of the High Resolution Si Channel
Figure 4.4-7 The Measured ESF and the Model Fitted ESF of Si Channel

Figure 4.4-8 The MTF Derived from ESF for the Si Channel

Figure 4.4-9 and Figure 4.4-10 are measured ESF and derived MTF for the high resolution long-wave (MCT) channel. The ESFs of other channels were also measured, there were essentially no difference between channels on the same focal plane with the same aperture size.
Figure 4.4-9 ESF and MTF of the High Resolution MCT Channel

Figure 4.4-10 shows the measured ESF of the imaging spectrometer. The ESFs were corrupted by random noise and low frequency interference. The model-based algorithm was able to derive the MTF from this noisy edge trace. The measured MTF is very close to the MTF predicted by the aperture function of a 2 mm fiber used to transmit radiation from the primary focal plane to the spectrometer.

Figure 4.4-10 ESF and MTF of the Imaging Spectrometer with 2 mm Fiber
4.4.4 The Measurement of Signal-to-Noise Ratio

Besides the system MTF, the SNR is another most important image quality factor. The MTF, SNR and the sampling resolution determine the information channel capacity of any imaging system (c.f. Section 4.6). The SNR of MISI were characterized by the noise equivalent reflectance (NEAR) for the reflective channels and noise equivalent temperature (NEAT) for the thermal infrared channels.

4.4.4.1 The Measurement of NEAR

The noise equivalent reflectance is given by

\[
NEAR = \frac{\text{Noise}(V)}{\text{Responsivity}(V/R)}
\]

The noise of the detector/pre-amplifier was measured as described in Section 4.3.1. The responsivity of each channel was measured by placing a 8" diameter Spectralon sample in front of the scan-mirror. Since the reflectance of Spectralon is approximately 100% [Feng et al. 1992], the response of the detector when viewing the Spectralon is the responsivity expressed as volts per unit reflectance. The responsivity of the two broadband Si detectors and the 12 channels of imaging spectrometer were measured and are shown in Table 4.4-1. The measurements were made at about 3:00 PM on September 29, 1995. It was a typical data collection day for remote sensing applications. The measured photo-current is about 20 nA for the spectrometer channels, which is within the range of predicted photo-current (10 ~ 30 nA) from published solar irradiance data.[RCA, 1974]. The NEARs are in the order of 0.0001 for the Si broad-band channels and 0.0007 for the spectrometer channels. The test conducted on October 10, 1995 with a 80% reflectance panel also support this results.
**Table 4.4-1 Noise Equivalent Reflectance**

<table>
<thead>
<tr>
<th>Channel</th>
<th>Spectral Band</th>
<th>Output Voltage (V)</th>
<th>Output Current (pA)</th>
<th>Noise Current (pA)</th>
<th>NEΔR</th>
</tr>
</thead>
<tbody>
<tr>
<td>Si 1</td>
<td>Broad-band</td>
<td>7.8</td>
<td>780000</td>
<td>65</td>
<td>0.0001</td>
</tr>
<tr>
<td>Si 2</td>
<td>Broad-band</td>
<td>7.6</td>
<td>760000</td>
<td>65</td>
<td>0.0001</td>
</tr>
<tr>
<td>740</td>
<td>10 nm</td>
<td>0.13</td>
<td>12600</td>
<td>15</td>
<td>0.0012</td>
</tr>
<tr>
<td>750</td>
<td>10 nm</td>
<td>0.21</td>
<td>20700</td>
<td>15</td>
<td>0.0007</td>
</tr>
<tr>
<td>760</td>
<td>10 nm</td>
<td>0.19</td>
<td>19350</td>
<td>15</td>
<td>0.0008</td>
</tr>
<tr>
<td>770</td>
<td>10 nm</td>
<td>0.21</td>
<td>21150</td>
<td>15</td>
<td>0.0007</td>
</tr>
<tr>
<td>780</td>
<td>10 nm</td>
<td>0.21</td>
<td>20700</td>
<td>15</td>
<td>0.0007</td>
</tr>
<tr>
<td>790</td>
<td>10 nm</td>
<td>0.23</td>
<td>22500</td>
<td>15</td>
<td>0.0007</td>
</tr>
<tr>
<td>800</td>
<td>10 nm</td>
<td>0.23</td>
<td>22500</td>
<td>15</td>
<td>0.0007</td>
</tr>
<tr>
<td>810</td>
<td>10 nm</td>
<td>0.23</td>
<td>22500</td>
<td>15</td>
<td>0.0007</td>
</tr>
<tr>
<td>820</td>
<td>10 nm</td>
<td>0.23</td>
<td>22950</td>
<td>15</td>
<td>0.0007</td>
</tr>
<tr>
<td>830</td>
<td>10 nm</td>
<td>0.23</td>
<td>22500</td>
<td>15</td>
<td>0.0007</td>
</tr>
<tr>
<td>840</td>
<td>10 nm</td>
<td>0.17</td>
<td>17100</td>
<td>15</td>
<td>0.0009</td>
</tr>
</tbody>
</table>

### 4.4.4.A The Measurement of NEAT

The noise equivalent temperature difference is given by

$$\text{NEAT} = \frac{\text{Noise}(V)}{\text{Responsivity}(V/K)}$$

4.4.2 where the rms noise of the MCT detector/pre-amplifier is listed in Table 4.3-2. The responsivity was measured by placing the blackbody in front of the scan-mirror. The blackbody temperature was changed from 15°C to 25°C and back to 15°C. The voltage differences were averaged to reduce DC drifting.

**Table 4.4-2 The Measured NEDT for the Long-wave Focal Plane**

<table>
<thead>
<tr>
<th>Channel</th>
<th>Spectral Band</th>
<th>ΔV (10 K)</th>
<th>Responsivity (V/K)</th>
<th>Noise (mV)</th>
<th>NEΔT</th>
</tr>
</thead>
<tbody>
<tr>
<td>Det 1</td>
<td>9.7-11.3</td>
<td>0.031</td>
<td>0.0031</td>
<td>0.3</td>
<td>0.1</td>
</tr>
<tr>
<td>Det 2</td>
<td>8.0-13.9</td>
<td>0.009</td>
<td>0.0009</td>
<td>0.22</td>
<td>0.24</td>
</tr>
<tr>
<td>Det 3</td>
<td>9.0-13.7</td>
<td>0.047</td>
<td>0.0047</td>
<td>0.39</td>
<td>0.08</td>
</tr>
<tr>
<td>Det 4</td>
<td>8.3-9.9</td>
<td>0.02</td>
<td>0.002</td>
<td>0.3</td>
<td>0.15</td>
</tr>
<tr>
<td>Det 5</td>
<td>11.7-14.4</td>
<td>0.0072</td>
<td>0.0072</td>
<td>0.22</td>
<td>0.3</td>
</tr>
</tbody>
</table>
The values of noise listed in Table 4.4-2 were measured with an electronic bandwidth of about 300 KHz. It is very possible that these values will decrease if a narrower electronic bandwidth is defined.11

The measured PSF confirms that the spatial resolution of MISI is detector limited as designed. The MTF derived from the LSF and ESF agrees very well with the MTF predicted from the model of the optical/electronic subsystems. The noise equivalent reflectance is below 0.1% and the noise equivalent temperature is around 0.1 K. The NEAT of some channels can be reduced because they are not detector limited. At the time of writing this thesis, detector 2 was replaced by a better detector with the performance closely matching that of the detector 3. It is expected that both detector 2 and 3 will have a NEAT of 0.1 K. The NEAT of detector 5 is expected to improve some with 10 mA bias and a higher order low-pass filter. All these NEAT values are in lines with the predicted values as listed in Table 3.4-1.

4.5.1 A Model-based Algorithm to Derive MTF from ESF

The optical transfer function (OTF) is a significant image quality descriptor of any imaging system. The modulus of OTF - the modulation transfer function (MTF) is often measured to be the figure of merit for electro-optical imaging system. Traditionally, the OTF of an imaging system can be measured with images of known input, such as sine-wave or square-wave targets, point sources or line sources, or edges. Edge targets are used extensively to derive the MTF of operational instruments because edges are readily available in the field test and the MTF derived from the knife-edge method represents its frequency response in operation. In the knife-edge method, the edge response of the instrument is measured. The line spread function is computed from the edge response, and the Fourier transform of LSF yields OTF. The power spectrum (squared magnitude of the spectrum) of an edge is a continuous function that falls off as the inverse square of frequency \((1/f)^2\). Thus signal-to-noise ratio is smallest at the same high frequencies where the OTF is most desirable to measure. Noise in the measurement not only causes the measured MTF to have random fluctuations, but also result in a bias at higher frequencies [Blackman, 1968]. Thus knife-edge (also called edge-gradient) analysis is

---

11Theoretically, the noise is proportional to the square-root of bandwidth, thus the rms values in Table 4.4-2 will be significantly reduced if MISI is operated at 40 Hz. But if detector noise is not white, the reduction in rms value may not obey the estimated.
among the least accurate method to determine the system MTF [Swing and McCamy, 1969]. In spite of this problem, knife-edge methods are still used very extensively for operational evaluation, because it is the only technique generally available. The edge is the only natural target that is useful in the absence of ground luminance information. Many papers in the literature describe methods to reduce the effect of noise [Blackman, 1968] [Jones and Yeadon, 1969] [Fischer and Holm, 1994], most of them involve averaging in the space or frequency domain. But averaging can not solve the fundamental problem of noise. Knife-edge analysis remains to be a widely used but unfortunately inaccurate way of characterizing imaging systems in the operational environment [Swing and McCamy, 1969].

In this thesis, a model-based algorithm was used to derive MTF from edge trace data. The algorithm is based on the fact that the overall MTF of any electro-optical imaging system is the products of optical transfer function of the optics, detector aperture function, and the transfer function of electronic signal processing. The transfer function of each subsystem can be described by a model with a few parameters. By using prior knowledge of the imaging system, the MTF of an imaging system can be derived from very noisy edge trace data.

4.5.1 Theory of knife-edge method

Figure 4.5-1 shows the block diagram of a typical electro-optical imaging system. The object is imaged at the focal plane of the optical system. The detector samples the optical image and converts optical radiation into an electrical signal, and this signal is further amplified and processed by the electronics and finally it is converted into a digital bit stream.

![Figure 4.5-1 Model of electro-optical imaging systems](image)

This process can be represented in mathematical form as
\[
\begin{align*}
\begin{align*}
\mathbf{s}(x,y) &= \left[K\mathbf{L}(x,y) \ast \tau_o(x,y) \ast \tau_d(x,y) \ast \tau_e(x)\right] \text{comb}(x,y) + n(x,y) \\
\end{align*}
\end{align*}
\]

(4.5.1)

where

- \(L(x, y)\) is the input radiance,
- \(s(x, y)\) is the output digital images,
- \(\tau_0(x, y)\) is the optical point spread function,
- \(\tau_e(x)\) is the impulse response function of the electronic signal process,
- \(\tau_d(x, y)\) is the detector point spread function
- \(n(x)\) is the noise of detector/pre-amplifier,
- \(K\) is the responsivity of the detector,

The symbol \(\ast\) denotes the convolution operation, and \(\text{comb}(x, y)\) is the sampling function with a sampling interval of 1

\[
\begin{align*}
\text{comb}(x, y) &= \sum_{n=-\infty}^{+\infty} \sum_{m=-\infty}^{+\infty} \delta(x - n, y - m) \\
\end{align*}
\]

(4.5.2)

The imaging process formulated by Eq. 4.5.1 can also be represented in the frequency domain \((u, v)\) which is usually computationally more efficient. Eq. 4.5.1 becomes

\[
\begin{align*}
\mathbf{S}(u,v) &= \left[K\mathbf{L}(u,v) \tau_o(u,v) \tau_d(u,v) \tau_e(u)\right] \ast \text{comb}(u,v) + n(u,v) \\
\end{align*}
\]

(4.5.3)

where \(u\) and \(v\) are spatial frequency in cycles/sample, or cycles/mm. It can be seen from Eq. 4.5.3 that the performance of the imaging system not only is limited by the blur caused by optics, detector aperture and noise, but also by the aliasing effect caused by insufficient sampling [Huck et al. 1980]. Since an ideal edge is not band limited, it is very important to sample the edge spread function at a resolution where the imaging system has a low frequency response in order to eliminate the aliasing. For systems with fixed sampling resolution, a super-sampling algorithms must be used to achieve sufficient sampling [Reichenbach et al. 1991]. With sufficient sampling, and perfect reconstruction, Eq. 4.5.3 becomes

\[
\begin{align*}
\mathbf{S}(u,v) &= K\mathbf{L}(u,v)T(u,v) + N(u,v) \\
\end{align*}
\]

(4.5.4)

where \(T(u,v) = \tau_o(u,v)\tau_d(u,v)\tau_e(u)\) is the total system transfer function. In knife-edge analysis, the edge used as a target, is defined as:
\[ e(x) = \text{step}(x) \quad (4.5.5) \]

The edge spread function (ESF) is the convolution of the step function with the point spread function (PSF) of the imaging system. In experiments to collect the ESF data, the measured edge is inevitably corrupted by noise from the imaging system. The measured ESF is given by

\[ \text{esf}(x) = \text{step}(x) \ast \text{psf}(x,y) + n(x) \quad (4.5.6) \]

Differentiating Eq. 4.5.5 yields

\[ \frac{d}{dx} \text{esf}(x) = \frac{d}{dx} [\text{step}(x) \ast \text{psf}(x,y)] + \frac{d}{dx} n(x) \quad (4.5.7) \]

since the derivative of the step function is the Dirac delta function \( \delta(u) \) and convolution of a delta function and the point spread function is the line spread function (lsf) [Gaskill, 1988], Eq. 4.5.7 becomes

\[ \frac{d}{dx} \text{esf}(x) = \delta(x) \ast \text{psf}(x,y) + n'(x) = \text{lsf}(x) + n'(x) \quad (4.5.8) \]

The Fourier transform of Eq. 4.5.8 is

\[ F\{\text{esf}'(x)\} = \int_{-\infty}^{\infty} (\text{lsf}(x) + n'(x)) e^{-j2\pi u x} dx = T(u) + j 2\pi u N(u) \quad (4.5.9) \]

where \( T(u) \) is the transfer function of the imaging system, and \( N(u) \) is the Fourier transform of noise. It is apparent from Eq. 4.5.9 that the signal decreases and the noise increases as frequency increases; the SNR of the MTF is inversely proportional to the square of frequency. One way to reduce the noise is to average \( N \) edge traces. If the noise is uncorrelated, the standard deviation is reduced by a factor of \( \sqrt{N} \). This results in a smoother curve, but it also broadens the ESF if alignment is imperfect. Eq. 4.5.10 shows the frequency domain representation of the averaging process:

\[ \frac{1}{N} \sum_i F\{\text{esf}''(x)_i\} = \frac{1}{N} \sum_i |T(u)| e^{j[\phi(u)+2\pi \Delta x_i u]} + \frac{1}{N} \sum_i j 2\pi u N_i(u) \quad (4.5.10) \]

Where \( |T(u)| \) is modulus of the transfer function \( T(u) \), \( \phi(u) \) is the phase of \( T(u) \) without registration error, and \( \Delta x_i \) is the amount of misregistration in edge traces. Since the noise most likely is uncorrelated, the noise term is reduced as \( N \) increases. But unfortunately
the phase of the transfer function is very sensitive to edge alignment. If there is a
misregistration of a quarter of a pixel, it will cause approximately 10% roll-off in the
measured MTF at the Nyquist frequency.

Another method used to smooth MTF is to average the MTF of many traces. Summation
the modulus of Eq. 4.5.9 yields

$$\sum_i |F\{\text{esf}(x)_i\}| = \sum_i \sqrt{(\text{Re}\{T_i(u)\} + 2\pi u \text{Im}\{N_i(u)\})^2 + (\text{Im}\{T_i(u)\} + 2\pi u \text{Re}\{N_i(u)\})^2}$$

(4.5.11)

And after rearranging Eq. 4.5.11, we have

$$\sum_i |F\{\text{esf}(x)_i\}| = \sum_j \sqrt{(\text{Re}\{T(u)\})^2 + (\text{Im}\{T(u)\})^2} \cdot \sqrt{\left[1 + \frac{4\pi u (\text{Re}\{T(u)\} \cdot \text{Im}\{N(u)\} + \text{Im}\{T(u)\} \cdot \text{Re}\{N(u)\})}{\text{Re}\{T(u)\}^2 + (\text{Im}\{T(u)\})^2} \right] + \frac{(2\pi u)^2 (\text{Re}\{N(u)\})^2 + (\text{Im}\{N(u)\})^2}{\text{Re}\{T(u)\}^2 + (\text{Im}\{T(u)\})^2}}$$

(4.5.12)

substituting $|T(u)|^2 = \text{Re}\{T(u)\}^2 + \text{Im}\{T(u)\}^2$ into Eq. 4.5.12, and approximating the
square-root with a first order Taylor series yields

$$\sum_i |F\{\text{esf}(x)_i\}| \approx \sum_i |T(u)| + \sum_i \frac{2\pi u (\text{Re}\{T(u)\} \cdot \text{Im}\{N(u)\} + \text{Im}\{T(u)\} \cdot \text{Re}\{N(u)\})}{|T(u)|^2} + \sum_i \frac{(2\pi u)^2 |N(u)|^2}{2|T(u)|^2}$$

(4.5.13)

Dividing both sides of Eq. 4.5.13 by the number of traces averaged gives

$$\frac{\sum_i |F\{\text{esf}(x)_i\}|}{N} = \frac{1}{N} \sum_i |T| + \frac{1}{N} \frac{2\pi u}{|T|^2} \left[ \sum_i (\text{Re}\{T\} \cdot \text{Im}\{N\} + \text{Im}\{T\} \cdot \text{Re}\{N\}) + \pi u \sum_i |N|^2 \right]$$

(4.5.14)

Since there is no correlation between transfer function and noise, the second term on the
right side of Eq. 4.5.14 goes to zero as $N$ becomes large. The measured MTF is given by
where \(|N(u)|^2\) is the noise power spectrum. The measured MTF is the system MTF plus a noise term which is the noise power spectrum multiplied by a term proportional to the square of frequency and inversely proportional to the square of MTF at that frequency. Averaging over MTF results in a true MTF plus a bias which is dependent not only on frequency and noise power spectrum, but also on the value of MTF at that frequency.

This analysis confirms the conclusion of Swing and McCamy that the edge-gradient method is inaccurate because of noise, and averaging over multiple edge traces can reduce the random fluctuation, but may introduce an error in the measured MTF [Swing and McCamy, 1969]. The problem becomes worse when this technique is applied to electro-optical imaging systems, where the signal-to-noise ratio of the system is lower than that of the typical microdensitometers used in the evaluation of the photographic system, and the contrast of the edges typically used are very low under operational conditions. It is imperative that we develop better techniques to derive MTF from edge response. In the following section, a model-based algorithm is described to derive MTF from the ESF.

### 4.4.3 Model-based algorithm to derive MTF from ESF

In the traditional knife-edge analysis, no prior information about the imaging system is assumed. This certainly is valid if no knowledge about the system to be measured is available. But in most cases, we do have some knowledge about the electro-optical imaging system being tested. Johnson [Johnson, 1970] measured the MTF of many types of electro-optical devices and found that it can be fitted to a function in the form:

\[ T(f) = \exp\left(-\frac{f}{f_c}\right)^N \]

Thus only two parameters are needed to characterizing the electro-optical device. For most electro-optical imaging systems, the transfer function of each subsystem do not vary arbitrarily, but are restricted, in a controlled fashion which is determined by a few parameters (Figure 4.5-1). For example, the frequency response of the detector is uniquely determined by its size and shape. The overall system transfer function is the product of subsystem transfer functions as given by

\[ T(u, v) = T_o(u, v)T_d(u, v)T_e(u) \]
Thus the system transfer function can be modeled with a few parameters of the three subsystems. The MTF can be determined by fitting the ESF with a set of known functions. The model approach creates redundancy in the data set which allows the use of non-linear regression method to derive MTF from edge trace data. The next section describes a subsystem transfer function model for electro-optical imaging systems.

4.5.3. A System transfer function model

The line spread function of the optical system can be approximated by the first three terms of the Fermite function \[ \text{Smith, 1972} \]

\[
\text{lsf}(z) = f(z) \left[ c_1 + c_2 z + c_3 \left( z^2 - 1 \right) \right]
\]

(4.5.17)

where

\[
f(z) = \frac{1}{\sqrt{2\pi}} \exp\left( -\frac{z^2}{2} \right)
\]

is a Gaussian function, and \( z = (x-x_0)/\sigma \), \( \sigma \) is the width of a Gaussian function.

Rearranging Eq. 4.5.17 gives

\[
\text{lsf}(z) = \left( c_1 - c_3 \right) f(z) + \frac{c_2}{j2\pi} (j2\pi z) f(z) + \frac{c_3}{(j2\pi)^2} (j2\pi z)^2 f(z)
\]

(4.5.18)

The Fourier transform of Eq. 4.5.18 is the optical transfer function. Based on the derivative properties of Fourier transform \[ \text{Gaskill, 1988} \], the OTF is given by

\[
\text{OTF}(\omega) = \left( c_1 - c_3 \right) F(\omega) + \frac{c_2}{j2\pi} F'(\omega) + \frac{c_3}{(j2\pi)^2} F''(\omega)
\]

(4.5.19)

where the ' denotes the derivative operation with respect to the spatial frequency \( \omega \), which is the conjugate quantity \( z \). The Fourier transform of a Gaussian function is another Gaussian function:

\[
F(\omega) = F\{ f(z) \} = \exp( -2\pi^2 \omega^2 )
\]

(4.5.20)

Substituting Eq. 4.5.20 into Eq. 4.5.19

\[
\text{OTF}(\omega) = (c_1 + j2\pi c_2 \omega - 4\pi^2 c_3 \omega^2) \exp(-2\pi^2 \omega^2)
\]

(4.5.21)
According to the scaling theorem of Fourier transform

\[ OTF(u) = \alpha e^{j2\pi c_1 u} (c_1 + j2\pi c_2 \sigma u - 4\pi^2 c_3 \sigma^2 u^2) e^{-2\pi^2 \sigma^2 u^2} \]  

(4.5.22)

Since MTF is normalized to 1 at \( u=0 \), this set \( c_1=1/\sigma \). In some cases, the optical transfer function closely resembles a Gaussian function, thus only the first term of the Fermite function may be needed to model the optical transfer function of the optical system.

The transfer function of the detector is given by

\[ T_d(u) = Sinc(bu) = \frac{\sin(\pi bu)}{\pi bu} \quad \text{for a rectangular aperture or} \]  

\[ T_d(u) = Somb(bu) = \frac{2J_1(bu)}{bu} \quad \text{for a circular aperture} \]  

(4.5.23)

\( J_1 \) is the first order Bessel function, and \( b \) is the width or diameter of the aperture respectively.

The transfer function of the electronic signal processing depends on the actual filters used. For an \( n \)-th order Butterworth filter with cutoff frequency \( u_c \), the magnitude of the transfer function is given by [Oppenheim and Schafer, 1975]

\[ |T_e(u)| = \frac{1}{\sqrt{1+2\pi(j\gamma u_c)^2}} \]  

(4.5.24)

The electronic impulse response function usually can be measured easily. The system frequency transfer function is controlled by eight parameters:

- \( x_0 \) Position of the edge (Eq. 4.5.22)
- \( \sigma \) Width of the Gaussian function in Eq. 4.5.17
- \( c_1, c_2, c_3 \) Optical system parameters, where \( c_1=1/\sigma \)
- \( b \) Detector size
- \( u_c \) and \( n \) Electronic cutoff frequency and filter order

With these eight parameters, the transfer function of an imaging system is completely determined. The process of deriving the MTF from edge-trace data becomes a nonlinear optimization process. The advantage of this approach is only 8 parameters need to be estimated to completely determine the MTF rather than the amplitude of each sample.
This introduces redundancy which allows a least mean-squares solution to the MTF calculation.

**4.5.3.B. Implementation of the Algorithm**

The model-based algorithm is implemented as follows:

1. Obtain a sampled edge trace using the imaging system at a very small sampling interval ( < 1/4 of the sampling interval of the imaging system). If the hardware does not support such super-resolution sampling, a super sampling technique described by Reichenbach *et al.* [1991] can be used to obtain a super resolution edge.

2. Bias the data to obtain an edge with a positive slope with amplitude 0 ≤ esf(x) ≤ 1. Correct any shading caused by illumination.

3. Set up the model. Determine if any of the eight parameters is not a variable in the optimization process. For example, if we know the detector size exactly, we do not need to have the detector size parameter b in the optimization process.

4. Generate a perfect edge (step function) of length M (M = 2^n > \( \frac{3N}{2} \)), where n is an integer number, N is the number of elements in the measured ESF. Convolve the step function with the known spread functions, i.e. the electronic filter function, or detector aperture function, or both if all are known. Compute the FFT of the edge.

5. Multiply the spectrum of the edge (from step 4) with the remaining elements in the estimated transfer function of the system (exclude the known transfer functions in step 4), and compute an inverse Fast Fourier transform (IFFT) to obtain an estimated ESF.

6. Calculate the mean-square error (MSE) between the estimated ESF and measured ESF.

7. Adjust the transfer function parameters, repeat step 5 to 7 until the MSE reaches a minimum.

8. Plugging all the parameter into Eq. 4.5.22, 23, and 24, we have the measured MTF in analytical form.

The minimization algorithm used is the simplex method described by Press *et al.* [1988]. The method only requires function evaluation, not derivatives. With N ≈ 50, it normally takes less than a minute to complete one optimization. The parameters are initialized as follows:

- **b, n, f_c,** interactive input from user (Known or estimated values)

\[ x_0 = \text{Mean}(x_i), x_i \in (0.8 > \text{ESF}(x_i) > 0.2) \]

\[ \sigma = \text{Max}\left[\text{Stdev}(x_i) - b / 2, 0.2\right], x_i \in (0.8 > \text{ESF}(x_i) > 0.2) \]

\[ c_2 = 0 \]

\[ c_3 = 0 \]
The edge position is initialized to be the mean position of the points inside the box as shown in Figure 4.5-2, and the $\sigma$ is estimated to be the standard deviation of the points inside the box.

$$T(u) = \exp\left[-\left(\frac{u}{\rho}\right)^2\right]\text{Sinc}(\beta u)\text{Sinc}(\gamma u)$$

(4.5.25)

where $\rho_c = 1/\sqrt{2}$, $\beta = 5/6$, and $\gamma = 1/6$. The test edge was generated by convolving a step function with the psf (inverse Fourier transform of OTF).

$$esf(x) = step(x) * \text{rect}\left(\frac{x}{\Delta}\right) * psf(x)$$

(4.5.26)

where $\text{rect}(x/\Delta)$ is the rectangle function of width of $\Delta$. In this simulation, the edge is sampled at 4 times the Nyquist frequency, i.e. $\Delta = 0.25$. Figure 4.5-3 shows the edge spread function and the OTF of the simulated system.
Figure 4.5-3 The ESF and OTF of the simulated imaging system

The simulated ESF was used to test the algorithm. Figure 4.5-4 shows the simulated ESF and the model-fitted ESF without noise. Although the transfer function model of simulated imaging system is not the same as the transfer function model used in this algorithm, there is almost a perfect fit between the simulated ESF and modeled ESF. The RMS error of fit is only 0.00005.

Figure 4.5-4 Simulated ESF and model fitted ESF
For comparison, the MTF of the simulated system was also derived using the traditional derivative technique. Figure 4.5-5 shows the OTF of the simulated system and its measured MTF using both the model-based algorithm and the derivative method. Since the OTF of the simulated system has no phase term, the OTF is the true MTF of the system. The RMS error of model-derived MTF is 0.001, which indicates that the model method is very accurate. The MTF from the derivative method is a little bit lower than the true MTF with RMS error of 0.01. The lower MTF is caused by the approximation of the derivative with a convolution kernel [-1 1]. Without noise, both methods yield good estimates of the MTF.

![OTF and MTF](image)

**Figure 4.5-5** The OTF of a simulated imaging system

In order to test the robustness of the algorithm against noise, a zero-mean Gaussian noise is added to the simulated ESF. Figure 4.5-6 shows the measured MTF at an SNR of 20:1. The MTF from the derivative method is very noisy with an RMS error of 0.16. There is a positive bias at higher frequency as predicted by Eq. 4.5.9. But the MTF derived from the model still is very close to the true MTF with an RMS error of 0.014. This indicates that the model-based algorithm is very robust against noise. Thus this method is well suited for deriving the MTF from ESF under operational conditions, where an edge is the only target available and noise always is present.
4.5.4 Application of the algorithm to electro-optical systems

The algorithm was used to derive the MTF of two real imaging systems. The first imaging system is a drum-type graphic scanner (Hell DC299). It uses a high quality microscopic lens, a circular aperture and a photomultiplier. Figure 4.5-7 shows the measured ESF and the calculated MTF. The scanner MTF was also measured with a tuned square-wave technique [Newell and Triplett, 1994]. The agreement between the two methods is very good.
Figure 4.5-7 (A) The ESF of the Hell Scanner with 90 μm aperture

Figure 4.5-7(B) MTF of the Hell scanner measured with square wave method and Model-based method

The second imaging system tested was the MISI. The SNR was quite low due to two factors: the reflectance between target and background were only a few percent in the available imagery, and the instrument was not yet fully optimized due to the nature of the testing environment. Figure 4.5-8 shows the measured ESF and the model-fitted ESF. The measured ESF is corrupted not only by random noise, but also by low-frequency interference from other instruments. The dashed line is the sampling clock from the optical encoder; where each clock pulse corresponds to one milliradian in spatial
resolution. System MTF analysis indicates that this spectrometer is primarily aperture limited due to the large optical fiber used to transmit energy from the primary focal plane to the entrance aperture of the spectrometer [Feng et al. 1994]. The diameter of the fiber corresponds to about 4 milliradian angular resolution, which yields a cutoff frequency (first zero of the Sombrero function) of 3.1 cycles/milliradian. Figure 4.5-8 (B) shows the MTF derived from the model-based approach where the MTF cutoff frequency is about 3.2 cycles/milliradian. The measured cutoff frequency is very close to the predicted cutoff frequency.

Figure 4.5-8 (A) The ESF of Mysi
4.6 Information Channel Capacity$^{12}$ - An Alternative Image Quality Criterion for Sampled EO Imaging Systems

Information channel capacity has been used widely as a figure-of-merit for communication channels. An informationally optimized communication channel maximizes the fidelity of optimally restored representations (Shannon 1948). Because of the inability of an optical systems to perform restoration (inverse filtering), an informationally optimized optical system does not ordinarily maximize the fidelity of the optical image. Therefore it seems that information can not be used as a valid figure-of-merit for optical systems, and without further reexamination, information criterion is not used in the design of image-gathering systems. But unlike optical systems, restoration of digital images is possible with the advent of computer technology. Thus it seems that information can once again be used as a figure-of-merit for EO imaging systems. In this section, we formulate information channel capacity for the MISI and show that the information channel capacity can be increased 1.5 bits by shaping the detector aperture.

---

$^{12}$ MISI is a line scanner which can be approximated to be a 1-D information channel. The ground information is collected by the scanner pixel-by-pixel. The ability of MISI to capture the ground information is defined to be information channel capacity.
and electronic filter. Computer simulations was used to test the validity of the information criterion.

4.6.1 Traditional Imaging System Design Rules

According to the traditional design rules, each IFOV (corresponding to a pixel in the image) defines the minimum resolution unit (Jaggi 1992; Seyrafi 1984). Every consecutive scan line is separated by 1 IFOV in the track direction. In the scan direction, each pixel (sample) is again 1 IFOV apart. Thus every point on the ground is sampled by the detector one time only by the detector. The electronic cutoff frequency is given by

\[ f_c = \frac{1}{2\tau_d} = \frac{f_s}{2} \]  

(4.6.1)

Where \( \tau_d \) is the detector dwell time and \( f_s \) is the sampling frequency. In most cases, a square detector is used which can provide 100% coverage of the ground.

4.6.2 Model of Line-scan Imaging Process

Fig. 4.6-1 illustrates the line-scan image gathering process of MISI where \( \text{comb}(x) \) and \( \text{comb}(y) \) representing the sampling process in the x and y direction. The ground radiance-field is scanned and sampled by the detector scan motion in the y-direction (track direction). The variation in the radiance field in the x-direction (scan direction) is converted to electrical signal, which is again processed and sampled by the electronics.

\[ s(x,y) = \left( \{ [K \ L(x,y) * \ \tau_o(x,y) * \ \tau_d(x,y)] \ \text{comb}(y) + n(x) \} * \tau_e(x) \right) \ \text{comb}(x) \]  

(4.6.2)

where the symbol * denotes the convolution operation, \( K \) is a responsivity of the detector, \( \tau_o(x,y) \) is the optical point spread function, \( \tau_d(x,y) \) is the detector point spread function.
\( \tau_e(x) \) is the impulse response function of the signal process electronics, \( n(x) \) is the noise of detector/pre-amplifier, and \( \text{comb}(x) \) and \( \text{comb}(y) \) are the sampling function with a sampling interval of 1.

\[
\text{comb}(x) = \sum_{n=-\infty}^{\infty} \delta(x-n)
\]

\[
\text{comb}(y) = \sum_{m=-\infty}^{\infty} \delta(y-m)
\]

The assumptions in Eq. 4.6.2 are that the noise is additive and both noise and electrical impulse response function are independent of \( y \). Both assumptions are valid for most line-scan type imaging systems.

The imaging process formulated by Eq. 4.6.2 can also be represented in the frequency domain \((u,v)\) which is usually computationally more efficient. Eq. 4.6.2 becomes

\[
\hat{S}(u,v) = [K \hat{L}(u,v) \hat{\tau}_o(u,v) \hat{\tau}_d(u,v) \hat{\tau}_e(u)] \ast \text{comb}(u,v) + \hat{n}(u) \hat{\tau}_e(u) \ast \text{comb}(u)
\]

(4.6.3a)

where

\[
\text{comb}(u,v) = \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} \delta(u-n, v-m)
\]

Eq. (4.6.3a) can be equivalently written in the form

\[
\hat{S}(u,v) = \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} K \hat{L}(u-n, v-m) \hat{\tau}_o(u-n, v-m) \hat{\tau}_d(u-n, v-m) \hat{\tau}_e(u-n) + \sum_{n=-\infty}^{\infty} \hat{n}(u-n) \hat{\tau}_e(u-n)
\]

(4.6.3b)

or

\[
\hat{S}(u,v) = K \hat{L}(u,v) \hat{\tau}_o(u,v) \hat{\tau}_d(u,v) \hat{\tau}_e(u) + \hat{n}(u) \hat{\tau}_e(u) + \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} K \hat{L}(u-n, v-m) \hat{\tau}_o(u-n, v-m) \hat{\tau}_d(u-n, v-m) \hat{\tau}_e(u-n) + \sum_{n=-\infty}^{\infty} \hat{n}(u-n) \hat{\tau}_e(u-n)
\]

(4.6.3c)

or

\[
\hat{S}(u,v) = \hat{S}_s(u,v) + \hat{n}_n(u) + \hat{S}_a(u,v) + \hat{n}_a(u)
\]

(4.6.3d)
where $S^n_s(u, v)$ and $n^n_n(u)$ are the continuous (i.e. unsampled) signal and noise frequency components, $S^n_a(u, v)$ and $n^n_a(u)$ are the sideband (aliased) components of the signal and noise that are generated by sampling. These aliased signal components can not be distinguished from the proper signal components, but they actually mask the spatial detail of the signal just as noise does.

### 4.6.3 Formulation: Information Channel Capacity of Line-Scan imaging System

The information channel capacity of an imaging system can be represented as (Fellgett and Linfoot 1955) (Huck and Park 1975):

$$h_i = \frac{1}{2} \int_B \frac{\Phi(u,v) + \Phi_n(u,v)}{\Phi_n(u,v)} \, du \, dv$$

(4.6.4)

where $\Phi(u, v)$ is the signal power spectrum, $\Phi_n(u, v)$ is the noise power spectrum, and $B$ is the sampling pass band. Note that the term inside the integral is actually the signal-to-noise ratio. The signal power spectrum is given by

$$\Phi(u,v) = K \Phi_L(u,v) |\tau_g(u,v)|^2$$

(4.6.5)

where $\tau_g$ is the transfer function of the image gathering system, given by

$$\tau_g(u,v) = \tau_o(u,v) \tau_d(u,v) \tau_e(u)$$

For most line-scan systems, the optics are nearly perfect as compared to the detector, so $\tau_o(u,v)$ is set to unity. The detector MTF is given by (Huck et al. 1980) (Katzberg et al. 1973)

$$\tau_d(u,v) = \frac{\sin(\pi au)}{\pi au} \frac{\sin(\pi bv)}{\pi bv}$$

(4.6.6a)

for a rectangular aperture

or

$$\tau_d(u, v) = \frac{\sin(\pi \frac{\sqrt{2}}{2} (au+bv))}{\pi \frac{\sqrt{2}}{2} (au+bv)} \frac{\sin(\pi \frac{\sqrt{2}}{2} (au-bv))}{\pi \frac{\sqrt{2}}{2} (au-bv)}$$

(4.6.6b)

for a diamond aperture
Figure 4.6-2 Detector Aperture Shape (a) Rectangular (b) Diamond

Figure 4.6-2 shows two types of detector apertures analyzed in this paper. The aperture parameters $a$ and $b$ are the effective width and height of the aperture size as illustrated in Figure 4.6-2. Figure 4.6-3 shows the MTF of two detector apertures. It is obvious that the MTF of the diamond aperture has smaller side lobes than the MTF of the rectangular aperture.

The electronic filter transfer function (Butterworth) can be represented by

$$
\tau_e(u) = \frac{1}{{\sqrt{1+(\frac{u}{u_c})^n}}} 
$$

(4.6.7)

where $u_c$ is the electronic 3 dB cutoff frequency, and $n$ is the order of the filter. Fig. 4.6-4 shows the transfer function of a electronic filter with different cutoff frequencies and filter orders.
The noise power spectrum consists of aliased signal, electronic noise, aliased noise (noise caused by under sampling of noise) and analog-to-digital conversion (ADC) noise. According to Eq. 4.6.3b, the aliased signal power spectrum is given by

$$\Phi_a(u,v) = \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} K_{L} (u-n,v-m)|\tau_g(u-n,v-m)|^2$$

$$\quad (u,v \in B) \quad (4.6.8)$$

The electronic noise consists of white noise (frequency independent noise) and colored noise (frequency dependent noise). The white noise is caused by detector shot noise and resistor noise (also called Johnson thermal noise). The white noise ($N_W$) is given by

$$N_w(f) = 4kT + 2i_b eA_d$$

where $k$ is the Boltzmann's constant, $T$ is the temperature in Kelvin, $i_b$ is the detector bias current, $e$ is the electron charge, and $A_d$ is the area of detector. For most high speed photovoltaic detectors (e.g., Si, InSb) using trans-impedance amplifiers, the limiting noise is the colored noise (called boosted voltage noise (Yang 1991)), which is given by

$$N_c(f) = 2\pi e_n C_d f A_d$$
where $e_n$ is the preamplifier input voltage noise, $f$ is the frequency in Hz, and $C_d$ is the detector junction capacitance. $A_d$ is included because the junction capacitance is proportional to the detector area. The Wiener spectrum of electronics noise becomes

$$
\Phi_{en}(u) = (4kT + 2i_e e A_d + 2\pi e_n C_d u \frac{2\pi f}{B} A_d \frac{2\pi f}{B} \tau_e(u))^2
$$

(4.6.9)

where $\Delta f$ is the electronic bandwidth in Hz, $B$ is the sampling bandwidth in cycles/sample, and $2\Delta f/B$ is unit conversion from Hertz to cycles/sample. The electronic noise is amplified and sampled together with the signal for digital transmission. Just as under sampling of the signal generates aliasing, undersampling of noise generates additional noise. Although this aliased noise is avoidable by proper shaping of the electronic frequency response, it could be significant as demonstrated later using the traditional design rules. According to Eq. 4.6.3b, the aliased electronic noise is defined as

$$
\Phi_{ea}(u) = \sum_{n=-\infty}^{\infty} \Phi_{en}(u-n)
$$

(4.6.10)

(u,v \in B)

The Wiener spectrum of the analog to digital conversion (ADC) noise is given by

$$
\Phi_{ADC}(u) = \frac{\sigma^2_{ADC}}{B}
$$

(4.6.11)

The total noise spectrum is given by

$$
\Phi_n(u,v) = \Phi_a(u,v) + \Phi_{en}(u) + \Phi_{ea}(u) + \Phi_{ADC}(u)
$$

(4.6.12)

Plugging Eq. 4.6.12 into 4.6.4 yields

$$
h_i = \frac{1}{2} \int_{-B}^{B} \int_{-B}^{B} \log(1 + \frac{\Phi(u,v)}{\Phi_a(u,v) + \Phi_{en}(u) + \Phi_{ea}(u) + \Phi_{ADC}(u)}) \, du \, dv
$$

(4.6.13)

In the next section, we are going to employ Eq. 4.6.13 to calculate the information channel capacity of a line-scan imaging system. The information channel capacity is used as a figure of merit for line-scan imaging system design tradeoff.
4.6.4. Calculation and Optimization

4.6.4.A. Assumptions and Imaging System Parameters for Computations

We use the same radiance field model as Huck et al. Assuming that the radiance field \( L(x, y) \) is a stationary process, and \( L(x, y) \) is a random set of two-dimensional pulses whose separation \( r \) obeys the Poisson probability density function with mean \( \mu_r \) and whose magnitude obeys the Gaussian probability density function with variance \( \sigma_L^2 \). The autocorrelation function \( R(x, y) \) and its associated Wiener spectrum are given by

\[
R(x, y) = \sigma_L^2 \exp(-\frac{r}{\mu_r}) \tag{4.6.14a}
\]

\[
\Phi_L(u, v) = \frac{2\pi \mu_r^2 \sigma_L^2}{[1+(2\pi\mu_r\rho)^2]^{3/2}} \tag{4.6.14b}
\]

where \( \rho = \sqrt{u^2+v^2} \) is the spatial frequency. Actual measurement confirms the generally exponential behavior of the autocorrelation function (Rittermam 1952). Fig. 4.6-4 show the Wiener spectra for scenes with different mean spatial detail \( (\mu_r) \). The smaller the mean separation \( (\mu_r) \), the more spread the corresponding Wiener spectrum.

Figure 4.6-5 Wiener Spectra of the Radiance Field
The following calculation is based on the MINSI's short-wave infrared channel (1.55-1.75 \(\mu m\)) with a spatial resolution of 2 mrad. The instrument parameters which are necessary for this calculation are listed in Table 4.6-1 (cf. Section 3.4).

Table 4.6-1. Scanner Instrument Parameters

<table>
<thead>
<tr>
<th>Detector type</th>
<th>InSb Photovoltaic</th>
</tr>
</thead>
<tbody>
<tr>
<td>Max. irradiance (E_{\text{max}})</td>
<td>2E-7 w/mm(^2)</td>
</tr>
<tr>
<td>Min irradiance (E_{\text{min}})</td>
<td>2.7E-8 w/mm(^2)</td>
</tr>
<tr>
<td>Detector responsivity K</td>
<td>1 A/w</td>
</tr>
<tr>
<td>Junction Capacitance (C_d)</td>
<td>400 pF/mm(^2)</td>
</tr>
<tr>
<td>Feedback Resistor (R_f)</td>
<td>300 k(\Omega)</td>
</tr>
<tr>
<td>Bias current (i_b)</td>
<td>1e-7 A</td>
</tr>
<tr>
<td>Preamplifier voltage noise (e_n)</td>
<td>4.5 E-9 V</td>
</tr>
<tr>
<td>Temperature (T)</td>
<td>300 K</td>
</tr>
<tr>
<td>Nyquist frequency (f_N)</td>
<td>115 kHz</td>
</tr>
<tr>
<td>Nyquist frequency (v_N)</td>
<td>0.5 cycle/sample</td>
</tr>
</tbody>
</table>

Since the magnitude of the radiance field is a Gaussian distribution, the variance of the scene is given by

\[
\sigma_L = \frac{(E_{\text{max}} - E_{\text{min}}) A_d}{6}
\]

where \(A_d = a \times b\) is the detector area in mm-square\(^*\). Assuming that an 8-bit analog to digital converter is used, the ADC noise is given by (cf. Table 4.6-1)

\[
\sigma_{ADC} = \frac{K E_{\text{max}} A_d}{\sqrt{12 \cdot 256}} = 2.25E-10 A_d
\]

The parameters used for design tradeoff are: (1) detector aperture parameters \((a, b)\), (2) electronic filter shape parameters (cut-off frequency \(f_C\) and filter order\(n\)).

We also use the design of Landsat Thematic Mapper as the baseline; the detector has a square aperture with size of the sampling interval and the electronic filter is a three-pole low-pass with cutoff frequency at the Nyquist frequency.

\* The signal level is proportional to detector area \((A_d)\), and noise is proportional to the square-root of \(A_d\) if the noise is white, thus the SNR is proportional to the square-root of \(A_d\). The SNR gain from \(A_d\) is not included in the analysis here because the noise of many detectors are not white.
4.6.4.B. Information Channel Capacity vs. Electronic Filter Parameters

The electronic filter can be used to shape the spectrum of signal and noise. Ideally, the transfer function of an electronic filter should be unity in the sampling passband (to retain the spatial detail) and 0 outside the sampling passband (to eliminate aliasing). But in reality, electronic filters can only be realized by an n-th order (or called n-pole) filter with the transfer function approximated by Eq. 4.6.7. The higher the filter order, the more complex the electronic circuit. We first consider the case that the mean spatial detail \( \mu_r = 1 \) and rectangular detector aperture with \( a = 1 \) and \( b = 1.5 \). The information channel capacity is plotted as a function of cut-off frequency for several filter orders in Fig. 4.6-6.

![Figure 4.6-6 Information channel capacity as Function of Electronic Cut-off Frequency](image)

The result in Fig. 4.6-6 shows that (1) as the filter order increases, the information channel capacity increases. At the Nyquist cutoff frequency, the information channel capacity varies from 2.8 bit/sample for \( n = 2 \) to 3.5 bit/sample for \( n = 20 \); (2) for every filter, there exists a cutoff frequency range from 0.2 for \( n = 2 \) to 0.45 for \( n = 20 \), which will maximize the information channel capacity. When the filter order approaches infinity, the informationally maximized cut-off frequency approaches the 3 dB cut-off frequency of the traditional design; and (3) there exists a tradeoff between filter order and cutoff frequency such that a lower-order filter can approach the information channel capacity of the ideal filter (\( n = \infty \)) by choosing a lower cutoff frequency (This is essentially a tradeoff between aliasing and blur). This result also shows that, for most practical filters (\( n < 5 \)).
using the Nyquist frequency as the cutoff frequency as specified by the traditional design rule will result in loss of information channel capacity of the imaging system.

Although this analysis is for µ=1 and for a specific rectangular aperture, the conclusion is valid for other radiance fields and apertures as well. Since 6th-order electronic filters can be realize without much difficult by cascading two active filter blocks, we choose the informationally optimized electronic filter to be a 6-pole low-pass with a cutoff frequency of 0.4 cycle/sample.

4.6.4.C. Information Channel Capacity vs. Detector Aperture

As shown in Fig. 4.6-3, the detector aperture acts as a low-pass filter. As a matter of fact, the aperture is the only protection against aliasing in the track direction. By varying the aperture size, we can tradeoff aliasing with blurring. Fig. 4.6-7 shows the information channel capacity as a function of aperture size.

The result shows that: (1) all three lines are close together, which means that the information channel capacity is not strongly dependent on the aperture width. This is because the frequency response in the scan direction is shaped pretty much by the electronic filter; (2) For the rectangular aperture, there exists an optimum b (aperture height) value of 1.5 which will have maximum information channel capacity, while for the diamond shaped aperture, the curves tend to flatten out after b ≥ 1.5; and (3) the diamond shaped aperture tends to have higher information channel capacity than the rectangular shape.

We conclude that for both the rectangular aperture and diamond aperture, the informationally optimized aperture should be a=1 and b=1.5.
4.6.5 Comparison: Informationally Optimized Design vs. Traditional Design

In this section, we are going to compare three designs: traditional, informationally optimized with rectangular aperture, and informationally optimized with diamond aperture. We will discuss the MTF, signal-to-noise ratio information, and the inevitable tradeoff between blur and aliasing.

4.6.5.A. Modulation Transfer Function

The modulation transfer function of three designs is plotted in Fig. 4.6-8 and 4.6-9, where $\text{MTF}_\text{trad}$, $\text{MTF}_\text{rect}$, and $\text{MTF}_\text{diad}$ represent the MTF for the traditional design, informationally optimized rectangular aperture, and informationally optimized diamond aperture respectively.
Figure 4.6-8 MTF of Three Designs in the Scan Direction

In the scan direction, the Nyquist MTF is about 0.47 for the traditional design and about 0.18 for the informationally optimized design. The MTF is almost zero beyond the frequency of 1 cycle/sample. In the track direction, the Nyquist MTF is about 0.6 for the traditional design, and about 0.25 for the two informationally optimized designs. But the side lobes for the rectangular apertures are quite significant, which will introduce aliasing if the scene is not band-limited. The blur is about the same for the two informationally optimized design, but the traditional design has much less blur. The diamond shaped aperture suppresses the high frequency aliasing much better than the rectangular aperture, thus the diamond aperture provides much better protection against aliasing.
4.6.5.B. Signal-to-Noise Ratio

Assuming the mean spatial detail of the radiance field $\mu_t=1$, the power spectra of the signal and the aliased signal for both the scan and track direction are plotted in Fig. 4.6-10 and 4.6.11. It is apparent that the aliased signal is quite significant for the rectangular apertures. The magnitude of the aliased signal is equal to the signal itself at the Nyquist frequency for all three designs, which suggests that it does not make much sense to specify the Nyquist MTF in the information sense, because the signal-to-noise ratio at Nyquist is always below 1.
Figure 4.6-10 The Power Spectrum of Signal and Aliased Signal in the Scan Direction

Figure 4.6-11 The Power Spectrum of Signal and Aliased Signal in the Track Direction

Fig. 4.6-12 is the SNR plotted in 3-D as a function of spatial frequency in the scan and track direction. The SNR is mostly aliasing limited at high frequency. The low SNR values at lower frequency for the informationally optimized rectangular aperture is caused by aliasing due to the second lobe in the track direction MTF. It is easy to see the overall SNR of the diamond shaped aperture is significantly better than the rectangular apertures.
4.6.5.C. Information Channel Capacity

The variation of information channel capacity $h$ with mean spatial detail $\mu_r$ is plotted in Fig. 4.6-13. For all three designs, information channel capacity is maximized at the mean spatial detail ($\mu_r$) matching the sampling interval. The informationally optimized design can always gather information better than traditional design over a broad range of radiance fields, and the informationally optimized design with diamond aperture can get up to 1.5 bit more information than the traditional design.

4.6.6 Digital Simulation of Information Analysis

The information analysis reveals that the traditional design of maximizing MTF in the sampling passband results in lower information channel capacity compared to the
informationally optimized design. In this section, the image-gathering process was simulated digitally for both the traditional design and the informationally optimized design with diamond shaped aperture. The image quality was evaluated using images fidelity and edge detection algorithms.

The image used in the simulation is a scanned analytical test pattern as shown in Figure 4.6-14. It consists of radial bar patterns and vertical and horizontal bars of varying frequency. The image-gathering process was simulated as follows:

1. A Gaussian random noise of standard deviation $\sigma=1/$SNR was added to the original image.
2. The original image was transformed into frequency space by FFT.
3. The spectrum of the image was multiplied by the OTF of the image gathering process.
4. The filtered spectrum was shifted vertically and horizontally by half of the original Nyquist frequency to simulate 2x sampling process as described by Eq. 4.6.3a.
5. Reconstruct the sampled image by multiplying the frequency response of the reconstruction filter.
6. Compare the reconstructed image with the original and calculate the fidelities.

![Test Image Used in Digital Simulation](image)

The main differences between the two designs are the OTF of the image-gathering system and the reconstruction process. The OTFs of the image-gathering process are described in Section 4.6.5A and plotted in Figure 4.6-8 and Figure 4.6-9. The reconstruction filter used for the traditional design is a $\text{sinc}^2$ filter (i.e. simple linear interpolation method).
The reconstruction for information optimized design is a modified Wiener filter [Huck et al. 1988], which is given by

$$
\psi_C(u,v) = \frac{\Phi_L(u,v) \tau_g(u,v)}{\Phi_L(u,v) |\tau_g(u,v)|^2 + \left[ \Phi_L(u,v) |\tau_g(u,v)|^2 * \sum_{n,m \neq 0} \frac{K \sigma_L}{\sigma_n} \right]^{2}}
$$

(4.6.1)

$\Phi_L$ is given by Eq. 4.6.14b with $\mu_r = 3$, and $\tau_g$ is the OTF of image-gathering process. The first term in the denominator is the aliased spectra and the second term is noise-to-signal ratio. The SNR used in the simulation is 50. Figure 4.6-15 shows the OTF and Wiener filter in x and y directions.

![OTF and Wiener Filter](image)

**Figure 4.6-15 OTF and Wiener Filter of the Information Optimized Design**

shauzly Figure 4.6-16 shows the reconstructed images. The aliasing caused by insufficient sampling is greatly reduced in the information optimized design. The fidelity (the complement of the normalized mean-square difference) is 0.913 for the traditional design and 0.941 for the information optimized design.
The image quality of these simulated images was also evaluated by extracting the edges from these images. The Sobel edge detection operators were applied to the original image and the simulated images [Gonzalez and Woods, 1992]. Figure 4.6-17 shows the edge detection maps. The images from the information optimized design have much better edge detection than the images from the traditional design. This simulation confirms the conclusion of Huck et al. [1988] that: "the combined process of image gathering and restoration behaves more as a communication channel in that informationally optimized design of the image-gathering system tends to maximize the fidelity of a variety of optimally restored representations ranging from images to edges".
Following the work of Huck et al., we formulated the information channel capacity of an airborne line-scan imaging system. The information criterion is attractive because it includes the effect of the spatial response (MTF), signal-to-noise ratio, sampling, and the statistical properties of the radiance field. By applying this information criterion to the design of an airborne line-scanner system, we derived the optimum electronic filter and detector aperture shape which will maximize the information channel capacity of the imaging system. Our analytical results show that: (1) the traditional design does not optimize the image-gathering system informationally, and (2) an informationally optimized design can gather up to 1.5 bit more information than the traditional design.

The digital simulation indicates that by combining image gathering and image processing, the information optimized design did maximize the fidelity of the image and edges. With the advent of digital imaging processing and computer vision, information criterion gives a promising alternative to the image quality metrics used today.

The information analysis presented here is based on a line-scan imaging system similar to the MISI and TM sensors, where the aperture shape of the detector can be non-rectangular. Its extension to push broom sensor is straight-forward as shown in Figure 4.6-17 Edge Extraction.
but the diamond shaped linear sensor will be costly to fabricate. The extension to a 2-D staring array is nearly impossible because of the overlapping aperture required for the information optimized design.

According to Huck et al. [1985], the information optimized design is to be preferred when computer processing is used. But even today, visual assessment of image is still one of the most important means of extract information from images. Traditionally, human visual system can accommodate the artifacts of aliasing but can not tolerate blur, which is in contradiction to the information optimized design where aliasing is treated as noise. This is probably caused by the fact that most images were not restorated digitally, and the images of lower aliasing appear dull (due to the lower MTF at the high frequency), and spatial details may fall below the visual detection threshold. McCormick et al. [1989] shows that by combining of digital image gathering and digital restoration, information optimized design is also optimized for visual image quality.
This thesis describes a method of design, modeling, analysis, and testing of a modular imaging spectrometer instrument (MISI), with special emphasis on system and subsystem MTF analysis. The optical system was modeled using optical ray tracing methods. The dynamic deformation of the scan-mirror was model using finite element analysis methods, and the image degradation due to the deformation was estimated using optical image formation theory. The detector and conditioning electronics were also modeled using the transfer function theory. This modeling approach was used as a tradeoff tool for the design of MISI. Laboratory experiments were conducted to test the performances of each sub-system on design criteria, and finally overall system performances were tested against the design specifications. The measured PSF confirms that the spatial resolution of MISI is primary detector limited as designed. The MTF derived from LSF and ESF agrees very well with the MTF predicted from the modeling of optical/electronic subsystems. The measured noise equivalent reflectance is below 0.1% for Si broad-band detectors and most of the spectrometer channels. The measured noise equivalent temperature is around 0.1 K. All these results indicate that MISI has achieved nearly all the image quality goals.

By combining FEA methods and optical image formation theory, the performance of a high-speed scan-mirror was modeled. The modeling approach bridges the gap between mechanical design and optical image quality.

A model-based algorithm was developed to derive the modulation transfer function of a imaging system from the edge spread function. The algorithm makes use of prior knowledge about the imaging system to be measured, thus it can derive MTF from very noisy edge trace data. By combining a third order Hermite function, detector aperture function, and electronic filter frequency response function, the algorithm can model the OTF of almost all electro-optical imaging systems. Digital simulation results indicate that this method is very robust against noise, and a comparison study between this method and the more accurate square-wave method yields very good agreement. Application of this method to a test instrument showed that this method is robust against not only random noise, but also low frequency periodic noise. This method would be an ideal choice to measure the MTF of operational instruments where perfect high contrast edge is not available, or in the test environment, where the instrument signal-to-noise ratio has not been optimized.
And finally, an alternative image quality metric - information criteria is evaluated. The information criteria lets us assess the image quality independent of process. It includes the effects of signal-to-noise ratio, spatial resolution, and sampling. It seems that this is the only appropriate criteria for a sampled imaging system. We formulated the information channel capacity of MISI, and found that the traditional EO image system design rules are not optimized for information. The information optimized design can provide up to 1.5 bits more information than the traditional design. Digital simulation confirms that by combining image-gathering and digital restoration, the informationally optimized EO system tends to maximize the fidelity of images and edges. We believe that with the advent of digital imaging processing and computer vision, the information criteria provide a very promising alternative to the traditional EO image quality metrics.
References


Blackman, E.S., "Effects of noise on the determination of photographic system modulation transfer functions", J. Phot. Sci. Eng., 12, 244-250 (1968).


Koger, David. “Landsat 6 images to provide more remote sensing muscle.” THE OIL AND GAS JOURNAL, 02/01/93 91 (5 1993):


Schott, John R. Remote Sensing –A Imaging Train Analysis ??


Appendix I  Specifications for the LW focal plane

I.A  LW detector #1

DETECTOR DATA SHEET

CUSTOMER NAME: RIT
SERIAL NUMBER: 90064
MODEL NUMBER: B12E-40-5A

CONTRACT NO.: A-0672
WINDOW TYPE: Germanium
FLUX: $1.76 \times 10^{-5}$ W/cm²

DETECTOR # 1

LENGTH: 1 mm  RESISTANCE (ROOM TEMP): 48.3 Ω  PEAK 10.9 μm
WIDTH: 1 mm  RESISTANCE (77K): 133.4 Ω  CUTOFF(50%): 11.3 μm
Platinum Diode (Room): 700  CUTOFF(20%): 11.5 μm
Temperature Sensor (77K): 1046

Specifications given with Filters

<table>
<thead>
<tr>
<th>BIAS CURRENT (mA)</th>
<th>I=5</th>
<th>I=10</th>
</tr>
</thead>
<tbody>
<tr>
<td>PEAK RESPONSIVITY (V/W)</td>
<td>7,100</td>
<td>14,200</td>
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<tr>
<td>D* PEAK</td>
<td>1 kHz</td>
<td>8.0 x 10¹⁰</td>
</tr>
<tr>
<td></td>
<td>10 kHz</td>
<td>1.2 x 10¹¹</td>
</tr>
</tbody>
</table>

***Graph***

MICRONS
**CUSTOMER NAME:** RIT  
**SERIAL NUMBER:** 90064  
**MODEL NUMBER:** B12W-40-5A  
**CONTRACT NO.:** A-0672  
**WINDOW TYPE:** Germanium  
**FLUX:** $1.76 \times 10^{-5}$ W/cm$^2$

### DETECTOR #2

| LENGTH: .5mm | RESISTANCE (ROOM TEMP): 64.2 Ω | PEAK: 11.1 μm |
| WIDTH: .5mm | RESISTANCE (77K): 112.3 Ω | CUTOFF (50%): 13.7 μm |
|             | Platinum Diode (Room): 700 | CUTOFF (20%): 14.1 μm |
|             | Temperature Sensor (77K): 1046 |              |

**Specifications given with Filters**

<table>
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<tr>
<th>BIAS CURRENT (mA)</th>
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<th>$D*_{PEAK}$ 10kHz</th>
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</thead>
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<tr>
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<td>I=10</td>
<td>11,800</td>
<td>$6.1 \times 10^{10}$</td>
<td>$8.2 \times 10^{10}$</td>
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![Graph showing spectral response with microns on the x-axis and intensity on the y-axis.](image)
CUSTOMER NAME: RIT
SERIAL NUMBER: 90064
MODEL NUMBER: B12H-40-5A

CONTRACT NO.: A-0672
WINDOW TYPE: Germanium
FLUX: $1.76 \times 10^{-5}$ W/cm$^2$

**DETECTOR #3**

| LENGTH: .5 mm | RESISTANCE (ROOM TEMP): 44.8 Ω | PEAK 11.1 μm |
| WIDTH: .5 mm | RESISTANCE (77K): 126.4 Ω | CUTOFF(50%): 14.0 μm |
|              | Platinum Diode (Room): 700 | CUTOFF(20%): 14.4 μm |
|              | Temperature Sensor (77K): 1046 |                    |

Specifications given with Filters

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<th>BIAS CURRENT (mA)</th>
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<th>I=5</th>
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<td>21,000</td>
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<td>$D^* \text{PEAK}$ (1kHz)</td>
<td>$6.3 \times 10^{10}$</td>
<td>$6.2 \times 10^{10}$</td>
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<tr>
<td>$\frac{cm}{\sqrt{Hz}}$ (10kHz)</td>
<td>$8.5 \times 10^{10}$</td>
<td>$8.4 \times 10^{10}$</td>
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</table>

**MICRONS**

161
I.D LW detector #4

CUSTOMER NAME: RIT
SERIAL NUMBER: 90064
MODEL NUMBER: B12M-40-5A

CONTRACT NO.: A-0672
WINDOW TYPE: Germanium
FLUX: $1.76 \times 10^{-5} \text{ W/cm}^2$

DETECTOR # 4

| LENGTH: 1mm | RESISTANCE (ROOM TEMP): 51.4 $\Omega$ | PEAK 10.9$\mu$m |
| WIDTH: 1mm | RESISTANCE (77K): 114.8 $\Omega$ | CUTOFF(50%): 11.9$\mu$m |
| Platinum Diode(Room): 700 | CUTOFF(20%): 12.2$\mu$m |
| Temperature Sensor (77K): 1046 |

Specifications given with Filters

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<td>cm$^2$/√Hz 1kHz</td>
<td>$7.0 \times 10^{10}$</td>
<td>$6.9 \times 10^{10}$</td>
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<tr>
<td>10kHz</td>
<td>$1.1 \times 10^{11}$</td>
<td>$1.0 \times 10^{11}$</td>
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CUSTOMER NAME: RIT
SERIAL NUMBER: 90064
MODEL NUMBER: B12W-40-5A

CONTRACT NO.: A-0672
WINDOW TYPE: Germanium
FLUX: 1.76 x 10^{-5} W/cm²

DETECTOR #5

LENGTH: 1mm  RESISTANCE (ROOM TEMP): 71.2 Ω  PEAK 13.1μm
WIDTH: 1mm  RESISTANCE (77K): 160.7 Ω  CUTOFF (50%): 14.4μm

Platinum Diode (Room): 700  CUTOFF (20%): 14.9μm
Temperature Sensor (77K): 1046

Specifications given with Filters

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<td>D* PEAK</td>
<td>5.2 x 10^{10}</td>
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<tr>
<td>1kHz</td>
<td>6.8 x 10^{10}</td>
<td>6.7 x 10^{10}</td>
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MICRONS
New Detector Elements

Detector #2 and #3 were replaced with two new elements with matched performance in November 1995. The manufacturer tested the D* and responsivity of all the detectors. The table below lists the performance tested on 11/21/95. Spectral response was not given for this test. The assumption is that they are the same as before since the filter sets were not changed.

**DETECTOR REPORT**

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<th>RIT</th>
<th>Serial Number:</th>
<th>90895</th>
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<td>814</td>
<td>Model Number:</td>
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<td>Type:</td>
<td>HgCdTe</td>
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<tr>
<td>Date:</td>
<td>11/21/95</td>
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<td>12</td>
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<td>Window type:</td>
<td>Ge</td>
<td>FLUX:</td>
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<td>.5 mm</td>
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<table>
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<td>$8 \times 10^{10}$</td>
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Appendix II Specifications for the Si diode array

Electrical characteristics of Hamamatsu S4114 Series Photodiode Arrays

<table>
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<th>Package</th>
<th>Sensitive Surface (per element)</th>
<th>Spectral Response</th>
<th>Characteristics (25°C, per element)</th>
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</thead>
<tbody>
<tr>
<td></td>
<td>Size (mm) Effective Area (mm²)</td>
<td>Number of Elements</td>
<td>Peak Wavelength (nm)</td>
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<tr>
<td>48 pin DIP</td>
<td>4.4 x 0.9 3.96</td>
<td>38</td>
<td>190 to 1000 800</td>
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<table>
<thead>
<tr>
<th>Characteristics (25°C, per element)</th>
<th>Maximum Ratings</th>
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<td>Shunt Resistance</td>
<td>Junction Capacitance</td>
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<tr>
<td>Min.</td>
<td>Typ.</td>
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<td>0.15</td>
<td>2</td>
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</table>

Figure 1: Typical Spectral Response

Figure 2: Typical Temperature Coefficient of Sensitivity

Figure 3: Example of Cross-talk (λ = 655nm, Vₐ=0V, Ta = 25°C)
Appendix III Simplified voltage amplifier

all resistor .1% mf 1/8 watt

c1 = 2uf

c2 = .1nf polystyrene or teflon

c = .01uf

rd = 180
r2 = 2k
r1 = 375
r3 = 2k pot
r4 = 100 pot
visheyl260

z = 1n4104@ 10v

1n4105@ 11v