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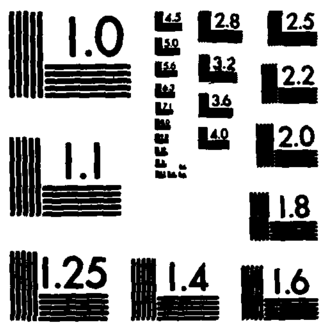
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TELEMETRY BANDWIDTH SELECTION FOR INFRARED SURVEILLANCE SENSORS

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This technical report has been reviewed and is approved for publication

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RANDALL E. MURPHY, Director
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Information theory, modelling and computer simulation are combined in this study to minimize the telemetry/recorder bandwidth requirements for IR sensors being developed by AFGL. The study included development and documentation of a computer code for optimizing the design of choppers, filters and electronics for IR radiometers.

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PREFACE

This work was funded by the Air Force Geophysics Laboratory Director's Fund (as Project No. LDF ILIR2J), and was performed under Utah State University Subcontract No. 82-047. We are grateful to Col. John Friel, who was the AFGL Commander during the project, and to Dr. A. T. Stair, Jr., Chief Scientist at AFGL, for their interest and encouragement. The expert computer programming skills of Mrs. Effi Huber-Buser are gratefully acknowledged. We also wish to thank Dave Newell of AFGL for his expertise in setting up a laboratory experiment to test the sampling methods addressed in this study.

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SECTION 1
INTRODUCTION AND SUMMARY

Several Air Force measurement programs presently in the hardware design stage, such as the CIRRIS 1A Space Shuttle program, the DNA sponsored SPIRIT probe experiment and the SD probe program, will obtain IR radiometric and spectral data which directly support the measurement needs of the Space Defense and Missile Surveillance Programs. However, the current capacity of state-of-the-art digital telemetry/recording systems is inadequate to handle the large data volume and high data rates associated with these experiments. The non-availability of a high speed telemetry system for Space Shuttle over the next few years has already resulted in undesirable reductions in planned experiment capabilities. The problem is compounded by the non-availability of data transfer satellites, by the limited number of ground stations, and by outages and reductions in ground contact times caused by Space Shuttle gravity gradient operations required for obtaining high quality background measurements.

More efficient methods of data compression are required to avoid further compromise of experiment objectives. On-board data processing, where feasible, analog signal conditioning, and selection of an optimum sampling rate are all potentially useful methods for minimizing data transmission rates. A significant reduction in rate would enhance the return from Air Force sensors that require data telemetry with minimum ground contact times.

The infrared radiometers and Fourier Transform Spectrometers (FTS) to be used in the cited Air Force programs will produce encoded waveforms that will be sampled, telemetered to ground and subsequently decoded, that is demodulated and/or analyzed for spectral content The radiometers use optical chopping to avoid problems associated with low frequency operation, such as detector 1/f noise and amplifier drift. The FTS produces an optically modulated signal, the interferogram, which is the Fourier transform of the desired IR radiation spectrum. Determination of sampling frequency and bandwidth requirements is fairly straightforward for the FTS, since the minimum and maximum frequencies in the interferogram and method of decoding are usually well defined. Similar requirements are more difficult to establish for the radiometers because the signals they will measure arise from spatio-temporal scene variations that are largely unknown, and because the chopping waveform is not known prior to detailed specification and/or design of the instrument, and because the effects of different types of analog signal conditioning and demodulation cannot be determined without extensive modelling and analysis. Because of these uncertainties, the number of data values recorded or transmitted in radiometric measurement programs has traditionally exceeded by far the number actually needed to define the input (or its frequency content). This is indicated by the two different paths in Fig. 1 that trace the flow of information from initial signal acquisition to the display of final results.

The objective of the present study has been to minimize the telemetry/recorder bandwidth requirements for IR sensors currently being designed at AFGL or under AFGL direction, and to provide the tools needed to accomplish the same optimization in future programs. This objective has been realized by following a technical approach that combines modelling, application of information theory and computer simulation (see Fig. 2). The study included development and documentation of an extensive computer code for optimizing sensor design. Included in the code package are subroutines representing a particular scene Power Spectral Density (PSD), sensor Modulation Transfer Functions (MTF's) and various chopping waveforms. These, or substitute models provided by the code user, comprise the modelling branch of the tree in Fig. 2. The study developed optimum sampling rules that guide the selection of code input parameters defining an initial sensor design. These represent the information theory branch.

The computer code developed in the study simulates the effects of (a) scanning a radiometric scene with a finite-width detector, (b) modification of the scene by the optics MTF, (c) chopping, (d) electrically filtering the chopped output, and (e) sampling the resultant signal. It also simulates analog (on-board) demodulation before sampling, or digital demodulation of the samples. The code produces plots comparing the power spectrum

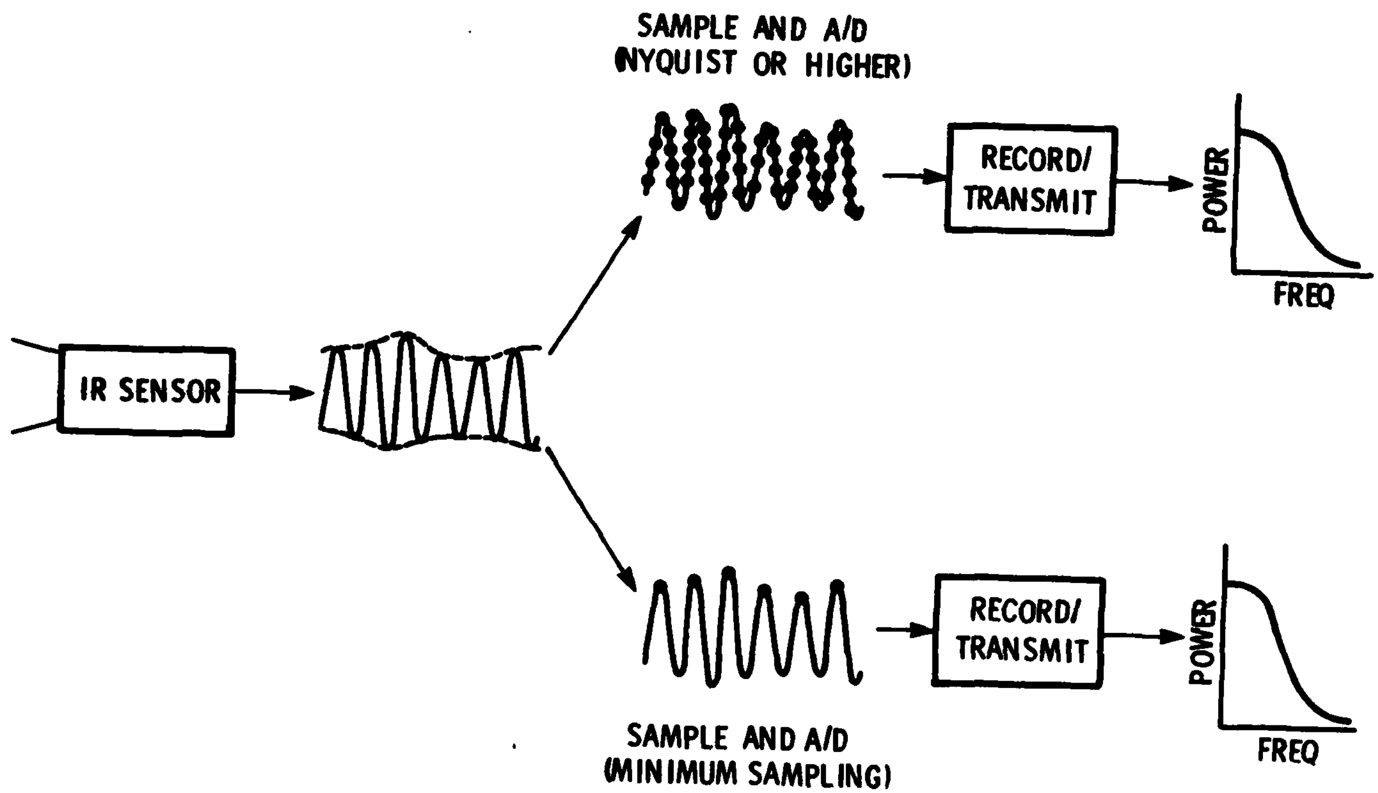

Figure 1. Signal Acquisition, Sampling and Display for a Chopped IR Sensor

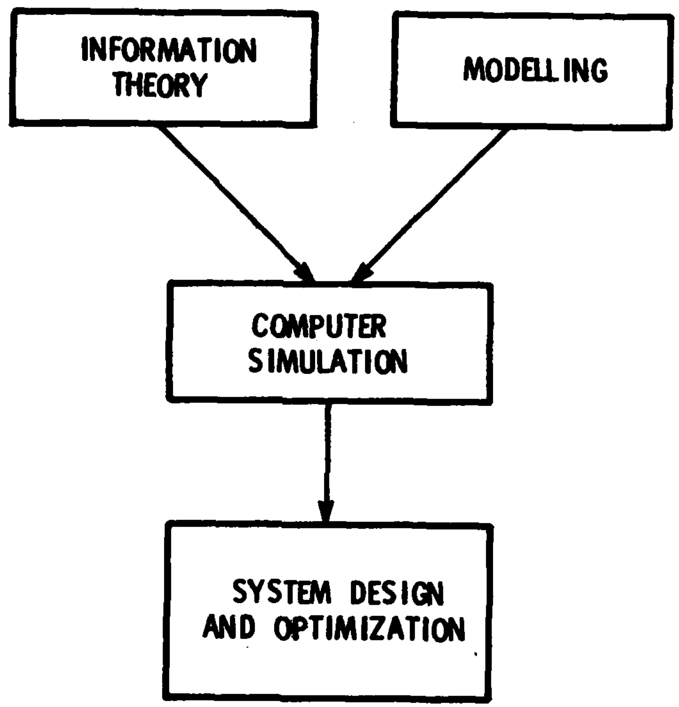

Figure 2. Technical Approach

(or amplitude spectrum) of the input radiometric signal and the spectrum of the demodulated output. Any significant differences between the two spectra generally indicate the need to revise the sampling and/or chopping frequencies, the filter design or the chopper design.

This LDF study provided some immediate payoffs. The code was used to redesign the sampling electronics of the ELIAS radiometer and to demonstrate that this redesign has substantially increased its measurement accuracy. The code was also used in the design of the chopper, filters and electronics of the CIRRIS-1A radiometer. It was found that with an optimized design only two tape recorders, rather than the planned four, would be required for 32 hours of data. This represents an estimated $200K-500K savings in flight hardware costs, and a significant reduction in the time and cost of post-flight analysis.

Section 2 of this report describes the general problem of choosing an optimum sampling rate for a chopped signal. Section 3 gives the optimum sampling rules in equation form and also as a nomograph. Section 4 describes the computer code for evaluating residual aliasing effects, and shows some results obtained during the design of the CIRRIS-1A radiometer.

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SECTION 2
THE ALIASING PROBLEM

This study is concerned mainly with the sampling of signals that have been chopped -- in effect, multiplied by a positive definite periodic waveform. The purpose of chopping is to beat the signal up to higher frequency, thereby avoiding problems associated with low frequency operation, such as detector 1/f noise and amplifier drift. Chopping alters the frequency spectrum of the original signal by introducing additional components at the chopping frequency and at all multiples of the chopping frequency. Perfect chopping, in which the chopping waveform is a constant plus a sunusoid, is an exception in that no harmonics are produced; i.e., the resultant spectrum is the original spectrum plus this spectrum shifted to the chopping frequency.

It is often desirable to electrically bandpass the chopped signal to a frequency range centered on the chopping frequency. The object is to suppress the introduced harmonics as well as the original low frequency signal components. These would otherwise give rise to additional, large alias components when the signal is sampled. The effects of aliases of the residual chopping harmonics will be discussed later in this section. We will first review the general effects of sampling a bandlimited signal.

Suppose Fig. 3a represents the amplitude spectrum of an input signal that will not be chopped, only sampled. The effect of sampling is to produce additional spectrum components (aliases) which are equal to the original spectrum shifted by all multiples (±1, ±2, ±3, ...) of the sampling frequency. Note that the spectrum of the original (real) signal is considered to have negative as well as positive frequencies; its real part is even about zero frequency and its imaginary part is odd. Thus, for a sampling rate greater than twice the highest original frequency, the amplitude aliases would have the positions indicated in Fig. 3b. The original signal can be recovered from the samples since none of the aliases overlap

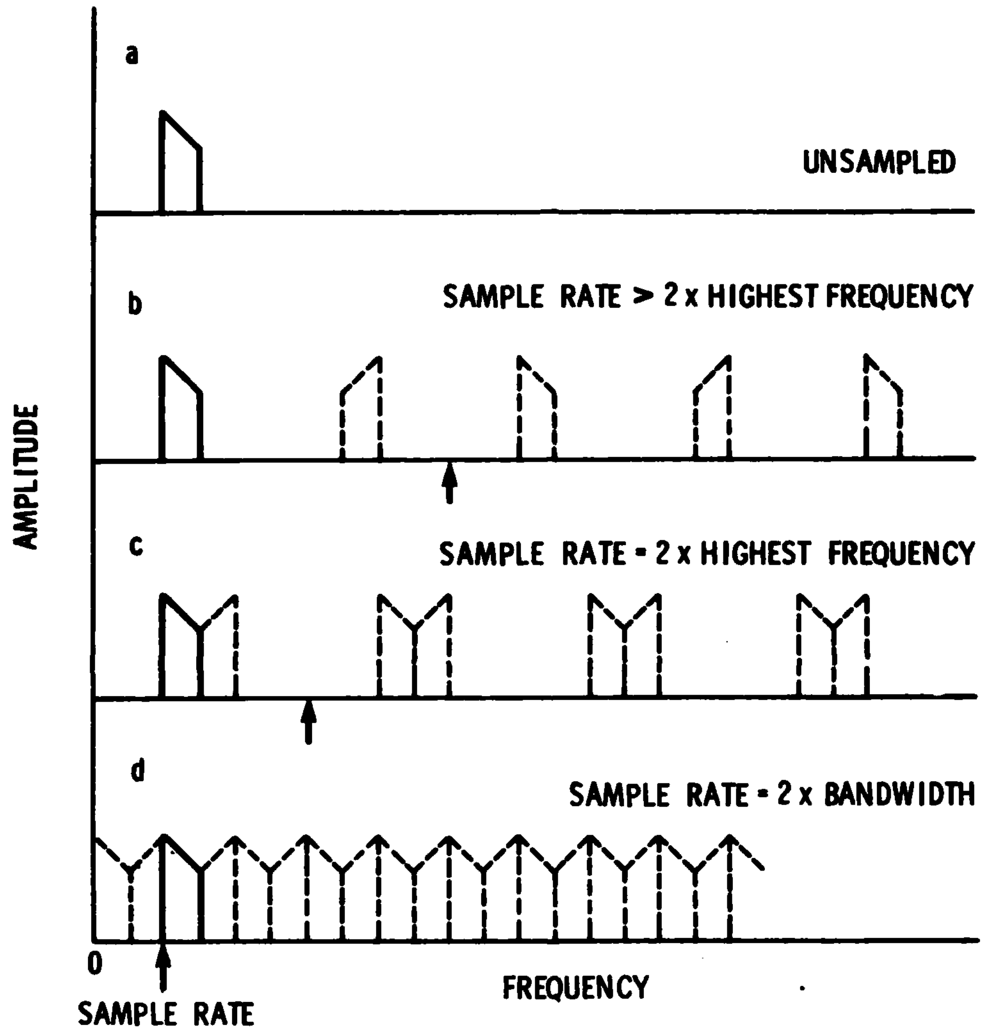

Figure 3. Spectrum of Sampled Bandlimited Signal

the signal band. In Fig. 3c the sampling rate is exactly the Nyquist rate, or twice the highest original frequency. It is seen that an alias abuts the original signal band. Again, the original signal can be recovered from the samples.

The signal represented in Fig. 3 has a "sparse" spectrum. That is, there is room between zero and the lowest frequency where aliases can occur that do not overlap the signal band. For the particular spectrum shown (Fig. 3a), the sampling rate can be twice the spectrum bandwidth, which is less than twice the highest frequency. The resulting aliases, shown in Fig. 3d, do not overlap the signal band.

It should be cautioned that it is not always possible to avoid the aliasing problem by choosing the sampling rate equal to twice the spectrum bandwidth. The spectrum shown in Fig. 3a is a special case in that there is room for an integral number of aliases below the lowest frequency. Generally it is safe to sample at twice the bandwidth when the lowest original frequency is exactly a multiple of the bandwidth. A general rule for sampling signals having sparse spectra with arbitrary maximum and minimum frequencies is given in Section 3. It is clear that a signal with a "very sparse" spectrum, i.e., with a bandwidth much less than the highest frequency, can be sampled at a rate equal to or only slightly higher than twice the bandwidth. This rate will be considerably less than the Nyquist rate.

The minimum sampling rule given in Section 3 for sparse spectra of the type shown in Fig. 3a will generally result in equi-spaced aliases like those pictured in Fig. 3d, although there may be space between the aliases depending on the bandwidth and sampling frequency.

In infrared Fourier Transform Spectroscopy (FTS), the spectra which are obtained (by taking the Fourier transform of a measured interferogram) are like the one shown in Fig. 3a. That is, they have a non-zero minimum frequency corresponding to the longwave cutoff of the IR detector or an optical filter and no particular symmetry about the median frequency in the sparse spectrum. Thus the sampling rule in Section 3 which applies to this type of spectrum (and ensures no overlap of the signal band and its aliases) can be used to select the interferogram sampling rate, Consider the amplitude spectrum shown in Fig. 4a. Unlike the one in Fig. 3a, it is symmetrical about its median frequency. This type of sparse spatial frequency spectrum can result from chopping the input of a scanning radiometer and then bandpass filtering around the chopping frequency (which becomes the median frequency). We are assuming that the bandpass filter has removed totally the second and higher harmonics due to chopping, and has also removed totally the spectrum of the unchopped input centered at zero frequency. The spectrum after chopping and ideal bandpass filtering has the same symmetry properties about the chopping frequency that the original unchopped spectrum had about dc, namely, an even real part and an odd imaginary part.*

These symmetry properties allow sampling at a rate that produces one or more aliases in exact registration with the signal band (spectral band shown in Fig. 4a). For the particular case represented in Fig. 4, in which the chopping frequency is 2.5 times the signal bandwidth, one can sample at a rate equal to the bandwidth. The resulting aliases are indicated in Fig. 4b. Note that this rate is half the minimum rate allowed when the spectrum does not have the requisite symmetry properties about its median frequency.

Figure 4 represents a special case in that there is room for an integral number of aliases below the lowest signal frequency. Thus, it is not always true that a sampling rate equal to the bandwidth will avoid aliasing problems for a chopped, ideally filtered signal. Generally, it is safe to sample at a rate equal to the bandwidth when the chopping frequency is an integer or half-integer times the bandwidth. A simpler way of saying this is that the samples may be taken at time intervals equal to an integer or half integer times the chopping period, provided this interval does not exceed the reciprocal of the bandwidth. The complete sampling rule is given in Section 3.

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*We define an "ideal" bandpass filter as one that has Hermitian symmetry; i.e., even amplitude and odd phase about its bandpass center.

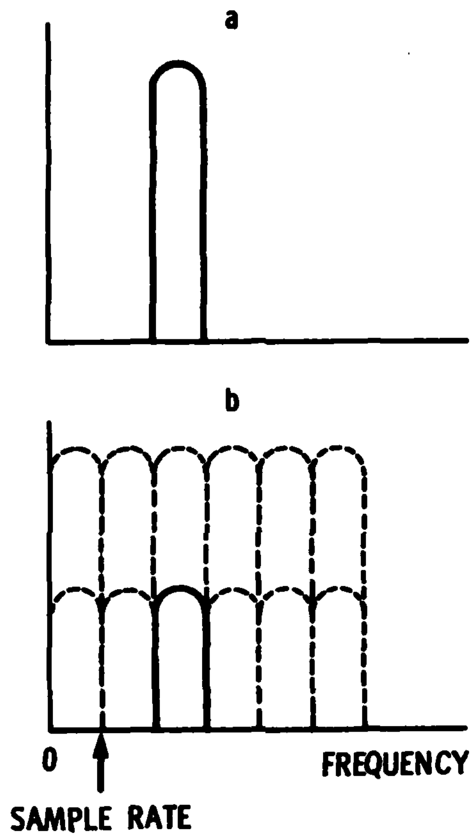

Figure 4. Sampling of a Bandlimited Signal with Symmetry about the Band Center

Figure 5 displays the combined effects, in the spectral domain, of chopping and sampling. The top panel shows partially the spectrum after chopping and before sampling. It consists of the original signal spectrum plus the fundamental, second and third harmonics introduced by chopping. The region of the spectrum from which information will be extracted (after sampling) is normally the region encompassing the fundamental.

The second panel in Fig. 5 shows the aliases introduced by sampling at a rate equal to twice the nominal highest frequency (the frequency at which the fundamental begins to roll off). The region of the fundamental is overlapped by an alias of the fundamental and two aliases of the third harmonic. This overlap represents an aliasing problem, i.e., a bad choice for the sampling frequency. Better choices are:

a.) a slightly lower frequency, equal to twice the chopping rate. This would put the three alias components mentioned above in exact registration with the fundamental.

b.) a somewhat higher frequency such that the alias of the fundamental has no overlap with the actual fundamental. This choice would require a lowpass filter that rejects the third chopping harmonic, assuming its alias still overlaps the fundamental after revising the sampling frequency.

c.) a sampling frequency that results in aliases of the fundamental that are situated at equal distances on either side of the fundamental. The aliases for this choice are displayed in the bottom panel of Fig. 5. For the case illustrated in Fig. 5, this sampling frequency is twice the bandwidth.* This choice is the lowest of the three mentioned. It would require that the chopped signal be bandpass filtered about the chopping frequency.

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*Note that we define "bandwidth" to mean the full width of the fundamental component, that is, twice the highest frequency in the signal spectrum before chopping and sampling.

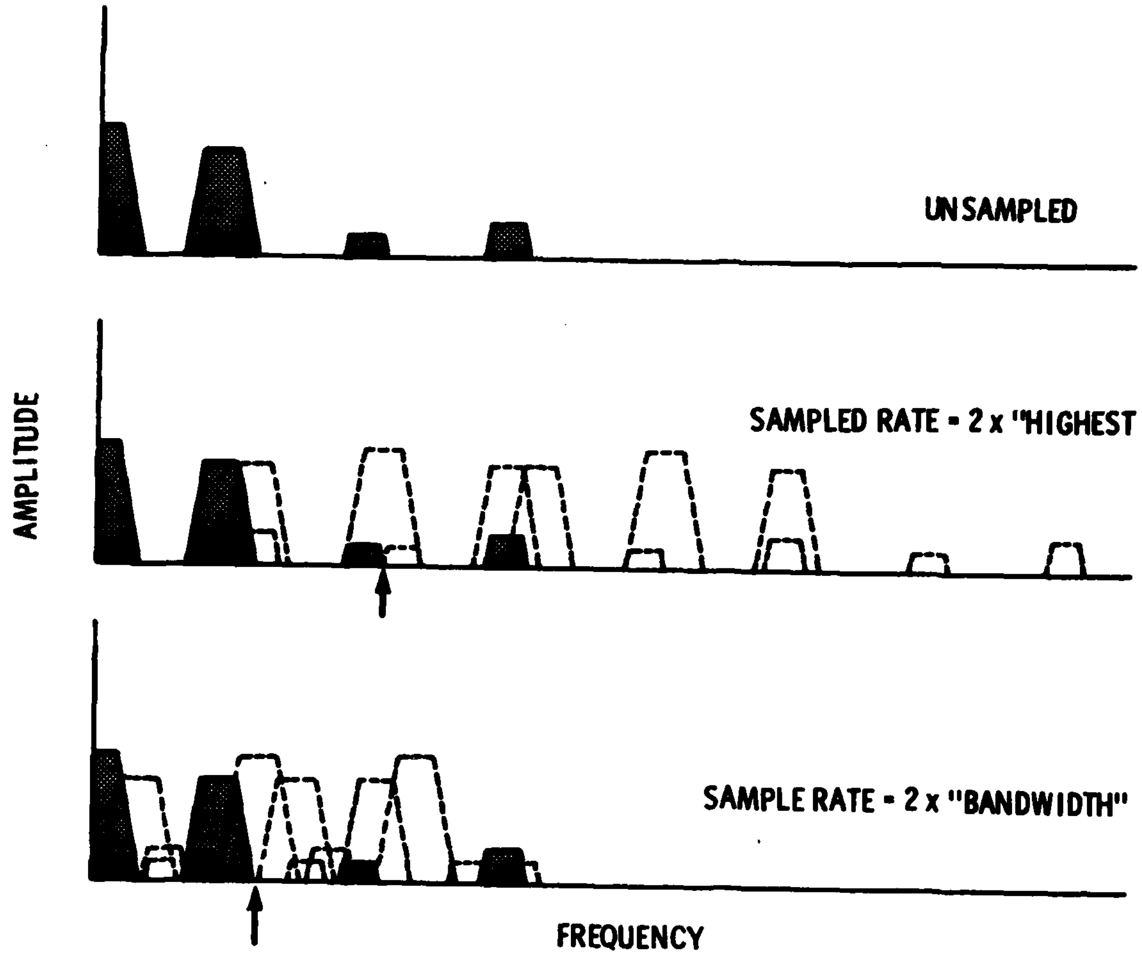

Figure 5. Spectrum of a Chopped Radiometer Signal

Note that different filter types are required for the two very different sampling rates in (b.) and (c.). These examples indicate the need to perform tradeoffs and to quantify the residual aliasing errors, i.e., the errors due to out-of-band filter leakage.

SECTION 3
SAMPLING RULES FOR NARROW-BAND SIGNALS

This section presents simple equations and a corresponding nomograph for choosing sampling rates for waveforms that have sparse band-limited spectra like the one shown in Fig. 6a. This spectrum is zero outside a frequency range of width (bandwidth) B centered at a frequency f_c that is greater than B/2.

Two types of sampling rules are given:

a.) One that results in aliases, which together with the information bands, comprise an array of equally spaced, non-overlapping bands, as in Fig. 6b.

b.) One that results in aliases that are in exact registration with other aliases and/or with the information bands (as in Fig. 4b). These alias groups are equally spaced.

The first type of rule is useful when a sparse spectrum similar to Fig. 6a results from some sort of filtering process; e.g. bandpass electrical filtering or the effect of an optical filter and longwave detector cutoff in a spectrum obtained by FTS. Usually, the filter or detector cutoffs are not sharp; hence, the aliases have weak "wings" that can overlap. The residual error (effect of the overlapping alias wings) is usually minimized when the aliases are equally spaced.

The second type of sampling rule is useful when the sparse spectrum is produced by chopping a bandlimited input signal and filtering out all spectral components except the fundamental centered at the chopping frequancy f_c. In this case the signal band and its aliases have the Hermitian symmetry properties noted in Section 2, and no problem is caused by having aliases in exact registration with the signal band. The minimum sampling rate achievable with the second rule is lower than that achievable with the first; however, other considerations may preclude sampling at the minimum rate.

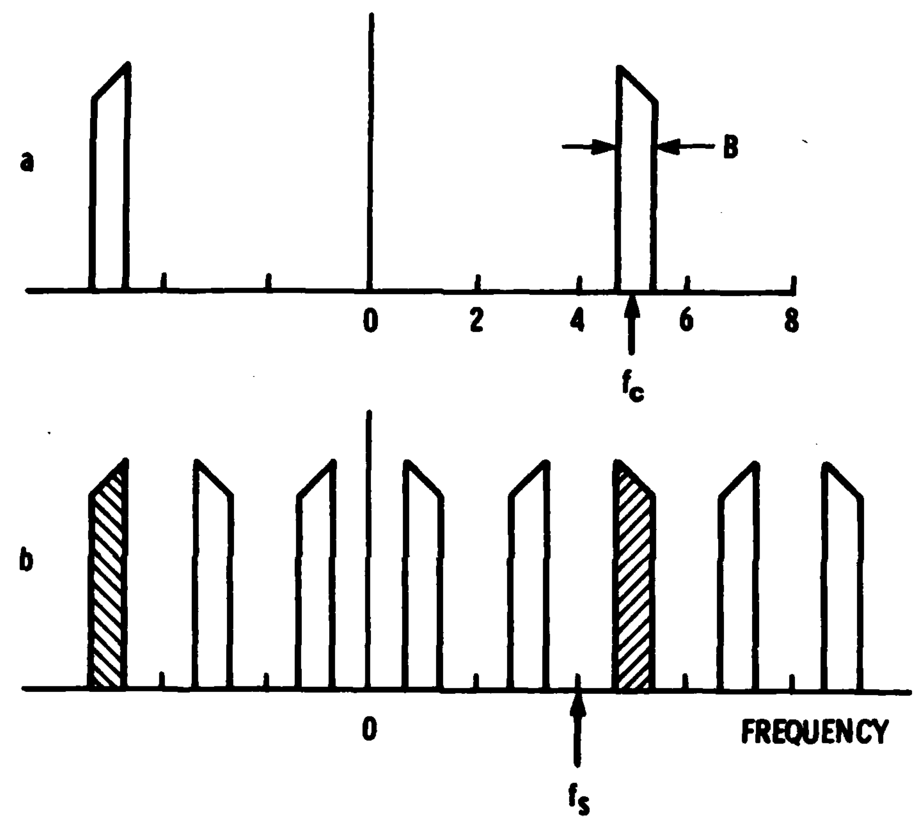

Figure 6. A Sparse Bandlimited Spectrum and Aliases Resulting from Sampling Rule 1

Note that these rules avoid or minimize effects of only the primary aliases. Any out-of-band components such as chopping harmonics, if not filtered out prior to sampling, will give rise to additional aliases that can cause problems.

3.1 First Type

A theorem of Fourier transform algebra states that the aliases introduced to the spectrum as a result of sampling can be obtained by shifting the original spectrum by all positive and negative multiples of the sampling frequency and summing the results. It is easy to show that the original spectrum in Fig. 6a plus its aliases, will comprise an array of equally-spaced bands (as in Fig. 6b) if the sampling frequency f_s and the original band center frequency f_c are related by

$$f_s = \frac{4}{2n+1} f_c , \tag{1a}$$

where n is any positive integer. Satisfaction of Eq. (1a) ensures that the successive bands have equal *distances between their center frequencies*. The additional requirement

$$f_s \geq 2B \tag{1b}$$

ensures that the bands will not overlap one another. The bands abut each other when $f_s = 2B$. The spacing of the band centers is $f_s/2$, and the center frequencies of the two bands closest to dc are $\pm f_s/4$.

Equations (1a) and (1b) comprise the rule for the first type of sampling. The allowable values of f_s as a function of f_c are plotted in Fig. 7 for this rule as well as for other sampling rules. Note that the first rule can be represented as a series of straight lines emanating from the origin but not beginning until $f_s \geq 2B$. The minimum sampling rate f_s for given f_c that results in equally spaced, non-overlapping aliases is obtained from the solid portions of the lines. However, sampling rates can also be selected anywhere on the upward extensions of the solid lines.

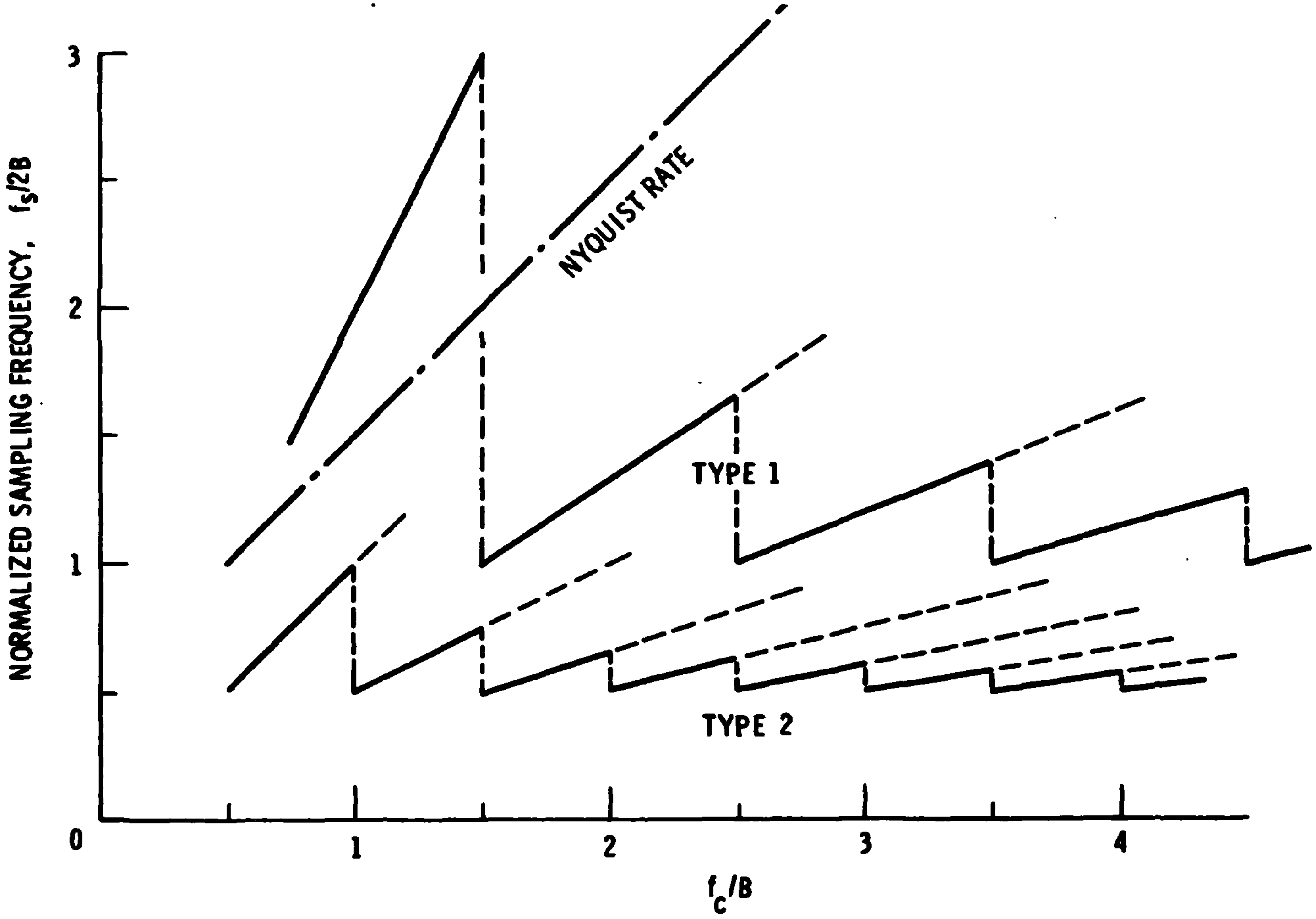

Figure 7. Sampling Rules 1 and 2 and the Nyquist Rate

Shown for comparison is f_s versus f_c as prescribed by the Nyquist theorem. The Nyquist sampling rate is twice the highest spectrum frequency:

$$f_s = 2(f_c + B/2) \; . \tag{2}$$

Note that aliasing is avoided for <u>any</u> rate greater than the Nyquist rate. Type 1 sampling, which avoids aliasing and also ensures equally-spaced aliases, is confined to rates given by the solid lines and their upward extensions.

3.2 Second Type

It is easy to show that when

$$f_s = 2f_c/n \; , \tag{3a}$$

where n is an integer, the signal band and its aliases comprise an array of groups, each group consisting of the signal band and/or aliases that are in exact registration (see Fig. 4). The group centers are uniformly spaced. The additional requirement

$$f_s \geq B \tag{3b}$$

ensures that the groups will not overlap one another. The conditions (3a) and (3b), which comprise the second sampling rule, are plotted in Fig. 7.

The second sampling rule says merely that the waveform can be sampled at intervals equal to n/2 times the chopping period $1/f_c$ without incurring detrimental aliasing effects. Clearly, it is best to sample at the chopping waveform maxima if the sampling interval is a multiple of the chopping period (i.e., if n is even), and at the chopping maxima and minima if n is odd.

Note that synchronous sampling (any even n except zero in Eq. 3a) effectively demodulates the chopped signal. If no noise is added by the detector/preamp, the samples are the same, within a multiplicative constant, whether or not the signal is chopped. By chopping and then filtering the chopped signal (detector/preamp output) before sampling, we accept a generally lower in-band noise level and eliminate an additional aliased noise contribution.

The spacing of the groups of registered aliases is equal to f_s. When n is even there is a group centered at zero frequency (the samples of the chopped signal are effectively demodulated). When n is odd the center frequencies of the two groups closest to dc are $\pm f_s/2$.

There are other useful sampling rules similar to the second type. For example, when

$$f_s = nf_c/2 \tag{4}$$

where n is an integer greater than 3, the spectrum after chopping and sampling (but no filtering) will contain groups of aliases that are in exact registration, <u>including</u> the aliases of the chopping harmonics. When n is large, only high-order harmonics will register with the fundamental. This may have an advantage, such as simplified filtering requirements, even though large n implies a high sampling rate.

3.3 Demodulation of the Samples

In some situations we require the spectrum or power spectrum of the input to the radiometer. Then it is unnecessary to demodulate the output samples if the signal was chopped. The spectrum can be obtained by computing the Fourier transform of the samples over the frequency range encompassing all or one-half of the lowest-frequency alias group. The frequency scale of the computed transform is then renumbered, i.e., a constant is added or subtracted to the scale. This general procedure, which recovers the spectrum of the input whether or not the input was chopped, is illustrated in Fig. 8. If the input was not chopped, we may compute the Fourier transform between frequency units 0 and 2; then four units are added to obtain the spectrum of the unsampled input. If the input was chopped, Fig. 8a represents the spectrum after filtering and before sampling. We may compute the transform of the samples between 1 and 2 frequency units,

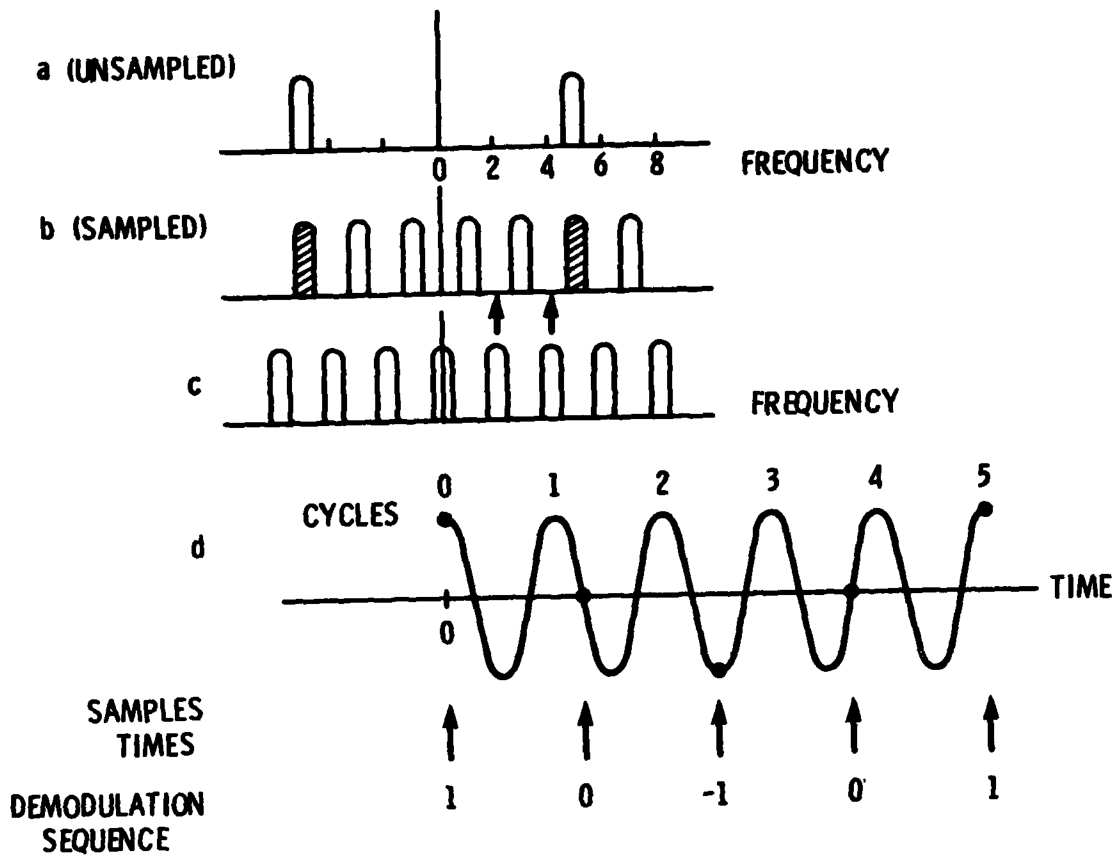

Figure 8. Example of Sampling a Chopped Signal and Demodulating the Samples

then subtract 1 to obtain the spectrum of the input. The center frequency of the lowest alias (or alias group) is determined by the sampling frequency and which sampling rule was used. When there is no special relationship between the sampling and chopping rates f_s and f_c, the lowest frequency alias in the spectrum is centered at the smallest positive value of

$$\begin{aligned} &nf_s \pm f_c \\ &n = 0, \pm 1, \pm 2, \ldots \end{aligned} \tag{5}$$

Figure 8 also helps to illustrate the process of demodulating the samples in the time domain. Suppose the chopping frequency f_c is five units as in Fig. 8a. Let the sampling frequency f_s equal four units, which obeys the type 1 sampling rule. Then the spectrum of the samples, given by Fig. 8b, has aliases spaced $f_c/2 = 2$ units apart. The spectrum in Fig. 8c can be obtained from Fig. 8b by shifting the later right by $f_s/4$, then left by the same amount, and averaging the shifted spectra. The equivalent operation in the time domain is to multiply the sample sequence by $\cos[2\pi(f_s/4)t] = \cos(\pi f_s t/2)$. The spectrum in Fig. 8c is what would be obtained by sampling the unchopped input at the rate $f_s/2$ (which is half the rate at which the chopped signal was actually sampled). Thus, in the time domain the samples can be demodulated by multiplying the sample sequence by

$$\cos(\pi f_s t/2) = \cos(\pi f_s k \Delta t/2) = \cos(\pi k/2) \tag{6}$$

where $\Delta t = 1/f_s$ is the actual sample interval and k is the sample index. The sequence of numbers given by Eq. (6) is

$$1, 0, -1, 0, 1, \ldots \tag{7}$$

corresponding to $k = 0, 1, 2, \ldots$. The zeros in this sequence correspond to zeros in the fundamental of the chopping waveform, as will be demonstrated. Thus, the demodulated sequence consists of alternate samples in the original set multiplied by the sequence $1, -1, 1, -1, \ldots$. These are spaced at twice the sampling interval Δt. The above arguments assume that every fourth sample coincides with a maximum (zero phase) in the fundamental of the chopping waveform.

Figure 8d shows the positions of the samples in the fundamental of the chopping waveform and the corresponding demodulation number sequence. Note that if the samples start out in phase with the chopping waveform, they always occur at the values 1, 0 or -1 for type 1 sampling. This happens because the ratio of sampling period to chopping period is an odd integer divided by four (see Eq. 1a), which is always an integer plus 1/4 or an integer plus 3/4. It is apparent from Fig. 8d that a different demodulation number sequence would be needed if the sampling had a different initial phase with respect to the chopping waveform.

Since alternate samples are zero in the present example, it is evident that we could have obtained the same information by sampling at one-half the rate used, i.e., at two rather than four frequency units. Then, since the chopping frequency is five units,

$$f_s = \frac{2}{5} f_c \; . \tag{8}$$

We are saying that the second sampling rule (Eq. 3a) with odd n could have been used instead of the first rule. With $f_s = 2$, the aliases in Figs. 8b and 8c are really pairs of aliases (of half the amplitude indicated) in exact registration. These pairs add to give exactly the same spectrum. Demodulation is accomplished simply by multiplying alternate samples by -1 whenever the second rule is used and n is odd.

In this example it is preferable to sample at the lower frequency prescribed by the second rule, since identical results are obtained for the demodulated samples. Note, however, that different results will be obtained by the two rules if the original spectrum (Fig. 8a) contains additional components, such as residual chopping harmonics.

If the input was not chopped, then, of course, demodulation is not required. Generally, the spectrum of the input would be of the type indicated by Fig. 6a rather than Fig. 8a. Sampling rule 2 would not be appropriate. If the input is sampled by rule 1, the input signal can be reconstructed in the time domain by convolving the sample sequence with the function

$$\cos(2\pi f_c t)\ \sin(\pi f_s t/2)/\pi t\ , \tag{9}$$

where f_c is the band center frequency and f_s is the sampling frequency.

SECTION 4
COMPUTER CODES

A set of computer codes was written to simulate instrument and sampling effects in a power density spectrum inferred from the output of a chopped radiometer. Some of these codes are subroutines that generate specific functions or parameters representing the scene power spectral density (PSD), the instrument's spatial responses (MTF's) and the chopping waveform. The user can revise these routines or replace them with others. The user can also specify an arbitrary chopping waveform by inputting its Fourier series coefficients.

The codes determine the apparent or inferred power density spectrum resulting from the following sequence of operations:

a.) modification of the scene spatial frequencies by the optics MTF.

b.) scanning the scene (in one direction) by a detector with finite (non-zero) instantaneous field of view.

c.) chopping of the radiometric input.

d.) noise-free detection of the chopped radiation.

e.) analog demodulation of the detector output (OPTIONAL).

f.) electrical filtering of the output.

g.) sampling.

h.) demodulation of the samples (if required) and computation of their PSD.

i.) normalization of the computed PSD to remove the effects of the system MTF's and electrical filter response.

The code produces plots that show the scene PSD and system MTF's and compare the apparent scene PSD to the actual scene PSD.

There are two main codes and a large set of subroutines. These are written in Fortran 77 for the Apollo computer. One of the main programs, NINPMI, accesses a user created file of input parameters or allows the user to operate interactively, supplying the input data in response to prompts. NINPMI computes the apparent PSD, and optionally will plot, all on the same set of axes,

a.) the original scene PSD.

b.) the square of the MTF of the detector (system IFOV).

c.) the square of the optics MTF.

d.) the product of a., b. and c., which is the PSD of the detector output in the absence of chopping, filtering and sampling.

The log-log plot is scaled automatically to accommodate the input scene PSD. The x-axis is labeled in frequency units supplied by the user. The code writes a mass storage file containing the apparent PSD and the original PSD.

The other main code, PSDPLT, produces a separate plot containing both the apparent PSD and original scene PSD. Inputs supplied by the user direct it to produce a full log-log plot, or a log-linear plot covering a reduced range of frequencies. The two programs produce their plots on the Apollo video display and also store the screen bit images as files that can be subsequently printed to obtain dot-matrix hard copies.

The required input parameters and available (default) subroutines that supply system MTF's and the scene PSD are defined in Section 4.1. General operating instructions and sample output plots are given in Section 4.2. We continue here with a general description of the different operating modes of the main program NINPMI.

There are four different operating modes. The most flexible mode, corresponding to input parameter TYPE = 4, allows the user to input up to 10 Fourier coefficients representing the chopping waveform or an analog-demodulated chopping waveform, and allows any choice for the chopping rate and sampling rate. The other three modes each correspond to particular chopping waveforms (or demodulated waveforms) and/or particular types of sampling.

The "synchronous demodulation mode" (TYPE = 1) assumes that the chopped signal undergoes imperfect analog synchronous demodulation (rectification) prior to sampling. The effective carrier after demodulation is sketched in Fig. 9. Note that a five percent imbalance in rectification is assumed. Except for the imbalance the wave is a rectified sinusoid. Three Fourier coefficients, representing dc through the second haromonic, are built into the code to represent this waveform. There is a relatively large second harmonic, due to the rectification, and a residual fundamental component due to the imbalance. This mode assumes synchronous sampling; i.e., the user should choose sampling and chopping frequencies that obey the second rule in Section 2. The code computes the apparent spectrum that begins at zero frequency.

The "minimum sampling mode" (TYPE = 2) assumes that the chopping waveform is a raised cosine $(1 + \cos 2\pi f_c t)$, except that the peaks and troughs are clipped within ±30 degrees of their maxima and minima. The clipping gives rise to a weak third harmonic. The code represents the wave by its dc through third harmonic Fourier coefficients. The user may choose any value for the sampling and chopping frequencies, although the use of rule 2 might be appropriate for this case. The code computes an apparent spectrum with origin at the alias center closest to zero frequency; i.e., it assumes that the samples have undergone digital demodulation.

The "stacked alias mode" (TYPE = 3) assumes square wave chopping of 100 percent efficiency; i.e., the wave consists of ones and zeros. The code uses the dc through fifth harmonic Fourier coefficients of the square wave. The user may specify any sampling and chopping frequencies, although the use of rule 1 with a sharp-cutoff bandpass filter might be appropriate. Again, the computation of the apparent spectrum assumes digital demodulation.

The modes represented by TYPE = 1 through TYPE = 3 represent common imperfections in the chopping waveform and/or analog demodulation process.

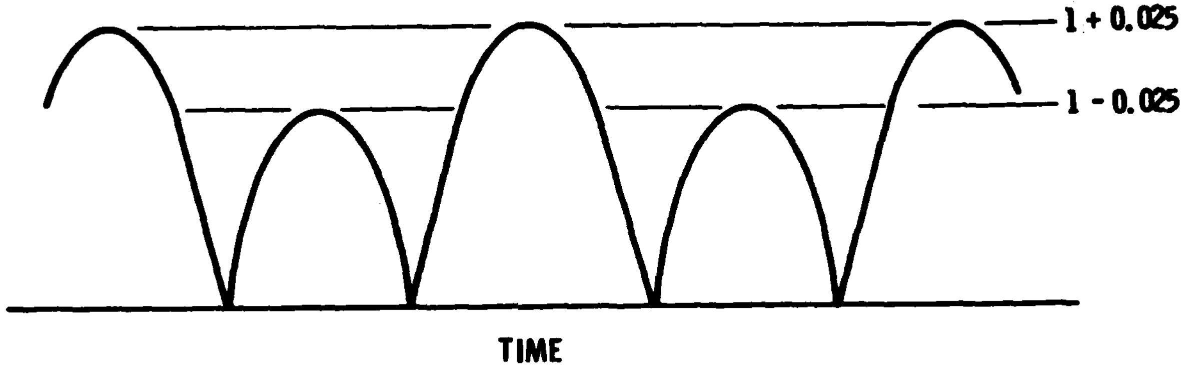

Figure 9. Synchronously Demodulated Chopping Waveform with Five Percent Imbalance in Rectification

They are merely default options that may or may not be appropriate for a particular chopper design or analog demodulation scheme.

As mentioned, the mode represented by TYPE = 4 allows the choice of any chopping waveform and sampling frequency, but in this mode the user must provide the coefficients representing the Fourier transform of the chopping waveform. When TYPE = 4, the user must also specify the additional parameter ITYP4 that indicates whether or not analog synchronous demodulation is used. If it is used (i.e., when ITYP4 = 1), the input coefficients are assumed to represent the rectified chopping wave, and the user would normally specify a sampling frequency that obeys rule 2; the code will compute the spectrum of the analog demodulated spectrum starting at zero frequency. When ITYPE = 2 it computes the spectrum beginning at the lowest frequency alias, which assumes correct digital demodulation of the sampled signal.

Fourier coefficients A(N) specified by the user should be computed from

$$\begin{aligned} A(0) &= \frac{2}{\pi}\int_0^{\pi} f(\theta)\, d\theta \\ A(N) &= \frac{1}{\pi}\int_0^{\pi} f(\theta)\cos(N\theta)\, d\theta \end{aligned} \tag{10}$$

where $f(\theta)$ is the chopping waveform defined over a half cycle ($\theta = 0$ to π). Note that the second formula gives values that are one-half those obtained using the usual definition of the Fourier coefficient. Equations (10) give the correct strengths of the impulses at dc and at positive and negative frequencies that comprise the Fourier cosine transform of a periodic waveform. The chopping waveform is assumed to have even symmetry about the point $\theta = 0$.

The code allows the choice of either a Butterworth or Bessel filter. The user specifies whether the filter is low pass, high pass or bandpass, its frequency cutoff(s), and the number of poles.

4.1 Input Parameters and Subroutines

4.1.1 Program NINPMI

The APOLLO command to execute the bound object module containing NINPMI and its subroutines must be appended with a file name. The code will store the original scene PSD and computed apparent PSD in this file. The user must also name an input file in the usual way, unless he will run the code interactively. Input is free format (list directed), but all character data must be entered in capital letters. When inputting in the interactive mode, the program supplies prompts and allows the user a chance to correct each line immediately after it is entered.

The following are the input parameters. Each line of parameters represents one input record. Frequency parameters can be either spatial or temporal but they must all be specified in the same units whether they refer to PSD's, MTF's, electrical filtering, sampling, etc.

J , JAPS , IBATCH

J($\leq$ 1000) is the number of frequencies at which the scene PSD and system MTF's are to be computed (by subroutines defined below). JAPS ($\leq$ 1000) is the maximum frequency index for the computed apparent PSD; it will cover the range 0 to JAPS. IBATCH is "1" if these and the remaining inputs are to be read from an input file.

UNIT , 'LABEL'

UNIT is the value of the first nonzero frequency represented in the scene PSD array, and LABEL is the name of the frequency unit; i.e., if the scene PSD as perceived by the scanning radiometer is computed at 0, 5 Hz, 10 Hz, ..., then UNIT would be "5" and LABEL could be Hz or HERTZ. The LABEL can have up to six characters.

FP , FD , FO

FP is the frequency of the half power point of the scene PSD. The ratio FP/UNIT is required to be a multiple of four when the sampling obeys rule 1 or obeys rule 2 with odd n. FD is the frequency of the first zero in detector MTF (which the default subroutine computes as a sin af/f function). FO is the frequency of the optics MTF at the 1/e point.

Y or N

Y or N indicates "yes" or "no". Specifying Y results in the plot of scene PSD and system MTF's that was described earlier.

Y or N

This Y or N is a response to the question "Shall plot cover full function arrays?". If the answer is N the next input record should give the lowest and highest frequency <u>indices</u> to be covered by the x-axis of the plot. The higher index may not exceed J-1 where J is the value submitted in the first input record.

Y or N

A Y response allows the user to change certain default values in the plot; this response evokes additional prompts. The action of the code in response to Y has not been thoroughly tested. The response must be N if input is from a file and the file created by this code will be used by the code PSDPLT.

Y or N

This Y or N determines whether or not the computation will continue; i.e., the response N would be entered if the code was run only for the purpose of obtaining a plot of the scene PSD and squares of the system MTF's.

If the input data is entered as a file (IBATCH = 1) and the first of the Y/N records contains N, then none of the additional Y or N records should be included in the file. In other words, if the first type of plot is not required, questions about the plot are not relevant, and the only purpose for executing NINPMI can be to obtain the apparent PSD, which requires additional input data.

FS , FC

FS is the sampling rate and FC is the chopping rate.

TYPE , N

TYPE specifies the mode of computation defined in the first part of Section 4, and N ($\leq$ 10) is the number of the highest order Fourier coefficient that will be used to represent the chopping waveform for the specified mode. We recall that the values of N used in the default subroutines are 2, 3 and 5 for the modes (TYPE) equal to 1, 2 and 3, respectively. The user may input an N that exceeds these values, in which case the undefined coefficients will be set to zero by the default subroutines. The user will want to specify higher values of N if he has modified the default subroutines to supply the higher-order Fourier coefficients.

ITYP4

This input is required only if TYPE is 4. Its meaning is defined in the *first part of Section 4.*

A(0)
A(1)
⋮
A(N)

These are the N+1 Fourier coefficients required if TYPE = 4. They are entered one per line.

ITYPFL

ITYPFL = 1 selects a Bessel filter. A value of "2" selects a Butterworth filter.

FON , FOF , NPOLE

These are the lower and upper filter cutoff frequencies and the number of poles, respectively. If the filter is lowpass, enter -1 for FON. NPOLE must be 10 or less.

4.1.2 Program PSDPLT

The command PSDPLT to execute the program must be appended with the name of the file created by NINPMI. This code will plot both the original scene PSD and computed apparent PSD. The inputs are

Y or N

Respond to the question "Do you want to change plot limits?". If the answer is N, a normal, full-size log-log plot is produced. If N is entered, no additional inputs except the last one (the plot label) are required.

IB1 , IB2

IB1 (> 0) and IB2 (< J-1) are the array limits that the plot will cover; i.e., they determine the range of frequencies of the x-axis.

Y or N

Submit Y to shrink the x-axis. (This option was not been tested.)

SHR

If above entry was Y, submit the percentage of the original size that should be used for the x-axis.

LOGPL

Submit "1" for LOGPL to obtain a log-log plot, or "0" to obtain a plot with log y-axis and linear x-axis.

´LABEL´

Submit up to 60 characters to be used as a label for the plot.

4.1.3 Subroutines

Subroutine POWER (PSD, N, FHJ) creates the array PSD representing the power spectrum

$$\mathrm{PSD}(J) = \frac{1}{1 + (J/\mathrm{FHJ})^2}\ ; \quad J = 0, 1, \ldots, N-1 \tag{11}$$

whose half power point is at the index FHJ. Recall that the input parameter FP defines the actual frequency represented by FHJ.

Subroutine MTFOPT (MTFO, N, FHJ, FPREL, FOREL) generates an optics MTF of the form

$$\mathrm{MTFO}(J) = \exp(-(J/\mathrm{AH})^2); \quad J = 0, 1, \ldots, N-1 \tag{12}$$

where AH ≡ FPREL·FOREL·FHJ is the index of the 1/e point. The user supplies the frequency of the 1/e point and other data to the main code, which determines FHJ, FPREL and FOREL. The Gaussian MTF given by Eq. (12) corresponds to a Gaussian blur circle. The two Gaussians are a Hankel transform pair.

Subroutine MTFDET (MTF, N, FHJ, FPREL) computes a detector MTF according to

$$\begin{aligned} \mathrm{MTF}(J) &= \sin(X)/X; \\ X &= \pi J/\mathrm{FHD}, \quad J = 0, 1, \ldots, N-1 \end{aligned} \tag{13}$$

where FHD ≡ FPREL·FHJ is the index of the first zero in the MTF*.

Some guidelines for choosing the input parameters required for the scene PSD and optics MTF are given in Section 4.2, entitled "Operating Instructions".

*It may seem that the two subroutines MTFOPT and MTFDET have an unnecessarily large number of parameters. This is true, and is a carryover from an earlier version of the code which used different conventions for specifying of the two system MTF's.

Subroutine FTSINI(A,N) returns the Fourier transform coefficients of a raised sine wave with flats within ±30 degrees of the maxima and minima. It supplies only the first four coefficients, A(0) to A(3), and sets A(4) to A(N) to zero. It also sets A(-J) = A(J).

Subroutine FTYSND(A,N) returns the Fourier transform coefficients of a synchronously demodulated sine wave with five percent imbalance in rectification. It returns A(0) through A(2), sets A(3) through A(N) to zero, and sets A(-J) = A(J).

The remaining subroutines would not normally be changed by the user and therefore are given very brief descriptions:

Subroutine FTSQW returns the Fourier transform coefficients of a fully modulated raised square wave.

Subroutine BESSEL returns the complex transfer function of a lowpass or highpass Bessel filter having a given number of poles.

Subroutine BWORTH returns the complex transfer function of a lowpass or highpass Butterworth N-pole filter.

Subroutine CONV performs the convolution of two symmetrical functions.

Subroutine PSDAPP computes the apparent power spectral density (PSD), given a complex filter response and convolved functions supplied by subroutine CONV.

Subroutine LMINA establishes minimum and maximum values of a function.

Subroutine NORMCO normalizes the apparent power spectrum to correct for the optics and detector MTF's, and the filter response, and writes the result and the original scene PSD to a file.

Subroutine SETCO initializes screen parameters for common block SCDAT.

Subroutine DLFUNP plots a function on logarithmic x and y axes.

Subroutine FRAME frames a figure with axes.

Subroutine LAXPL draws a logarithmic, labelled axis.

Subroutine LINAXP draws a linear, labelled axis.

4.2 Operating Instructions and Sample Outputs

The name of the bound object module containing the main code NINPMI and its subroutines is NINPMI. Thus, to execute the code, one inputs the command

NINPMI FILE1

The appended file name identifies the file into which the code will write output data. One must identify an input file in the usual manner unless the user will enter data in the interactive mode. The second code PSDPLT is executed by giving Apollo the command

PSDPLT FILE1

The appended file name must, of course, be the same for both runs.

The input instructions have been given in Sections 4.1.1 and 4.1.2. An abbreviated listing of these instructions for code NINPMI is contained in the file NINPUT. A similar file for code PSDPLT is named INPPSP.

In the event the user wishes to change or replace any of the default subroutines, he must create or "build" a new object module containing the new binaries. The two command files BNINPMI and BPSDPLT may be used to bind the two existing main codes with all of their required subroutines and a required library of primitive plot routines (/USER.LIB/GPLOTLIB).

All of the codes, subroutines, instruction listings and command files are in the CGS Apollo directory //CGS58/DISK_BETA/CONTRACTORS/USERS/ARC/HUBER/PROJ1.

Code NINPMI will occasionally print the error message

"ICAFT = XXXX EXCEEDS LIMIT OF ICAFAL = 5000"

which aborts execution. This message concerns the upper limit, ICAFAL, of the index ICAFT used in the conveolution of the input power spectrum with the array of impulses representing the Fourier spectrum of the chopping waveform. To avoid excessive computing time and unnecessarily large arrays, this limit has been set at 5000; i.e , it is required that

$$(J-1) + JN \cdot JC \leq 5000 \ ,$$

where J is the number of points in the scene PSD, JN is the number of the highest-order chopping waveform coefficient and JC is the integer value of FC/UNIT, where FC is the chopping frequency and UNIT is the frequency spacing of points in the scene PSD. The message indicates that the user must select a larger value of UNIT and/or fewer Fourier coefficients and/ or smaller J.

4.2.1 Selection of Input Parameters

This subsection concerns the selection of input frequencies defining the scene PSD and optics MTF, when the default subroutines are used to compute these functions.

The default PSD function defined by Eq. (11) in Section 4.1.3 has one free parameter which is defined by the user input FP, the frequency of the half power point. This PSD is the Fourier transform of the spatial auto-correlation function $\exp|-2\pi x \cdot (FP)|$; i.e., the correlation distance in the scene is $1/2\pi \cdot (FP)$. For the altitudes to be probed by limb sounding in the CIRRIS 1A flight, there is some evidence[1] that correlation lengths are of the order of 10 km and that the default PSD has the correct high frequency roll-off; i.e., slope = -2 on a log-log plot. Thus the frequency FP in spatial units is

$$FP \simeq 1/(2\pi \cdot 10) \text{ cy/km} \ .$$

To convert this to the units cy/mr, we multiply by the number of kilometers subtended by one milliradian at the tangent height. For one of the CIRRIS 1A radiometer detectors this number of 1.62 km/mr. Finally, to convert to temporal frequency, we multiply by the spatial scan rate, which in the CIRRIS 1A fast scan mode will be 20 mr/sec. Thus, we obtain

$$FP \simeq \frac{1}{2\pi \times 10} \frac{cy}{km} \times 1.62 \frac{km}{mr} \times 20 \frac{mr}{sec} \simeq 0.516 \text{ Hz}$$

In the computer run described below, we chose FP = 0.4 Hz, which is a reasonable approximation considering that the correlation distance and PSD rolloff are not known to high accuracy.

The default optics MTF function defined by Eq. (12) in Section 4.1.3 corresponds to a Gaussian blur circle, and has one free parameter which is defined by the user input FO, the frequency of the 1/e point in the MTF. One can show that the MTF function

$$\exp[-(F/0.48302)^2] ,$$

where F has units cy/mr, corresponds to a blur circle in which 90 percent of the encircled energy is contained within a radius of one mr. Assuming this optical quality for the CIRRIS 1A radiometer and a scan rate of 20 mr/sec, the input temporal frequency would be

$$FO = 0.48302 \frac{cy}{mr} \times 20 \frac{mr}{sec} = 9.66 \text{ Hz}$$

4.2.2 Sample Outputs

The above input values were used to model the final design for the CIRRIS 1A radiometer. An additional input value was FD = 40 Hz corresponding to a detector size of 0.5 mr and scan rate of 20 mr/sec. Figure 10 shows the plot produced by code NINPMI, which displays the source PSD and the squares of the two system MTF's, and their product (the dotted curve).

The chopping waveform was represented over 0 to π (a half-period) by $\sin\theta$ clipped at amplitude 0.25. The wave is zero over the second half-period. This waveform was represented by its Fourier transform coefficients,

A(N) = 0.4598, 0.3149, 0.03930, -0.09625,
-0.03685, 0.04806, 0.03040, -0.02525
-0.02823

corresponding to N = 0 to 8. The selected chopping rate was 84.6 Hz and the sampling rate was 550/sec. The selected filter was a lowpass Butterworth with two poles, and cutoff at 169 Hz. Figure 11 shows the plot

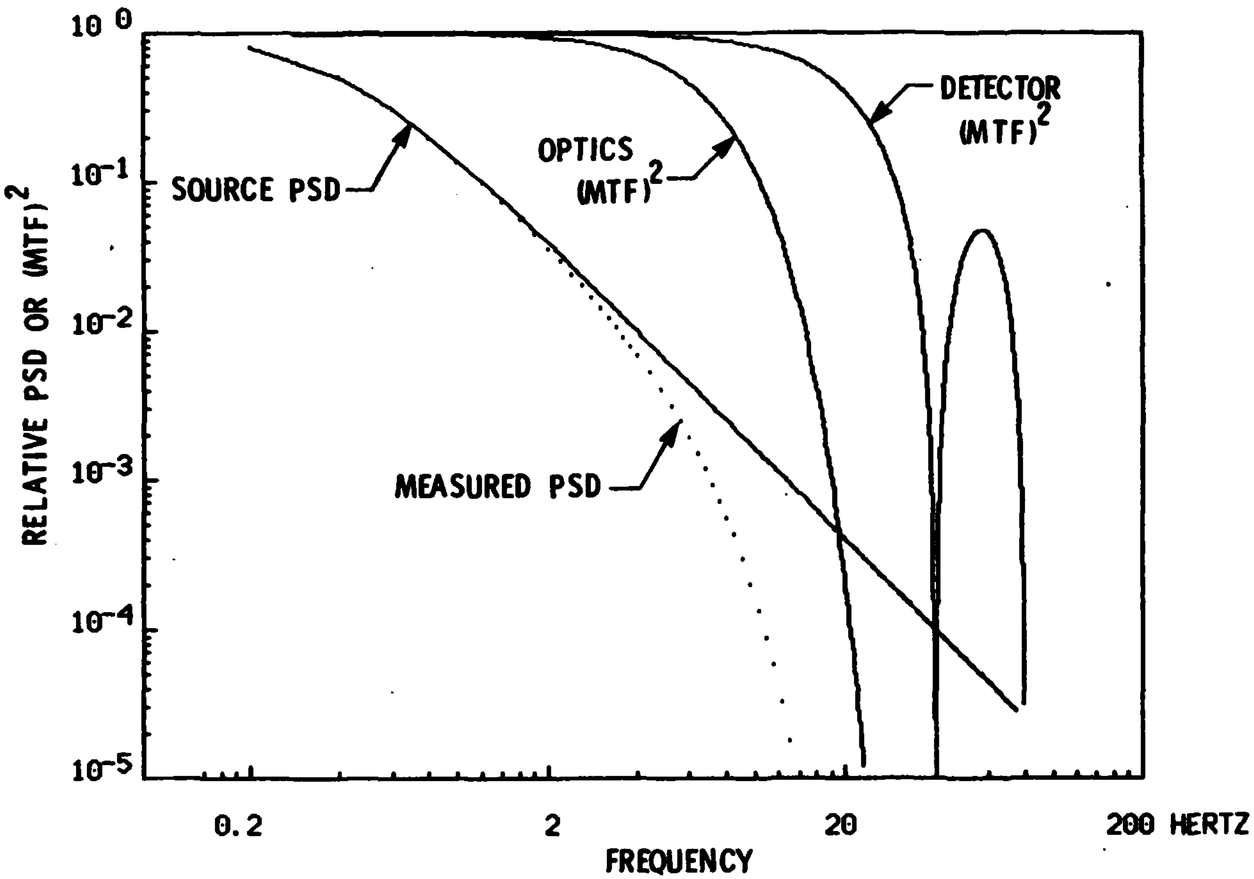

Figure 10. CIRRIS 1A Radiometer Spatial Frequency Response

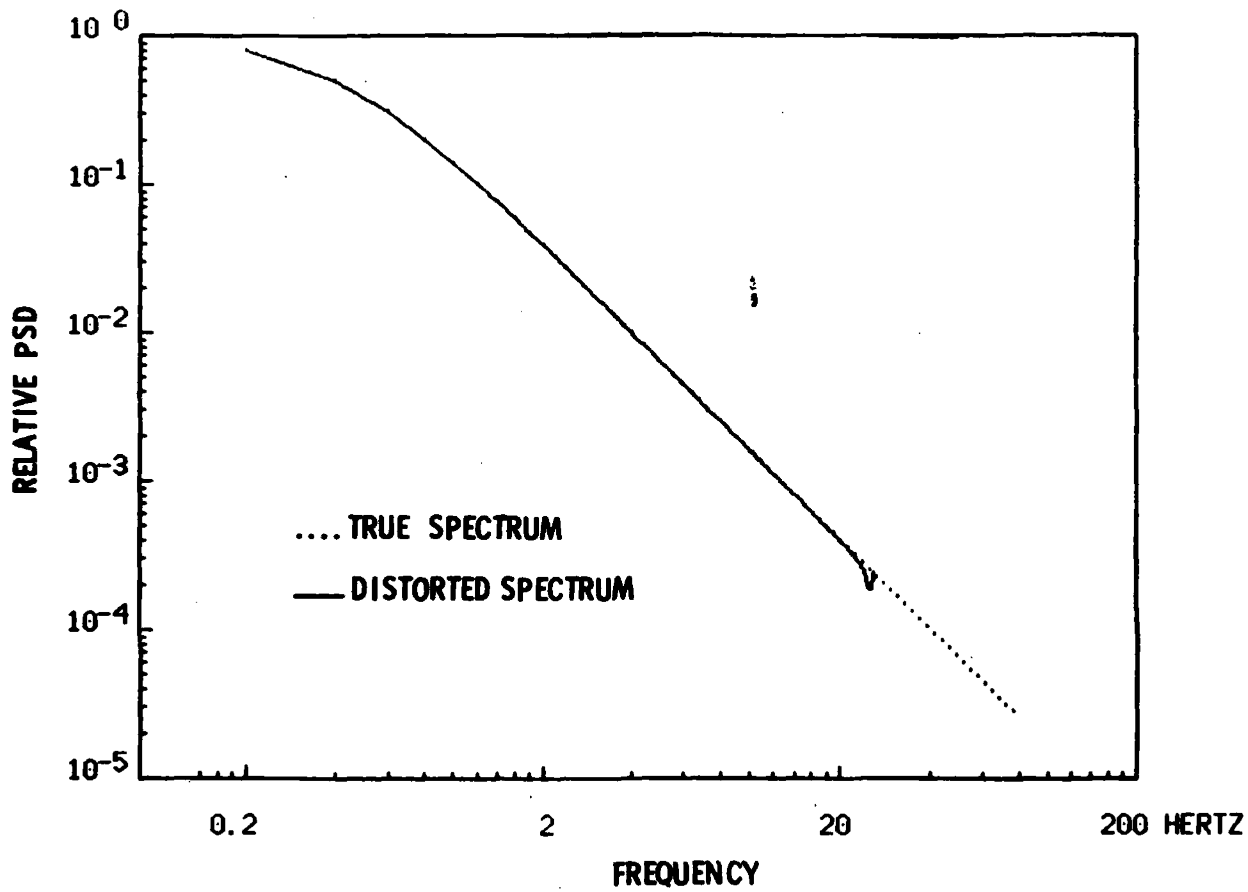

Figure 11. CIRRIS 1A Radiometer Alias Distortion of Spatial PSD

produced by code PSDPLT, which compares the true scene PSD to the computed apparent PSD. Note that the effects of the optics MTF's have been normalized out of the apparent PSD.

Figure 12, which is unrelated to the computer runs for CIRRIS 1A, demonstrates the ability of code PSDPLT to produce blow-up plots on a log-linear scale. The nearly vertical discontinuity in the curve representing the distorted PSD corresponds to a zero in the detector MTF, where the MTF normalization cannot be applied.

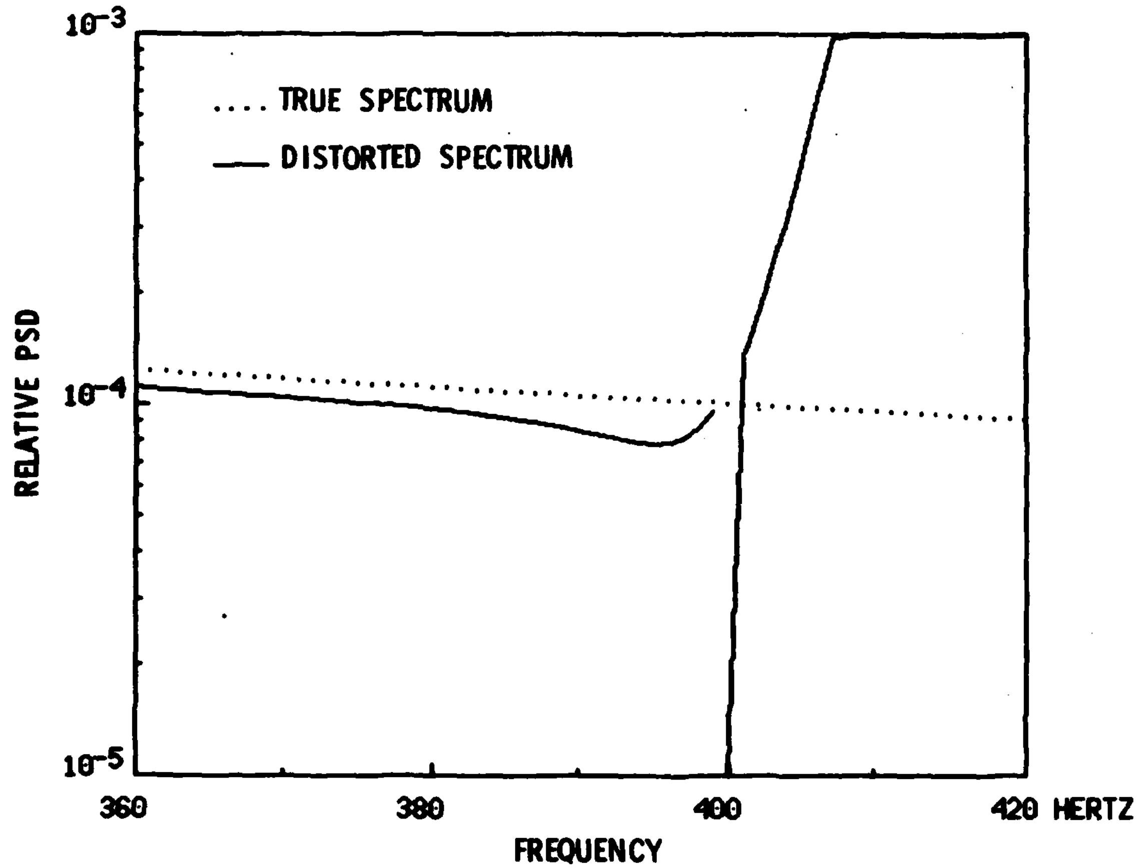

Figure 12. Example of Log-Linear "Blow-up" Plot by Code PSDPLT

REFERENCE

1. C. H. Humphrey, R. M. Nadile, C. R. Philbrick and D. R. Smith (Editors), *Atmospheric Infrared Radiance Variability*, 27 May 1981, AFGL-TR-81-0207. AD A109928.