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# The Telecommunications and Data Acquisition Progress Report 42-72

October - December 1982

E C. Posner  
Editor

February 15, 1983



National Aeronautics and  
Space Administration

Jet Propulsion Laboratory  
California Institute of Technology,  
Pasadena, California



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## Preface

This quarterly publication provides archival reports on developments in programs managed by JPL's Office of Telecommunications and Data Acquisition (TDA). In space communications, radio navigation, radio science, and ground-based radio astronomy, it reports on activities of the Deep Space Network (DSN) and its associated Ground Communications Facility (GCF) in planning, in supporting research and technology, in implementation, and in operations. Also included is TDA-funded activity at JPL on data and information systems and reimbursable DSN work performed for other space agencies through NASA. The preceding work is all performed for NASA's Office of Space Tracking and Data Systems (OSTDS).

In geodynamics, the publication reports on the application of radio interferometry at microwave frequencies for geodynamic measurements. In the search for extraterrestrial intelligence (SETI), it reports on implementation and operations for searching the microwave spectrum. The latter two programs are performed for NASA's Office of Space Science and Applications (OSSA).

Finally, tasks funded under the JPL Director's Discretionary Fund and the Caltech President's Fund which involve the TDA Office are included.

This and each succeeding issue of the TDA Progress Report will present material in some, but not necessarily all, of the following categories:

### OSTDS Tasks

- DSN Advanced Systems
  - Tracking and Ground-Based Navigation
  - Communications, Spacecraft-Ground
  - Station Control and System Technology
  - Network Data Processing and Productivity
- DSN Systems Implementation
  - Capabilities for New Projects
  - Networks Consolidation Project
  - New Initiatives
  - Network Sustaining
- DSN Operations
  - Network Operations and Operations Support
  - Mission Interface and Support
  - TDA Program Management and Analysis
- GCF Operations and Implementation
- Data and Information Systems

### OSSA Tasks

- Search for Extraterrestrial Intelligence
- Geodynamics
  - Geodetic Instrument Development
  - Geodynamic Science

### Discretionary Funded Tasks



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# Description and Overview of an Instrument Designed to Measure Line-of-Sight Delay Due to Water Vapor

G M Resch

Tracking Systems and Applications Section

M C Chavez and N I Yamane

Microwave Observational Systems Section

*Eight dual-channel microwave radiometers have been constructed as a research and development effort for the Crustal Dynamics Project and the Deep Space Network. These instruments, known as water vapor radiometers, are primarily intended to demonstrate that the variable path delay imposed by atmospheric water vapor can be calibrated in microwave tracking and distance measuring systems but could also be used in other applications involving moist air meteorology and propagation studies. They are being deployed to various stations and observatories that participate in Very Long Baseline Interferometry (VLBI) experiments including the Deep Space Stations in Spain and Australia. In this paper we review the development history of these instruments, outline the theory of operation and overall design considerations, and sketch the instrumental parameters and performance characteristics.*

## I. Introduction

It takes longer for a radio wave to traverse an atmospheric path  $L$  relative to the time it would take to traverse the same path in a vacuum. The electrical path length  $L_e$  is just the integral of the refractive index of the atmosphere along the path. The difference  $\Delta L$  between the electrical path length and the physical path length is simply,

$$\Delta L = 10^{-6} \int N ds \quad (1)$$

where  $N$  is the refractivity at the point  $s$  along the path. The refractivity for the atmosphere as given by Smith and Wein-

traub (Ref 1) is composed of two terms. The first and largest term is called the dry term and is proportional to the integrated dry air density. For most purposes it is sufficient to "weigh" the atmosphere at the zenith with a barometer in order to estimate the "dry" delay  $\Delta L_d$  and then scale this zenith quantity to the line-of-sight using a cosecant elevation law. The second term contributing to the refractivity is a function of the atmospheric water vapor. The excess path delay in centimeters for this term of the refractivity is of the form,

$$\Delta L_v = 0.1723 \int (\rho_v/T) ds \quad (2)$$

where  $\rho_v$  is the vapor density in  $\text{gm/m}^3$ ,  $T$  is the temperature in kelvin, and  $s$  is in meters. Unfortunately (for our applica-

tions) atmospheric water vapor is a highly variable and not well-mixed atmospheric constituent. As a result it is impossible to estimate the path delay in Eq (2) with high accuracy from surface measurements alone.

In interferometric systems our primary observable is the differential time-of-arrival of a radio wave at the stations comprising the interferometer. In a ranging system the primary observable is the round-trip signal time. For both systems atmospheric water vapor along the signal path will impose an additional delay between 3 to 60 cm (one way) depending on how much vapor is in the atmosphere and the elevation angle of the observations. Typically, the baseline or range determination is derived from observations at several different elevation angles so that the mapping of propagation effect error into baseline or position is rather complex. In general, if one requires a system accuracy better than 10-12 cm then the vapor along the signal path must be measured. Of the several techniques that could be used to measure line-of-sight vapor delay, the most cost effective utilizes the techniques of passive remote sensing with a dual-channel microwave radiometer. The water vapor molecule emits spectral radiation at a frequency of 22.235 GHz, whose intensity is approximately proportional to the number of molecules (or the delay) along the line of sight. The presence of liquid water complicates the measurement problem as it contributes significantly to the intensity of the atmospheric emission but contributes very little to the excess path delay. However, with a second microwave channel it is possible to separate the vapor and liquid effects in all but the most severe weather situations.

## II. History

At its inception, it was immediately recognized that the technique of Very Long Baseline Interferometry (VLBI) could be used to investigate a wide variety of geodetic and geophysical phenomena (e.g., see Ref 2). The basic instrumentation appeared capable of centimeter-level measurement precision. Shapiro and Knight (Ref 3) pointed out that the overall accuracy of VLBI would be limited by systematic effects. In particular, the variable delay imposed by tropospheric water vapor seemed to be the limiting error source.

In 1972 the Jet Propulsion Laboratory began a research task called ARIES (Astronomical Radio Interferometric Earth Surveying) under the leadership of P. F. MacDoran. The objective of this task was to demonstrate that VLBI could be used to study the deformation of the earth's crust. The idea was to use a transportable 9-m antenna as one end of an interferometer, with base stations at Goldstone and the Owens Valley Radio Observatory. If the transportable station was moved around a network of geophysically interesting sites in southern California it would be possible to compile a time history of

three-dimensional baseline vectors whose changes in time would tell us something about the crustal deformation around major faults in this tectonically complex region. The accuracy goals for this task were beyond the capability of conventional surveying techniques. Our aim was to measure three-dimensional vector baselines up to 1000 km long with 3- to 5-cm accuracy in each component and 1- to 2-cm accuracy in length. Ong et al (Ref 4) demonstrated that the ARIES instrumentation was capable of 3-cm accuracy by measuring a 300-m baseline between the 9-m antenna and the 64-m antenna at Goldstone. Progress toward the reduction or elimination of other major error sources was underway. By 1974 it seemed that three-dimensional baseline accuracy of a few cm would be possible in just a few years — if only the water vapor problem could be solved.

During the summer of 1974 J. W. Waters and R. Longbotham compared a spacecraft prototype microwave radiometer with radiosondes launched at the El Monte Airport in an experiment to determine atmospheric delay using the techniques of passive remote sensing. Their regression analysis against radiosonde data indicated a level of agreement better than 2 cm (units of excess path delay). Later work by Winn et al (Ref 5) and Moran and Penfield (Ref 6) extended this basic result. It was clear that this device, which became known as a water vapor radiometer (WVR) could be used to calibrate vapor-induced delay along the line of sight of a nearby antenna. If the water vapor delay along the line of sight at each station could be calibrated to 2 cm, then 3- to 5-cm VLBI system accuracy seemed possible. During this same period a VLBI data acquisition system was being developed at the Haystack Observatory that was well suited for the small mobile VLBI antennas that we intended to use in ARIES. It was clear that ARIES needed a WVR that would operate as an integral part of this much larger, fully automated VLBI data acquisition system. The WVR would have to operate unattended and require a minimum of operator attention. It must be well calibrated, stable, able to operate in all but the most severe weather conditions, and provide a vapor path correction that was accurate to better than 2 cm. There was also a need to understand the random and systematic error budget of the WVR and to develop an inversion algorithm to convert the WVR observable (brightness temperature) to path delay correction.

The technical aspects of demonstrating the performance of the WVR and specification of its system operation turned out to be the easy part of the problem. Finding the funding to actually build the kind of WVR that ARIES needed was a much more complicated, time consuming, and discouraging task. In late 1976, with the help of E. J. Johnston, ARIES was able to borrow the engineering model of a spacecraft radiometer and began a low key effort to modify and repackage the

device for VLBI support. In 1977, the VLBI group at Goddard Space Flight Center (GSFC) purchased two microwave receivers from Sense Systems Inc (corresponding to the two channels of a WVR) with the intention of building their own WVR. Of course everyone realized that WVR calibration was needed on both ends of the interferometer. Both the ARIES and GSFC groups used the Owens Valley Radio Observatory (OVRO) as a base station and they simultaneously perceived a rare opportunity to act in their mutual self-interest and equip the observatory with a WVR. Support was obtained from the Caltech President's Fund to purchase one of the microwave receivers necessary to make a WVR, and GSFC purchased the second channel as well as two positioners. The ARIES group agreed to put all of these assorted pieces together to make three reasonably compatible WVRs. Our intention was to use these WVRs as tools to demonstrate the validity of the technique as it applies to VLBI support and then build units that were an optimum design for our application. In early 1978, with the support of the Goddard group, we received partial funding to package three WVRs. Several months later we received directions to construct two additional WVRs for the Crustal Dynamics Project (which was then coming into existence).

The situation in July 1978 was very chaotic. We had some of the parts for three WVRs, directions to put together five WVRs, no clear funding pattern, and were restricted from carrying over money from one fiscal year to the next. We decided that it would be less expensive and faster to construct the two additional WVRs "in house" and immediately began the purchase of long-lead-time parts. In the latter part of 1978 the Deep Space Network (Advanced Systems and Implementation Offices) added two more radiometers to what had now become a production line. A year later Project ARIES requested another WVR (its second unit). These units were given several numbers, R01, R02, ..., R08 according to the order in which we began work on the hardware. The "R" designation served to emphasize the R&D nature of the effort.

We were directed by JPL management to construct these WVRs in a manner that was consistent with the needs of both JPL and non-JPL users. This presented a problem in that we had already inherited the pieces of three radiometers and were constrained to use off-the-shelf components for the new units. Our approach to this problem was dictated by our stated goal - to build research tools that were capable of demonstrating that vapor delay could be calibrated in VLBI measurements. We did *not* attempt research in radiometer construction, but instead decided to adopt the "safe" design of a Dicke switched instrument taken largely from a spacecraft design. It was decided to standardize on the two frequencies 20.7 and 31.4 GHz, use doubly stabilized Dicke radiometers, and discard the cold load calibration in the Sense Systems

receivers. The latter point was dictated by three factors: (1) the load was inaccurate, (2) it was operationally cumbersome to keep filled with liquid nitrogen, and (3) it presented severe packaging problems.

For three reasons we decided to mount the WVRs on their own positioner. First, we assumed that if the WVR would be used for any application other than VLBI support it should not tie up the VLBI antenna. Second, the procedure that we use to establish and check the temperature scale calibration (i.e., the tipping curve) is rather awkward and time consuming when performed on a large antenna. Third, preliminary analysis suggested that the only difference between the broad beam WVR and the smaller beam radio telescope would be in the short-period fluctuations of water vapor. Since most VLBI observations are the result of incoherently averaging up to 15 minutes of data, it effectively smoothes the fast but small vapor fluctuations. In order to simplify the interfacing we added a microcomputer controller to which the user can send simple commands and have the WVR point, take data, etc.

There were several minor areas of design, assembly, and/or construction in the first three instruments that we felt could be improved. The changes that were finally incorporated into units R04-R08 were dictated by the requirement that all WVRs appear identical to the user at the controller interface. Unfortunately it proved to be impractical to remove all of the differences between the different units. The most obvious remaining difference is that the vapor frequency of WVR-R02 and R03 is centered at 21 GHz (as compared to 20.7 GHz for all other units) and the slew rate is approximately a factor of 4 slower than the other radiometers. Thus the eight radiometers that have been deployed consist of three different microwave packages, two different positioners, and three slightly different microcomputer controllers. However, thanks to the software that resides in the microcomputer these differences are largely transparent to the user. The current plan is to mount units R01 and R07 on the edge of the antenna; they are intended to calibrate and to use the fast mounts from these units to retrofit units R02 and R03.

In July 1979 the unit designated WVR-R03 was delivered to the Haystack Observatory. According to our original agreement with GSFC (which now managed the Crustal Dynamics Project), the WVRs would be merely bench tested at JPL. Each observatory would be responsible for engineering tests and calibration of the WVR it received. This arrangement was found to be impractical, so in late 1979 at the request of the Crustal Dynamics Project we agreed to test and calibrate each WVR prior to shipment. In addition, we would assist with installation of the WVR at those observatories that needed help, keep a stock of spare parts, assist with maintenance problems, provide a minimal level of documentation (i.e.,

consistent with R&D instruments), and provide the algorithm necessary to convert WVR output into a delay correction. In order to provide a backup capability for vapor calibration as well as a check on WVR operation, we also agreed to construct, test and calibrate a solar hygrometer to go along with each WVR. These extensions of our responsibilities turned out to be more difficult and time consuming than was originally anticipated. WVR-R02 was installed at OVRO in September 1980, unit R05 was installed at the Haystack Observatory in January 1981, unit R06 was installed on the ARIES electronics van (Mobile Van-2) also in January, unit R04 was installed at DSS 13 in February, unit R01 was delivered to ARIES (Mobile Van-1) in May, and unit R08 was installed at Ft Davis in June 1981. As per our agreement with the Crustal Dynamics Project, we have purchased a modest stock of spare parts, some of which have been used to solve problems at several installations. We have installed the WVRs where it was necessary.

This paper is the first of a series that will constitute the "minimal" level of documentation. In later articles we will discuss the control module, the microwave package, the positioner, testing and calibration, the algorithms, and the solar hygrometers.

### III. Remote Sensing of Water Vapor

The physics of atmospheric water vapor and the techniques used to sense it are well developed in Refs 7 and 8. We shall merely summarize a few pertinent points in this paper. In Fig 1 we see the apparent brightness temperature of the zenith atmosphere in the frequency interval 10 to 40 GHz. The lower curve illustrates the behavior of the brightness temperature in the complete absence of atmospheric water vapor. The middle curve in Fig 1 is drawn for the case of 2 gm-cm<sup>-2</sup> of precipitable water vapor  $M_v$ , and the upper curve represents the spectrum for an additional 0.1 gm-cm<sup>-2</sup> of liquid  $M_L$  that is assumed to exist as small droplets. This is a columnar measure, that is, if we were to take a column 1 cm<sup>2</sup> at the base that extended through the entire atmosphere and precipitated all of the vapor, we would wind up with 2 grams of water. For our application a "handy dandy" relationship to remember is

$$\Delta L_v \approx 6.3 M_v \quad (3)$$

or, the excess path delay  $\Delta L_v$  (in cm) is about 6 times the precipitable vapor content  $M_v$  (expressed in gm/cm<sup>2</sup>). Thus the middle curve of Fig 1 refers to a zenith vapor delay of 12.6 cm, representing roughly average conditions over much of the United States. The reader should note that in the literature, precipitable vapor is often measured in centimeters (instead of gm/cm<sup>2</sup>). This results in some confusion when

instruments that measure precipitable vapor are compared with instruments that measure path delay.

The bump in the brightness temperature curve of Fig 1 at the frequency of 22.235 GHz is due to an emission line from the water vapor molecule. If there is not a lot of water vapor along the line of sight (i.e., the line is not saturated), then the brightness temperature is roughly proportional to the columnar density plus the background radiation. The background radiation is due to three sources: (1) the black-body radiation of the cosmos, which is known to be 2.9 K, (2) emission from the lower wings of higher frequency lines, in particular the band of oxygen lines near 60 GHz, and (3) emission and scattering of radiation from liquid water droplets that occur in clouds or rain. The emission from oxygen is calculable and can be removed from the observations. The presence of liquid water has little effect on the path delay but considerable effect on the brightness temperature, thereby confusing the vapor measurement. With measurements at two frequencies we can separate the vapor and liquid contributions.

If the average diameter of the water droplets is small compared to observing wavelength, then the spectrum of the radiation from the liquid component will vary as frequency squared (Ref 9). At some frequency  $f_1$  (near the vapor emission line) the brightness temperature  $T_{b1}$  will be proportional to  $k_1 \Delta L_v + k_2 M_L$ , where  $k_1$  and  $k_2$  are constants, and  $M_L$  is the integrated liquid content. At some frequency  $f_2$  (off the vapor emission line) the brightness temperature  $T_{b2}$  will be proportional to  $k_1' \Delta L_v + k_2' M_L$ . If we observe these two brightness temperatures  $T_{b1}$  and  $T_{b2}$ , then we have two equations in two unknowns. Given the frequency-squared dependence of the liquid emission, these equations can be solved simultaneously for  $\Delta L_v$  and  $M_L$ . We find that both the path delay and the precipitable liquid can be expressed as a linear combination of the observed brightness temperatures,

$$\Delta L_v \approx a_0 + a_1 T_{b1} + a_2 T_{b2} \quad (4)$$

where  $a_0$ ,  $a_1$ , and  $a_2$  are "constants." This equation summarizes the basic measurement approach and is useful in considerations of performance of the measurement system. In actual use where observations are made at low elevation angles and in cloudy conditions, Eq (4) is not a good approximation. A slightly more complicated expression analogous to the approach used in Ref 10 will be derived in a later paper.

### IV. The Instrumentation

A water vapor radiometer is simply a device for measuring sky brightness temperature at two frequencies on and near the

emission line at 22.2 GHz. It consists of two independent Dicke radiometers (see Ref. 11 for an explanation of the operation of a Dicke radiometer), one tuned to 20.7 GHz and the other tuned to 31.4 GHz. The RF electronics, switching circuitry, and power supplies comprise the microwave package. The microwave package is mounted atop a positioner that can point in azimuth and elevation. Both the microwave electronics and positioner are interfaced to the controller package that consists of a microcomputer and local control panel from which one can monitor any of the analog signals from either the microwave electronics or the positioner. The ability to monitor and control all WVR functions from the local control panel is very useful for diagnostic purposes. Under normal operating conditions, the microcomputer controls and acquires data from the microwave package and the positioner. The microcomputer can be interfaced to another computer (i.e., a host computer) or to a terminal via an RS-232 serial interface. A set of simple instructions is available to command the microcomputer to point the WVR, switch to internal loads, take data, etc. Figure 2 shows all three WVR modules, Figure 3 is a block diagram of the complete WVR.

## V. System Requirements

In VLBI the basic observable is a group delay (Ref. 12), i.e., the differential time of arrival of a radio wave at the stations comprising the interferometer. Differing amounts of atmospheric water vapor surrounding the stations cause the observed group delay to differ from the calculated group delay (e.g., calculated from the geometry). Of course there are many other sources of error that lead to differences between observed and calculated delay. Our ability to either calibrate or "solve-for" these error sources sets the limit on the accuracy of the VLBI technique to measure vector distance. As discussed in Ref. 13, the analysis of VLBI error sources suggests that baseline accuracies of 3 to 5 cm are possible if the line-of-sight water vapor effects can be calibrated to  $\pm 2$  cm. At the 2-cm level there are several comparable error sources, so that it simply does not make sense to invest a large amount of resources in the reduction of any single error source. Keep in mind that the 2-cm vapor delay accuracy refers to a differential measurement, so that the accuracy of an individual measurement should be  $2/\sqrt{2}$  cm. We would like the precision of the vapor measurement to be a factor of 3 or 4 better in the hope that the increased precision could be used to spot systematic effects. Hence the overall system requirement for the WVR is that it measure vapor delay with accuracy of  $\pm 1.5$  cm and precision of  $\pm 0.4$  cm. Our problem in 1978 was to convert this requirement on system performance into a general set of design specifications.

As a response to a directive from the JPL Program Office for the Crustal Dynamics Project we formed a design team to

review the design specifications for a water vapor radiometer. The conclusions and recommendations of this team were issued as a JPL internal report, which was written in the very early stages of construction and became the engineering specifications document. Table I summarizes the specifications of the water vapor radiometers.

### A. Physical Characteristics and Environment

Since we started with a stock of existing parts the physical characteristics of the WVR were somewhat predetermined. The physical characteristics of units R01 and R04 to R08 are listed in Table 1. Units R02 and R03 have slightly larger volume but more or less the same physical characteristics. The environmental factors were specified from consideration of the meteorological conditions for both the mobile and fixed antenna sites that were under consideration by the Crustal Dynamics Project. At temperatures below  $-10^{\circ}\text{C}$  it was felt that the atmospheric vapor content would be very small and hence there would be little need for the vapor correction. The WVRs will survive temperatures as low as  $-40^{\circ}\text{C}$  but below  $-10^{\circ}\text{C}$  it becomes increasingly more difficult for the internal heaters to thermally stabilize the electronics. Similarly, at wind speeds above 65 km/hr the VLBI antenna would probably shut down. At wind speeds higher than 95 km/hr it is recommended that the WVR be stowed in the horizontal position in order to present the minimum surface to wind loading. A particularly irksome problem that was not solved was the problem of dew accumulation on the cover of the horn antennas.

### B. Operating Frequencies

One of the critical specifications in the WVR design is the choice of operating frequencies. This is because the water vapor line centered at 22.235 GHz is pressure broadened. Thus the line profile will depend on the altitude distribution of the vapor. A given amount of vapor that was concentrated at low altitude would produce a wide, flat profile relative to the same amount of vapor that was concentrated at high altitude. Most of the vapor signal is produced at 22.235 GHz but this is also the frequency that is most sensitive to the shape of the line. However, if we pick a frequency on the wing of the line, e.g., between 20 and 21 GHz, we can minimize this sensitivity to vapor height (Ref. 14 and 15).

The components of the NEMS (unit R01) radiometer operated at 22.2 and 31.4 GHz, and the two Sense Systems radiometers operated at 21.0 and 31.4 GHz. Preliminary bench tests indicated that the NEMS components could possibly be used as low as 20.7 GHz. There is a decided advantage of having all WVRs operate at identical frequency pairs, so we decided to standardize on 20.7 GHz as the vapor sensitive channel for all new WVRs and attempt to retune the existing components.

### C. Frequency Stability

The specification of frequency stability is important because the vapor channel operates on the part of emission line where the rate of change of brightness temperature with respect to frequency is largest. The frequency will change because the Gunn diode local oscillator is temperature-sensitive and the WVR enclosure does not act as a perfect oven. The sensitivity is approximately 3 MHz/°C and since temperature changes of ±4°C inside the WVR are possible, this implies frequency shifts of ±0.12 GHz. At a precipitable vapor content of 2 gm/cm<sup>2</sup> the rate of change of brightness temperature with respect to frequency is 3.3 K/GHz, so that 0.04 K errors are possible in the presence of moderate amounts of vapor. These errors are below the quantization level of the A/D converter for most observing conditions.

### D. Bandwidth and Integration Time

The bandwidth specification, like the operating frequency of 20.7 GHz, was set by equipment in-hand. Both the NEMS unit and the two Sense Systems units are double sideband receivers with an IF passband from 10 to 110 MHz. The Dicke switch rate is 1000 Hz on all of the radiometers. The minimum integration time is set by a lowpass filter on the output of the synchronous detector whose characteristics were based on three considerations. First, the time constant should be long enough to span many cycles of the Dicke switch and thereby average out transients and any short term dithering in the switch reference frequency. Second, the time constant should be large enough to reduce the random noise fluctuations to a value that is several times the quantization level of the analog-to-digital converter. Third, the time constant should not be so large as to impose tedious wait states between changes of the input. We chose a lowpass filter with an equivalent time constant of 100 millisecond and thereby average 100 cycles of the Dicke switch. The rms fluctuation level is approximately 0.4 K compared to a quantization level of 0.1 K. The wait state after a mode change is determined by the time it takes the previous level to decay to less than the quantization level, which for our design is 9 time constants or 0.9 sec. This delay is enforced in the controller software – the WVR simply will not respond for 0.9 sec after a mode change. The effective integration time can be lengthened by software averaging techniques.

### E. Antenna

Ideally, one would like the antenna beamwidth of the WVR to be comparable to the beamwidth of the radio telescope to be calibrated. However, as indicated previously, this restricts the use of the WVR in other applications. In addition, a comparable WVR beam would entail considerable cost either as hardware or in development, in that it would mean a WVR aperture whose diameter is comparable to the radio telescope with all of the attendant problems of pointing, spillover, etc.

These problems are solvable given enough time and money, but this is unnecessary in our application.

We chose to use a horn antenna for the WVR that has a relatively broad beamwidth (7 degrees, full width at half maximum) for three reasons: (1) it has virtually no sidelobes, (2) the dimensions could be easily scaled to give the same beamwidth at both operating frequencies, and (3) the temporal/spatial resolution of the broad beam was more than sufficient in our applications. The beam pattern for this horn is shown in Fig. 4 taken from Ref. 16. In order to have a WVR beam that was comparable to that of the 64-m antenna operating at a 3.8-cm we would need a 16-m aperture. We could either use a separate 16-m paraboloid for the WVR or we might consider mounting the WVR in such a way as to utilize some or all of the 64-m aperture. In either case we would have to devote considerable engineering effort toward minimizing the problems of sidelobes and spillover that produce serious systematic error in the delay determination. Furthermore, since we use a variety of radio telescopes in our application we would require a custom WVR aperture or feed for each installation. Clearly this is impractical – if we need eight WVRs perhaps a single design is most cost effective. However, if one is to use a composite design, then the question that arises is how do we relate the output of the WVR to the output of the radio telescope when they have unequal beams?

Formally, there is an important distinction between the antenna temperature (which is what the WVR measures) and the sky brightness temperature (which is what we wish to measure). The antenna temperature  $T_a$  (neglecting losses) is given by a convolution of the brightness temperature distribution  $T_b(\theta, \phi)$  with the antenna beam pattern  $P(\theta, \phi)$  as,

$$T_a \sim \int T_b(\theta, \phi) P(\theta - \theta_0, \phi - \phi_0) d\Omega \quad (5)$$

where  $\theta, \phi$  are spherical coordinates and  $(\theta_0, \phi_0)$  is the pointing direction. If the beam pattern has unwanted responses in directions other than the pointing direction then we must correct for them. A typical sky brightness temperature of 30 K would be considerably corrupted by a 100 K cloud or the 300 K earth in a sidelobe even if the sidelobe responses were 20 dB down from the main beam. The virtue of having an antenna beam with no sidelobes (i.e., very high main beam efficiency) is that we can equate the antenna temperature to the brightness temperature (except for a small pointing correction) and save a great deal of data “massaging”.

When we ask how the vapor affects a large radio telescope we must keep in mind that the beam of the radio telescope



takes a relatively large distance to form. The extent of the near field of an aperture is

$$L_{nf} \sim D^2/\lambda \quad (6)$$

where  $D$  is the diameter of the aperture and  $\lambda$  is the observing wavelength. This quantity is tabulated in Table 2 for some of the antennas used in our applications. Since the scale height of water vapor is typically 2 km, this means that most of the vapor is in the near field of a large radio telescope. Thus fluctuations in phase for the radio telescope are determined by the vapor inhomogeneities that pass through a cylindrical column of length  $L_{nf}$  and diameter  $D$ .

Let us assume that the water vapor exists in the form of “blobs” that are randomly distributed in size. If we sample equal volumes of atmosphere containing  $N_1, N_2, \dots$  etc., then we would expect to find  $N$  blobs on the average with an rms variation of  $\sqrt{N}$ . We would obviously expect less variation if we sampled with a large volume as compared to sampling with a small volume. If we normalize  $N$  to be the number of blobs per unit area, then the average value will be independent of sampling volume but the rms variations about the average will be a function of the sampling volume. Thus we can roughly estimate the difference in fluctuation between the WVR and radio telescope by simply computing the average vapor contained in the two beams. If we will assume that on the average the water vapor is exponentially distributed with scale height  $H_0$ , and the WVR beam is approximated by a cone of full-width  $\beta$ , then it is straightforward to calculate the ratio  $R$  of the volume of vapor in the WVR beam to the volume of vapor in the near field of the radio telescope,

$$R = V_{wvr}/V_{cyl} = 8[(H_0/D) \tan(\beta/2)]^2 \quad (7)$$

This ratio is shown in Table 2 using the 7-deg beamwidth of the WVR. If we neglect instrumental noise then we would expect the ratio of the fluctuations for the WVR to the fluctuations from the radio telescope to vary as the square root of  $R$ . The WVR, since it samples a larger volume, will vary less, but on the average it will determine the correct delay for the radio telescope.

An alternate way to view this problem is to note that Eq (5) is a function of time. The brightness temperature describes the two-dimensional projection of an assemblage of water vapor inhomogeneities that are constantly rearranged by winds aloft. Thus the quantity  $T_a$  is really a time series. If we had a pencil beam, i.e.,  $P(\theta, \phi)$  were a delta function, then the time variations of  $T_a$  and  $T_b$  would be identical. The fact that we have a beam of finite width that is convolved with  $T_b$  implies that the antenna acts as a lowpass filter by spatially averaging the details of the sky brightness distribution. There-

fore, we expect the output of the WVR to be a smoothed representation of the output from the radio telescope. On very short time scales the radio telescope may exhibit fluctuations that the WVR will not detect.

The calculation of the time scale that relates the WVR to radio telescope fluctuations is quite complex, involving the geometry and the three-dimensional spectrum of vapor inhomogeneities. We shall resort to a more intuitive approach. Consider that at the zenith the width of the WVR beam is approximately

$$D_w = 2H \tan(\beta/2) \quad (8)$$

where  $H$  is the height. For a height of 2 km (the nominal scale height of water vapor) and a beam width of 7 deg we calculate  $D_w = 244$  m or roughly 10 times the near field width from a 25-m radio telescope. For a given blob diameter  $D$  and radio telescope diameter  $D_R$ , we can distinguish three different regimes. Regime 1  $D < D_R$ , the ratio of the time that the blob spends in the WVR beam to the time it spends in the radio telescope beam is  $T \sim D_w/D_R$ . Regime 2  $D_R < D < D_w$ , the time scale is  $T \sim D_w/D$ . Regime 3  $D > D_w$ , the ratio of the times becomes  $T \sim 1$ . Thus in this example the very small blobs of vapor will spend about 10 times longer in the WVR beam as they do in the radio telescope beam, but as the size of the blob grows the ratio of time spent in the two beams approaches unity. If we further assume that the horizontal and vertical dimensions of the blob are roughly equivalent then it follows that the small blobs will cause small phase perturbations and large blobs will cause large perturbations. The WVR will not “see” the short-period fluctuations in the radio telescope output that are due to water vapor but it can detect the larger long-term effects. In reality, nature is not nearly so simple and all that can be safely said is that there is a rough equivalence between the spatial averaging of the WVR and temporal averaging from the radio telescope.

## F. Calibration

Two types of calibration are required for the WVR: (1) a frequent calibration of the relative gain of the instrument, and (2) a less frequent calibration of the absolute gain. The reference load of the Dicke switch is kept at the same temperature as the rest of the electronics. This means that the Dicke receiver is not fully stabilized (e.g., see Ref. 11) and we are susceptible to gain changes. Determination of the relative gain is accomplished by setting internal switches in the WVR so that the input of the Dicke receiver is connected to either of two waveguide terminations. One termination is kept at an elevated temperature (100°C) and is called the hot load while the other is kept at the same temperature as the rest of the electronics (43°C) and is called the base load. The physical

temperature of these loads is measured with thermistors and denoted by  $TH$  and  $TB$  respectively. The output of the receiver is digitized and presented to the user as a number  $N$  that we shall term "counts". In order to determine an unknown antenna temperature  $TA$  we can (in principle) go through the observing sequences – hot load, base load, antenna – and observe the outputs  $TH$ ,  $TB$ ,  $NH$ ,  $NB$ , and  $NA$ . For an ideal radiometer the antenna temperature is simply

$$TA = TB + [(TH - TB)/(NH - NB)] (NA - NB) \quad (9)$$

where the quantity in square brackets is called the relative gain or simply the gain of the instrument and is expressed in units of kelvin/count. For all of the WVRs, the value of the gain is approximately 0.1 K/count and is determined by the product of the gains from all circuit elements extending from the Dicke switch to the analog-to-digital converter. Figure 5 illustrates the typical fluctuations in the gain from one of the WVR channels. We plot the relative gain (i.e. by subtracting out the average) vs time for a 40-hour period. A diurnal variation of  $\pm 4\%$  is quite obvious and is due primarily to temperature changes inside the microwave package. Superimposed on the diurnal variation are short-term fluctuations due primarily to radiometer noise which can be reduced by averaging or smoothing.

Equation (9) describes an ideal radiometer. A real instrument suffers various shortcomings such as attenuation, less than perfect switch isolation, reflections, etc. The calibration of absolute gain involves finding the correction factor that puts Eq (9) closer to the absolute temperature scale. One of the procedures that can be used to find the instrumental correction factor is called a "tip curve", it is used to derive a correction to the hot load. Let us assume that we can rewrite Eq (9) for a real radiometer as

$$TA = TB + [TH - TB + \Delta TH] N' \quad (10)$$

where  $N' = (NA - NB)/(NH - NB)$  is the normalized count in the antenna mode, and  $\Delta TH$  is the hot load correction. The brightness temperature of the sky can be written

$$T_b = T_c e^{-\tau} + T_m (1 - e^{-\tau}) \quad (11)$$

where  $T_c = 2.9$  K is the cosmic blackbody background,  $T_m$  is the mean radiating temperature of the atmosphere, and  $\tau$  is the opacity. The mean radiating temperature exhibits some seasonal and site variations but with a reasonable degree of approximation we can assume that it is constant and equal to 275 K. For a stratified atmosphere the opacity can be expressed as,

$$\tau = \tau_0 (AM) \quad (12)$$

where  $\tau_0$  is the zenith opacity and  $AM = 1/\sin(\text{elevation})$  is the air mass. Suppose we make a series of observations, first observing the internal loads to get  $TB$ ,  $TH$ ,  $NB$ , and  $NH$ , and then moving to different elevation angles to get  $NA_1$ ,  $NA_2$ , etc., at positions  $AM_1$ ,  $AM_2$ , etc., and assume that  $TA = T_b$ . Equations (10), (11), and (12) can be combined to give,

$$N'_i = [T_M + T_B + (T_c - T_m) \exp(-\tau_0 AM_i)] / (TH - TB + \Delta TH) \quad (13)$$

Thus we have expressed the observable from the WVR as the dependent variable in terms of the independent variable  $AM_i$  and the two parameters  $\tau_0$ , and  $\Delta TH$ . Given two or more observations at different  $AM_i$  it is straightforward to solve for  $\tau_0$  and  $\Delta TH$  that are "best" in the least squares sense. Figure 6 shows tip curve data taken with WVR-RO7 (31.4 GHz) located at Goldstone in May 1981. All WVRs are calibrated in this manner prior to shipment. In addition, the temperature scale of each radiometer channel is checked by observing hot and cold aperture loads, which serves as a consistency check.

The proof of WVR performance lies in a comparison of the WVR with some independent (and one would hope more accurate!) technique of measuring vapor path delay. The most convenient comparison technique is to use a radiosonde to provide a vertical profile of temperature and relative humidity that can be integrated to give an estimate of the zenith path delay which we then compare with the delay estimate from the WVR that is observed at the zenith. Figure 7 shows two such comparisons, the first done in May 1974 and the second from May 1975 that was reported in Ref 5. Figure 8 shows previously unpublished data taken at Pt Mugu, California in 1976, and Fig 9 shows the histogram of the residuals for this experiment. During the Pt Mugu experiment we had an instrumented aircraft fly various slant paths up to a maximum altitude of 3 km, which provided a larger range of path delay estimates than are normally obtained from radiosondes. This data is shown in Fig 10.

The rms residuals from these comparisons are 1.6 cm from 1974, 1.1 cm from 1975, 1.5 cm from 1976 WVR/radiosonde, and 1.4 cm from 1976 WVR/aircraft (excluding data at 10 deg elevation). If the two techniques are truly independent then we expect the rms residual to be the root-sum-square of the error from each technique. If we take the average rms residual to be 1.5 cm, and assume that the radiosonde has a 10% accuracy, the average path delay to be 1.0 cm, then we infer the accuracy of the WVR to be 1.1 cm.

The WVR used in the experiments just described operates at 22.2 and 31.4 GHz. Unfortunately none of the new WVRs

have undergone such comparison. If anything, the older radiometers should be noisier due to the greater sensitivity to the vertical distribution of vapor. Guiraud et al. in Ref. 17 report a WVR/radiosonde comparison for a WVR that operates at 20.6 and 31.6 GHz and infer a WVR accuracy of 0.5 cm. We hope to demonstrate comparable performance with the new WVRs.

The instrumental correction factor should be constant for long periods of time (e.g., years). There are circumstances that could cause it to change slowly or dramatically. For instance, it is possible that the waveguide walls could deteriorate slowly in time due to atmospheric constituents. If the enclosure is leaking and there is a sudden change from warm, damp conditions to cold conditions, then liquid water or ice could form in the waveguide or switches or on the cover of the horn antenna. In either case the performance of the instrument will change and the predetermined value of  $\Delta TH$  will not be valid. It is recommended that tip curves be performed periodically and the results kept as a history of the instrumental performance.

### G. Temperature Control

The quantity that correlates most strongly with gain variations is temperature, so that it is important that the WVR have enough thermal inertia to resist sudden changes in temperature. Conversely, the gain should be monitored on time scales shorter than the time scale for real temperature changes. The important specification is the restriction that the physical temperature of any load must not change by more than the quantization level during the time it takes to measure the gain. Thus, the specification  $\pm 0.1^\circ\text{C}/\text{minute}$  assumes that it will never take longer than one minute to measure the gain.

Almost all of the microwave components are thermally connected to a large aluminum mounting plate. A mercury thermostat controls strip heaters attached to this plate and regulates its temperature to  $43^\circ\text{C}$ . A small fan circulates air around the few components that are not attached to the plate. The entire microwave package is enclosed by a sheet metal box that is lined with styrofoam and provides a thermal attenuation of approximately 6-7 dB. The box is attached to the positioner with a base plate made from a thermally insulating material. Thermal baffles are mounted on standoffs on the four sides of the box that expose the largest area. These baffles, which are covered with a paint that has high reflectivity in the infrared, were found to be necessary for summer operation in desert-type environments.

Figures 11 and 12 show the warmup characteristics of the two waveguide terminations (i.e., the hot and base loads). Typically, the hot load requires 1.5 to 2 hours to thermally stabilize and will remain constant to  $\pm 0.2^\circ\text{C}$  thereafter. The

base load, which is representative of the remainder of the electronics package, takes up to 3 hours to reach quasi-equilibrium. Figure 13 shows the typical diurnal variation of the enclosure temperature, which correlates very well with the gain variations.

### H. Positioning Module

The specifications for the positioner follow from the requirement that the WVR point along the same line of sight as the radio telescope that is being calibrated. This is most easily done with an Az/E1 mount and conversions from right ascension/declination done in software if necessary. Due to the 7-deg bandwidth of the horn antennas, we have not included a capability for tracking – only pointing. Also note that the pointing capability is  $\pm 1.0$  deg but the readout precision is  $\pm 0.1$  deg. Analysis of the pointing sensitivity indicates that positioning errors of  $\pm 0.1$  deg will give worst-case errors of 0.5 K in brightness temperature and 3 mm in the path delay in high opacity/low elevation angle conditions. Using these instruments in a typical VLBI experiment one would point to within  $\pm 1$  deg of the line of sight of the radio telescope (keeping 10 deg away from obstructions). Then one would request the data necessary to calculate the path delay, and note the actual position of the WVR so that the path delay could be mapped to the correct line of sight by assuming a cosecant elevation angle scaling.

The slew rate of the positioner even for the two “slow” units is generally much faster than the associated radio telescope. This allows the possibility of more complex observing strategy with the WVR than simply following along behind the radio telescope. Details of the operation of the positioner will be given in a later paper of this series.

### I. Control and Interface Module

This module consists of a local control/monitor panel and microcomputer. Using the local control panel the user can manually point the WVR, switch modes, or monitor any of the signal lines from either the microwave or positioner modules. These signals are displayed on a built-in digital voltmeter and are available via BNC jacks on the panel in order to drive a chart recorder. The control panel is intended primarily to assist in diagnosis of problems. In normal operation the control panel is switched to the REMOTE position and all control and data acquisition is exercised by the microprocessor. The interface to the microprocessor is set by the requirement of compatibility with the MK III data acquisition system. In early talks with colleagues at Goddard Space Flight Center and at the Haystack Observatory it was decided that the WVR would simply be treated as one of several devices on a daisy-chain type of serial interface. This interface has come to be known as the MAT (Microprocessor ASCII Transceiver) bus.

Originally it was believed that the new RS422/423 interface standard would be necessary to implement the long lines to the WVR that we thought might occur at some installations. Unfortunately, reliable components to implement the RS422/423 standard were not available in time to be implemented. A simple three wire RS-232 is now in use with short-haul modems for the WVR interface lines.

The baud rate of the WVR is set to any standard rate from 75 to 19,200 baud by a combination of software and jumper changes on the microcomputer board. At the present time it is set to 9600 baud. Implementation of the operating modes and command/respond sequence is primarily a matter of sufficient software in the microprocessor. In the current version of this software only the OPERATE mode has been implemented along with a set of primitive commands. Details of this software will be discussed in a later paper. The microcomputer contains an analog I/O subsystem that is used to digitize the various analog signals from the microwave package and the positioner. Our desire for 0.1 deg of position resolution dictated the choice of a 12-bit A/D converter (i.e.,  $2^{-11} > 0.1/360 > 2^{-12}$ ). Similarly, the quantization of the physical and brightness temperature measurements is 0.1 K. As indicated previously, the fluctuation level from the radiometer is on the order of 0.4 K. It is possible to average many samples in order to reduce these fluctuations. However, Clark (Ref 18), has pointed out this must be done with caution if one attempts to reduce the fluctuations below the quantization level.

## VI. In Retrospect

Table 3 summarizes the hardware costs and time expended as a result of our work on the WVRs. Hardware costs are referred to epoch October 1981 and for the most part must be inflated to the current date. An exception to this is the microprocessor hardware where many components are being reduced in price. The hardware, assembly, testing, and calibration costs are listed on a per unit basis. The development of the algorithm, microprocessor control software, and the data acquisition software is summarized as a total effort. The final accounting of hardware cost and assembly time was very close to our original estimates for this task in spite of severe price inflation. This was due largely to the fact that we were more efficient in assembly than originally estimated.

Having several WVRs under construction meant that some jobs could be done on a "production line" basis and there were always lots of little jobs that kept everyone busy.

The software part of the WVR task cost a great deal more than we originally estimated. There were essentially three reasons for this overrun. First, purchase and delivery of the JPL MK III data acquisition system was delayed, which meant that we had to purchase additional hardware for host computers other than the MK III computer. These additional microcomputers were intended to serve as "host" machines and were not in our original budget. In order to save dollars we purchased lowest-bidder type of equipment, which was a serious mistake. We saved a few dollars on the hardware but spent much more later in maintenance costs and schedules that had to be slipped because equipment was not working. The second reason for the software overrun was due to the fact that we simply underestimated the amount of system-level support we would need to keep several development systems running. The third and major reason for the overrun was due to the personnel turnover among the people working on the software. Over the three years of the task there were eight people who worked on the software or algorithm development. All worked part time and five left after several months on the job – an extremely inefficient way to develop software.

Had we realized at the time we started building these instruments that we would construct eight of them, we probably would not have used the conventional Dicke design. One can imagine a spectrum of design possibilities that range from a fully compensated Dicke radiometer utilizing full digital control and data acquisition to a single-horn total power instrument. The goal in any redesign effort should be to lower the parts count, thereby decreasing the hardware cost, reducing assembly time, lowering weight, size and power requirements, and increasing instrumental stability. At some point in the design spectrum one is inevitably faced with the possibility of cost/performance tradeoff that can limit the applicability of the instrument. It is our educated guess that the cost of the WVR could be reduced by 15-30% by simply reengineering the current design but keeping the same overall characteristics of the current instruments. Potential cost reductions of a factor of 2 are possible but entail some technical risk that would require a design study at the least and possibly the construction and testing of a prototype.

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**Table 1. Water vapor radiometer specifications**

(1) Physical characteristics	Weight, kgm	Volume, m <sup>3</sup>	Power, W
Microwave package	45	0.12	300-600
Positioner	34	0.074	0-450
Control module	26	0.15	120
<b>(2) Environmental</b>			
Operating range			
Ambient temperature	-10 to +50°C		
Wind speed	up to 65 km/hr		
Relative humidity	0 to 95% or until liquid water precipitates onto the horn cover		
Survivability			
Ambient temperature	-40 to +60°C		
Wind speed	up to 160 km/hr in stowed position		
Relative humidity	100%, sealed from rain and dust		
<b>(3) Microwave module</b>			
Operating frequencies	20.7 and 31.4 GHz		
Frequency stability	±12 MHz over operating range		
RF bandwidth	200 MHz		
IF bandwidth	100 MHz		
Integration time	software selectable in steps of 0.1 sec		
Signal range	3 to 400 K		
Antenna beamwidth	7° (full width at half power)		
Beam efficiency	>99.9% for ±15° around beam center supplied by two waveguide terminations held at fixed temperatures and supplemented with tip curve observations		
Calibration			
Base load	315 K		
Hot load	370 K		
Temperature stability	>±0.1 K/min		
<b>(4) Positioner</b>			
Coordinates	azimuth and elevation		
Range	0 to 355° in Az, 5 to 175° in El		
Slew rate	> 1.5 deg/sec (both axis)		
Position accuracy	1° both axes		
Readout accuracy	0.1° both axes		
<b>(5) Control and interface module</b>			
Operating modes			
(a) Local	all data and control lines available for monitoring by built-in digital voltmeter		
(b) Remote			
Interface	RS-232, twisted pair of cables		
Baud rate	110 to 19200 baud, selectable by software and hardware jumpers compatible with MAT <sup>a</sup> bus		
Protocol			
Data storage	temporary storage only, mass storage is the responsibility of the host computer		

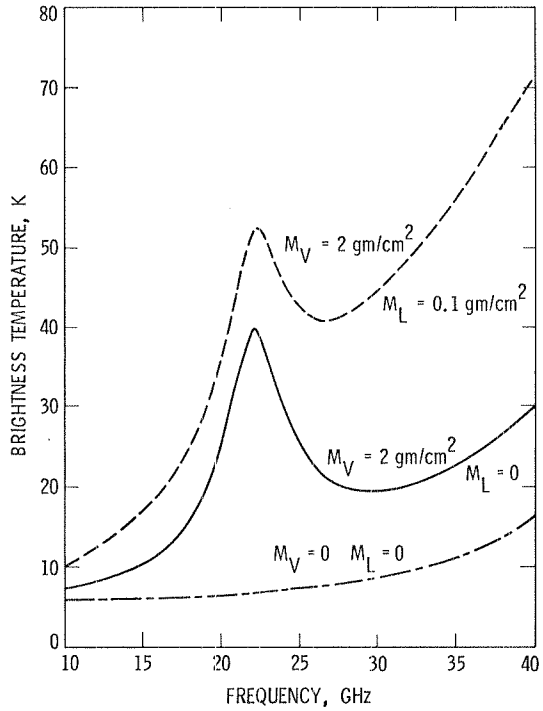
**Table 2. Extent of the near-field and beam volume ratios for WVR and radio telescope**

Diameter, m	Frequency, GHz	$L_{nf}$ , km	$R$	$\sqrt{R}$
64	8.4	108.00	29.2	5.4
64	2.3	32.00		
40	8.4	42.00	74.8	8.6
40	2.3	12.30		
25	8.4	16.40	191.5	13.8
25	5.0	10.40		
25	2.3	4.80		
9	8.4	2.10	1477.9	38.4
9	2.3	0.60		
4	8.4	0.40		
4	2.3	0.01		

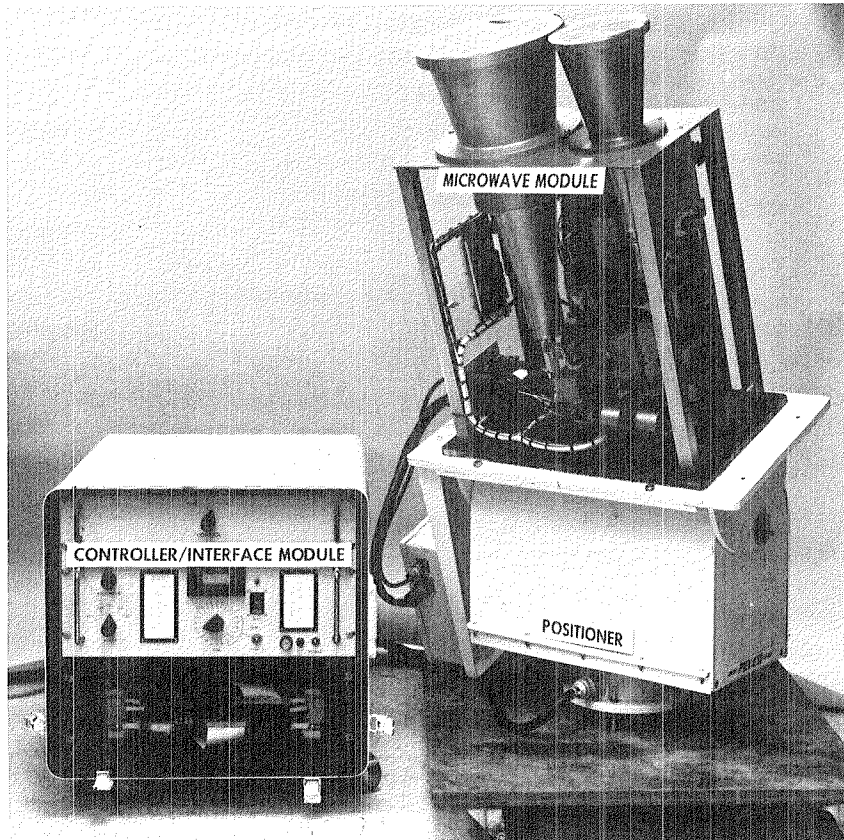
**Table 3. Cost summary**

A Cost per radiometer (as of Oct 1981)			
Components	K\$	Tasks	Man-month
Ferrite switches	7.3	Fabrication	3.0
LO/mixer/IF	8.5	Assembly and wiring	4.0
Horn antennas	3.1	Bench testing	3.0
Waveguide components	1.0	Engineering tests	0.8
Power supply	1.7	Calibration	2.0
Detectors	0.1		
Temp sensors and control	0.4		12.8
Enclosure (microwave)	0.6		
Misc components	1.2		
Control panel	0.8		
Microprocessor	4.5		
Positioner	2.8		
Cables and connectors	0.5		
	<u>\$32,500.0</u>		
B Support tasks			
Task		Man-years	
Controller software (3 versions)		1.2	
System support		0.8	
Data acquisition S/W		1.0	
Environmental test		0.2	
Algorithm development		2.0	
		<u>5.2 MY</u>	

<sup>a</sup>MAT = Microprocessor ASCII Transceiver used as the control/communicator standard in the Haystack/GSFC Mark III data acquisition system.



**Fig. 1. Zenith brightness temperature of the atmosphere between 15 and 40 GHz**



**Fig. 2. Photograph of the three WVR modules (cover removed from the microwave package)**



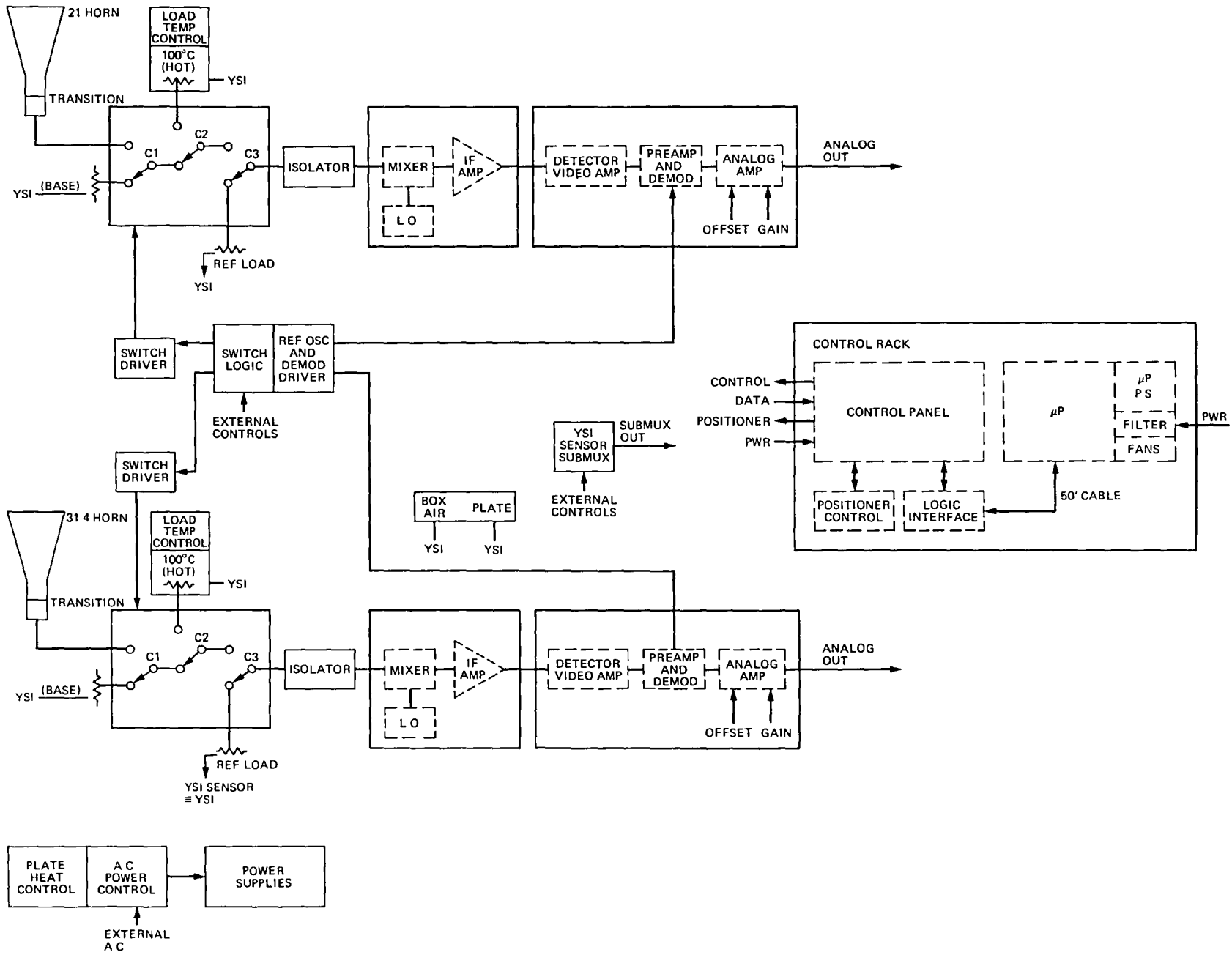


Fig. 3. Block diagram of the WVR

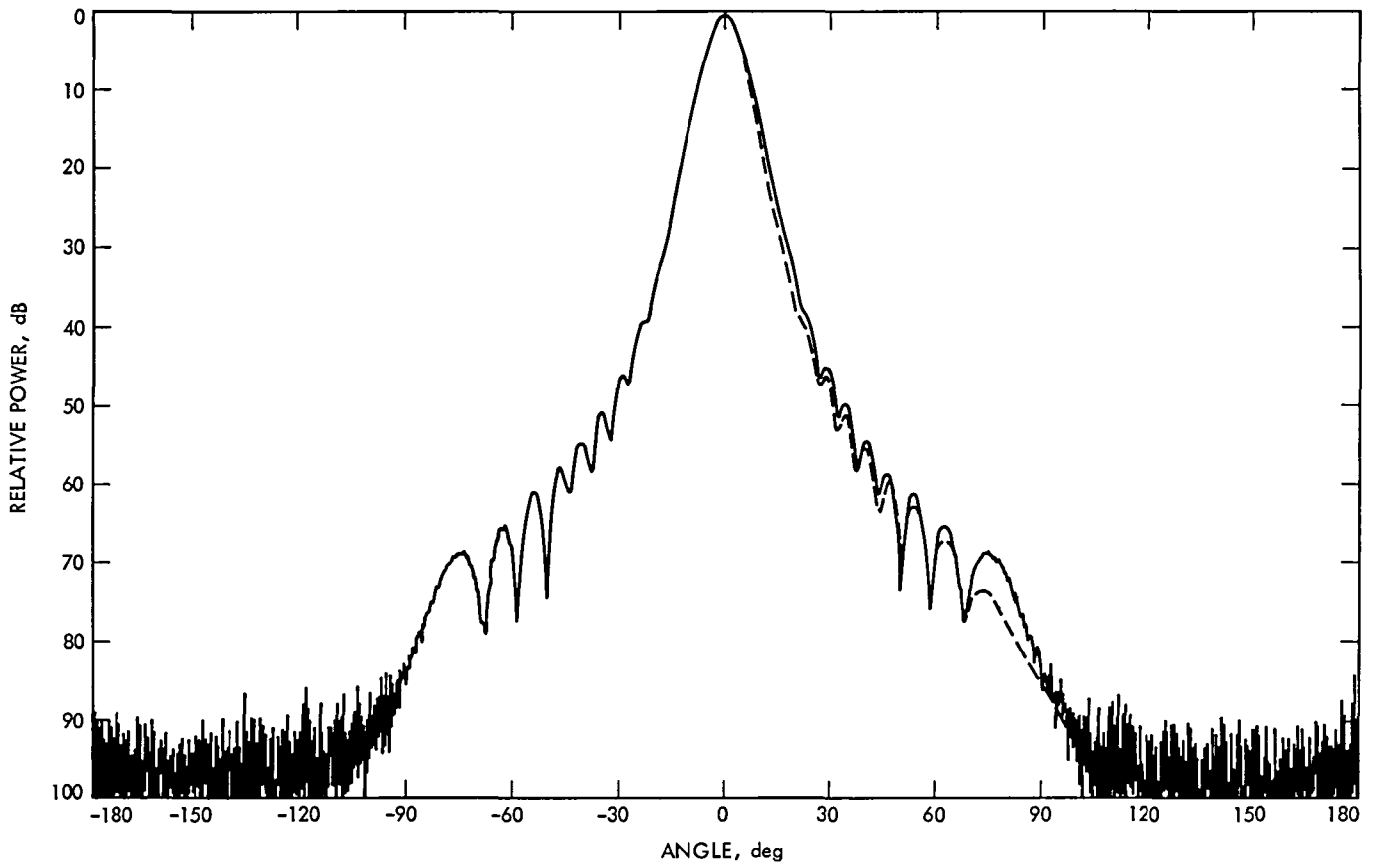


Fig 4 Antenna beam pattern of the WVR horn antenna

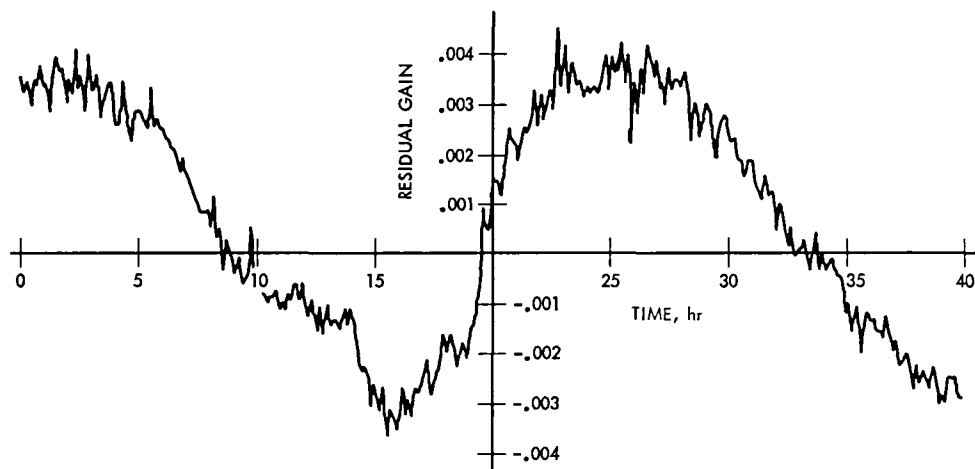


Fig 5 Gain of one of the WVR channels vs time

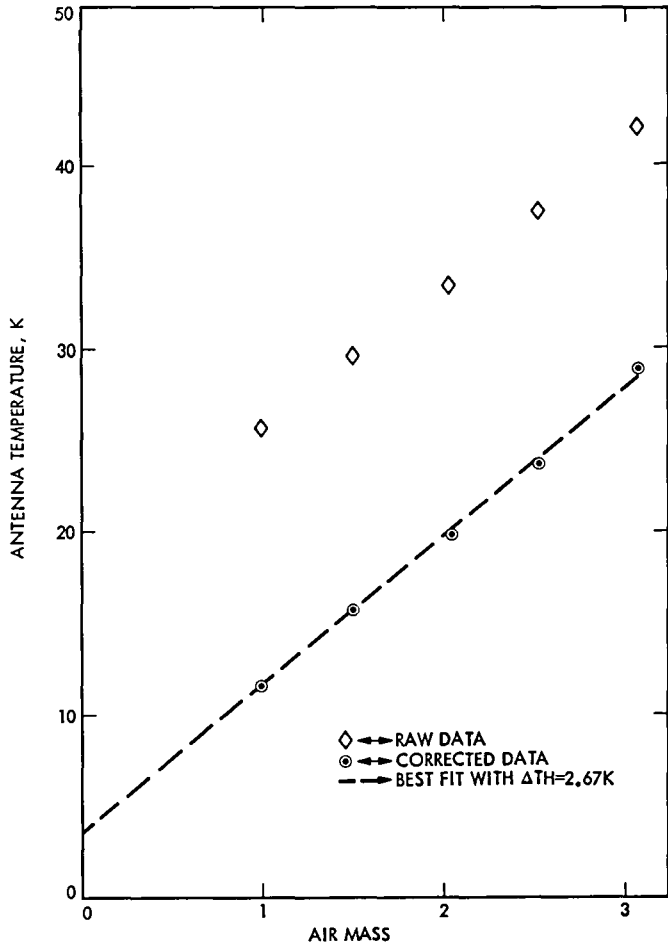


Fig. 6. Tip curve determination of the hot load correction

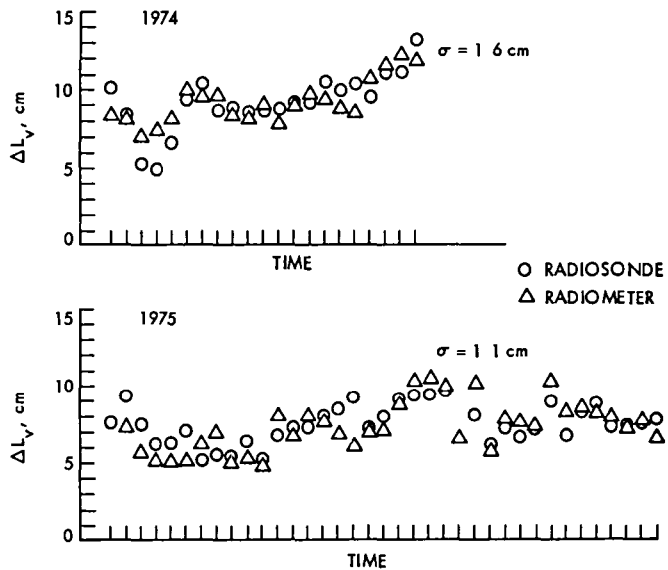


Fig. 7. Comparison of path delay as determined by radiosonde and WVR (El Monte, 1974 and 1975)

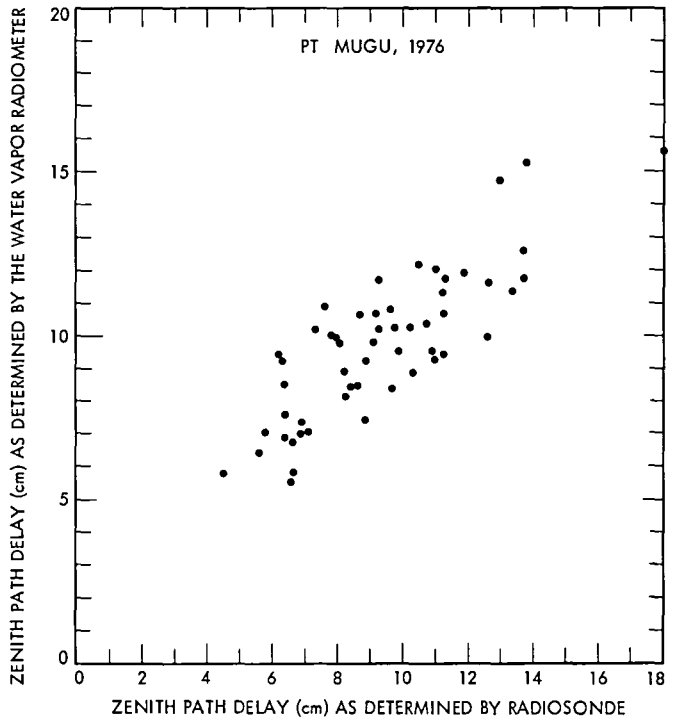


Fig. 8 Comparison of path delay as determined by radiosonde and WVR (Pt. Mugu, California, 1976)

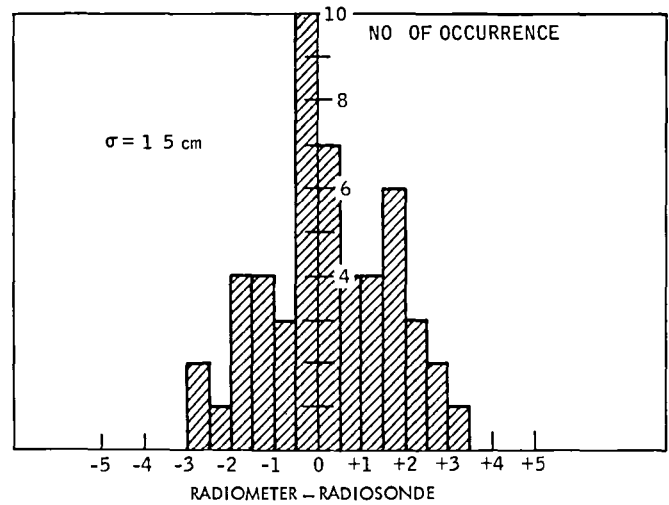
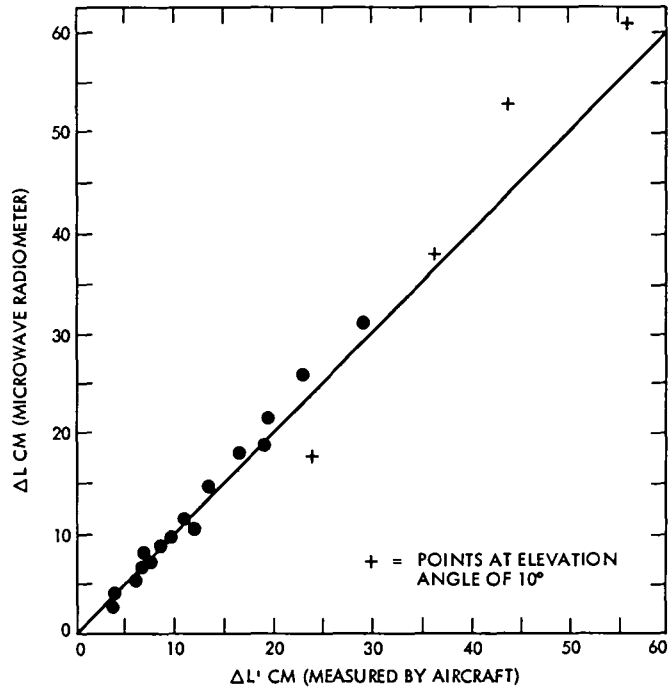
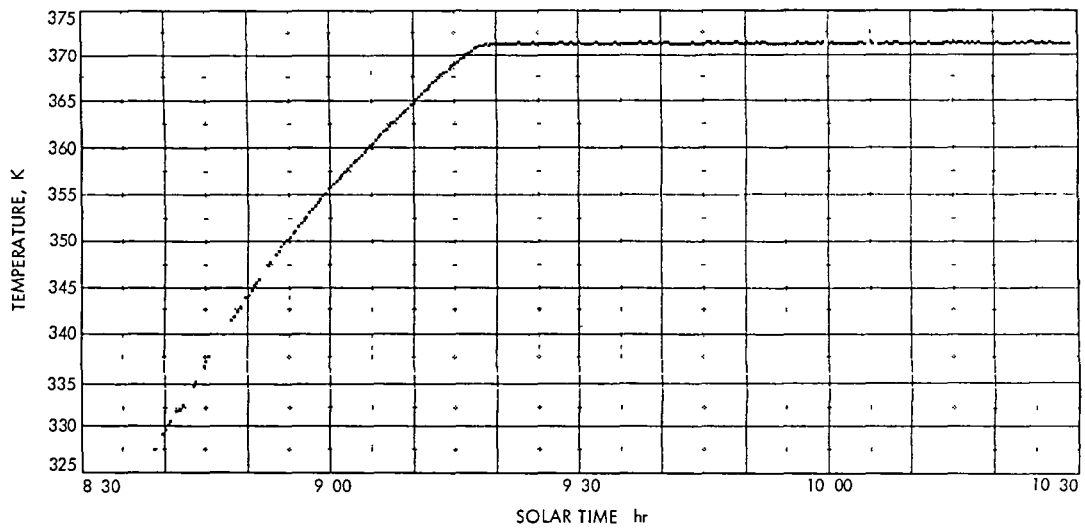


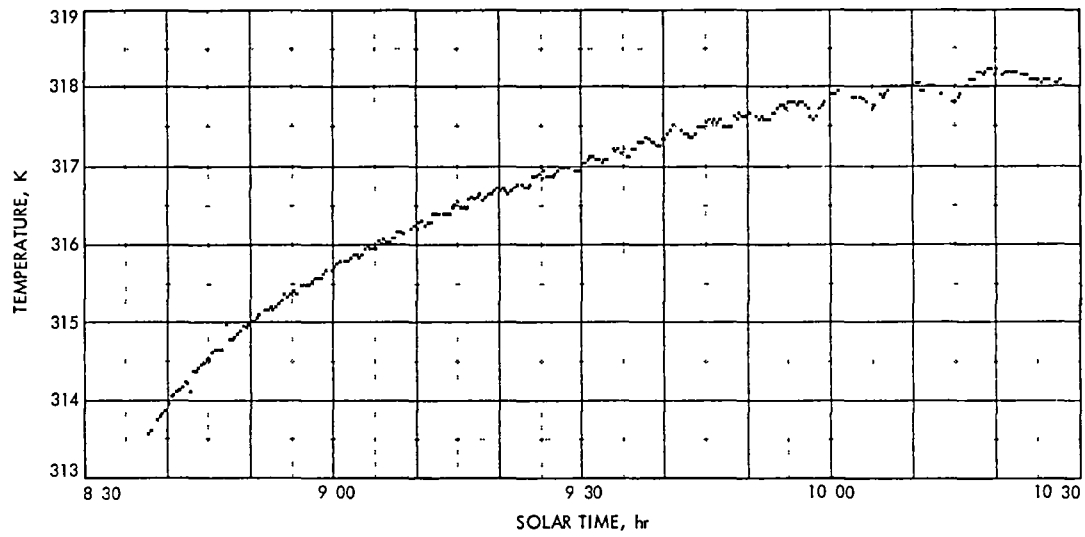
Fig. 9. Histogram of path delay residuals from the Pt. Mugu data



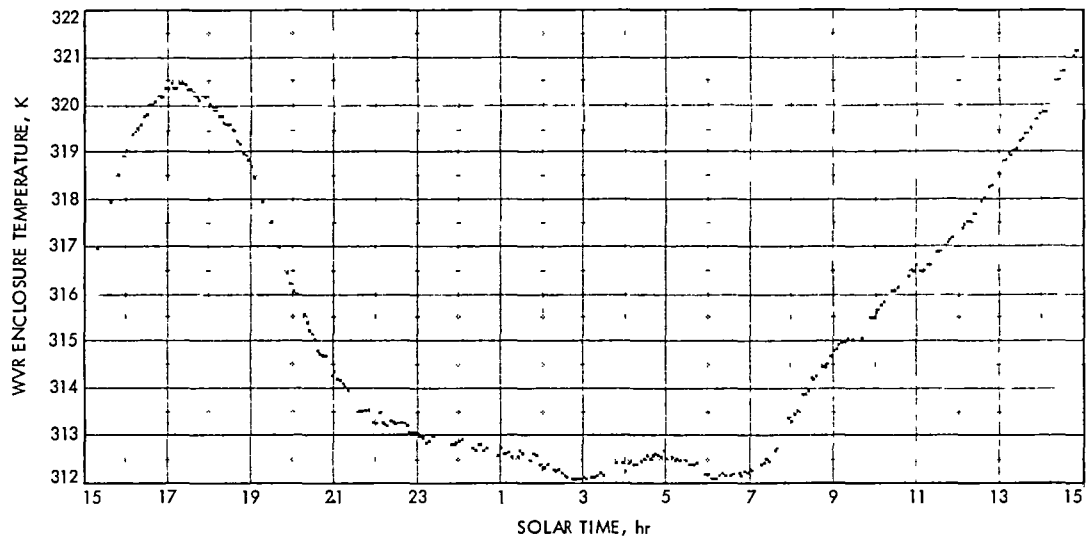
**Fig. 10 Comparison of path delay as determined by an instrumented aircraft and the WVR (Pt. Mugu)**



**Fig. 11 Warmup characteristics of the hot load**



**Fig. 12. Warmup characteristics of the base load**



**Fig 13. Enclosure temperature vs time**

# VLBI Collimation Tower Technique for Time-Delay Studies of a Large Ground Station Communications Antenna

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*A need for an accurate but inexpensive method for measuring and evaluating time delays of large ground antennas for VLBI applications motivated the development of the collimation tower technique described in this article. This article discusses supporting analytical work which was performed primarily to verify time delay measurement results obtained for a large antenna when the transmitter was at a collimation distance of 1/25 of the usual far-field criterion. Comparisons of theoretical and experimental results are also given.*

## I. Introduction

As accuracies of Very Long Baseline Interferometry (VLBI) systems for the measurement of time delay changes have continued to improve to current levels of less than 0.1 ns, a need has existed for studies to be made of accuracy-limiting errors associated with the microwave antenna itself. Two types of VLBI errors attributed to the antenna are (1) errors produced by unwanted multiple reflections (multipath signals) within the antenna propagation media and (2) undesired phase delay changes produced when the antenna is mispointed from boresight of the radio source.

The subject of antenna time delay and errors contributed by multipath signals has been studied in recent years both analytically and experimentally (Refs. 1–3). Results of studies made on some types of Deep Space Network (DSN) antennas tend to show that multipath problems can be mitigated by increasing the VLBI spanned bandwidth.

The second type problem, which is concerned with phase delay errors due to antenna mispointing, is one to which little attention has been given in the past. This subject is important because antenna mispointing can occur in practice due to intentional mispointing as in the case of consscanning VLBI systems, or unintentionally due to practical limitations on the accuracies of antenna pointing systems. Although some far-field VLBI antenna mispointing experiments have been performed with a 64-m DSN antenna, poor reproducibility of the results and insufficient experimental data have thus far prevented any definite conclusion to

be drawn concerning the effects of antenna mispointing on VLBI time delay data. There has been a strong need for quite some time now for a rigorous theoretical analysis to help predict the order of magnitude of phase delay changes that can occur for various VLBI type antennas when the mainbeam is pointed off the boresight coordinates of the radio source.

This article is concerned with supporting analytical work which was performed primarily to verify time delay measurement results obtained for a large antenna when the transmitter was at a collimation distance of 1/25 of the usual far-field criterion. The unexpected spinoff from this work was that since a collimation tower geometry is basic to many experimental configurations, the same theoretical equations can also be used to study phase delay changes as a function of antenna pointing offset angles from boresight. The analysis is applicable, not only for near-field but also for far-field studies.

## II. Theory

Application of the Lorentz Reciprocity Theorem yields the voltage coefficient of mutual coupling between two lossless, matched antennas to be (Ref. 4).

$$S_{12} = \frac{e^{j\Phi}}{4\sqrt{P_1 P_2}} \int_S (\bar{E}_1 \times \bar{H}_2 - \bar{E}_2 \times \bar{H}_1) \cdot \hat{n} dS \quad (1)$$

where  $\Phi$  is an arbitrary phase reference,  $(\bar{E}_1, \bar{H}_1)$  and  $(\bar{E}_2, \bar{H}_2)$  are the transmitted fields of each antenna,  $P_1$  and  $P_2$  are the radiated powers, and  $S$  completely encloses one of the antennas. This statement of Robieux's Theorem can be rewritten

$$S_{12} = \frac{e^{j\Phi} \int_S (\bar{E}_1 \times \bar{H}_2 - \bar{E}_2 \times \bar{H}_1) \cdot \hat{n} dS}{4 \left[ \frac{1}{2} \operatorname{Re} \int_{A_1} (\bar{E}_1 \times \bar{H}_1^*) \cdot \hat{n} dS \right]^{1/2} \left[ \frac{1}{2} \operatorname{Re} \int_{A_2} (\bar{E}_2 \times \bar{H}_2^*) \cdot \hat{n} dS \right]^{1/2}} \quad (2)$$

which, when squared, yields

$$\frac{P_R}{P_T} = |S_{12}|^2 = \frac{\frac{1}{4} \left| \int_S (\bar{E}_1 \times \bar{H}_2 - \bar{E}_2 \times \bar{H}_1) \cdot \hat{n} dS \right|^2}{\left[ \operatorname{Re} \int_{A_1} (\bar{E}_1 \times \bar{H}_1^*) \cdot \hat{n} dS \right] \left[ \operatorname{Re} \int_{A_2} (\bar{E}_2 \times \bar{H}_2^*) \cdot \hat{n} dS \right]} \quad (3)$$

In the event that one or both of the antennas does not have a clearly defined aperture, Eq. (1) can be rewritten

$$S_{12} = \frac{e^{j\Phi}}{4\sqrt{P_1 P_2}} \int_V (-\bar{E}_2 \cdot \bar{J}_1 + \bar{H}_2 \cdot \bar{J}_{m1}) dV \quad (4)$$

where  $\bar{J}_1$  and  $\bar{J}_{m1}$  are the source currents completely enclosed by  $S$  and  $\hat{n}$  (in Eq. 1) points out of  $V$ .

As a relatively simple application of Eqs. (1)-(4), consider a paraboloid and an on-axis electric dipole (Fig. 1). The paraboloid is fed by a focused, point-source feed located at its focus. The fields of the feed are

$$\bar{E}_f = V_0 e^{-j\beta L} \frac{e^{-jk\rho}}{\rho} \left( \frac{2F}{1 + \cos \theta_f} \right) (\hat{a}_{\theta_f} \cos \phi_f - \hat{a}_{\phi_f} \sin \phi_f) \quad (5)$$

Located at  $P(0, 0, R)$  on the Z-axis is an electric dipole  $\hat{a}_I d$ . Application of Eq (4) over the paraboloid aperture then yields

$$\frac{P_R}{P_T} = \frac{G_d G_p}{\left(\frac{4\pi R}{\lambda}\right)^2} \quad (6)$$

where the dipole gain is

$$G_d = 1.5 \quad (7a)$$

and the Fresnel-region gain of the paraboloid is

$$G_p = \left(\frac{\pi D}{\lambda}\right)^2 \left\{ \frac{\sin [k(D/2)^2/4R]}{[k(D/2)^2/4R]} \right\}^2 \quad (7b)$$

where  $D$  is the paraboloid diameter.

Equations (1)–(3) can be applied to the problem of coupling between two paraboloids (Fig 2)

- (1) Test paraboloid #1 is a large reflector with a point-source feed located at its focus. The feed pattern consists of arbitrary E- and H-plane patterns. Central blocking is included, but strut blocking is not. This reflector lies in the far-field of paraboloid #2.
- (2) Collimation paraboloid #2 is a smaller reflector with an axially-defocused point-source feed. Various feed patterns may be specified. Central blocking is included, but strut blocking is not. This reflector need not lie in the far-field of paraboloid #1, and it can be oriented arbitrarily with respect to #1.

Voltage coupling between the two reflectors is computed by means of Eq (1) and the corresponding power coupling by means of Eq (3). The integration surface  $S$  is a large coaxial circle located in front of paraboloid #1 and perpendicular to its axis. This surface is generally near to #1 and sufficiently large that the fields  $\bar{E}_1$  and  $\bar{H}_1$  are assumed to be negligible beyond the limits of the circle. In principle,  $S_{12}$  should be independent of the location of this surface, in practice the size and location of the circle must be determined empirically. The fields  $\bar{E}_1$  and  $\bar{H}_1$  on  $S$  are determined by geometrical optics and geometrical edge diffraction (Refs 5, 6).

The integration surface  $S$  is assumed to lie in the far-field of collimation paraboloid #2 so that the fields of this antenna may be expressed

$$\bar{E}_2 = \frac{e^{-jkR}}{R} [E_{2a}(\theta_{2a}) \cos \phi_{2a} \hat{a}_{\theta_{2a}} - H_{2a}(\theta_{2a}) \sin \phi_{2a} \hat{a}_{\phi_{2a}}] \quad (8)$$

which can be decomposed into rectangular components  $E_{2x}$ ,  $E_{2y}$ ,  $H_{2x}$ , and  $H_{2y}$  for integration in Eqs (1) and (3). The above formulations are contained in an appropriate computer program.

Provision is also made to rotate paraboloid #1 (Fig 3) about an offset azimuth axis ( $\Delta z_{az}$ ) and an offset elevation axis ( $\Delta x_{el}$ ,  $\Delta z_{el}$ ). In the rotated coordinate system the integration surface  $S$  rotates with #1 so that it remains symmetrically positioned on a plane normal to the axis of #1. However, it is no longer symmetric or normal with respect to the line between #1 and #2.

### III. Experimental Setup

Figure 4 shows a block diagram of the collimation tower experimental setup for time delay measurements using VLBI measurement techniques and associated equipment. At the collimation tower, a simulated VLBI radio source signal, provided by a broadband noise source, is fed into a power divider. One of the power divider outputs is a transmit-reference signal which is recorded



on a magnetic tape recorder simultaneously with clock signals furnished by a portable rubidium frequency standard. The other power divider output is fed to the collimation tower transmitter antenna which is 1.83 m (6 ft) in diameter. The signal is transmitted over a distance of about 1.6 km and illuminates both a 1.83-m (6-ft) antenna and a 26-m (85-ft) antenna. The smaller antenna is a receive-reference antenna whose time delay (including feed and interconnecting transmission lines) has been pre-calibrated while the 26-m antenna is the antenna whose delay is being measured.

For this test, the center frequency was 8.440 GHz and the VLBI spanned bandwidth 40 MHz. When boresighted to each other, the distances between the focal points of the transmit and receive antennas were approximately  $0.08 D^2/\lambda$  and  $16 D^2/\lambda$ , respectively, for the 26-m and 1.83-m receive antenna paths where  $D$  corresponds to the applicable receive antenna diameter and  $\lambda$  is the free space wavelength. Figure 5 shows the collimation tower at DSS 13 and the 1.83-m transmit antenna used in the test, while Fig. 6 shows the 26-m antenna (being calibrated) and the receive-reference 1.83-m antenna referred to in this article.

The experimental procedure was to first record the received signal for a duration of 2 minutes for the receive-reference antenna and then switch to the 26-m antenna under test and record its received signal for 2 minutes and then switch back to the reference antenna. This procedure was repeated for a period of 1 hour during which the downconverted received signals were recorded on magnetic tape. Simultaneously recorded were clock signals provided by the station hydrogen maser frequency standard.

The recorded data was later postprocessed on the Caltech-JPL VLBI correlator where received signal recordings were correlated against the transmit-reference signal data recorded at the collimation tower. The final experimental value was determined after correcting the VLBI correlator output data for the precalibrated delays of the receive-reference antenna and all interconnecting transmission lines and waveguides of the receive-reference antenna and the 26-m antenna up to the common switching port.

The advantages of switching between a reference antenna and the antenna under test is that common mode changes such as propagation media changes, multipath, and system drifts tend to cancel out. It is reasonable to expect that better accuracy and precision can be obtained from comparisons made with a calibrated delay standard than from direct absolute delay measurement over long distances.

#### IV. Comparisons of Theoretical and Experimental Data

The basic theory of antenna time delay for Cassegrain antennas has been discussed in Ref. 1. To compute the theoretical value of antenna delay for the results of this paper, the output phase of integrated fields received by the antenna is computed at the two microwave frequencies involved in the VLBI experiment (resolving any multiples of  $2\pi$  radian ambiguities). Then group delay is computed by the "VLBI bandwidth synthesis" equation (Ref. 7) which is equivalent to the "phase-slope" method of determining group delay (Ref. 8).

One of the significant results of this experiment was that a measurement precision of 10 ps (corresponding to 0.3 cm) was achieved. A second important result was that a significant difference was found between the theoretical 26-m antenna delay determined from mutual coupling theory and the value calculated from the optical ray pathlength of a single ray traveling from the collimation tower antenna to the feed horn phase center of the 26-m antenna. The near-field delay, based on mutual coupling theory described in Section II of this article, was calculated to be 0.6 ns longer than the value determined from the ray optical pathlength calculation. A third significant result was that the experimentally determined delay of the 26-m antenna agreed with the mutual coupling theoretical value to within  $0.2 \pm 0.2$  ns probable error. The agreement between mutual coupling near-field and experimental results is excellent when taking into consideration the relatively small VLBI spanned bandwidth of the 40 MHz that was used in the experiment.

It should be pointed out that the ray pathlength calculation method is only approximately valid for a focused antenna and at far-field distances, but is a method commonly used erroneously for near-field geometries as well. The mutual coupling theoretical result is based on numerical integrations over the entire 26-m-dia aperture using subreflector pattern data (Ref. 9) applicable to the X-band feed system that was installed on the 26-m Cassegrain antenna during the experiments. Since the subreflector pattern is referenced to the focal point of the paraboloid, it is applicable to the theory and computer program described in this paper.

Many collimation tower test setups are plagued with ground multipath problems. However, a previous study made by the Howland Co. of Atlanta, Ga., showed that for the particular ground terrain involved in this collimation tower setup and with the receive-reference antenna mounted on top of the 26-m antenna as shown in Fig. 6, ground reflected multipath signals would not interfere with accurate measurements of the delays of the two receive antennas. In addition, independent time delay measurements made on the 26-m antenna with a time domain technique similar to the FM/CW technique described in Ref. 10 confirm the VLBI results discussed in this paper. With the time domain technique, it was possible to identify and separate out the ground multipath signals and measure the delay of the primary signal accurately. Although time domain techniques can also be used to measure time delays, the FM/CW instrumentation is primarily a diagnostics tool and is not part of the standard DSN instrumentation.

Figure 7 shows a plot of power pattern for an elevation angle scan for a test frequency of 8.420 GHz and at the collimation tower geometry and distances described above for the transmit 1.83-m antenna and the receive 26-m antenna. Both theoretical and experimental data are shown. The theoretical pattern is based on uniform illumination of the 26-m antenna aperture. Later comparisons will be made where the theoretical pattern is based on subreflector pattern data (Ref. 9). The experimental data was interpolated from a linear power pattern recording which had dB calibration marks provided through the use of a rotary vane attenuator in series with the transmitter waveguide. Since the data had to be extracted from the linear power recordings, it is estimated that the experimental data shown in Fig. 7 is accurate to only  $\pm 0.5$  dB above -6 dB,  $\pm 1.0$  dB between -6 dB and -10 dB, and  $\pm 1.5$  dB below -10 dB. It is significant to note in the power pattern that both experimental and theoretical results show three bumps on the main beam. Also note that the main beam beamwidths between the -10 dB points are close to 0.8 degree for both mutual coupling theory and experimental results at this near-field distance of  $0.08 D^2/\lambda$ . For comparison purposes, when the separation of the transmit antenna and the 26-m receive antenna changes from this near-field distance to far-field distances, the beamwidth between -10 dB points changes from the 0.8 deg value to about 0.15 deg. This far-field beamwidth value was determined by performing numerical integration of the X-band subreflector pattern over the 26-m antenna aperture at 8.420 GHz and computing far-field patterns.

Also of interest is the value of  $P_{R1}/P_{T2}$  calculated from mutual coupling theory (Eq. 3) as compared to the value calculated from Friis transmission loss formula (Ref. 11). For the near-field collimation tower distance involved in the experiment, the 26-m antenna received power level was calculated from mutual coupling theory (and use of Ref. 9 subreflector data) to be smaller than the received power calculated from the Friis formula by 13.6 dB. The experimentally measured received power was less than the Friis value by  $13.5 \pm 1.5$  dB. The large uncertainty associated with the measured value is due to uncertainty in the efficiency of the large antenna at 8.420 GHz during earlier installation periods, and uncertainty as to whether the receive and transmit antennas were optimally boresighted to each other at the particular time of this measurement. Despite the large uncertainty in this particular experimental result, it is significant that the agreement with the mutual coupling theoretical value is still quite good.

At the time that initial boresighting and power measurements were being made, it was thought that the large difference between actual received power and the value based on Friis formula was a major experimental leakage problem. Had the theoretical results of this article been available prior to this measurement, a greater understanding of the near-field measurement procedure would have been available and greater care would have been exercised in boresighting procedures. Termination of funding for this work and recent conversion to an updated feedcone at DSS 13 make it difficult to repeat this series of experiments in the near future.

If at some time in the future, funding does become available, it would be of great interest and value to the DSN to perform additional experimental measurements with the collimation tower technique and make accurate comparisons to theory, especially with regards to studying phase delay change vs offset from boresight angles.

## V. Conclusion

This article has described the development of theoretical equations specifically developed for practical collimation tower test setup geometries. A brief description has been given of the VLBI collimation tower technique that was used for the measurement of antenna delay. An agreement of 0.2 ns was obtained between theoretical and experimental values of time delay for the DSS 13 26-m antenna at a collimation tower distance that was  $1/25$  the accepted  $2D^2/\lambda$  far-field criterion. These results apply to measurements performed with VLBI equipment at 8.440 GHz and 40 MHz spanned bandwidth.

Based on the results presented in this article, one can conclude that the collimation tower technique has the potential of being very useful for checkout and diagnostics of the performance of a large ground station antenna. Now that the necessary theoretical equations are available, theoretical values calculated for actual near-field geometries can be used as a criterion for expected performance. Once the near-field performance of an antenna is verified experimentally, then transformations can be made on the near-field data to predict performance in the far-field.

## Acknowledgments

The VLBI collimation tower technique described in this article was suggested by Hugh Fosque of NASA. Measurements of received power vs elevation angle were made by Juan Garnica of Bendix Field Engineering Corporation.

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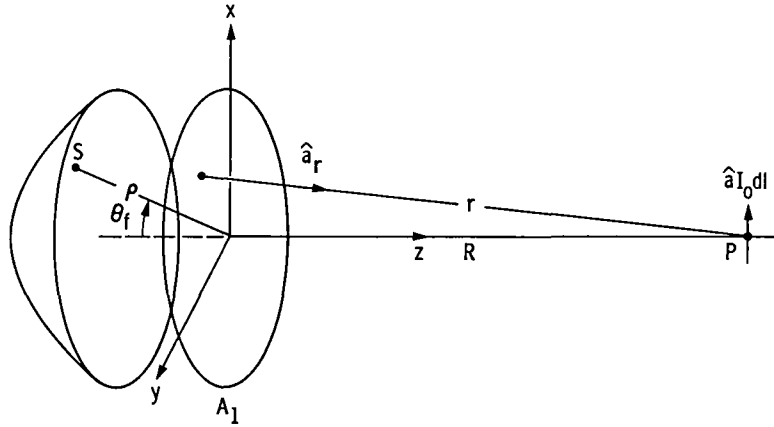


Fig 1. Coupling between paraboloid and on-axis electric dipole

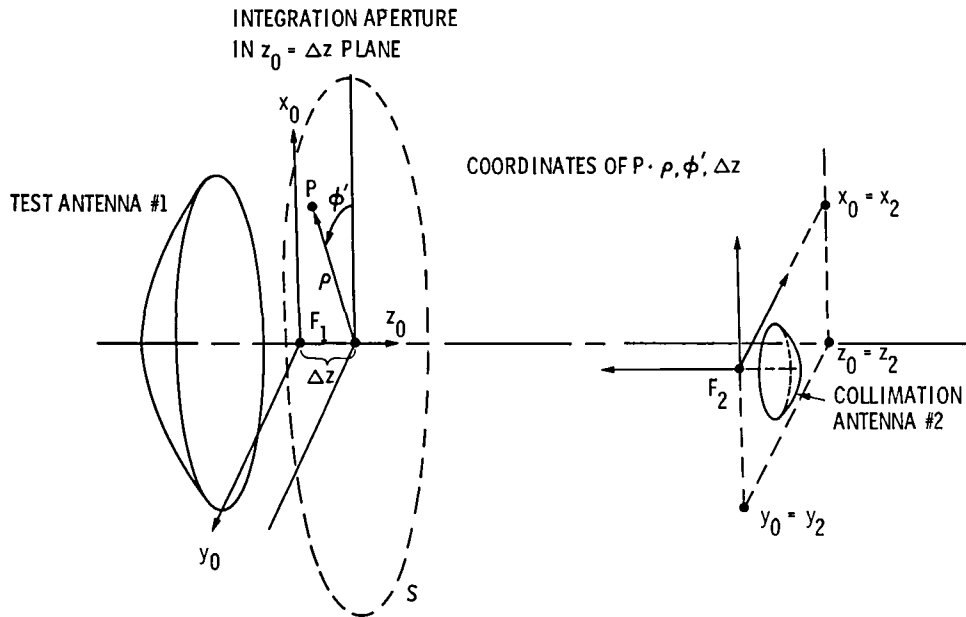


Fig 2. Coupling geometry for two paraboloids

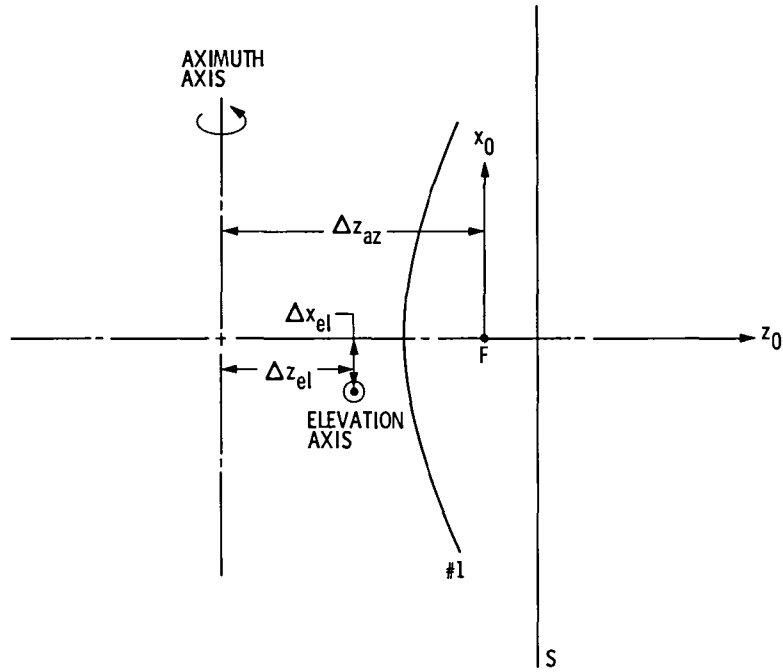


Fig. 3. Geometry of Az-EI rotation

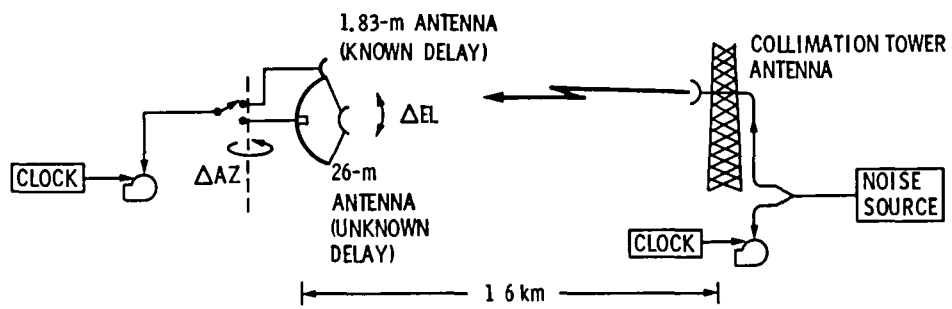
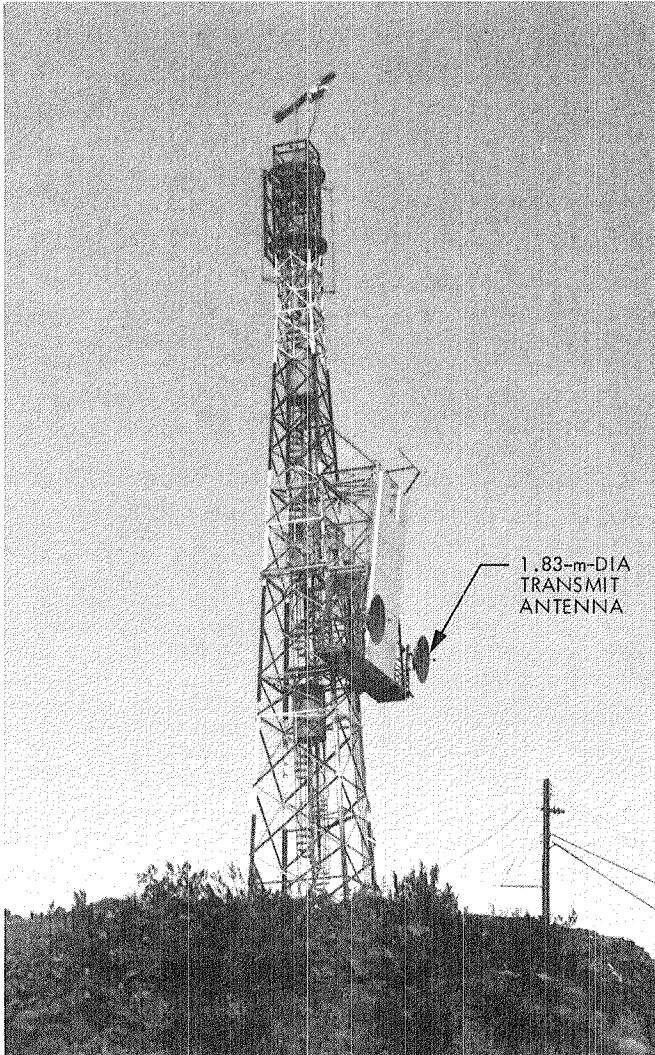
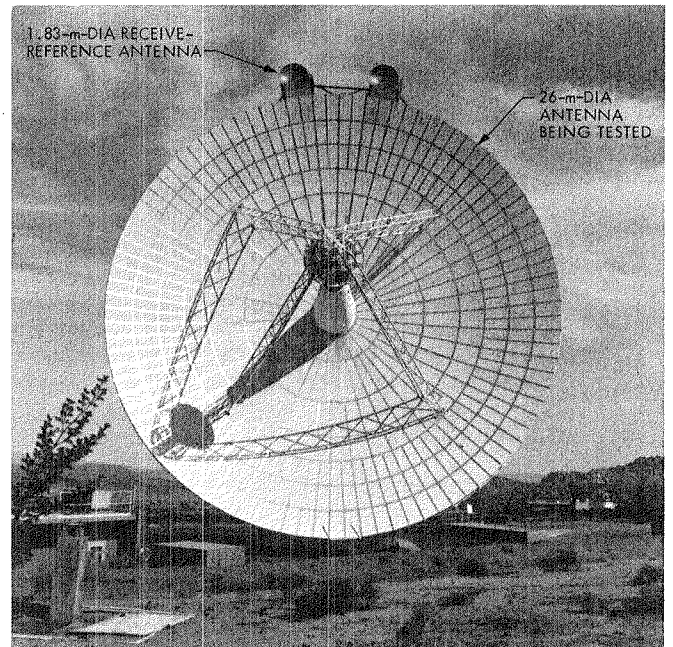


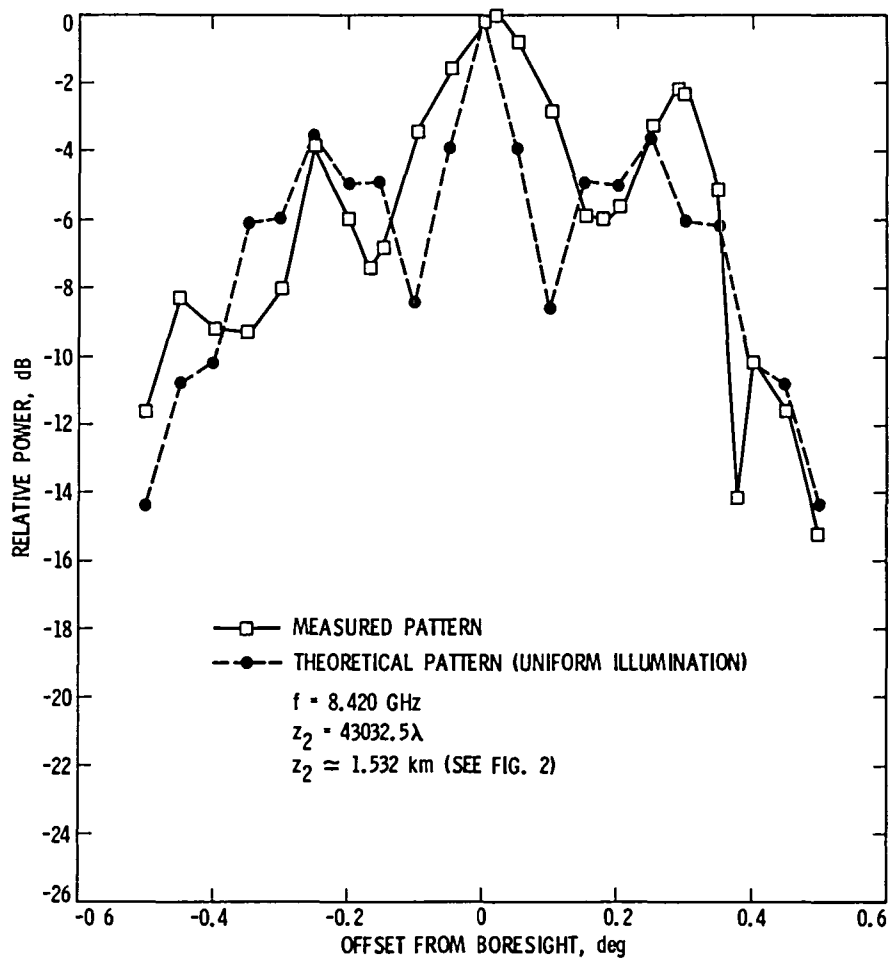
Fig. 4. Collimation tower geometry for the measurement of antenna time delay using VLBI technique and instrumentation



**Fig. 5. DSS 13 collimation tower and 1.83-m (6-ft) diameter transmit antenna. The S-band feed shown in photograph was replaced by an X-band feed**



**Fig. 6. 26-m-dia. antenna and 1.83-m-dia. receive-reference antenna at DSS 13. The S-band feed of the 1.83-m-dia. antenna shown in photograph was replaced by an X-band feed**



**Fig. 7. Received power vs offset angle scanning in the elevation angle plane at 8 420 GHz when the 26-m paraboloidal antenna was located 1 53 km distance from the 1.83-m-dia transmit (collimation tower) antenna**

# A Near-Optimum Receiver Structure for the Detection of *M*-ary Optical PPM Signals

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*A class of receivers called “conditionally nulling receivers” is defined for quantum noise limited optical communications. These receivers have the ability to decide at each moment in time whether or not to coherently combine a predetermined local oscillator field with the received optical field, prior to performing an energy measurement (photodetection) on the combined field. Conditionally nulling receivers are applicable to pulse position modulation (PPM) and related modulation schemes, which have the property that, at each moment in time, the transmitted signal is in one of only two states, ON or OFF. The local oscillator field which may or may not be added by the receiver is an exact replica of the negative of the received ON field, hence, the receiver can exactly null the ON signal if the ON signal is present and the receiver chooses to use the local field.*

*An ideal conditionally nulling receiver achieves very nearly the same error probability (within a multiplicative factor varying from 1 to 2.15) as the optimum quantum measurement for quantum noise limited detection of *M*-ary PPM signals. In contrast, other known receiving methods, such as direct, heterodyne, and homodyne detection, are exponentially suboptimum.*

*The performance of receivers which are only approximately conditionally nulling receivers, due to imperfect nulling, is also investigated for the quantum limited PPM detection problem. Imperfect nulling is assumed to be caused by a phase discrepancy between the received ON field and the local nulling field. The performance of the imperfect conditionally nulling receiver is found to degrade rapidly to that of a direct detection receiver in the presence of nonzero phase error.*

## I. Introduction

In many optical communications applications, the predominant “noise” that limits system performance is the fundamental uncertainty of measurement predicted by quantum theory. However, explicit calculation of the ultimate

communication inaccuracy legislated by quantum noise has been accomplished for very few problems (Refs 1-4), and the specification of physical receiving structures for achieving optimum performance is virtually nonexistent (except, see Ref 5). Conventional receiver structures based on familiar physical devices exhibit peculiar quantum noise limited



behavior of their own, but their performance falls short of the fundamental quantum limit

In this article a new receiver structure is defined for quantum limited optical communications. Receivers with this structure have the ability to decide at each moment in time whether or not to coherently combine a predetermined local oscillator field with the received optical field, prior to performing an energy measurement (photodetection) on the combined field. These receivers are termed “conditionally nulling receivers,” because the local oscillator field which may or may not be added by the receiver is an exact replica of the negative of one of the possible received fields. An ideal conditionally nulling receiver achieves very nearly the same error probability (within a multiplicative factor varying from 1 to 2.15) as the optimum quantum measurement for quantum noise limited detection of  $M$ -ary PPM signals. In contrast, other known receiving methods, such as direct, heterodyne, and homodyne detection, are exponentially suboptimum.

The purposes of this article are threefold: (1) to set forth a general definition of the class of conditionally nulling receivers, (2) to broaden the definition to include receivers which are only approximately conditionally nulling receivers, and (3) to compare the performance of ideal and imperfect conditionally nulling receivers applied to the quantum noise limited  $M$ -ary PPM detection problem.

## II. Applicable Modulation Schemes

The useful application of conditionally nulling receivers is restricted to PPM and related modulation schemes, which have the following characteristic: at each moment in time, the transmitted signal is in one of only two states, ON or OFF. The OFF state is assumed to correspond to a field of amplitude zero, and the ON state corresponds to a single nonzero field. At any given instant in time, there may be  $m$  possible signals in the ON state and  $m_o$  possible signals in the OFF state, and  $m, m_o$  may vary with time. Multiple amplitude levels to help distinguish among the  $m$  ON signals are not permitted, but the common ON state amplitude level may vary with time. Any receiver attempting to decipher this type of signaling can be viewed as an assembler of *binary* information, favoring either the  $m$  ON signals or the  $m_o$  OFF signals collectively, extracted from the received signal at each instant in time.

$M$ -ary PPM signaling fits nearly in this category, because at every moment in time ( $M - 1$ ) of the  $M$  possible signals are OFF and only one possible signal is ON. PPM is presented as the prototypical modulation scheme for conditionally nulling receivers, and all of the performance analysis in this article is based on PPM.

Three examples of applicable modulation schemes are shown in Fig. 1a, b, c. Figure 1a shows standard 4-ary PPM, and Fig. 1b shows a PPM-complementary modulation scheme, which is discussed at length in the next section. Figure 1c shows a 4-slot coding scheme in which 6 possible messages are represented by 6 different pulse doublets.

## III. An Unconditional Role Reversal Strategy

It is well known that direct detection does not distinguish symmetrically between ON and OFF signals. For example, under quantum noise limited conditions, the detection of a single quantum of energy is sufficient to positively identify the ON signal, but equally conclusive confirmation of the OFF signal is not possible. It is reasonable to ask whether simple direct detection optimally utilizes this inherent asymmetry as it attempts to distinguish at each moment in time between  $m$  ON signals and  $m_o$  OFF signals. Perhaps more useful information could be obtained if the  $m$  signals were OFF and the  $m_o$  signals were ON, rather than vice-versa.

A simple, complete reversal of the roles of the ON and OFF signals does not necessarily lead to improvement. For example, consider quantum limited minimum error probability detection of the “role-reversed”  $M$ -ary PPM signal set shown in Fig. 1b. With standard PPM signaling, if the  $j$ th hypothesis  $H_j$  is true, the signal is ON during the  $j$ th signaling slot and OFF during the remaining ( $M - 1$ ) slots. On the other hand, with “role-reversed” PPM, the signal is OFF during the  $j$ th slot and ON during the remaining ( $M - 1$ ) slots, under hypothesis  $H_j$ . A direct detection receiver is certain to record a zero energy measurement during at least one slot (the OFF period of the true signal), and it errs only if additional zero energy measurements are also obtained. A performance expression for this problem is easy to write down, if equal a priori probabilities and equal ON signal slot energies are assumed.

$$P'_\epsilon = \sum_{m=0}^{M-1} \frac{m}{m+1} \left(\frac{M-1}{m}\right) p^m (1-p)^{M-1-m} \quad (1)$$

Here,  $P'_\epsilon$  is the error probability,  $p$  is the probability of measuring zero energy from the ON signal during a single slot ( $p = e^{-E}$ , where  $E$  is the mean detected ON signal slot energy in units of photons), and  $m$  indexes the number of slots in addition to the true signal slot in which no photons are observed. The sum in (1) may be collapsed to the form

$$P'_\epsilon = \frac{1}{Mp} [(1-p)^M - 1 + Mp] \quad (2)$$

The corresponding performance expression for straightforward PPM detection is well known,

$$P_{\epsilon}^o = \frac{M-1}{M} p \quad (3)$$

and it can be shown that

$$P_{\epsilon}^o \leq P_{\epsilon}^{\prime}, \text{ for all } p \in [0, 1] \quad (4)$$

Thus, “role-reversed” PPM appears to be inferior to straightforward PPM, even with no consideration given to its apparently increased requirements for average transmitter power (requiring  $(M - 1)E$  detected photons per symbol instead of  $E$ )

## IV. The Ideal Conditionally Nulling Receiver

### A. General Description

It turns out that the conclusion of the previous section is no longer valid if the ON-OFF role reversal can take place at the receiver and if the receiver is able to continually choose whether or not to perform the reversal, based on its prior observations. Note that the receiver can accomplish the role reversal on uncorrupted ON-OFF signals by coherently adding the negative of the ON signal prior to detection, i.e., by nulling the ON signal. Exact role reversal at the receiver may prove impossible under two nonideal conditions: (1) if the received signal has been corrupted by a noisy channel, or (2) if the receiver cannot produce an exact replica of the ON signal.

Any receiver which has the power to precisely reverse the roles of the ON and OFF signals and, in addition, is able to decide at each instant in time  $t$  whether or not to perform this operation, based on its own observations prior to  $t$ , is termed an ideal conditionally nulling receiver, or ideal conditional nuller for short. A receiver which tries to act like a conditional nuller, but falls short due to an inability to perform the ON-OFF role reversal exactly, is called a conditionally nulling receiver with imperfect nulling.

A block diagram of the conditionally nulling receiver structure is shown in Fig 2. The received field at any given time  $t$  is either the ON signal, denoted  $S(t)$ , or else 0. At the receiver a replica of the negative of  $S(t)$  is generated. Provision is made to coherently combine this field with the received field, but the combining is subject to an on-off switch. The combined field (or received field only) impinges upon a photodetector, and the output of the photodetector (with assumed conditionally Poisson statistics) is used both to control the

position of the on-off switch and eventually to infer which message was sent.

### B. The Quantum Limited PPM Detection Problem

It has been shown elsewhere (Ref 6) that there exists a conditionally nulling receiver for quantum limited detection of  $M$ -ary PPM signals which approaches the error probability performance of the optimum quantum measurement within a *multiplicative* factor no greater than 2.23, over all ranges of  $M$  and slot energy  $E$ . This result is significant because the performance of the direct detector in (3) is *exponentially* inferior to the optimum. In this section, the structure of the ideal conditional nuller and its comparative performance are briefly described.

**1 Nulling strategy and decision algorithm** The ideal conditionally nulling receiver for the quantum limited PPM problem is defined as follows. During the first signaling slot, the ON signal is nulled and an energy measurement is performed. If no photons are detected, this is considered a partial confirmation of the first hypothesis relative to the remaining  $(M - 1)$ , and the receiver will continue to believe this hypothesis unless it subsequently obtains sufficiently conclusive evidence in favor of another. Given this state of affairs at the end of the first signaling slot, the receiver decides to forgo nulling the ON signal for the remaining  $(M - 1)$  slots and henceforth to simply direct detect, so that any future evidence impeaching the first hypothesis will automatically confirm with certainty one of the others. If, on the other hand, one or more photons are detected during the first signaling slot, the first hypothesis is completely contradicted and the receiver's task is reduced to the  $(M - 1)$ -ary version of the same problem, since no information discriminating among the remaining  $(M - 1)$  hypotheses has yet been obtained. In this case, the receiver proceeds to the second signaling slot and again nulls the ON signal, this time looking for partial confirmation of the second hypothesis if no photons are detected or complete denial of it otherwise. In the former case the receiver discontinues nulling, while in the latter it proceeds to use nulling to test for the third hypothesis, and so forth. The receiver simply continues to null the ON signal until such time as it obtains partial confirmation of a specific hypothesis by the measurement of zero energy throughout the corresponding PPM slot, and afterward it direct detects through all remaining time slots.

The nulling strategy and decision algorithm for the ideal conditional nuller can be represented by a binary tree diagram, as illustrated in Fig 3 for the case  $M = 4$ . The four levels of branches in the tree correspond to the four PPM slots. Each node in the tree is labeled “N” or “D” to denote whether nulling mode or direct detection mode is used during the next

slot. The branches are labeled “ $\geq 1$ ” or “0” to designate whether or not at least one photon is detected during the corresponding slot. Some paths (such as those containing two “ $\geq 1$ ” branches from “D” nodes) have probability zero under the assumed ideal conditions, and these are marked “impossible”. The end nodes in the tree which are reached with non-zero probability are all marked with the corresponding optimum decision.

**2. Performance** The performance of the ideal conditional nuller in quantum limited conditions is easy to calculate recursively, and a simple closed form expression is obtained for the case of equal a priori probabilities and equal slot energies  $E$ , for details, see Ref. 6

$$P_e = \frac{1}{M} [(1-p)^M - 1 + Mp] \quad (5)$$

where  $p = e^{-E}$  as before. Note that the error probability in (5) is smaller by a factor of  $p$  than the error probability obtained for the unconditional nulling strategy in (2).

The error probability achievable by the optimum quantum measurement is also known for this problem, and it is given by (Ref. 3)

$$P_e^* = \frac{M-1}{M^2} [\sqrt{1+(M-1)p} - \sqrt{1-p}]^2 \quad (6)$$

The three error probabilities  $P_e$ ,  $P_e^*$ ,  $P_e^o$  are plotted in Fig. 4a, b, c as a function of the average detected slot energy  $E$ , for  $M = 2, 16, 256$ . The conditional nuller's performance is seen to track the performance of the optimum quantum measurement very closely in all situations. In fact, it can be demonstrated numerically that the deviation is never more than a multiplicative factor of 2.15, i.e.,

$$1 \leq P_e/P_e^* \leq 2.15 \quad \text{for all } E, M \quad (7)$$

On the other hand, the performance of the direct detection receiver is exponentially inferior to that of the optimum measurement. This inferiority is most apparent in the  $M = 2$  graph, in which the (logarithmic) slope of the near-optimum  $P_e$  curve is exactly double the (logarithmic) slope of the  $P_e^o$  curve. As  $M$  increases, larger values of  $E$  are needed before the direct detection curve begins to diverge from the other two.

## V. Conditionally Nulling Receivers with Imperfect Nulling

The potential performance advantage of conditionally nulling receivers for quantum limited PPM detection has been demonstrated, but there remains a delicate practical problem in implementing the nulling operation. Perfect amplitude and phase coherence is required in order to exactly null the ON signal, and such precision is probably impossible for the receiver to achieve. Thus, it is important to investigate how the performance of a conditional nuller degrades with imperfect nulling.

### A. The Effect of Imperfect Nulling

A general definition of the class of conditional nullers with imperfect nulling will not be given here. As stated earlier, imperfect nulling at the receiver might be inevitable in principle due to random channel disturbances imparted to the transmitted signal, or else it may result simply from the receiver's inability to produce an exact replica of a deterministic ON signal. Only the second case is considered in this article.

A conditionally nulling receiver with imperfect nulling differs from an ideal conditional nuller in one important respect. In the absence of nulling, both receivers continually obtain by direct detection information distinguishing an ON signal of mean slot energy<sup>1</sup>  $E$  from an OFF signal of slot energy 0. After perfect nulling, the ON signal is converted to slot energy 0 and the OFF signal to mean slot energy  $E$ , a perfect reversal of roles. After imperfect nulling, on the other hand, the ON signal is reduced to some nonzero slot energy  $E'_o$ , and the OFF signal is imparted slot energy  $E'$  not necessarily equal to  $E$ . It is assumed that the nulling operation is at least accurate enough that  $E'_o < E'$ . However, the slot energies  $E'$ ,  $E'_o$  will always be less distinguishable via direct detection than the slot energies  $E$ , 0, and so the efficacy of attempting an imperfect nulling operation is drawn into question.

### B. The Quantum Limited PPM Detection Problem

The remainder of the article evaluates the impact of imperfect nulling on the performance of the ideal conditionally nulling receiver found to be near-optimum for the quantum limited PPM detection problem.

**1. A reasonable nulling strategy** A slight generalization of the ideal conditional nuller's operation is required in order

<sup>1</sup>For simplicity, it is assumed that the mean slot energies  $E$ ,  $E'_o$ ,  $E'$  are derived from mean conditional Poisson intensity functions which are constant over a slot time. This assumption eliminates the need to worry about count data records over time intervals finer than the slot interval.

to adapt it to imperfect nulling. With imperfect nulling, the nulled signal can never be entirely contradicted, and in fact may actually be corroborated relative to the opposite signal by the detection of a small, nonzero number of photons. Thus, a natural generalization of the conditionally nulling receiver described earlier involves setting a threshold  $\theta$  to determine whether to continue or discontinue nulling after observations during each signaling slot. The generalized receiver starts off in the nulling mode as before, only now the nulling operation is imperfect. At the end of any signaling slot in which nulling was used, a decision is made to continue nulling if the number of photons detected during the slot exceeds  $\theta$  and to discontinue nulling if the number does not exceed  $\theta$ . As with the ideal conditional nuller, nulling, once discontinued, is never resumed. On the other hand, a downgraded hypothesis, corresponding to a signaling slot in which nulling was used and counts exceeded threshold, could later be revived and chosen as most likely by the receiver if more damaging negative evidence is ultimately obtained against all the other  $(M - 1)$  hypotheses.

The conditional nuller with imperfect nulling just described is not necessarily the best conditional nuller subject to the same nulling inaccuracy. It is simply a straightforward adaptation of the ideal conditional nuller with perfect nulling that was found to be near-optimum for the given detection problem.

The description in this section has detailed only the basic *structure* of the imperfect conditional nuller, the rules by which it prescribes what type of measurement to perform, i.e., to null or not to null prior to detection. It still remains to specify the algorithm by which the receiver optimally combines the outcomes of all its measurements to arrive at a decision. For the ideal conditional nuller, the decision rules and nulling rules went hand-in-hand, and no additional analysis was needed.

**2 The MAP decision rule** For the nonideal conditional nuller, the derivation of the optimum decision rule is still somewhat simplified (for the quantum limited problem) by the hard decisions that take place if photons are ever detected in direct detection mode. The receiver must combine the always inconclusive information obtained in nulling mode with the possibly conclusive evidence obtained in direct detection mode. To minimize error probability, the receiver uses a maximum a posteriori probability (MAP) decision rule.

The MAP rule for the nonideal conditional nuller is derived in the Appendix, under the assumptions of equal a priori probabilities, equal slot energies, constant nulling inaccuracy from slot to slot, and constant (but selectable) threshold level

$\theta$ . A slightly suboptimum version of the rule is used here. It can be stated as follows:

- (1) If no switchover from nulling to direct detection occurs, choose the slot with the fewest counts. (In case of a tie for fewest counts, select randomly from the tied slots.)
- (2) If a switchover occurs after slot  $k$ , and nonzero energy is later detected in slot  $i$ ,  $i > k$ , choose  $H_i$ .
- (3) If a switchover occurs after slot  $k$ , and no further counts are recorded, choose  $H_k$ .

This statement of the decision rule is exactly optimum if the nulling strategy threshold parameter  $\theta$  satisfies

$$\theta \leq \theta_o \equiv \frac{E + E' - E'_o}{\ln(E'/E'_o)} \quad (8)$$

If  $\theta$  exceeds  $\theta_o$ , the true MAP decision rule consists of rules 1 and 2 exactly as stated, together with a refined version of rule 3:

- (3\*) If switchover occurs after slot  $k$ , and no further counts are recorded, choose  $H_k$ , unless  $k < M$  and the number of counts during slot  $k$  exceeds  $\theta_o$ . In the latter case, randomly select from  $H_{k+1}, \dots, H_M$ .

The difference between rules 3 and 3\* is minor, and it does not significantly affect the numerical results. Rule 3 is used in this paper because it allows a much clearer presentation of the performance analysis than rule 3\*.

A tree diagram depicting the behavior of the nonideal conditional nuller is given in Fig 5 for the case  $M = 4$ . The notation conventions are the same as in Fig 3. The branches leaving the direct detection nodes are the same as before, but the branches leaving the nulling nodes are now marked " $> \theta$ " and " $\leq \theta$ " to indicate the nonzero threshold used to determine when to discontinue nulling. The decisions taken at the end nodes are the same as those in Fig 3, with the exception that the top node on the tree is now reachable with nonzero probability and an optimum decision rule at that node is now well-defined.

**3 Performance** The error probability  $\tilde{P}_e$  achieved by the nonideal conditional nuller can now be calculated. Use as a conditioning variable  $k$ , the slot which triggers a switchover from nulling to direct detection, and let  $n_i$ ,  $1 \leq i \leq M$ , denote the number of counts recorded during the  $i$ th slot. With the assumed nulling strategy, slot  $k$  triggers the switchover if and only if  $n_k \leq \theta$  and  $n_i > \theta$  for  $1 \leq i \leq k - 1$ . Note that the possibility that  $k = M$  is included in this definition, even

though in this case the switchover occurs too late to actually be implemented. There is one remaining possibility, that of no switchover at all (i.e.,  $n_i > \theta$  for  $1 \leq i \leq M$ ), and this case is treated separately below.

The error probability  $\tilde{P}_\epsilon$  is expanded in the form

$$1 - \tilde{P}_\epsilon = \gamma + \frac{1}{M} \sum_{j=1}^M \sum_{k=1}^M \Pr(k|H_j) \Pr(C|k, H_j) \quad (9)$$

where

$$\gamma = \frac{1}{M} \sum_{j=1}^M \Pr(\text{correct detection and no switchover}|H_j) \quad (10)$$

and  $\Pr(C|k, H_j)$  is the conditional probability of correct detection, given  $H_j$  and switchover after slot  $k$ . The latter probability is easily calculated from decision rules 2 and 3

$$\Pr(C|k, H_j) = \begin{cases} 1 - e^{-E} & , k < j \\ 1 & , k = j \\ 0 & , k > j \end{cases} \quad (11)$$

The conditional probability for the switchover location, given  $H_j$ , is evaluated as

$$\Pr(k|H_j) = \begin{cases} [1 - P(\theta, E')]^{k-1} P(\theta, E') & , k < j \\ [1 - P(\theta, E')]^{k-1} P(\theta, E'_o) & , k = j \\ [1 - P(\theta, E')]^{k-2} [1 - P(\theta, E'_o)] P(\theta, E') & , k > j \end{cases} \quad (12)$$

where  $P(N, x)$  denotes the cumulative Poisson distribution function with mean  $x$

$$P(N, x) = \sum_{n=0}^N p(n, x) \quad (13)$$

where

$$p(n, x) = \frac{x^n}{n!} e^{-x} \quad (14)$$

After insertion of (11) and (12) into (9) and considerable algebraic manipulation, the expression for  $\tilde{P}_\epsilon$  collapses to

$$\tilde{P}_\epsilon = (1 - P(\theta, E'_o) - e^{-E}) \frac{1 - [1 - P(\theta, E')]^M}{MP(\theta, E')} + e^{-E} - \gamma \quad (15)$$

The final item needed for  $\tilde{P}_\epsilon$  is the evaluation of the term  $\gamma$ . Expand this probability by conditioning on  $n_j$ , then apply decision rule 1. For an exact calculation, a second conditioning variable is also needed. This variable,  $m$ , denotes the number of slot counts  $n_i$ ,  $i \neq j$ , which exactly tie  $n_j$  for lowest count

$$\gamma = \frac{1}{M} \sum_{j=1}^M \sum_{n_j=\theta+1}^{\infty} p(n_j, E'_o) \sum_{m=0}^{M-1} \binom{M-1}{m} \frac{1}{m+1} \\ \times [p(n_j, E')]^m [1 - P(n_j, E')]^{M-1-m} \quad (16)$$

The sum over  $m$  is evaluated in the same manner as the one in (1) and the sum over  $n_j$  is independent of  $j$ . Thus, (16) reduces to

$$\gamma = \frac{1}{M} \sum_{n=\theta+1}^{\infty} \frac{p(n, E'_o)}{p(n, E')} \{ [1 - P(n-1, E')]^M - [1 - P(n, E')]^M \} \quad (17)$$

Note that, for the ideal conditional nuller,  $\theta = 0$ ,  $P(\theta, E'_o) = 1$ ,  $P(\theta, E') = e^{-E}$ , and  $p(n, E'_o) = 0$  for  $n \geq \theta + 1$ . Thus,  $\gamma = 0$  and (15) reduces to (5).

**4 Numerical results for imperfect nulling due to phase error only** The error probability expression (15) is a function of the number of PPM slots  $M$ , the mean slot energy  $E$ , the two nulling mode slot energies  $E'_o$ ,  $E'$ , and the nulling strategy threshold parameter  $\theta$ . In this section, the dimensionality of this problem is reduced from five to three by (numerically) optimizing the nulling strategy threshold and by assuming that the nulling field is subject to phase inaccuracy only. In this case, the nulling mode slot energies are related to the signal slot energy by

$$E'_o = E |1 - e^{j\phi}|^2 = 2E(1 - \cos\phi) \\ E' = E \quad (18)$$

where  $\phi$  is the phase discrepancy (assumed constant) between the nulling field generated at the receiver and the negative of the received ON field.

Error probabilities were numerically evaluated for various values of phase error  $\phi$ , with the nulling strategy threshold  $\theta$  optimized. Some representative curves are shown in Fig 6a, b, c, for  $M = 2, 16, 256$ , respectively. The cusps evident in these curves occur at transitions between discrete values of the optimum threshold.

Superimposed on the phase error performance curves in Fig 6a, b, c are the corresponding curves from Fig 4a, b, c for the direct detection receiver and the ideal conditionally nulling receiver. It is seen from Fig 6a, b, c that small values of phase error are sufficient to cause much of the performance advantage of the ideal conditional nuller to disappear. For example, approximately half the performance advantage (measured logarithmically) of the ideal conditional nuller is erased by phase errors of  $\phi = 4^\circ, 2^\circ, 1/2^\circ$ , for  $M = 2, 16, 256$ , respectively.

It is interesting to note that the error probability curves for the imperfect conditional nuller (with optimized threshold) are upper bounded by the error probability curves for the direct detection receiver. This is a general characteristic of the assumed nulling strategy that also holds for larger amounts of nulling inaccuracy than those assumed in Fig 6a, b, c, even though for large nulling error the information extracted during nulling mode is very small. This property results from the fact that the error probability performance of the direct

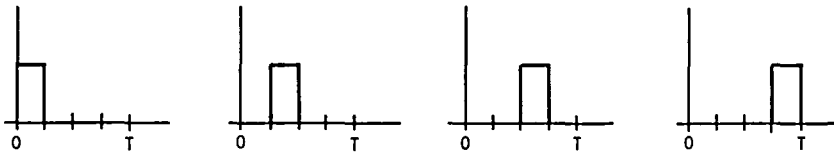
detection receiver for the quantum limited PPM detection problem is unaffected if data from the first slot is unavailable. An imperfect conditional nuller starting in nulling mode can always set its nulling strategy threshold high enough that direct detection mode is essentially guaranteed for the final  $(M - 1)$  slots. Therefore, it can always at least match the performance of the direct detection receiver, with optimized threshold, there will be some improvement. Thus, it always pays, however slightly, for the imperfect conditional nuller to start out in nulling mode.

## VI. Conclusions

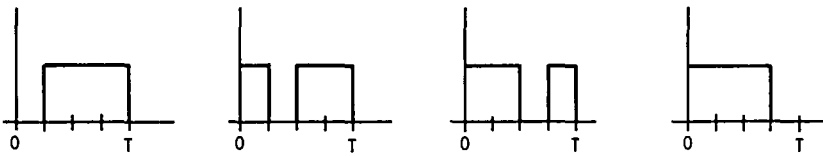
In this article a general class of optical receivers called "conditionally nulling receivers" was postulated for PPM and related modulation schemes to take advantage of the inherent asymmetry in the information obtained by direct detection of ON and OFF signals. An ideal conditional nuller achieves essentially the same error probability as the optimum quantum measurement for quantum noise limited PPM detection. Analysis of the conditionally nulling receiver structure was extended to assess the effects of imperfections in the nulling process. It was found that the near-optimum performance of the conditionally nulling receiver degrades rapidly, albeit gracefully, to the performance of a direct detection receiver in the presence of small nulling field phase errors.

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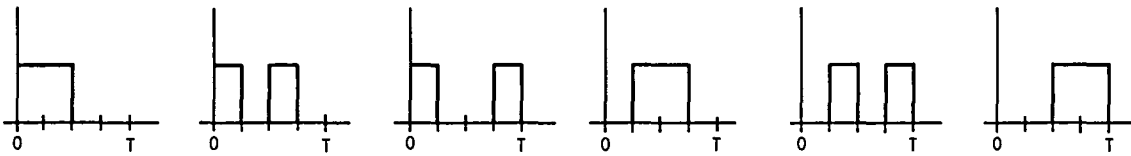
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(a) 4-ARY PPM



(b) 4-ARY 'ROLE-REVERSED' PPM



(c) 6-ARY PULSE DOUBLET CODING ON 4 SLOTS

**Fig 1 Examples of generalized ON – OFF modulation schemes**

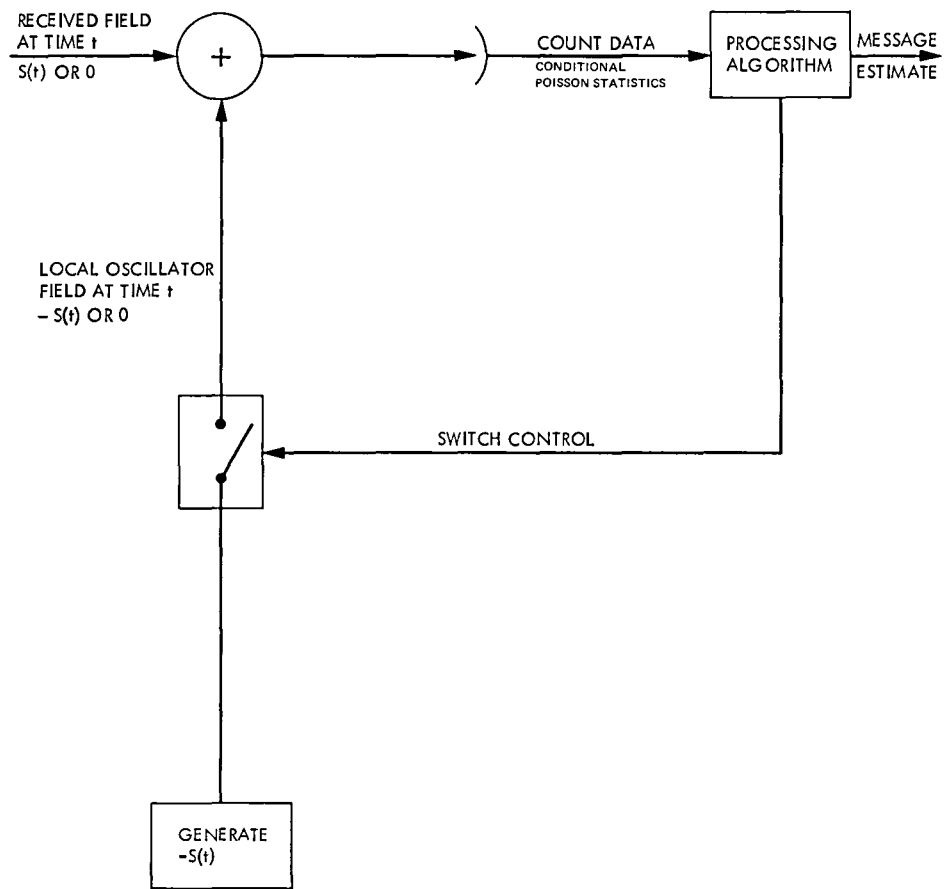
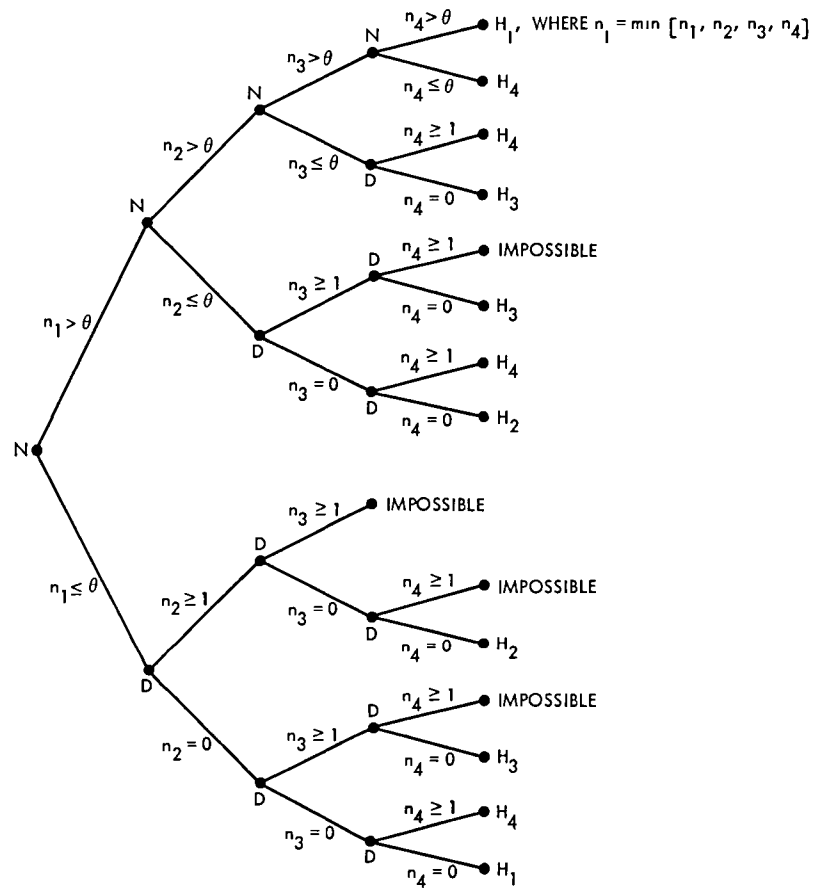


Fig. 2 Conditionally nulling receiver block diagram

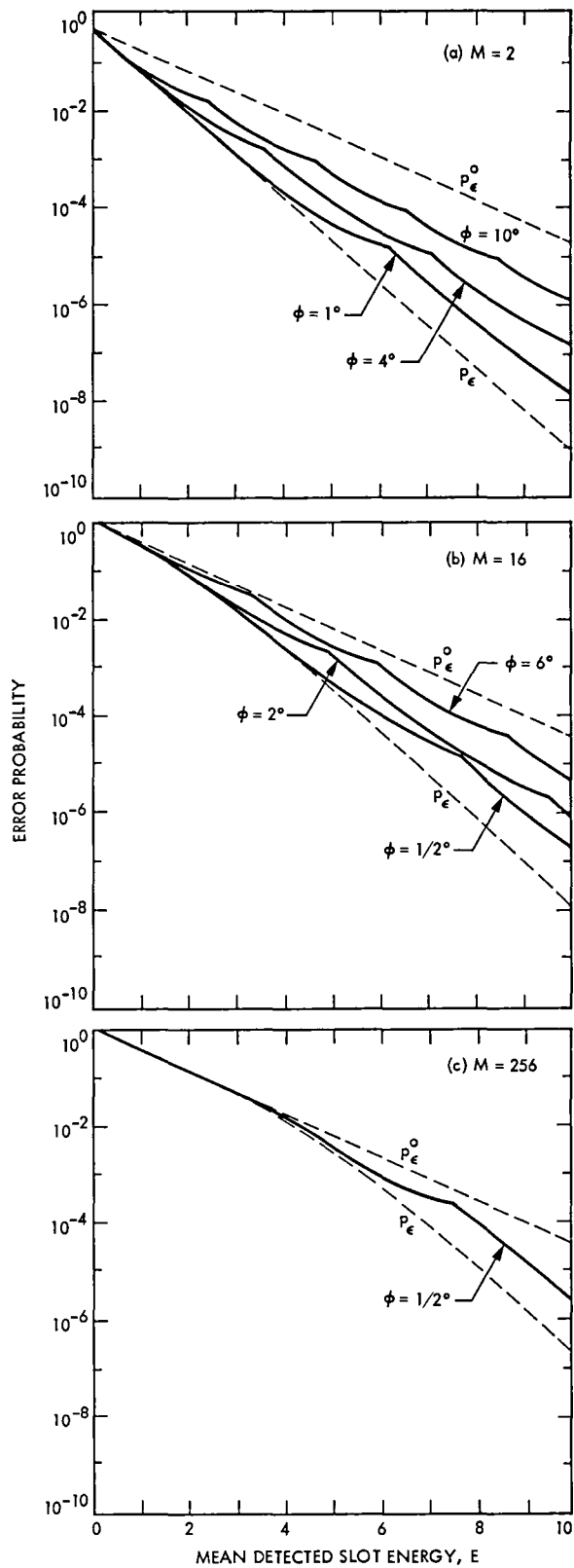




N NULLING MODE FOR NEXT SLOT  
 D DIRECT DETECTION MODE FOR NEXT SLOT  
 $n_i$  NUMBER OF COUNTS RECORDED DURING SLOT  $i$



**Fig 5 Nulling strategy/decision algorithm of conditional nuller with imperfect nulling for quantum limited PPM detection problem ( $M = 4$ )**



**Fig. 6 Deterioration of the performance of the conditionally nulling receiver in the presence of nulling field phase error**

## Appendix

The following three observations lead directly to the optimum decision rule stated in the text

- (1) If photons in any positive number are detected during a signaling slot in which nulling is *not* used, the corresponding hypothesis has a posteriori probability unity and is automatically selected. If no photons are detected during the signaling slots in which nulling is not used, then all of the hypotheses corresponding to these slots have equal a posteriori probabilities.
- (2) Among the hypotheses corresponding to signaling slots in which nulling *is* used, the one with maximum a posteriori probability is the one whose signaling slot yielded the fewest detected photons. In case of ties in count totals, the corresponding a posteriori probabilities are equal. In particular, since the threshold  $\theta$  is constant, this implies that any slot which triggers a switchover from nulling to direct detecting is “king of the hill” relative to all preceding slots.
- (3) If a switchover from nulling to direct detecting occurs after the  $k$ th slot and no photons are recorded in direct detection mode during the remaining  $(M - k)$  slots, then the decision algorithm depends only on the number of detected photons in the  $k$ th slot,  $n_k$ . There is a threshold

$$\theta_o = \frac{E + E' - E'_o}{\ln(E'/E'_o)} \quad (\text{A-1})$$

such that  $H_k$  is selected if  $n_k \leq \theta_o$ , and a random choice among  $H_{k+1}, \dots, H_M$  is made if  $n_k > \theta_o$ . In particular, this implies that  $H_k$  is always chosen in this situation if the switchover threshold  $\theta$  is no greater than the decision threshold  $\theta_o$ .

Justification of the above statements is based on pairwise comparison of the a posteriori probabilities of hypotheses within two separate categories, corresponding to signaling slots in which nulling is or is not used. Note that, for the PPM signal set, a pairwise comparison between the a posteriori probabilities of two hypotheses depends only on data obtained during the two corresponding signaling slots. Thus, the most probable hypothesis within each category can be determined by successive application of binary PPM MAP decision principles. The rule for the direct detection category (statement 1) is the usual rule for direct detection of PPM signals under noise-free conditions. The rule for the nulling category (statement 2) is similar to the rule for PPM signals in background noise, except here the true signal slot is characterized by lowest energy instead of highest energy. The final test to complete the MAP decision is to compare the a posteriori probabilities of the leaders from each category. The likelihood ratio  $\Lambda_{ki}$  between hypotheses  $H_k$  and  $H_i$ ,  $k + 1 \leq i \leq M$ , is based entirely on the data from slots  $k$  and  $i$ . Under the conditions of statement 3, during slot  $k$  nulling is used and  $n_k$  counts are observed, while during slot  $i$  nulling is not used and 0 counts are observed. Thus

$$\Lambda_{ki} = \frac{\Pr(n_k, 0 | H_k)}{\Pr(n_k, 0 | H_i)} = \frac{\frac{(E'_o)^{n_k}}{n_k!} e^{-E'_o} \cdot 1}{\frac{(E')^{n_k}}{n_k!} e^{-E'} \cdot e^{-E}} \quad (\text{A-2})$$

This expression leads to the threshold test

$$\Lambda_{ki} \geq 1 \quad \text{if and only if} \quad n_k \leq \theta_o \quad (\text{A-3})$$

with  $\theta_o$  defined in (A-1)

# Capacity and Cutoff Rate of $(M+1)$ -ary Decision Rules for Noisy $M$ -ary Optical PPM Channel

P. J. Lee

Communications Systems Research Section

*The channel capacity  $C$  and the cutoff rate  $R_0$  of two  $(M+1)$ -ary decision rules for noisy  $M$  slots/symbol optical pulse position modulation (PPM) with ideal photon counting are computed and compared. Also the values of the optimum thresholds needed to minimize the signal energy requirements are given. With a minor increase in hardware complexity, the symbol-by-symbol threshold decision rule is shown to be superior to the slot-by-slot threshold detection-and-decision rule in two aspects. First, it saves more than 0.5 dB in signal energy for the very noisy cases of more than one noise photon per slot (for low noise cases it also saves signal energy, but a negligibly small amount). Second, it is more robust to variations in the noise level.*

## I. Introduction

In this article, the theoretical performance of coded systems for the noisy optical PPM channel with two different  $(M+1)$ -ary decision rules will be studied. The system model is depicted in Fig. 1. From the information bits the encoder selects channel input symbols  $x$  from an  $M$ -ary alphabet  $\{1, \dots, m, \dots, M\}$ . When  $x = m$ , the optical PPM modulator emits an optical pulse only at the  $m$ th time slot during a symbol time period which consists of  $M$  time slots. Depending upon the numbers of the received noise and/or signal photons, the optical detector produces output  $y_{m'}$  at the end of each  $m'$ th time slot,  $m' = 1, 2, \dots, M$ . Ordinarily, hard decision  $M$ -ary symbols would be produced by the PPM demodulator, and passed to the decoder, which uses  $M$ -ary symbols as inputs. However, if we provide, by an  $(M+1)$ -ary decision rule, an additional symbol called "erasure" which denotes the fact that no confident decision can be made, and if we use a decoder which can use  $(M+1)$ -ary symbols as inputs, then

an additional gain can be obtained. Hence,  $(M+1)$ -ary decision (or decision with erasure) rules have been investigated.

An  $(M+1)$ -ary decision rule based on the slot-by-slot threshold detection, which requires very little complexity, has been proposed and used (Ref. 1). In this case, each  $y_{m'}$ ,  $m' = 1, \dots, M$ , is compared to a given threshold  $\gamma$ ,  $\gamma \geq 0$ , and the corresponding slot is declared active ( $z_m = 1$ ) if  $y_m > \gamma$ , or inactive ( $z_m = 0$ ) if  $y_m \leq \gamma$ . Then a nonerased decision is made if and only if there is only one active slot in a PPM symbol time, and the erasure decision is made for all the other cases. This decision rule is shown in Fig. 2(a).

More recently, the optimum and a near optimum  $(M+1)$ -ary decision rules for the  $M$ -ary orthogonal input channel have been developed (Ref. 2). The near optimum  $(M+1)$ -ary decision rule, which uses symbol-by-symbol threshold decision, can be applied to the case of the noisy  $M$ -ary input optical PPM channel. The decision rule produces a nonerased

symbol if and only if the maximum of  $y_m$  is greater than all the other  $y_{m'}$ , by a given threshold  $\delta$ ,  $\delta \geq 0$ . In all other cases, the erasure decision is made. This decision rule is shown in Fig 2(b). We can see that the decision rule of Fig 2(b) is only slightly more complex than that of Fig 2(a).

The same  $(M + 1)$ -ary decision rule with symbol-by-symbol threshold decision was also considered, independently, but without showing the near optimality, by Divsalar, et al (Ref 3) under the different name of "delta-max demodulation". They compared this decision rule to the slot-by-slot threshold detection-and-decision rule by calculating the decoded bit error rate with a Reed Solomon code in each case. Their results, however, apply to the specific set of codes studied and cannot be easily extended to other situations.

Notice that with any  $(M + 1)$ -ary decision rule, the inside of the encoder-decoder pair (inside of the dashed line in Fig 1) becomes an  $M$ -ary input,  $(M + 1)$ -ary output discrete memoryless coding channel. The channel capacity  $C$  and the cutoff rate  $R_0$  of a given coding channel have long been recognized as valuable performance indicators of the coding channel. Also, these quantities are independent of the actual codes to be used over the coding channel. Furthermore, their evaluation is usually much easier than computing the decoded bit error rates with specific complex codes.

Therefore, in this article we will compute  $C$ 's and  $R_0$ 's with more variety of situations (noise levels and/or channel input alphabet sizes) than Ref 3 for more detailed comparisons of the two decision rules. We will show that the performance gain of the symbol-by-symbol threshold decision rule, over the slot-by-slot threshold detection-and-decision rule, is larger than 0.5 dB under strong background noise conditions. As a by-product of the evaluation we will, as in Ref 3, determine the optimum threshold settings. We will also show that under weak background noise conditions (such as in Ref 1), the improvement is negligibly small.

## II. Evaluation of $C$ and $R_0$

Let  $G(a) = \Pr \{y_m \leq a \mid x = m\}$  and  $F(a) = \Pr \{y_{m'} \leq a \mid x = m \text{ and } m' \neq m\}$  represent the cumulative probability distribution functions of the receiver output value for the slot with and without signal pulse, respectively. Let  $P_c(\epsilon)$ ,  $P_s(\epsilon)$ , and  $P_e(\epsilon)$  be the probabilities of correct decision, decision error, and erasure decision, respectively, with a given threshold  $\epsilon$ , which is either  $\gamma$  or  $\delta$  depending on the decision rule. Recall that  $\gamma$  is the threshold for the slot-by-slot detection-and-decision rule and  $\delta$  corresponds to the symbol-by-symbol decision rule. For the slot-by-slot threshold detection-and-decision, these probabilities are given by (Ref 1)

$$P_c(\gamma) = P_{ds}(\gamma) \cdot P_{dn}^{M-1}(\gamma),$$

$$P_s(\gamma) = (M - 1) \cdot (1 - P_{ds}(\gamma)) \cdot (1 - P_{dn}(\gamma)) \cdot P_{dn}^{M-2}(\gamma),$$

and

$$P_e(\gamma) = 1 - P_c(\gamma) - P_s(\gamma),$$

where

$$P_{ds}(\gamma) = 1 - G(\gamma) \quad \text{and} \quad P_{dn}(\gamma) = F(\gamma)$$

For the symbol-by-symbol threshold decision (Refs 2, 3),

$$P_c(\delta) = \int F^{M-1}(t) \cdot dG(t + \delta),$$

$$P_s(\delta) = (M - 1) \cdot \int G(t) \cdot F^{M-2}(t) \cdot dF(t + \delta),$$

if  $y_m$  takes on continuous values, or

$$P_c(\delta) = \sum_k F^{M-1}(k) \cdot g(k + \delta + 1),$$

$$P_s(\delta) = (M - 1) \cdot \sum_k G(k) \cdot F^{M-2}(k) \cdot f(k + \delta + 1),$$

if  $y_m$  takes on integer values (in which case the threshold  $\delta$  should also be an integer), with

$$g(t) = G(t) - G(t - 1) \quad \text{and} \quad f(t) = F(t) - F(t - 1)$$

Finally,

$$P_e(\delta) = 1 - P_c(\delta) - P_s(\delta)$$

The capacity  $C$  and cutoff rate  $R_0$  in [infor bits/channel bit] for the resulting  $M$ -ary input,  $(M + 1)$ -ary output coding channel, which can be described with  $P_s(\epsilon)$ ,  $P_e(\epsilon)$ , and  $P_c(\epsilon)$ , are given by (Ref 2),

$$C = \max_{\epsilon \geq 0} C(\epsilon)$$

and

$$R_0 = \max_{\epsilon \geq 0} R_0(\epsilon)$$

where

$$C(\epsilon) = P_c(\epsilon) + P_s(\epsilon) + (P_c(\epsilon) \cdot \log [P_c(\epsilon)] + P_s(\epsilon) \cdot \log [P_s(\epsilon)] - [P_c(\epsilon) + P_s(\epsilon)] \cdot \log [P_c(\epsilon) + P_s(\epsilon)] - P_s(\epsilon) \cdot \log [M - 1]) / \log [M],$$

and

$$R_0(\epsilon) = 1 - \log [1 + (M - 1) \cdot P_e(\epsilon) + (M - 2) \cdot P_s(\epsilon) + \sqrt{4 \cdot (M - 1) \cdot P_c(\epsilon) \cdot P_s(\epsilon)}] / \log [M],$$

and  $\epsilon$  is either  $\gamma$  or  $\delta$ , depending on the decision rule (Note that the values of the optimum threshold need not be the same for each different criterion)

We assume here an ideal photon counting receiver. That is, we neglect thermal noise and assume a constant gain PMT (photomultiplier tube). In this simple case,  $f(i)$  and  $g(i)$  are given by (Ref 3)

$$f(i) = \exp(-\bar{N}_n) \cdot \bar{N}_n^i / i! \\ g(i) = \exp(-\bar{N}_n - \bar{N}_s) \cdot (\bar{N}_n + \bar{N}_s)^i / i! \quad i = 0, 1, 2, \dots$$

where  $\bar{N}_n$  is the average number of the received background noise photons per PPM slot and  $\bar{N}_s$  is the average number of the received signal photons per PPM symbol

When  $M$  and  $\bar{N}_n$  are given, we can calculate  $C$ 's and  $R_0$ 's as functions of a single parameter of  $\bar{N}_s$ . Since  $C$  and  $R_0$  are monotonically increasing functions of  $\bar{N}_s$ , they are invertible. The coding theorem (Ref 4) says that there exists a code with code rate  $r$  [info bits/channel bit] which gives arbitrary small decoded error rate provided that

$$r < C(\bar{N}_s)$$

Or equivalently, since  $C$  is invertible, the condition can be rewritten as

$$\bar{N}_s > C^{-1}(r)$$

Since the parameter of interest is usually the average number of received signal photons per information bit  $\bar{N}_b$  [photons/info bit], we can obtain it from  $\bar{N}_s$  by the simple relation  $\bar{N}_b = \bar{N}_s / (r \cdot \log_2 M)$ . Therefore, the condition can again be rewritten as

$$\bar{N}_b > C^{-1}(r) / (r \cdot \log_2 M)$$

Hence, we call  $C^{-1}(r) / (r \cdot \log_2 M)$  the "required average number of signal photons per information bit, or simply photons per information bit, to achieve capacity  $C$  with a given code rate  $r$ ". Similarly,  $R_0^{-1}(r) / (r \cdot \log_2 M)$  is called the "required average number of photons per info bit to achieve a cutoff rate  $R_0$  with a given code rate  $r$ ".

For  $M = 64, 256,$  and  $1024$  (or  $\log_2 M = 6, 8,$  and  $10$ ), the required  $N_b$  [signal photons/info bit] to achieve (a) the channel capacity  $C$  and (b) the cutoff rate  $R_0$  vs code rate  $r$  plots are given in Figs 3, 4, and 5 for various noise levels of  $N_n$  [noise photons/slot] for the two different  $(M + 1)$ -ary decision rules. The curve marked  $N_n = 0$  corresponds to the noiseless channel use.

### III. Discussions and Conclusions

First, we notice the nonsmoothness of the curves in the plots. This is due to the discreteness of the observation values and threshold values. Also the discreteness of the observation and threshold values gives large gaps in performance between the two decision rules near the nonsmooth points. This phenomenon is more noticeable for the low noise level cases, due to the small threshold values (Note that there is a much bigger difference when using threshold values of 0 or 1 than the difference corresponding to thresholds of 22 or 23). If, instead of using an ideal photon counting receiver assumption, we had considered a more realistic random gain PMT with additive thermal noise model (Ref 1), then the nonsmoothness and the large gaps for some rates would disappear and hence the differences in performance between the two decision strategies would be almost the same for all values of  $r$ . (We can see this for very noisy cases where the curves tend to be much smoother, as well as for  $\bar{N}_n = 10^{-2}$  of Fig 4(a) where accidentally the optimum threshold levels do not vary for almost all values of  $r$ .)

As was concluded in Ref 3, the symbol-by-symbol threshold decision rule saves more signal energy than the slot-by-slot threshold detection-and-decision rule as the noise level increases. For the noiseless case there is no gain at all. For the 2.5 bit/photon demonstration program (Ref 1), 256-ary PPM with rate 3/4 or 7/8 coding is used. The typical value of background noise is  $10^{-5}$  [noise photons/slot]. In this case also, we calculated  $C$ 's and  $R_0$ 's. The gain in signal energy of using the symbol-by-symbol threshold decision over using the slot-by-slot threshold detection-and-decision is less than 0.03 dB for almost all code rate, hence totally negligible. With this noise level, the performance is about 0.2 dB poorer than the noiseless case. However, for the very noisy case of more than

one noise photon per slot, the gain is more than 0.5 dB for all values of  $M$  considered. We also observed as in Ref. 3 that the symbol-by-symbol threshold decision rule is more robust to variations in the noise level.

In conclusion, we have found that the symbol-by-symbol decision strategy always requires less signal energy than the

simpler slot-by-slot detection-and-decision strategy, with greater benefits both in energy savings and in robustness to noise level variations for the high background noise cases. In the weak background noise case, the differences in performance are negligibly small, and hence the symbol-by-symbol decision will not be attractive since it requires additional complexity.

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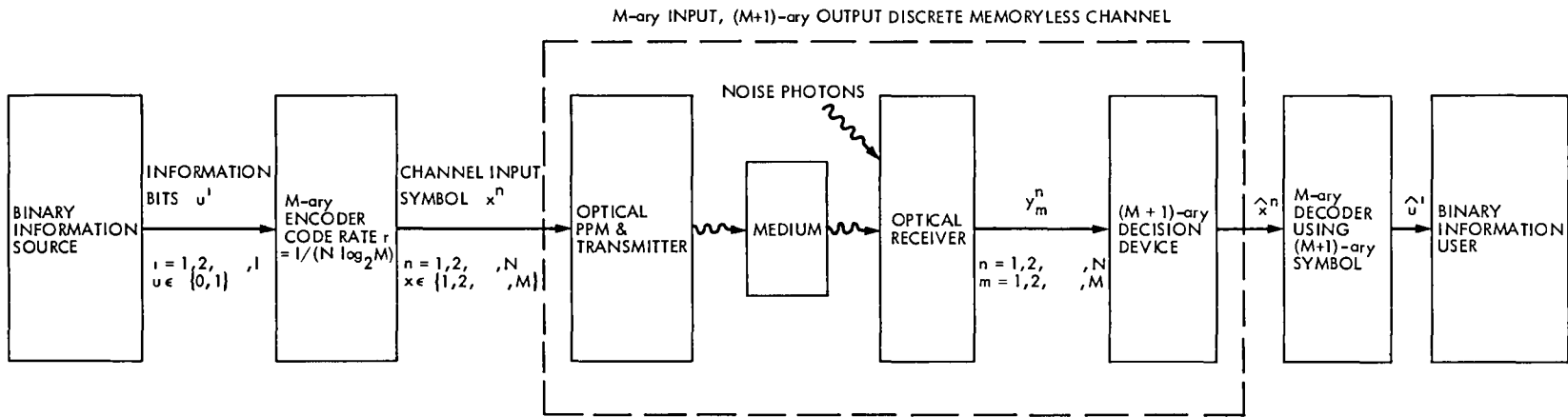


Fig. 1. Coding system for  $M$ -ary optical PPM channel with  $(M+1)$ -ary decision rule

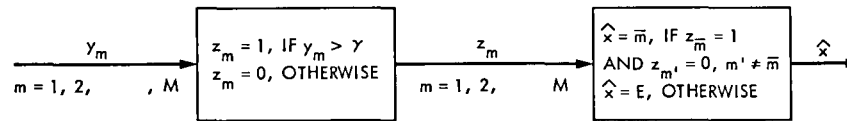


Fig. 2(a). Slot-by-slot threshold detection-and-decision device

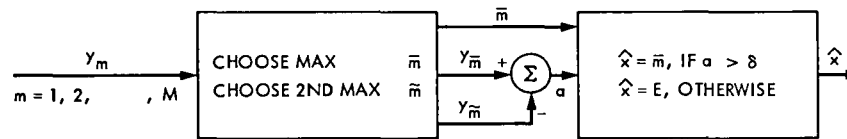


Fig. 2(b). Symbol-by-symbol threshold decision device

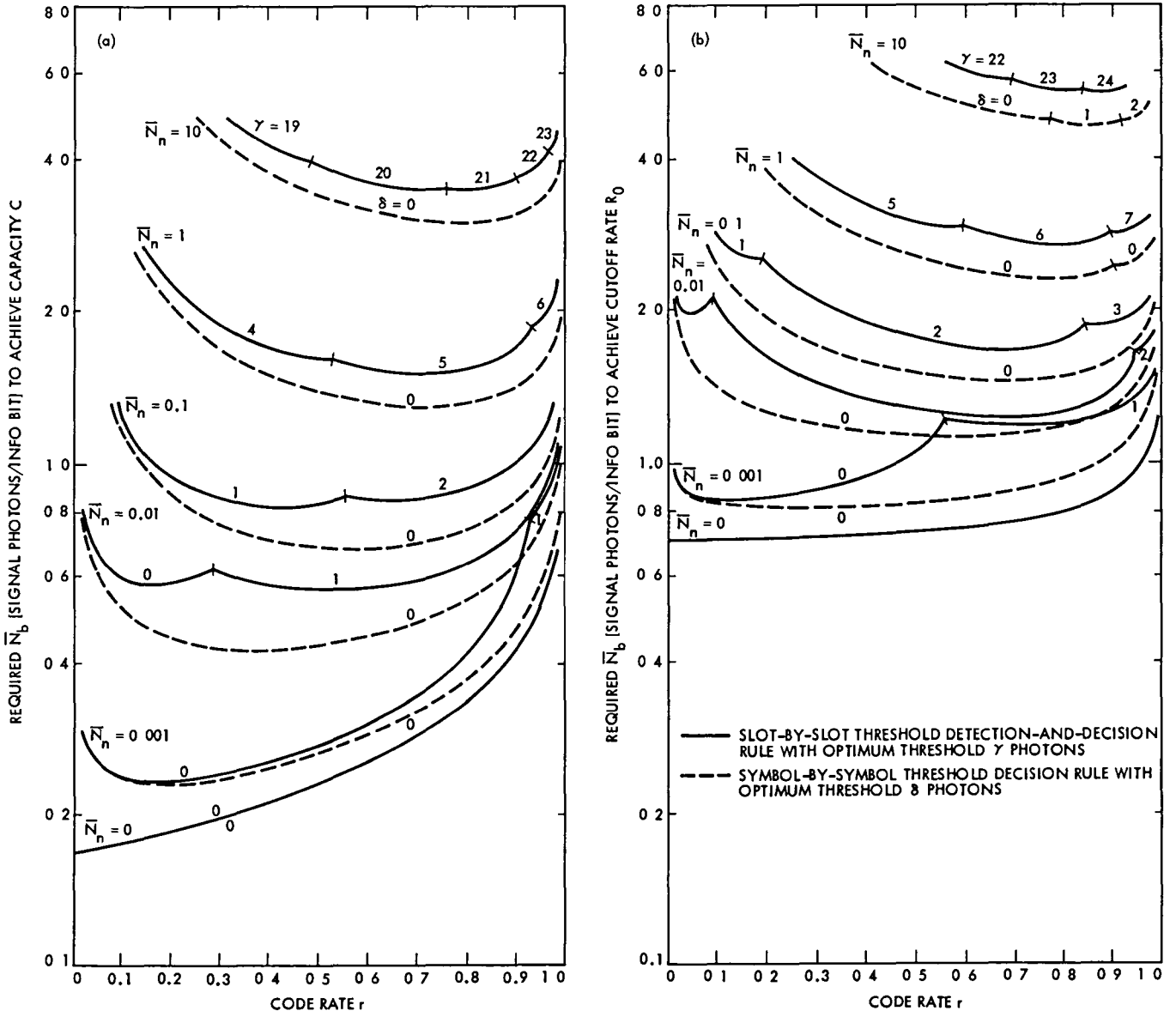


Fig. 3. Required  $N_b$  [signal photons/info bit] to achieve (a) channel capacity  $C$  and (b) cutoff rate  $R_0$  vs code rate  $r$  for two different  $(M+1)$ -ary decision rules on ideally direct-detected  $M = 64$ -ary optical PPM channel with several background noise levels  $N_n$  [noise photons/PPM slot]

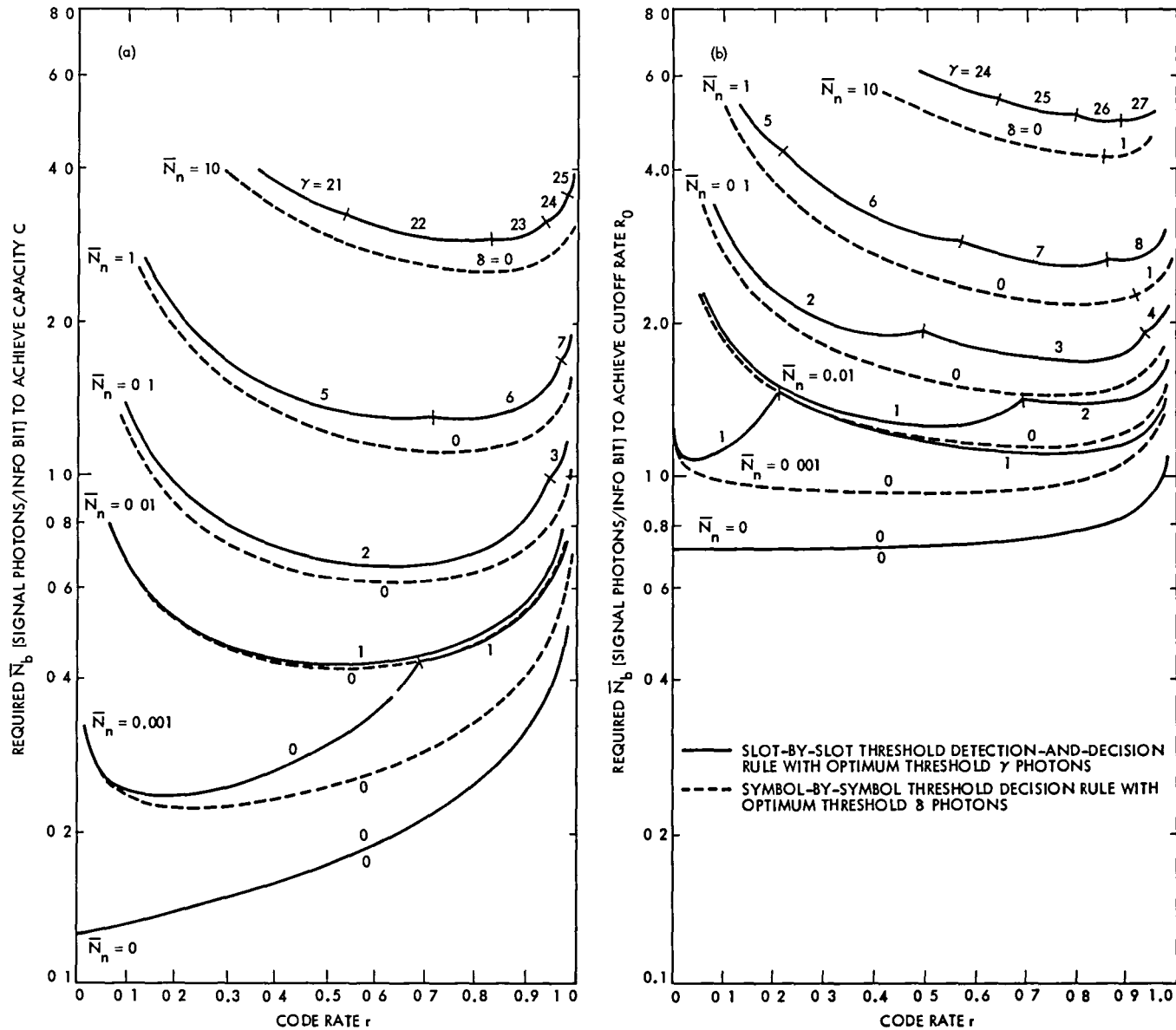


Fig. 4. Required  $\bar{N}_b$  [signal photons/info. bit] to achieve (a) channel capacity  $C$  and (b) cutoff rate  $R_0$  vs code rate  $r$  for two different  $(M+1)$ -ary decision rules on ideally direct-detected  $M = 256$ -ary optical PPM channel with several background noise levels  $N_n$  [noise photons/PPM slot]

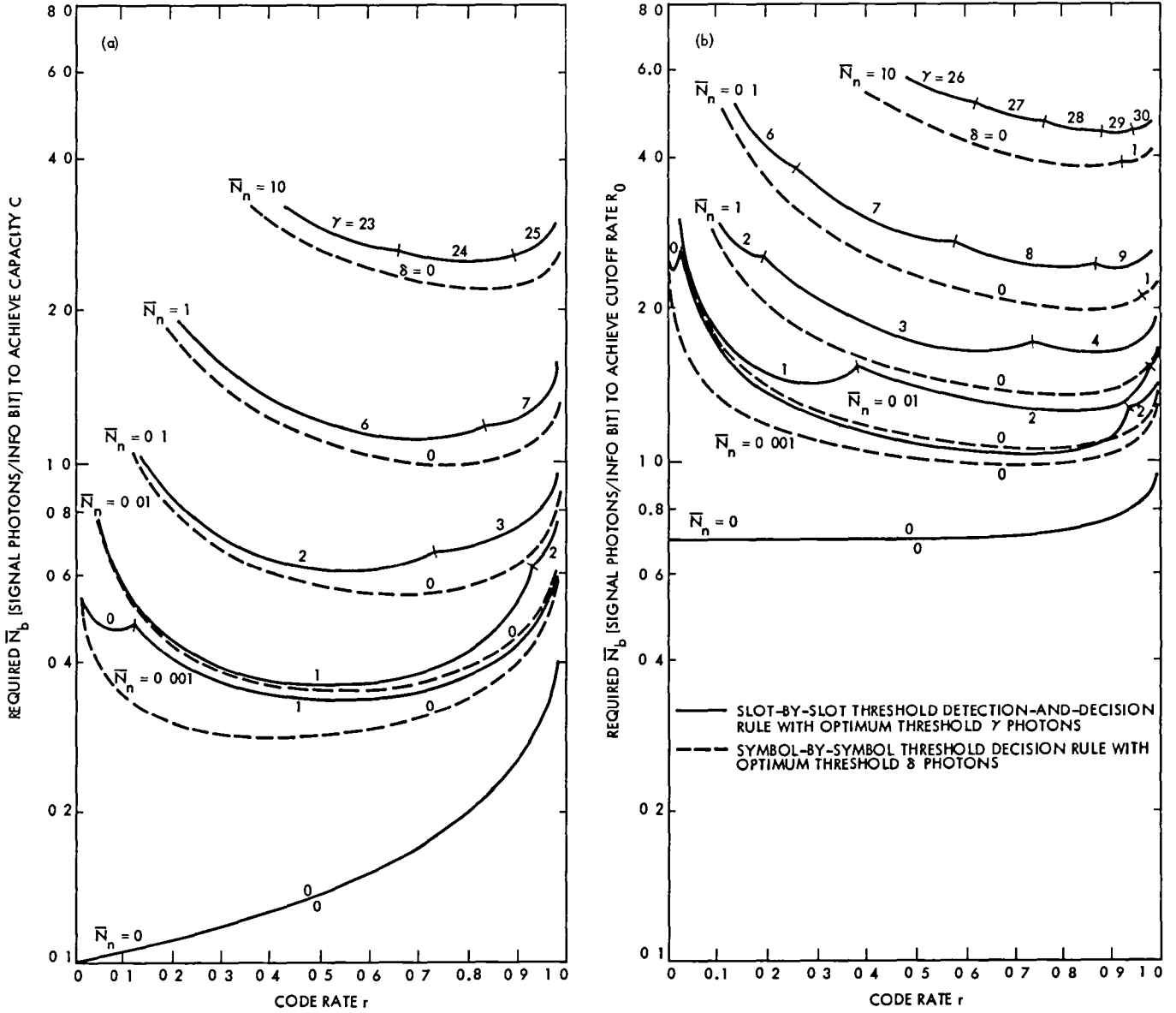


Fig 5. Required  $\bar{N}_b$  [signal photons/info bit] to achieve (a) channel capacity  $C$  and (b) cutoff rate  $R_0$  vs code rate  $r$  for two different  $(M+1)$ -ary decision rules on ideally direct-detected  $M = 1024$ -ary optical PPM channel with several background noise levels  $\bar{N}_n$  [noise photons/PPM slot]

# An Update on the Use of the VLA for Telemetry Reception

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*This article modifies a previous analysis to incorporate the actual structure of the command signal system of the Very Large Array (VLA). In particular, in addition to the 1-ms command signal of the previous article, there is a "data invalid" signal that is generated. The command signals are transmitted to the antennas during the period in which the "data invalid" signal is on. This means that the gaps in the received data are really 1.6 ms long rather than 1 ms long. Simulation results with this taken into account show that the VLA will not support (7, 1/2) convolutionally encoded telemetry at acceptable error rates at any of the Voyager telemetry data rates. VLA will support Voyager encounters provided that either concatenated coding is implemented, VLA is arrayed with another receiving site (such as Goldstone), or VLA is reconfigured so that the gaps are rotated.*

## I. Introduction

The performance of the Very Large Array (VLA) as a receiver for Voyager 2 coded telemetry data was recently determined (Ref 1). Since that time, however, more information concerning the operation of the VLA has come to light. This additional knowledge does affect the previous results. In fact, new simulations and modeling show that the VLA will not be capable of supporting convolutional coded Voyager data at any of the data rates that are being considered for Uranus or Neptune encounters.

The VLA is a phased array consisting of 27 independently pointable parabolic dishes in the New Mexico desert. Each dish measures 25 meters in diameter. The array is configured in three linear arrays of nine antennas each that radiate from a common centerpoint. The antennas may be moved along these lines by means of railroad-type tracks. Communication between the antennas is by means of a connecting network of waveguides. The total aperture of the VLA is about twice that of a DSN array consisting of a 64-m antenna and

three 34-m antennas. The VLA was designed for use as a radio astronomy observatory. Some additional equipment will be needed in order for the VLA to receive the X-band Voyager telemetry transmissions.

The main problem with using the VLA as a telemetry receiver is that signals are not transmitted continuously from the antennas to the processing facility. Instead, the VLA operates in a duty cycle mode. Every so often, at regular intervals, communications from the antennas are suspended momentarily so that calibration signals may be sent to the antennas. In Ref 1 it was stated that the VLA collects data for 51 ms, followed by a 1-ms calibration period. This turns out to be only partially true (Ref 2). There is indeed a 1-ms calibration signal. However, data is only collected for 50.4 ms of the 52-ms cycle. The relevant signals are illustrated in Fig 1. The 1-ms calibration signal is transmitted during the time that the "data invalid" signal is on. This means that for the purposes of this study, the VLA must be treated as a system that collects data for 50.4 ms followed by a 1.6-ms gap.

If uncoded telemetry data were received by the VLA then, for data rates over a few thousand bits per second, the best possible error rate would be  $(1/2) (1/52) = 1.54 \times 10^{-2}$ . Coding must be used in order to achieve bit error rates below the  $5 \times 10^{-3}$  level required for good image quality. In Ref. 1 it was stated that VLA could support Voyager 2 convolutionally encoded transmissions at a data rate of 10.8 kbps at both Uranus and Neptune encounters. In light of the new information mentioned above, this is no longer the case. In order to achieve good quality imaging data from either encounter at the VLA, additional coding must be used or the VLA must be reconfigured to at least partially resolve the gap problem.

## II. Best Possible Performance of VLA with Convolutional Coding Only

In this section, the performance of the VLA for the reception of convolutionally encoded data is derived in the case of infinite bit SNR. This represents the best possible performance of the system. The formula that was derived in Ref. 1 is still valid

$$P_b = 0.5 (R_b T_2 - k + 1) / [R_b (T_1 + T_2)]$$

As before,  $k$  is the constraint length of the convolutional code,  $R_b$  is the data rate in bits per second,  $T_1$  is the time in seconds spent in that portion of the cycle representing a perfect channel, and  $T_2$  is the time in seconds spent in each gap. It is assumed that  $R_b$  is large enough so that the gaps are at least  $k$  bits long. In the case of interest in this report,  $k = 7$ ,  $T_1 = 50.4$  ms, and  $T_2 = 1.6$  ms. A graph of  $P_b$  as a function of  $R_b$  is shown in Fig. 2. This represents the best possible performance of the VLA with  $(7, 1/2)$  convolutional coding. It is evident from the figure that in this scenario, the VLA will not support any of the planned Voyager data rates at either encounter.

## III. Software Simulations of VLA Performance

The performance of the VLA under various scenarios was determined by using the simulation software that was described in Ref. 1. Five scenarios were simulated. For each scenario (except the second) two data rates were considered: 10.8 and 19.2 kbps. These data rates are representative of those that can be expected at Voyager's future planetary encounters. In addition, the performance of VLA was determined both for  $(7, 1/2)$  convolutionally encoded data and for  $(7, 1/2)$  coded data concatenated with a  $(255, 223)$  Reed-Solomon code. This latter coding scheme, which will be referred to as *concatenated coding*, is an option for Voyager. A detailed description of concatenated codes may be

found in Ref. 3. The scenarios that were considered are as follows:

- (1) *Ideal performance* In this scenario, it was assumed that no gaps were present in the data. This represents a best possible case and it can be used to measure the magnitude of losses that occur in the other scenarios.
- (2) *Normal VLA* This scenario represents the VLA as described in the introduction. The gaps are assumed to be periods of time in which the channel  $E_b/N_0$  is arbitrarily small. Since it is already known from the remarks in Section II that this scenario will not suffice for Voyager encounters, only the 10.8-kbps data rate was simulated. The results give an indication of just how far off the mark the performance really is.
- (3) *VLA arrayed with an equal aperture* The gaps in this scenario are 3.0 dB lower in  $E_b/N_0$  than the rest of the data. This scenario can be interpreted as arraying half of the VLA with the Goldstone DSN complex. The combining is assumed to be perfect combined carrier referencing (CCR). Two different methods of achieving nearly ideal CCR performance are described in Refs. 4 and 5.
- (4) *Equally spaced rotating gaps* This is the first of two scenarios that involve a reconfiguring of the VLA. In this scenario, each of the three arms of the VLA receives its 1.6-ms gaps at a different time. These times are equally spaced so that the entire array has a duty cycle of 52/3 ms with 1.6-ms periods of 1.7-dB attenuation in the received data.
- (5) *Clumped rotating gaps* In this scenario, as in 4, each arm receives its 1.6 ms of control signals at a different time. In this case, however, these individual arm gaps are consecutive in time. This means that the VLA would still have a duty cycle of 52 ms but there would be a  $3 \times 1.6 = 4.8$  ms part of each cycle that is attenuated by 1.7 dB. It turns out that although one could expect better performance from scenario 4, this scenario is easier to implement in the present VLA hardware.

For the purposes of displaying the results of the simulations, the expression  $E_b/N_0$  represents the bit SNR during the good part of the VLA cycle at all times in the figures.

The results of the simulations for the convolutional-only case are shown in Figs. 3-7. The performance of the VLA in scenarios 2, 3, and 4 is slightly worse than that reported in Ref. 1 due to the longer gap lengths. Also, one can clearly see the slight advantage that scenario 4 has over scenario 5. This is because the Viterbi decoder can handle several short gaps better than a single long one.

The performance of the VLA with concatenated coding was calculated from these simulations using the formulas described in Ref. 1. These results are shown in Figs. 8-12. In the case of concatenated coding, a bit error rate of  $10^{-5}$  is considered necessary for good image quality.

#### IV. VLA Throughput at Voyager Encounters

The most reasonable measure of performance for a telemetry receiving system such as the VLA is the amount of data throughput that it is capable of handling. The particular measure of throughput used in this study is the number of "good bits per day." During each day of an encounter, Voyager will be "visible" from the VLA for some number of hours. Depending on the data rate and coding that are implemented, only a fraction of this viewing time may support reception at the required bit error rate or better. This is due to the fact that received power is a function of the elevation angle of the antennas.

The raw data in this section comes from design control tables for Voyager 2 Uranus encounter. The curve shown in Fig. 13 represents the total power-to-noise-spectral-density ratio ( $P_T/N_0$ ) that is expected to be incident at the Goldstone complex on day 34 of 1986 (Uranus encounter). This curve includes a 90% weather confidence level and the effects of antenna elevation angle. Since the VLA is at approximately the same latitude as Goldstone, the values of  $P_T/N_0$  for the

VLA at Uranus encounter were derived from these by simply scaling them up by the ratio of the apertures involved. For Neptune encounter, 3.5 dB was subtracted to account for the additional space loss. It was assumed that the maximum viewing time is 8.3 hours.

The throughputs for the various scenarios and data rates were calculated according to the formulas that were derived in Ref. 1 and the data from the simulations of Section III. The results are shown in Fig. 14. A comparison with the corresponding figures in Ref. 1 shows the extra loss from the 0.6-ms additional gap length.

#### V. Conclusions

The main point of this study is that the VLA cannot be used as a telemetry receiver for Voyager 2 Uranus and Neptune encounters in its present configuration with only convolutional coding. There are three workable solutions to this problem. The first and easiest is to use concatenated coding. Voyager 2 does have this capability on board and there is already Reed-Solomon decoding hardware at JPL. The second is to array the VLA with another receiver that does not have this gapped behavior. The prime candidate for such an array would be the Goldstone complex due to its relatively close proximity. The third solution is to reconfigure the VLA so that the gaps occur at different times in the different apertures. Two such methods were described in Section III.

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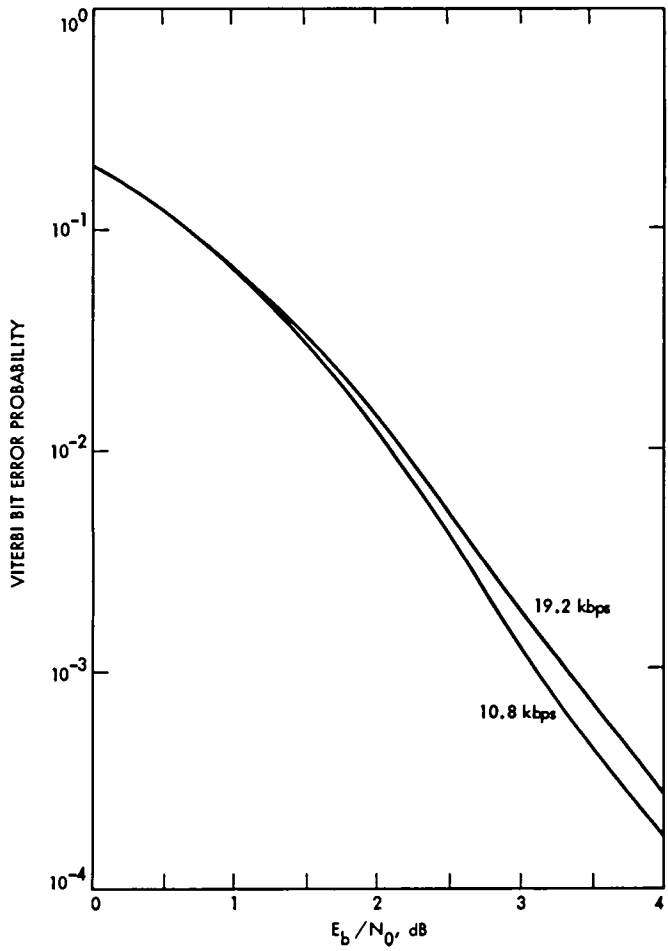


Fig 6. Simulated performance of VLA with equally spaced rotated gaps (Scenario 4) (7, 1/2) convolutional coding only

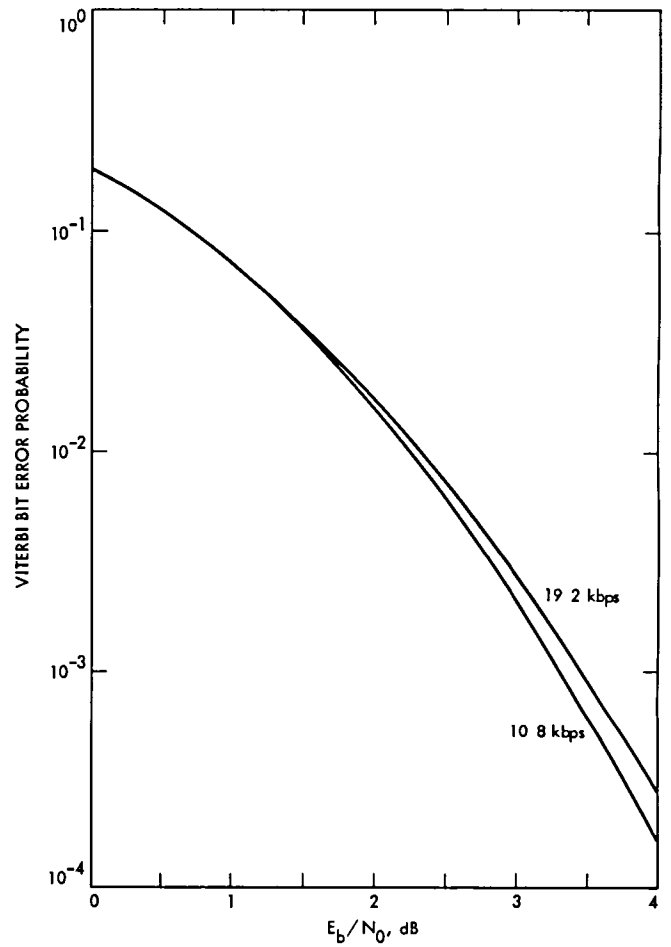


Fig 7. Simulated performance of VLA with clumped rotated gaps (Scenario 5) (7, 1/2) convolutional coding only



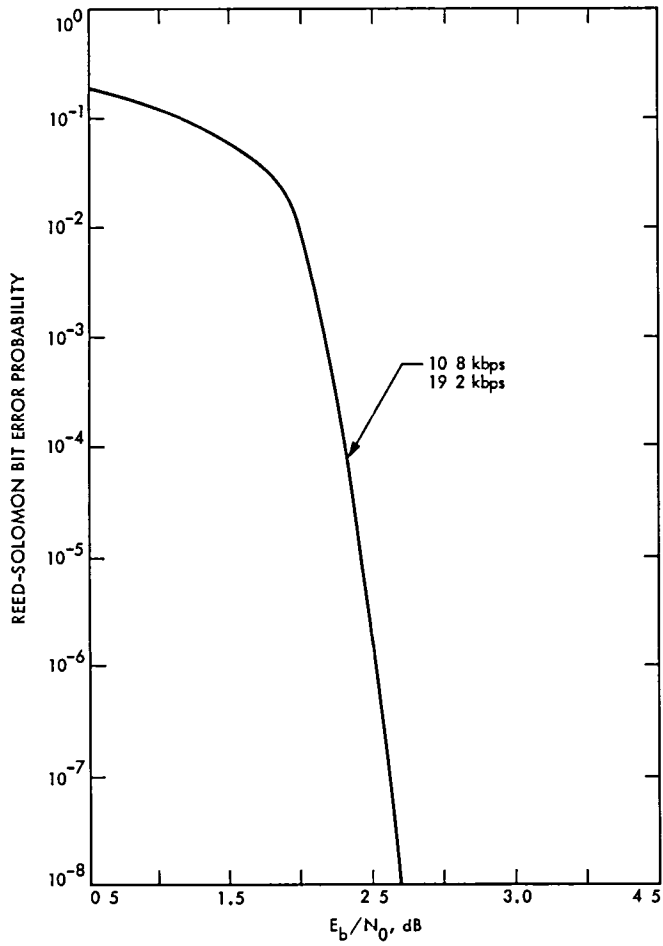


Fig 8 Simulated performance of an ungapped receiving system (Scenario 1) Concatenated coding

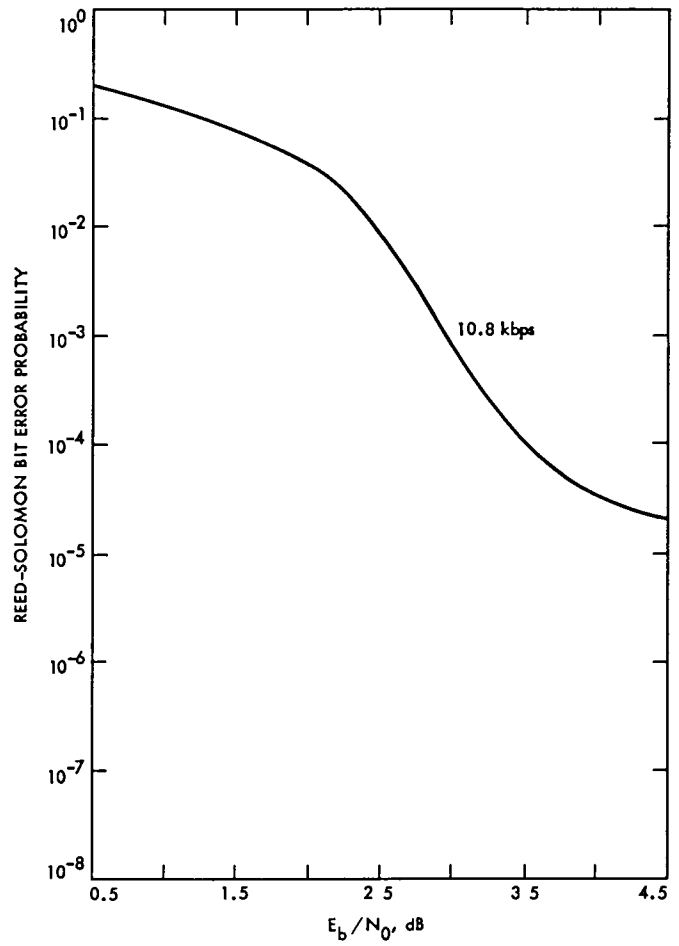


Fig. 9. Simulated performance of VLA with 1.6-ms gaps with no signal (Scenario 2) Concatenated coding

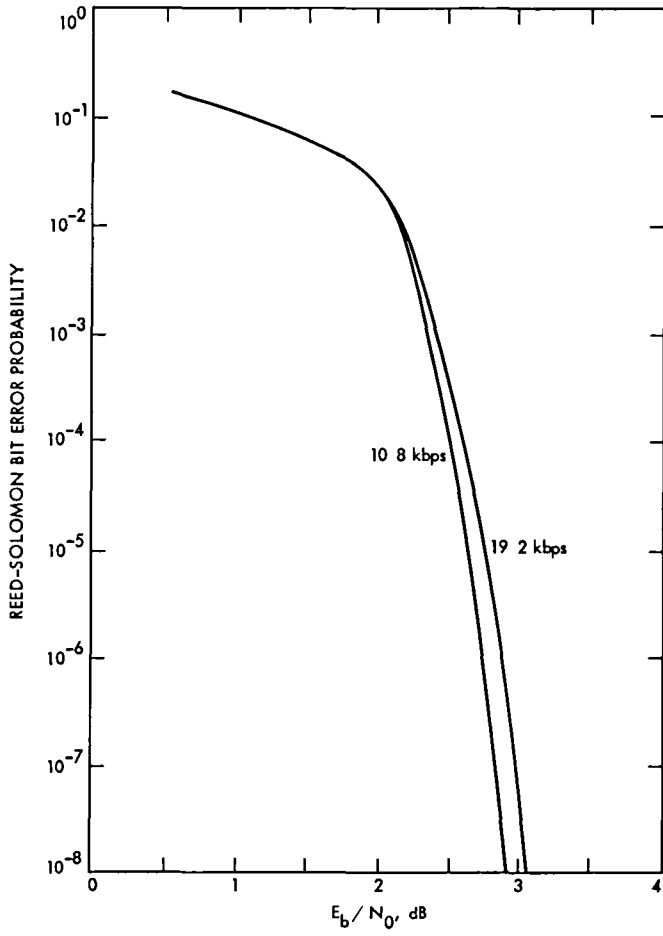


Fig. 10. Simulated performance of VLA with 1.6-ms gaps attenuated by 3 dB (Scenario 3) Concatenated coding

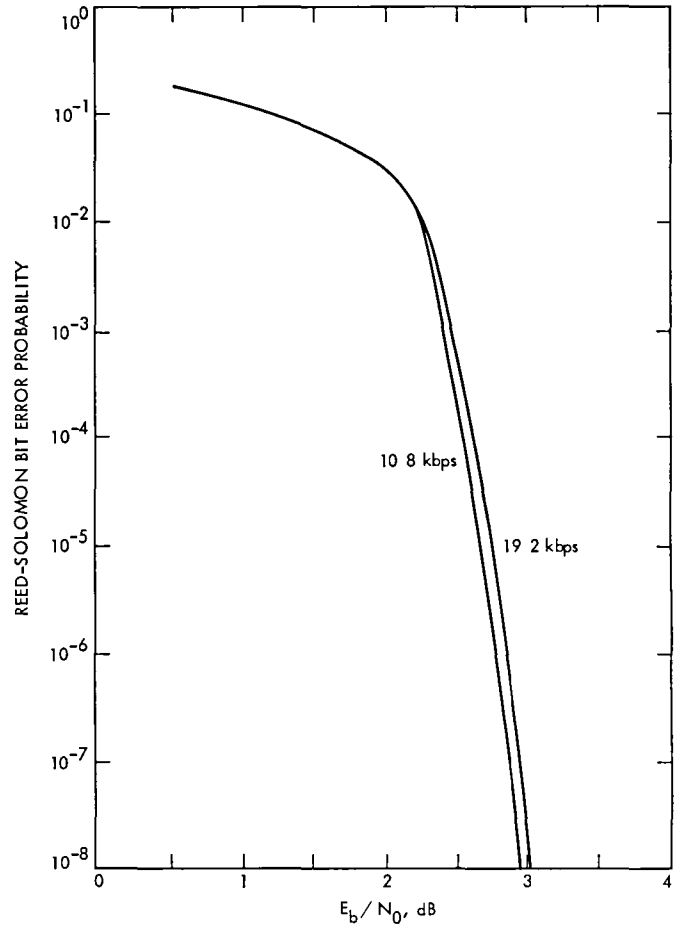
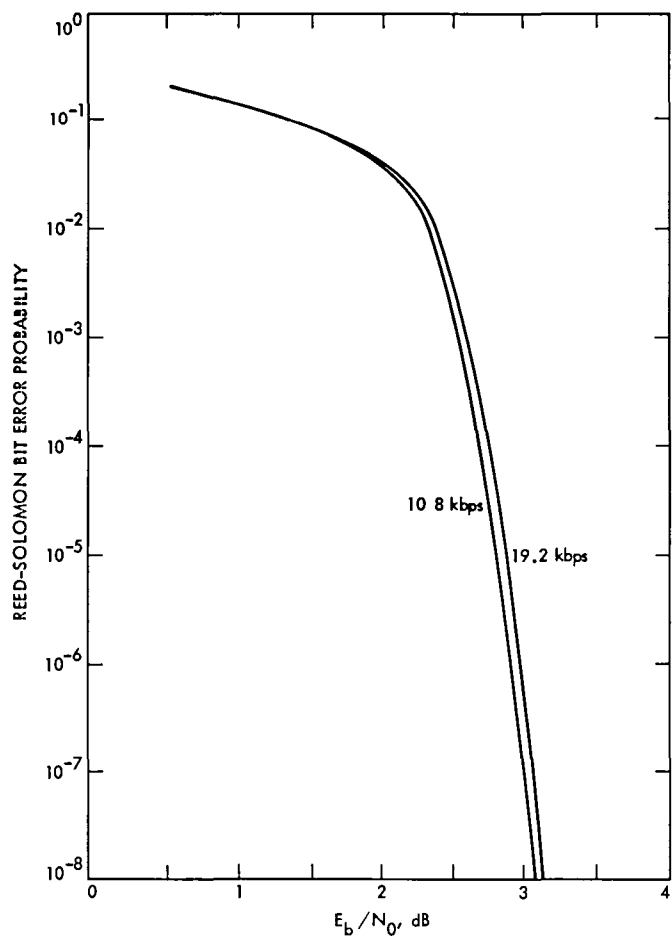
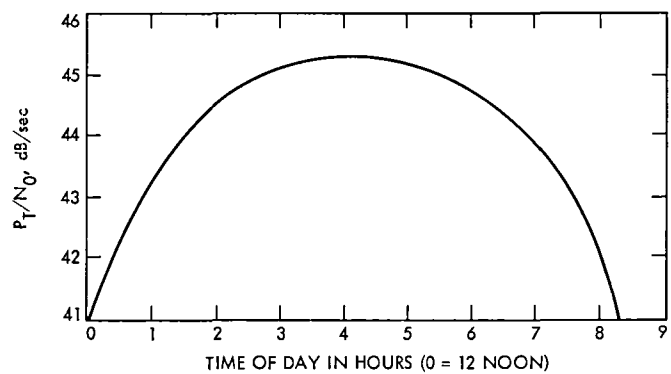


Fig. 11 Simulated performance of VLA with equally spaced rotated gaps (Scenario 4) Concatenated coding



**Fig. 12. Simulated performance of VLA with clumped rotated gaps (Scenario 5). Concatenated coding**



**Fig. 13. Baseline performance of Goldstone array at Voyager 2 Uranus Encounter, array = 64 m + 34 m + 34 m, day 34, 1986, 90% weather**

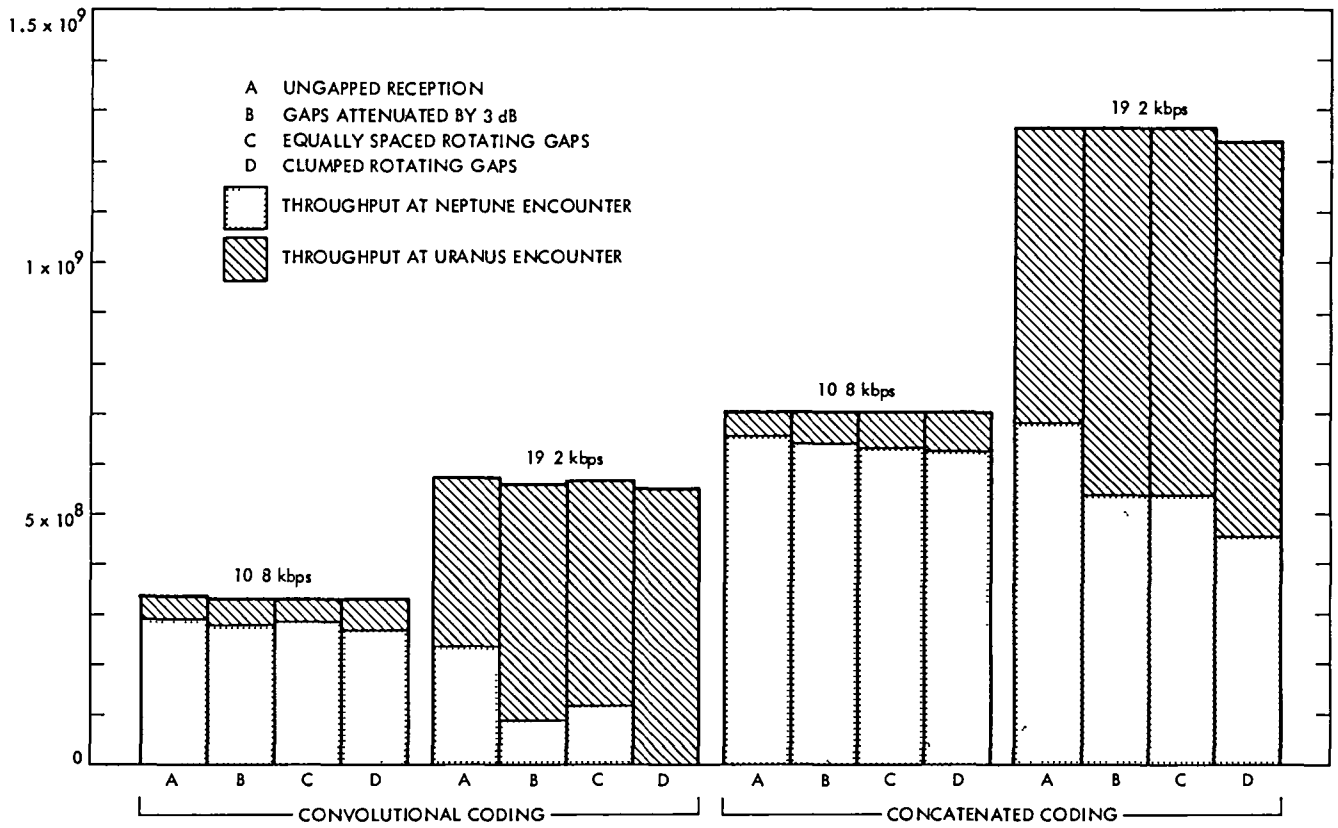


Fig. 14. VLA throughput at Voyager 2 Uranus and Neptune encounters

# A VLSI Implementation of a Multicode Convolutional Encoder

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*This article describes a VLSI architecture and layout for a convolutional encoder. This architecture allows a single chip implementation of an encoder that is capable of handling many different convolutional codes including all the convolutional codes that are presently used by NASA for deep space missions.*

## I. Introduction

Most present and all planned deep space missions make use of convolutional coding. One reason for this is that the encoder for these codes is very simple and can be made relatively small and light even with off-the-shelf components. A block diagram of a convolutional encoder is shown in Fig. 1. The convolutional code implemented in the figure is the NASA standard  $(7, 1/2)$  code. The input information bits  $i_1 i_2 i_3$  are shifted into a register. This register is used to generate two parity checks each time a new information bit is entered. A commutator interleaves these parity checks to produce the coded output stream  $a_1 b_1 a_2 b_2 a_3 b_3$ .

In a general convolutional decoder, the shift register is  $K$  bits long. The number  $K$  is called the "constraint length" of the code. In the most general situations, there can be more than one  $K$ -bit shift register in the encoder. For the purposes of this article, however, it will be assumed that there is only a single register. If there are  $P$  parity checks that are generated by the encoder, then the resulting code rate will be  $1/P$ . The code would then be referred to as a  $(K, 1/P)$  convolutional code. This is not enough to define the code uniquely. The particular code is determined by the parity checks.

Consider a single parity check that is generated in the encoder. Let the positions in the shift register be labeled  $1, 2, 3, \dots$  with the position 1 being the one that the data bits enter first. It follows that this parity check can be described by a  $K$ -bit binary number where a one in position  $i$  indicates the presence of a parity check tap after state  $i$  of the shift register. The parity checks themselves can be ordered so that "parity check #1" is the first to be scanned by the commutator and the "parity check # $P$ " is the last. Using these conventions, any  $(K, 1/P)$  convolutional code can be represented by an ordered set of  $P$   $K$ -bit binary numbers. For the  $(7, 1/2)$  code described above, the numbers are

1011011

1111001

If one reads the columns of this binary matrix as binary numbers, this can be reduced to simply 3233013. This sequence of integers, together with the more common  $(7, 1/2)$  notation, completely describes the convolutional code.

Sometimes one or more of the parity checks are inverted before the multiplexing operation of the commutator. This

is done so that additional channel symbol transitions are introduced in the case that the information bits consist of long strings of either all zeroes or all ones. These transitions are needed in order to synchronize the symbols in the receiver before they can be sent to a decoder. These inversions have no effect on the performance of the code itself but their presence in any usable encoder is essential.

The performance of convolutional codes has been studied in Refs 1-3. The problem of node synchronization (determining what the order of the parity checks is in the decoder) was treated in Refs 3-5.

## II. A Functional Description of the Chip

The chip, as implemented, is capable of supporting three codes. These are the standard NASA (7, 1/2) code 3233013, the standard NASA (7, 1/3) code 7376147, and a longer constraint length (10, 1/3) code 7461776427. For each of these codes, the capability of inverting any or all of the parity checks has been included in the design.

There are 22 pins that will be connected to the chip. These are

### INPUTS

Vdd	The voltage connection to the chip — nominally -5 Vdc
Gnd	The ground connection to the chip
Sub	A connection to allow biasing of the chip substrate
1/3	Selects the code rate. A “one” signifies that a rate 1/3 code has been selected, while a “zero” means the rate is 1/2.
SEVEN	Selects the code constraint length. A “one” indicates a constraint length seven code while a “zero” indicates a constraint length of ten.
NRM1	This signal determines whether or not the first parity check is inverted. A “one” indicates the noninverted state.
NRM2	The same as NRM1 for the second parity check.
NRM3	The same as NRM1 for the third parity check. This signal has no effect if a rate 1/2 code has been selected.

BITIN	This is the data input
PHI1IN	The input for the first phase of a two-phase clock
PHI2IN	The input for the second phase of a two-phase clock
CLOCKIN	The input for a one-phase clock
CLKSEL	Selects between a one- and two-phase clock. A “one” indicates that a one-phase external clock has been selected.

### OUTPUTS

OUTPUT	The output for the encoded channel symbols
PHI1OUT	Output of first phase of the internal two-phase clock. If a two-phase external clock has been selected (CLKSEL=0), then this is the same as PHI1IN.
PHI2OUT	Similar to PHI1OUT for the second clock phase.
S1	Indicates when the commutator is scanning the first parity check.
S2	Indicates when the commutator is scanning the second parity check.
S3	Indicates when the third parity check is being scanned. In the case that a rate 1/2 code is selected, S3 is always zero.
P1OUT	The first phase of the shift register advancing clock.
P2OUT	The second phase of the shift register advancing clock.
X#9	The output of the shift register. This is simply BITIN delayed by 10 bit times.

As implemented, there are two ways of getting clocking information into the chip. A two-phase clock can be wired to the device by using PHI1IN and PHI2IN and grounding the input CLKSEL. Also, a one-phase clock can be used on the CLOCKIN input. In the latter case, the input pin CLKSEL must be wired to voltage. This two-choice system is possible because a two-phase clock generator has been included in the chip itself.

A timing diagram that shows the relationship between the time-varying signals is shown in Fig 2. It is assumed, for the moment, that the selected code rate is  $1/3$ . Notice that the clock frequency must be  $P$  times the input data rate (recall that  $P$  is the inverse of the code rate). The data that is entered on pin BITIN is clocked through the shift register by the signals P1 and P2. P1 and P2 are simply the logical "and" of S1 and the signals PHI1 and PHI2 respectively. The data in the register will remain stable during the three clock times that it takes to scan the three parity checks. The particular parity checks that are scanned are also functions of the input signal SEVEN. The signals S1, S2, and S3 represent the position of the commutator. It is important that the input bits must be synchronized with the commutator so that a new bit is entered whenever P1 is high. The output symbols on the pin OUTPUT are synchronized with PHI2. Since all the important timing signals are available as pins, it would be an easy task to interface this chip with data sources.

In the case that the selected code rate is  $1/2$ , then the clock frequency is twice that of the input data rate and the signal S3 is always low. This case is indicated in the timing diagram of Fig 3.

Although the actual clock speeds of the chip cannot be determined until the fabricated device is tested, it is likely that a maximum clock rate of at least 1-2 MHz will be achievable. Such clock rates are nominal for the NMOS technology that is being used for this project.

### III. Chip Architecture

The VLSI multicode convolutional encoder logic is divided into two major sections. The first is the timing signal generator and the second is the convolutional encoder array.

The timing signals that were described in Section II are generated in the timing signal generator subblock. A programmable logic array (PLA) (Ref 6) is used to implement a counter for generating the commutator signals S1, S2, and S3. If the input "1/3" is high, then the counter is set to count modulo three. If "1/3" is low, then it counts modulo two. P1 and P2 are generated by ANDing S1 with PHI1 and PHI2 respectively. This subblock also contains an output latch that synchronizes the output channel symbols with the signal PHI2 before they are sent to the output pad OUTPUT.

The actual encoding takes place in the convolutional encoder array subblock. It is shown conceptually in Fig 4. This figure also indicates the geometric layout of this section of the chip. The input bits from BITIN are shifted through the ten-bit parallel output register shown along the bottom of the figure. The output of each register cell is available at the top as

well as the right end. Above each cell are six blocks, each of which is either a PRTY (or "parity") cell or a NPRTY (or "no parity") cell. Each row of parity cells corresponds to a particular parity check on the ten outputs of the register. The presence of a PRTY cell in row  $i$  and column  $j$  means that the  $i^{\text{th}}$  parity check has a tap at position  $j$  in the register. The top three parity checks are for the  $(10, 1/3)$  code. The lower three are for the  $(7, 1/3)$  code. The  $(7, 1/2)$  code uses the first two parity checks from the  $(7, 1/3)$  code.

The parity checks are not implemented as exclusive OR operations as is done in a conventional encoder. Instead, there are two signal paths through each row of the parity check matrix. At the extreme left of the array, a "one" is connected to one of these and a "zero" to the other. Each time this pair of signals passes through a PRTY cell, they exchange places if the output of the corresponding register cell is a "one". A NPRTY cell has no effect on these signals (circuit diagrams for these two cells appear in Fig 5). In this way, the pair of signals has gone through a number of path exchanges equal to the number of ones in the shift register at the taps that correspond to that particular parity check. If the parity is even, then the signals are in the same place as when they started. If the parity is odd, they have come out reversed. Also, since a complementary pair of signals is used, the complement of the parity check, as well as the check itself, is available at this point.

The signals NRM1, NRM2, and NRM3 are used to select either the parity information or its complement at the end of each row of the parity array. The signal SEVEN is then used to select either the top three or bottom three parity checks. The commutator is implemented as a multiplexer in which the signals S1, S2, and S3 are used to select the appropriate parity check.

This architecture is a good example of how an algorithm that works well in a SSI system is not nearly optimal in VLSI. In the conventional encoder, it would be impractical to provide all the interconnections that are necessary for implementing the above algorithm. This VLSI architecture is compact, fast, and easily extendable to longer constraint lengths and more parity checks. In fact, it would be a relatively simple matter to write a silicon compiler (Ref 7) that would automatically lay out masks for similar encoders for arbitrary sets of convolutional codes.

It should be mentioned that the overall size of this chip was determined by the number of pads rather than the size of the logic. This means that there is some extra space available. Some of this space is filled with a long array of "butting contacts". A butting contact is a structure for connecting the polysilicon and diffusion layers in NMOS. There is some doubt as to whether this structure will work reliably in the 2.0-

micron NMOS technology. By including this test array it will be possible to gather some additional data on this subject. The two ends of the array of contacts are connected to the pins BUTT-TEST1 and BUTT-TEST2. It should be noted that the butting contact was not used in the remainder of the chip.

#### IV. The Design Process

The multicode convolutional encoder chip was designed according to the NMOS design rules that appear in Ref. 6. The design philosophy was influenced by the JPL VLSI design course as taught by Dr. G. Lewicki and Dr. R. Nixon.

The actual chip layout was performed using the interactive graphics editor "CAESAR" that was developed at the University of California at Berkeley by Dr. J. Ousterhout (Ref. 8). This software design system allows the user to develop a chip in a hierarchical fashion with full editing capabilities. There are also a number of programs including a PLA generator and optimizer that can be used in conjunction with CAESAR. A picture of the finished layout as produced with CAESAR appears in Fig. 6. The UC Berkeley software also includes a circuit extractor called MEXTRA that works in conjunction with CAESAR to create a file that can be used with a number of simulation packages. One very nice feature of the extractor is that it keeps track of labels that the user creates with CAESAR so they can be referred to during simulation. A complete description of the Berkeley software can be found in

Ref. 9. The chip was completely simulated on a logical level using the program ESIM that was developed at the Massachusetts Institute of Technology. Although it was not used for this project, the circuit level simulator SPICE (Ref. 10) is also integrated with CAESAR and MEXTRA.

#### V. Conclusions

It is evident, even in a simple project such as this one, that VLSI technologies allow the implementation of much more efficient algorithms than conventional circuit design. This is because all intermediate signals that are generated in a circuit can be made available to other portions of the circuit without a proliferation of pins and drivers that can greatly reduce the computational speed of the system.

The chip described here was sent out over the ARPANET for fabrication on December 2, 1982. When the completed chips return they will be evaluated and tested.

These chips will probably be used as test function generators in research on convolutional decoding. They will not be radiation hardened and so they cannot be flight qualified. However, the architecture described in this article could be easily used to produce flight qualified chips if the need ever arises. Also, the experience gained on this project will be invaluable in the design of the complex decoding systems that are planned for future projects.



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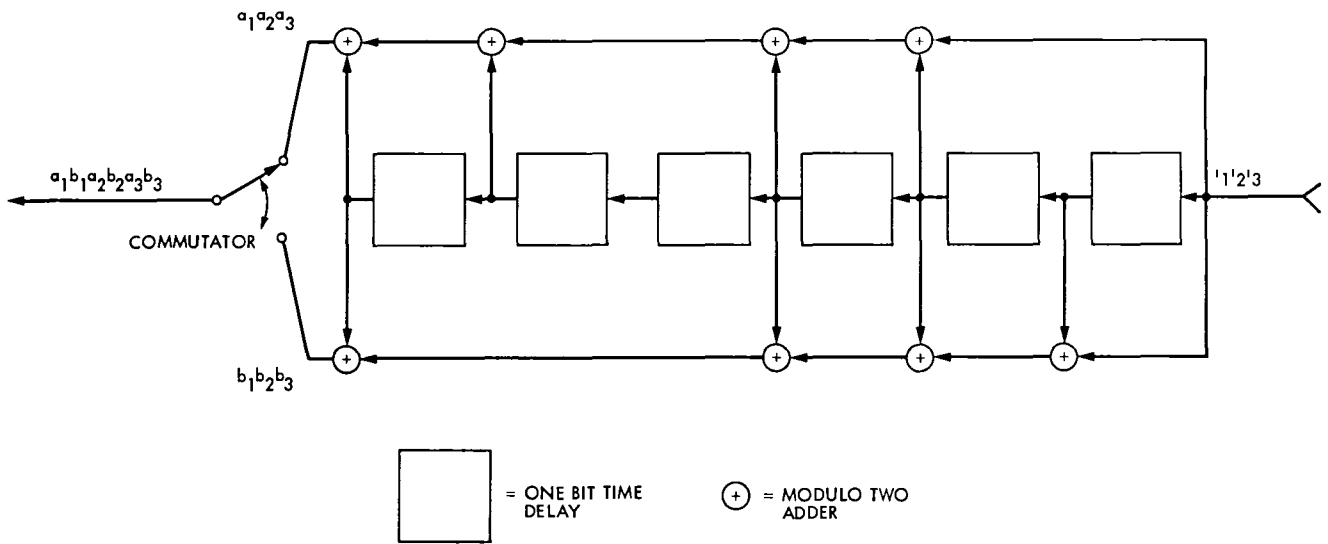


Fig 1 Conceptual diagram of a (7, 1/2) convolutional encoder

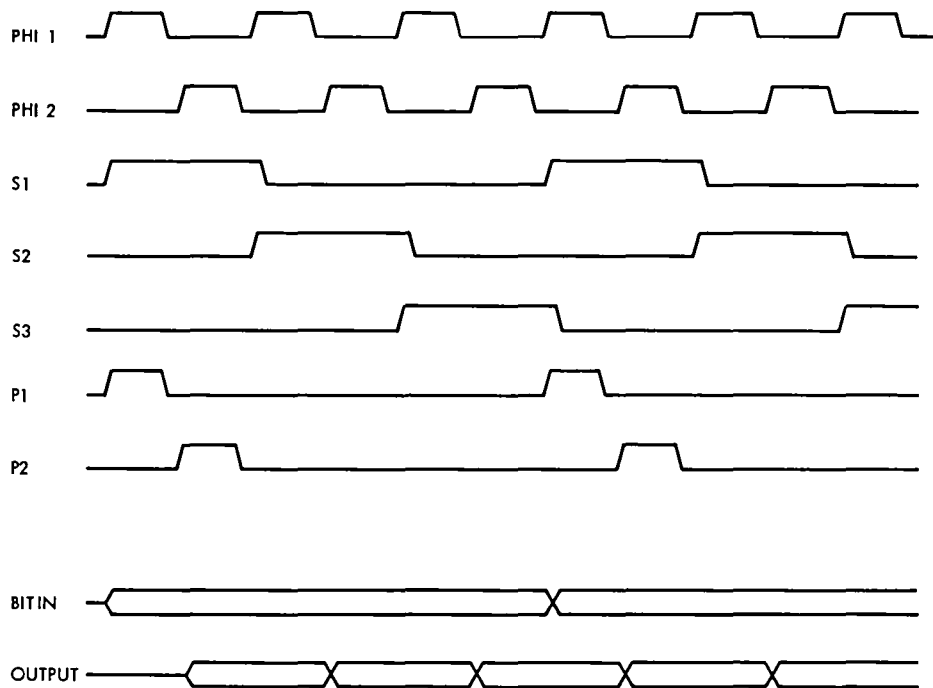
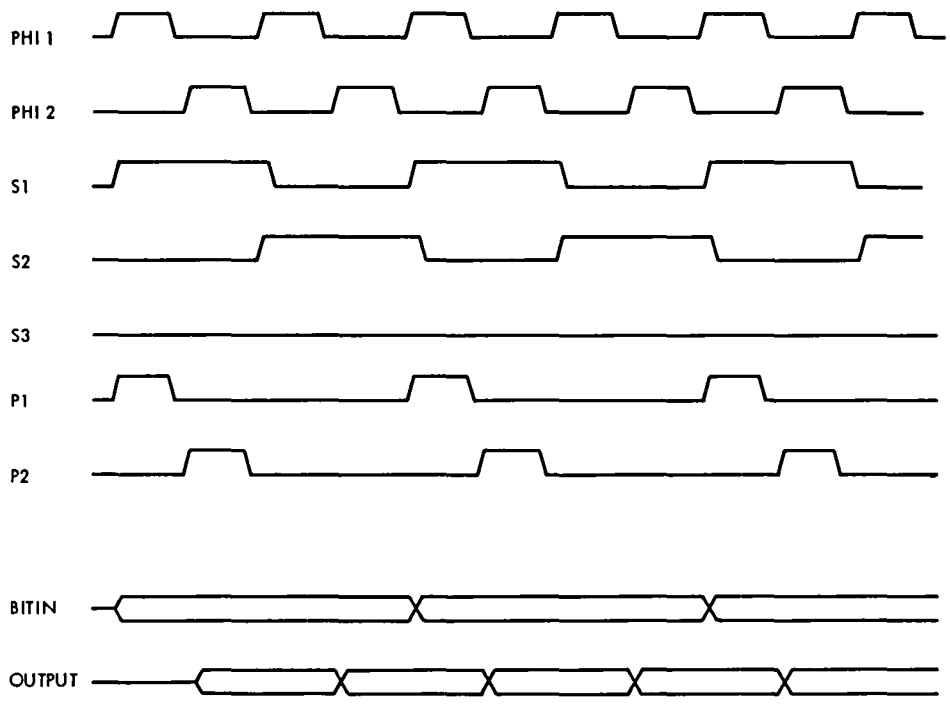


Fig 2 Timing signals for the convolutional encoder, rate 1/3 codes



**Fig 3 Timing signals for the convolutional encoder, rate 1/2 codes**

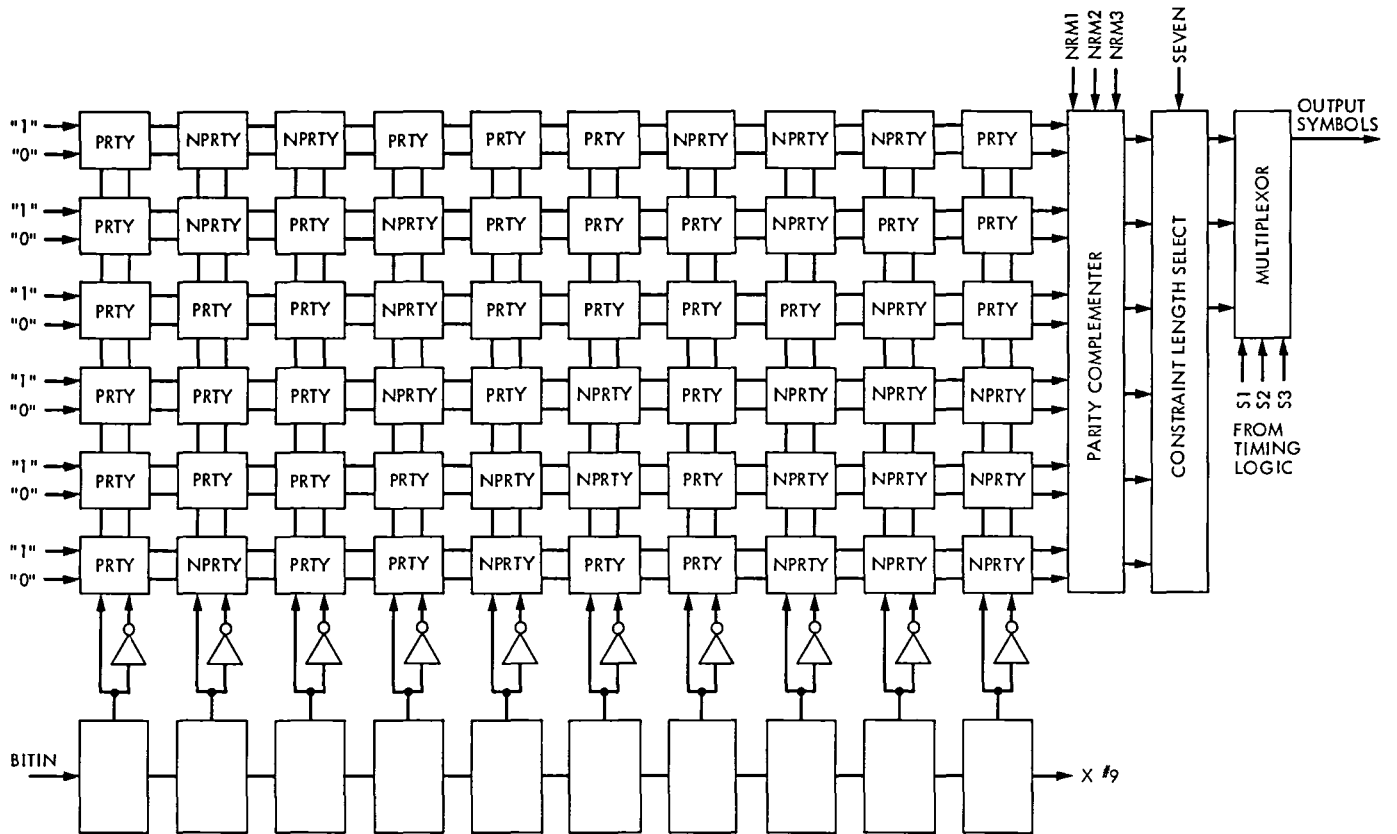


Fig 4 Conceptual diagram of the encoder logic array

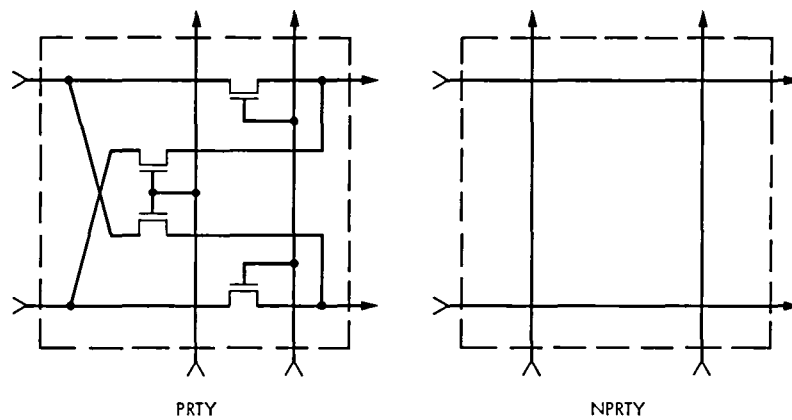


Fig 5 The PRTY and NPRTY cells

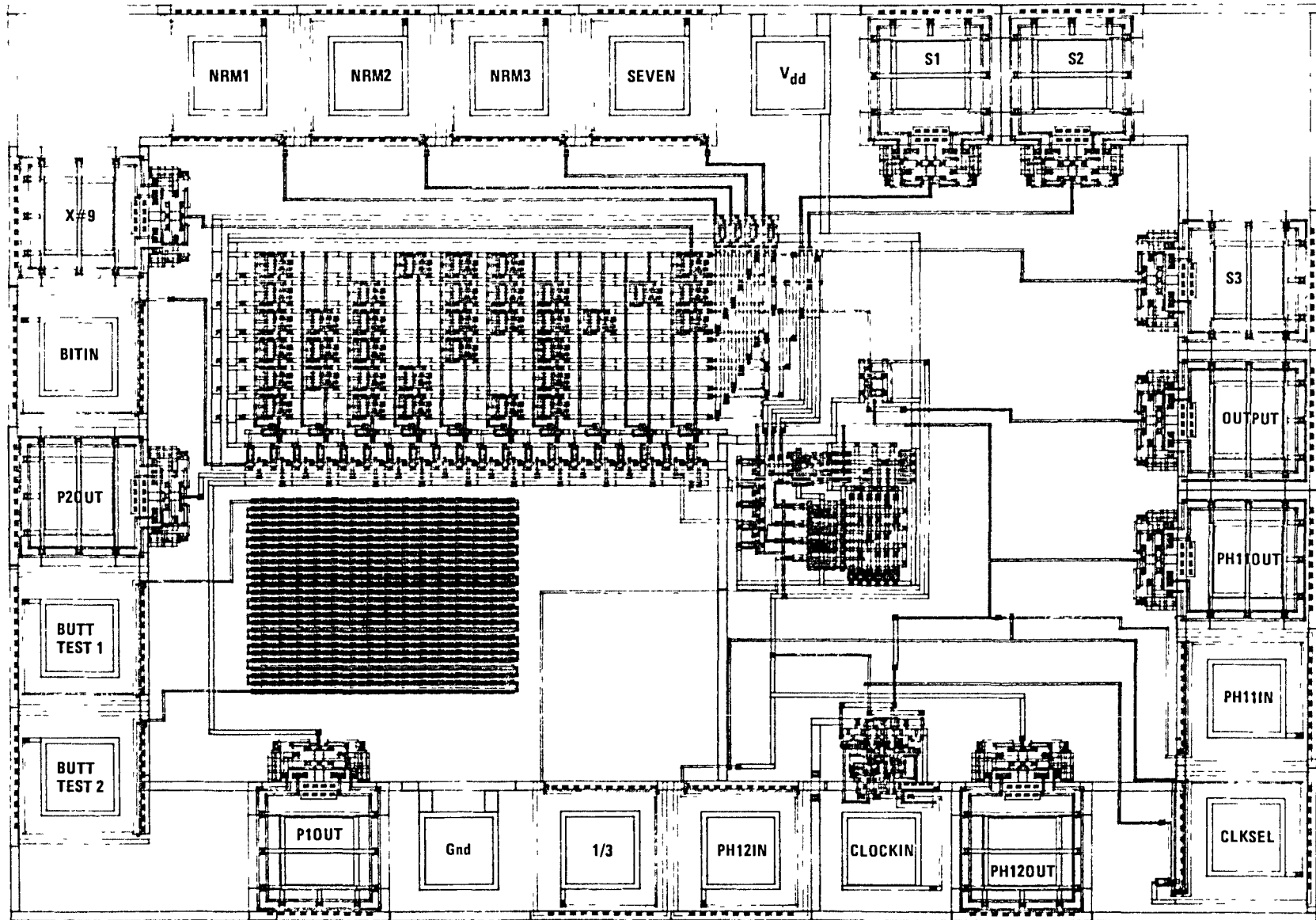


Fig. 6 Layout of the multicode convolutional encoder chip

# Staffing Implications of Software Productivity Models

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*This article investigates the attributes of software project staffing and productivity implied by equating the effects of two popular software models in a small neighborhood of a given effort-duration point. The first model, the “communications overhead” model, presupposes that organizational productivity decreases as a function of the project staff size, due to interfacing and intercommunication. The second, the so-called “software equation,” relates the product size to effort and duration through a power-law tradeoff formula. The conclusions that may be reached by assuming that both of these describe project behavior, the former as a global phenomenon and the latter as a localized effect in a small neighborhood of a given effort-duration point, are that (1) there is a calculable maximum effective staff level, which, if exceeded, reduces the project production rate, (2) there is a calculable maximum extent to which effort and time may be traded effectively, (3) it becomes ineffective in a practical sense to expend more than an additional 25-50% of resources in order to reduce delivery time, (4) the team production efficiency can be computed directly from the staff level, the slope of the intercommunication loss function, and the ratio of exponents in the software equation, (5) the ratio of staff size to maximum effective staff size is directly related to the ratio of the exponents in the software equation, and therefore to the rate at which effort and duration can be traded in the chosen neighborhood, and (6) the project intercommunication overhead can be determined from the staff level and software equation exponents, and vice versa. Several examples are given to illustrate and validate the results for use in DSN implementation.*

## I. Introduction

Brooks (Ref 1), in *The Mythical Man-Month*, proposed a simple model of software project intercommunication to show that, if each task of a large project were required to interface with every other task, then the associated intercommunication overhead would quickly negate the believed advantage of partitioning a large task into subtasks. While not meant to be an accurate portrayal of an actual project, the model effectively illustrated an increasing inefficiency symptomatic of projects too large to be performed by a single individual.

Putnam (Ref 2), in a 1977 study of software projects undertaken by the US Army Computer Systems Command,

discovered a statistical relationship among product Lines of code, Work effort, and Time duration for those projects whose best-fit formula was a power-law relationship, now referred to as the “software equation,”

$$L = c_k W^{0.33} T^{1.33}$$

(I have taken the liberty of changing Putnam’s notation in order to be consistent with my notation in the remainder of the article.)

One rather startling extrapolation one may make from the software equation is that in order to halve the duration of any

one of the projects studied, it would have taken 16 times the resources actually used! I say "extrapolation" because I suspect the software equation is more likely to be applicable incrementally—that is, if one were to require a 5% shortening of the schedule, then a 20% (actually 21.5%) increase in resources would be required

In this paper, I will generalize both of these models parametrically, and suppose that both do describe the statistical trends of software projects in small neighborhoods about a chosen project situation. By equating the model behaviors in these neighborhoods, we shall be able to see how the parameters of one model relate to the parameters in the other. In addition, we shall discover some rather interesting facts about some actual projects for which published data exists

## II. A Generalized Intercommunication Overhead Model

Let us suppose that a software project is to develop  $L$  kilo-lines of executable source language instructions, and that this number remains fixed over all our considerations of effort, duration, staffing, etc. That is, we shall suppose that the product size is invariant over the neighborhood of variability in these parameters—a project utilizing greater effort attempting to shorten the schedule slightly would produce the same program as a smaller effort requiring somewhat more time

Let us denote by  $W$  the Work effort (in person-months) to be expended in the production of the  $L$  lines of code, and let the Time duration, in months, be denoted by  $T$ . Then the average full-time equivalent Staff size  $S$  in persons, is

$$S = \frac{W}{T}$$

and the overall team productivity can be defined as the number

$$P = \frac{L}{W} \text{ (kilo-lines/person-month)}$$

Let us further suppose that the average fraction of time that each staff member spends in intercommunication overhead is dependent on the staff size alone, within a particular organizational structure and technology level, and let this fraction be denoted by  $t(S)$

$$t(S) = \frac{\text{(intercommunication time/mo)}}{\text{(hours/mo worked)}}$$

Generally speaking, one intuitively expects  $t(S)$  to increase monotonically in  $S$  due to the expanding number of potential

interfaces that arise as staff is increased. But the *individual average productivity* of the staff, defined as the individual productivity during nonintercommunication periods,  $P_i$ , is somewhat greater than  $P$ , being related to it by

$$P = P_i [1 - t(S)]$$

The relationship between the number of kilo-lines produced, the effort, and the staffing is

$$L = P_i W [1 - t(S)]$$

Let us denote by  $W_0$  and  $T_0$  the effort and time, respectively, that would be required by a single unencumbered individual to perform the entire software task (assuming also that it could be done entirely by this individual, no matter how long it took). Then, with respect to the actual  $W$  and  $T$ , there is the relationship

$$W_0 = \frac{L}{P_i} = W [1 - t(S)] = T_0$$

Thus  $W_0$  represents the least effort that must be expended, and  $T_0$  is the maximum time that will be required. By substituting  $W/T$  for  $S$ , one obtains an effort-time tradeoff relationship

$$\omega = \frac{1}{[1 - t(\omega/\tau)]}$$

where  $\omega = W/W_0$  and  $\tau = T/T_0$  are "normalized" effort and duration, respectively

The rate at which an increase in staffing results in an increase in normalized work effort is then

$$\frac{\partial \omega}{\partial S} = \omega^2 t'(S) > 0$$

where  $t'(\ )$  refers to the derivative of  $t$  with respect to  $S$ . Because of the monotone character of  $t(S)$ , an increase in staff leads to an increase in effort

The overall staff production Rate  $R$  is the number of kilo-lines of code *per month* produced by the entire team of  $S$  persons,

$$R = P_i S [1 - t(S)]$$

The factor

$$\eta = [1 - t(S)]$$

is then the *team production efficiency*. Note that the normalized task effort is the inverse of the production efficiency,

$$\omega = \frac{1}{\eta}$$

The maximum rate of software production will occur when the derivative of  $R$  with respect to  $S$  becomes zero, a condition requiring a value  $S_0$  that will satisfy the relationship

$$t'(S_0) = \frac{[1 - t(S_0)]}{S_0}$$

We shall refer to this staffing level as the *maximum effective staff*. Two particular examples of  $t(S)$  will serve to illustrate the characteristics of the intercommunication overhead model

### A. Linear Intercommunication Overhead

Let us assume first, as did Brooks, that the overhead is linear in staff,

$$t(S) = t_0(S - 1)$$

That is, there is no overhead for 1 person working alone, but when there are  $S - 1$  other people, then each requires an average fraction  $t_0$  of every other individual's time. Under these assumptions, the maximum effective staff level is

$$S_0 = \frac{(1 + t_0)}{(2 t_0)}$$

This value yields a maximum team production rate of

$$R_{max} = \frac{P_t S_0^2}{(2 S_0 - 1)}$$

and team efficiency

$$\eta_0 = \frac{(1 + t_0)}{2 S_0} (2 S_0 - 1) \approx 0.5$$

This perhaps alarming result states that a team producing at its maximum rate is burning up half its effort in intercommunication overhead! The behavior is illustrated in Fig. 1

The normalized effort-duration tradeoff equation for this model takes the form

$$\tau = \frac{t_0 \omega^2}{(1 + t_0) \omega - 1}$$

which has its minimum value at the maximum-production-rate point,

$$\tau_{min} = \frac{4 t_0}{(1 + t_0)^2} \approx 4 t_0$$

at which point the normalized effort is

$$\omega_0 = \frac{2}{(1 + t_0)} < 2$$

Figure 2 shows the characteristic of this tradeoff law at  $t_0$  values of 0.1 and 0.2 for illustrative purposes

According to this model, it *never* pays to expend more than twice the single-individual effort. Moreover, even though the  $\omega$  producing the shortest schedule is less than 2, the cost-effective range is much less than this, as shown in the figure. Effort can be traded for schedule time realistically only up to about 1.25  $W_0$ , and a factor of 2 reduction in time can only come about if the individual intercommunication can be kept below about 15% per interface.

### B. Exponentially Decaying Intercommunication Overhead

One unsettling aspect of the linear intercommunication overhead model is that, at some staffing level, the production rate goes to zero, and beyond, unrealistically into negative values. Perhaps a more realistic model is one which assumes that  $t(S)$  tapers off, never exceeding unity, at a rate proportional to the remaining fraction of time available for intercommunication as staff increases, or

$$t'(S) = t_1 [1 - t(S)]$$

Then we are led to the form

$$t(S) = 1 - \exp[-t_1(S - 1)]$$

The maximum effective staff in this case becomes

$$S_0 = \frac{1}{t_1}$$

and the maximum production rate is

$$R_{max} = P_t S \exp\left(-1 + \frac{1}{S}\right) \approx \frac{P_t S}{e}$$



The team efficiency at this rate is

$$\eta_0 = \exp\left(-1 + \frac{1}{S}\right) \approx \frac{1}{e}$$

Now this is perhaps even more alarming a revelation than before, because it says that when producing software at the maximum team rate, that team is burning up 63% of its time in intercommunication! The consolation, as shown in Fig 1, is that the team performance under this assumed model is superior to that of the linear-time team model. More staff can be applied before the maximum effective staff level is reached.

The effort-duration tradeoff equation according to this model is

$$\tau = \frac{t_1 \omega}{[t_1 + \ln(\omega)]}$$

The minimum  $\tau$  occurs at

$$\omega_0 = \exp(1 - t_1) < e$$

and the minimum value is

$$\tau_{min} = t_1 \exp(1 - t_1) \approx e t_1$$

The form of this tradeoff is shown in Fig 3 for  $t_1$  values of 0.1 and 0.2, for illustrative purposes. Note that the minimum  $\tau$  is much broader in this model, so that, although the actual minimum occurs when  $\omega$  is about  $e$  in value, the practical effective range for  $\omega$  is less than about 1.5. That is, it is not cost-effective to expend more than about 1.5 times the single-individual effort  $W_0$  in an attempt to reduce the schedule time. A reduction in schedule by a factor of 2 is possible only when the individual intercommunication factor  $t_1$  can be kept below 0.2.

### C. Conclusions from Intercommunication Overhead Models

Both of the examples of intercommunication overhead above bespeak a maximum effective staffing level at which the project is 37-50% efficient. Beyond this point, further staffing is counterproductive. Both examples conclude that the maximum practical extent to which added effort is effective in buying schedule time is limited to about 25-50%. Significant schedule reduction factors are possible only when the intercommunication factors can be kept below 15-20%.

## III. Matching the Software Equation Model

Let us generalize the Putnam Software Equation as the form

$$L = c_k W^p T^q$$

and let us define  $r = q/p$ , the exponent ratio. As in the previous section,  $L$  is held constant with respect to effort-duration tradeoff considerations. The value of  $p$  is assuredly positive; it generally requires more work at a given  $L$  to reduce  $T$ . If  $q$  is positive, effort can be traded to decrease the schedule time required to deliver a given  $L$ . The larger  $r$  is, the larger the increase in effort required to shorten the schedule, and the larger the team production inefficiency. If  $q$  is zero, then  $L$  is a function of  $W$  alone,  $T$  is determined solely by the staffing level,  $T = W/S$ , and no additional effort is required to reduce schedule time (in the neighborhood in which the  $p$  and  $q = 0$  are valid). If  $q$  were ever to be negative, then an increase in  $W$  would render an increase in  $T$ , a situation indicating overmanned projects.

Substitution of  $T = W/S$ , differentiation with respect to  $S$ , and normalization of the software equation produces the result

$$\frac{\partial \omega}{\partial S} = \omega \frac{r}{[S(1+r)]} = \tau \frac{r}{(1+r)}$$

Let us now suppose that both the software equation and the intercommunications overhead model agree at the point  $(L, W, T)$ . The two models can be equated by suitable choices of the "technology constant"  $c_k$  and individual productivity  $P_i$ . Then, in addition, let us suppose that the derivatives of effort with respect to staff level for both models also agree at this point. Such can only be attempted when  $r > 0$ , because the derivative in the intercommunication overhead model is always positive. When this is the case, the two models may be said to agree in the neighborhood of the point  $(L, W, T)$ .

Thus, by equating the derivatives, we arrive at a relationship between the parameters of the two models

$$\frac{S t'(S)}{[1 - t(S)]} = \frac{r}{1+r}$$

or

$$\eta = S t'(S) \frac{r+1}{r}$$

Let us now examine this relationship for the two examples of the interface overhead model

## A. Linear Intercommunication Overhead

Substitution of the linear  $t(S)$  form into the neighborhood agreement condition yields

$$S = \left[ \frac{2r}{1+2r} \right] \left[ \frac{1+t_0}{2t_0} \right] = S_0 \frac{r}{r+0.5}$$

This equation states that the staffing level is related to the maximum effective staff point through the software exponent ratio  $r$ . At the Putnam value,  $r = 4$ , the staffing level is 89% of the maximum effective level, and the team efficiency is

$$\eta = 0.55(1+t_0) \approx 55-65\%$$

$$\omega = \frac{1.8}{1+t_0} \approx 1.5-1.8$$

As seen in Fig 2, projects having this high an  $\omega$  are at the point that extra effort is very ineffective

## B. Exponentially Decaying Intercommunication Overhead

By substituting the exponential form for  $t(S)$  into the neighborhood agreement condition, we find

$$S = \frac{r}{[t_1(1+r)]} = S_0 \frac{r}{(1+r)}$$

Again, we see that the staffing level is related to the maximum effective staff via the exponent ratio. The Putnam value  $r = 4$  produces

$$S = 0.8 S_0$$

$$\eta = \exp \left[ -\frac{(S-1)}{S_0} \right] = \exp [-0.8 + t_1] \approx 45\% - 55\%$$

$$\omega = \frac{1}{\eta} = \exp [0.8 - t_1] \approx 1.8 - 2.2$$

Although this example indicates a somewhat more comfortable margin below maximum effective staffing than did the linear model, it nevertheless shows an alarmingly low cost inefficiency

## IV. Examples Using Available Data

Several data sets of project resource statistics published in the literature readily show that Putnam's value of  $r = 4$  is not

universal. Specifically, Freburger and Basili (Ref 3) publish data which yield the following 3-parameter best power-law fits

$$L_0 = 1.24 W^{0.95} T^{-0.094} \quad (r = -0.1)$$

$$L_1 = 0.22 W^{0.78} T^{0.78} \quad (r = 1.0)$$

in which  $L_0$  is kilo-lines of delivered code, and  $L_1$  is *developed* delivered code. It is interesting here to note that the former relationship is nearly independent of  $T$ , whereas the latter shows a definite beneficial  $W - T$  tradeoff characteristic. The negative  $q$  in the former relationship indicates that, on a *delivered code basis*, added resources in one of the projects would have extended the schedule! An equivalence between the software equation and the intercommunication overhead model cannot be established when  $r$  is zero or negative

This data set is not the only one to show a negative  $q$ . Boehm (Ref 4), in his *Software Economics* book, has a data base used to calibrate his COCOMO software cost model. A 3-parameter best power-law fit to the adjusted data produces the relationship

$$L = 0.942 W^{0.675} T^{-0.028} \quad (r = -0.041)$$

and, on the unadjusted data,

$$L = 0.957 W^{0.646} T^{0.0555} \quad (r = 0.086)$$

Gaffney (Ref 5), on the other hand, did a 3-parameter best power-law fit of IBM data (Federal Systems Division, Manassas) to arrive at the relationship

$$L = c_k W^{0.63} T^{0.56} \quad (r = 0.88)$$

This last value of  $r$  aligns more closely with the Freburger-Basili value for developed delivered code

Figure 4 shows plots of staffing normalized to maximum effective levels as a function of  $r$ , for both examples of the intercommunication overhead model. Efficiency curves are shown in Fig 5

## V. Conclusion

This article has shown that when there is a positive effort-duration tradeoff relationship in a software project, it is possible to estimate the team production efficiency and proximity to maximum effective staffing. These figures can be

used to advantage by software managers who must judge the effectiveness of increasing resources in order to shorten schedules. It points out the necessity of keeping accurate records of software project statistics, so that the parameters in the model can be estimated accurately.

Low values of  $r$  in an organization are a mark to be proud of, showing efficiency in terms of structuring subtasks for clean interfaces. High (or negative) values of  $r$  may be indicative of overall task complexity, volatility of requirements, organizational inefficiency, or any number of other traits that tend to hinder progress. The value of  $r$  may thus be

treated as a figure of merit—a measurable statistic indicative of the efficiency of a set of projects in performance of assigned tasks.

The ratio  $S/S_0$  is another indicator for management. When low, it indicates that adding resources can potentially help a project in schedule trouble. If closer to unity, it is a warning that adding resources may not help, will not appreciably shorten the schedule, will incur expense at a low return in productivity, and, if applied often in other projects, will thereby contribute to an organizational reputation for expensive software.

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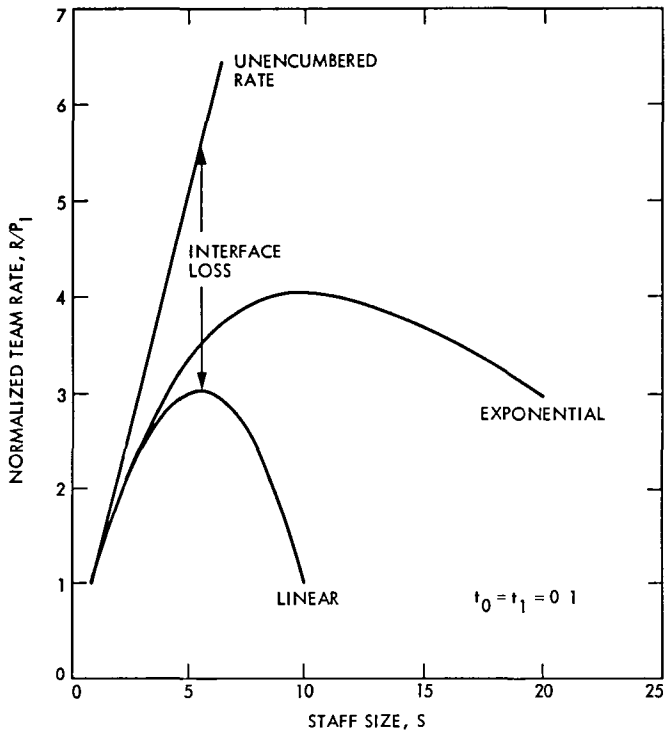


Fig 1. Software team production rate, intercommunications overhead model

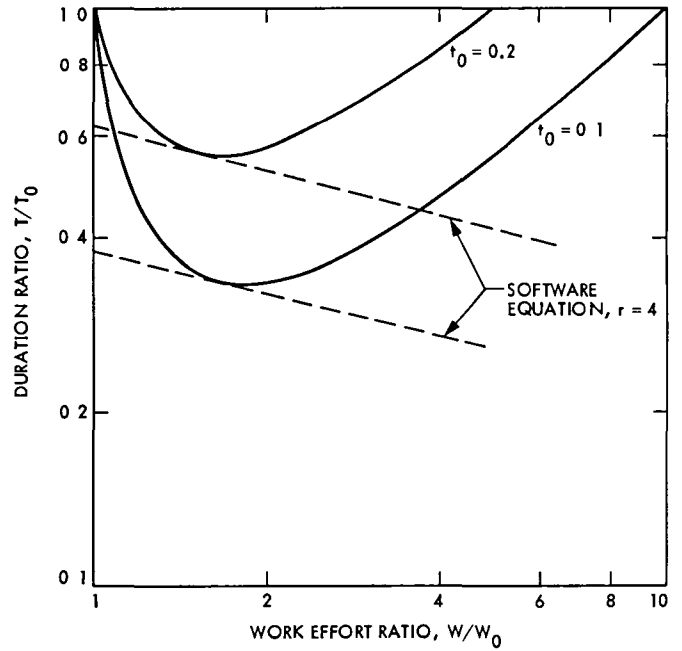


Fig. 2. Time duration vs effort tradeoff curve for linear intercommunication overhead model

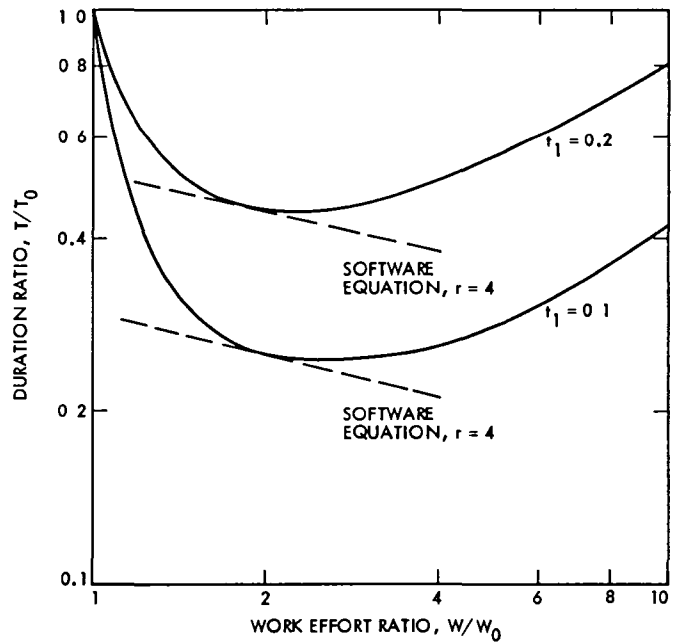


Fig 3. Time duration vs work effort tradeoff characteristic for the decaying exponential intercommunication overhead model

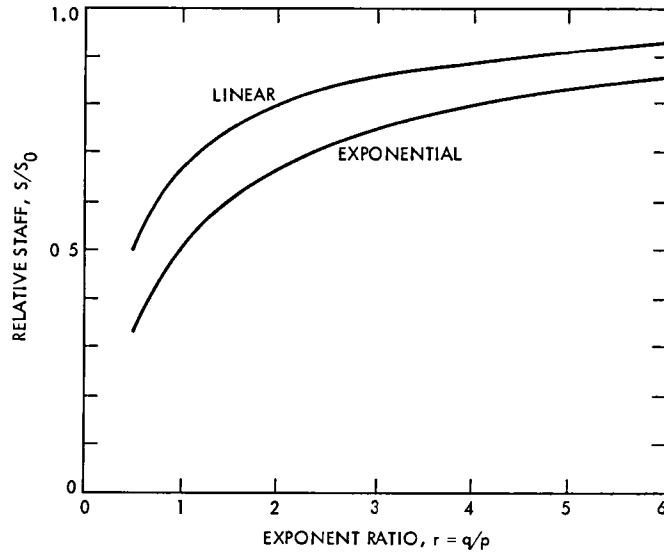


Fig. 4. Normalized staff levels for the intercommunication overhead model

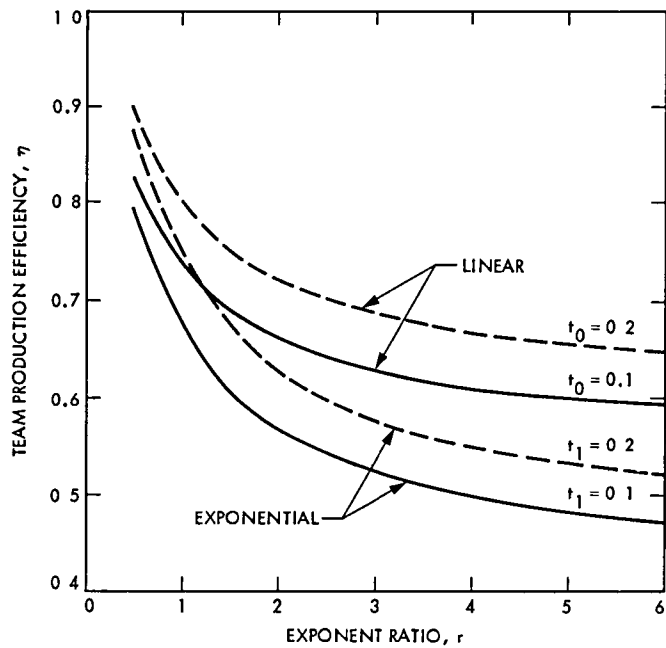


Fig 5 Team production efficiency of projects as a function of the exponent ratio and intercommunications overhead parameter

# Applications of Telemetry Arraying in the DSN

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*The individual large antennas at a DSN complex have been used in arrayed configurations to improve the telemetry return from very distant spacecraft. The technique has been applied to support critical spacecraft events when good telemetry is most important. An expanded array configuration, using the Parkes, Australia, Radio Astronomy Observatory will be provided to support the 1986 Voyager encounter with Uranus. This report reviews the development and use of arraying by the DSN for spacecraft mission support. Application of the experience to the 1986 design is discussed.*

## I. Introduction

For support of the 1986 Voyager 1 encounter with Uranus, the DSN is planning to array telemetry signals from the Parkes, Australia, Radio Astronomy Observatory 64-m antenna with those of the 64-m and two 34-m antennas of the Tidbinbilla DSN complex. During the process of assessing the array configuration system design for that application, the previous DSN operational experience with arrayed configurations was reviewed. The purpose of the review was to ensure that lessons from experience were not neglected.

The operational applications of arraying for telemetry reception by the DSN have been described in past TDA Progress Reports. The reporting covers all applications, except for the Voyager 2 Saturn encounter, which will be reported in a future issue. This material is an abstraction of those reports and of discussions with some of the authors. It also includes discussion of anticipated new factors involved with the integrated participation of the Parkes Facility for critical scientific deep space mission support.

## II. Arraying Phase 0

In 1970, conceptual levels of combining were described and their basic performance analyzed. The levels described were carrier, baseband, bit stream, and processed data. Baseband and bit stream combining were configured with two 26-m antennas and demonstrated using the signals from the Pioneer 8 spacecraft. The results showed approximately 2.5 dB (vs 3 dB maximum possible) improvement with baseband and bit stream combining. Time delay tracking of the baseband signals was not required or used because of the low subcarrier frequency of the Pioneer signals. The work was done by the personnel of the Spanish complex using available station equipment and the DSS 61/62 26-m antennas (Ref 1).

The next effort was demonstration of RT baseband combining of high-rate telemetry via microwave links of the signals from the DSS 12 and DSS 13 26-m antennas with the signal from DSS 14 (Ref 2). The array was used to enhance the telemetry signals from the MVM spacecraft during its September 1974 second Mercury encounter. The system was

designed, built, and operated by Advanced Systems Program personnel with support from Operations personnel. Installation and initial testing of the array system at Goldstone was straightforward. Array performance was evaluated using spacecraft signals from far encounter TV sequences. These tests indicated that the performance of the Subcarrier Demodulation Assembly (SDA) processing string, which was hard-wired to the array combiner signal stream, performed about 1/2 dB poorer than the string which was hard-wired to the 64-m-only signal stream. In the time available, it was not practical to alter or improve the situation. As a result, during the encounter pass, the arrayed signal stream was only 0.3–0.4 dB improved over the 64-m-only signal stream, rather than the  $\approx 0.7$  dB, which was in principle achievable. The  $\approx 0.7$  dB (vs 0.8 theoretical) was demonstrated by deleting briefly the 26-m antenna signals from the combiner input and observing the increase in image data bit error rate.

During a portion of the encounter pass over Goldstone, the spacecraft transmitted at maximum data rate (117 kbs vs 22 kbs), which was only supportable by the arrayed configuration. Although the full 0.7 dB capability of the array could not be exploited, the 0.3–0.4 dB improvement did significantly enhance the mission imaging data yield. The experience showed the difficulty in achieving full performance of an arrayed system, and the importance of a few tenths of a dB of performance.

### III. Arraying Phase I

The 1973-74 R&D demonstration project just described established technical feasibility of baseband arraying of very weak high-rate signals. In 1977, the DSN started to develop an operational arraying capability for the network. The first application was for Voyager encounters of Jupiter and Saturn. The first phase was providing a prototype system at Goldstone to be used for enhancement of the Jupiter encounters (Refs 3, 4).

A prototype baseband real-time combiner (RTC) based on the analysis and design techniques developed by the earlier R&D activity was completed in the fall of 1978. It was designed to combine signals from DSS 14 and DSS 12. The signals from DSS 12 were sent to DSS 14 via the existing microwave link. The RTC was installed at the Compatibility Test Area (CTA 21) for initial testing. The tests were conducted in November and December 1978. The tests indicated that the RTC performed according to its analytical model. Conducting the test program at CTA 21 was surprisingly difficult and time consuming, finally, the equipment in the telemetry strings had to be carefully realigned and recalibrated and the test data taken at night to avoid electrical transients. Most

of the problems encountered were external to the RTC—in the receiver/telemetry strings and test instrumentation.

The RTC was installed at DSS 14 in mid-December 1978. After checkout and alignment, the system was tested by receiving Voyager signals. Performance on the initial Voyager tests in December was inconclusive. After further off-line testing and alignment of the system, array gain on Voyager signals, ranging from 0.4 to 1.0 dB, was obtained in mid-February 1979. With intensive work by the design and operations personnel, the array system was brought on-line for the Voyager 1 Jupiter encounter during the first 10 days in March. It was brought up and used again for the Voyager 2 Jupiter encounter for two weeks in early July 1979.

The performance of the prototype array system during the Voyager Jupiter encounter operations was good. It provided approximately 1.0 dB S/N improvement vs the 64-m-only, with an availability of  $\approx 0.95$ . The Voyager Project used the data from the array in preference to the on-line alternate 64-m-only stream for most of each tracking pass.

The prototype array system was next configured to provide enhanced support of the early September 1979 encounter of Saturn by the Pioneer 11 spacecraft (Ref 5). Pioneer 11 had two important data rate options for the encounter/near encounter period, 1024 and 512 bps. The higher rate was greatly desired, but its support appeared marginal. To enhance the support capability, the DSN installed ultra-low noise listen-only front ends and selected the best performing of normally equal telemetry strings at the 64-m stations, also, it adapted the DSS 14-12 RT array for Pioneer S-band parameters.

At S-band, the S/N performance of the DSS 12 34-m system relative to the DSS 14 64-m system is poorer than at X-band. As a result, the potential arraying gain was only 0.5 dB for Pioneer vs approximately 1.0 dB for Voyager. Also, the DSS 12 Pioneer signals were so weak as to not support automatic tracking of their time delay by the RTC. The varying time delay between the DSS 14 and DSS 12 signals during a pass was calculated and input to the RTC (that open-loop mode was provided for in the prototype RTC design, and it worked satisfactorily on the relatively narrow baseband Pioneer signal).

The initial tests in early August of the array on the Pioneer signals were not successful. An intensive effort was required to resolve the performance problems with the array system. On August 25, the first day of planned mission support by the array, it worked well. It provided 0.4 to 0.5 dB performance improvement relative to the 64-m-only configuration and supported the 1024 bps rate. On the next pass, it was no better than the 64 m's. On the following pass, after

inconclusive testing and diagnosis, it was back working properly

The 10-day encounter sequence was hectic for the Pioneer Project and the DSN, unexpected solar noise and satellite RFI complicated the data rate strategy. The efforts to maximize imaging data (by using 1024 bps) was impacting non-imaging experiments when the rate was unsupportable. By two days before encounter, the array performance had stabilized fairly well. This allowed use of the 1024 rate during mid-pass over the Goldstone array, 512 was used at other times and places.

The experience on Voyager and Pioneer with the prototype RTC and array configuration showed the need for central control and monitoring of the array system, and again, the importance of having *all* elements of the array—receivers, antennas, instrumentation—operating completely correctly. Also, array performance testing on live spacecraft signals was next to impossible if the spacecraft and the ground stations were not in stable configurations dedicated to supporting the array tests—array testing on a noninterference basis was usually unproductive. This experience was applied to the design of the equipment and use procedures for the next phase of arraying system development.

#### IV. Arraying Phase II

The objective of this phase was to provide operational arraying system configurations to combine the signals from the 64-m antenna and the 34-m antenna at each of the three DSN complexes. The arrayed systems were committed for operational support of the Voyager 1 (fall, 1980) and Voyager 2 (summer, 1981) encounters with Saturn (Ref 6).

The design of the operational RTC's and arraying system was a direct upgrade of the Goldstone prototype just described. Modifications to the system and its operating procedures were based on the experience with the Goldstone prototype. The principal modifications to the system included

- (1) Displays of 34-m station receiver in-lock and AGC were provided at the 64-m stations
- (2) The Block IV receiver telemetry detectors were modified to improve performance and stability
- (3) A routine procedure for measurement of microwave link path delay was developed (DSS 12-14 use)
- (4) Pretrack countdown time for the RTC's was reduced
- (5) The RTC was modified so that it could be restarted or accept new parameter inputs, without requiring reinitialization

- (6) The procedures for testing the array performance on spacecraft signals were improved
- (7) Simultaneous displays of symbol SNR for the 64-m-only, 34-m-only and the arrayed signals were provided at both the 64-m station and at the JPL network advisor's position

The first operational RTC was installed at Goldstone in mid-May 1980. Testing of the DSS 14-12 array on Voyager signals was started on May 19. Eight performance tests and training sessions were conducted in the following month's time. Installation of the systems at the overseas stations was then done, followed by a similar test and training sequence. The process of bringing up the operational systems was difficult, but much easier than the previous experiences with the prototype system.

Beginning in mid-August 1980, the arrayed configurations were used at all stations at regular intervals for Voyager 1 Saturn far-encounter operations support, and for DSN test and training. For the near-encounter period, the arrayed configuration was used for all passes (Ref 7).

During the first several days of the period the array performance of the complexes was only "fair". At that point (November 1, 1980, 12 days before encounter) a procedure for assessing array system performance at regular frequent intervals was established. The procedure was at 1-hour intervals the JPL network advisor computed a 10-minute average of the symbol SNR from the 34-m, the 64-m, and the arrayed signal streams of the active DSN complex.

The "observed" arraying gain was calculated as  $\text{SNR}(\text{array})/\text{SNR}(64\text{-m})$ , the "theoretical" or "loss-free" arraying gain was calculated as  $1 + \text{SNR}(34\text{-m})/\text{SNR}(64\text{-m})$ . If the ratio of the two gain numbers exceeded 0.3 dB, the nominal combining loss plus tolerance, the DSN complex was alerted to look for possible system problems. That procedure was used on all remaining passes through encounter, and good stable array performance was achieved. The average array gain for the encounter period was  $0.62 \pm 0.15$  dB relative to the 64-m-only and the array availability was approximately 0.95 (Cf Ref 7).

In preparation for the late August 1981 Voyager 2 Saturn encounter, array system test and training was started by the Network in early April (Ref 8). The work went fairly smoothly. It was supported by arrayed stream, 64-m-only stream, and 34-m-only stream symbol SNR performance monitoring in real-time at the complexes and at JPL.

All complexes supported the beginning of the observatory phase on June 5, 1981, in the arrayed configuration, with



very good results. Array training and performance validation tracks on the Voyager signals were conducted at regular intervals until the start of near-encounter operations.

During the 30-day near-encounter period, the arrayed configuration was used extensively. The procedure of hourly collection of 10-min average symbol SNRs and calculation of the combining loss was used throughout the 30-day period. It was the basic procedure developed during the Voyager 1 Saturn encounter. To reduce operations staffing required, the 10-min averages were automatically calculated in the station Telemetry Processor Assemblies (TPAs) and reported verbally to the Network Analysis Team at JPL. Also, the near-real-time (within 10 minutes) image display at the stations was found to be useful in keeping the arraying performance near optimum at all times.

The results during the Voyager 2 Saturn encounter were very satisfactory, clearly the best yet achieved. The *network* average of observed array gain relative to 64-m-only was 0.8 dB with a standard deviation of 0.2 dB. The *network* average combining loss was  $0.2 \pm 0.05$  dB. Those results are consonant with the design predictions/specifications. Out of a total of 79 passes scheduled for arraying support during the near-encounter operations, satisfactory array capability was provided for 75 of them, giving an availability of  $\approx 0.95$ . These above results are from a preliminary analysis of operations data.

The experience with the operational RTC and 64-m–34-m arraying systems for the Voyager 1 and 2 Saturn near-encounter operations showed the following:

- (1) The array system works methodically according to its design principles.
- (2) Array availability data are sparse and subject to interpretation, but some conclusions can be drawn. Downtimes are the sum of the 64-m string, the 34-m string and the combining system downtimes in that order of contribution, a logical result. The availability of the total array from about 7 weeks of encounter operations is  $\approx 0.95$ , not a super good result, but probably representative of the present system capability.
- (3) Extensive real-time visibility into the performance of the individual strings and the total system has been provided by the operations organization during the evolution of the use of the arrayed configurations. It is concluded that was done because it was needed.
- (4) Bringing up and validating a new array configuration is difficult but can be systematically accomplished, culminating with 6-10 scheduled spacecraft live data exercises.

## V. Arraying Phase III and Beyond

The next application of DSN arraying will be support of the Voyager 2 Uranus encounter in January 1986. At the Australia and Goldstone complexes, the 64-m and two 34-m's will be arrayed, at Madrid, the 64-m and the single 34-m will be arrayed. Also, at Australia, it is planned to array the Parkes 64-m with the DSN complex. The Parkes facility is about 280 km distant from the DSN Tidbinbilla complex.

The design and the operational use planned for the 1985-86 arraying system for the DSN complexes is based on the previous configurations. It will be baseband with RT-only combining at the DSN complexes. The combining method, RT-only, NRT-only or hybrid, for the Parkes signals is being reviewed—the current baseline design includes both RT and NRT.

A new Baseband Assembly (BBA) is being designed for the DSN complexes. The BBA will provide the functions of the RTC, the SDA, and the Symbol Synchronizer Assembly (SSA) in a single integrated unit. The new design is expected to improve the operational performance, stability, and monitor capability of the combiner/telemetry channel.

Also, the Mark IVA DSN will incorporate improvements in the antenna pointing system and monitor and control of the station equipment. Without these kinds of improvements, the availability of the planned three-element DSN arrays would clearly be less than the  $\approx 0.95$  of the present two-element configurations. The design availability for the Mark IVA Australian DSN complex 64-m –  $2 \times 34$ -m array is  $\approx 0.97$ , an ambitious objective.

In the baseline design for the planned Parkes-Tidbinbilla array, both RT and NRT combining configurations are provided. At critical times they would be used in parallel. The basic function of the NRT capability is for backup or fail-soft protection to the RT microwave link failure node. The basic technical specification on the NRT configuration is to provide records from Parkes and Tidbinbilla which can be played into the RTC with degradation  $\leq 0.2$  dB. That is the same performance specification as for the RT link.

The DSN is developing a recording and playback system to meet the specification in a station environment. It uses the Mark III data acquisition terminals (DATS) with special signal processing electronics to accommodate the weak telemetry signals. The Mark III DATS is used by the DSN and many radio observatories for VLBI recording.

DSN engineers have had experience in the development and application of sophisticated recording systems (e.g., VLBI

recording and playback and the recording and playback systems for the Pioneer Venus multiprobe and DLBI wind experiment support), there is high confidence in the NRT telemetry system development. A software simulation of the output spectrum reconstruction electronics has been successfully run. That is an important accomplishment. The recording system will be demonstrated in the fall of 1983. Output combining of two recorder systems through an RTC will be demonstrated in the spring of 1984, field demonstration at Goldstone will be in the summer of 1984.

The plan to include Parkes with the Tidbinbilla array presents new technical and managerial challenges to the DSN. Compared to the Goldstone DSS 14-12 configuration, Parkes is truly remote from the Tidbinbilla complex. Its expected contribution to the array performance is next in significance to the DSN 64 m. Its availability, as currently predicted, will be significantly less than for a DSN array element, so that its proper operational use is expected to be more difficult. The use of Parkes will be the DSN's first experience with a non-NASA owned facility tightly coupled

with critical mission support. Hence, the operational procedures and interfaces will be relatively immature compared to those that have been brought to bear on past arrayed mission support operations.

The GIOTTO Project will use the Parkes facility for prime support of its mission, and it will use the DSN Tidbinbilla complex for backup support. The GIOTTO spacecraft encounter with Halley's Comet occurs in mid-March 1986, following the Voyager 2 Uranus encounter by less than two months. This is perceived as an advantage (perhaps an enabler due to shared facility and equipment use at Parkes), and at the same time, an operational complication in the use of Parkes for Voyager support. While there does not appear to be mutual conflict of critical project schedules at Parkes and Tidbinbilla during late 1985 and early 1986, they are interlaced without much contingency during January and February. However, if the two spacecraft and the ground systems behave reasonably well during that period, the continuous usage of the ground facilities under tight configuration management could benefit both projects and the DSN.

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# Antenna Arraying Performance for Deep Space Telecommunications Systems

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*Antenna arraying will be a crucial Deep Space Network technique in maximizing the science return of planetary and comet encounters in the 1980's. This article develops the equations which describe the total figure of merit for a multiple system of arrayed antennas. An example is given for three Canberra DSN antennas and the Parkes 64-m antenna to be arrayed for the Voyager 2 Uranus flyby.*

## I. Introduction

As the Deep Space Network prepares for the decade of the 1980's, the science return for critical spacecraft encounters such as Voyager 2 at Uranus and Neptune and GIOTTO at Halley's Comet is expected to be significantly enhanced through the use of multiple antenna arraying. Antenna arraying concepts were developed in the 1960's for radio astronomy and were successfully demonstrated for space telecommunications with Pioneer 8 (Ref 1) and during the Mariner 10 Mercury encounter in 1974 (Refs 2 and 3) and the Voyager 2 Saturn encounter in 1981 (Ref 4).

The figure of merit of a single aperture receiving system is given by (Ref 5)<sup>1</sup>

$$M = G_R/T_{op}, K^{-1} \quad (1)$$

<sup>1</sup>For convenience,  $M$  is usually used in computations as shown with units  $K^{-1}$  and in discussion with units of dB where  $M$  (dB) = 10 log  $M$ , it is understood that this is relative to a reference figure of merit of  $1 K^{-1}$ .

where

$G_R$  = receiving antenna gain, ratio

$T_{op}$  = system noise temperature, K

A profile of the figure of merit of the Deep Space Network (DSN) receiving systems from 1960 to 1982 is shown in Fig 1.

The figure of merit required for a receiving system to support a given communication link is (Ref 6)

$$M \geq 4\pi kRLD^2 (E_b/N_o)_T/P_T A_T \quad (2)$$

where

$D$  = distance between the transmitting and receiving antennas, m

$R$  = data bit rate, bps

$A_T$  = effective area of transmitting antenna, m<sup>2</sup>

$L$  = total link losses (includes polarization loss, pointing loss, atmospheric loss, demodulation loss, etc), ratio

$(E_b/N_0)_T$  = threshold value (Ref 5) of  $(E_b/N_0)$  for a given bit error rate (BER), ratio

$P_T$  = transmitter power, W

$k$  = Boltzmann's constant,  $1.3806 \times 10^{-23}$ , J/K

For example, Voyager 2 at Saturn had the parameter values shown in Table 1 for transmitting imaging data. The threshold figure of merit  $M_T$  from Eq (2) is calculated to be 55 dB. This results in a communications link margin of 3 dB (Fig 1). This link margin is based on mean parameter values, adequate planned margin is required to allow for variations in the actual parameter values, particularly the change of  $T_{op}$  as a function of weather.

To meet future requirements for Voyager 2 communications at Uranus and Neptune, it will be necessary to improve the figure of merit with a system of arrayed apertures. This report reviews the equations which describe the figure of merit for a multiple array of antennas.

## II. Antenna Arraying

The figure of merit required for the receiving system can be obtained with an array of antennas each with a separate figure of merit ( $M_i = G_i/T_i$ ). Refs 1, 2, and 3 developed the array performance analysis and optimum combiner strategy in terms of the receiver-detected SNR. The array figure of merit could be deduced from this analysis or derived directly. For completeness, an analysis for  $M$  in terms of  $M_i$  follows.

Assume an array of antennas (Fig 2) where the voltage outputs of the individual channels are weighted by  $\beta_i$  and optimally combined<sup>2</sup> to maximize the output signal-to-noise ratio (SNR). Using subscript 1 for the "reference" channel, summing the coherent signal voltages ( $\sqrt{G_i g_i} \beta_i$ ) and the assumed incoherent<sup>3</sup> noise powers ( $T_i g_i \beta_i^2$ ),

<sup>2</sup>This analysis assumes perfect combining and does not allow for unequal atmospheric losses, combiner losses, differential time delays, and other nonideal effects.

<sup>3</sup>The assumption of incoherent noise cannot be made for that portion of received noise that is due to a hot body in view of all elements of the array.

$$\frac{S(\beta_i)}{S_1} = \frac{\left( \sum_{i=1}^n \sqrt{G_i g_i} \beta_i \right)^2}{G_1 g_1 \beta_1^2} \quad (3)$$

and

$$\frac{N(\beta_i)}{N_1} = \frac{\sum_{i=1}^n T_i g_i \beta_i^2}{T_1 g_1 \beta_1^2} \quad (4)$$

where

$S(\beta_i)/S_1$  = arrayed received signal power as a function of  $\beta_i$ , relative to the reference channel, ratio

$N(\beta_i)/N_1$  = arrayed noise power output as a function of  $\beta_i$ , relative to the reference channel, ratio

$G_i$  = gain of  $i$ th receiving antenna, ratio

$g_i$  = gain of  $i$ th receiver, ratio

$T_i$  = operating system noise temperature of  $i$ th antenna, K

$\beta_i$  = voltage weighting function of  $i$ th channel, ratio

Then the signal-to-noise ratio as a function of  $\beta_i$  is given by

$$SNR(\beta_i) = \frac{(SNR)_1}{M_1} \frac{\left( \sum_{i=1}^n \sqrt{G_i g_i} \beta_i \right)^2}{\sum_{i=1}^n T_i g_i \beta_i^2} \quad (5)$$

The value for  $\beta_i$  to maximize the SNR can now be obtained.

Differentiating with respect to  $\beta_i$  and setting the result equal to zero,<sup>4</sup>

<sup>4</sup>The power weighting is given by

$$(\beta_i/\beta_1)^2 = \left( \frac{T_1}{T_i} \right)^2 \frac{G_i g_i}{G_1 g_1} = \frac{T_1 g_1}{T_i g_i} \frac{(SNR)_i}{(SNR)_1}$$

This can be obtained (Ref 7) in an operational system by first setting each channel receiver gain for equal noise level ( $g_i = g_1 T_1/T_i$ ) and then weighting each channel by an additional factor  $(SNR)_i/(SNR)_1$ . Techniques are available (Ref 8) for monitoring  $(SNR)_i$ .

$$(\beta_i/\beta_1) = (T_1/T_i) \sqrt{G_i/g_i/G_1g_1} \quad (6)$$

and

$$SNR = \frac{(SNR)_1}{M_1} \sum_{i=1}^n M_i \quad (7)$$

Using  $SNR/(SNR)_1 = M/M_1$  results in

$$M = \sum_{i=1}^n M_i \quad (8)$$

Expanding Eq (7) and using  $(SNR)_i/(SNR)_1 = M_i/M_1$ ,

$$SNR = \sum_{i=1}^n (SNR)_i \quad (9)$$

In terms of antenna efficiency,  $\epsilon$  and physical diameter  $D$  [using  $G_i = \epsilon_i(\pi D_i/\lambda)^2$ ],

$$M = (\pi/\lambda)^2 \sum_{i=1}^n \epsilon_i D_i^2 / T_i \quad (10)$$

Relative to the highest performance antenna with figure of merit  $M_1$ , using<sup>5</sup>  $\Delta M_i(\text{dB}) = M_1(\text{dB}) - M_i(\text{dB})$ ,

$$\Delta M(\text{dB}) = 10 \log \sum_{i=1}^n 10^{-\Delta M_i(\text{dB})/10} \quad (11)$$

The following section presents some examples

### III. Discussion

Consider an idealized array of one 64-m antenna and  $N$  34-m antennas with equal antenna efficiencies and system noise temperatures. Figure 3 shows the improvements of  $N$  34-m antennas relative to the single 64-m antenna (upper curve) and relative to the single 64-m antenna plus  $(N - 1)$  34-m antennas (lower curve). This illustrates that adding multiple antennas to an array has a smaller percentage effect as  $N$  grows large.

A successful DSN array configuration was used at the Voyager 1 and 2 Saturn flybys using a 64-m antenna and a 34-m antenna (Table 1 and Fig 1). This resulted in an average measured increase in signal-noise ratio of about 0.6 dB relative to the 64-m antenna only (Ref 4). Assuming 50% efficiency for both antennas and no array losses, the potential improvement was about 1.1 dB (Fig 1). Future refinements of the array techniques should reduce the implied  $\approx 0.5$  dB average array loss.

Finally, consider the optimum improvement at 8.42 GHz of the following array relative to the DSS 43 antenna: DSS 43 (64-m), DSS 42 (34-m,  $G/T = -6.0$  dB relative to DSS 43), DSS 45 (34-m,  $G/T = -4.5$  dB relative to DSS 43), and Parkes (64-m,  $G/T = -1.1$  dB relative to DSS 43). Using Eq (11), the estimated improvement in potential figure of merit is

$$\begin{aligned} \Delta M &\approx 10 \log (1 + 0.25 + 0.35 + 0.78) \text{ dB} \\ &\approx 3.8 \text{ dB} \end{aligned}$$

This is the array configuration and potential performance improvement presently planned for the Voyager 2 1986 Uranus encounter (Refs 9 and 10).

<sup>5</sup> $M(\text{dB}) = 10 \log M$

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**Table 1. Tabulated downlink parameters for Voyager 2 spacecraft, Saturn (August 1981) flyby, X-band (8.42 GHz)**

Parameter	Value
Transmitter power $P_T$ , W	21.3
Spacecraft antenna effective area $A_T$ , m <sup>2</sup>	5.4
Distance $D$ , m	$1.557 \times 10^{12}$
Total link loss $L$ , ratio	1.1
Data rate $R$ , bps	$4.48 \times 10^4$
Threshold signal-to-noise ratio for $5 \times 10^{-3}$ BER ( $E_b/N_0$ ) <sub>T</sub> , ratio	1.8
Threshold figure of merit, calculated from Eq (2) $M_T$ , dB	55

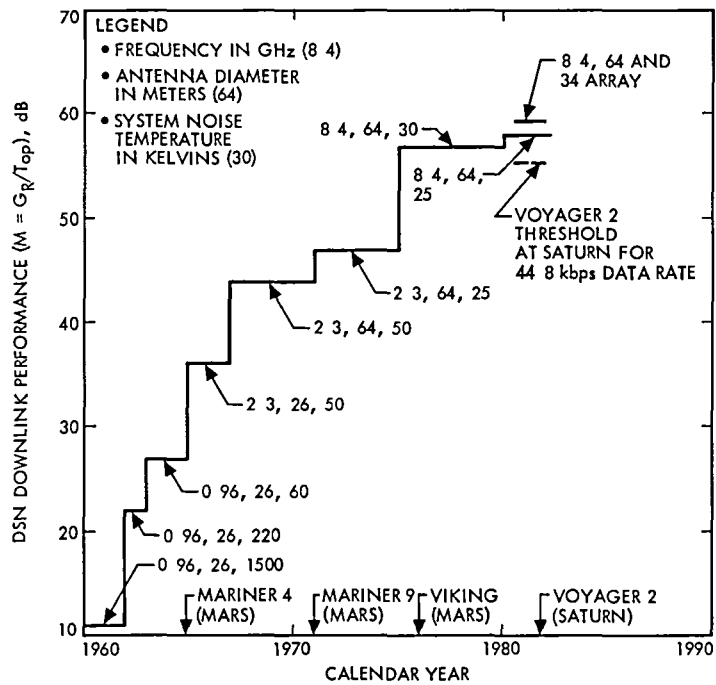


Fig 1 Profile of the DSN downlink performance ( $M = G_R/T_{op}$ ) from 1960 to 1982

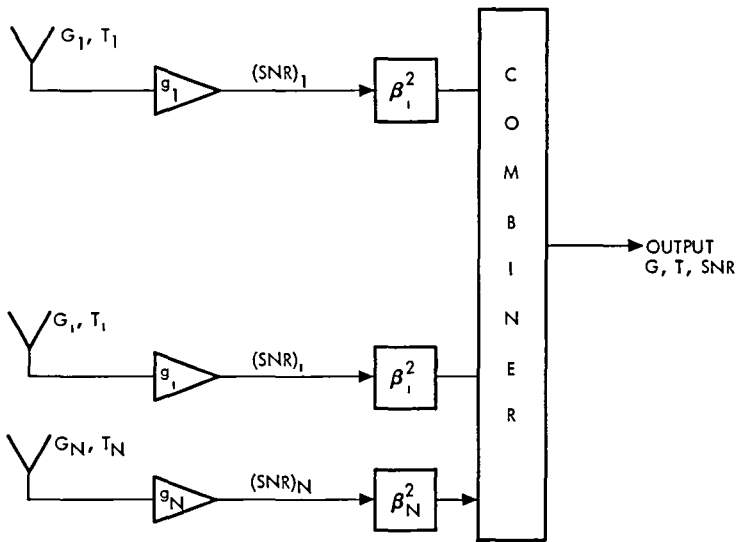


Fig. 2 Antenna array configuration

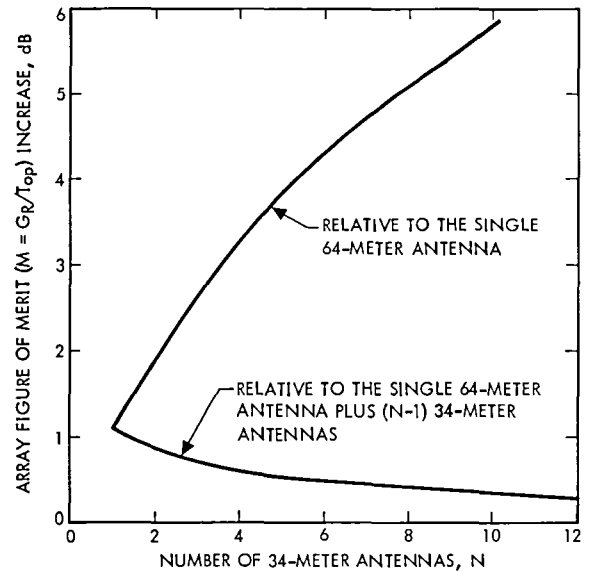


Fig 3 Ideal figure of merit increase of an antenna array consisting of one 64-m antenna and  $N$  34-m antennas



# S-Band Aperture Efficiency of the DSS 13 26-Meter Antenna with the S-X Common Aperture Feed

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*Measurements of the S-band performance of the DSS 13 26-meter antenna at Goldstone, equipped with the second-generation S-X common aperture feed, are reported. Aperture efficiency and overall antenna gain results are compared with predicted values. The elevation dependence of the aperture efficiency is reported.*

## I. Introduction

Precision measurements of astronomical radio sources have been analyzed to evaluate the gain and aperture efficiency of the 26-m antenna at DSS 13. The observations were made during January and February 1982 at 2270 MHz and 2295 MHz with the second generation S-X common aperture feed (Ref 1), which was installed in November 1981. The results have been used to compare the performance of the antenna before and after the installation of the new feed cone. Slight changes in the elevation dependence of the aperture efficiency have been detected and the peak antenna gain was measured.

This recalibration of the DSS 13 antenna was conducted to maintain the high precision capability required for performance evaluations of DSN antennas and for radio astronomy research programs. An example of the latter is the Jupiter Patrol (Ref 2), by which variations in the radio emission from the Jovian radiation belts have been successfully monitored for the past decade. The one-sigma precision of Jupiter's flux-density measurements has been  $\pm 2\%$  ( $\pm 0.1$  dB). The effective gain of the antenna near the elevation of transit for the ISEE-3 spacecraft was also determined, and these

results were used to evaluate carrier power measurements from the spacecraft.

## II. Observations

A selected list of five calibration sources were observed for several hours on January 26, January 29, and February 16, 1982. The observing sessions were planned so that the sources could be measured over a wide range of elevation angles. The intensity of each source was measured several times each night using the drift-curve technique described in Ref 2.

The one-sigma uncertainty of each source-temperature measurement was less than  $\pm 0.1$  dB. To achieve this precision, corrections were applied to account for residual pointing errors and a noise-adding radiometer (NAR) subsystem was used to provide receiver gain stabilization. The feed subsystem was switched to the low-noise listen-only configuration (Ref 1) to optimize the system sensitivity.

The absolute calibration is determined by the flux density scale published by Baars et al (Ref 4). This work is widely

used by the radio astronomy community. The primary reference source is Virgo A (3C274), which unfortunately was not visible during the available observing time. However, the flux densities of the five sources that were observed were calculated from the previously observed ratios of their intensities relative to Virgo A (Refs 3 and 5)

The flux density adapted for Virgo A at 2295 MHz is 1400 Jy. The flux densities of Virgo A and the calibration sources were calculated at 2270 MHz from the fluxes at 2295 MHz and the spectral indices given in Ref 4.

The source list and the flux densities used at both frequencies are given in Table 1. The number,  $N$ , of independent measurements is given in the third column. The correction factor  $C_R$ , listed in the fifth column, is applied to the measurements to account for the loss of intensity due to the angular resolution of the antenna beam, which partially resolves the spatial structure of some radio sources. The effect is small, but not negligible, for the sources included in this report. The source temperatures and the effective aperture efficiency  $\eta$  are discussed below.

### III. Effective Aperture Efficiency

The effective aperture efficiency is equal to the ratio of the observed source temperature and the theoretical source temperature for a 100% efficient antenna,  $T$  ( $\eta = 100\%$ ). The latter is given by the expression

$$T(\eta = 100\%) = \frac{\pi D^2}{4} \frac{S}{2k C_R}, \text{ kelvin} \quad (1)$$

where  $D$  is the antenna diameter,  $S$  is the flux density of the source,  $k = 1.3806 \times 10^{-23} \text{ W s}^{-1} \text{ K}^{-1}$  is the Boltzmann constant and  $C_R$  is the resolution correction factor. For  $D = 25.91 \text{ m}$  (85.0 ft) and flux density  $S$  expressed in Jy ( $1 \text{ Jy} = 10^{-26} \text{ W m}^{-2} \text{ Hz}^{-1}$ )

$$T(\eta = 100\%) = 0.19095 \frac{S}{C_R}, \text{ kelvin} \quad (2)$$

The values of  $T$  ( $\eta = 100\%$ ) are given in the 6th column of the table, adjacent to the observed source temperatures. The calculated effective aperture efficiencies are given in the last column. The weighted averages of  $\eta$  for the two frequencies are also given beneath the entries in the last column. The weighted average of the effective aperture efficiency for all seven entries is  $0.594 \pm 0.002$ . The uncertainty is the one-sigma standard error calculated from the scatter in the measurements.

The measured efficiency is the effective aperture efficiency because the data are affected by systematic attenuation in the clear atmosphere as well as certain losses in the feed system. With the NAR, all system temperatures are calibrated with respect to the temperature of the NAR reference diode, which is injected into the system via a waveguide coupler at the input to the maser. Consequently, the radio source temperatures are referenced to the waveguide flange at the maser input.

## IV. Elevation Dependence of Aperture Efficiency and Gain

### A. Experimental Results

The individual data that resulted in the average results given in Table 1 are plotted in Fig 1 as a function of source elevation angle. The curve represents an empirically determined function of the elevation dependence. The equation is given by

$$\eta(EI) = \eta(37^\circ) A e^{-\tau \sec Z} [(1 - a) + a \cos(EI - 30^\circ)] \quad (3)$$

where  $\eta(EI)$  is the effective aperture efficiency at elevation  $EI$ ,  $A$  is a normalization factor,  $\tau = 0.007$  is the atmospheric loss factor at S-band frequencies,  $Z$  is the zenith angle of the source ( $Z = 90^\circ - EI$ ). Note that  $\tau$  is the atmospheric attenuation in nepers at zenith and that  $1 \text{ neper} \approx 4.343 \text{ dB}$ . The coefficient  $a = 0.098$  was empirically determined from the data in Fig 1.

Written in this form, Eq (3) is the expression for the elevation dependence of the effective efficiency as observed by an experimenter. The equation is normalized to  $EI = 37^\circ$  to match the observed peak in the efficiency measurements. The factor  $A = 1.0110$  normalizes the curve to unity at  $EI = 37^\circ$ . The last factor  $[(1 - a) + a \cos(EI - 30^\circ)]$  represents the loss in efficiency due to gravity loading. The measured symmetry about  $EI = 30^\circ$  suggests that the peak efficiency of the antenna should occur near  $30^\circ$  elevation when atmospheric losses are removed.

The average values of the effective aperture efficiency for each source (Table 1) are the average values of  $\eta(37^\circ)$ , which were found by solving Eq (3) for each measurement. The effective aperture efficiency was converted to antenna gain for a frequency of 2295 MHz. The equivalent antenna gain scale (in dB) is plotted along the right-hand ordinate. Using this scale one can see in Fig 1 that the peak antenna gain, in the presence of atmospheric attenuation, occurs near  $37^\circ$  elevation. The maximum value of the antenna gain, referred to the reference flange at the input to the maser, is  $53.65 \pm$

0.16 dB at 2295 MHz. The corresponding peak gain at 2270 MHz, not shown in the figure, is  $53.56 \pm 0.16$  dB.

The uncertainties quoted for the peak gain results are the one-sigma error estimates for the two frequencies. The error budget includes the standard error on the measured values of  $\eta$  ( $\pm 1\%$ ), the absolute uncertainty of the flux-density calibration scale ( $\pm 3\%$ ), and the systematic error of the NAR temperature calibration ( $\pm 2.0\%$ ). If these errors are uncorrelated, the resultant error is  $\pm 3.7\%$ , or  $\pm 0.16$  dB. The peak effective aperture efficiency, with its total one-sigma error estimate, is  $\eta(37^\circ) = 0.594 \pm 0.022$ .

## B. Comparison with Calculated Predictions

In order to compare the effective aperture efficiency reported in this article with efficiency calculations performed for the S-X common aperture feed, corrections must be applied to account for attenuation through the clear atmosphere at Goldstone and for attenuation between the input flange of the maser and the aperture of the feed. The first step is to write Eq (3) for  $EI = 30^\circ$  (the peak in the antenna gain curve without atmospheric losses)

$$\eta(30^\circ) = \eta(37^\circ) A e^{-\tau \sec(90^\circ - 30^\circ)} \quad (4)$$

Next, multiply by  $e^{+\tau \sec(90^\circ - 30^\circ)}$  to remove atmospheric attenuation and divide by  $f_L$  (the waveguide loss factor)

$$\eta^*(30^\circ) = (A/f_L) \eta(37^\circ) \quad (5)$$

where  $\eta^*(30^\circ)$  is the maximum aperture efficiency for this antenna referred to the phase center of the common aperture feed. From J. Withington (private communication)  $f_L = 0.98 \pm 0.002$  ( $-0.09$  dB) is the loss factor in the waveguide that includes two switches, three bends, and a rejection filter at the X-band transmitter fundamental frequency. With  $A = 1.0110$ , the solution to Eq (5) is

$$\eta^*(30^\circ) = 1.0316 \eta(37^\circ) = 0.613 \pm 0.022 \quad (6)$$

$\eta^*(30^\circ)$  can be compared with calculations of the overall aperture efficiency for this feed, prepared by Williams and Reilly (Ref 6). Their result is 0.578.

A review (private communication, D. A. Bathker) of the predicted aperture efficiency has been done. Three factors

have been identified as principal sources of uncertainty in the predictions. The largest uncertainty is the spar (or quadripod) blockage, which was assumed to consist of the optical shadow area inflated by a factor of 1.2 to account for diffraction at S-band wavelengths. A likely  $3\sigma$  tolerance on the 0.861 spar blockage factor used in the predictions is  $^{+0.04}_{-0.00}$ . The next largest uncertainty is the forward spillover for this dual-band feed, which is higher than older designs at 1.1%. A likely  $3\sigma$  tolerance on this prediction is  $^{+0.02}_{-0.00}$ .

Finally, the predictions accepted an estimate of 1.5 mm (0.06 in.) rms for the surface tolerance of both reflectors. This is conservative, and results in a prediction of 0.979 surface efficiency, a likely  $3\sigma$  tolerance is  $^{+0.01}_{-0.00}$ . The net (conservative)  $3\sigma$  tolerance on the overall predicted efficiency is thus  $^{+0.07}_{-0.00}$ , the one-sigma tolerance is  $^{+0.023}_{-0.000}$ .

The results of the measurements reported here and the predicted values discussed above are summarized in Table 2. The ratio of the measured and predicted values of antenna-design aperture efficiency at S-band is  $(0.613/0.578) = 1.0606$ . The measured efficiency is 6.1% (0.25 dB) greater than that predicted from component evaluation calculations. Although this difference is not statistically significant when the one-sigma error estimates are considered, the difference may be large enough to warrant additional investigation.

## V. Discussion and Conclusions

The elevation dependence of the S-band aperture efficiency and the peak gain of the DSS 13 antenna appear to have changed only slightly with the installation of the S-X common aperture feed. The peak efficiency now occurs near  $37^\circ$  elevation, where the effective efficiency is 59.4%. Similar calibration data taken on several occasions between 1972 and 1979 yielded curves with maxima near  $50^\circ$  elevation and peak effective efficiencies of 60.5%. Independent measurements by Freiley, Batelaan, and Bathker (Ref 5) gave consistent results with a peak effective efficiency of 60.6%.

The shift in the efficiency curve toward lower elevation angles is probably caused by a slight change in focus. The Z-axis subreflector focus was measured soon after the new feed was installed. A proper S-band focus was found when the Z-axis readout, which is a relative measure of distance between the subreflector and the feed, was set at 350 (arbitrary units). The previous value of the S-band focus was 330.

## Acknowledgments

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**Table 1. DSS 13 antenna aperture efficiency, common aperture feed: 1982 January and February data (calibrated with noise adding radiometer)**

Source designation		N	Flux density, Jy	$C_R$ (26-m ant)	Source temperature,		$\eta$ (EI = 37°)
(IAU)	(Other)				T( $\eta = 100\%$ ), K	T(observed), K	
2295 MHz							
0433 + 295	3C123	10	31.94	1 000	6 10	3.61 ± 0.01	0 592
1648 + 05	3C348	8	28 10	1 006	5 33	3.22 ± 0 02	0 604
1957 + 40	Cygnus A	19	915 00	1 009	173.20	103.30 ± 0.02	0 596 (0 597) ± 0.003
2270 MHz							
0433 + 295	3C123	2	32 24	1.000	6.6	3 61 ± 0.17	0.586
0134 + 329	3C48	3	10 80	1.000	2 06	1.23 ± 0 06	0.596
1717 - 00	3C353	2	40.64	1 010	7.68	4 50 ± 0 06	0 586
1957 + 40	Cygnus A	6	927.50	1 009	175 00	105.00 ± 0 40	0.598 (0 592) ± 0 004

**Table 2. DSS 13 aperture efficiency S-band results with S-X common aperture feed**

Aperture efficiency type	Peak efficiency ( $\pm 1\sigma$ )	Elevation of peak	Reference plane	Notes
Effective, measured	$\eta = 0 594 \pm 0 022$	37°	Maser input	User effective aperture includes atmospheric and waveguide losses
Antenna design, inferred from measurements	$\eta^* = 0 613 \pm 0 022$	30°	Phase center of feed	Does not include atmospheric or waveguide losses
Antenna design, predicted <sup>1</sup>	$\eta^* = 0 578 \begin{smallmatrix} +0 023 \\ -0 000 \end{smallmatrix}$	—	Phase center of feed	Does not include atmospheric or waveguide losses

<sup>1</sup>Includes feed pattern integration, subreflector and spar blockage, and surface roughness (Ref 6, and D A Bathker, private communication)

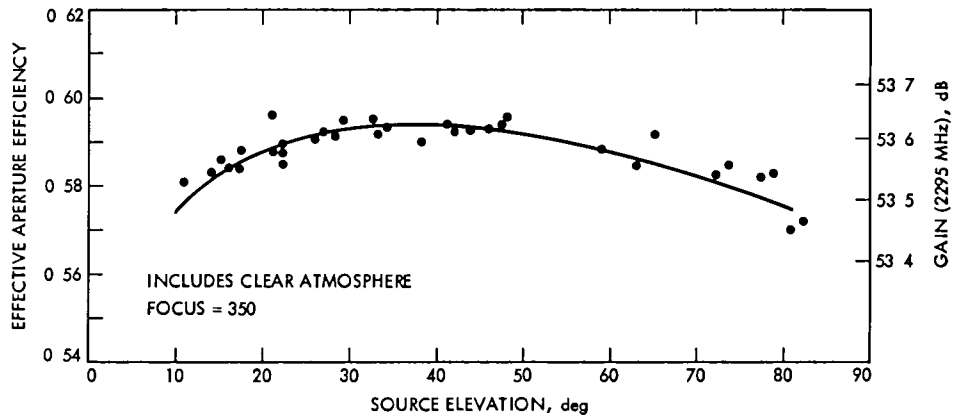


Fig 1. Elevation dependence of aperture efficiency of the DSS 13 26-m antenna, with S-X common aperture feed, 2295 MHz (1982 January and February data)

# The Effect of Partial Coherence In Receiving System Noise Temperature on Array Gain for Telemetry and Radio Frequency Carrier Reception for Receiving Systems with Unequal Predetection Signal-to-Noise Ratios

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*Signal-to-noise ratio improvement realized by arraying receiving systems or stations for coherent reception is reduced when some portion of the predetection noise is coherent in the array. This article examines this effect for arrayed receiving systems with unequal apertures including expected performance for selected planets of the solar system within the array beamwidth.*

## I. Introduction

An earlier report (Ref 1) considered the effect of partial coherence in system noise temperature on predetection signal-to-noise ratio improvement and equivalent radio frequency (RF) carrier signal-to-noise ratio improvement for coherent reception and demodulation for an array of similar receiving systems with essentially equal predetection noise and equal size antenna apertures. This report examines the effect of partial coherence in system noise temperature on predetection signal-to-noise ratio improvement and RF carrier margin improvement for coherent reception and demodulation for an array of receiving systems with unequal predetection noise and unequal antenna apertures. The analytic expressions given represent the general case for the configuration examined here.

Performance is presented in this report for the case which is representative of 34-meter-diameter antenna receiving systems arrayed with a 64-meter-diameter antenna receiving system which also has transmit capability. For the performance shown here, the coherent portion of predetection noise (which is coherent periodically) results from a planet in the solar system which is within the beamwidth of the antennas of the array. RF carrier array gain measurements made in the Laboratory for simulated 64-meter/34-meter antenna receiving system arrays are compared with theoretical results.

This report also provides information on the expected periodic reduction in array gain at a reception frequency of 8420 MHz due to antenna feed spacing with the planets Jupiter or Venus in the array beamwidth. The expected periodic reduction in array gain with these planets in the array beamwidth is less than 0.2 dB for a two-aperture array and less than 0.25 dB for a three-aperture array for the antenna spacings and orientation planned for the 64- and 34-meter antennas at the three DSN complexes. Expected reduction in array gain with the planets Uranus or Neptune in the array beamwidth is negligible.

## II. Predetection Signal-to-Noise Ratio and Resultant Carrier Tracking Loop Phase Noise

Figure 1 illustrates the configuration considered herein which provides predetection signal-to-noise ratio improvement and equivalent RF carrier signal-to-noise ratio improvement (RF carrier margin improvement) for high rate telemetry with residual RF carrier. The RF carrier margin improvement also provides a corresponding improvement in minimum RF carrier level for radio metric tracking. Figure 1 and a modification of Fig. 1 (so that much larger antenna separations for the array can be handled conveniently) were presented in Ref. 1 with a discussion of predetection noise resulting from operating equivalent system noise temperature  $T_{op}$ .

With receiving system 2 connected to the summing junction (see Fig. 1), the RF carrier predetection signal-to-noise ratio in receiving system 1 at the output of the summing junction is (from Expression 3, Ref. 1)

$$\frac{P_{c1\Sigma 12}}{P_{n1\Sigma 12}} = \frac{(A_1 \cos m_{pd} + \beta_2 A_2 \cos m_{pd})^2}{\left[ NBW_{F_{A1}} (1 - \epsilon_1) N_{o1} + \beta_2^2 NBW_{F_{A2}} (1 - \epsilon_2) N_{o2} \right] + \left[ (NBW_{F_{A1}} \cdot \epsilon_1 N_{o1})^{1/2} + \beta_2 (NBW_{F_{A2}} \cdot \epsilon_2 N_{o2})^{1/2} \right]^2} \quad (1)$$

where  $NBW_{F_{A1}}$  represents the noise bandwidth of second IF filter  $F_{A1}$ ,  $NBW_{F_{A2}}$  is equal to  $NBW_{F_{A1}}$ ,  $N_{o1}$  is the one-sided noise spectral density for receiving system 1 related to  $T_{op1}$ , and  $N_{o2}$  is the noise spectral density of receiving system 2 related to  $T_{op2}$ . The term  $\beta_2$  is the voltage coupling of receiving system 2 relative to receiving system 1 at the summing junction, and  $\gamma_2^2$  is the carrier power-to-noise spectral density ratio of receiving system 2 relative to receiving system 1. The statistically independent portion of predetection noise is  $(1 - \epsilon_1)$  in receiving system 1 and  $(1 - \epsilon_2)$  in receiving system 2. The corresponding portion of predetection noise that is coherent is  $\epsilon_1$  and  $\epsilon_2$  in the two receiving systems. Expression 1 can be rewritten as

$$\frac{P_{c1\Sigma 12}}{P_{n1\Sigma 12}} = \frac{P_{c1}}{NBW_{F_{A1}} \cdot N_{o1}} \cdot \frac{(1 + \beta_2 \gamma_2)^2}{\left[ (1 - \epsilon_1) + (1 - \epsilon_2) \frac{N_{o2}}{N_{o1}} \beta_2^2 \right] + \left[ \epsilon_1^{1/2} + \epsilon_2^{1/2} \left( \frac{N_{o2}}{N_{o1}} \right)^{1/2} \beta_2 \right]^2} \quad (2)$$

Note that (from Ref. 1) the predetection carrier signal-to-noise ratio in receiving system 1 (with receiving system 2 switched out of the summing junction) is  $P_{c1}/(NBW_{F_{A1}} \cdot N_{o1})$ . Consequently, the improvement in predetection signal-to-noise ratio for two receiving systems arrayed ( $n_2''$ ) is

$$n_2'' = \frac{(1 + \beta_2 \gamma_2)^2}{\left[ (1 - \epsilon_1) + (1 - \epsilon_2) \frac{N_{o2}}{N_{o1}} \beta_2^2 \right] + \left[ \epsilon_1^{1/2} + \epsilon_2^{1/2} \left( \frac{N_{o2}}{N_{o1}} \right)^{1/2} \beta_2 \right]^2} \quad (3)$$



For  $N$  receiving systems arrayed, the predetection carrier signal-to-noise ratio in receiving system 1 becomes

$$\frac{P_{c1\Sigma 1, \dots, N}}{P_{n1\Sigma 1, \dots, N}} = \frac{P_{c1}}{NBW_{F_{A1}} \cdot N_{o1}} \cdot \frac{(1 + \beta_2 \gamma_2 + \beta_N \gamma_N)^2}{\left[ (1 - \epsilon_1) + (1 - \epsilon_2) \frac{N_{o2}}{N_{o1}} \beta_2^2 + \dots + (1 - \epsilon_N) \frac{N_{oN}}{N_{o1}} \beta_N^2 \right] + \left[ \epsilon_1^{1/2} + \epsilon_2^{1/2} \left( \frac{N_{o2}}{N_{o1}} \right)^{1/2} \beta_2 + \dots + \epsilon_N^{1/2} \left( \frac{N_{oN}}{N_{o1}} \right)^{1/2} \beta_N \right]^2} \quad (4)$$

The improvement in predetection signal-to-noise ratio for  $N$  receiving systems arrayed is then

$$n_N'' = \frac{(1 + \beta_2 \gamma_2 + \beta_N \gamma_N)^2}{\left[ (1 - \epsilon_1) + (1 - \epsilon_2) \frac{N_{o2}}{N_{o1}} \beta_2^2 + \dots + (1 - \epsilon_N) \frac{N_{oN}}{N_{o1}} \beta_N^2 \right] + \left[ \epsilon_1^{1/2} + \epsilon_2^{1/2} \left( \frac{N_{o2}}{N_{o1}} \right)^{1/2} \beta_2 + \dots + \epsilon_N^{1/2} \left( \frac{N_{oN}}{N_{o1}} \right)^{1/2} \beta_N \right]^2} \quad (5)$$

It should be noted (as discussed in Ref 1) that for the condition where the varying group delay on the telemetry sidebands is tracked out among the various receiving systems of the array (which is accomplished at baseband for the configuration shown in Fig 1), the improvement in telemetry predetection signal-to-noise ratio that can be obtained is also represented by expressions (3) and (5) for the corresponding voltage coupling  $\beta$  in the signal spectrum combining process and ratio of signal-to-noise spectral density ratios  $\gamma^2$  for telemetry. However, the predetection signal-to-noise ratio improvement obtained for telemetry does not include loss in the telemetry signal spectrum combining process and the loss (radio loss) due to carrier tracking loop phase noise.

The resultant rms phase noise ( $\sigma_{\phi_{n1\Sigma 1,2}}$ ) at the output of the RF carrier tracking loop (i.e., on the first local oscillator signal) in receiving system 1 due to the predetection signal-to-noise ratio within the closed-loop noise bandwidth of the RF carrier tracking loop becomes (see expression 7, Ref 1)

$$\sigma_{\phi_{n1\Sigma 1,2}} = \frac{N_{o1}}{2} \cdot 2B_{L1} \cdot \frac{1}{P_{c1}} \cdot \frac{1}{n_2''} \left[ \frac{1 + \frac{P_{c1} n_2''}{NBW_{F_{A1}} \cdot N_{o1}} \cdot \exp\left(\frac{N_{o1} \cdot B_{L1}}{P_{c1} \cdot \eta_2''}\right)}{0.862 + \frac{P_{c1} \cdot \eta_2''}{NBW_{F_{A1}} \cdot N_{o1}}} \cdot \frac{\sinh\left(\frac{N_{o1} \cdot B_{L1}}{P_{c1} \cdot \eta_2''}\right)}{\sinh\left(\frac{N_{o1} \cdot B_{L1}}{P_{c1} \cdot \eta_2''}\right)} \right]^{1/2} \text{ rad, rms} \quad (6)$$

for two systems arrayed where  $n_2''$  is defined in expression (3) above. For  $N$  receiving systems arrayed, the resultant rms phase noise ( $\sigma_{\phi_{n1\Sigma 1, \dots, N}}$ ) becomes

$$\sigma_{\phi_{n1\Sigma 1, \dots, N}} = \frac{N_{oN}}{2} \cdot B_{L1} \cdot \frac{1}{P_{c1}} \cdot \frac{1}{n_N''} \left[ \frac{1 + \frac{P_{c1} \cdot n_N''}{NBW_{F_{A1}} \cdot N_{o1}} \cdot \exp\left(\frac{N_{o1} \cdot B_{L1}}{P_{c1} \cdot n_N''}\right)}{0.862 + \frac{P_{c1} \cdot n_N''}{NBW_{F_{A1}} \cdot N_{o1}}} \cdot \frac{\sinh\left(\frac{N_{o1} \cdot B_{L1}}{P_{c1} \cdot n_N''}\right)}{\sinh\left(\frac{N_{o1} \cdot B_{L1}}{P_{c1} \cdot n_N''}\right)} \right]^{1/2} \text{ rad, rms} \quad (7)$$

where  $n_N''$  is defined in expression (5) above. The rms phase noise  $\sigma_{\phi_{n1}}''$  for receiving system 1 with receiving systems 2 through  $N$  switched out of the summing junction is shown in Ref 1 (expression 2)

Note that the total phase noise at the output of the principal carrier tracking loop (i.e., on the first local oscillator) becomes

$$\left[ \sigma_{\phi_{n1\Sigma 1,2}}^2 + \left( \frac{\beta_2 \sigma_{\phi_{n2}}}{1 + \beta_2} \right)^2 \right]^{1/2} \quad (8)$$

for two receiving systems arrayed (see Ref 1) For  $N$  receiving systems arrayed, the total rms phase noise on the first local oscillator is

$$\left[ \sigma_{\phi_{n1\Sigma 1, \dots, N}}^2 + \left( \frac{\beta_2 \sigma_{\phi_{n2}}}{1 + \beta_2} \right)^2 + \dots + \left( \frac{\beta_N \sigma_{\phi_{nN}}}{1 + \beta_N} \right)^2 \right]^{1/2} \quad (9)$$

Expressions for rms phase noise in systems 2 through  $N$  were developed in Ref 1

As developed in Ref 1, the total rms phase noise in expressions (8) and (9) can be considered as due to an equivalent carrier predetection signal-to-noise ratio within the closed-loop noise bandwidth of the RF carrier phase tracking loop Comparison of this equivalent carrier signal-to-noise ratio with the initial carrier predetection signal-to-noise ratio in a single receiving system (i.e., system 1) alone provides the improvement in carrier margin due to radio frequency carrier arraying for the high rate telemetry configuration where some portion of the predetection noise is not statistically independent among the various receiving systems of the array

### III. Performance

Expressions (3) and (5) represent improvement in predetection signal-to-noise ratio in receiving system 1 under the condition that the phase shift and group delay in the various receiving systems of the array are essentially matched in a predetection bandwidth that encompasses the signal of interest In addition for telemetry (as described in Ref 1), expressions (3) and (5) apply for the additional condition that the varying group delay on the signal modulation sidebands is tracked out among the receiving systems of the array by a telemetry signal spectrum combiner during a station pass Note again that the predetection signal-to-noise ratio improvement for telemetry does not include the telemetry signal combiner loss and the loss (radio loss) due to carrier tracking loop phase noise Reference 2 provides information on radio loss

Improvement in receiving system 1 predetection signal-to-noise ratio is shown in Fig 2 as a function of the statistically independent (noncoherent) portion of predetection noise ( $1 - \epsilon_1$ ) in receiving system 1 for an array consisting of a 64-meter-diameter antenna with a 50% aperture efficiency (system 1) and a 34-meter-diameter antenna listen-only, with a 60% aperture efficiency As pointed out in Section I, for the performance shown in this report, the coherent portion of predetection noise results from a planet in the solar system which is within the beamwidth of the antennas of the array Consequently, the coherent noise for the 34-meter-diameter listen antenna is scaled relative to the 64-meter-diameter antenna by  $(34/64)^2 \times (0.60/0.50)$  Performance is shown for  $\gamma_2$  values of 0.55, 0.61, and 0.66 for the case where  $\beta_2 = \gamma_2$  As developed earlier in this report,  $\gamma_2^2$  is the signal-power-to-noise-spectral-density ratio of receiving system 2 relative to the receiving system 1, and  $\beta_2$  is the voltage coupling of receiving system 2 relative to receiving system 1 for the combining process Improvement in predetection signal-to-noise ratio is maximum for  $\beta = \gamma$  when the predetection noise is all statistically independent or  $1 - \epsilon_1 = 1$ , which is representative for the DSN most of the time A  $\gamma_2$  value of 0.61 represents the nominal value for this array and provides a 1.46-dB improvement in predetection signal-to-noise ratio at  $\beta = \gamma$  for  $1 - \epsilon_1 = 1$  A  $\gamma_2$  value of 0.66 represents operation of the antennas at low (10-15 deg) and high (approximately 80 deg) elevation angles, whereas a  $\gamma_2$  value of 0.55 represents a below nominal signal-to-noise ratio for the 34-m antenna (listen-only) The ratio of noise spectral densities ( $N_{o2}/N_{o1}$ ) used in Fig 2 for noncoherent predetection noise ( $1 - \epsilon_1 = 1$ ) is 0.925 (18.5K/20.0K) A  $N_{o2}/N_{o1}$  value of 1 for  $1 - \epsilon_1 = 1$  reduces the improvement in predetection signal-to-noise ratio shown in Fig 2 by about 0.1 dB and less than 0.1 dB for  $1 - \epsilon_1$  less than 1<sup>1</sup> The resultant decrease in predetection signal-to-noise ratio improvement is shown in Fig 2 as an increasing portion of predetection noise becomes coherent

<sup>1</sup>A  $N_{o2}/N_{o1}$  of 0.925 (18.5/20.0) represents operation at high elevation angles, while at low elevation angles,  $N_{o2}/N_{o1}$  approaches a value of 1

( $\epsilon_1$ ) or as the statistically independent portion of predetection noise ( $1 - \epsilon_1$ ) decrease for receiving system 1. An increase in aperture efficiency for the 34-meter-diameter listen antenna to 65% reduces the improvement in predetection signal-to-noise ratio shown in Fig 2 by about 0.1 dB at  $1 - \epsilon_1 = 0.3$  and has no effect when  $1 - \epsilon_1 = 1$  for a given  $\gamma_2$  value.

Figure 3 shows receiving system 1 predetection signal-to-noise ratio improvement as a function of the statistically independent portion of system 1 predetection noise ( $1 - \epsilon_1$ ) for an array of two receiving systems utilizing a 64-meter-diameter antenna and a 34-meter-diameter antenna (transmit/receive) (with a 54% aperture efficiency), for  $\gamma_2$  values of 0.48, 0.53, and 0.58. Coherent noise for the 34-meter-diameter transmit/receive antenna is scaled relative to the 64-meter-diameter antenna by  $(34/64)^2 \times (0.54/0.50)$ . A nominal  $\gamma_2$  value of 0.53 provides a 1.0-dB improvement in predetection signal-to-noise ratio for  $1 - \epsilon_1 = 1$ . The preceding discussion for  $\gamma_2$  values of 0.55, 0.61, and 0.66 for Fig 2 applies to the  $\gamma_2$  values of 0.48, 0.53, and 0.58 respectively for Fig 3. The ratio of noise spectral densities ( $N_{o2}/N_{o1}$ ) is 1.075 (21.5K/20.0K) for noncoherent predetection noise ( $1 - \epsilon_1 = 1$ ) in Fig 3. In this case, a  $N_{o2}/N_{o1}$  value of 1 for  $1 - \epsilon_1 = 1$  increases the improvement in predetection signal-to-noise ratio shown in Fig 3 by 0.1 dB (or less). A decrease in aperture efficiency for the 34-meter-diameter transmit/receive antenna to 50% increases the improvement in predetection signal-to-noise ratio shown in Fig 3 by about 0.1 dB at  $1 - \epsilon_1 = 0.3$  and has no effect when  $1 - \epsilon_1 = 1$  for a given  $\gamma_2$  value.

Figures 4 and 5 show improvement in receiving system 1 predetection signal-to-noise ratio for an array of three receiving systems as a function of the statistical independent portion of predetection noise ( $1 - \epsilon_1$ ) in receiving system 1 for the case where  $\beta_2 = \beta_3 = \gamma_2$ . Figure 4 represents an array utilizing a 64-meter-diameter antenna (system 1), a 34-meter-diameter antenna listen-only (system 2) and a 34-meter-diameter antenna with transmit/receive capability (system 3). Nominal values of  $\gamma_2 = 0.61$  and  $\gamma_3 = 0.53$  provide a 2.17-dB improvement in predetection signal-to-noise ratio for  $1 - \epsilon_1 = 1$  (Fig 4). For this figure, the ratio of noise spectral densities  $N_{o2}/N_{o1}$  and  $N_{o3}/N_{o1}$  are 0.925 and 1.075 respectively at  $1 - \epsilon_1 = 1$ . A value of 1 for  $N_{o2}/N_{o1}$  and  $N_{o3}/N_{o1}$  provides essentially the same performance as shown in Fig 4. Figure 4 represents performance for aperture efficiencies of 50, 60, and 54% for receiving systems 1, 2, and 3 respectively for the three sets of values for  $\gamma_2$  and  $\gamma_3$  shown in Fig 4. An increase in aperture efficiency to 65% for the 34-meter-diameter listen antenna and a decrease in aperture efficiency to 50% for the 34-meter-diameter transmit/receive antenna reduce the improvement in predetection signal-to-noise ratio shown in Fig 4 by about 0.05 dB at  $1 - \epsilon_1 = 0.3$  and has no effect when  $1 - \epsilon_1 = 1$  for a given set of  $\gamma_2$  and  $\gamma_3$  values. Figure 5 represents an array utilizing a 64-meter-diameter antenna (system 1) and two 34-meter-diameter antennas listen-only (systems 2 and 3) with 60% aperture efficiencies. Nominal values of  $\gamma_2 = \gamma_3 = 0.61$  provide a 2.55-dB improvement in predetection signal-to-noise ratio for  $1 - \epsilon_1 = 1$  and  $N_{o2}/N_{o1} = N_{o3}/N_{o1} = 0.925$ . A value for 1 for  $N_{o2}/N_{o1}$  and  $N_{o3}/N_{o1}$  decreases the improvement in predetection signal-to-noise ratio shown in Fig 5 by 0.15 dB (or less). An increase in aperture efficiency to 65% for systems 2 and 3 reduces the predetection signal-to-noise ratio improvement shown in Fig 5 by about 0.2 dB at  $1 - \epsilon_1 = 0.3$  and has no effect when  $1 - \epsilon_1 = 1$  for a given set of  $\gamma_2$  and  $\gamma_3$  values.

Figures 6, 7, 8, and 9 show the effect on predetection signal-to-noise ratio improvement as the voltage coupling  $\beta$  in the signal combining or summing process is varied for two and three receiving systems arrayed. The array parameters for Figs 6, 7, 8, and 9 are as described above for Figs 2, 3, 4, and 5 respectively. Also shown in Figs 6, 7, 8, and 9 is the effect of predetection signal-to-noise ratio improvement when the portion of predetection noise that is statistically independent ( $1 - \epsilon_1$ ) is 1.0, 0.7, or 0.3 in receiving system 1.

Consider next the equivalent RF carrier predetection signal-to-noise ratio improvement (RF carrier margin improvement) as determined from the total rms phase noise on the first local oscillator (see expressions (6), (7), (8), and (9) and associated discussion in Section II). The following sets of design parameters apply for the performance presented in this report. The sets of parameters for receiving system 1 are

Threshold two-sided noise bandwidth			
$2B_{L_{o1}}$	12	152	30 Hz
Predetection IF filter noise bandwidth			
$NBW_{FA1}$	2200	2200	2000 Hz

while the corresponding sets of parameters for receiving system 2 through  $N$  are

Threshold two-sided noise bandwidth $2B_{L_{o2, \dots, N}}$	0.1	1.0	0.3 Hz
Predetection IF filter noise bandwidth $NBW_{F_{A2, \dots, N}}$	2200	2200	2000 Hz

Using the bandwidth parameters above and the array parameters described in conjunction with Fig 2, Fig 10 shows the RF carrier margin improvement as a function of summing junction voltage coupling  $\beta$ . Figure 10 represents an array of two receiving systems with a 64-meter-diameter antenna (system 1) and a 34-meter-diameter antenna listen-only (system 2). Performance is shown for  $\gamma_2$  values of 0.55, 0.61, and 0.66 with  $1 - \epsilon_1$  values of 1.0, 0.7, and 0.3 as in Fig 6. The information in Fig 10 can be rearranged to show RF carrier margin improvement as a function of the statistically independent portion of predetection noise in system 1 ( $1 - \epsilon_1$ ). Figure 11 shows this characteristic for a voltage coupling  $\beta_2 = \gamma_2$  for  $\gamma_2$  values of 0.55, 0.61, and 0.66.

Figures 12 and 13 show similar RF carrier margin improvement characteristics for an array of two receiving systems with a 64-meter-diameter antenna and a 34-meter-diameter antenna, transmit/receive. Similar characteristics are shown in Figure 14 and 15 for an array of three receiving systems with a 64-meter-diameter antenna (system 1), a 34-meter-diameter antenna listen-only (system 2) and a 34-meter-diameter antenna (transmit/receive) (system 3). Figures 16 and 17 show RF carrier margin improvement characteristics for an array of three receiving systems with a 64-meter-diameter antenna (system 1) and two 34-meter-diameter antenna (listen-only).

Some initial measurements of RF carrier margin improvement have been made in the laboratory, for two and three receiving systems arrayed, by measurement of rms phase noise ( $\sigma_{\phi_n}$ ) on the first local oscillator signal in system 1 and utilizing this  $\sigma_{\phi_n}$  in conjunction with expressions (6) and (7) in Section II. For these measurements,  $\gamma_2$  and  $\gamma_3$  values of 0.61 were set up in the laboratory to simulate reception with 34-meter-diameter antenna(s) (listen-only) arrayed with a 64-meter-diameter antenna (system 1). Measurements were made using the following set of parameters. The predetection IF filter noise bandwidth was 2200 Hz with a  $2B_{L_o}$  of 152 Hz for receiving system 1 and  $2B_{L_o}$  of 1 Hz for receiving systems 2 and 3.

The statistically independent noise ( $1 - \epsilon$ ) coupled into the input of each receiving system from separate random noise sources was equal so that  $N_{o2}/N_{o1} = N_{o3}/N_{o1} = 1$  for  $1 - \epsilon_1 = 1$ . Coherent noise ( $\epsilon$ ) coupled into the input of the simulated 34-meter receiving systems was scaled by the ratio  $(34/64)^2 \times (0.6/0.5)$  relative to the simulated 64-meter receiving system and adjusted for correlation. The values 0.6 and 0.5 represent the aperture efficiencies, discussed earlier, of the 34-meter-diameter antenna listen-only and the 64-meter-diameter antenna respectively. Table 1 shows the experimental measurements relative to the calculated results.

## IV. Discussion

The situation where a planet in the solar system is within the beamwidth of an array with equal antenna apertures was examined in Ref 1 (Section V). The receiving systems were arrayed for coherent reception of a signal from a spacecraft in deep space. The same situation is considered herein except that the antennas of the array have unequal apertures as presented in the earlier sections of this report.

Consider the case (as in Ref 1) of two receiving systems arrayed for coherent reception of the spacecraft signal. A two-aperture (antenna) array will see the hot body (noise) radiation from the planet as an interferometer operating at a reception frequency wavelength  $\lambda$  with an antenna spacing  $B_{1,2}$ . Designate the component of  $B_{1,2}$  that is perpendicular to the line of sight to the planet as  $B_{r1,2}$ , then the resultant fringe spacing of the interferometer is  $\lambda/B_{r1,2}$  radians.

A large part of the following material was presented in Ref 1 and is included here for clarification and continuity of discussion. Consider for the moment, a situation where the planet appears as a point noise source. That is, the angular size of the planet as seen from Earth is very small compared to the fringe spacing ( $\lambda/B_{r1,2}$ ). Within the fringe spacing (as the Earth rotates), the noise radiation from the planet as seen in the predetection bandwidth of the array is in phase for the two receivers at the center of an angular segment  $\lambda/2B_{r1,2}$ , orthogonal at each edge of the angular segment, and 180 deg out of phase at the center of the adjacent angular segments  $\lambda/2B_{r1,2}$  due to continuing change in path length to the two antennas. This in-phase/out-of-phase situation continues alternately during a station pass as the array tracks the spacecraft and it is superimposed on the statistically independent portion of predetection noise. It should be noted again that the varying group delay on the signal modulation sidebands is tracked out among the receiving systems of the array during a station pass (see Section III, first paragraph). For this situation of a planet appearing as a point noise source and representing a given fractional part ( $\epsilon$ ) of operating equivalent system noise temperature ( $T_{op}$ ), the decrease in predetection signal-to-noise ratio improvement shown in Figs 2 and 3 (for the corresponding  $(1 - \epsilon)$ ) corresponds to the in-phase predetection noise condition at the center of an angular segment  $\lambda/2B_{r1,2}$ . The orthogonal predetection point noise source condition corresponds to a signal-to-noise ratio improvement for  $(1 - \epsilon) = 1$ , while the 180-deg out-of-phase point noise source condition corresponds to a signal-to-noise ratio improvement with a portion of the point source contributions removed (cancelled). The discussion above also applies to RF carrier margin improvement as shown in Figs. 11 and 13 for two systems arrayed. Cross plots of the performance shown in Figs 6 and 7 and in Figs 10 and 12 would provide similar information for other signal summing voltage couplings.

A planet for the solar system may not appear as a point noise source for the array considered here. In fact, it can have an angular size that is larger than the fringe spacing  $\lambda/B_{r1,2}$ . The following development addresses this consideration. In general, for the application presented in this report, the planets (as viewed from Earth) can be characterized as disks with a uniform brightness distribution  $D$  at the reception frequency. The brightness distribution will be a function of the reception frequency. Utilizing information in Ref. 3 and writing the brightness transform as a function of the angular size  $\nu$  of the planet and the inverse of fringe spacing, the brightness transform  $R$  becomes

$$R_{1,2} = \frac{1}{\nu} \int_{-\nu/2}^{+\nu/2} D(\omega) e^{-2\pi i \left( \frac{B_{r1,2}}{\lambda} \right) x} dx \quad (10)$$

with integration from the centerpoint of the planet disk out to the edges. Integration of expression (10) provides

$$R_{1,2} = D(\omega) \cdot \frac{\sin \left( \pi \frac{B_{r1,2}}{\lambda} \nu \right)}{\pi \frac{B_{r1,2}}{\lambda} \nu} \quad (11)$$

For the situation where the planet appears as a point noise source, the  $\sin x/x$  type of expression above approaches 1. The brightness transform becomes the total power of the noise source (planet) as seen in the predetection bandwidth, and the preceding discussion relating to a point noise source applies. Figure 18 shows expression (11) plotted as a function of the ratio of planet angular size to fringe spacing [ $\nu/(\lambda/B_{r1,2})$ ]. Note that the above integral is centered ( $\nu = 0$ ) on the in-phase detection noise situation at the center of an angular segment  $\lambda/2B_{r1,2}$  that was discussed earlier in this section of the report. Figure 18 is the resultant amplitude of the brightness transform for this particular path length situation to the two antennas for various ratios of planet angular size to fringe spacing. This figure shows the magnitude of the correlated portion of noise power relative to total noise power from the planet, which is defined as fringe visibility for radio interferometry measurements (Ref 3). Consequently, Fig 18 provides the information necessary to scale the coherent portion of predetection noise relative to a point noise source for any ratio of planet angular size to fringe spacing [ $\nu/(\lambda/B_{r1,2})$ ]. Any point on Fig 18 can be treated as an equivalent reduced-in-magnitude point noise source representing a reduced  $\epsilon$ , and the resulting reduction in predetection signal-to-noise ratio and RF carrier margin improvement can be determined from Figs 2, 3, 6, 7, 10, 11, 12 and 13.

Consider next three receiving systems (1, 2, and 3) arrayed for coherent reception of a spacecraft signal with a planet within the beamwidth of the three antennas. For the moment, consider a situation where the planet appears as a point noise source which represents a given fractional part  $\epsilon$  of the operating equivalent system noise temperature  $T_{op}$ . At those times when the noise radiation from the planet as seen in the predetection bandwidth of the array is in phase from receivers 1 and 3 in an angular segment  $\lambda/2B_{r1,3}$  at essentially the same time as it is from receivers 1 and 2 in an angular segment  $\lambda/2B_{r1,2}$ , the decrease in predetection signal-to-noise ratio improvement shown in Figs 4 and 5 applies for the corresponding  $(1 - \epsilon)$  value. At all other times, the decrease in predetection signal-to-noise ratio improvement will be less. This discussion also applies to RF carrier margin improvement shown in Figs 15 and 17 for three systems arrayed. Cross plots of the performance shown in Figs 8 and 9 and in Figs 14 and 16 would provide similar information for other signal summing voltage couplings.

For the situation where the planet is not a point source, the normalized brightness transforms for three receiving systems become

$$R_{1,2,3} = \frac{D(\omega)}{2} \left( \frac{\sin \left( \pi \frac{B_{r1,2}}{\lambda} \nu \right)}{\pi \frac{B_{r1,2}}{\lambda} \nu} + \frac{\sin \left( \pi \frac{B_{r1,3}}{\lambda} \nu \right)}{\pi \frac{B_{r1,3}}{\lambda} \nu} \right) \quad (12)$$

Note again in this case that the above expression (12) represents a result centered ( $\nu = 0$ ) on the in-phase predetection noise situation at the center of angular segment  $\lambda/2B_{r1,2}$  and simultaneously at the center of angular segment  $\lambda/2B_{r1,3}$ . For this particular path length situation and given fringe spacings  $\lambda/B_{r1,2}$  and  $\lambda/B_{r1,3}$ , expression (12) can be evaluated and the effect of the planet on decrease in predetection signal-to-noise ratio and RF carrier margin improvement can be determined from Figs 4, 5, 8, 9, 14, 15, 16, and 17.

In order to illustrate the effect of a planet in the solar system on signal-to-noise ratio improvement for an array of receiving systems as described in Sections II and III of this report, consider reception at 8420 MHz ( $\lambda = 3.56$  cm). Table 2 shows nominal radio disk temperatures  $T_R$  of the planets in the solar system at a reception wavelength of 3.55 to 3.6 cm, angular size as seen from Earth, and increase in system noise temperature due to the planet in the antenna beamwidth of the 64-meter-diameter antenna. References 4, 5, 6, 7, and 8 provide the basis for the information shown in Table 2. Increase in system noise temperature  $\Delta T_A$  due to a planet in the antenna beamwidth is obtained (except for Jupiter, due to its radiation belts) from the expression

$$\Delta T_A = T_R \times \frac{\text{planet (source) solid angle}}{\text{antenna beam solid angle}}$$

where the antenna beam solid angle is calculated from the expression  $(4/3)(\text{half-power beamwidth})^2$  (Ref 4). Increase in system noise temperature due to Jupiter is obtained from the normalized flux density at 4.04 AU using the expression

$$\Delta T_A = \frac{\text{Antenna Geometrical Area} \times \text{Aperture Efficiency} \times \text{Flux Density}}{2 \times (\text{Boltzmann's constant})}$$

and scaling  $\Delta T_A$  for minimum and maximum distance from Earth by (ratio of distances)<sup>2</sup> (Ref 8). It is interesting to note that  $\Delta T_A$  for the outer planets Saturn, Uranus, and Neptune can also be obtained from their normalized flux densities (as with Jupiter), utilizing the information in Ref 8. The  $\Delta T_A$  for these three planets obtained by this method essentially agree with the  $\Delta T_A$  obtained by using  $T_R$  and solid angles as described above. It should be noted that expressions (11) and (12) above assume that the planet disk has a uniform brightness distribution  $D(w)$  at the reception frequency. Nonthermal radio emission from Jupiter's radiation belts contributes to Jupiter's brightness temperature at a wavelength of 3.56 cm (Refs 5 and 6), and in addition, the brightness temperature changes with time at this wavelength. Consequently, the above assumption of uniform

brightness distribution is not strictly true for Jupiter and a small error results. Note that the planets Jupiter and Venus produce the largest increase in system noise temperature.

Consider two receiving systems (a 64- and 34-meter-diameter antenna) arrayed for reception of a spacecraft signal. A planet is within the beamwidth of the antennas. The following set of components of antenna feed spacing perpendicular to the line of sight to the planet ( $B_{r_{1,2}}$ ) provides a corresponding set of fringe spacings ( $\lambda/B_{r_{1,2}}$ ) as shown below.

$B_{r_{1,2}}$ , meters	Fringe spacing ( $\lambda/B_{r_{1,2}}$ ), arc sec
100	73 50
200	36 75
300	24 50
400	18 37
500	14 70

Consider a situation where a  $B_{r_{1,2}}$  of 200 meters represents operation of a 64-meter-diameter antenna receiving system (system 1) at an antenna elevation angle of about 60 deg (fringe spacing is 36.75 arcsec). The system noise temperature is 20 + 1 or 21 kelvins (excluding noise contribution from a planet). The planet Jupiter is in the beamwidth of each of the two antennas. Jupiter is at closest approach to the Earth so that it subtends an angle of 50 arcsec and its contribution to the system noise temperature of the 64-meter-diameter antenna receiving system is 13 kelvins (Table 2). The ratio of 50/36.75 equals 1.36 which, from Fig. 18, provides a reduction in relative magnitude of correlated noise power from one to 0.21, which represents 2.7 kelvins. The operating system noise temperature (system 1) is 21 + 13 or 34 kelvins. The statistically independent portion ( $1 - \epsilon_1$ ) of predetection noise power is  $(34 - 2.7)/34$  or 0.92 for system 1. For an array of two receiving systems utilizing a 34-meter-diameter transmit/receive antenna in conjunction with a 64-meter-diameter antenna (Fig. 3), the improvement in predetection signal-to-noise ratio is reduced from 1.0 dB for  $1 - \epsilon_1 = 1.0$  to 0.91 dB for  $1 - \epsilon_1 = 0.92$  or 0.09 dB ( $\gamma_2 = 0.53$ ). The reduction in RF carrier margin improvement (Fig. 13) is also 0.09 dB. This 0.09-dB reduction in array performance at  $1 - \epsilon_1 = 0.92$  essentially applies over the range of  $\gamma_2$  values shown in Figs. 3 and 13 and also to Figs. 7 and 12 over a large portion of the voltage coupling ( $\beta_2$ ) range shown. Note that, in light of the discussion above relative to expressions (10) and (11), this 0.09-dB reduction applies at those times when the path length to the two antennas represents an in-phase predetection noise situation (2.7 kelvins) due to the planet Jupiter. At those times when the path lengths represent an out-of-phase predetection noise situation for the 2.7 kelvins, the predetection signal-to-noise ratio and carrier margin improvement will be larger than the values shown in Figs. 3 and 13 for  $1 - \epsilon_1 = 1.0$ . The periodic reduction in predetection signal-to-noise ratio and RF carrier margin improvement for the situation discussed above is included as the first array configuration (first line) in Table 3 under the column titled Maximum Periodic Reduction in Array Gain.

Table 3 shows periodic reduction in array gain (as defined in the discussion above) at a reception frequency of 8420 MHz ( $\lambda = 3.56$  cm) for two receiving systems arrayed (64- and 34-meter-diameter antenna) with the planets Jupiter or Venus in the antenna beamwidth at either minimum or maximum distance from Earth. The table includes a few selected values of  $B_{r_{1,2}}$  (component of antenna feed spacing perpendicular to the line of sight to the planet) at antenna elevation angles of approximately 60 and 20 deg. The  $B_{r_{1,2}}$  value of 100 meters (shown with an asterisk) is not representative of planned antenna spacings for the DSN. It is included to illustrate the effect on array gain reduction due to the planets for closer antenna feed spacings. The planets Jupiter and Venus were chosen since their presence in the antenna beamwidth produces the largest effect on periodic array gain reduction.

The calculated performance shown in Table 3 is expanded in Figs. 19 through 26 to provide a more graphic presentation of the maximum periodic reduction in array gain as a function of  $B_{r_{1,2}}$  for Jupiter (Figs. 19, 20, 23, and 24), or for Venus (Figs. 21, 22, 25, and 26) in the array beamwidth at their minimum and maximum distances from Earth. Planned antenna spacings and orientation relative to the spin axis of the Earth provide a  $B_{r_{1,2}}$  in the range from 195 to 290 meters for a 34-meter-diameter

transmit/receive antenna arrayed with a 64-meter-diameter antenna and a  $B_{r_{1,2}}$  in the range from 445 to 550 meters for a 34-meter-diameter listen-only antenna arrayed with a 64-meter-diameter antenna

Note that for a 34-meter transmit/receive and 64-meter-diameter antenna array, a  $B_{r_{1,2}}$  in the 195 to 290 meter range results in a maximum periodic reduction in array gain of 0.1 dB or less for Jupiter in the array beamwidth and 0.17 dB or less for Venus in the array beamwidth. For a 34-meter listen-only and 64-meter-diameter antenna array, a  $B_{r_{1,2}}$  in the range from 445 to 550 meters results in a maximum periodic reduction in array gain of 0.05 dB or less for Jupiter in the array beamwidth and 0.1 dB or less for Venus in the array beamwidth.

Table 4 shows periodic reduction in array gain at a reception frequency of 8420 MHz for an array of three receiving systems with the planets Jupiter or Venus in the antenna beamwidth as in Table 3. The array utilizes a 64-meter-diameter antenna (system 1), a 34-meter-diameter listen-only antenna (system 2) and a 34-meter-diameter transmit/receive antenna (system 3) with antenna feed spacings (component perpendicular to the line of sight to the planet) of 500 meters ( $B_{r_{1,2}}$ ) and 200 meters ( $B_{r_{1,3}}$ ). This is representative of the planned three station array in Australia at an elevation angle of approximately 50 deg.

It should be noted that with the planets Uranus or Neptune in the antenna beamwidth (see Table 2) their effect on array gain is negligible even though (for the planned antenna feed spacings discussed above) most of the noise contribution from Uranus or Neptune will be coherent periodically.

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**Table 1. RF carrier array gain measurements**

( $\gamma_2 = \gamma_3 = 0.61, N_{o2}/N_{o1} = N_{o3}/N_{o1} = 1$  for  $1 - \epsilon_1 = 1$ )

Two receiving systems				Three receiving systems			
$1 - \epsilon_1$	$\beta_2$	Array gain, dB		$1 - \epsilon_1$	$\beta_2 = \beta_3$	Array gain, dB	
		Theory	Measured			Theory	Measured
1.0	0.5	1.3	1.3	1.0	1.0	1.9	1.8
1.0	0.43	1.23	1.2	1.0	0.97	1.94	1.9
0.73	0.5	0.92	0.9	1.0	0.61	2.28	2.3
0.73	0.43	0.88	0.8	0.73	1.0	1.33	1.2
0.21	0.5	0.28	0.2	0.73	0.61	1.55	1.6
0.21	0.41	0.24	0.2	0.21	0.59	0.46	0.4

**Table 2 Increase in system noise temperature due to the planets;  
64-meter-diameter antenna;  $\lambda = 3.55$  to  $3.6$  km**

Planet	Nominal radio disk temperature ( $T_R$ ), K	Subtended angle from earth, arcsec		Increase in system noise temperature, K	
		max	min	max	min
Mercury	330	13	5	1.8	0.3
Venus	600	67	17	85.0	5.5
Mars	200	24	4	3.6	0.1
Jupiter	230	50	32	13.0	5.0
Saturn	165	20	14	1.4	0.7
Uranus	210		3.6		0.08
Neptune	200		2.2		0.03

**Table 3. Maximum periodic reduction in array gain due to a planet within beamwidth of two-aperture array (two receiving systems arrayed for coherent reception, reception frequency 8420 MHz,  $\lambda = 3.56$  cm)**

Array configuration	Antenna elevation angle, deg	Planet in array beamwidth	Planet subtended angle from earth $\nu$ , arcsec	$B_{r1,2}$ m	$\nu / \left( \frac{\lambda}{B_{r1,2}} \right)$	Maximum periodic reduction in array gain, dB
64- and 34-m transmit/receive	~60	Jupiter	50	200	1.36	0.09
	~20		50	200	1.36	0.07
	~60		32	200	0.87	0.04
	~20		32	200	0.87	0.03
	~60		50	100 <sup>a</sup>	0.68	0.18
	~20		50	100 <sup>a</sup>	0.68	0.14
	~60		32	100 <sup>a</sup>	0.44	0.16
	~20		32	100 <sup>a</sup>	0.44	0.12
	~60	Venus	67	200	1.82	0.09
	~20		67	200	1.82	0.08
	~60		17	200	0.46	0.16
	~20		17	200	0.46	0.12
	~60		67	100 <sup>a</sup>	0.91	0.09
	~20		67	100 <sup>a</sup>	0.91	0.08
	~60		17	100 <sup>a</sup>	0.23	0.22
	~20		17	100 <sup>a</sup>	0.23	0.16
64- and 34-m listen-only	~60	Jupiter	50	200	1.36	0.13
	~20		50	200	1.36	0.09
	~60		32	200	0.87	0.05
	~20		32	200	0.87	0.04
	~60		50	500	3.40	0.05
	~20		50	500	3.40	0.04
	~60		32	500	2.18	0.03
	~20		32	500	2.18	0.02
	~60	Venus	67	200	1.82	0.13
	~20		67	200	1.82	0.11
	~60		17	200	0.46	0.22
	~20		17	200	0.46	0.15
	~60		67	500	4.56	0.09
	~20		67	500	4.56	0.08
	~60		17	500	1.16	0.04
	~20		17	500	1.16	0.03

<sup>a</sup>Not planned for the DSN

**Table 4. Maximum periodic reduction in array gain due to a planet within beamwidth of three-aperture array (three receiving systems arrayed for coherent reception; reception frequency 8420 MHz;  $\lambda = 3.56$  cm)**

Array configuration	Antenna elevation angle, deg	Planet in array beamwidth	Planet subtended angle from earth $\nu$ , arc sec	$B_{r1,2}$ , m	$\nu / \left( \frac{\lambda}{B_{r1,2}} \right)$	$B_{r1,3}$ , m	$\nu / \left( \frac{\lambda}{B_{r1,3}} \right)$	Maximum periodic reduction in array gain, dB
64- and 34-m	~60	Jupiter	50	500	3.40	200	1.36	0.15
listen-only and	~20		50	500	3.40	200	1.36	0.12
34-m transmit/	~60		32	500	2.18	200	0.87	0.06
receive	~20		32	500	2.18	200	0.87	0.04
	~60	Venus	67	500	4.56	200	1.83	0.18
	~20		67	500	4.56	200	1.83	0.17
	~60		17	500	1.16	200	0.46	0.23
	~20		17	500	1.16	200	0.46	0.17

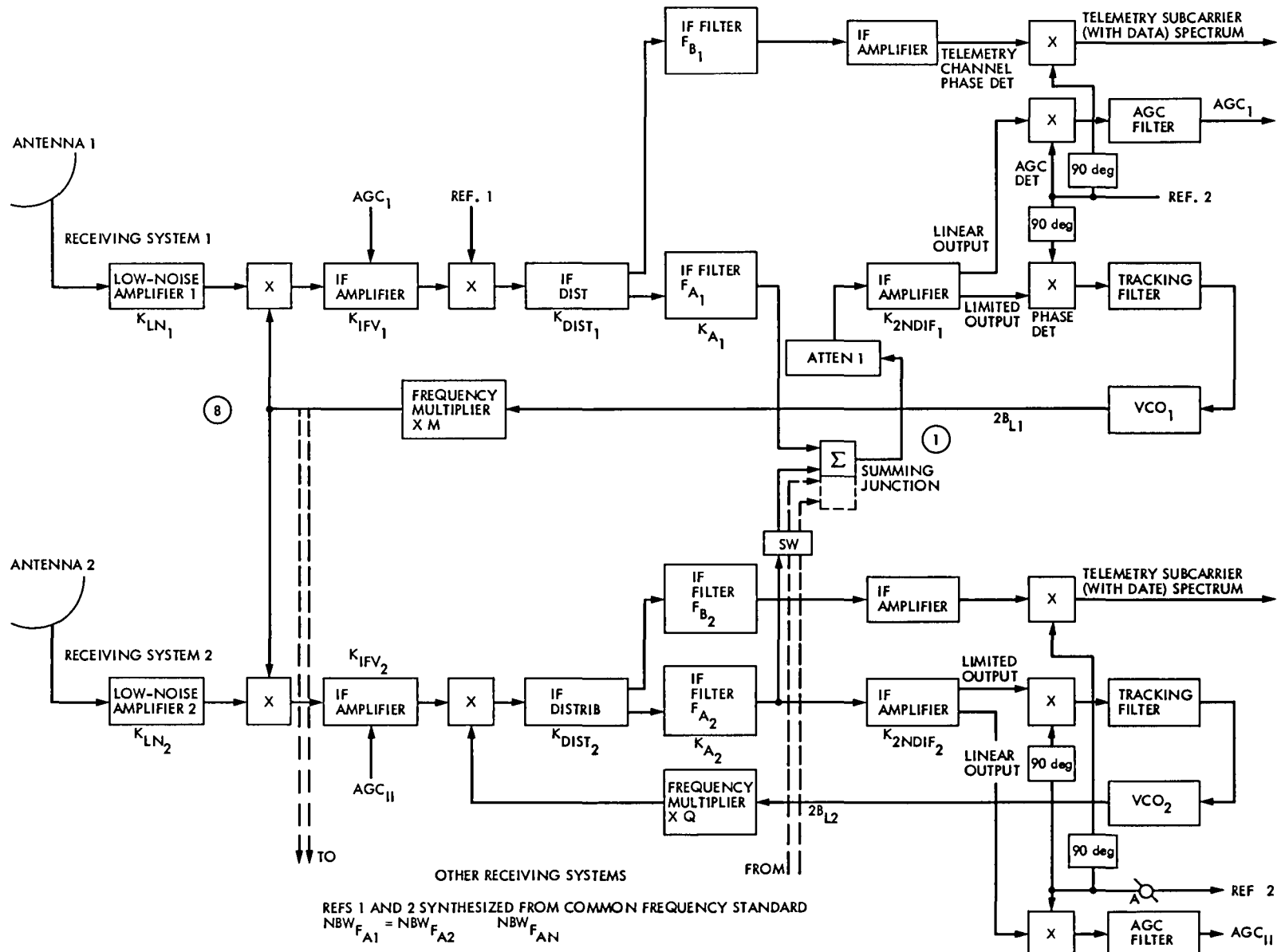


Fig. 1. Radio frequency carrier arraying for high-rate telemetry reception

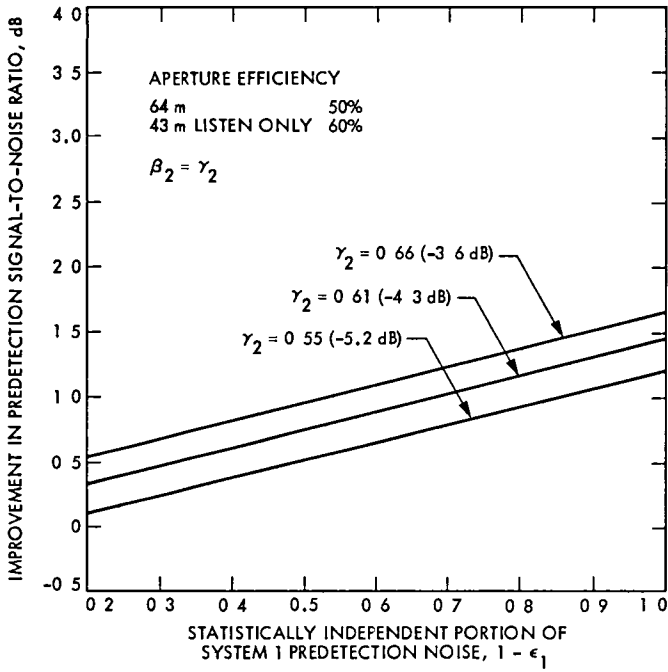


Fig. 2. Effect of partial coherence in system noise temperature on predetection signal-to-noise ratio improvement, two receiving systems arrayed, 64- and 34-m-diameter listen-only antennas

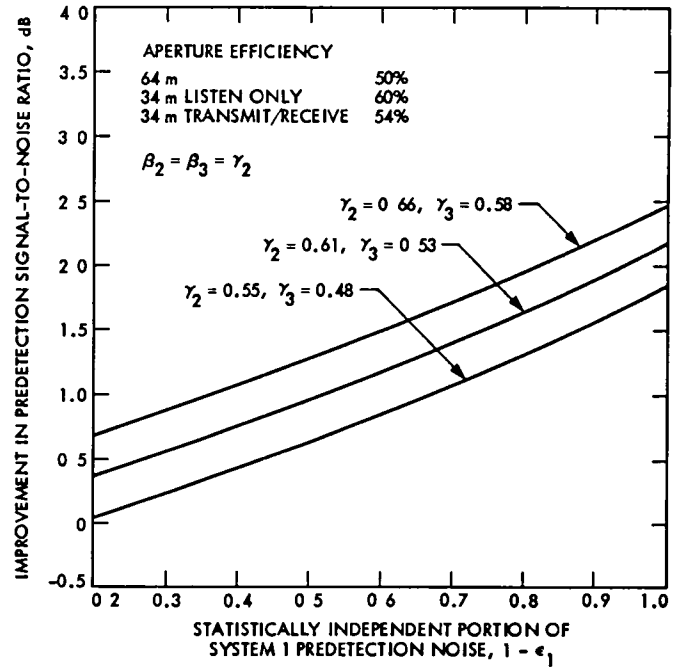


Fig. 4. Effect of partial coherence in system noise temperature on predetection signal-to-noise ratio improvement, three receiving systems arrayed; 64- and 34-m-diameter listen-only and 34-m-diameter transmit/receive antennas

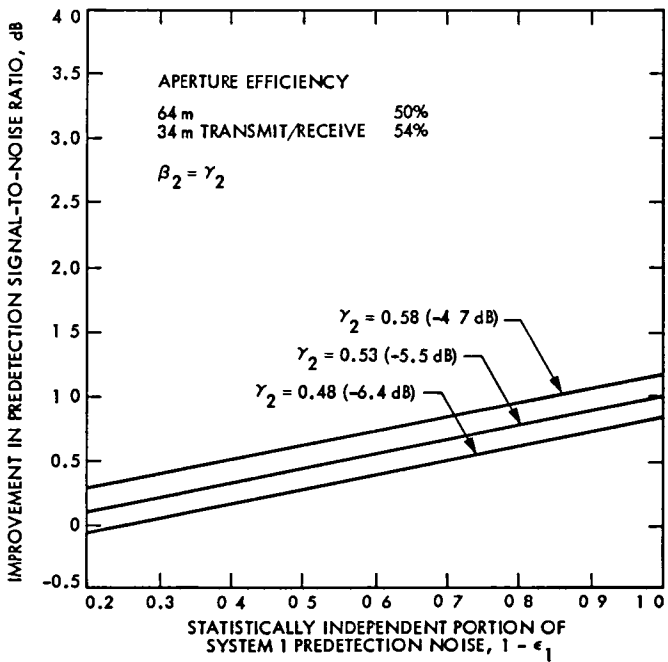


Fig. 3. Effect of partial coherence in system noise temperature on predetection signal-to-noise ratio improvement, two receiving systems arrayed, 64- and 34-m-diameter transmit/receive antennas

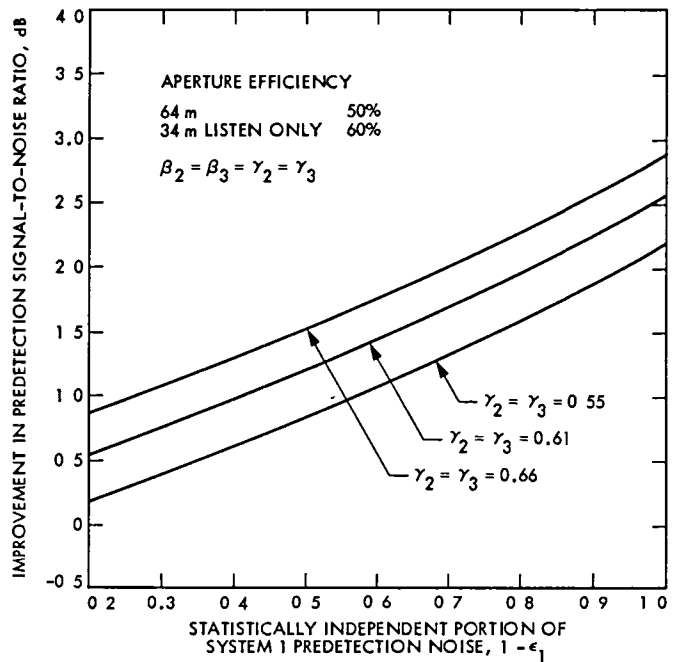


Fig. 5. Effect of partial coherence in system noise temperature on predetection signal-to-noise ratio improvement, three receiving systems arrayed, 64-m-diameter and two 34-m-diameter listen-only antennas

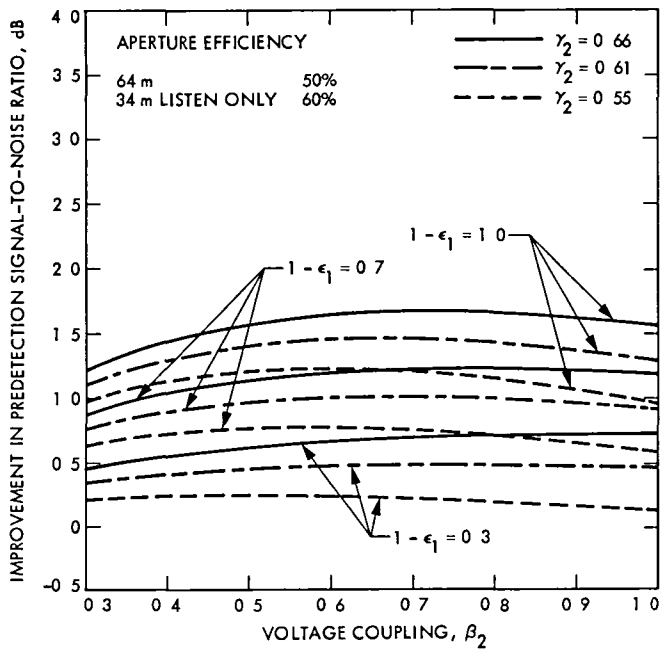


Fig. 6. Effect of signal summing voltage coupling on predetection signal-to-noise ratio improvement with partial coherence in system noise temperature, two receiving systems arrayed, 64- and 34-m-diameter listen-only antennas

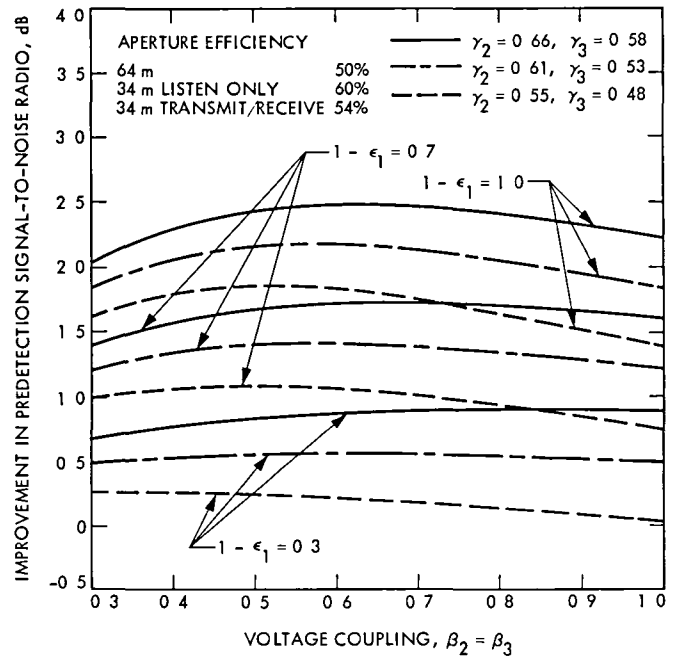


Fig. 8. Effect of signal summing voltage coupling on predetection signal-to-noise ratio improvement with partial coherence in system noise temperature, three receiving systems arrayed; 64-m-diameter, 34-m-diameter listen only, and 34-m-diameter transmit/receive antennas

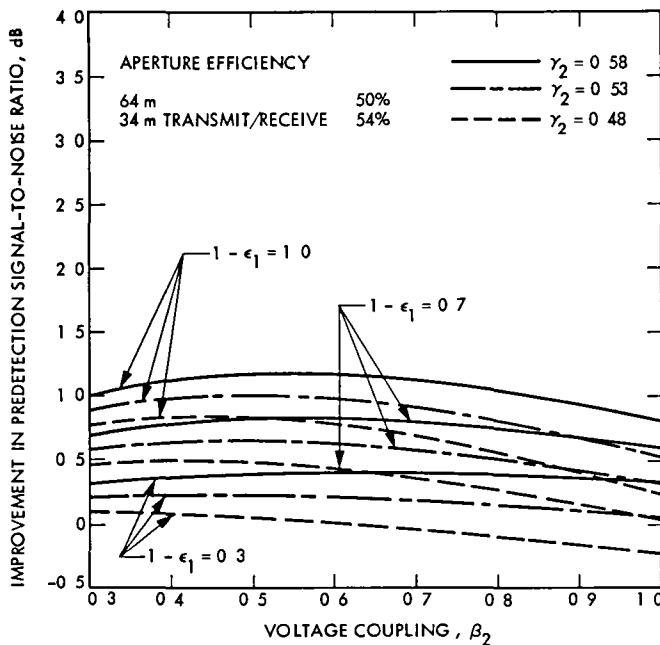


Fig. 7. Effect of signal summing voltage coupling on predetection signal-to-noise ratio improvement with partial coherence in system noise temperature, two receiving systems arrayed, 64- and 34-m-diameter transmit/receive antennas

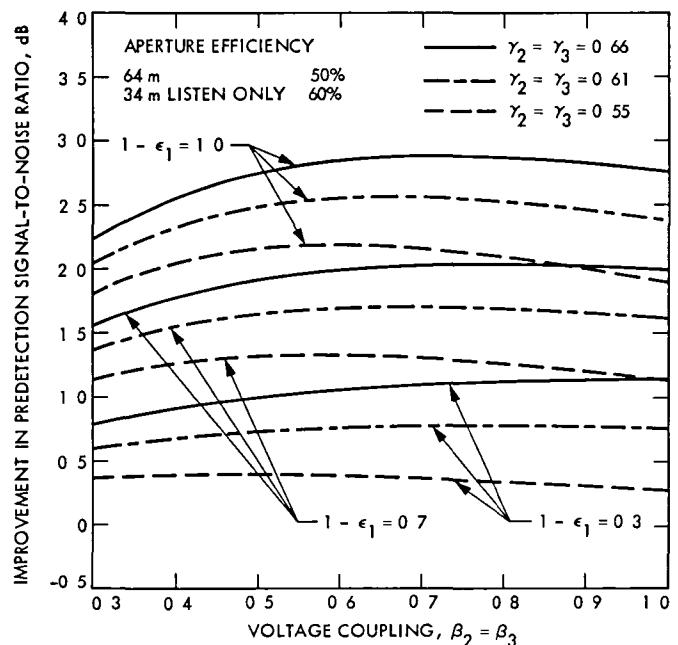


Fig. 9. Effect of signal summing voltage coupling on predetection signal-to-noise ratio improvement with partial coherence in system noise temperature, three receiving systems arrayed, 64-m-diameter and two 34-m-diameter listen-only antennas

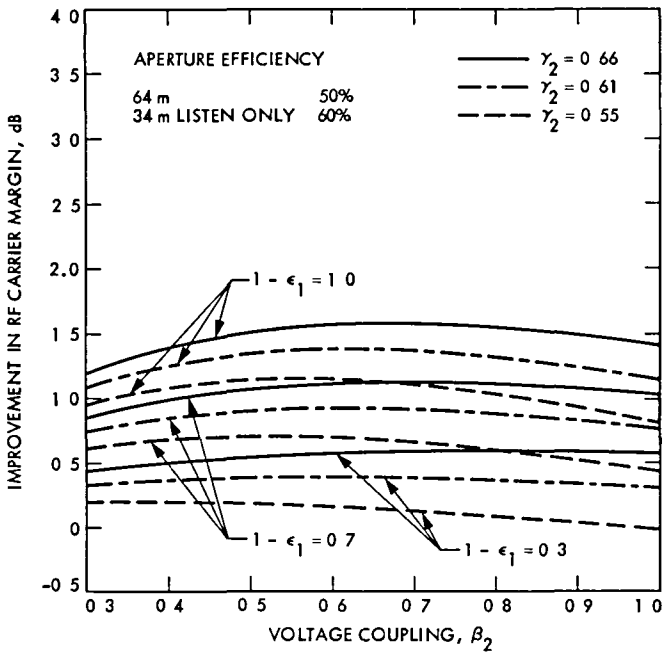


Fig. 10. Effect of summing junction voltage coupling on RF carrier margin improvement with partial coherence in system noise temperature, two receiving systems arrayed, 64- and 34-m-diameter listen-only antennas

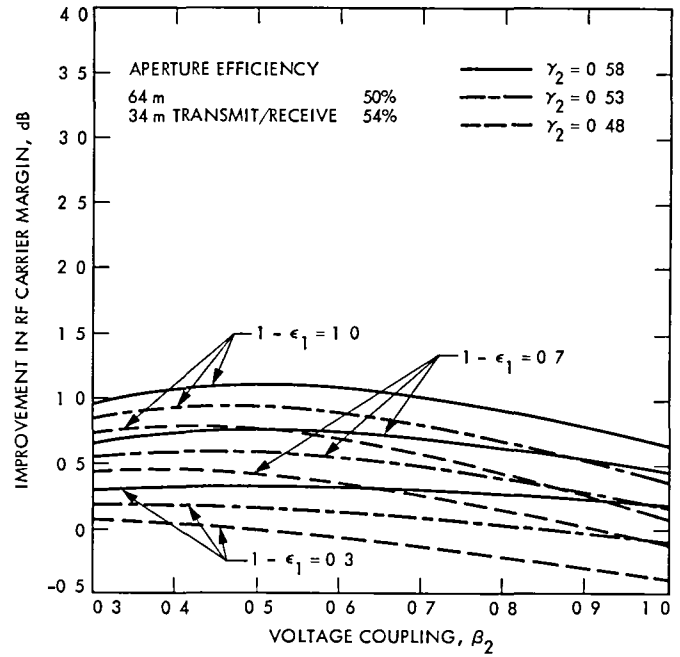


Fig. 12. Effect of summing junction voltage coupling on RF carrier margin improvement with partial coherence in system noise temperature, two receiving systems arrayed, 64- and 34-m-diameter transmit/receive antennas

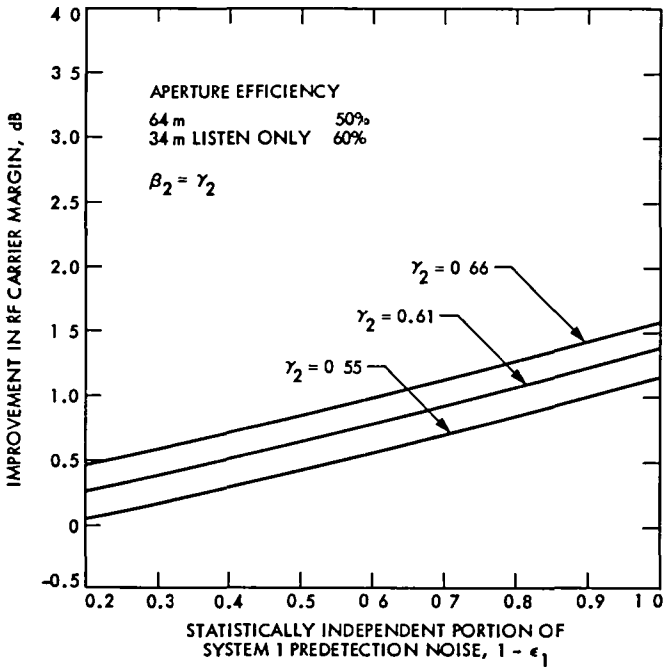


Fig. 11. Effect of partial coherence in system noise temperature on RF carrier margin improvement, two receiving systems arrayed, 64- and 34-m-diameter listen-only antennas

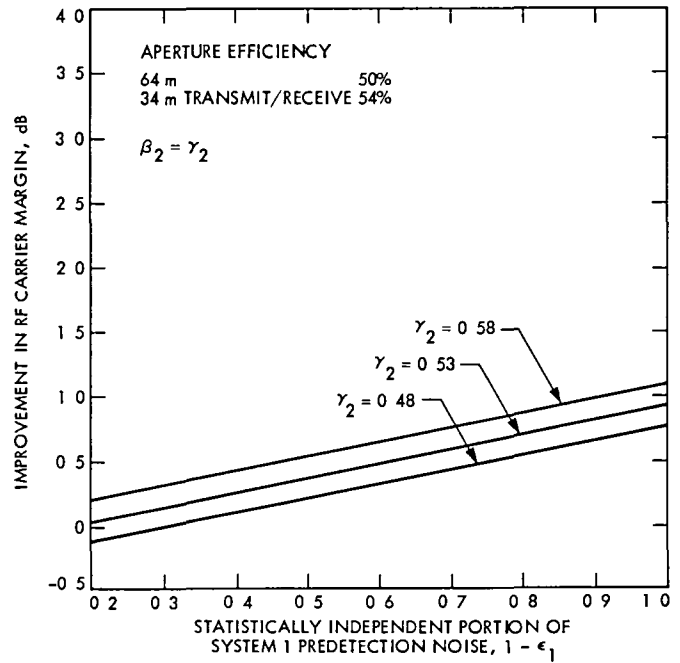


Fig. 13. Effect of partial coherence in system noise temperature on RF carrier margin improvement, two receiving systems arrayed, 64- and 34-m-diameter transmit/receive antennas

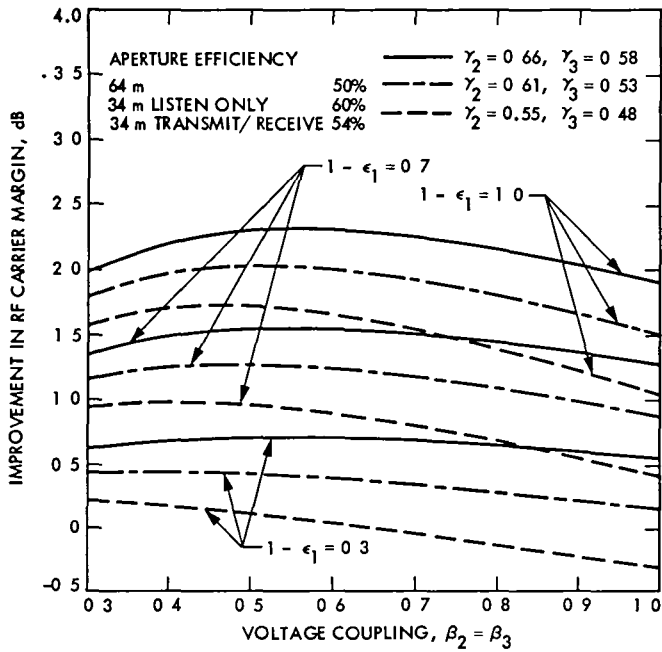


Fig. 14. Effect of summing junction voltage coupling on RF carrier margin improvement with partial coherence in system noise temperature, three receiving systems arrayed, 64-m-diameter, 34-m-diameter listen-only, and 34-m-diameter transmit/receive antennas

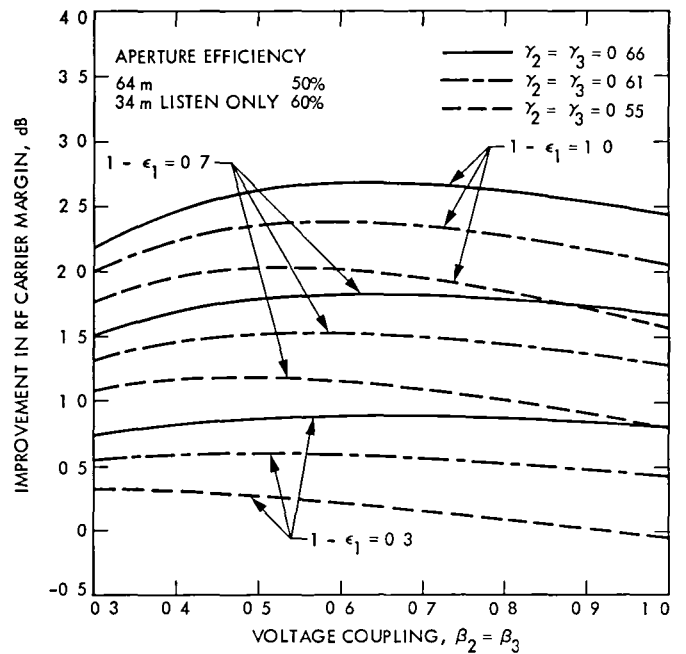


Fig. 16. Effect of summing junction voltage coupling on RF carrier margin improvement, with partial coherence in system noise temperature, three receiving systems arrayed; 64-m-diameter and two 34-m-diameter listen-only antennas

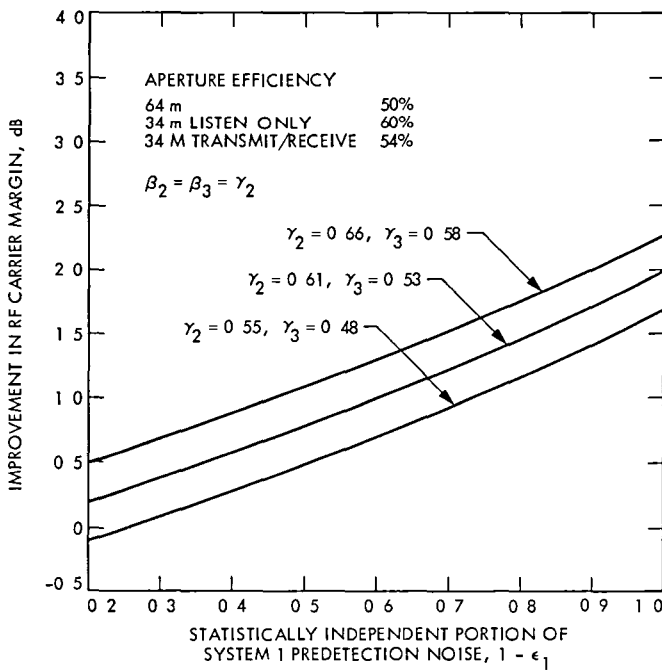


Fig. 15. Effect of partial coherence in system noise temperature on RF carrier margin improvement, three receiving systems arrayed; 64-m-diameter, 34-m-diameter listen-only, and 34-m-diameter transmit/receive antennas

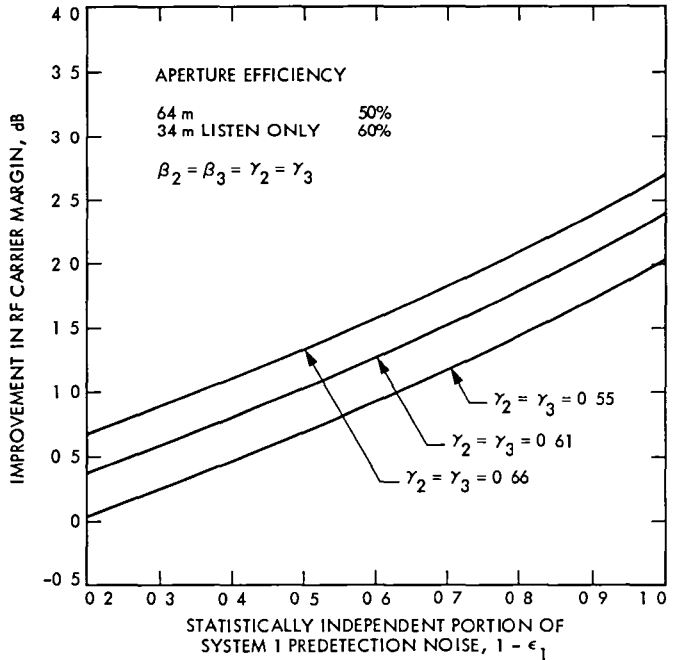


Fig. 17. Effect of partial coherence in system noise temperature on RF carrier margin improvement, three receiving systems arrayed; 64-m-diameter and two 34-m-diameter listen-only antennas



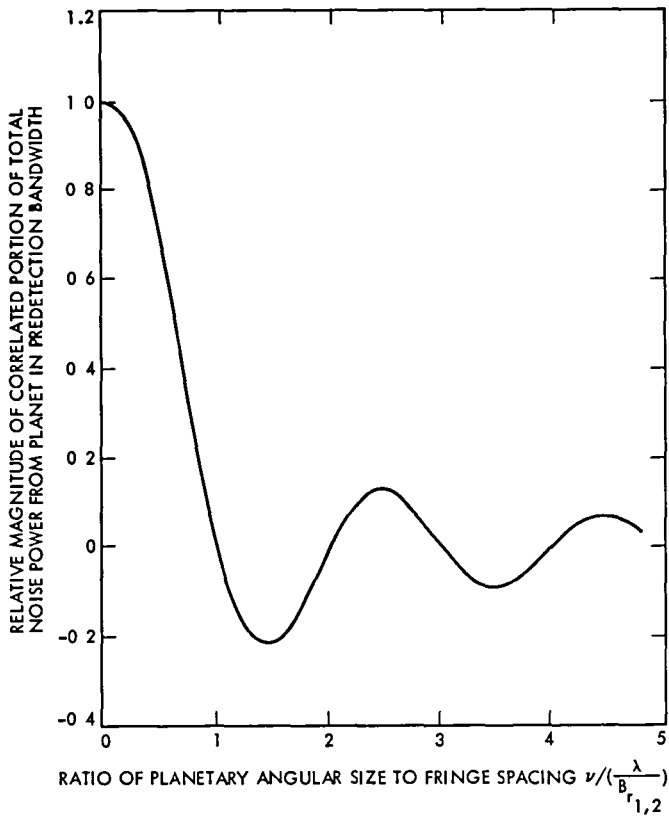


Fig. 18. Relative effect of ratio of planet angular size to fringe spacing on correlated predetection noise, two-antenna array

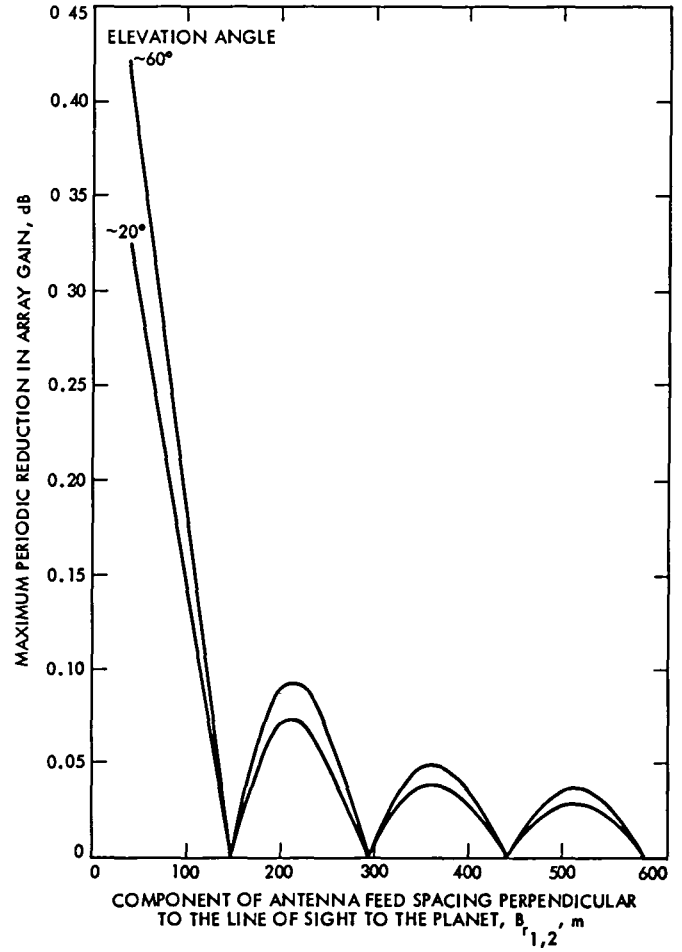
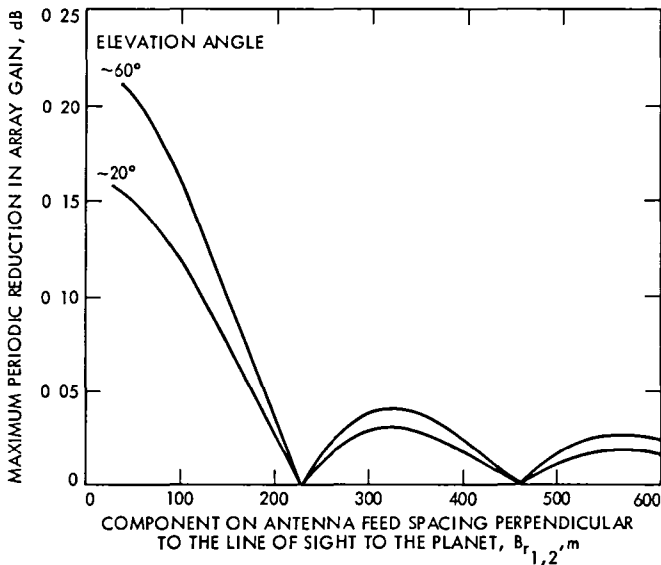
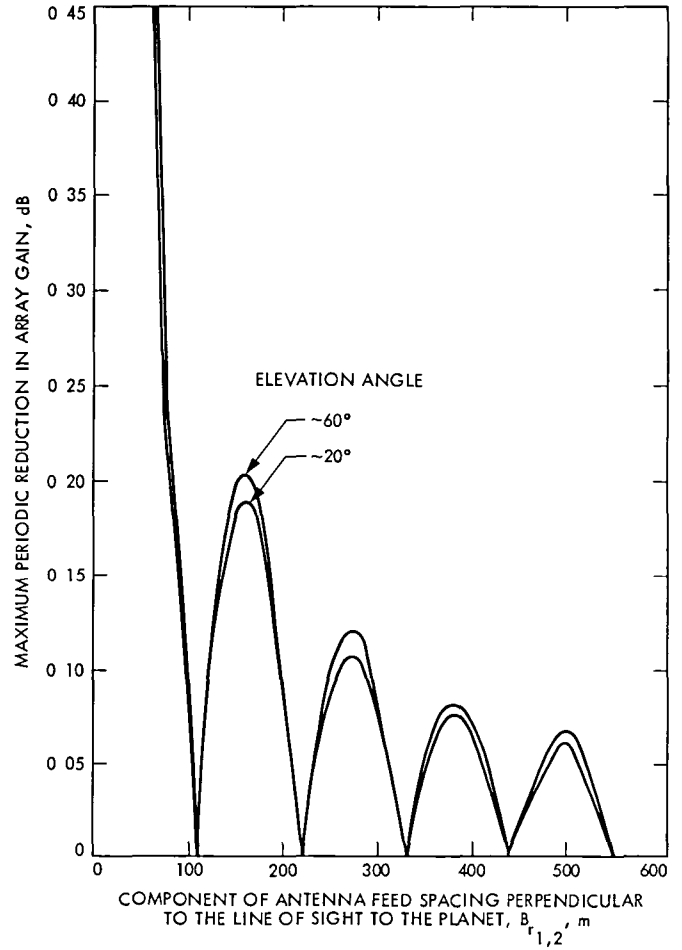


Fig. 19. Maximum periodic reduction in array gain due to Jupiter (at minimum distance) within beamwidth of two receiving systems arrayed for coherent reception vs effective feed spacing; 64- and 34-m-diameter (transmit/receive) antennas, reception at 8420 MHz



**Fig. 20. Maximum periodic reduction in array gain due to Jupiter (at maximum distance) within beamwidth of two receiving systems arrayed for coherent reception vs effective feed spacing, 64- and 34-m-diameter (transmit/receive) antennas, reception at 8420 MHz**



**Fig. 21 Maximum periodic reduction in array gain due to Venus (at minimum distance) within beamwidth of two receiving systems arrayed for coherent reception vs effective feed spacing, 64- and 34-m-diameter (transmit/receive) antennas, reception at 8420 MHz**

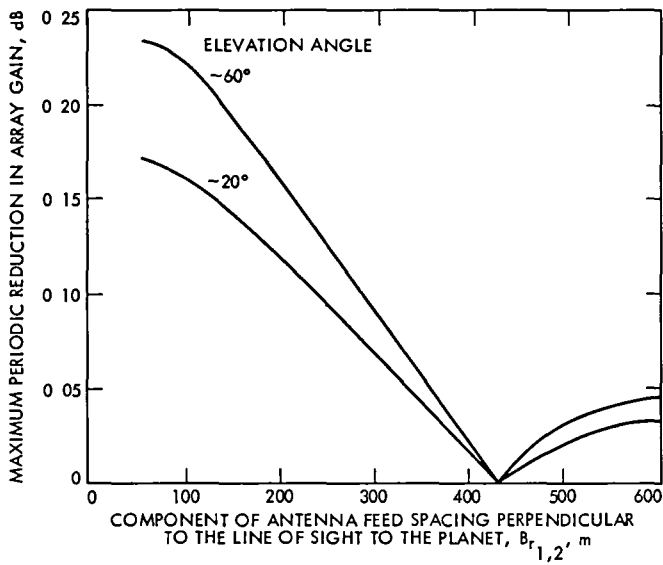


Fig. 22. Maximum periodic reduction in array gain due to Venus (at maximum distance) within beamwidth of two receiving systems arrayed for coherent reception vs effective feed spacing; 64- and 34-m-diameter (transmit/receive) antennas, reception at 8420 MHz

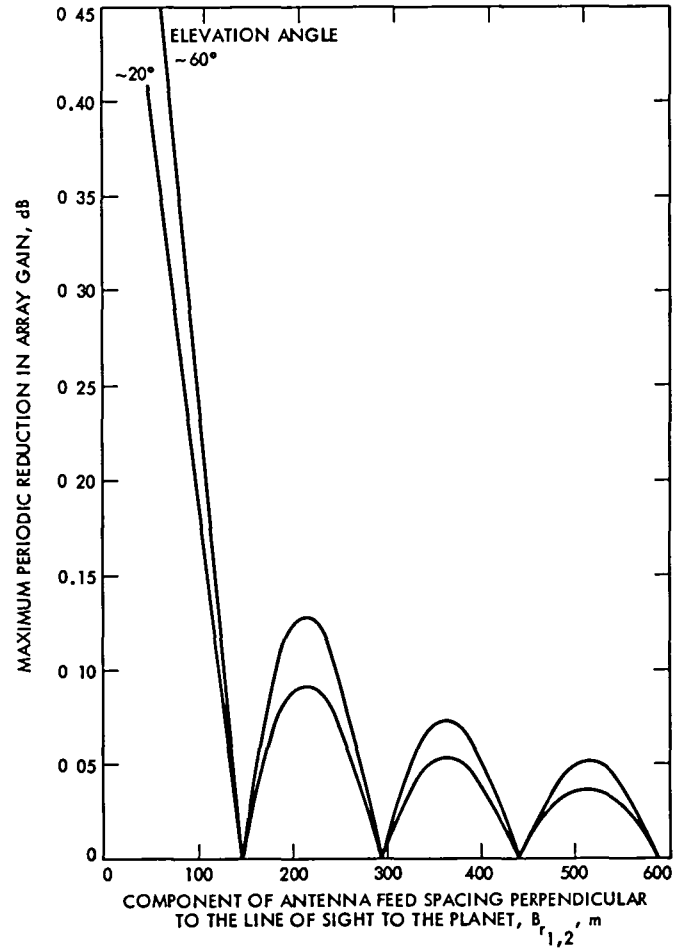
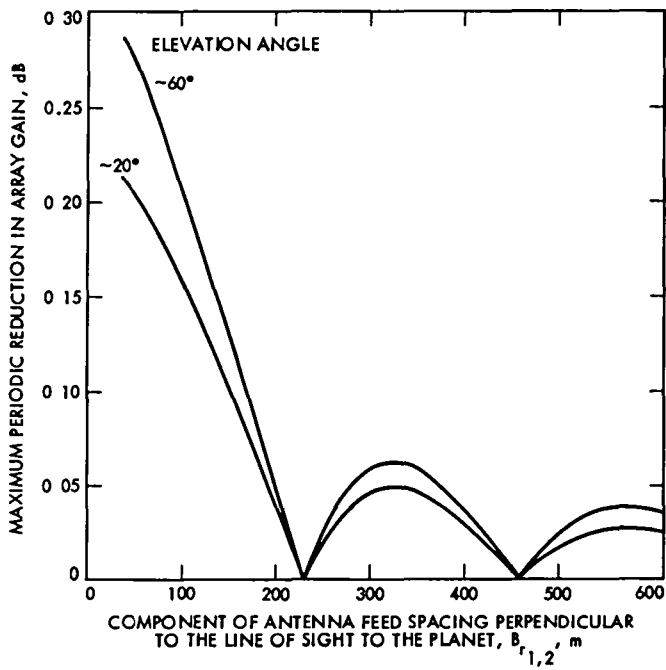
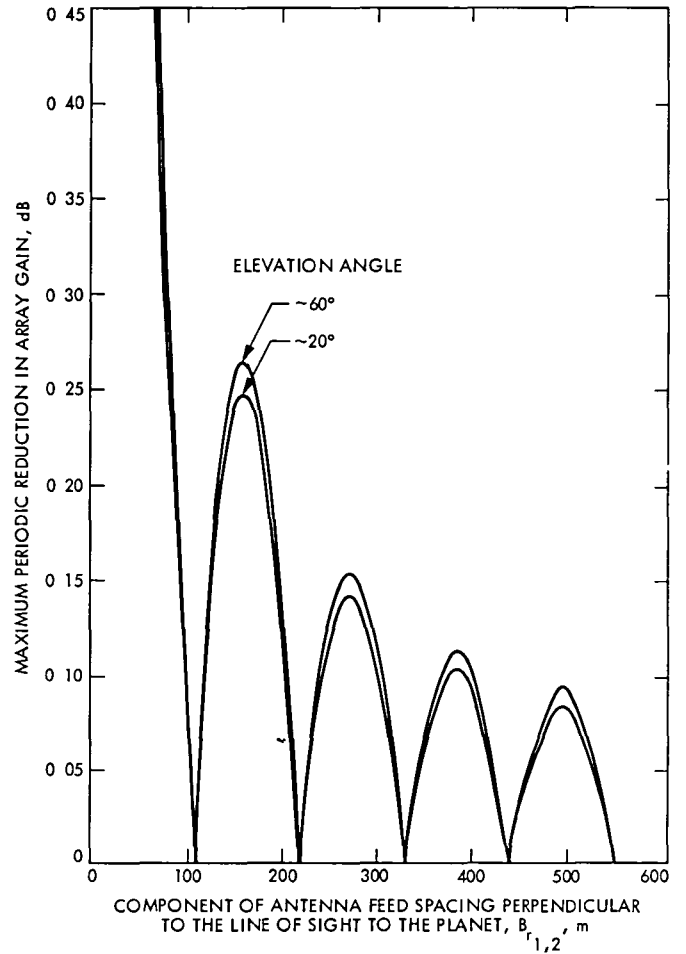


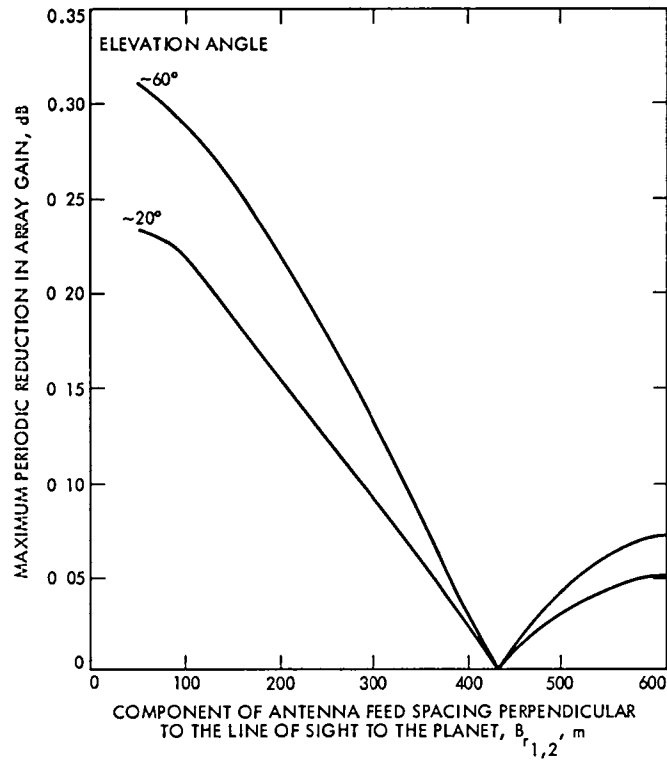
Fig. 23. Maximum periodic reduction in array gain due to Jupiter (at minimum distance) within beamwidth of two receiving systems arrayed for coherent reception vs effective feed spacing; 64- and 34-m-diameter (listen-only) antennas, reception at 8420 MHz



**Fig 24. Maximum periodic reduction in array gain due to Jupiter (at maximum distance) within beamwidth of two receiving systems arrayed for coherent reception vs effective feed spacing; 64- and 34-m-diameter (listen-only) antennas, reception at 8420 MHz**



**Fig 25. Maximum periodic reduction in array gain due to Venus (at minimum distance) within beamwidth of two receiving systems arrayed for coherent reception vs effective feed spacing, 64- and 34-m-diameter (listen-only) antennas, reception at 8420 MHz**



**Fig 26 Maximum periodic reduction in array gain due to Venus (at maximum distance) within beamwidth of two receiving systems arrayed for coherent reception vs effective feed spacing; 64- and 34-m-diameter (listen-only) antennas, reception at 8420 MHz**

# A Two-Year History of Atomic Frequency Standards Syntonization in the Deep Space Network

S C Ward

Operations Sustaining Engineering Section

*The NASA-JPL Frequency and Timing System (FTS) of the Deep Space Network (DSN) consists of a collection of three sets of clocks (located in Australia, California, U S A , and Spain) driven by independent atomic oscillators. It is the synchronization of the output frequencies (syntonization) of these oscillators (reference frequency standards) that is reported here. Prior to 1980 there was no DSN direct specification of syntonization. However, there was an implied specification of a  $\pm 5.5 \times 10^{12}$  related to the DSN time synchronization specification of a  $\pm 100$  microseconds. Both the syntonization within the three sets and the syntonization of the sets to the international standard [International Atomic Time (TAI), UTC (BIH)] are considered.*

## I. Establishing Syntonization

The original calibration was established during the third quarter of calendar 1980 by syntonizing a specially prepared transportable cesium atomic clock (Ref 1) to the standard frequency as maintained by the National Bureau of Standards at Boulder, Colorado. This "calibrated transfer standard" was then transported to each of the three sets of atomic frequency standards, where it was used to measure their frequency offset with respect to the NBS standard (designated as UTC-NBS). In this manner each of the reference frequency standards were syntonized to UTC-NBS and, to a lesser degree, syntonized to each other.

Each atomic reference frequency standard set consists of a Smithsonian Institution Astrophysical Observatory (SAO) model VLG-10B hydrogen maser (H2M) and two Hewlett-Packard model 5061-A option 004 cesium beam frequency standards. The offset of each H2M's output frequency from the hydrogen hyperfine spectral line was established with re-

spect to the transported transfer standard. By means of phase comparators, the H2M calibration is transferred to the two cesium beam standards. Thus, the syntonization between the three sets and between any set and UTC was established (Ref 2) within less than  $\pm 3 \times 10^{-13}$ .

## II. Maintaining Syntonization

To make effective use of syntonization it must be maintained on a continual basis, e.g., the tuning of an orchestra to A above middle C is a syntonization maintenance application. The DSN makes use of four resource techniques for maintaining syntonization. They are LORAN-C, traveling clocks, television vertical sync pulse, Very Long Baseline Interferometry (VLBI/ $\Delta$ DOR<sup>1</sup>). Figure 1 illustrates how the DSN is con-

<sup>1</sup>A technique equivalent in accuracy to VLBI that makes use of an extragalactic radio source along with a spacecraft to gather data from which time sync information can be derived.

figured to utilize these resources, making use of a combination of (1) Mediterranean chain LORAN, traveling clocks and VLBI/ $\Delta$ DOR in Spain, (2) television, microwave links, traveling clocks and VLBI/ $\Delta$ DOR in Australia, (3) microwave links, West Coast chain LORAN, traveling clocks<sup>2</sup> and VLBI/ $\Delta$ DOR in California

Table 1 lists the comparative merits of the resources used for maintaining syntonization. For all comparisons, it has been assumed that the resource has been in use for gathering data for periods longer than one year and that at least two out of three reference frequency standards in a set have been in continual operation over that period.

### III. Syntonization to UTC

Figure 2, a plot of  $\Delta f/F$  vs time, presents a two-year history of the syntonization of the DSS 63<sup>3</sup> reference frequency standards with respect to UTC-USNO. The "SPEC wrt UTC (NBS)" annotated dotted lines in the vicinity of UTC (USNO) plus and minus  $10E-12$  refers to System Functional Requirements Document 821-14 III B 2. The data were collected through the use of LORAN-C and have been adjusted through the use of USNO Bulletin 4 corrections. With the exception of routine frequency standard maintenance (H2M tuning) periods, 98% of the data presented applies to H2M, SAO VLG-10B serial #07. The data indicate that the reference frequency standards were syntonized to UTC-USNO within less than 1 fractional part in  $10^{12}$  over the entire two-year period.

Figure 3, a plot of  $\Delta f/F$  vs time, presents a two-year history of the syntonization of the DSS 43<sup>4</sup> reference frequency standards with respect to UTC-AUS<sup>5</sup>. The "SPEC wrt UTC (NBS)" annotated dotted lines in the vicinity of UTC(USNO) plus and minus  $10E-12$  refers to System Functional Requirements Document 821-14 III B 2<sup>4</sup>. The data were collected through the daily use of the TV simultaneous view technique. The reference to which syntonization is made is the Department of National Mapping (NATMAP) clock ensemble (four clocks) in the local area custodian of UTC-AUS. It is equivalent to and serves the same function for DSS 43 as the Deep Space Network Master Clock ensemble (NMC) does for DSS

<sup>2</sup>At Goldstone, California, traveling clocks serve a double role: (1) to synchronize the Network Master Clock (NMC) ensemble of seven cesium clocks to UTC-NBS, and (2) to synchronize the DSS 14 clocks to the NMC ensemble on each VLBI time sync operation.

<sup>3</sup>Deep Space Tracking Station 63 (DSS 63) is located in the Madrid area of Spain (Fig. 1).

<sup>4</sup>DSS 43 is located in the Australian Capital Territory (A.C.T.) near Canberra.

<sup>5</sup>UTC-AUS is traceable to UTC-USNO within  $\pm 5 \times 10^{-13}$  through the use of semiannual traveling clock trips between Canberra, A.C.T., and Washington, D.C.

14 at Goldstone, California, USA. The data have been adjusted per published NATMAP Bulletin E corrections. With the exception of the routine frequency standard maintenance (H2M tuning) periods, 83% of the data applies to H2M, SAO VLG-10B serial #06. The data indicate that the reference frequency standards were syntonized to UTC-USNO within a part in  $10^{12}$  over the entire two-year period.

Figure 4 presents a two-year history of the syntonization of the DSS 14 reference frequency standards with respect to UTC-NMC<sup>6</sup>. The data were adjusted per weekly NMC closure reports. Over this two-year report period there were occasions when UTC-MC and/or UTC-DSN were not available<sup>7</sup>. Continuity of the syntonization history of H2M SAO VLG-10B serial #05 was maintained at those times through the use of output frequency drift data from the hydrogen hyperfine line obtained during tuning maintenance. Greater than 98% of the data presented applies to H2M SAO VLG-10B serial #05, and it indicates that the syntonization of the DSS 14 reference frequency standards was maintained within a part in  $10^{12}$  of UTC(NBS) over the entire period. Performance was borderline during April and May 1982.

### IV. Syntonization Within the DSN

Figure 5 presents a plot of  $\Delta f/F$  vs time for a two-year history of the syntonization of the DSS 14 reference frequency standard with respect to the DSS 63 reference frequency standard. The data were collected through the use of VLBI and  $\Delta$ DOR time synchronization operations (Ref. 3). These syntonization data were derived from sets of blunder-free time offset data points, where a set consists of a minimum of three "good" points<sup>8</sup>. Further, where there was a lack of "good" VLBI data,  $\Delta$ DOR data sets were used. The data indicate that the DSS 63 reference frequency standards remained syntonized to the DSN master at DSS 14 within  $\pm 3 \times 10^{-13}$  over the two-year report period.

Figure 6 presents a plot of  $\Delta f/F$  vs time for a two-year history of the syntonization of the DSS 14 reference frequency

<sup>6</sup>UTC-NMC is syntonized to UTC-NBS, and to lesser degree to UTC-USNO, within a part in  $10^{13}$  through the use of traveling clocks and West Coast LORAN-C. UTC-NMC is syntonized to UTC-DSN within less than a part in  $10^{13}$  through the use of weekly traveling clock trips and microwave links. Thus UTC-DSN is syntonized to UTC-NBS within a part in  $10^{12}$ .

<sup>7</sup>Extended power outages and airconditioning failures in the building housing the NMC ensemble and extensive repair work on the DSS14 antenna caused the loss of UTC-DSN and UTC-NMC.

<sup>8</sup>A point is defined as "good" if it is within less than  $3\sigma$  of the average rate computation ( $\Delta f/f$ ), where the number of data points used is greater than 2.

standard with respect to the DSS 43 reference frequency standard. The data were collected through the use of VLBI and  $\Delta$ DOR time synchronization operations (Refs 3, 4). These synchronization data were derived from sets of blunder-free time offset data points, where a set consists of a minimum

of three "good" points. Further, where there was a lack of "good" VLBI data,  $\Delta$ DOR data sets were used. The data indicate that the DSS 43 reference frequency standards remained synchronized to the DSN master of DSS 14 within  $\pm 3 \times 10^{-13}$  over the two-year report period.

## References

- 1 Ward, S , "The Operational Performance of Hydrogen Masers in the DSN," *TDA Progress Report 42-63*, pp 203-218, Jet Propulsion Laboratory, Pasadena, Calif , June 15, 1981
- 2 Ward, S , "The Operational Performance of Hydrogen Masers in the DSN," Proceedings of the Twelfth Annual Precision Time and Time Interval (PTTI) Applications and Planning Meeting, pp 616-617, December 2-4, 1980, Goddard Spaceflight Center, Greenbelt, Md
- 3 Roth, M G , "International Clock Synchronization with the Block I VLBI System," Proceedings of the Thirteenth Annual Precise Time and Time Interval (PTTI) Applications and Planning Meeting, pp 489-500, December 1-3, 1981, Goddard Spaceflight Center, Greenbelt, Md



**Table 1 Comparative accuracy of syntonization maintenance resources**

	Osc to osc (DSS to DSS) <sup>a</sup>	Osc to listed Ref <sup>a</sup>	Osc to UTC- USNO <sup>a</sup>	Osc to UTC- NBS <sup>a</sup>
LORAN-C (daily meas )	1 1E -12	±5E -13 LORAN	±7E -13	±7E -13
LORAN-C (simult view)	1 0E -12	±3 5E -13 LORAN	±5E -13	±6E -13
Traveling clocks (2 trips/yr)	1 4E -13 <sup>b</sup>	±7 5E -14 Clock <sup>b</sup>	DNA	DNA
Traveling clocks (4 trips NBS/yr)	DNA	±7 8E -14 UTC-NMC	±2 5E -13	±1 3E -14
Television (daily meas )	1 05E -12	±2 3E -13 UTC-AUS	±5 5E -13	±7 4E -13
VLBI X-band (7 to 10 day)	5 4E -15	±3 8E -15 UTC-DSN	DNA	DNA
VLBI S-band (7 to 10 day)	1 4E -14	±9 9E -15 UTC-DSN	DNA	DNA
Microwave (3 meas /day)	2 46E -12	±2 2E -15 UTC-NMC (1)	DNA	DNA

<sup>a</sup>The number following E designates the exponent or power of 10 e.g , ±7 5E -14 is ±7 5 × 10<sup>-14</sup>

<sup>b</sup>These are frequency or phase measurements All others are time measurements



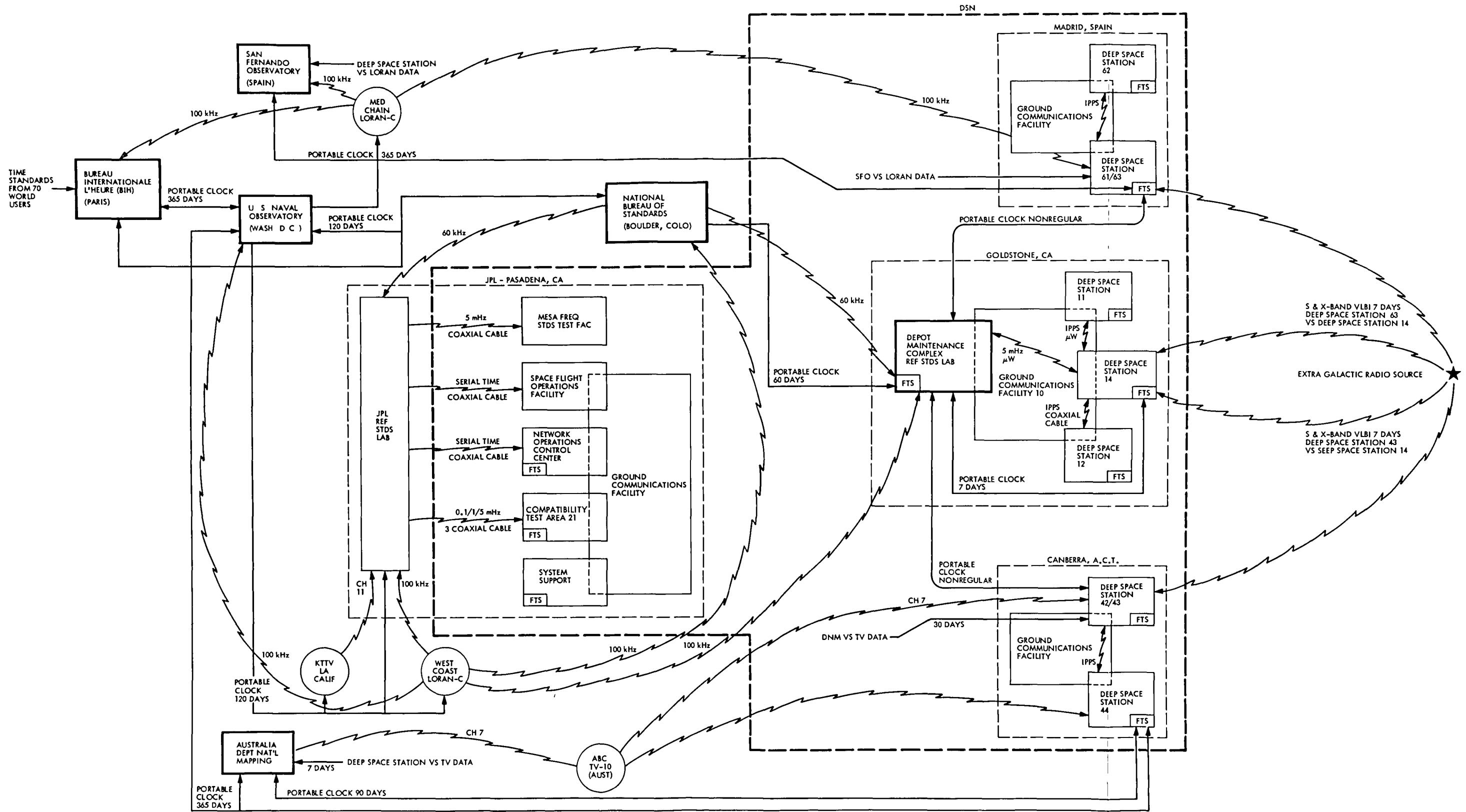


Fig 1 NASA/JPL intra/extra DSN frequency and time sync system

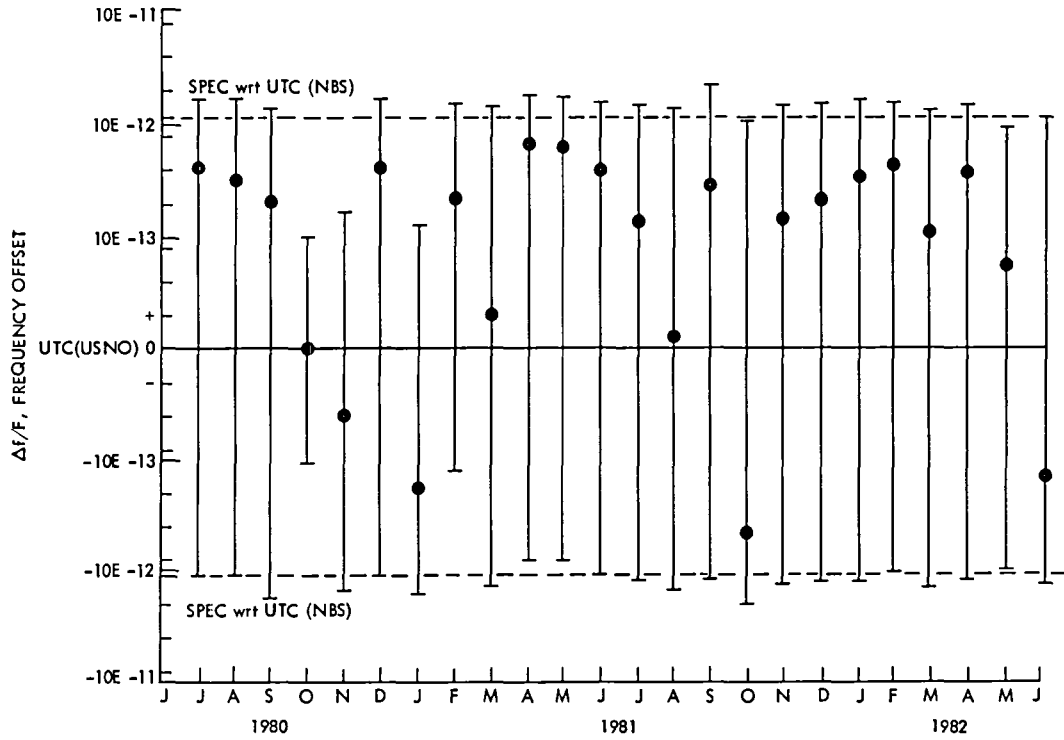


Fig. 2. DSS 63 syntonization to UTC/(USNO) through the use of LORAN-C

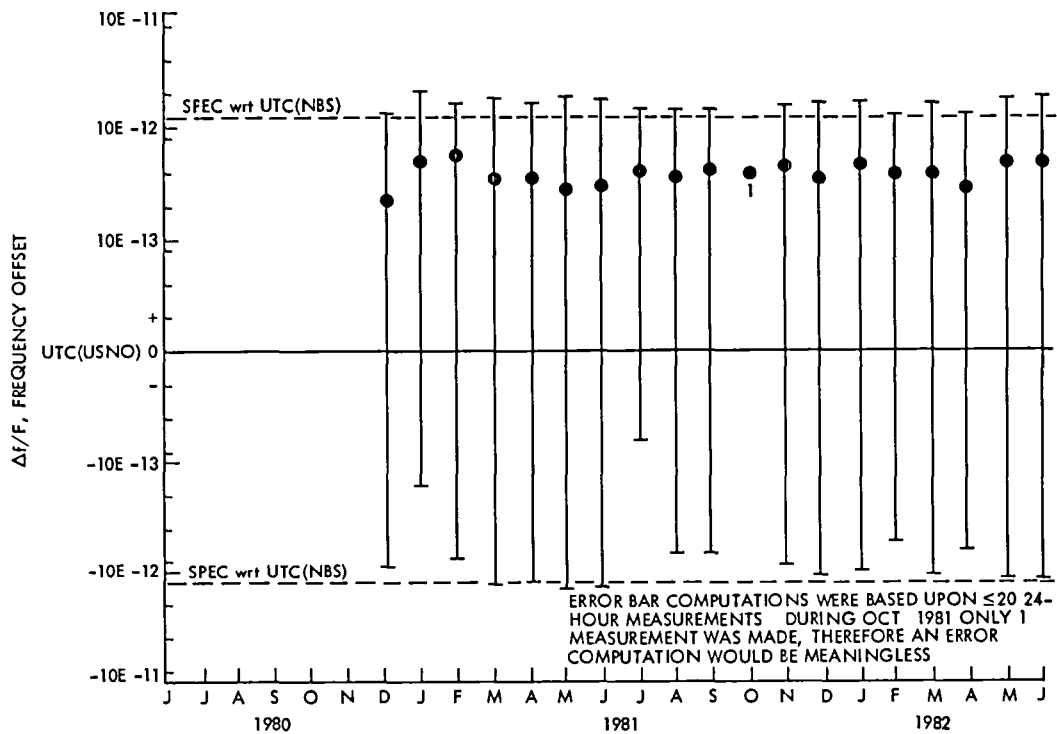


Fig. 3. DSS 43 syntonization to UTC/(USNO) through the use of television

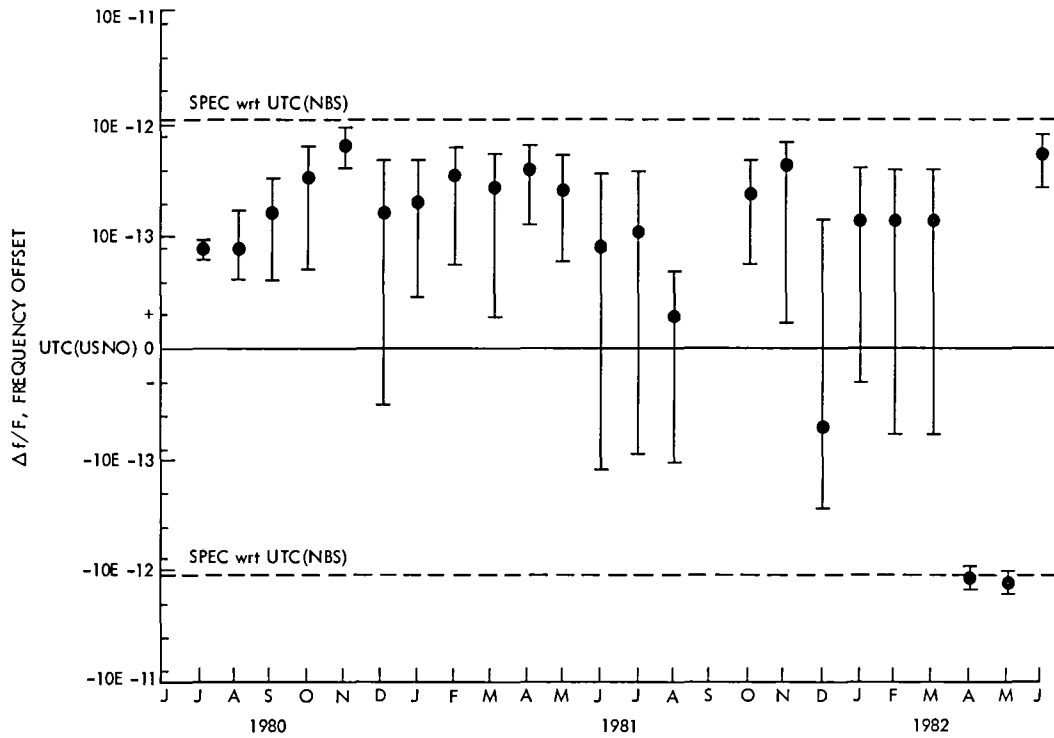


Fig. 4 DSS 14 syntonization to UTC/(USNO/NBS) through the use of microwave and traveling clocks

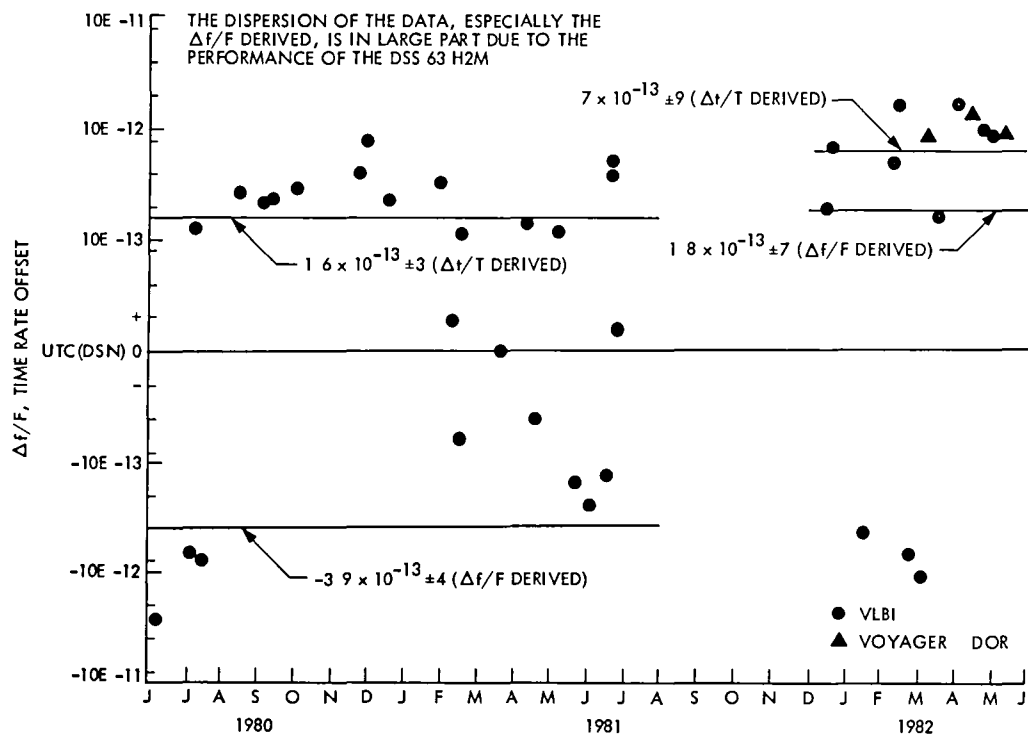


Fig 5 DSS 63 vs DSS 14 syntonization through the use of VLBI

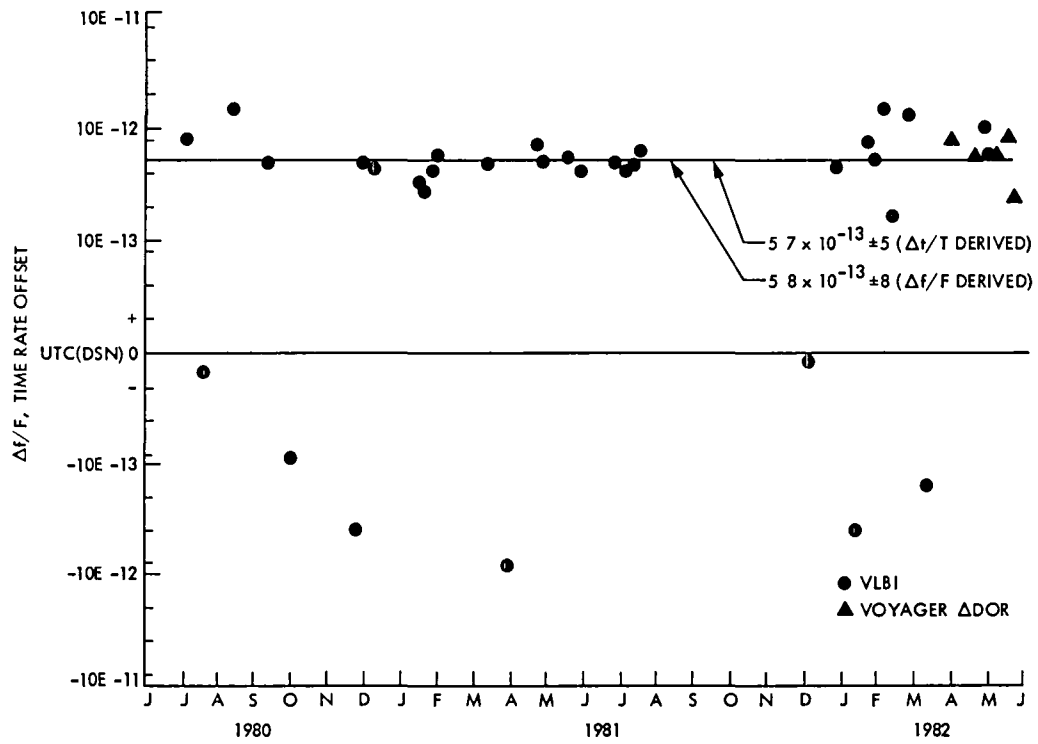


Fig. 6 DSS 43 vs DSS 14 syntonization through the use of VLBI

# DSN Test Support System, Mark IV-85

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*As part of the Network Consolidation Program, the present DSN Test and Training System, Mark III-77, will be upgraded during the Mark IVA implementation planned for 1984-85. This new upgraded system will be referred to as the DSN Test Support System, Mark IV-85. This article provides descriptions and functional capabilities that exist or will be implemented in the subsystems which comprise the DSN Test Support System.*

## I. System Definition

### A. General

The DSN Test Support System, as part of the Network Consolidation Program (Ref 1), provides support capabilities for testing the DSN data systems at each of the three Deep Space Communications Complexes (DSCC) (Goldstone, California, Canberra, Australia, Madrid, Spain) and at the Network Operations Control Center (NOCC). The system also provides support to flight projects for testing and training activities with the DSN facilities. The system includes capabilities for

- (1) Facility-level system performance testing of each DSN system
- (2) Maintenance and diagnostic testing for the DSCC subsystems
- (3) End-to-end testing of each DSN system, including related DSCC, Ground Communications Facility (GCF), and NOCC functions

The system consists mainly of a number of hardware and software components that are normally parts of other DSN systems and subsystems. However, it includes two subsystems that are dedicated to the test support: DSCC Test Support Subsystem and NOCC Test Support Subsystem.

Figure 1 describes the system's functions and interfaces. This article updates the system description published in Ref 2.

### B. Key Characteristics

Design goal key characteristics of the DSN Test Support System are

- (1) Centralization of test support capabilities at each DSCC
- (2) Capability to support testing and maintenance activities concurrently with operational activities
- (3) Capability to function without alteration of DSN operational configurations
- (4) Accommodation of flight-project-supplied simulated data via GCF
- (5) Capability to support network loading tests with a combination of actual and simulated data streams
- (6) Testing functions to be controlled by Monitor and Control Subsystems
- (7) Accommodation of other data sources, as follows
  - (a) Spacecraft test data via JPL Compatibility Test Area (CTA 21).

- (b) Spacecraft prelaunch data via Merritt Island, Florida, Spacecraft Compatibility Monitor Station (STDN, MIL 71)
- (8) DSCC Test Support Subsystem will use mission-independent hardware and software

### C. System Usage

Since the system capabilities can be used to support various DSN activities, the following are provided

- (1) Calibrations and prepass readiness verification
- (2) Real-time diagnostics and fault isolation
- (3) DSN implementation activities and performance testing of DSN systems, DSCC subsystems, and NOCC subsystems.
- (4) DSN operational verification tests to prepare for mission support.
- (5) Flight project ground data system tests, mission simulations, and operational readiness tests.

## II. System Implementation

### A. Description

The Mark IVA DSN plan establishes a Signal Processing Center (SPC) at each of the three Deep Space Communications Complexes. In each SPC, a new DSCC Test Support Subsystem will be implemented. Included within this subsystem are two assemblies, Telemetry Simulation Assembly (TSA) and Maintenance Support Assembly (MSA), that provide major support capabilities for the DSN Test Support System. Also, within the SPC, a new Monitor and Control Subsystem will be implemented that will provide control for operational activities. The backup Monitor and Control Subsystem will be used to support new system performance test programs.

In addition to the major implementation described in the above paragraph, additional changes will be implemented in other DSCC subsystems that support the DSN Test Support System. The test support functions provided by all DSCC subsystem are described in Section III.

### B. New Capabilities

This paragraph lists some of the new support features that are different from those provided by the Mark III-77 Test and Training System as described in Ref 2.

- (1) DSCC Test Support Subsystem. In support of the telemetry simulation and conversions functions the DSCC Test Support Subsystem will

- (a) Provide status to Mission Control and Computing Centers (MCCC) during remote simulation tests
- (b) Provide capability to store and replay project-generated telemetry data
- (c) Provide project tables that contain the data parameters for establishing unique spacecraft telemetry test signals
- (d) Generate delayed-telemetry test signals
- (e) Generate telemetry test signals to support the High Earth Orbiters mission that will be tracked by the Mark IVA DSN
- (f) Provide test signal generation and distribution capability to support multiple telemetry equipment groups

- (2) Maintenance support. At the DSCCs, the Maintenance Support Assembly will provide maintenance support carts to interface with various subsystem controllers via maintenance interface connectors
- (3) System performance test support. With the implementation of a local area network as part of the DSCC Monitor and Control Subsystem, testing activities and operational activities can be supported concurrently. System performance test programs and procedures will be provided for each of the eight DSN systems, as part of the Mark IVA project.

### C. Schedule

The Mark IVA Network implementation plan is phased with the Goldstone facility, being completed in February 1985, followed by Canberra in May 1985 and Madrid in August 1985.

## III. Deep Space Communication Complex Functions

### A. DSCC Test Support Subsystem (DTS)

As a major subsystem of the Test Support System, the following functions are provided:

- (1) Telemetry simulation and conversion functions are shown in Fig 2. The Telemetry Simulation Assembly (TSA) provides the functions that include the various digital and analog capabilities as itemized in Table 1.
- (2) Maintenance Support functions include the following:
  - (a) Accept equipment assignments for maintenance



- (b) Maintain status of all equipment assigned for maintenance
- (c) Provide portable maintenance support equipment
- (d) Interface with subsystem-supplied diagnostics

## **B. Receiver-Exciter Subsystem (RCV)**

The RCV provides the following test support functions

- (1) Generation of simulated S-band and X-band downlink test carriers Specific capabilities are as follows
  - (a) 64-meter and 34-meter transmit-receive stations provide one X-band carrier and one S-band carrier.
  - (b) 34-meter receive-only stations provide a single X-band test carrier
  - (c) CTA-21 and MIL-71 provide the same capability as the 64-meter stations
- (2) Modulation of telemetry subcarriers and data streams from the TDA onto simulated test carriers
- (3) Variable attenuation of each test carrier signal level
- (4) Translation of S-band exciter uplink frequencies to S- and X-band downlink frequencies
- (5) Measurement of carrier power levels and modulation levels

## **C. Antenna Microwave Subsystem**

The Antenna Microwave Subsystem provides the following test support functions

- (1) Routing of simulated downlink carriers to masers and/or receivers
- (2) Mixing of simulated S-band downlink carriers

## **D. Transmitter Subsystem**

The Transmitter Subsystem includes provision for feeding the transmitter output into a dummy load to support Command System and Tracking System test operations

## **E. Frequency and Timing Subsystem**

The Frequency and Timing Subsystem provides the following support functions

- (1) Time code and reference frequencies
- (2) Generation and distribution of a simulated time signal which can be substituted for the true GMT input to the various DSCC subsystems This capability is provided to support realistic mission simulations in flight project testing and training activities

## **F. DSCC Digital Communication Subsystem**

The Digital Communications subsystem provides the following test support functions

- (1) Transmission of data blocks of simulated telemetry data and DTS control messages from the NOCC or MOC
- (2) Transmission of status and alarm messages from the DTS to NOCC and/or MOC
- (3) Transmission of system performance test data blocks and simulated response data to the NOCC for DSN systems testing and project testing and training
- (4) Provision of on-site loop-back of data blocks for testing

## **G. DSCC Monitor and Control Subsystem**

The Monitor and Control Subsystem (DMC) provides the following test support functions

- (1) Generating and transmitting control messages to the DTS and to all subsystem controllers during system performance testing and flight project training activities
- (2) Displaying DTS alarm and status information for DSCC operations personnel
- (3) Supporting system performance test programs for DSCC operations personnel
- (4) Providing DSCC subsystem status data to DTS upon request
- (5) Transferring control of DSCC subsystem controllers to DTS for maintenance support activities

## **H. DSCC Subsystem/Assembly Controllers**

All DSCC subsystem controllers and/or assembly controllers provide the following functions in support of the DSCC maintenance support capability

- (1) Diagnostics and test procedures to support the DSN maintenance program
- (2) Standard maintenance interface connector

## **IV. Ground Communications Facility Functions**

The DSN Test Support System utilizes the Ground Communications Facility Subsystems for communicating data and information between Network Operations Control Center (NOCC) or any Mission Operations Center (MOC) and the

Deep Space Communications Complexes, CTA-21, and MIL-71 The following capabilities are supported

- (1) Transmission of control messages, simulated telemetry, and simulated command data to any DSCC from NOCC or from any MOC Simulated telemetry data rates will be less than 30 kbps
- (2) On-site loop-back of test data for system performance testing
- (3) Voice communications for purpose of test coordination and training

## V. Network Operations Control Center Functions

Test support capabilities presently implemented in NOCC are as follows

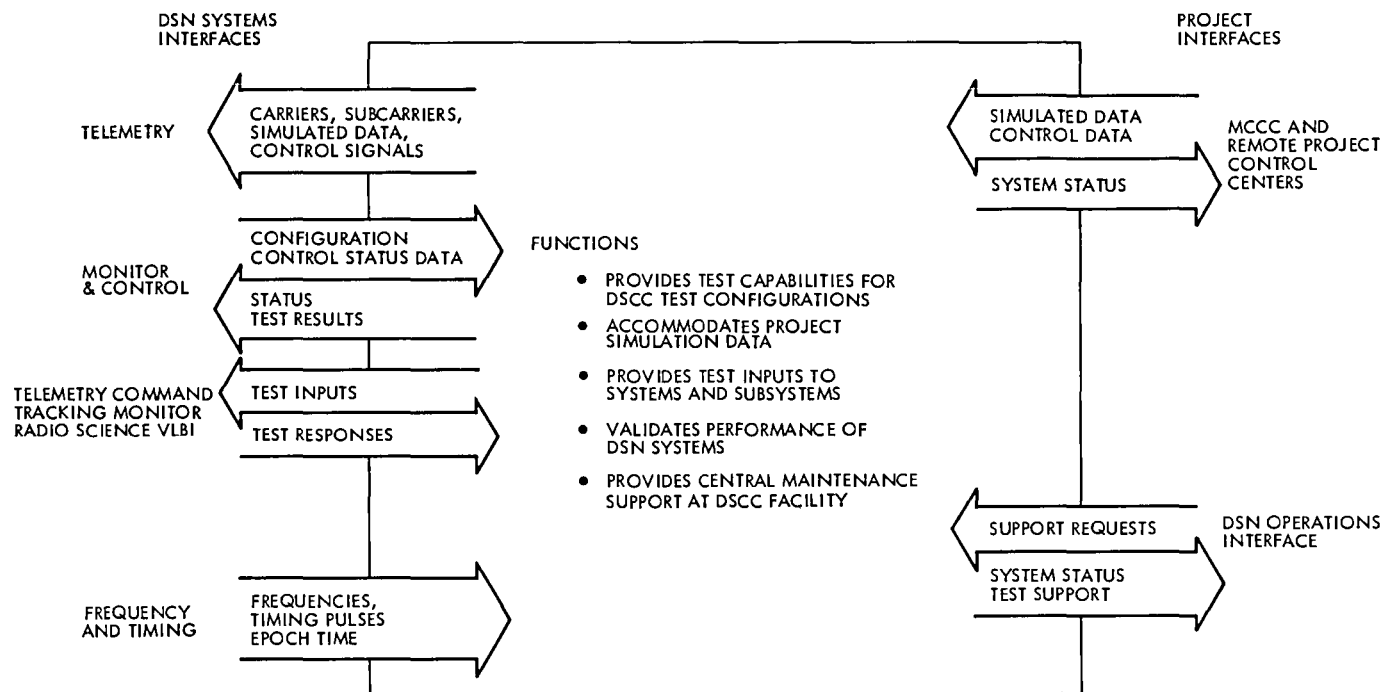
- (1) Off-line generation of recordings of high-speed data blocks for testing of the real-time monitors in the NOCC Tracking, Telemetry, Command, Radio Science, and Monitor and Control Subsystems
- (2) Selection of stored data blocks and output of the data to the DSS for system readiness verification

## References

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- 2 Herrman, D T , "DSN Test and Training System, Mark III-77," *DSN Progress Report 42-50*, pp 7-13, Jet Propulsion Laboratory, Pasadena, Calif , Feb 15, 1979

**Table 1 DSCC Test Support Subsystem digital telemetry simulation capabilities**

Functional capability	Description
Real-time data streams	6 independent channels
Variable data rates	6 SPS to 4 MSPS without subcarrier 6 SPS to 2 MSPS with subcarrier
Block-to-serial conversion	2 remote streams, combined rate 30 kbps
Block coding	1 channel, biorthogonal (32, 6)
Long constraint length convolutional codes	6 channels, $K=32, r=1/2, 1/3$ $K=24, r=1/2, 1/3$
Short constraint length convolutional codes	6 channels, $K=7, r=1/2, 1/3$
Selectable frame sizes	
Insert frame sync codes	Up to 48 bits
Increment data fields	Maximum of 3 fields per frame
Data patterns	Selectable per channel PN sequence, fixed formats, or project provided
Pulse coded modulation	Non-return-to-zero level (NRZL) or non-return-to-mark (NRZM) or bi-phase-L
Subcarrier frequency output	100 Hz to 2 MHz, 1/4 Hz resolution
Modulation index control	0 to 89 deg on each subcarrier
Subcarrier mixing	Maximum of two telemetry sub-carriers
Delayed test signals	Five channels can be delayed from a reference channel, delays are selectable from 0 to 128 microsec



**Fig 1. Test Support System, functiona and interfaces**

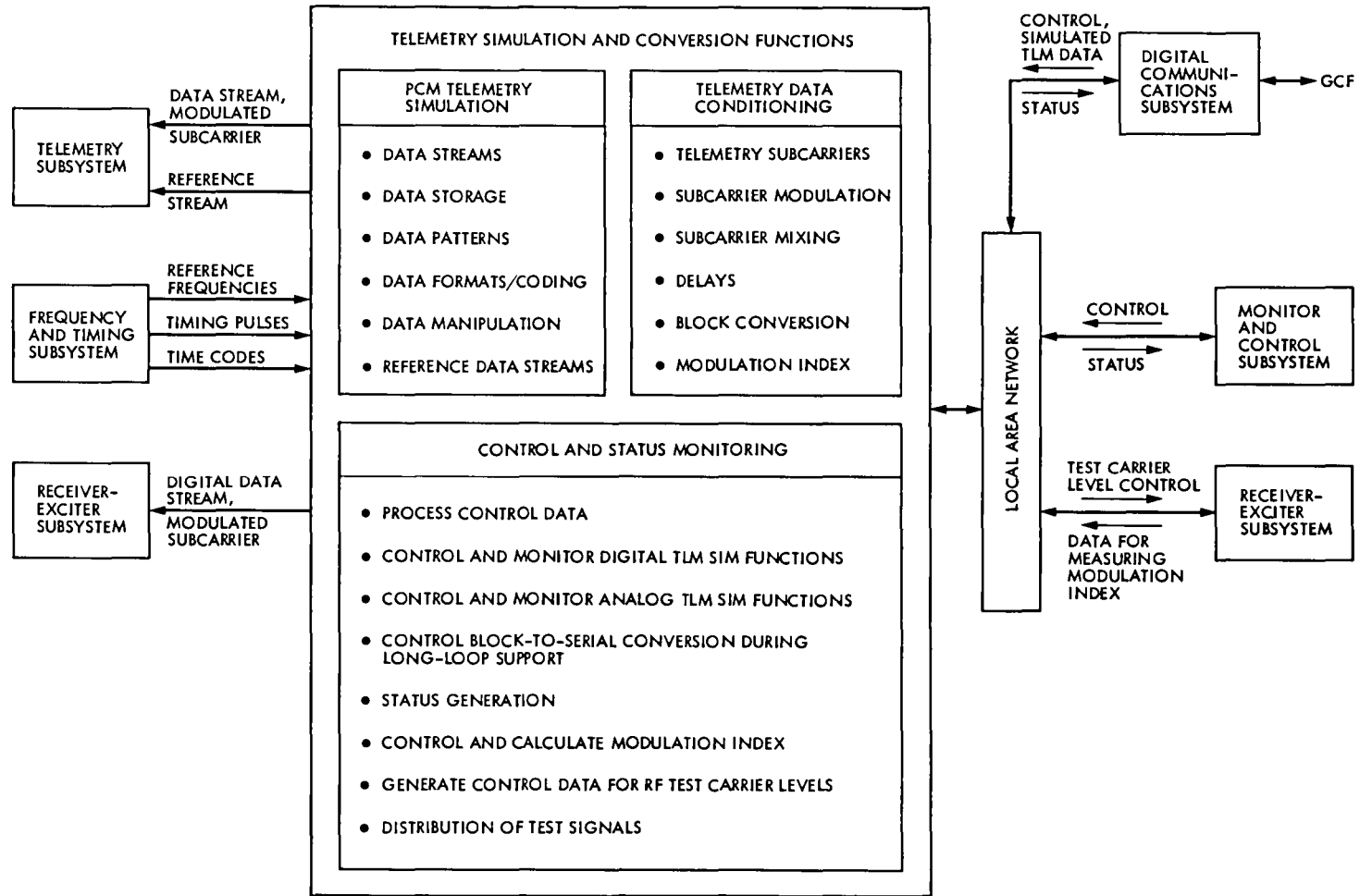


Fig 2 Telemetry functions, interfaces, and data flow

# An Overview of the Goldstone Energy Systems Study

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*The primary objective of the Goldstone Energy Systems Study (GESS) was to develop a systems planning methodology for analyzing and synthesizing Deep Space Network (DSN) energy systems. The resultant product, the Deep Space Network Planning and Analysis Methodology (DESPAM), can support DSN energy planning evaluations. It addresses a broad spectrum of tradeoff and dispatching scenarios. It evaluates a variety of energy generation configurations and also includes a capability to evaluate conservation measures.*

## I. Introduction

In recent years, energy costs have taken increasingly large portions of the Deep Space Network (DSN) annual operations budget. There exists the possibility that energy supplied to the stations could be cut off or reduced during times of world crisis. In response to these factors, several energy goals have been developed by NASA and the DSN. These include reducing energy consumption at all NASA installations by 50% from 1973 levels by 1985, minimizing life-cycle costs in balance with other costs through energy use reductions at the Goldstone Deep Space Communications Complex (DSCC), the attainment of a 90% non-utility, on-site energy generation capability, and an initiative to carry out demonstration projects in order to provide insight and data on existing low-risk technologies. The Office of Telecommunications and Data Acquisition (TDA) implemented a number of studies which evaluated DSN energy consumption as well as future options for alternative energy use. The Goldstone Energy Systems Study (GESS) was one of the TDA energy conservation projects.

The primary objective of GESS was to develop a systems planning methodology for analyzing and synthesizing DSN energy systems. The resultant product can support DSN short and long-term energy planning evaluations. It addresses a broad spectrum of strategic tradeoff and tactical dispatching scenarios. It evaluates a variety of energy generation configurations and also includes a capability to evaluate conservation measures. By implementing the methodology, the user is able to address the following types of issues:

- (1) The definition of optimal energy dispatch strategies for current operations (optimal can be lowest cost or other user-defined attributes such as minimum grid reliance)
- (2) The identification of conservation measures that would have the most favorable impact on operations and costs
- (3) The determination of the value of new energy generation sources. The identification of the best time for implementing new energy alternatives as well as the

dispatch strategy to be used and the quantity of the new source to use

- (4) The calculation of the net cost benefits of one energy generation scenario vs another. The calculation of the annual cash flows of varying scenarios

## II. Approach

The tool for assessing the energy planning questions is the DSN Energy System Planning and Analysis Methodology (DESPAM). It consists of several key components. The first is the characterization of new, alternative energy systems that have the potential to be implemented by the DSN in combination with existing generation sources such as grid and diesel. This representation includes performance curves, cost equations, reliability factors, and the selection of a timeframe for implementation which will have the best chance for lowering system costs. The next component relates to energy utilization by the DSN as well as physical conditions at the site. Here, DSN energy loads are characterized, meteorological conditions are specified, and any desired conservation measure is defined. The energy generation data and utilization parameters are then evaluated by implementing the DSN Energy System Simulation Model (DSNX). The next step of the DESPAM methodology is to use the DSNX model to evaluate actual test cases which replicate existing or potential energy generation scenarios of the DSN. The last step is to select those strategies and options which the user deems best for actual implementation.

The remainder of this article focuses on a description of the DSNX simulation model and on an end-to-end application of the DESPAM methodology. For a detailed discussion of the other components of the DESPAM methodology, see the Goldstone Energy Systems Study Final Report (an internal document).

## III. DSNX Description

As stated earlier, the main purpose of DSNX is to serve within the overall DESPAM methodology as a tool for analyzing various operating policy scenarios and to perform multi-year generation mix analyses with respect to energy utilization for the DSN. It does this by parametrically evaluating varying quantities of energy generation levels for different sources so that optimum configurations can be determined. In addition, it is possible to evaluate the effects of different timeframes on each scenario evaluated by making multiple runs (the multiple run aspect of the code is covered in the next section). The model has the capability to address new, emerging technologies such as solar thermal, wind, photovoltaics, or solar ponds

in conjunction with conventional utility grid or diesel-generated energy.

When the model addresses these new, emerging technologies, it is usually necessary to know insolation and meteorological data pertinent to a specific geographical location. This input can be provided to DSNX in the form of data tapes. The hourly energy generation level of any new technology selected by the user is then estimated. (A generalized flow chart of the entire model is shown in Fig. 1.)

The user then provides load data for the DSN facility in the form of data tapes, or he can create his own typical day load profile. It is also possible to adjust load data. Additionally, sizing parameters for each energy generation source are required. This includes any new energy technology to be implemented in addition to those sources currently available to the Goldstone DSCC, namely, utility grid and diesel (the diesel capability can also be used to model fuel cells). It is possible to address as many as four new energy technologies in addition to the utility grid and diesel. Grid or diesel or any alternative can also be evaluated individually.

The program then matches demand against energy availability for each source over a 1-year timeframe. The user specifies the dispatch strategy. Varying energy generation sizes and grid penetration levels can be evaluated within a single simulation run. At the user's option, the utility grid maximum usage level can change in accordance with the billing period. Only one diesel size per run is allowed.

The program determines alternative source energy generation, excess alternative generation, grid generation, excess grid capacity, diesel generation, total energy generation, unmet demand, and total demand over the entire period simulated. In addition, utility grid energy and power consumption levels are also determined on a monthly basis according to billing period. All outputs are determined for all configurations evaluated.

These outputs are then combined with user inputs of capital costs, maintenance and operations costs, fuel costs, subsystem replacement and/or rebuild, and utility billing charges, in order to determine one-time and recurring costs for each configuration under evaluation. In addition, inputs of system life, escalation rates, cost of capital, tax parameters, and a time frame for new technologies enable DSNX to determine life-cycle cost and levelized busbar energy cost for each configuration. The model also provides a complete yearly cash flow table. Lastly, a summary table for all configurations is given and a data file is formed which can be accessed from a Hewlett Packard Graphics Terminal for use in creating output plots.

#### IV. Implementation of DSN Energy System Planning and Analysis Methodology (DESPAM)

In the Approach Section it was explained that the overall methodology that covers all analysis components (characterization of generation systems, characterization of weather and load data, DSNX simulation, interpretation of results) is known as DESPAM. In this section, a typical end-to-end application of DESPAM is given

Let us suppose we want to evaluate a new technology (solar thermal) for use at the Mars site at the Goldstone DSCC. We want to look at varying penetration levels from 100 kWe to 1500 kWe. We want to integrate the new technology with the currently existing diesel-grid operation. Our criteria for the "best" system will be lowest life cycle cost. A secondary preference is to reduce reliance on grid-generated energy.

The Goldstone Energy Systems Study Final Report provides the following necessary input information:

- (1) The solar thermal system will have a point focusing distributed receiver and a centrally located steam Rankine engine. This system will be assumed to need no major overhauls during a 24-year life and there will be no performance degradation over time.
- (2) The new system will be implemented in 1988. It will be assumed to meet DSN reliability standards at that time.
- (3) Insolation data for Barstow, California, for 1976 will be used.
- (4) The Southern California Edison demand tape for 1980 for the Mars site will be used to characterize the energy load. During those periods when the diesels are in use (0 values on the tape), it will be assumed that 850 kWe of energy is used.
- (5) Diesel efficiency is 12 kWeh/gal. There is no initial capital cost (it is assumed that the diesels are already in place). Diesel M&O cost is \$ 003/kWeh. Diesel fuel price is \$ 915/gal. Diesel subsystems are overhauled after every 20,000 hours of operation at a cost of \$14,390.
- (6) The electric utility grid cost is based on actual 1982 bills (\$0 069/kWeh). There will be no sell-back of any excess generation of solar thermal energy to the grid (this energy is lost).
- (7) The nominal (including inflation) annual cost escalation rates are:

Inflation rate <sup>1</sup>	10 0%
Capital cost	9.3
M&O	8 9
Discount rate <sup>1</sup>	10 0
Fuel	10 4
Purchased electricity	8 4

We now need to make assumptions regarding the dispatch strategy, baseline cases, and technology size.

The dispatch strategy will be to use solar-generated energy first when available, the utility grid second, and diesel third. In the baseline case (no solar, grid vs diesel only), five different diesel penetration levels will be tested.

Scenario	Grid billing period		
	Off-peak	Mid-peak	On-peak
A	All grid	All grid	All grid
B	All grid	All grid	Grid up to 500 kWe, then diesel
C	All grid	Grid up to 1000 kWe, then diesel	All diesel
D	All grid	Grid up to 500 kWe, then diesel	All diesel
E	All grid	All diesel	All diesel

The current Goldstone DSCC operation of all grid, all grid, all diesel was not tested. However, Scenario C is very similar. These baseline scenarios will be simulated from 1982 to 2011 (Run 1).

The nominal solar thermal sizes tested are 100, 300, 1000, and 1500 kWe. These sizes will be tested in conjunction with the five grid-diesel combinations. It will be assumed that the grid-diesel combinations are used without solar from 1982-1987, and that solar is added in 1988 (construction takes place in 1987). This second case consists of two simulation runs, 1982-1987 (Run 2), and 1988-2011 (Run 3). Each run evaluates performance for a single year separately.

<sup>1</sup>When making multiple DSNX runs with differing timeframes, which this case requires, the economic calculations are simplified when inflation equals discount rate.



## V. Results

The lowest cost combination in Run 1 (diesel vs grid for 30 years) was Scenario C. Its life cycle cost in 1982 dollars is \$11.64 million. Its annual cash outflow is shown graphically in Fig. 2. This scenario derives 82% of its energy from the grid and 18% from diesel.

In Run 3 the lowest configuration is 300 kWe of solar combined with Scenario C for diesel and grid. Its life cycle cost is \$8.69 million as compared to \$8.98 million for the no-solar case over the same timeframe (1988-2011).<sup>2</sup> This best case only requires 72% of the energy to be generated by grid (16% comes from diesel and 12% is derived from solar thermal). At large solar penetration levels (1000 and 1500 kWe), reliance on the grid decreases but the monetary savings are lost now since

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<sup>2</sup>The Run 3 costs would have been somewhat lower had we assumed that excess solar generation was sold back to grid.

large portions of solar-generated energy were wasted (It appears that a solar size of about 500 kWe would give the lowest cost.) Figure 3 shows the results of Run 3 in terms of energy grid displacement as a function of life cycle cost. The cash flow for the lowest cost configuration of Run 3 (300 kWe of solar) is shown in Fig. 4. When the cash flow for the best scenario in Run 2 (Scenario C) is combined with the one from Run 3 and is compared to the baseline (Run 1), we get the net savings shown in Fig. 5. From 1982 to 1987 the net change is 0 since for those six years the baseline is replicated. The solar installation results in a negative net flow of about \$130,000 annually from 1988 to 1992, but turns positive from 1993 to 2011 and results in an overall net life cycle cost savings of about \$300,000. The energy generation breakdown for our optimal case from 1982 to 2011 is given in Fig. 6. We now know that it might be a wise idea to investigate solar thermal systems more fully. Furthermore, if it is possible to acquire a solar system as described, and if the remaining assumptions hold, the DSN may be able to cut grid reliance and reduce net overall operating costs at the same time.

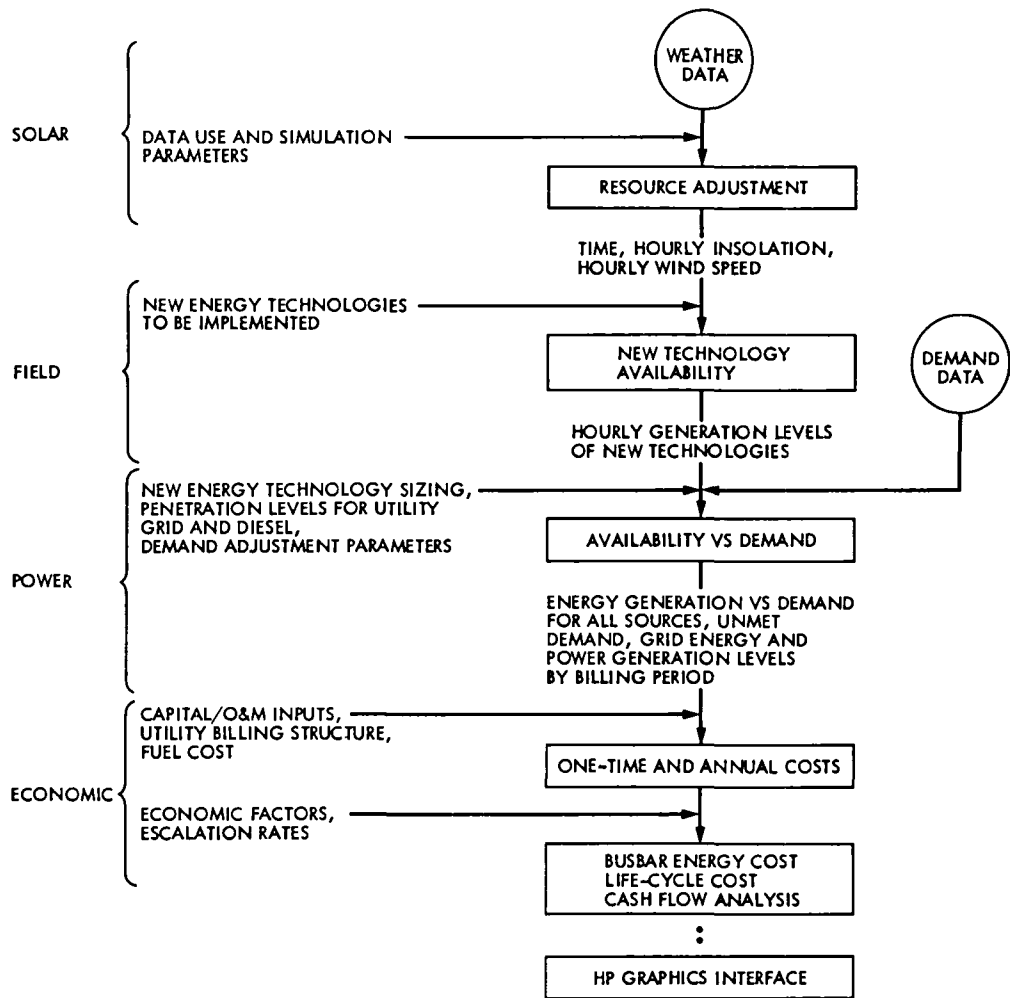


Fig 1 DSNX simulation sequence

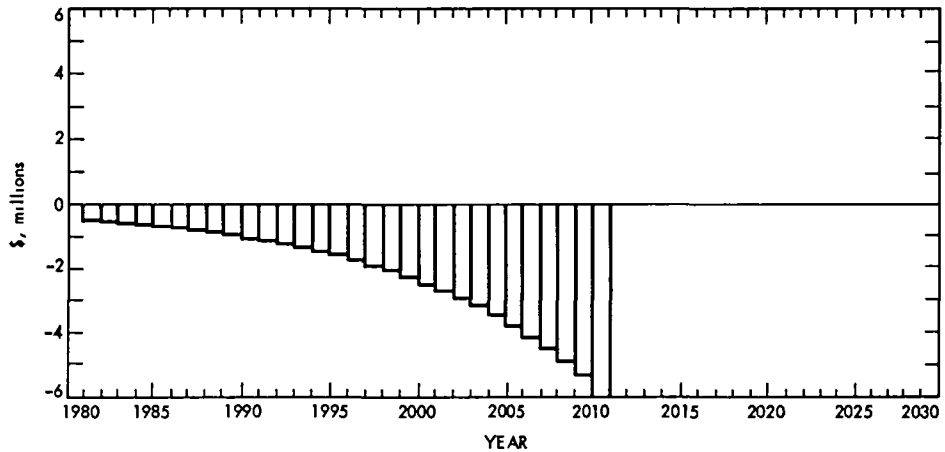


Fig 2 Net cash flow; diesel vs grid—Run 1, Scenario C

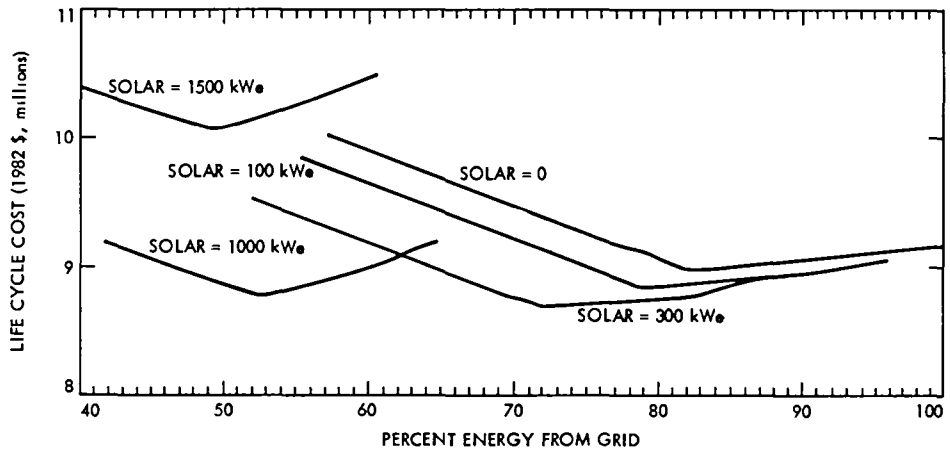


Fig 3 Grid displacement vs life cycle cost for various solar thermal/grid/diesel scenarios (Run 3)

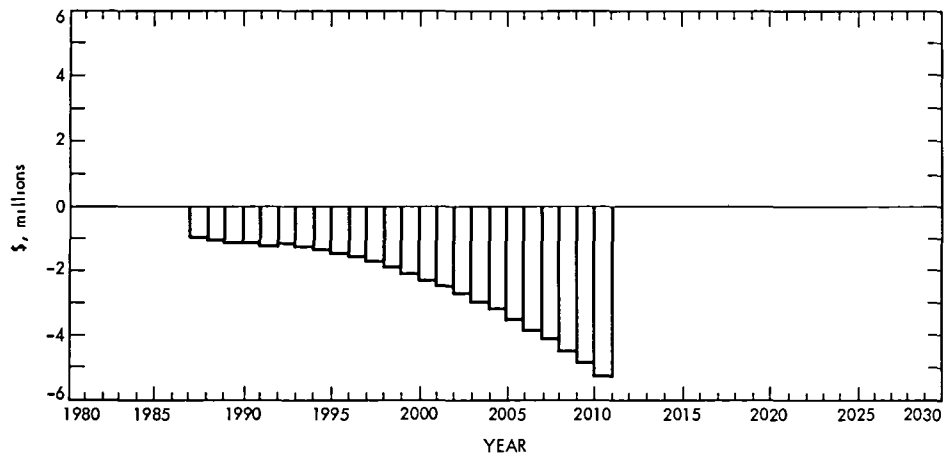
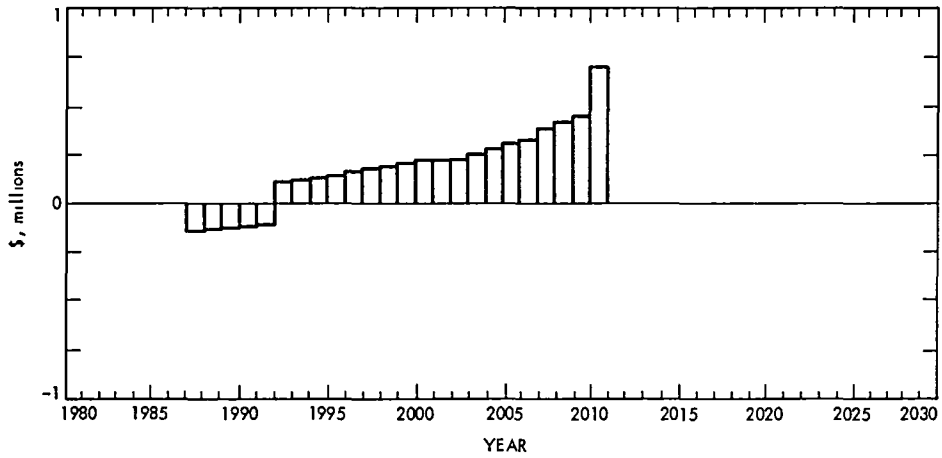
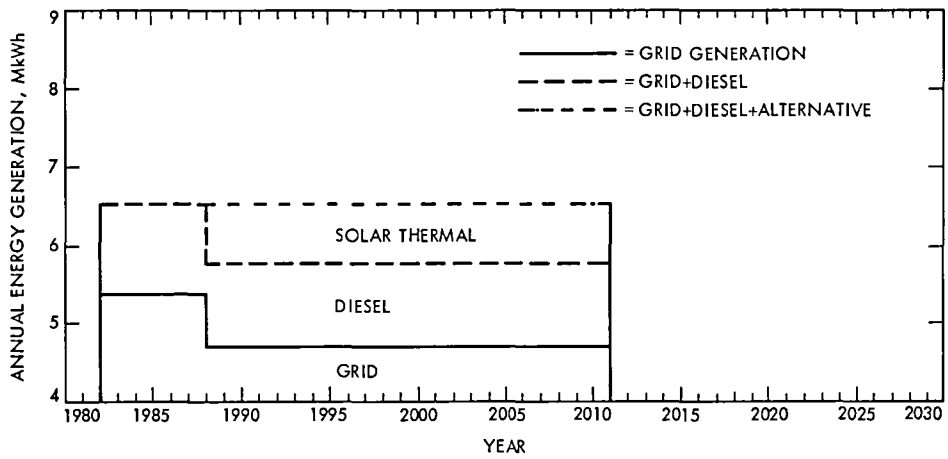


Fig 4 Net cash flow, Run 3--lowest cost configuration (solar power level = 300 kWe)



**Fig 5 Net cash flow; solar strategy vs nonsolar**



**Fig. 6 Energy generation by source (addition of solar to grid and diesel in 1988)**

# Performance Simulation of the JPL Solar-Powered Distiller

## Part I. Quasi-Steady-State Conditions

C S Yung and F L Lansing  
DSN Engineering

*A 37 85-m<sup>3</sup> (10,000 gallons) per year (nominal) passive solar-powered water distillation system has been installed and is operational in the Venus Deep Space Station at Goldstone, California. The system replaced an old, electrically powered water distiller. The distilled water produced with its high electrical resistivity is used to cool the sensitive microwave equipment. A detailed thermal model has been developed to simulate the performance of the distiller and study its sensitivity under varying environment and load conditions. The quasi-steady-state portion of the model is presented together with the formulas for heat and mass transfer coefficients used. Initial results indicated that a daily water evaporation efficiency of 30% can be achieved. A comparison made between a full day performance simulation and the actual field measurements gave good agreement between theory and experiment, which verified the model.*

### I. Introduction

A 37 85-m<sup>3</sup> nominal (10,000 gallons) per year passive solar-powered distillation (solar still) system has been operating successfully in the Venus station (DSS 13) of the Deep Space Network (DSN) tracking complex at Goldstone, California, since the initial installation in December 1981. Although this solar still system was intended to be an experimental system, only a few problems have been encountered. Those problems were minor and were corrected without impairing the operation of the system. Continuous observation of the installed system has indicated that it is highly productive, requiring only minimum maintenance. The potential application of similar solar-powered distillers in other DSN stations is being examined to replace present electrically powered distillation systems.

In order to complement the experimental investigation, a detailed analytical model of the solar still module is developed. This report presents the quasi-steady-state portion of the model to simulate the module performance in a simple, yet accurate manner. A computer program was written using the model characteristic equations to speed up the parameterization study.

A literature search was made earlier (Refs 1–4) to gather available information on existing simulation tools for solar-distillation systems. Dunkle (Ref 1) appeared to be a pioneer researcher on solar-stills. The steady-state solution was tried by Dunkle (Ref 1) and Morse and Read (Ref 2). Sayigh (Ref 3) presented a gross steady-state model of solar still, neglecting the internal heat exchanges and the influence of

makeup water Cooper (Ref. 4) presented a transient model where heat flow to the ground beneath the still was strongly emphasized. A finite-difference method was used to determine the temperature distribution below the ground surface. The multiple reflections between the glass and the water were not considered in the model.

In order to provide an accurate simulation of the JPL solar distiller, the model is developed based on the JPL design configuration (Ref 5) and includes all the influencing factors such as multiple reflections effect, internal heat exchange, and makeup water.

## II. Design Details

Figure 1 illustrates the design details (Ref 5) of the JPL solar distillation module. The system consists of 8 modules. The still module consists of a molded fiberglass tray with glass glazing on top sealed with special silicone rubber (RTV). The bottom of the still module is insulated with rigid foam and reinforced with marine plywood. The base of the wetted tray is painted black to serve as the solar absorber for evaporation enhancement. The remaining interior surface of the tray is colored white to promote reflection of sunlight into the water layer. The solar still is mounted on a welded steel frame elevated 3.35 m (11 ft) above ground level to facilitate the gravity feed and to eliminate shading caused by nearby structures. The solar still is oriented due south for maximum solar exposure.

Local well water with a high mineral content is supplied to the solar still by gravity from a storage tank located on a hill 91 m (300 ft) above the solar still inlet. The distilled water is collected in a 322-liter (85 gallon) temporary holding tank and then released to a large 5.7-m<sup>3</sup> (1500-gallon) underground storage tank for use in the microwave klystron-tube cooling system.

## III. Model Assumptions

The following assumptions are made for a one-dimensional thermal model.

- (1) The thermal losses from all four sides of the still module are neglected due to the presence of side air gaps. The side surfaces are painted a light color which enhances the reflection of the incoming solar radiation toward the water body. The sides temperature is assumed to be identical to the internal air temperature, with negligible heat loss to the ambient air.
- (2) The distillation mechanism is viewed as follows: natural convection takes place while the water vapor

is diffused uniformly from the water surface at  $T_w$  through the air space and condenses at the glass temperature  $T_g$ . The bottom surface of the still is painted black and absorbs most solar energy received.

- (3) Due to the small inclination angle of the glass cover, the optical effect of the system is modeled as two parallel transparent media with an opaque bottom surface for the bottom medium.
- (4) The internal air space is assumed to be 100% transparent to solar energy. No air leakage is assumed to take place to or from the solar still.

## IV. Quasi-Steady-State Model

Using the above assumptions, a quasi-steady-state model has been developed, the details of the heat balance equations are presented in Appendix A. Appendix B gives the details of the heat transfer coefficients used in the model. Appendix C gives the mass transfer and diffusion coefficients. Appendix D illustrates the model use by a numerical example. Since the heat capacitance of the solar still components is not included in a steady-state solution, only the glass cover, the water layer, and the bottom layer of the fiberglass tray are included. From Eqs (A 8), (A 15), and (A 17), the temperatures of these three components are expressed as

$$T_w = [(B_1 + B_2 + m_w C_w) T_g - (E_1 + B_1 T_0)] / (B_2 + m_w C_w)$$

$$T_f = (E_3 + h_{c,fw} T_w) / B_4 \quad (1)$$

$$T_g = (X_0 + X_1 T_0 + X_2 T_m) / (X_1 + X_2)$$

where the  $B$ 's are equivalent heat transfer coefficients and the  $X$ 's are characteristic constants that depend on the evaporation rate, thermal conductance, and the solar absorptivity of various components. Detailed descriptions of the  $B$ 's and  $X$ 's are given in Appendix A.

The nonlinearities in the radiation energy and water diffusion terms make the equations difficult to solve explicitly. An iteration method was used to obtain the final solution which uses a set of initially given temperatures  $T_g$ ,  $T_w$  and  $T_f$  for the glass cover, the water layer, and the fiberglass bottom, respectively. The temperature-dependent terms such as the Grashof number  $Gr$ , saturated vapor pressure  $P_w$ , latent heat of evaporation of water  $L$ , heat convection coefficients for air and water, and the linearized radiation heat transfer coefficients can be calculated. The new set of temperatures can be determined using Eq (1). By recomputing the temperature-dependent parameters using the new temperatures, a more

accurate set of temperatures can be determined, and the procedure is repeated until convergence is reached

## V. Evaporation Rate

Water evaporates isothermally from the top surface of the water layer at temperature  $T_w$  and diffuses upward through the stagnant air layer to the glazing. Water vapor condenses on the interior surface of the glazing at a lower temperature  $T_g$ . This diffusion process is caused by the differential saturated water vapor pressures in the air layer, at the boundary of the water layer, and at the boundary of the glazing. The evaporation rate of water, per unit still area  $m_w$ , is given in Appendix C as

$$m_w = \frac{DPM_w}{RT_a \delta_a} \ln \frac{(P - P_{wg})}{(P - P_{ww})} \quad (2)$$

where  $D$  is the diffusion coefficient ( $m^2/hr$ ),  $P$  is the system pressure ( $N/m^2$ ),  $M_w$  is the molecular weight of water,  $\bar{R}$  is the universal gas constant,  $\delta_a$  is the average air spacing between the water surface and the glass cover, and  $P_{wg}$  and  $P_{ww}$  are the saturated water pressure at the glass temperature and the water temperature, respectively

If the leakage of air and water vapor is assumed negligible, the amount of water produced by evaporation-condensation would be completely replenished by the makeup water at an inlet temperature  $T_m$ . Hence, Eq (2) represents the amount of water evaporation rate or the makeup water rate

## VI. Distillation Efficiency

The instantaneous (or steady-state) distillation efficiency based on a constant solar radiancy on the projected area is defined by

$$\eta = \frac{m_w L}{I} \quad (3)$$

where  $L$  is the heat of vaporization of water. In general, the evaporation efficiency increases with increasing the solar energy intensity where it reaches maximum efficiency at noon. Although the instantaneous (or steady-state) efficiency gives an indication of the system performance under constant environmental conditions, it is not a complete characterization of the distiller performance over a full day or a year period. The other alternative to an accurate transient simulation is to have a sequence of 24-hour quasi-steady-state simulation for each day. Accumulated daily efficiency can be ob-

tained by dividing the accumulated distilled water production up to time  $t$  by the accumulated solar incident energy up to time  $t$ , i.e.,

$$\eta(t) = \frac{\sum_0^t m_w(t) L}{\sum_0^t I(t)} \quad (4)$$

## VII. Simulation Results

In order to facilitate the simulation process, a computer program was written to speed up the analysis of the system under many different operating conditions. Appendix D gives a sample one-hour simulation. When the sample solar still module operates at 299.85 K ( $\sim 80^\circ F$ ) ambient temperature with 800  $W/m^2$  of solar intensity, it would yield 0.73 kg/hr (0.19 gal/hr) of distilled water. The temperature of the glass cover, water layer, and the bottom surface would be 317.1 K ( $111^\circ F$ ), 343.8 K ( $159.2^\circ F$ ), and 347.2 K ( $165.30^\circ F$ ), respectively. The steady-state or instantaneous efficiency would be 28.6%.

An experimental measurement program was conducted on the JPL solar still to determine its actual performance. Under a solar intensity of about 800  $W/m^2$  and  $32^\circ C$  ( $89.6^\circ F$ ) ambient temperatures, the measured water temperature was approximately  $71^\circ C$  ( $159^\circ F$ ). With this actual field measurement in good agreement with quasi-steady-state simulations, we proceeded using the model to determine the effect of the major solar still performance parameters.

In addition to an instantaneous performance simulation, a full-day simulation was performed for the month of June 1982 to predict the daily distilled water production rate. In the full-day simulation, the hourly solar intensities are calculated by using the ASHRAE model (Ref. 6) on an average day on June 21 at Goldstone, California (latitude is  $35^\circ N$ ). The outdoor air temperature profile was measured and the wind speed is assumed to be constant for the day at 4.47 m/sec (10 miles/hr). Since in a full-day simulation, steady-state conditions are assumed to be reached at the end of each hour of operation, the summation of the hourly water production rate would give the accumulated daily water production. The results of the full-day simulation of the JPL design indicated that the water production during 12 sunshine hours (from 6:00 a.m. to 6:00 p.m.) in June would yield 61.3 kg/day (16.2 gallons/day) of purified water vs the measured 64 kg/day (17 gallons/day), representing a 4.2% difference between the two values. Part of this deviation could be due to applying the quasi-steady-state method instead of the true transient method. Water production trails the solar intensity profile by a time-

lag Errors in estimating material optical and thermal properties are responsible for another part of the deviation. The idealization and the assumptions listed in Section III contribute to the rest of the deviation.

A graphical presentation of the daily performance of one solar-powered distillation module for the month of June is given in Fig. 2. As shown in this graph, the water purification rate reached a peak value of 1.11 kg/hr (or 0.53 kg/(hr-m<sup>2</sup>)) at noon, representing a maximum instantaneous efficiency of 35%. Since the measured daily temperature profile was higher

in the afternoon than in the morning, the module produced larger quantities of distilled water in the afternoon. The accumulated efficiency increased asymptotically to 29.7% at 2:00 p.m. and then dropped slightly to give a daily efficiency of 29.1% as a typical daily efficiency in June.

In addition to the above quasi-steady-state model, which provides accurate simulation, a transient model and a detailed sensitivity study are planned for the distiller. The results of the comparison between quasi-steady and transient simulations and parameterization will be presented in a future report.

## Acknowledgment

The authors are indebted to F. W. Stoller, D. Kuma, R. Z. Toukdarian, and F. Menninger for their technical management in the implementation of the JPL solar still at the Venus Deep Space Station. Acknowledgement is extended to R. Elder and J. May for providing valuable experimental data on weather and water purification rates.

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## List of Symbols

<p><math>A</math> surface area</p> <p><math>C</math> specific heat</p> <p><math>D</math> water vapor, air diffusion coefficient</p> <p><math>F</math> radiation view factor</p> <p><math>g</math> gravitational constant</p> <p><math>Gr</math> Grashof number</p> <p><math>h_c</math> convective heat transfer coefficient</p> <p><math>h_d</math> mass transfer coefficient</p> <p><math>H'</math> enthalpy of saturated water liquid</p> <p><math>H''</math> enthalpy of saturated water vapor</p> <p><math>I</math> solar radiation intensity</p> <p><math>k</math> thermal conductivity</p> <p><math>l</math> characteristic length</p> <p><math>L</math> latent heat of vaporization</p> <p><math>Le</math> Lewis number</p> <p><math>m</math> rate of makeup water or distilled water per unit area</p> <p><math>M</math> molecular weight</p> <p><math>Nu</math> Nusselt number</p> <p><math>P</math> pressure</p> <p><math>Pr</math> Prandtl number</p> <p><math>\bar{R}</math> universal gas constant</p> <p><math>Sc</math> Schmidt number</p> <p><math>Sh</math> Sherwood number</p> <p><math>t</math> time</p> <p><math>T</math> absolute temperature</p> <p><math>V</math> wind speed</p>	<p><math>X</math> constants</p> <p><math>\alpha</math> absorptivity</p> <p><math>\bar{\alpha}</math> thermal diffusivity</p> <p><math>\beta</math> volumetric expansion coefficient</p> <p><math>\delta</math> thickness</p> <p><math>\epsilon</math> emissivity</p> <p><math>\eta</math> efficiency</p> <p><math>\mu</math> dynamic viscosity</p> <p><math>\nu</math> kinematic viscosity</p> <p><math>\rho</math> reflectivity</p> <p><math>\bar{\rho}</math> density</p> <p><math>\sigma</math> Stefan Boltzman constant</p> <p><math>\tau</math> transmissivity</p> <p><math>\phi</math> extinction coefficient</p> <p style="text-align: center;">Subscripts</p> <p><math>a</math> airspace between glass and water</p> <p><math>c</math> convection</p> <p><math>e</math> effective</p> <p><math>f</math> bottom fiberglass</p> <p><math>g</math> glass</p> <p><math>i</math> bottom insulation</p> <p><math>o</math> ambient air</p> <p><math>r</math> radiation</p> <p><math>s</math> bottom plywood support</p> <p><math>w</math> water</p>
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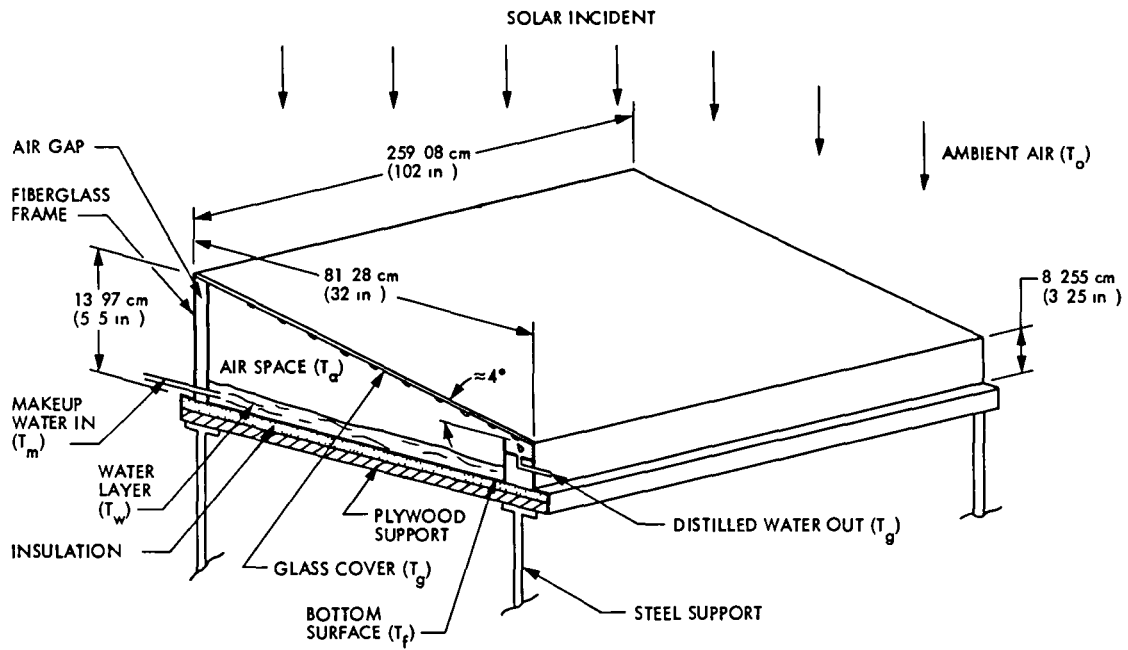


Fig 1 Layout of JPL solar-powered distillation module

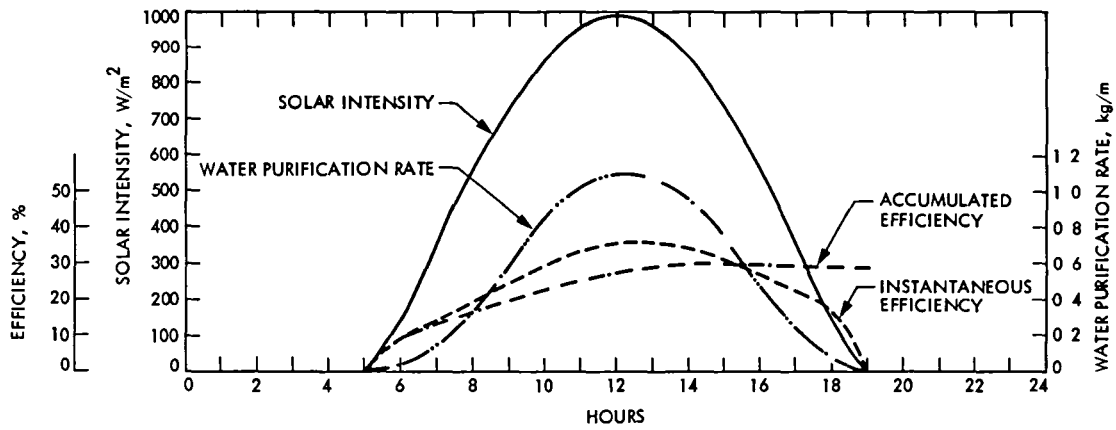


Fig 2 Daily performance of one module in June

## Appendix A

### Derivation of Quasi-Steady-State Equations

The derivations of the governing heat and mass transfer equations for the quasi-steady state conditions are made following the assumptions outlined in Section III. The heat transfer coefficients are given in Appendix B and the mass transfer coefficients are given in Appendix C. Figure A-1 depicts the energy flow to and from the major components the glass cover, the water layer and the bottom of the still.

#### A. Optical Properties

The optical effect of the glass cover and the water layer of the solar still can be modeled as two parallel transparent media with an opaque bottom surface for the second medium representing the black interior bottom of the still. The expressions of the "effective" absorptivity, reflectivity, and transmissivity due to multiple ray reflections are taken from Ref 7 as follows

For the glass cover, the effective absorptivity is

$$\left. \begin{aligned} \alpha_{g,e} &= \alpha_g + \frac{\alpha_g \tau_g \rho_w}{1 - \rho_g \rho_w} \\ \text{for the water layer} \\ \alpha_{w,e} &= \alpha_w \tau_g / (1 - \rho_g \rho_w) \\ \text{for the bottom surface} \\ \alpha_{f,e} &= \alpha_f \tau_g / (1 - \rho_g \rho_w) \end{aligned} \right\} \quad (\text{A-1})$$

where the single-layer absorptivities  $\alpha_g, \alpha_w, \alpha_f$  for the glass cover, water layer, and the bottom surface, respectively, are given by the expressions

$$\left. \begin{aligned} \alpha_g &= (1 - r_g)(1 - a_g) / (1 - a_g r_g) \\ \alpha_w &= (1 - r_w)(1 - a_w)(1 + a_w b) / (1 - a_w^2 b r_w) \\ \alpha_f &= a_w(1 - b)(1 - r_w) / (1 - a_w^2 b r_w) \end{aligned} \right\} \quad (\text{A-2})$$

where  $r_g$  and  $r_w$  are the ray reflection coefficient from the upper surface of glass and water respectively,  $b$  is the reflection coefficient of the opaque bottom surface of water and  $a$  is the ray absorption coefficient determined by

$$\left. \begin{aligned} a_g &= e^{-\phi_g \delta_g} \\ a_w &= e^{-\phi_w \delta_w} \end{aligned} \right\} \quad (\text{A-3})$$

where  $\phi$  is the extinction coefficient (in  $\text{m}^{-1}$ ). Note that the reflectivities for the glass cover  $\rho_g$  and for the water layer  $\rho_w$  are computed from

$$\left. \begin{aligned} \rho_g &= r_g + \frac{r_g a_g^2 (1 - r_g)^2}{(1 - r_g^2 a_g^2)} \\ \rho_w &= r_w + [a_w^2 b (1 - r_w)^2] / (1 - a_w^2 b r_w) \end{aligned} \right\} \quad (\text{A-4})$$

By using Eqs (A-2) and (A-4), the transmissivity of the glass cover  $\tau_g$  must satisfy the energy equation

$$\tau_g = 1 - \rho_g - \alpha_g \quad (\text{A-5})$$

#### B. Heat Balance Equations

Figure (A-1) depicts the energy flux to and from the still components. At steady-state conditions the heat balance equations are divided as follows

1 **The glass cover** By selecting a control volume as shown in Fig A-2, the energy balance equation is

$$\begin{aligned} \alpha_g I + m_w H''_{T_w} + (h_{c,wg} + h_{r,wg})(T_w - T_g) \\ = (h_{c,go} + h_{r,go})(T_g - T_o) \left( \frac{A_g}{A_w} \right) + m_w H'_{T_g} \end{aligned} \quad (\text{A-6})$$

where  $m_w$  is the mass flow rate of water per unit area, and  $I$  is the solar radiation intensity on the projected area of glass cover  $A_g$  (which is equal to the water surface area  $A_w$ ). The enthalpy terms can be combined as

$$H''_{T_w} - H'_{T_g} = L_{T_a} + C_w (T_w - T_g) \quad (\text{A-7})$$

where  $L_{T_a}$  is the latent heat of vaporization evaluated at the internal air space temperature  $T_a$ .  $T_a$  is taken as the average

temperature between  $T_w$  and  $T_g$ . Combining Eqs (A-6) and (A-7) gives

$$T_w = [(B_1 + B_2 + m_w C_w) T_g - (E_1 + B_1 T_o)] / (B_2 + m_w C_w) \quad (\text{A-8})$$

where

$$\left. \begin{aligned} B_1 &= (h_{c,go} + h_{r,go})(A_g/A_w) \\ B_2 &= h_{c,wg} + h_{r,wg} \\ E_1 &= \alpha_g I + m_w L_{Ta} \end{aligned} \right\} \quad (\text{A-9})$$

**2 The water layer** For the water layer, the energy equation can be derived by using the control volume as shown in Fig A-2 as

$$\alpha_w I + h_{c,fw}(T_f - T_w) + m_w H'_{Tm} = (h_{c,wg} + h_{r,wg})(T_w - T_g) + m_w H''_{Tw} \quad (\text{A-10})$$

where  $H'_{Tm}$  is the enthalpy of the make up water at  $T_m$ , and  $H''_{Tw}$  is the enthalpy of the saturated water vapor at  $T_w$ . The difference in enthalpy ( $H''_{Tw} - H'_{Tm}$ ) can be written as

$$H''_{Tw} - H'_{Tm} = L_{Tw} + C_w (T_w - T_m) \quad (\text{A-11})$$

where  $L_{Tw}$  is the latent heat of vaporization of water at  $T_w$  (or taken at  $T_a$  approximately) Equations (A-10) and (A-11) give the temperature  $T_f$  as

$$T_f = (B_3 T_w - B_2 T_g - E_2) / h_{c,fw} \quad (\text{A-12})$$

where

$$\left. \begin{aligned} B_3 &= h_{c,fw} + B_2 + m_w C_w \\ E_2 &= \alpha_w I + m_w C_w T_m - m_w L \end{aligned} \right\} \quad (\text{A-13})$$

**3 Bottom fiberglass surface** For the bottom fiberglass surface

$$\alpha_f I = h_{c,fw}(T_f - T_w) + h_{fo}(T_f - T_o) \quad (\text{A-14})$$

or

$$T_f = (E_3 + h_{c,fw} T_w) / B_4 \quad (\text{A-15})$$

where

$$\left. \begin{aligned} B_4 &= h_{c,fw} + h_{fo} \\ E_3 &= \alpha_f I + h_{fo} T_o \end{aligned} \right\} \quad (\text{A-16})$$

By manipulating Eqs (A-8), (A-12), and (A-15), the temperature  $T_g$  can be expressed as

$$T_g = (X_0 + X_1 T_o + X_2 T_m) / (X_1 + X_2) \quad (\text{A-17})$$

where

$$\begin{aligned} X_0 &= (m_w L + \alpha_g I) \left[ \frac{h_{c,fw} h_{fo}}{h_{c,fw} + h_{fo}} \right] \\ &+ I(B_2 + m_w C_w) \left[ \alpha_g + \alpha_w + \left( \frac{\alpha_f h_{c,fw}}{h_{c,fw} + h_{fo}} \right) \right] \end{aligned} \quad (\text{A-18})$$

$$X_1 = (B_1 + B_2 + m_w C_w) \left[ \frac{h_{c,fw} h_{fo}}{h_{c,fw} + h_{fo}} \right] + B_1 (B_2 + m_w C_w)$$

$$X_2 = m_w C_w (B_2 + m_w C_w)$$

where the latent heat  $L$  is taken at the air space temperature  $T_a$

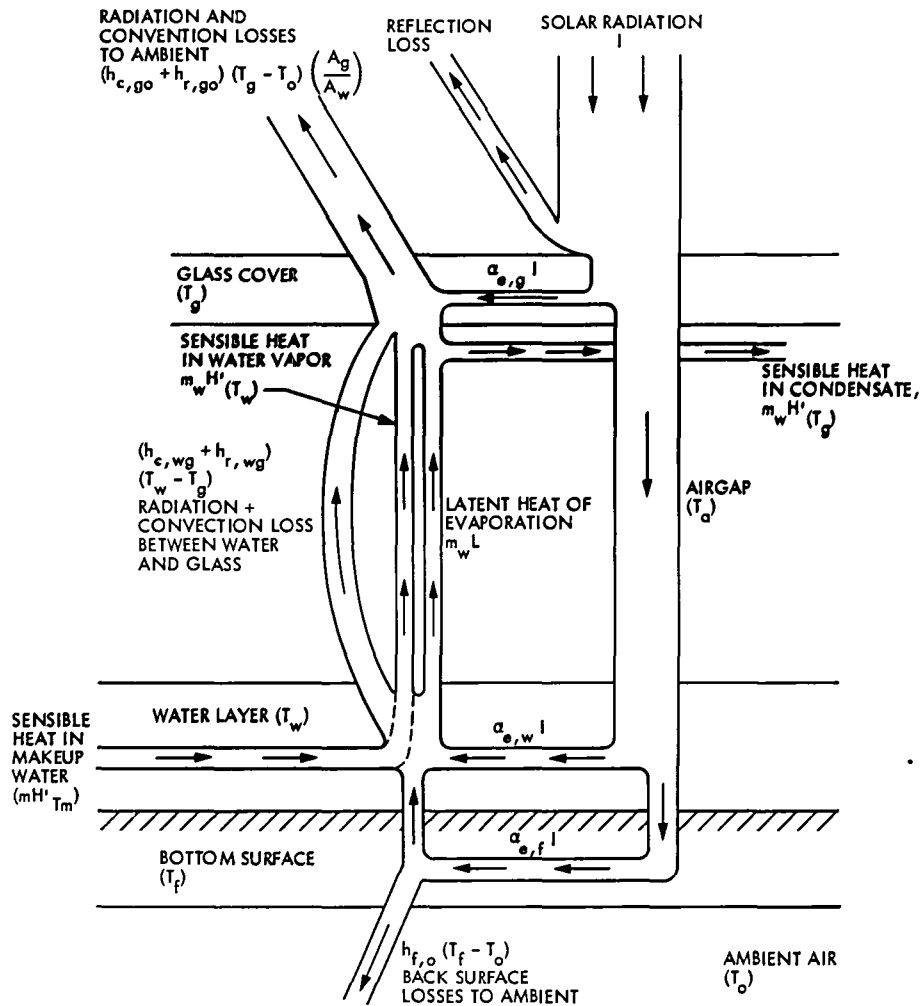


Fig A-1. Energy flux through still components

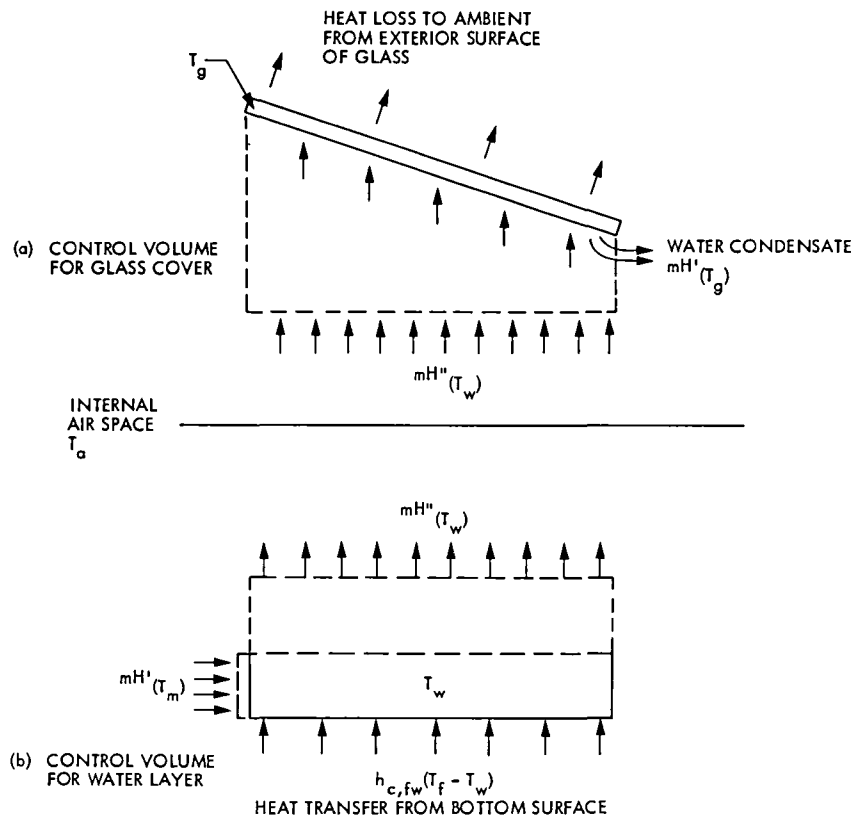


Fig A-2. Selected control volume for energy equations

## Appendix B

### Heat Transfer Coefficients

The development of the temperature distribution of the major solar still components as given in Appendix A uses a set of mass transfer, convection, and radiation heat transfer coefficients. The heat transfer coefficients are grouped as follows

**1 Radiation coefficient between glass cover and ambient air  $h_{r,go}$**  The coefficient  $h_{r,go}$  is obtained by linearizing the expression of radiation exchange between the glass cover (taken as a gray surface with emissivity  $\epsilon_g$ ) and a black ambient air (treated as a black body), hence

$$h_{r,go} = \epsilon_g \sigma (T_g^4 - T_o^4) / (T_g - T_o)$$

or

$$h_{r,go} = \xi_g \sigma (T_g^2 + T_o^2) (T_g + T_o) \quad (\text{B-1})$$

where  $\sigma$  is Stefan-Boltzmann constant ( $= 5.729 \times 10^{-8} \text{ W} / (\text{m}^2 \cdot \text{K}^4)$ )

**2 Radiation coefficient between glass cover and the water layer  $h_{r,wg}$**  The radiative coefficient  $h_{r,wg}$  is obtained by linearizing the expression of radiation exchange between the glass cover and the water layer, treated as two gray surfaces

$$h_{r,wg} = \epsilon_{wg} \sigma (T_w^4 - T_g^4) / (T_w - T_g)$$

or

$$h_{r,wg} = \epsilon_{wg} \sigma (T_w^2 + T_g^2) (T_w + T_g) \quad (\text{B-2})$$

where the effective emissivity  $\epsilon_{wg}$  is given by Holman (Ref 8)

$$\frac{1}{\epsilon_{wg}} = \left( \frac{1 - \epsilon_g}{\epsilon_g} \right) \left( \frac{A_w}{A_g} \right) + \left( \frac{A_w}{A_g F_{wg}} \right) + \frac{1 - \alpha_w}{\epsilon_w} \quad (\text{B-3})$$

where  $F_{wg}$  is the radiation view factor between the water and glass surfaces. Denoting the top, bottom, and side surfaces of

the still module as shown in Fig (B-1) and making use of the view factors expressions between two perpendicular rectangles (Ref 9),

$$F_{2-4} = \frac{1}{2} \left( 1 + \frac{h_4}{w_2} - \sqrt{1 - \left( \frac{h_4}{w_2} \right)^2} \right) \quad (\text{B-4})$$

$$F_{2-3} = \frac{1}{2} \left( 1 + \frac{h_3}{w_2} - \sqrt{1 - \left( \frac{h_3}{w_2} \right)^2} \right) \quad (\text{B-5})$$

since surface 2 does not "see" itself,

$$F_{2-2} = 0$$

$$F_{2-1} = 1 - F_{2-3} - F_{2-4} = F_{wg} \quad (\text{B-6})$$

**3 Convection coefficient between warm water surface and cold glass cover  $h_{c,wg}$**  For a horizontal air space enclosed between a warm plate at the bottom and a cold plate at the top, Holman (Ref 8) gave the natural convection coefficient as

$$h_{c,wg} = 0.195 Gr_{\delta_a}^{0.25} K_a / \delta_a \quad 10,000 < Gr < 400,000 \quad (\text{B-7})$$

$$h_{c,wg} = 0.068 Gr_{\delta_a}^{0.333} K_a / \delta_a \quad Gr > 400,000$$

where  $Gr_{\delta_a}$  is the Grashof number for the air,  $\delta_a$  is the thickness of the air space, and  $k_a$  is the thermal conductivity of the air evaluated at temperature  $T_a$ .  $k_a$  in W/m-K is determined by a curve-fitted expression using tabulated data in Ref 10 as

$$k_a = 0.0003623 T_a^{0.7488} \quad (\text{B-8})$$

The Grashof number is computed with  $\delta_a$  as the characteristic length by

$$Gr_{\delta_a} = g \beta (T_w - T_g) \delta_a^3 / \nu_a^2 \quad (\text{B-9})$$



where  $g$  is the gravitational constant ( $9.81 \text{ m/sec}^2$ ),  $\beta$  is the volumetric expansion coefficient (equals  $1/T_a$  for an ideal gas at  $T_a$ ),  $\nu_a$  is the kinematic viscosity of air (in  $\text{m}^2/\text{sec}$ ) evaluated at  $T_a$ . An expression for  $\nu_a$  is obtained, also, by curve-fitting tabulated data from Ref 10 as

$$\nu_a = 9.253 \times 10^{-10} T_a^{1.709} \quad (\text{B-10})$$

**4 Convection coefficient between the warm bottom surface and the relatively cold water layer  $h_{c,fw}$**  The average natural convection coefficient between a horizontal flat plate and a liquid is given by Holman (Ref 8) as

$$h_{c,fw} = 0.27 k_w (Gr_l Pr_w)^{0.25} / l_w \quad (\text{B-11})$$

where  $k_w$  (in  $\text{W/m-K}$ ) is the thermal conductivity of water evaluated at  $T_w$ ,  $l$  is the characteristic length taken as the mean of the two sides of the rectangular bottom surface of the still, and  $Pr_w$  is the Prandtl number for water. By curve fitting tabulated data in Ref 10 one obtains

$$k_w = 0.557 + 0.0011799 (T_w - 273.15) \quad (\text{B-12})$$

and

$$Pr_w = -0.222365 + \frac{200.90503}{(T_w - 273.15)} - \frac{1115.8828}{(T_w - 273.15)^2} \quad (\text{B-13})$$

The Grashof number of water is computed from

$$Gr_w = 0.00018g (T_f - T_w) l^3 / \nu_w^2 \quad (\text{B-14})$$

where the volumetric expansion coefficient for water  $\beta$  is  $0.18 \times 10^{-3} \text{ K}^{-1}$ . The kinematic viscosity of water ( $\nu_w$ ) in  $\text{m}^2/\text{sec}$  is determined by curve-fitting tabulated data in Ref 10 as

$$\nu_w = 1.1853606 \times 10^{-10} + \frac{32.562357 \times 10^{-6}}{(T_w - 273.15)} - \frac{248.87299 \times 10^{-6}}{(T_w - 273.15)^2} \quad (\text{B-15})$$

**5. Convection coefficient between glass cover and the ambient air  $h_{c,go}$**  The forced convection heat transfer coefficient  $h_{c,go}$  in  $\text{W/m}^2\text{-K}$  is given by Duffie and Beckman (Ref 11) as

$$h_{c,go} = 5.7 + 3.8V \quad (\text{B-16})$$

where  $V$  is the wind speed in  $\text{m/sec}$

**6 Overall heat transfer coefficient from the bottom surface to the ambient air** As shown in Fig 1, the bottom portion of solar still is composed of three layers: a fiberglass layer, an insulation layer, and a plywood layer. A general expression for the overall heat transfer coefficient  $h_{fo}$  through this composite wall can be written as

$$\frac{1}{h_{fo}} = \frac{\delta_i}{k_i} + \frac{\delta_s}{k_s} + \frac{1}{h_{c,fo}} \quad (\text{B-17})$$

where  $\delta_i$  and  $\delta_s$  are the thickness of the insulation and the plywood support, respectively, and  $k_i$  and  $k_s$  are the thermal conductivity of the insulation and the plywood support, respectively. The convection coefficient between the bottom plywood surface and the ambient air is  $h_{c,fo}$

**7. Latent heat of vaporization for water** The latent heat of vaporization expression given by Cooper (Ref 4) in English units was converted to SI units as

$$L = 883 - 0.6687T \quad (\text{B-18})$$

where  $T$  is the absolute water temperature in kelvin and  $L$  is given in  $\text{Whr/kg}$

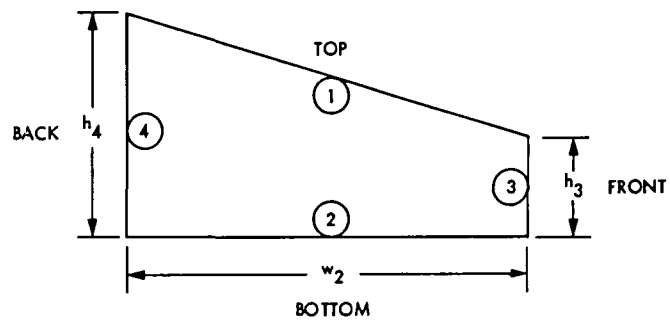


Fig B-1 Internal solar still radiation view factor

## Appendix C

### Mass Transfer and Diffusion

The mass transfer is significant in the model since it indicates the water purification rate and consequently the overall efficiency of the solar still. Two approaches for evaluating the mass transfer rate,  $m_w$ , are presented as follows:

#### I. First Approach: Stefan's Law of Diffusion

Following the Stefan's Law of Diffusion (Ref. 8), the total mass flow of water vapor per unit area evaporating from the upper water surface to the still air above is

$$m_{evaporation} = \frac{DPM_w}{\bar{R}T_a(\delta_a/2)} \ln \frac{P - P_{wa}}{P - P_{ww}} \quad (C-1)$$

where  $P$  is the total pressure for air-vapor mixture,  $\bar{R}$  is the universal gas constant (8.315 kJ/kmol-k),  $(\delta_a/2)$  is half the air spacing between the water surface and glazing,  $D$  is the water-air diffusion coefficient,  $T_a$  is the temperature of air spacing,  $P_{wa}$  and  $P_{ww}$  are the saturation vapor pressures for water at temperatures  $T_a$  and  $T_w$ , respectively. Also, for the condensation process at the glazing surface at  $T_g$ ,

$$m_{condensation} = \frac{DPM_w}{\bar{R}T_a(\delta_a/2)} \ln \frac{P - P_{wg}}{P - P_{wa}} \quad (C-2)$$

since at steady-state,

$$m_{condensation} = m_{evaporation} = m_w$$

By eliminating the pressure  $P_{wa}$  from Eqs. (C-1) and (C-2) the mass flow  $m_w$  for evaporation or condensation at steady state is

$$m_w = \frac{DPM_w}{\bar{R}T_a \delta_a} \ln \frac{P - P_{wg}}{P - P_{ww}} \quad (C-3)$$

where  $P_{wg}$  and  $P_{ww}$  are the saturated vapor pressure at temperature  $T_g$  and  $T_w$ , respectively. An empirical equation for the saturated vapor pressure for water in KN/m<sup>2</sup> or ( $kP_a$ ) is given by Kays (Ref. 12) from 10°C to 150°C as

$$P_w = 22105.8 \times 10^{-6} \left[ \frac{X(a+bx+cx^3)}{T(1+dx)} \right] \quad (C-4)$$

where  $T$  is the saturated water temperature (K) and

$$\left. \begin{aligned} a &= 3.2437814 \\ b &= 5.86826 \times 10^{-3} \\ c &= 1.1702379 \times 10^{-8} \\ d &= 2.1878462 \times 10^{-3} \\ x &= 647.27 - T \end{aligned} \right\} \quad (C-5)$$

The diffusion coefficient  $D$  between water vapor and air in cm<sup>2</sup>/sec is taken from Gilliland's semiempirical relation for any two gases  $A$  and  $B$  (Ref. 8) as

$$D = 577.7 T_a^{3/2} \sqrt{\frac{1}{M_A} + \frac{1}{M_B}} / [P(\bar{V}_A^{1/3} + \bar{V}_B^{1/3})^2] \quad (C-6)$$

where  $P$  is the total pressure of the air-vapor mixture in N/m<sup>2</sup>, and  $\bar{V}_A$  and  $\bar{V}_B$  are the molecular volume of  $A$  and  $B$ . For a mixture of air and water vapor the values of  $\bar{V}$  are 29.9 and 18.8, respectively. The molecular weights  $M$  for air and water vapor are 28.9 and 18, respectively. At atmospheric pressure  $P = 1.0132 \times 10^5$  N/m<sup>2</sup>, Eq. (C-6) is reduced to

$$D = 0.04977 (T_a/100)^{3/2} \quad (C-7)$$

Equations (C-3) and (C-7) are combined to give the distilled water production rate  $m_w$  in the text.

#### II. Lewis Number Approach

The mass transfer (or diffusion) of water vapor is the isothermal evaporation of water at  $T_w$  and the subsequent diffusion through the stagnant air layer above at  $T_a$ . By analogy to heat transfer, the mass transfer per unit surface area is expressed as

$$m_w = h_d(C_w - C_{wa}) \quad (C-8)$$

where  $h_d$  is the mass transfer coefficient,  $C_w$  and  $C_{wa}$  are the water vapor concentration (or density  $\bar{\rho}$ ) in saturated air-vapor mixture at temperature  $T_w$  and  $T_a$ , respectively

Assuming that water vapor at low pressure follows an ideal gas equation of state, the density  $\bar{\rho}_w$  is given by

$$\bar{\rho}_w = M_w P_w / (\bar{R} T) \quad (C-9)$$

where  $M_w$  is the molecular weight of water, and  $P_w$  is the partial pressure of water vapor in the saturated air layer above, hence

$$m_{evaporation} = h'_d (M_w / \bar{R} T_a) (P_{ww} - P_{wa}) \quad (C-10)$$

where  $h_d$  is an evaporation mass transfer coefficient, and  $P_{ww}$  and  $P_{wa}$  are the saturation pressures of water vapor at the temperatures  $T_w$  and  $T_a$ , respectively. Similarly, the condensation of water vapor taking place at the glass surface at temperature  $T_g$  is a result of the concentration difference between water vapor in two saturated air mixtures, one at temperature  $T_a$  and the other at a lower temperature  $T_g$ . An analogous expression to Eq (C-10) can be written for the condensing mass rate, as

$$m_{condensation} = h''_d (M_w / \bar{R} T_a) (P_{wa} - P_{wg}) \quad (C-11)$$

where  $h''_d$  is the mass transfer coefficient during condensation. At steady state, the make-up water will compensate for the condensate leaving the still, then

$$m_w = m_{evaporation} = m_{condensation}$$

Equating Eqs (C-10) and (C-11) and eliminating  $P_{wa}$ , one obtains the mass flow rate per unit surface area as

$$m_w = M_w (P_{ww} - P_{wg}) / \left[ \bar{R} T_a \left( \frac{1}{h'_d} + \frac{1}{h''_d} \right) \right] \quad (C-12)$$

Furthermore, if we assume that,

$$h'_d = h''_d = h_d \quad (C-13)$$

where  $h_d$  is the mass transfer coefficient at either the interface between the water and air space or between the air space and the glazing, then

$$m_w = (M_w h_d / 2 \bar{R} T_a) (P_{ww} - P_{wg}) \quad (C-14)$$

The mass transfer coefficient  $h_d$  is obtained following similar relationships between convective heat and mass transfer. These are given by the dimensionless numbers below

$$\left. \begin{aligned} \text{(a) Schmidt number } Sc &= \frac{\nu}{D} \text{ in mass transfer} \\ \text{Prandtl number } Pr &= \frac{\nu}{\alpha} \text{ in heat transfer} \end{aligned} \right\} \quad (C-15)$$

and

$$\left. \begin{aligned} \text{(b) Sherwood number } Sh &= \frac{h_d \delta}{D} \text{ in mass transfer} \\ \text{Nusselt number } Nu &= \frac{h_c \delta}{k} \text{ in heat transfer} \end{aligned} \right\} \quad (C-16)$$

where  $\bar{\alpha}$  is the thermal diffusivity ( $= k / \bar{\rho} c$ ),  $\nu$  is the kinematic viscosity ( $= \mu / \rho$ ), and  $D$  is the diffusion coefficient

If heat and mass transfer occur simultaneously as is the case in the solar still, the coefficients  $h_c$  and  $h_d$  are found to be related by Lewis number  $Le$  (Ref 8) as

$$\left. \begin{aligned} h_c / h_d &= \bar{\rho}_a C_a Le^{2/3} \\ Le &= \frac{\bar{\alpha}_a}{D} = \frac{Sc}{Pr} \end{aligned} \right\} \quad (C-17)$$

Knowing the value of Lewis number, the physical properties of air-water vapor mixture and the convective heat transfer coefficient  $h_c$  will enable the mass transfer coefficient  $h_d$  to be computed from Eq (C-14). This approach was abandoned, however, in favor of the first, simpler, approach which uses direct calculation of the diffusion coefficient  $D$ .

## Appendix D

### Numerical Example

A short computer program was developed to speed up the simulation of a solar still module using the equations presented in Appendices A, B, and C. The total water production rate of the distillation system can be obtained by multiplying the output of one module by the total number of modules in the system.

#### I. Input Variables

The variables used in the computer program are grouped as follows:

##### A. Optical Properties

Glass reflection coefficient <sup>1</sup> $r_g$	0.043
Water reflection coefficient <sup>2</sup> $r_w$	0.0201
Glass absorption coefficient <sup>3</sup> $a_g$	0.9730
Water absorption coefficient <sup>4</sup> $a_w$	0.7065
Bottom surface reflection coefficient <sup>5</sup> $b$	0.05
Emissivity of water surface $\epsilon_w$	0.9565
Emissivity of glass $\epsilon_g$	0.925
View factor $F_{wg}$	0.8533

- (1) Based on an index of refraction of 1.526
- (2) Based on an index of refraction of 1.33
- (3) Using an extinction coefficient of  $6.85 \text{ m}^{-1}$  and Eq. (A-3)
- (4) Using an extinction coefficient of  $6.85 \text{ m}^{-1}$  and Eq. (A-3)
- (5) Assuming an absorption coefficient of 0.95 for the opaque black-painted bottom

##### B. Module Geometry

Average spacing between water surface and the glass cover	6.00 cm
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Length of the still base	259.08 cm
Width of the still base	81.28 cm
Height of the front side of the still	8.25 cm
Height of the back side of the still	13.97 cm
Thickness of glass $X_g$	0.4 cm
Thickness of water layer $X_w$	5.08 cm
Thickness of fiberglass layer $X_f$	0.3175 cm
Thickness of insulation $X_i$	1.905 cm
Thickness of support wood frame $X_s$	1.905 cm
Surface area of glass $A_g$	2.111 $\text{m}^2$
Surface area of water $A_w$	2.1059 $\text{m}^2$

##### C. Thermodynamic Properties

Prandtl number of air $Pr_a$	0.7
Conductivity of insulation $k_i$	0.151 w/m-K
Conductivity of wood $k_s$	0.2942 w/m-K
Conductivity of fiberglass $k_f$	0.433 w/m-K
Specific heat of water $C_w$	1.163 Wh/Kg-K
Total pressure $P$	97.7 KN/ $\text{m}^2$ (KP <sub>a</sub> ) or 14.2 psi

##### D. Weather

Solar radiance $I$	800 W/ $\text{m}^2$
Wind speed $V$	4.47 m/sec (10 mph)
Ambient temperature $T_0$	299.85 K (80.1°F)

##### E. Temperatures

Makeup water temperature $T_m$	299.85 K (80.1°F)
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## II. Sample Results

The following is a partial list of the results printed by the computer using the input data presented above and the equations given in Appendices A, B, and C.

<sup>1</sup>Based on an index of refraction of 1.526

<sup>2</sup>Based on an index of refraction of 1.33

<sup>3</sup>Using an extinction coefficient of  $6.85 \text{ m}^{-1}$  and Eq. (A-3)

<sup>4</sup>Using an extinction coefficient of  $6.85 \text{ m}^{-1}$  and Eq. (A-3)

<sup>5</sup>Assuming an absorption coefficient of 0.95 for the opaque black-painted bottom

### A. Optical Properties

Effective absorptivity of glass $\alpha_{e,g}$	0.02805
Effective absorptivity of water $\alpha_{e,w}$	0.26693
Effective absorptivity of bottom $\alpha_{e,f}$	0.58950

### B. Thermal Properties

Conductivity of air $k_a$	0.02787 W/m-K
Conductivity of water $k_w$	0.6403 W/m-K
Kinematic viscosity of air $\nu_a$	$1.8664 \times 10^{-5} \text{ m}^2/\text{sec}$
Kinematic viscosity of water $\nu_w$	$4.1132 \times 10^{-7} \text{ m}^2/\text{sec}$
Prandtl number of water $Pr_w$	2.3988
Latent heat of evaporation at $T_a$ $L_{Ta}$	662.15 Whr/kg
Saturated vapor pressure at $T_g$ $P_{wg}$	8.91133 KN/m <sup>2</sup>
Saturated vapor pressure at $T_w$ $P_{ww}$	31.9833 KN/m <sup>2</sup>

### C. Heat Transfer Coefficients

Radiation coefficient between glass cover and ambient air $h_{r,go}$	6.2165 W/m <sup>2</sup> -K
Radiation coefficient between water layer and glass cover $h_{r,wg}$	6.3690 W/m <sup>2</sup> -K
Forced convection coefficient between glass cover and ambient air $h_{c,go} = h_{c,fo}$	22.686 W/m <sup>2</sup> -K
Natural convection coefficient between water layer and glass cover $h_{c,wg}$	2.503 W/m <sup>2</sup> -K
Overall heat transfer coefficient between the bottom surface and the ambient air $h_{fo}$	4.25549 W/m <sup>2</sup> -K
Natural convection coefficient between the bottom surface and the water layer, $h_{c,fw}$	81.245 W/m <sup>2</sup> -K

Combined heat transfer coefficients

$B_1$	28.9725 W/m <sup>2</sup> -K
$B_2$	8.8717 W/m <sup>2</sup> -K
$B_3$	90.5190 W/m <sup>2</sup> -K
Equivalent coefficient $B_4$	85.5011 W/m <sup>2</sup> -K

### D. Temperatures

Glass cover temperature $T_g$	317.1 K
Water layer temperature $T_w$	343.8 K
Bottom surface temperature $T_f$	347.2 K

### E. Performance

Water production rate per module	0.727 kg/hr
Quasi-steady-state efficiency	28.6%

From the energy balance equations presented in Appendix A, an energy flux distribution of the solar still can be developed. Figure D-1 shows the energy flux through the module. Assuming that 100 units of solar energy are radiated upon the still projected surface, 11.6 units of energy will be reflected back to the ambient air. The glass cover absorbs 2.8 units of energy and transmits the remaining 85.6 units to the still interior. The water layer absorbs 26.7 units of energy and the remainder is absorbed by the bottom surface. The makeup water for this example contributes 0.9 units of sensible heat loss from the solar still. The thermal loss of the top surface is more than twice that of the bottom. The back loss could be reduced further if the still is better insulated.

Water distillation rate depends mainly on incident solar radiation. To investigate this effect, the ambient temperature  $T_0$  was kept constant at 26.7°C (299.85 K) and several simulations were made using various solar intensities. The results are illustrated in Fig. D-2, showing a nonlinear relationship between the solar radiation intensity and the water purification rate. At 200 W/m<sup>2</sup> solar intensity the water distillation rate is approximately 0.05 kg/(hr-m<sup>2</sup>), which increases to 0.35 kg/(hr-m<sup>2</sup>) at 800 W/m<sup>2</sup>. The water distillation sensitivity is accelerated as the solar radiation becomes higher than 400 W/m<sup>2</sup>. Since the average daily solar radiation is higher in summer than in winter, the water distillation rate will be high in summer and low in winter, as shown in Fig. D-2.

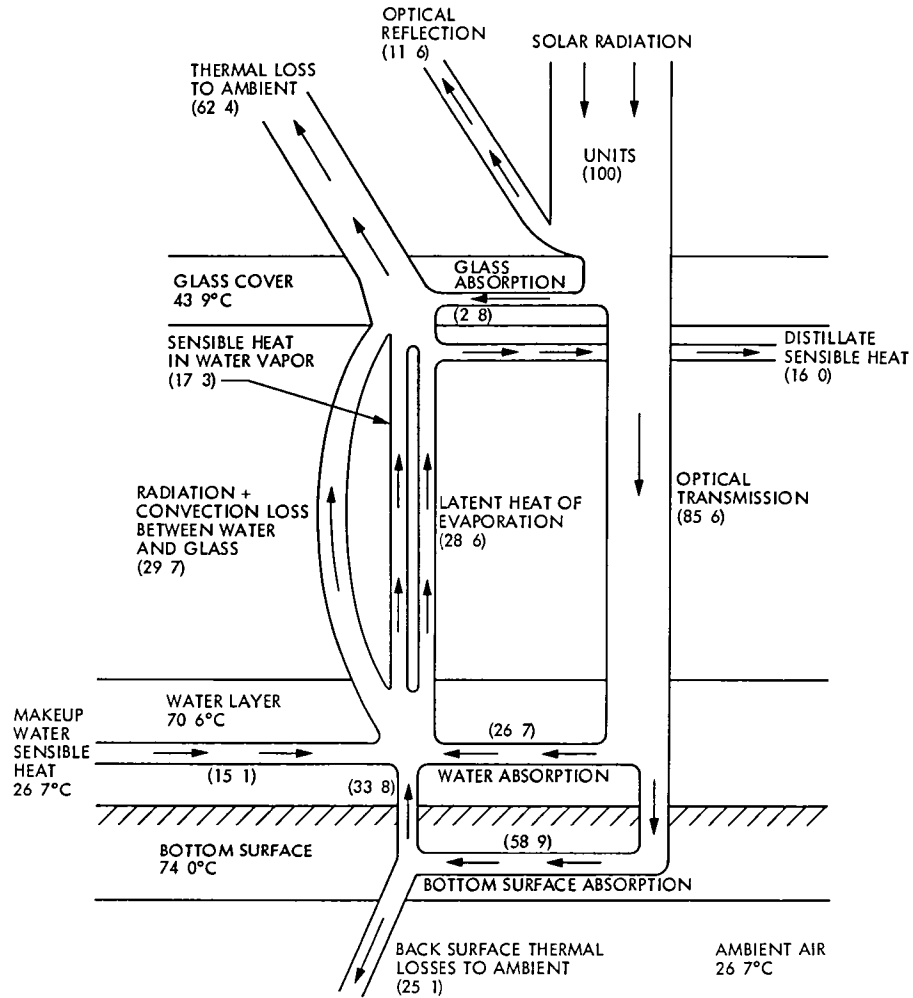


Fig D-1 Energy flux through still components for the sample 1-hour simulation

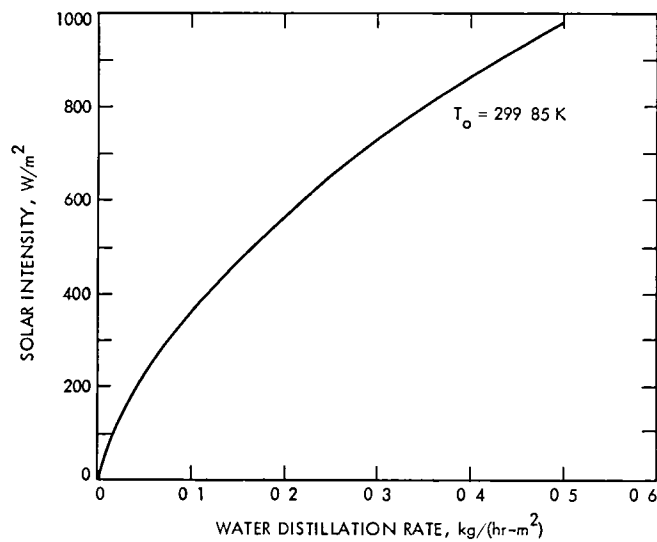


Fig D-2 Solar radiation versus water distillation rate

# Efficient Computerized Model for Dynamic Analysis of Energy Conversion Systems

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*In searching for the optimum parameters that minimize the total life cycle cost of an energy conversion system, one must examine various combinations of components and study the resulting system performance and associated economics. The Systems Performance and Economics Simulation computer program (SPECS) was developed to fill this need. The program simulates the fluid flow, thermal, and electrical characteristics of a system of components on a quasi-steady-state basis for a variety of energy conversion systems. A unique approach is used in which the set of characteristic equations is solved by the Newton-Raphson technique. This approach eliminates the tedious iterative loops which are found in comparable programs such as TRNSYS or SOLTES-1. Several efficient features were also incorporated such as the centralized control and energy management scheme, the analogous treatment of energy flow in electrical and mechanical components, and the modeling of components of similar fundamental characteristics using generic subroutines. Initial tests indicate that this model can be used effectively with a relatively small number of time steps and low computer cost.*

## I. Introduction

The satisfactory development and design of efficient and economical energy conversion systems require a thorough understanding of the interactions of the energy sources, system components, exterior environment, load distribution and automatic control devices. It is necessary to select components from among many proposed configurations that vary widely in performance characteristics and affect the overall system cost. Since experiments on new configurations are expensive and time-consuming, properly formulated computer simulations can provide the same information at much less time and cost.

As part of the continuing engineering efforts to upgrade the performance of the NASA-owned antenna facilities of the Deep Space Network (DSN) at JPL, it is desirable to obtain not only efficient "working" solutions but also "optimized" ones that minimize total system cost. Two steps should be taken in order to obtain optimum solutions. First, for a given energy conversion system configuration, the optimized performance parameters must be determined. Because the design configurations of components are often extensive (Ref 1), some bounds and size limits should be set on these parameters. Reliability and availability constraints are also added. Energy conservation measures, which reduce the energy consumption of the facility and subsequently affect



its operation and maintenance costs, must also be taken into account. Second, tradeoff studies of several optimized configurations must be made to select the optimum one.

At the supply side of an energy conversion system, energy is generated from a set of power plants, at the demand side, the energy is distributed among facility loads. It is essential, therefore, to match and balance, with minimum energy losses, the flow of energy between the generation and distribution sides. With the above energy balance, reliability, and cost constraints in mind, a detailed dynamic (i.e., time-varying) simulation of energy conversion systems is needed to analyze the system behavior and costs on a time-dependent basis.

It is common in system design to select equipment by assigning single values of key properties (such as pressures, temperatures, flow rates, voltages, currents, etc.) at critical network stations and select the appropriate size of components that meet given specifications. The more efficient dynamic simulation model, in contrast, assumes that the system configuration is defined, that all component sizes are selected and that the performance characteristics of all components are known. The simulation model then computes the station properties (pressures, temperatures, flow rates, voltages, and currents) at other than design loads. The model hence suits the following purposes:

- (1) It can be used as an analytical tool for equipment design for large energy conversion systems where the cost of computer simulation represents a small portion of the total cost. The program can assist the designer in avoiding oversized or undersized systems and in selecting optimum equipment size. After completing a trial component selection based on one set of loadings, one can apply the model to other partial or overloading conditions to determine if the original design is adequate.
- (2) It serves as a basis for developing "short cut" techniques after making a large number of detailed simulations on many different designs. These shorter solutions can be helpful in making initial assessments.
- (3) It provides the field operator with a learning and diagnostic aid for understanding the dynamic performance of the operating system and the interrelationships between components at off-design conditions.
- (4) It can be used to study the effects of add-on components and analyze the corrections sought for some operating problems if an existing system is expanded or reconfigured. The benefits of any modification can be assessed by analyzing the simulation results before and after making a change.

## II. Survey of Existing Programs

### A. Selection Criteria

The need to have such a time-varying simulation model has been encountered in many facility projects. The criteria described below are used in selecting the proper computer model in our literature search.

- (1) The model should be a well-documented and low cost tool with an easily understood input, incorporating basic and fundamental analytical expressions, and thereby being available to a wide variety of users.
- (2) The model should be modular and open-ended to simulate either small or large problems. The system size is determined by its number of components, and the model therefore should accommodate systems with arbitrary arrangements of series and parallel branches and have no limitations on these arrangements.
- (3) The model should have the capability to simulate steady-state conditions for a system without capacitive components as well as quasi-steady-state conditions for a system with capacitive components (such as storage tanks, batteries, etc.). Quasi-steady-state simulation is done with time steps that are much larger than the largest time constant of all components. A true transient simulation involves time steps that are smaller than the smallest characteristic time constant of all components. A true transient simulation should be avoided since most of the components encountered in facility energy conversion systems, excluding capacitive components, respond to abrupt changes in local conditions in a relatively short time. This time response varies from a few seconds to several minutes, which makes computations using large time steps (such as every half-hour or every hour) suitable for quasi-steady-state conditions. The quasi-steady-state approach reduces the computation time tremendously without a significant loss of accuracy (Refs 2, 3).
- (4) The model should treat mechanical and electrical components in an analogous fashion. Examples are a fluid pump vs an electrical generator, a pipe vs a connecting cable, a fluid storage tank vs a battery, etc. This analogous treatment will simplify the energy management and control schemes and reduce computer costs.
- (5) The model should have a central monitor and control software, as in real systems, to monitor changes in the system parameters and send the appropriate signals to respond to internal constraints as determined by the component's behavior or to external controls imposed by the user.

The validity of any program depends primarily upon the formulation of the equations which it solves. Fundamental laws and relations such as energy balances, momentum balances, mass balances, and properties of substances in combination with the performance characteristics of components must all be written in the proper form. In general, one method of solving the equations is not necessarily as effective as another in saving time and computer cost. Among the many programs written in system dynamics the following ones are examined.

## B. Available Programs

A detailed literature search showed that several computer programs which have been written for the dynamic simulation of solar-powered energy systems are relevant. TRNSYS (Ref 4) for example, was developed by the University of Wisconsin to simulate solar-powered water heating and cooling systems. Sandia Laboratories, Albuquerque, New Mexico, has developed SOLSYS (Ref 5), which is similar to TRNSYS but also includes simple Rankine-power cycles. Sandia later on developed SOLTES-1 (Simulator of Large Thermal Energy Systems (Ref 6)), which is a combined version of SOLSYS and TRNSYS. SOLTES-1 simulates, in a modular approach, Rankine-power cycles, solar powered heating and cooling systems. Though flexible and modular, TRNSYS and SOLTES-1 are limited to solar-powered systems only.

Other dynamic models such as LASL and CAL-ERDA were developed by Los Alamos Scientific Laboratory (LASL), Los Alamos, New Mexico (Ref 7). These are limited also to the dynamic description of a double-glazed flat-plate solar-collector, a hot water energy storage, an absorption water chiller, an air-handler, and a building shell and its interior loads for solar-heated and solar-cooled buildings.

In computer hardware real-time simulation, the Utility Systems Group at JPL (Ref 8) has developed an Electric Power System Simulator (EPSS) for the US Department of Energy (DOE) Office of Electric Energy Systems, which is coordinated with NASA's Energy Systems Division. The simulator is a microprocessor-based device which is designed to illustrate the complexities of operating a real utility power system that consists of several combinations of electrical power generation units. It is a portable, table-top device consisting of display and video monitors portraying the total system performance for different combinations of energy generation, energy storage, and load-varying effects. Costs of energy are user-provided for each component according to the load curves and operating characteristics.

W. Stoecker at the University of Illinois, Urbana-Champaign, (Ref 2), has developed a dynamic simulation using a different approach for centralized chilled water circuits. He has combined the Newton-Raphson technique for solving nonlinear

pressure balance equations with a simple Gaussian elimination technique for solving linear heat-balance equations. The extension of this approach to complex energy conversion systems employing phase-change fluids was not tried elsewhere, although it is worth investigating because of its simplicity and time-saving potential.

In addition to the above programs, other models were developed by independent groups: DYNOSIM by Honeywell Energy Resources Center, Minnesota (Ref 9), SOLCOST by Martin Marietta Corporation, Colorado, HISPER by NASA's Marshall Space Flight Center (MSFC), Alabama, and STEPS by Colorado State University, Colorado (Ref 10). Differences in accuracies of modeling a given problem have varied from +2% to +20% (Ref 11), which shows the need for program unification and standardization.

After comparing the above available computer programs with their differences regarding handling of equations, completeness of documentation, operating time and cost, complexity of input data, and ability to be reconfigured to accommodate additional components, it was decided that writing a new program instead of modifying an existing one was necessary in order to obtain an engineering tool which would have the desired capabilities and efficiency. The Systems Performance and Economics Simulation Program (SPECS) was developed to fill this need. This article describes the following aspects of the SPECS computer program: the method used for modeling a system of components, the mathematical representation of these components, the numerical technique used in solving the resulting system of equations, and the program input and output.

## III. Modeling Theory

Since an energy conversion system is a set of components interconnected to accomplish a specific task, it is by definition modular. It is possible to simulate its performance by collectively simulating the performance of its components. A large problem can be reduced to a number of smaller problems, each of which can be solved independently. Mathematically, the problem consists of constructing a set of nonlinear equations and solving these equations simultaneously using the Newton-Raphson technique instead of the commonly used consecutive approach. This approach avoids iteration of configurations with cyclic (closed loops), acyclic (open ended) and multicyclic (joint loops) calculations. The simultaneous solution process is repeated at each "selected" time step during a representative day of each month in the simulation period. A detailed economic analysis of the system follows its performance evaluation to establish a base for further optimization study. Additional features of the program are presented in the remainder of this section.

## A. Definitions

It is necessary to define the terms “station,” “streamline,” and “component,” since each has a definite meaning in the program description

A “station” is a unique position on the electrical or fluid network. A station is defined by 2-digit integers (e.g., 27), a maximum of 99 stations can be defined

A “streamline” is the path of energy (or fluid) flux between two adjacent stations. A streamline is identified by two stations, one upstream and one downstream

A “component” is the smallest constituent of the system which performs a well-defined physical function when assembled with other components. Components as shown in Fig. 1 can be simple, with one streamline (one inlet station and one outlet station) or compound, with two or more streamlines (one or more inlet stations and one or more outlet stations). Components are distinguished by two identifiers and described by a set of performance parameters. The first identifier is for the component “type” which is a 2-digit integer (a maximum of 99 types could be identified) specifying whether the component is a pump, a pipe, a 2-way valve, a water tank, a heat exchanger, a wind turbine, etc. The second identifier is a component “position” identifier giving the relative component location in the system and its associated links with other stations and streamlines. No two components can have the same position identifier. A particular component type can appear in more than one place as needed to complete the function of the system. Additional component description parameters include size and flow characteristics and efficiencies or control strategies as described by the program inputs

## B. Newton-Raphson Technique

The Newton-Raphson technique (Ref. 2) used in the SPECS program is based on the linear approximation of the set of nonlinear functions  $[F(X)] = 0$  around the root  $[X^*]$  such that

$$\left. \begin{aligned} f_1(X) &= f_1(x_1, x_2, \dots, x_N) = 0 \\ &\vdots \\ f_N(X) &= f_N(x_1, x_2, \dots, x_N) = 0 \end{aligned} \right\} \quad (1)$$

or in matrix form

$$\begin{aligned} [F] &= [f_1, f_2, \dots, f_N] \\ [X] &= [x_1, x_2, \dots, x_N], \end{aligned} \quad (2)$$

where  $[X]$  is the vector of unknown variables. The Jacobian matrix  $[J]$  is the “slope” matrix whose element  $J$  in the  $i$ th row and  $j$ th column is the derivative

$$J_{ij} = \partial F_i / \partial x_j \quad (3)$$

At the  $n$ th iteration the Taylor series expansion for  $[F]^{(n)}$  around  $[X]^{(n)}$  gives the following matrix equation

$$\begin{aligned} [F]^{(n+1)} - [F]^{(n)} &= \left[ \frac{\partial F}{\partial x} \right]^{(n)} \{ [X]^{(n+1)} - [X]^{(n)} \} \\ &+ 0 [\Delta X]^2 \end{aligned}$$

By truncating the series to drop terms of the order  $\Delta X^2$  and higher and evaluating the derivative matrix at the  $n$ th iteration

$$[F]^{(n+1)} - [F]^{(n)} \cong [J]^{(n)} \{ [X]^{(n+1)} - [X]^{(n)} \} \quad (4)$$

or

$$[X]^{(n+1)} \cong [X]^{(n)} + [J]^{-1(n)} \{ [F]^{(n+1)} - [F]^{(n)} \} \quad (5)$$

At an  $(n+1)$ th iteration, which is assumed to bring the trial  $[X]^{(n+1)}$  close to the root  $[X^*]$ , the function  $F^{(n+1)}$  approaches zero and the non-zero residual ( $[F]^{(n)}$ ) is used to obtain the updated value of  $[X]^n$ . Equation (5) then gives

$$[X]^{(n+1)} \cong [X]^{(n)} - [J]^{-1(n)} [F]^{(n)} \quad (6)$$

Note that if  $[X]^{(n)} \neq [X^*]$ , the function set  $[F]^{(n)}$  will not equal zero, rather, it equals a “residual” which tends to be close to zero as the iteration proceeds. Instead of inverting the Jacobian matrix, a modified Gauss-Jordan elimination technique is used to solve Eq. (6)

Convergence is accepted when the condition

$$| [X]^{(n+1)} - [X]^{(n)} | \leq [\epsilon] \quad (7)$$

is met or when  $n$  exceeds a user-specified upper limit. The value of the variable  $[X]$ , computed at the end of one time step, is used as the initial guess for the iteration at the next time step and the above process is repeated until the simulation period is completed

Note that the variables sought at the end of each time step can be classified as (a) “equilibrium” variables that are free to take on any value during the time step and therefore they are determined by the program after satisfying all conservation

laws, and (b) externally controlled (or source/sink) variables that must be kept at a fixed value during each iteration step as determined by the user's input controls. The structure of the Jacobian matrix includes only the first type of variables

### C. Equations and Unknowns in a Fluid Network

The set of nonlinear equations needed to describe a fluid network can be obtained by characterizing each fluid station in the fluid network by a vector  $[V]$  which is composed of seven elements

$$[V] = [\dot{m}, P, h, T, v, s, x] \quad (8)$$

where  $\dot{m}$  is the fluid mass flow rate,  $P$  is the absolute pressure,  $h$  is the specific enthalpy,  $T$  is the absolute temperature,  $v$  is the specific volume,  $s$  is the specific entropy, and  $x$  is the quality (or dryness fraction). These seven fluid properties constitute  $7S$  unknowns for a fluid network that consists of  $S$  number of fluid stations. Energy is transferred across the boundary of each component in the network in three possible forms: heat,  $\dot{Q}$ , electrical work,  $\dot{W}_e$ , and mechanical (or shaft) work,  $\dot{W}_m$ , thus adding  $3C$  unknowns for  $C$  components. The  $(7S + 3C)$  equations needed to solve these  $(7S + 3C)$  unknowns are grouped as follows

- (1) For each fluid station where the fluid is a pure substance, only two independent thermodynamic properties are needed out of the six properties  $p$ ,  $h$ ,  $T$ ,  $v$ ,  $s$ , and  $x$ . The thermodynamics-property subroutines are designed to compute the remaining four thermodynamic properties once two properties are specified. The property subroutines thus act as  $4S$  implicit equations for a network of  $S$  fluid stations.
- (2) The continuity and momentum equations form a set of  $2(S - 1)$  equations for  $S$  fluid stations in a closed fluid loop. These equations are written along streamlines only. The number of equations depends upon the structure of the fluid network, two equations for each streamline minus two equations for closing the loop.
- (3) If the first law of thermodynamics is applied to each component boundary or control volume, a maximum set of  $C$  equations for  $C$  components can be obtained. Additional component equations (equals  $2C$ ) can be obtained from the second law of thermodynamics for the efficiency or effectiveness expressions and from the energy constraints for each component (such as no  $\dot{Q}$ , no  $\dot{W}_e$ , or no  $\dot{W}_m$ ). A total of  $3C$  equations are thus obtained.
- (4) The remaining  $(S + 2)$  equations can be written to control the properties at certain fluid stations by assigning

values to temperature limits (for two-phase flows), pressure limits, quality limits, or mass flow limits. Control equations can be used to define load management as needed by the control software.

### D. Structure of the Jacobian Matrix and Component Equations

Although the Jacobian matrix  $[J]$  could take on various arrangements, the one sketched in Fig. 2 is one option. The elements  $(\partial F_i / \partial x_j)$  are grouped so that the seven variables of the first fluid station are followed by the variables of the second station and so on until the  $S$ th station is reached. Next to the station property variables, the derivatives of three variables ( $Q$ ,  $W_e$ ,  $W_m$ ) of the first component are placed, followed by the three variables of the second component and so on until the  $C$ th component is completed. Other arrangements can be adopted if found more advantageous.

Only components with a time constant much smaller than the iteration time step are included in the Jacobian structure since these components are treated as having Steady-State, Steady-Flow (SSSF) processes. Any capacitive component with a large time constant (such as a fluid storage tank) will be dealt with separately at the end of each time step, where the component is treated as having Uniform-State, Uniform-Flow (USUF) processes.

To avoid having discontinuous function derivatives, especially for the  $4S$  implicit thermodynamic relations, the SPECS program includes a special treatment explained as follows. Once two thermodynamic properties ( $p$  and  $v$ , say) are known at the  $n$ th iteration step, the other corresponding four properties (in this case  $h$ ,  $s$ ,  $T$  and  $x$ ) are determined by the thermodynamics property subroutine. These four properties are treated as "fixed" or constants throughout the  $(n + 1)$ th iteration step. At the end of the  $(n + 1)$ th iteration, new values for the two properties ( $p$ ,  $v$ ) are obtained and the process is repeated.

If a fluid property  $y$  is assigned a temporary constant value  $y_0$ , an equation can be written to reflect this process

$$y - y_0 = 0 \quad (9)$$

Accordingly, the element  $J(i, j)$  of the Jacobian matrix corresponding to  $i$ th equation and the  $j$ th fluid property will be equal to 1 and all the other elements in the  $i$ th row become zero. If desired, these fluid properties, which are assumed to be temporarily fixed, could be excluded from the set of unknown equilibrium variables in the Jacobian, hence reducing the size of the Jacobian matrix. Additional tests, however, are

still required to determine whether these fixed variables in equations similar to Eq (9) may save computation time

As an illustration of the structure of the Jacobian matrix, consider the continuity, momentum and energy equations written for a compound stationary component ( $\sigma$ ) with multiple inlet stations ( $i$ ) and outlet stations ( $o$ ) as sketched in Fig 3

For USUF conditions, the continuity equation becomes

$$\sum_i \dot{m}_i(t) - \sum_o \dot{m}_o(t) = \frac{m_\sigma(t + \Delta t) - m_\sigma(t)}{\Delta t} \quad (10)$$

The momentum - impulse vector equation at time  $t$  becomes

$$\begin{aligned} \sum_i \dot{m}_i [U_i] + \sum_i [P_i A_i] - \sum_o [P_o A_o] \\ + [m_\sigma g] + [R] = \sum_o \dot{m}_o [U_o] \quad (11) \end{aligned}$$

where [ ] indicates a vector of multiple components [R] is a boundary-surface force which could be negative (as a friction force opposing the flow) or positive (as a result of forced agitation by a blower or a pump)

The energy equation is written as

$$\begin{aligned} \dot{Q} + \sum_i \dot{m}_i \left( h_i + \frac{U_i^2}{2} + gZ_i \right) - \dot{W}_m - \dot{W}_e \\ - \sum_o \dot{m}_o \left( h_o + \frac{U_o^2}{2} + gZ_o \right) \\ = \frac{E_\sigma(t + \Delta t) - E_\sigma(t)}{\Delta t} \quad (12) \end{aligned}$$

where  $E_\sigma$  and  $m_\sigma$  are the internal energy and retained mass of the control volume  $\sigma$ , and  $t$  is the time. Equations (10 through 12) are used for USUF components only. The velocity  $U$  could be further expressed in terms of the mass flow rate as

$$U_i = v_i \dot{m}_i / A_i$$

$$U_o = v_o \dot{m}_o / A_o$$

For SSSF components which appear in the Jacobian matrix, Eqs (10) through (12) are reduced to the following

Continuity equation

$$\sum_i \dot{m}_i - \sum_o \dot{m}_o = 0 \quad (13)$$

Momentum equation for a streamline

$$P_i - P_o - (a_0 + a_1 \dot{m} + a_2 \dot{m}^2) = 0 \quad (14)$$

where the pressure change ( $R/A$ ) is expressed by a polynomial in  $\dot{m}$ , and  $a_0, a_1, a_2$  are constants determined by the component type. Energy equation for a SSSF component or a control volume

$$\begin{aligned} \dot{Q} + \sum_i \dot{m}_i \left( h_i + \frac{U_i^2}{2} + gZ_i \right) - \dot{W}_m - \dot{W}_e \\ - \sum_o \dot{m}_o \left( h_o + \frac{U_o^2}{2} + gZ_o \right) = 0 \quad (15) \end{aligned}$$

The coefficients derivatives of Eq (13) appear in the Jacobian matrix as 1, corresponding to an inlet station mass flow rate, -1 to the outlet station mass flow rate, and zero otherwise. Other simple relations can be derived from Eqs (14) and (15) in a similar manner.

#### IV. Program Structure

The SPECS program is structured in a modular fashion which will allow future modification of any aspect of the program. It consists of four blocks: (1) an input block, (2) a library block, (3) a control block, and (4) an output block as shown in Fig 4. The input block contains the user-provided data. The library block contains the subroutines which perform various analytical and numerical computations. The library block computes all parameters for a given system configuration according to the data and specifications given in the input block. The control block monitors and manages the flow of information, solves the system of nonlinear equations at each time step, and controls the computational stability requirements. The output block prints for the user the final performance and economics results of a specific system. Note that the output block is controlled by user options for short, intermediate, or detailed printout.

The program language is FORTRAN IV. Use of FORTRAN "NAMELIST" is made to classify the input data into groups of common feature. Descriptions of program blocks and sub-routines are as follows

## A. Input Block

The input to the SPECS program is divided into seven major groups

- (1) System configuration data
- (2) Station properties data
- (3) Components characteristics data
- (4) Control data
- (5) Outdoor environment/site data
- (6) Energy loads data
- (7) Economics data

Each group is further described below

**1 System configuration data** This group defines the number, the type and location of components, the number and location of inlet and outlet fluid stations, the number of streamlines, etc. This information is used by the program to set limits of DO loops and determine the size of the problem under study

**2 Station properties data** This group includes the type of working fluid at each station, externally controlled properties, initial conditions and setpoints. All the pressures are computed relative to a user-assigned pressure at a reference point. The program also allows for differences in elevation of each station above some arbitrary datum

**3 Component characteristics data** This group includes the diameter and equivalent friction length for pipes and joints, the head-flow rate curve for a pump, the heat transfer coefficient times surface area product for a heat exchanger, the coefficient of velocity for a fully open valve, etc

**4 General control data** This group includes the case title, convergence criteria for termination of program, the time-step, the total simulation period, the type of results to be printed out for the user and options to read weather and load data

**5 Outdoor environment/site data** This group includes outside air dry-bulb temperature, outside air wet-bulb temperature, relative humidity, dewpoint temperature, wind speed and direction, site latitude, site elevation, and solar radiation components (direct normal, diffuse, and ground reflected) as

measured by a pyrometer or tracking pyrheliometer. A 24-hour profile for one representative day of each month is selected throughout a year's simulation. The solar radiation data can be taken from field measurements or otherwise simulated using the deterministic ASHRAE model (Ref. 12) which is built into the program

**6 Energy loads data** This group includes space and water heating load, space cooling load, lighting load and other electrical energy loads consumed by the facility. Water consumption data can also be included. Since it is impractical to run a large model for each hour of the 8760 hours of a year, an appropriate compactness is required to save computation time. Therefore, loads data for two representative days for each month are satisfactory, the first is a load profile for weekdays and the second is a load profile for weekends and holidays. Twenty-four hourly values for each representative day are required for energy loads

**7 Economics data** For detailed life cycle cost and cash flow analyses, this group includes the installation, operation, and maintenance costs of all components in the system, purchased energy costs, energy escalation rate, discount rate, and salvage values, in addition to preimplementation costs such as site preparation, surveying, study and design costs, etc

## B. Library Block

This block contains five sets of subroutines

- (1) Thermodynamic properties subroutines
- (2) Component description subroutines
- (3) Gauss-Jordan subroutine
- (4) Solar-radiation simulation subroutine.
- (5) Life cycle cost and cash flow subroutine

Each set is described as follows

**1 Thermodynamic properties subroutines** The driver routine for the thermodynamic properties computation is called by the Control Block to obtain values of fluid properties at any fluid station. This routine forms, from a library file, an array containing tables of saturation data only for each fluid used in the network being analyzed. The array does not contain data for all 16 fluids, listed in Table 1, which SPECS recognizes. This subroutine computes the values of four properties ( $p$ ,  $h$ ,  $s$  and  $x$ ), given the values of the two independent properties ( $T$ ,  $v$ ) for a fluid according to the flow chart in Fig. 5

If the fluid is in the 2-phase saturated region ( $v_f < v < v_g$ ) where the subscripts  $f$  and  $g$  denote saturated liquid and

saturated vapor, respectively, the driver routine computes the values of  $p$ ,  $h$ ,  $s$ , and  $x$  by interpolating according to  $T$  at constant  $v$ . If the fluid is in the compressed region ( $v < v_f$ ), the four properties are calculated based on approximations using the saturated liquid conditions.

If the fluid is in the superheated region ( $v > v_g$ ), general analytical expressions from Ref. 13 are used to calculate  $p$ ,  $h$ ,  $s$ , and  $x$  as a function of  $T$  and  $v$ . A set of fluid-specific subroutines calculates the two ideal gas properties ( $h^*$ ,  $s^*$ ) followed by the generalized expressions to compute the departure from ideal gas properties.

After the computation of all four properties is complete, the values are returned to the Control Block. Thus this subroutine acts as 4S implicit equations for a system having  $S$  number of fluid stations.

In addition to the six thermodynamic properties analyzed in this subroutine, three other physical properties are stored for each of the 16 fluids: the thermal conductivity,  $k(T)$ , dynamic viscosity,  $\mu(T)$ , and specific heat,  $C_p(T)$ . These three physical properties are given for both liquid and superheated vapor forms.

**2 Component description subroutines** This set is a library of component model subroutines from which the user can draw to configure the specific system under study. The components are divided into generic categories according to function: Mass, momentum, energy conservation equations, and performance effectiveness relations can be written regardless of differences in design, operating conditions, and working fluids. These differences are taken into account by assigning different values for similar parameters in each component. This process eliminates superfluous component models of similar type.

For example, a counterflow recuperator, a parallel flow recuperator, and a heat regenerator could be simply grouped under one component type: a heat exchanger in which heat transfers from one medium to another. The distinct behavior of each component, however, is accounted for by different analytical expressions for the effectiveness, heat capacities of exchanging media, number of heat transfer units, and the conductance heat transfer area product.

Distinction between mechanical components is made based on the number of streamlines and inlet and outlet stations as shown in Fig. 1. Electrical components are modeled by simple relations between  $\dot{W}_m$ ,  $\dot{W}_e$ , and  $\dot{Q}$ . Some of the generic component types to be modeled include but are not limited to the following:

- (1) A heat exchanger (boiler, recuperator, and regenerator with phase change in one or two media)
- (2) A fluid pump (rotary, screw, and positive displacement)
- (3) A compressor (for air, refrigerant, etc.)
- (4) A turbine (for steam, organic fluid, gas, wind, water)
- (5) A simple heater (for air or water) or a simple cooler without phase change of either medium
- (6) A solar collector (flat plate, parabolic trough, paraboloid concentrator, etc.)
- (7) A fluid storage tank (compressed air, water, steam accumulator, etc.)
- (8) Electrical battery
- (9) Electrical motor
- (10) Electrical generator or alternator
- (11) Photovoltaic cell (flat plate or concentrated type)
- (12) A mixing valve
- (13) A splitting valve
- (14) A pressure reducing valve
- (15) A fluid pipe (insulated or noninsulated)
- (16) An underground cable or electrical transmission line
- (17) Building structure
- (18) Air-handler for air conditioning (dual duct, variable air volume type, etc.)
- (19) A thermoelectric cell for power generation or refrigeration
- (20) A thermionic cell for power generation
- (21) A fuel cell
- (22) An absorption chiller (using lithium bromide-water, aqua-ammonia, etc.)

**3 Newton-Raphson iteration subroutine** This subroutine solves the system of nonlinear equations using the Gauss-Jordan reduction method with diagonal pivoting instead of directly inverting the Jacobian matrix. This subroutine is written in double precision arithmetic to overcome the problem of oscillatory convergence during iteration.

**4 Solar radiation simulation subroutine** Based on the deterministic model set by ASHRAE (Ref. 12), the total solar

radiation intensity (including the direct, diffuse, and ground-reflected components) falling on a surface is analytically simulated for the user's convenience. The solar elevation, azimuth, and declination angles are computed for 24 hours for one representative day of each month. The information available in this subroutine pertains to various solar collector types, selected as follows

- (1) Tilted flat plate, with arbitrary tilt angles
- (2) Horizontal flat plate
- (3) Single-axis tracking surface, with east-west orientation
- (4) Single-axis tracking surface with north-south orientation
- (5) Single-axis tracking surface with polar mount, i.e., the surface is oriented toward north-south line with a tilt angle to the horizontal which equals latitude
- (6) Two-axis tracking surface, always perpendicular to the sun's direct rays

For northern latitudes, the surfaces are oriented toward the south, and for southern latitudes, toward the north. Additional surface configurations can be added in this subroutine provided they follow the SPECS format

**5 Life cycle cost subroutine** This subroutine calculates, for the system under study, the yearly maintenance and operation costs and the life cycle cost and presents a detailed cash-flow analysis. The methodology used in the subroutines follows the National Bureau of Standards' economic criteria as outlined in Ref. 14

### C. Control Block

This is the main routine which executes, monitors, controls and drives the whole program as sketched in Fig. 6. It enters the input data and sends the desired output. The main routine calls the thermodynamic properties subroutines in order to determine the properties of flow for each station. It also calls the set of component description subroutines to compute equation residuals and partial derivatives, which compose the Jacobian matrix, and executes the Newton-Raphson software to solve the system of nonlinear equations.

As in a real system, the control block is thought of as receiving input signals at the end of each iteration step, represented in the values of the variables (such as temperature, pressure, mass flow rate, etc.). It compares these input signals to one another or to set points and transmits the resulting "logical" signal to a logic element where a transformation occurs to generate output signals to actuators (represented by a set of analytical expressions or tabular data) for the

operation of dampers, valves, pumps, boilers, etc. A "truth-table" is another way of representing the function of the controller-actuator in handling the input/output signals in the simulation.

### D. Output Block

The output of the program is controlled by several input variables. The computation results consist of the value of the different thermal and electrical loads at that particular hour, the thermo-dynamic properties at each station, the energy flux across each component of the system, and the overall performance, efficiency, and economics of the system. Several user options are introduced to obtain the desired level of printout detail.

### V. Simulation Steps

The philosophy of the program logic can be visualized as an attempt to perform in a stepwise manner what the real system does in a continuous mode of operation following a startup. To simulate a particular system configuration, the major steps of program operation are as follows:

**Step 1 (user's input data)** After all input data are entered into the program, the user can "echo" or print-back all input data for rechecking.

**Step 2** The dynamic simulation starts by the control block which groups the built-in characteristic equations of the components of the system being simulated. Because the Newton-Raphson method is iterative, it is essential to choose initial trial values of the variables to start iteration at each time step. Fluid valves can be initially set at their half-open position, and continuously adjusted as iteration proceeds. At the end of one time step, the results are used as the initial guess for iteration at the following time step, and so on. The above process is repeated until the simulation period has elapsed.

**Step 3** The check for convergence is based on the changes occurring between one iteration step and the preceding step. If the changes in all the variables are less than a specified tolerance,  $[\epsilon]$ , the computation is considered convergent and the program proceeds to Step 5, otherwise the program proceeds to Step 4 for another iteration.

**Step 4** Revise the initial setting of temperature and pressure controllers, valves, set points, etc. Diagnostic messages are printed if the variables attain unreasonable values or if convergence is not attained after a prescribed number of iterations. The program returns to Step 2 with updated values of the parameters.



Step 5 The program now prints out complete energy and flow rate profiles for the whole system under daily, monthly, and yearly operation Detailed life-cycle cost and cash flow analyses are made over a given number of years

Step 6 Additional optimization and parameter sensitivity will result from employing a search-type optimization technique at different design or operating conditions by repeating Steps 1 through 5

## VI. Program Summary

The SPECS program in its present form is structured to follow the above block descriptions Current activities relate to the following

- (1) Incorporating a detailed economics package for the system life cycle costs and cash flow analysis
- (2) Improving and expanding the tabulated results in the output block.
- (3) Expanding the components subroutines in the library block to include all the candidate components in Section IV-B-2
- (4) Debugging, testing, and documenting the final version of the program in order to extend its use to other design projects

In summary, the SPECS model was developed to suit the following purposes

- (1) To be used as an analytical tool for equipment design for large energy conversion systems where the cost of computer simulation represents a small portion of the total cost
- (2) After completing a trial equipment selection based on one set of loadings, the designer can apply the model to other partial or overloading conditions to determine if the original design is adequate
- (3) To serve the basis for developing "short cut" techniques after making a large number of detailed simulations on many different designs
- (4) To provide the field operator with diagnostic aid for understanding the dynamics of the operating system and the relationships between components at off-design conditions
- (5) To study the effects of add-on components and analyze the corrections sought for some operating problems if an existing system is expanded or reconfigured

The article has described these aspects of the SPECS computer program the method used for modeling a system of components, the mathematical representation of these components, the numerical technique used in solving the resulting system of equations, and the program input and output

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**Table 1. List of fluids used in SPECS**

Type No	Fluid	Chemical symbol
1	Water	H <sub>2</sub> O
2	Ammonia	NH <sub>3</sub>
3	Helium	He
4	Hydrogen	H <sub>2</sub>
5	Refrigerant 11 (trichlorofluoromethane)	CCl <sub>3</sub> F
6	Refrigerant 12 (dichlorodifluoromethane)	CCl <sub>2</sub> F <sub>2</sub>
7	Refrigerant 13 (chlorotrifluoromethane)	CClF <sub>3</sub>
8	Refrigerant 14 (carbontetrafluoride)	CF <sub>4</sub>
9	Refrigerant 22 (chlorodifluoromethane)	CHClF <sub>2</sub>
10	Refrigerant 23 (trifluoromethane)	CHF <sub>3</sub>
11	Refrigerant 114 (dichlorotetrafluoroethane)	CClF <sub>2</sub> CClF <sub>2</sub>
12	Refrigerant 318 (octafluorocyclobutane)	C <sub>4</sub> F <sub>8</sub>
13	Refrigerant 500 (azeotropes (73.8/26.2))	CCl <sub>2</sub> F <sub>2</sub> /CH <sub>3</sub> CHF <sub>2</sub>
14	Refrigerant 502 (azeotropes (48.8/51.2))	CHClF <sub>2</sub> /CClF <sub>2</sub> CF <sub>3</sub>
15	Refrigerant 503 (azeotropes (40.1/59.9))	CHF <sub>3</sub> /CClF <sub>3</sub>
16	Sodium	Na

## List of Symbols

$A$	cross-section area	$s$	entropy
$c_p$	specific heat	$S$	number of fluid stations
$C$	number of SSSF components	$T$	absolute temperature
$f$	function symbol	$U$	velocity
$g$	local acceleration of gravity	$v$	specific volume
$h$	enthalpy	$V$	vector of station properties
$J$	Jacobian matrix	$W$	work energy
$k$	thermal conductivity	$z$	elevation above sea level (or arbitrary datum)
$l$	streamline index	$\epsilon$	error limit for convergence
$m$	mass flow rate	$\mu$	dynamic viscosity
$n$	iteration step index	Subscripts	
$N$	number of unknown variables	$i$	inlet
$P$	absolute pressure	$o$	outlet
$Q$	heat rate	$m$	mechanical
$R$	boundary force due to friction or forced agitation	$e$	electrical

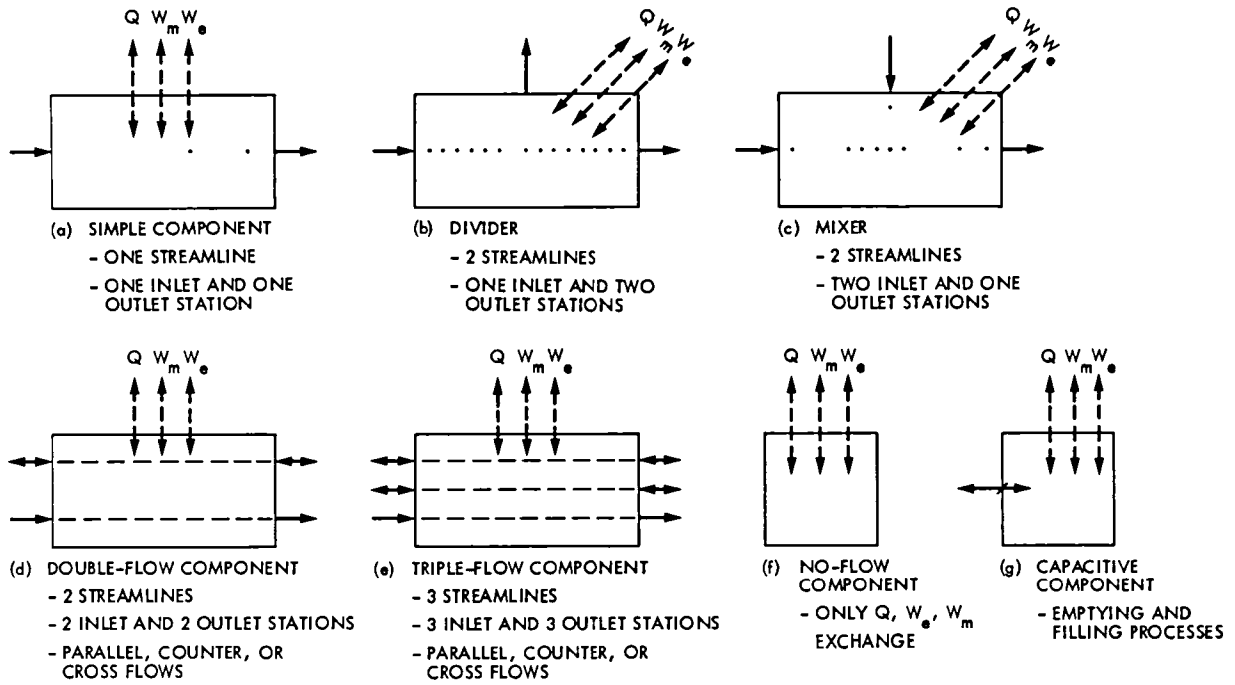


Fig 1 Simple and compound components

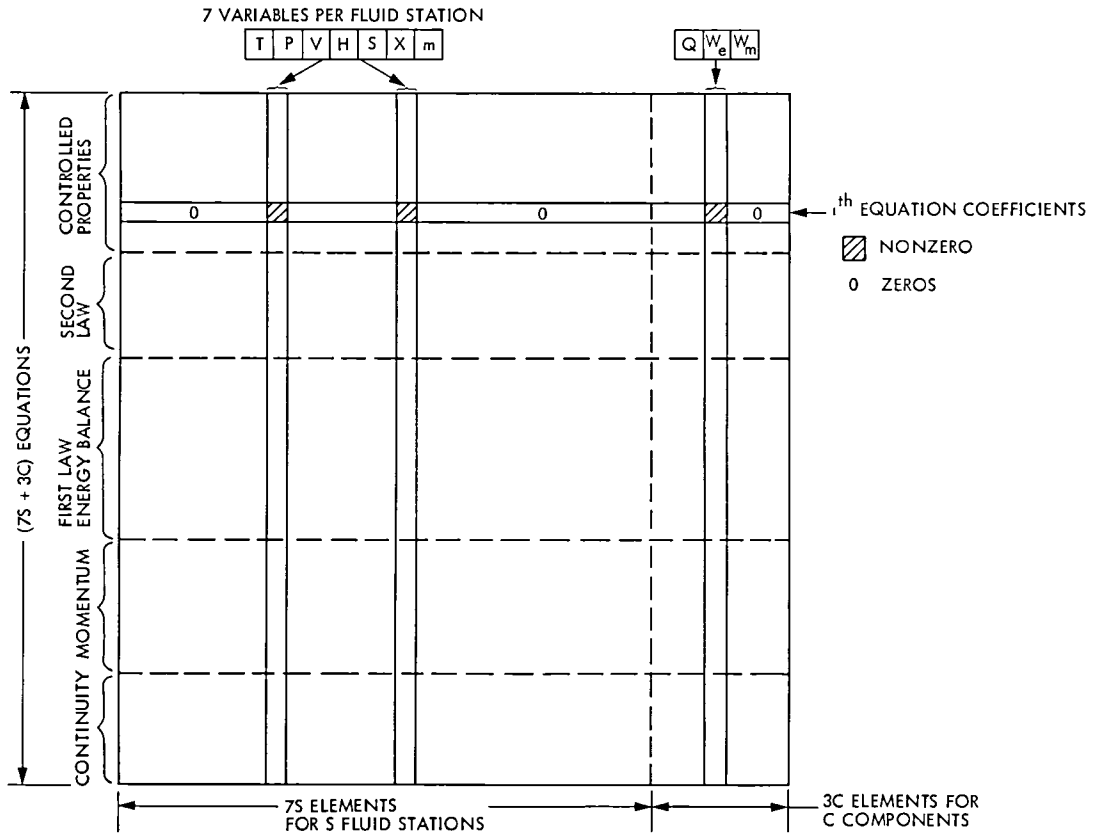


Fig 2 Details of the Jacobian matrix [J]

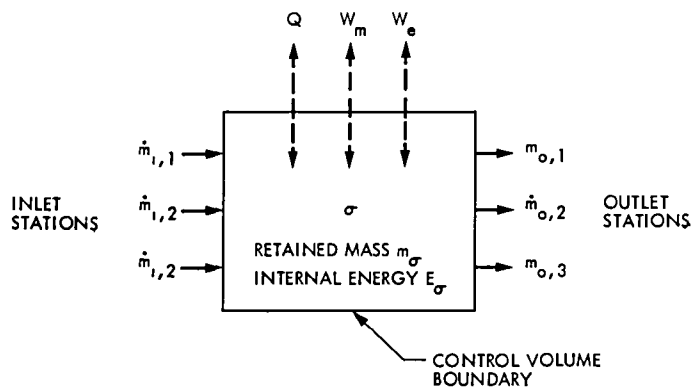


Fig 3 Compound component  $\sigma$  with multiple inlet and outlet ports

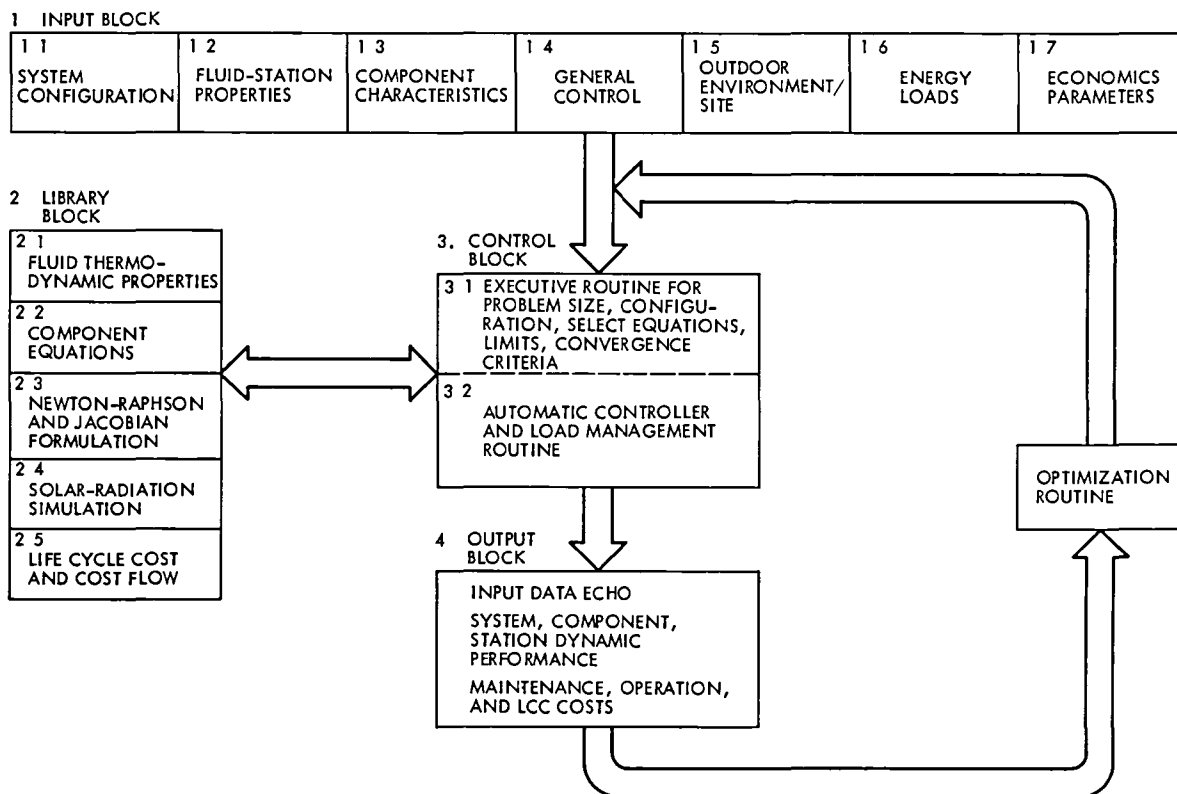


Fig 4 Overall structure of SPECS program

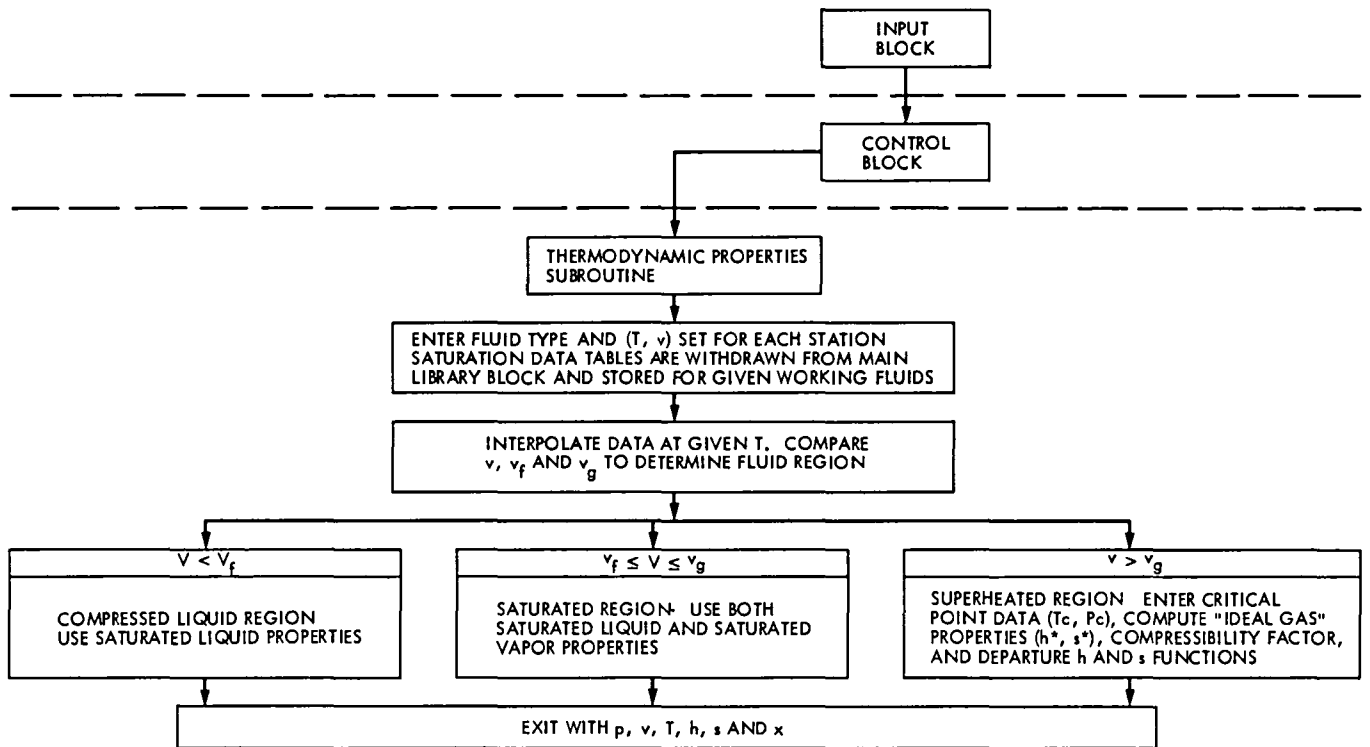


Fig 5 Flow chart for thermodynamic properties subroutine

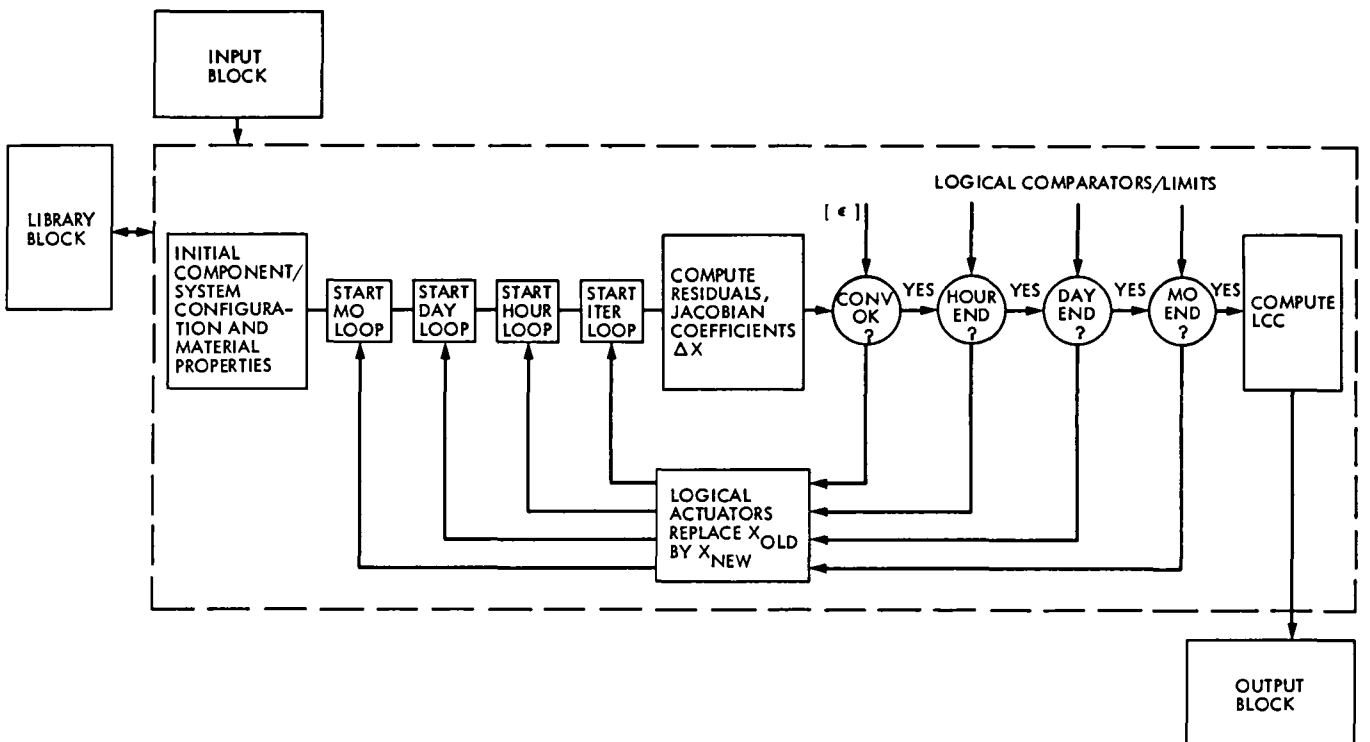


Fig 6 Information flow in control block

# The Measurement of Operator Workload in the Mark IVA DSCC Monitor and Control Subsystem

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*An operator workload measurement methodology is presented which will be used in support of the Mark IVA Operational Test and Evaluation (OT&E) Plan. Three operator workload measures are suggested: operator ratings, primary task work measures, and information processing time measures. A method of validating the workload measures using secondary task work measures is presented. We can expect that operations testing using these measures will assist in establishing the time required to perform essential operational activities and will indicate high risk operations areas due to potential operator overload.*

## I. Introduction

The goal of the Mark IVA DSN design is to provide a single cost-effective ground tracking and data acquisition network that will be capable of supporting both deep space and highly elliptical earth orbiter missions. To achieve this goal, existing tracking stations at Canberra, Goldstone, and Madrid will each be centralized so that the necessary antennas will be collocated near the signal processing equipment, which will be housed within a single building. Operating personnel will be reorganized and retrained to the extent necessary for them to be effective in their new roles.

Typically each of the present day Mark III DSN or GSTDN tracking stations is operated separately by its own staff. The Mark IVA DSN design will reduce the total amount of central-

ized equipment needed in conjunction with the antenna group to be developed at each 64-meter antenna site either by the relocation of an existing 34-meter antenna or by new construction.

A station supervisor in a signal processing center will be able to assign tasks to a control room crew consisting of a Complex Monitor and Control (CMC) operator working together with three Link Monitor and Control (LMC) operators. The centralized arrangement of the complex will permit a chosen level of tracking activities to be supported while the requirements for personnel and redundant equipment will have been reduced.

The CMC operator will be responsible for allocating assignable equipment resources to a link and for monitoring and



controlling the global equipment that provides services to the complex. An LMC operator will be responsible for the acquisition of tracking data during a scheduled project support activity.

The LMC operator workload will be offset in the initial implementation by a combination of equipment automation and some additional roving operators who will perform specific tasks on manually operated equipment in the equipment room in response to LMC operator directions.

This article reports on a project whose purpose was to develop a method of measuring control room operator workload. This methodology is an essential part of the OT&E plan for Mark IVA operations testing. In an effort to estimate operator working times on the Mark IVA DSCC Monitor and Control Subsystem (DMC) a series of operations sequence diagrams (OSDs) have been drawn up. These are detailed task descriptions with estimates of the amount of time needed to complete each subtask by novice, experienced, and expert operators. These estimates are added to obtain a prediction of the amount of time needed for completion of a whole task such as a VLBI clock sync prepass. These times will be used as part of an effort to assess the actual workload on individual operators during task performance. This is necessary since it is likely that the occurrence of some events, particularly alarms, will result in a sudden, significant increase in workload which may, in turn, increase operator error rate, with an attendant risk of data loss. Minimizing the risk in the complex user-computer interface of the DMC is the purpose of the present effort.

Although much attention in recent literature has been given to the user-computer interface (Refs 1, 2, and 3), most of it has not been aimed at gauging the effect of an increase in workload in an on-line, interactive system, with the possible exception of studies of pilot and aircrew performance, and this is a somewhat unique case of a system which is only partly, though increasingly, an information processing system. Moreover, most of the studies cited in an extensive search (Ref 3) have been done in existing systems. The present problem is to assess workload on a system which is still in the design stage and to predict its effect on system performance. The measures developed to do this can be implemented in an operational test, which will show where human operator overload difficulties are likely to arise.

There is extensive literature on the measurement of workload (Refs 3 and 4), and many possible measures have been proposed. These have been classified into four general categories (Ref 3) operator ratings, spare mental capacity, primary task performance, and physiological measures.

A method of evaluating workload assessment measures based on this classification, also presented in Ref 3, was applied to the operator's tasks as outlined in the OSDs, and a combination of operator ratings and primary task performance measures was chosen, with a spare mental capacity technique being used to appraise the validity of the measures. Physiological measures were not considered because they are expensive to implement, may be intrusive on the operator, and are not practical for ordinary on-line application. Some characteristics of the chosen measures are considered as follows.

### **A. Operator Ratings**

Previous studies have shown this to be a reliable measure which is consistently related to performance (Refs 5 and 6). It is relatively easy to obtain, and has the "face validity" of getting the operator's own estimate of the subjective phenomenon of mental workload. With careful attention to design (Refs 7 and 8), rating scales yield quantitative results which may be used to gauge "crossover" points beyond which the chance for error may increase greatly or other system parameters may be affected.

### **B. Primary Task Performance**

This is perhaps the most direct way to assess workload, since it involves the measurement of the effect of workload on the performance of the task of interest. The time needed to start and/or complete the task is the most relevant (Refs 8 and 9), although a number of errors may also be considered. These measures must be carefully designed to reflect increasing workload, however, since human operators tend to adapt to the task and hold their performance constant over a broad range of conditions. Thus, operators' initial starting times for a particular subtask are generally taken as a measure of their strategy for dealing with increased load. The starting time reflects the speed at which incoming subtasks are processed mentally (Ref 10), and is a linear function of the amount of information processed by the operator.

Additionally, successful completion of a particular subtask or set of tasks in a stated amount of time may itself be considered a measure of workload (Ref 11), and putting increasingly stringent time requirements on the operator until task completion becomes impossible may reveal operator strategies as well as workload limits for particular subtasks. Therefore, both a measure of time to start (reaction time) and time to complete a subtask will be used to assess operator workload.

### **C. Secondary Task Measures**

Since human operators tend to hold their task performance constant under increasing workload, primary task performance measures may change very little in the normal course of opera-

tion If, however, a secondary task is introduced, so that the operator is required to divide attention between two tasks, then the secondary task may be considered a measure of "spare mental capacity" (Ref 4) and, as the demands of the primary task increase, performance on the secondary task decreases, so it is an indirect measure of operator workload This is the standard way of using a secondary task in workload measurement

For on-line, operational measurement, the imposition of a secondary task is impractical, since it could interfere with normal operations However, a variation of the secondary task technique can be used to validate the primary task measures in a trial or pilot study Such a study is necessary in order to demonstrate that the workload measures used do, in fact, increase with increasing objective workload and that this in turn results in performance degradation and possible data loss

For this purpose, the secondary task should be similar to tasks that would normally be performed by the operator, and it should be possible to make the task more or less time consuming Log keeping in writing and by voice is such a task It, in turn, will affect primary task performance as its requirements result in increased workload (Ref 9) Then the primary task workload measures should rise, and this will result in a decrement in system performance

Three specific ways of implementing rating and primary task workload measures are described below, along with a secondary task technique for validating the measures

## II. The Measures

Three measures are proposed for use in gauging operator workload in the DMC They are interrelated, and should be collected simultaneously on a single task They are

- (1) Sequential operator ratings of workload on subtasks
- (2) The ratio of the time required to the time available to do a particular subtask
- (3) Information processing time, or the time taken to initiate the physical (keystroke) portion of a subtask

The latter two measures are keyed to the operator rating measure, which is taken at the end of each subtask and is recorded with a short free-form input (FFI)

### A. Operator Ratings

**1 Procedure** A single rating of workload, on a scale from 1 to 7 with a 7 meaning the highest workload, will be obtained from each operator during the performance of a task The

instructions to the operator (reproduced in Fig 1) are given only once Thereafter, at the completion of each subtask, the words "difficulty rating" and "time pressure rating," will appear in the SOE, and the operator will respond with an FFI text message and a rating of from 1 to 7

A subtask is defined as a listed task with an estimated performance time in an operations sequence diagram (OSD) This will result in the operator's giving a rating at approximately 2-4 minute intervals, depending on operator experience

The operator rating will be recorded on the operations log, along with the time at which it was made This will provide a record of the time at which each subtask is finished (to be used in the other two measures described below), as well as of the operator's rating

**2 Analysis** The procedure will result in operator ratings on each of a number of subtasks, so that those points in the task at which overload occurs can be quickly located

The operator ratings are obtained from the operations log and classified according to subtask and operator experience Therefore, there should be some means of identifying the particular subtask (e g , "Display and Compare APA Predicts against VLBI Source Table") on the log The classification of operator experience (novice, intermediate, or expert) should be done by an experienced supervisor, with a set of guidelines taking into account the amount of time on the job and previous experience, if any

Then the mean rating can be obtained for each subtask and each operator classification, and plotted as a function of task, resulting in a graph similar to that illustrated in Fig 2 The peaks on such a graph will show the subtasks on which the operator is working at a higher subjective workload In conjunction with the measures described below, this will indicate tasks on which errors might be expected to occur in the event of a sudden increase in workload such as might result from an event notice or alarm

In this type of testing, the differences in ratings obtained for each subtask may be relatively small, the differences among the three types of operators may also be small but of interest In order to test the significance of the differences, the rating data should be subjected to an analysis of variance (ANOVA), which shows whether or not there are real differences among tasks and/or operators, and whether there is an interaction between them, i e , whether some tasks are perceived as having a higher workload by novice operators, and others by experienced operators, etc

In addition, after the completion of a task, such as a Delta DOR Prepass operation or 64-meter antenna configuration,

operators will rate several aspects of the job as a whole on a form like the one illustrated in Fig 3 This is a very simple form, and can be analyzed easily, whether it is done on the log or by hand Mean ratings on each of the first 10 questions should be obtained for each operator class These should be examined individually as indicative of problem sources perceived by operators The open-ended question 11 should also be used in the same way The questionnaire will complement the subtask ratings by providing a general operator rating of workload in question 1 This rating should be correlated with the mean of the subtask ratings for each subject (i.e.,  $\bar{x}_i$  = mean of subtask ratings for  $S_i$ ,  $Y_i$  = whole task workload rating from question 1 for  $S_i$ ) It will also serve as an indication of problems perceived by operators in particular *systems*, as opposed to the problems of particular *subtasks* pointed out in the repeated workload ratings obtained after each subtask

## B. Primary Task Workload Measure

**1. Procedure** This measure will be similar to standard system reliability workload measures, in that it will consist of a ratio of time required to time available Thus, workload is defined as

$$W = T_r/T_a$$

where

$W$  = workload

$T_r$  = time required by the operator to perform a specific subtask

$T_a$  = time estimated in the OSDs for that particular subtask

Since we are estimating mental workload, there may be no obvious action recorded on the operations log to signal the start and end of a subtask Therefore, the operator will be asked to signal the start of each task with a FFI which will be used to time that task on the log

**2. Analysis** The time between the FFI messages which indicate the start of a subtask and the operator's rating of that task can be taken as the actual or required time to perform the subtask This time,  $T_r$ , will be available in printout form along with the operator ratings It can then be compared with the times estimated in the OSDs for that subtask

If subtasks are always performed in the proper sequence, particular tasks can be identified on this basis However, an identifying code for each subtask, either a mnemonic or a numeral, would insure that the proper task was being considered with its time and rating, and would aid data analysis

The resulting workload measure for each subtask will be analyzed in a fashion similar to that used for the ratings Mean workload will be plotted as a function of task, with separate curves for each operator classification, and an analysis of variance performed on the data to determine if different subtasks yield significantly different workloads In addition, the correlation between the workload measure and the operator rating for each subtask should be obtained

If the obtained mean workload for a particular task is greater than 1, then the time needed for the task is greater than that originally allowed in the estimate If it is above 0.7 or 0.8, it is approaching the operator's maximum capacity In either case, if the estimates are to be considered as the time allowed for the task, then some provision must be made to relieve operator workload on these subtasks, perhaps by task rearrangement, extra help from personnel or computer, or simply increasing the time allowed Otherwise, there will be an increased risk of operator error occasioned by sudden increases in workload, such as the occurrence of an event or alarm requiring operator action during an already overloaded subtask

## C. Information Processing Time

**1 Procedure** The third value to be used does not measure workload directly, but is an attempt to measure the time required for mental processing of the information needed to perform a control function

Many of the subtasks in the OSDs require only monitoring functions, in this case, since there is no overt action, but only information processing (e.g., comparing source time and ID against VLBI source table for compatibility), this information processing or "think" time will be the same as  $T_r$  (time required) in the second measure However, on most subtasks, an overt response is required The response may be part of the normal SOE or be contingent on the detection of some anomaly such as a high system temperature value However, the majority of tasks do require some sort of physical action, and the time to start such action is called a choice reaction time (CRT), since the operator must choose the action to be taken from among a number of possible actions It varies directly with the number of choices or actions, and is a measure of uncertainty, or the amount of information the operator must process The longer the CRT, the more complex the information processing task and the greater the workload Thus, information processing time, as measured by CRT, is also a measure of workload

One task follows another immediately in the OSDs However, this may not always be the case in practice, since there may be interruptions or operator recovery time from the previous task It is reasonable to include this extra time in the

direct workload measure (the second measure), since it becomes part of the workload and adds to it. However, it is obviously not part of the information processing time for a particular subtask. Therefore, it will be necessary, as for the second measure, for the operator to signal the start of each subtask with an FFI text message which can be used, in conjunction with the next action taken (keystroke or voice communication), in determining the information processing time.

**2 Analysis** The time between the FFI text message which signals the start of a task and the next action is obtained from the operations log for each subtask for each operator. The mean of these values is obtained for each class of operators. This data is then subjected to the same type of analysis as before, i.e., mean information processing time is plotted as a function of task for each class of operators, so that those tasks which carry a heavy information processing load can be identified. The data is also subjected to a 3 (operator class)  $\times$   $n$  (number of tasks)  $\times$   $S$  (subjects) ANOVA, to determine the significance of the observed differences.

In addition, when all three measures have been obtained, a grand mean across all three classes of operators should be calculated for each measure on each task. If all three tasks are measuring workload, then there should be some degree of correlation among them. In fact, since the information processing time is a fraction of the required time in the workload measure, these two will probably be highly correlated. This makes the calculation of a multiple correlation coefficient inappropriate. However, it is appropriate to calculate the correlation coefficients between the operator rating and each of the other measures. These should be relatively high (above 0.5), which would indicate that all three measures are measuring different aspects of the same thing (workload) and would be evidence for the validity of the measures.

### III. Validation of the Measures

The three proposed measures seem logical, and they are based on measures that have been used for similar purposes in the literature (Refs 3 and 4). Nevertheless, their validity and reliability need to be evaluated for the use in the operational testing. It is possible, for example, that one or two of the measures will show no variation, even when operator workload is obviously markedly increased, whereas a minor change in the data collection method would yield usable results. This is a common occurrence in experimental trials with human operators. The only way to determine whether or not a measure of human performance will reflect variations in external working conditions is to run a "pilot study", i.e., try the measure out using a small sample of people performing a relevant task.

In addition to demonstrating that the measure varies with task parameters, pilot studies can also be used to assess the effect of increasing workload on performance. The purpose of measuring operator workload, after all, is first to detect tasks on which the workload is so high that the task becomes impossible (this would be relatively rare) and second to detect those tasks or parts of tasks in which the workload is high enough so that, if a further burden is added in the form of, say, an unforeseen event, operator errors are likely to occur. If the measures are valid, i.e., if workload is really being measured, then they should increase with increasing task demand, and this should result in increased operator errors. The pilot studies, or trials, are designed to assess this.

The trials must be run before full-scale operational testing, both to allow for changes in the measures and because they will involve "loading" the operator with extra, or secondary, tasks until the error rate starts to increase, a technique that would be undesirable during full-scale operations testing.

**1. Procedure** From three to six people, preferably those with some experience on the Mark III or a similar system, should serve as subjects for the trials. They need not be rated as novice, intermediate, or expert, since the purpose of the trials is simply to see whether the measures respond to increasing objective workload.

A task of about 30-45 minutes duration will be designed for the trials so as to be as similar as possible to the Mark IVA LMC operator's task, i.e., an interactive computer task. Although the LMC operator's tasks may be intermittent, with periods of high workload alternating with relatively light workload, the trial task should occupy the operator continuously.

The operators will perform the task under each one of three conditions. In the first condition, the primary task alone will be performed by the operator (low workload), in the second condition, a moderate amount of a secondary task, to be done at the same time as the primary task, will be assigned (medium workload), in the third condition, a still heavier amount of the secondary task will be assigned the operator (high workload). The purpose of the secondary task is twofold. First, it should demonstrate that as the secondary task increases the workload on the operator, the three workload measures rise, and second, performance on the secondary task should be negatively correlated with the workload measures, since, as primary workload goes up, the operator has less "spare mental capacity" to work on the secondary task. Both these effects serve as checks on the validity of the workload measures.

There will, of course, be no secondary task in the first condition. In the second condition, the secondary task will consist in the operators' keeping a written log of all of the opera-

tions and parameters. This should be done on a prepared form, similar to forms currently in use. In the third condition, in addition to keeping the written log, the operator will keep a voice log by reporting the data over the voice circuit to a receiving operator, similar to the NOCC operator. These two tasks have three characteristics: (1) they are designed to overload the operator in two increasing steps so as to produce an increase in the workload measures and possibly an increase in procedural lapses (errors of a type that would not result in data loss) on the part of the operator, (2) they are designed to be similar to secondary tasks normally performed by operators, so they will not be perceived as simply "busy work" that can be ignored, and (3) the amount of work involved in them can be manipulated in terms of the amount of detail required. Thus, if on first presentation the overloading tasks are found to be easily handled by the operator, and make no difference in the workload measures or in lapses, the amount of record keeping required should be increased until an effect is shown.

The operators should practice on the simulated system before the trials until their performance is asymptotic in terms of the amount of time they require to do the job. Then the trials with the three workload conditions can begin. Each person who serves as a subject should experience all three conditions, so that performance differences among conditions can be attributed to the effects of the workload variation and not to individual differences. Each individual should experience the workload conditions in a different order chosen randomly from among the six possible orders of three conditions.

**2 Analysis** The three workload measures described above will be taken for each individual for each subtask under each condition. Means will be calculated for each condition, and an analysis of variance run separately for each measure. This should demonstrate that the means of the workload measures increase significantly with the increase in imposed workload over the three conditions, thus providing evidence for the validity of the measures.

It would also be of interest to gauge the effect of workload on operator performance. Generally, operators compensate for increased workload by increased efficiency, so that errors which actually result in lost data are rare, and probably would not occur at all in a 30-45 minute trial. However, anomalies such as an incorrect input, failure to detect a significant reading on a display, or failure to understand a verbal communication are fairly common. They are usually corrected in a short time (so human operators are sometimes called "self-correcting") but they may significantly add to the workload itself and, more importantly, if the workload is already high, they may lead to further errors which may actually result in missed data. Therefore, an increase in operator anomalies with

increasing workload must be taken seriously, even though the anomalies are corrected.

In the trials described here, then, all operator anomalies should be recorded in the course of data collection. They can then be subjected to the same type of analysis as the workload measures, i.e., an analysis of variance which should demonstrate that anomalies increase as imposed workload increases. Further, a detailed examination of the subtask in which anomalies occur, and their workload measures, should yield useful data on the possible effect of workload on the probability of operator anomalies, on which some predictions may be made about the probability of data loss given a particular level of workload.

Some operator anomalies which should be considered are

- (1) Failure to notice a signal change or update
- (2) Hitting the wrong input key
- (3) Failure to find the source of an error message
- (4) Voice communication failure (incorrect reception or transmission), including requests to "say again"
- (5) Incorrect information recorded on the log
- (6) Incorrect information transmitted to the voice log

Furthermore, since such anomalies are inherently unpredictable, it is not possible to define them completely in advance, so that any anomaly that occurs during data collection should be recorded. This will involve some on-the-spot judgment on the part of the test conductor. The slightest anomaly should be recorded so as to increase the data pool.

If this procedure is followed, a useful and meaningful (in terms of probability of data loss due to operator error) measure or measures of human operator workload can be obtained. The measures can then be used in an operations test as a part of systems evaluation and periodically thereafter whenever system testing is needed.

## IV. Implementation

The operator workload measurement methodology would use specially developed SOEs for operations testing. The SOEs would indicate when an operator rating is required. The rating will be input to the system by means of a text message which is timed, tagged and recorded on the operations log. The analysis will use the data recorded on the operations log.

Validating the operator workload measurement with the secondary task method would be accomplished before the formal Mark IVA operational workload testing is begun

## V. Summary

Operator overload is identified as a potential problem in Mark IVA system operations. Three methods for measuring operator overload have been presented. The methods are operator ratings, primary task work measures, and information processing time measures. A method of validating these three

workload measures by secondary task performance measures is suggested.

These operator workload measures would be an integral part of the OT&E plan for operations testing of the DMC. Operations testing in this form would provide indications of where in the operations of the Mark IVA system we can expect system failures due to operator overload.

The operator workload measurement methodology described here allows the problem of operator overload to be addressed directly. It provides a general capability to measure operator workload in future systems.

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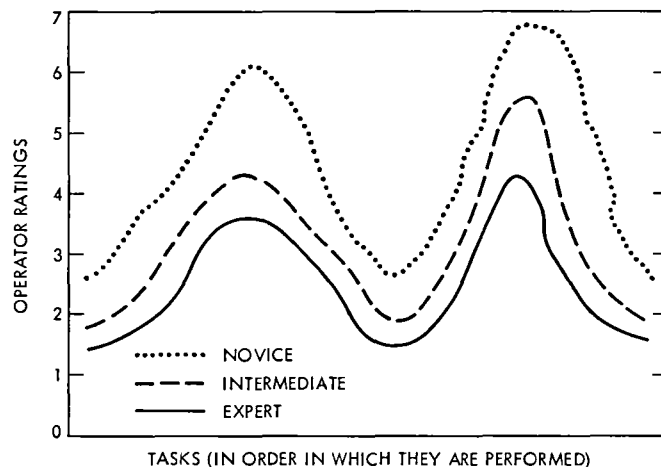
We are interested in obtaining information about specific points in the sequence of operations where the workload or amount of effort required by you, the operator, approaches the level where the time allowed is not sufficient to get the job done. In order to do this, we will ask you for a series of ratings of the difficulty of the task and the amount of workload you experience on each of many of the small subtasks of a particular job. At frequent intervals while you are doing a job, a request for two ratings will appear in the SOE. At these times, please respond as fast as you can with the appropriate FFI text message which includes a value of from 1 through 7 to indicate your ratings of the difficulty of the task and the workload you have experienced since the last rating (or since starting the job). Use the following scale.

The task was	There was
(1) Extremely easy to do	(1) Plenty of time to spare
(2) Very easy to do	(2) Some time to spare
(3) Easy to do	(3) A little time to spare
(4) Neither hard nor easy to do	(4) Time allowed is just enough to get the job done
(5) Hard to do	(5) Hard to finish in the time allowed
(6) Very hard to do	(6) Very hard to finish in the time allowed
(7) Extremely hard to do	(7) Impossible to finish in the time allowed

1 2 3 4 5 6 7	1 2 3 4 5 6 7
increasing difficulty —	increasing time pressure —
Difficulty rating	Time pressure rating

**Fig. 1. Instructions and scale for operator workload ratings**



**Fig. 2. Mean workload ratings for each subtask**

The following questionnaire should be given once after the completion of a trial and/or after a reasonable familiarization period during operational testing

Please indicate your rating of the following statements by writing a number from 1 to 7 in the space provided

1 – Very strongly disagree

2 – Strongly disagree

3 – Disagree

4 – Neither agree nor disagree

5 – Agree

6 – Strongly agree

7 – Very strongly agree

- 1 The workload required of the operators of this system is too high \_\_\_\_\_
- 2 The system provides all the displays that are needed for the job \_\_\_\_\_
- 3 In general, each display is adequate for its job \_\_\_\_\_
- 4 The working group was able to respond to the demands placed on it \_\_\_\_\_
- 5 Information can move freely and accurately among the LMC, roving, and NOCC operators \_\_\_\_\_
- 6 It is easy to find the information needed in the system documentation \_\_\_\_\_
- 7 System event reporting makes clear what action is to be taken \_\_\_\_\_
- 8 Error messages describe each fault in sufficient explanatory detail \_\_\_\_\_
- 9 Error messages make clear what action is to be taken \_\_\_\_\_
- 10 There are too many meaningless messages on the screen \_\_\_\_\_
- 11 There were other problems in operating the system (specify) \_\_\_\_\_

**Fig 3 Operator workload questionnaire**



# The 1986 Launch of the Galileo Spacecraft via the Space Transportation System

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*Beginning with the Galileo spacecraft launch in 1986, deep space payloads will be launched via the Space Shuttle. This change from the previous use of expendable launch vehicles will introduce large changes in procedures and data flow configurations for both the flight project and the Deep Space Network during the launch period. This article describes the planned Galileo launch period sequence of events and telemetry and command data flow configurations.*

## I. Introduction

Starting with the launch of the Galileo spacecraft, now scheduled for late May 1986, deep space payloads will be launched via the Space Shuttle vehicle of the Space Transportation System (STS), in sharp contrast to all previous launches of deep space payloads via expendable launch vehicles. This very significant change in the method of launch will result in large procedural changes for both flight projects and the Deep Space Network (DSN), in its capacity as lead support network for all deep space missions. In a previous article (Ref 1), the Space Shuttle impact on the DSN initial acquisition was described, in this article, procedural differences for the flight project and DSN during the pre- and postlaunch periods are examined. In particular, major subphases of the pre- and post-launch period are identified, and telemetry and command data flow configurations are presented for each subphase. The Galileo mission sequence of events and data flow configura-

tions here described can be considered typical of deep space launches planned for the Shuttle era.

The information here described was first presented in Ref 2, under the assumption that the upper stage was to be the Inertial Upper Stage (IUS), this article now updates the information for the currently planned Centaur upper stage.

Major differences for the flight project and DSN in the upcoming Space Shuttle era are the larger number of subphases in the launch period (e.g., Shuttle on-orbit phase of several hours, for which there was no similar phase in the expendables era) and direct launch involvement of additional NASA centers (e.g., Johnson Space Center and Lewis Research Center).

Section II describes the nominal Galileo sequence of events during the launch period. Section III identifies major launch

period subphases, while Section IV identifies the telemetry and command data flow configurations for each of the launch period subphases

## II. The Nominal Galileo Sequence of Events

As previously mentioned, the Galileo spacecraft is currently scheduled for a late May 1986 launch. The overall launch period begins with the transporting of the spacecraft to the Kennedy Space Center (KSC), extends through liftoff, and terminates with a successful DSN initial acquisition. Typical prelaunch (prior to liftoff) and postlaunch (following liftoff) Galileo sequences of events are described in detail below.

### A. Nominal Galileo Prelaunch Sequence of Events

The Galileo spacecraft is transported from the Jet Propulsion Laboratory (JPL) to the Kennedy Space Center (KSC) by a United States Air Force C-5A aircraft. During all spacecraft testing at KSC, command and telemetry support will be provided via the DSN Facility (MIL 71) with operations control from the Mission Support Area (MSA) at JPL. After being unpacked in the Spacecraft Assembly and Encapsulation Facility (SAEF), the spacecraft is inspected to determine if any damage was sustained during transit. Baseline tests, including the use of S- and X-band radio frequency (RF) subsystems, are evaluated. At the conclusion of these tests, Radioisotope Thermoelectric Generators (RTGs) are installed on the spacecraft and tests are rerun using the RTGs as the source of electrical power. When these tests are completed, all power is removed from the spacecraft to ensure that the installation of pyrotechnic devices and the loading of consumables in the propulsion system can be carried out in a safe environment.

At the conclusion of the propulsion loading and pyrotechnics installation, the spacecraft will again be activated for further testing and will be prepared for transfer to the Vertical Processing Facility (VPF). At the VPF the spacecraft will be mated to the Centaur Tests via the Merritt Island Launch Area (MILA) will be conducted to verify Space Shuttle end-to-end communications using the Cargo-Integration Test Equipment (CITE) to provide simulated Shuttle data interfaces. A series of operational tests will also be conducted. Upon satisfactory completion of these tests, the spacecraft will be placed into the storage mode to await shipment to the launch pad.

About 30 days before launch, the spacecraft will be removed from storage and transported to the launch pad, preceding the Shuttle arrival by two days. After cargo preparation procedures are completed, the spacecraft will be installed in the Shuttle Orbiter Bay. About four days before launch, the RTGs

are re-installed in the spacecraft. Final end-to-end communications tests using MILA and Space Transportation System (STS) Tracking Data Relay Satellite (TDRS) communications links will then be conducted. These tests should last about two days. At their conclusion, the launch countdown will commence.

Figure 1 presents the nominal Galileo prelaunch sequence of events for a May 1986 launch.

### B. Nominal Galileo Postlaunch Sequence of Events

Galileo launch mode telemetry will be available in near-real-time via the STS-TDRS communications link, from liftoff to Shuttle-Centaur separation. Based on the telemetry data, a decision to continue with the planned flight to Jupiter must be made by Launch plus 3 hours.

If a "go" decision is made, the Shuttle-Centaur separation should occur during the third Shuttle orbit, at about four hours after launch. During the more favorable periods of the launch window, the separation could be delayed until the fourth or the fifth orbit without jeopardizing the objectives of the Galileo mission. After separation, a Centaur S-band link can be used to route data between the spacecraft and Shuttle or TDRS. The maximum useful range of the Centaur to Shuttle link is 10 kilometers.

About 45 minutes after Shuttle-Centaur separation, the Centaur main engine will burn for approximately 10 minutes. After main engine cutoff (MECO), the Centaur will initiate a slow thermal roll of 0.1 rpm, and the Galileo spacecraft will start deployment of the RTG, science and magnetometer booms. The spacecraft transmitters will then be turned on to provide a downlink through TDRS just prior to Centaur-spacecraft separation. Centaur will turn the spacecraft to point 8 degrees to the earth side of the sun and will spin-up the spacecraft to 2.9 rpm. The spacecraft will then separate from Centaur, which will maneuver to avoid the same trajectory path as the Galileo spacecraft. The spacecraft transponder will now be the only means of exchanging data between the earth and the spacecraft.

Immediately after separation, the high-gain antenna will be deployed and the retro propulsion module (RPM) pressurized. During these events, DSN acquisition of the downlink will be established (approximately 15 minutes after Centaur-spacecraft separation). After RPM pressurization, the Galileo spacecraft will perform a sun acquisition (approximately one hour after Centaur-spacecraft separation).

Figure 2 presents the nominal Galileo postlaunch sequence of events.

### III. Major Launch Subphases During a Shuttle Deep Space Payload Launch

The overall launch period of a deep space payload launch is conveniently subdivided into three major categories, as follows

- (1) Prelaunch phase
- (2) Shuttle attached phase
- (3) Shuttle detached phase

#### A. Prelaunch Phase

This phase starts with spacecraft assembly in the Spacecraft Assembly and Encapsulation Facility (SAEF), and includes that time the spacecraft spends in test at the Vertical Processing Facility, and finally the move to the launch pad. This phase terminates with liftoff from the pad. Subphases during the prelaunch phase are identified as

- (1) Spacecraft assembly and encapsulation facility
- (2) Vertical processing facility
- (3) Launch pad

#### B. Shuttle Attached Phase

The phase starts at the moment of liftoff, and proceeds through the Shuttle ascent and the Shuttle on-orbit operations. This phase is terminated when the Centaur spacecraft is placed outside the Shuttle Orbiter. Subphases during the shuttle attached phase are identified as

- (1) Shuttle ascent
- (2) Shuttle on-orbit

#### C. Shuttle Detached Phase

This phase starts when the Centaur spacecraft is placed outside the Shuttle vehicle, and proceeds through the Centaur burn and spacecraft-Centaur separation. This phase terminates with a successful DSN initial acquisition. Subphases during the Shuttle detached phase are identified as

- (1) Centaur burn
- (2) Post Centaur separation

### IV. Galileo Telemetry and Command Data Flow Configurations During the Launch Period

There are four major telemetry and two major command data flow configurations during the launch period, and these are detailed as follows

#### A. Galileo Spacecraft – DSN

This path exists for both telemetry and command data. The link between the spacecraft and the Merritt Island Launch Area (MILA) DSN facility (MIL 71) is both radio frequency (RF) and hardline. Communications from MIL 71 to the JPL Mission Control and Computing Center (MCCC) is via the JPL Ground Communications Facility (GCF).

#### B. Galileo Spacecraft – TDRSS

This path is for telemetry data only. The link between the spacecraft and the MILA Ground Spacecraft Tracking and Data Network (GSTDN) station is RF. From the GSTDN station an RF uplink is established to the Tracking and Data Relay Satellite (TDRS). Alternatively, an RF link can be established directly from the spacecraft to TDRS. From TDRS, an RF downlink is established to the White Sands Ground Station (WSGS). From there, the data is transmitted via domestic satellite (DOMSAT) to the Goddard Space Flight Center (GSFC) NASA Communications (NASCOM) switching center, and thence through DOMSAT to JPL MCCC.

#### C. Galileo Spacecraft – Centaur

This path is for telemetry data only. Spacecraft telemetry is embedded in Centaur telemetry. An RF link is established from the Centaur to the TDRS. From TDRS, an RF downlink is established to WSGS. From there, the data is transmitted via DOMSAT to the GSFC NASCOM switching center, and thence through DOMSAT to the Centaur Payload Operations Control Center (POCC), at the Lewis Research Center (LRC). From LRC POCC, Galileo telemetry is stripped out and transmitted via GSFC NASCOM to JPL MCCC.

#### D. Galileo Spacecraft – Shuttle

This path is for both telemetry and command. For telemetry the link can be direct from the Galileo spacecraft to the Shuttle Orbiter, or embedded in Centaur telemetry data to the Shuttle Orbiter. From the Shuttle the link is RF to TDRS to WSGS. From WSGS, the data is transmitted via GSFC NASCOM to Johnson Space Center (JSC) Mission Control Center (MCC). From JSC MCCC, Galileo telemetry is stripped out and transmitted via GSFC NASCOM to JPL MCCC. Alternatively, Centaur Galileo telemetry is transmitted via GSFC NASCOM to LRC POCC, where Galileo telemetry is stripped out and provided through GSFC NASCOM to JPL MCCC.

For command, the link begins with the LRC Centaur POCC, in conjunction with voice communication from JPL

MCCC From the LRC POCC, the link is first to JSC MCC through GSFC NASCOM, then to WSGS through GSFC NASCOM, and finally RF to TDRS to the Shuttle Orbiter. From the Orbiter, the link is either hardline or RF (Centaur Shuttle Orbiter distance <20 km) to Centaur, and finally, to

the Galileo spacecraft. For this mode, only a limited series of commands ("discrete commands") is possible.

Figures 3 through 11 illustrate the above data flow paths for the various launch subphases.

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## Note Added During Publication

In December 1982, significant modifications were made to the Galileo launch support plan, primarily, these modifications involved locating the Centaur Payload Operations Control Center (POCC) at Kennedy Space Center (KSC) and performing the DSN initial acquisition prior to spacecraft-Centaur separation. The impact of these changes on the launch phase configurations and sequences of events will be described in a future article.

SPACECRAFT EVENT	FACILITY	MIL 71	ACTIVITY	1986			
				FEBRUARY	MARCH	APRIL	MAY
• INSPECTION..... BASELINE TEST	S/C ASSEMBLY AND ENCAPSULATION	✓	S- AND X-BAND T/R	□			
• INSTALL TEST RTG.....	S/C ASSEMBLY AND ENCAPSULATION	✓	S- AND X-BAND T/R COAX TLM	□			
• PROPULSION..... PREPARATIONS	S/C ASSEMBLY AND ENCAPSULATION	✓	NO POWER FOR SPACECRAFT	▬			
• BASELINE TEST.....	S/C ASSEMBLY AND ENCAPSULATION AND VERTICAL PROCESSING	✓	S-BAND RECEIVE COAX (TLM, TDRS, CENTAUR)		□		
• CENTAUR MATE AND TEST	VERTICAL PROCESSING	✓	S-BAND RECEIVE COAX (TLM)			□	
• STORAGE AND BASELINE TEST	VERTICAL PROCESSING	✓	S-BAND RECEIVE COAX (TLM)				□
• INSTALL IN SHUTTLE CARGO BAY	LAUNCH PAD	✓	NO POWER FOR SPACECRAFT				□
• RTG FINAL INSTALL AND TEST	LAUNCH PAD	✓	S-BAND RECEIVE COAX (TLM)				□
• END-TO-END TEST.....	LAUNCH PAD	✓	S-BAND RECEIVE COAX (TLM) CENTAUR, TDRS VIA SHUTTLE				□
• LAUNCH.....	LAUNCH PAD	✓	TRACKING AND TELEMETRY VIA CENTAUR AND TDRSS NETWORK				▽

Fig 1 Nominal Galileo prelaunch sequence of events for a May 1986 launch

LAUNCH + HOURS	1986											
	1	2	3	4	5	6	7	8	9	10	11	
• JPL GO-NO GO.....			△									
• SHUTTLE - CENTAUR SEPARATION*				△								
• CENTAUR BURN.....					□							
• SPACECRAFT BOOM DEPLOYMENT.....					▬							
• SPACECRAFT TRANSMITTER ON*.....					△							
• SPACECRAFT SPINUP AND SUN POINT.....					□							
• CENTAUR - SPACECRAFT SEPARATION.....						△						
• DEPLOY HGA AND PRESSURIZE RPM.....							□					
• DSN DOWNLINK ACQUISITION*.....								△				
• SUN AND DSN UPLINK ACQUISITION*.....									▬			
STS ORBIT		2	3	4	5	6	7	8				

\* THIS SEQUENCE COULD BE DELAYED 1 OR 2 ORBITS

Fig 2 Nominal Galileo postlaunch sequence of events for a May 1986 launch

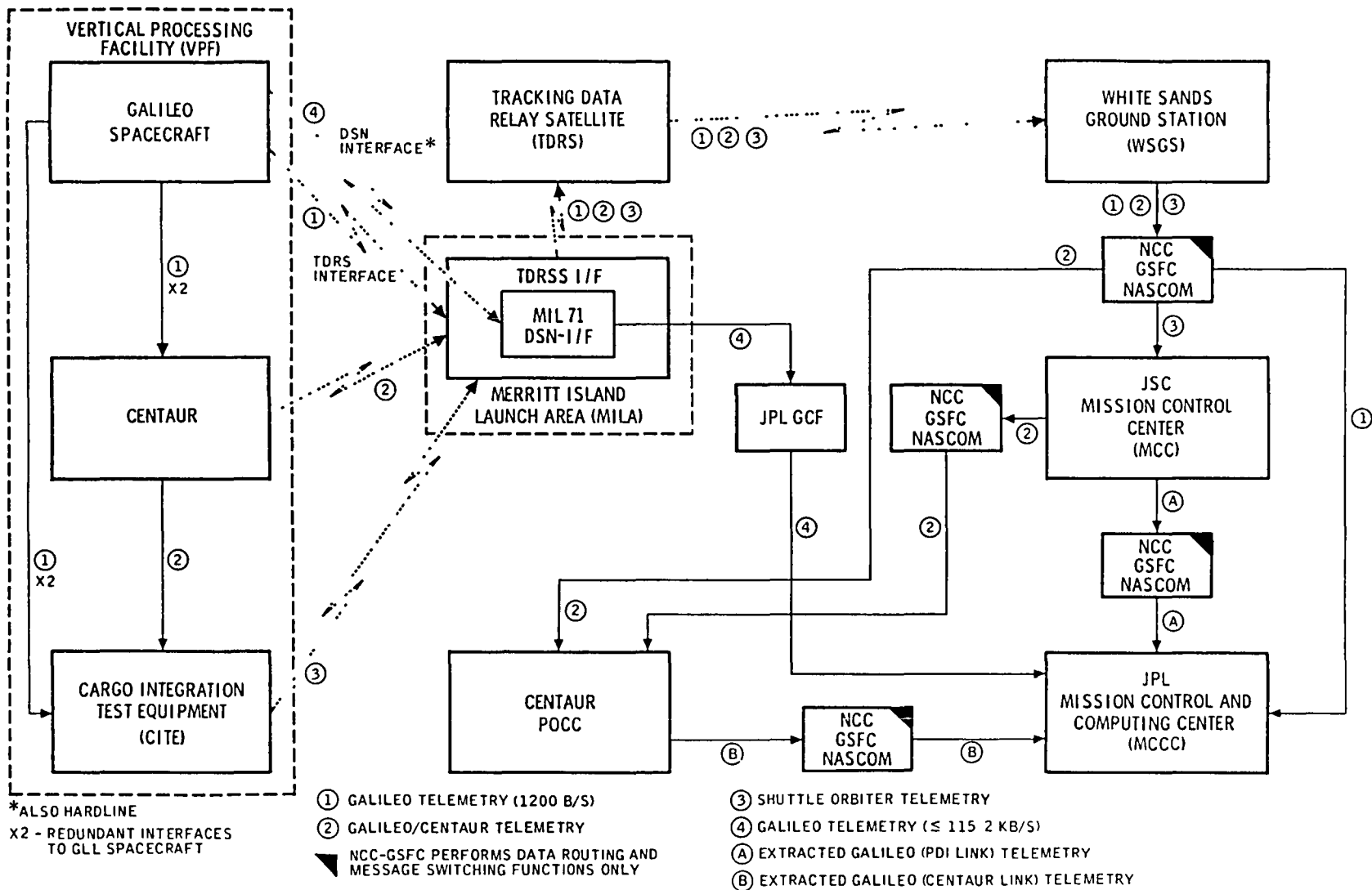


Fig 3 Telemetry data flow in the prelaunch phase – Vertical Processing Facility subphase

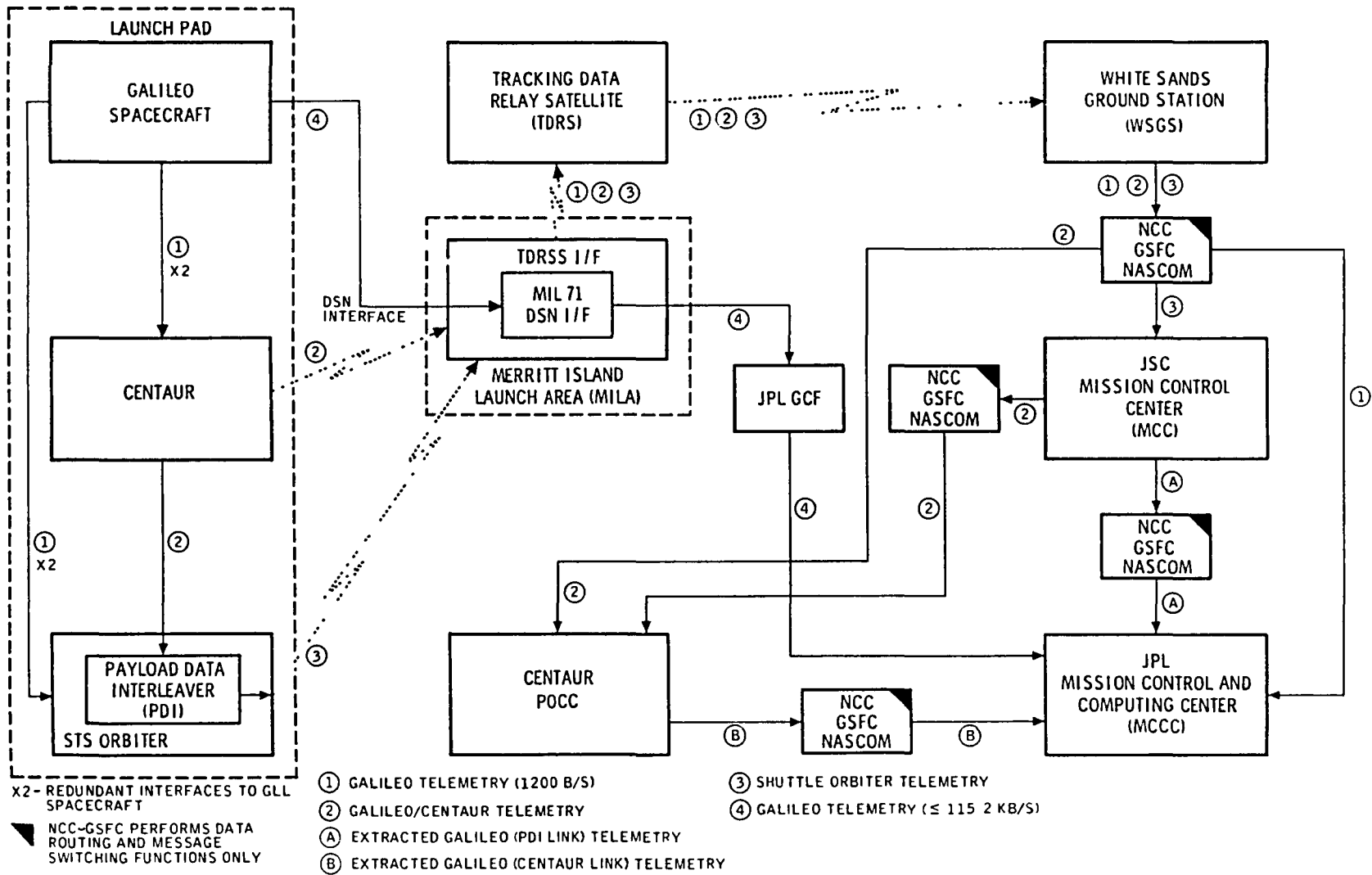


Fig 4 Telemetry data flow in the prelaunch phase – launch pad subphase

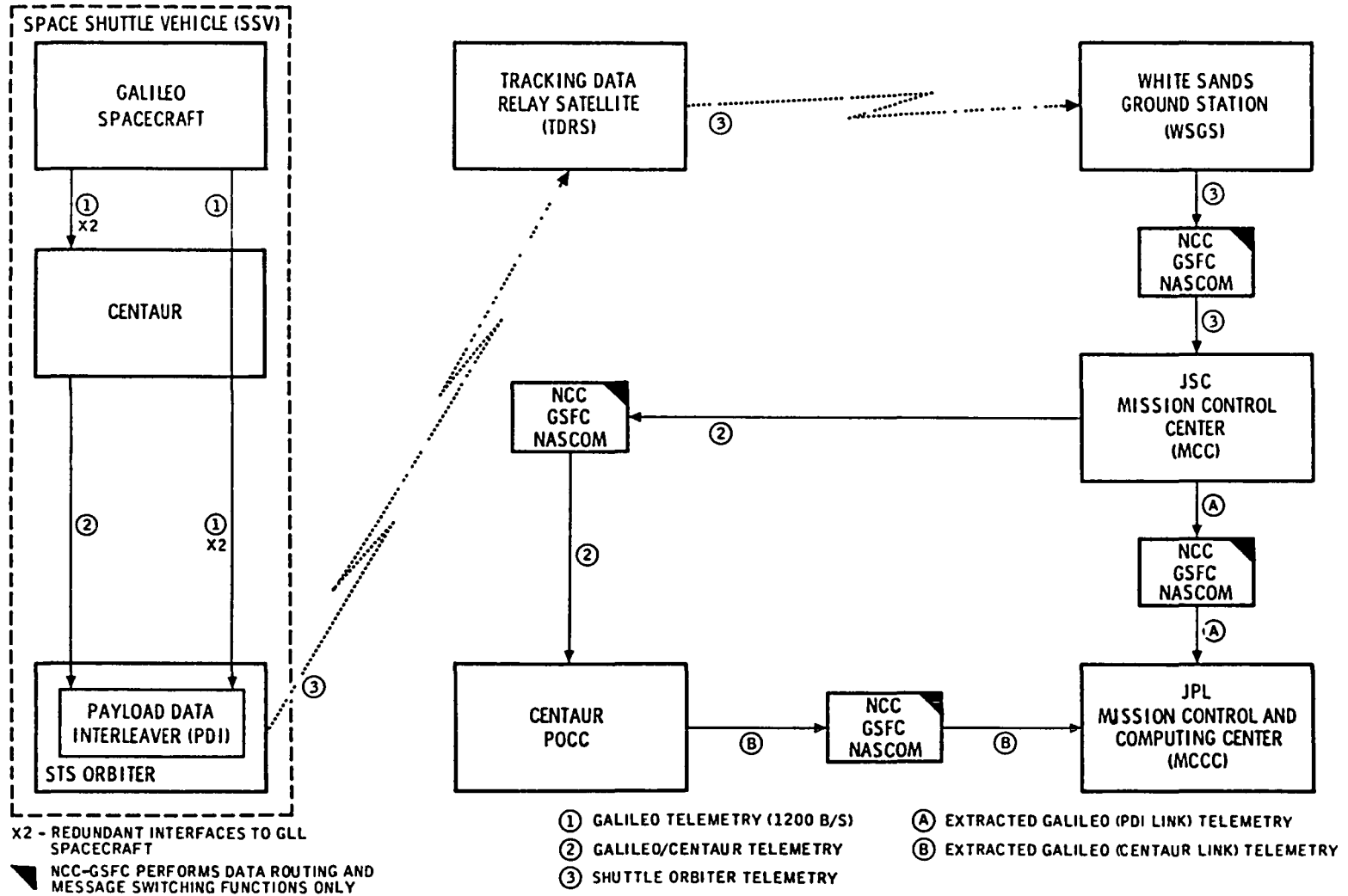


Fig. 5. Telemetry data flow in the Shuttle attached phase – ascent subphase



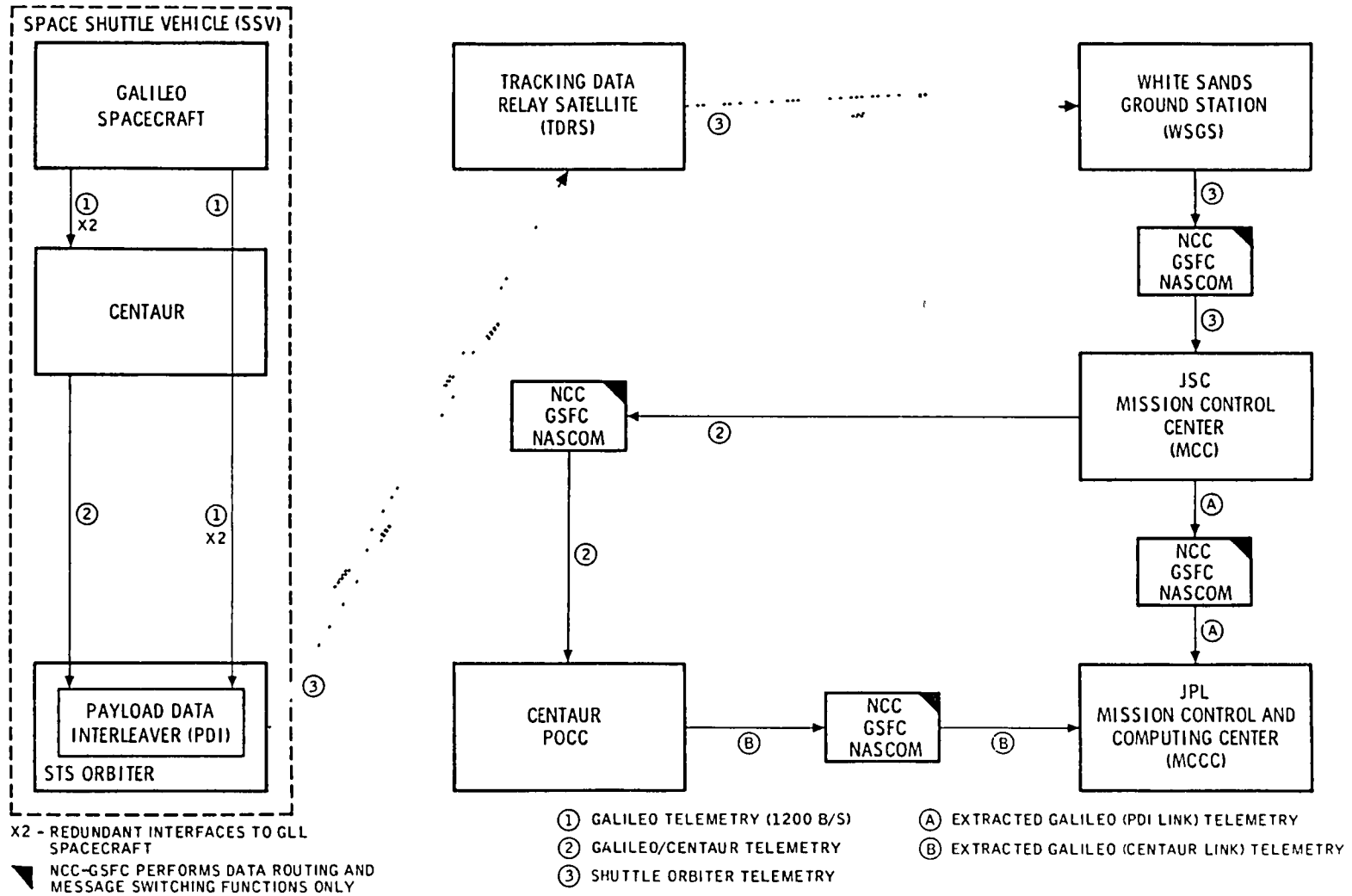


Fig. 6 Telemetry data flow in the Shuttle attached phase – Shuttle on-orbit subphase

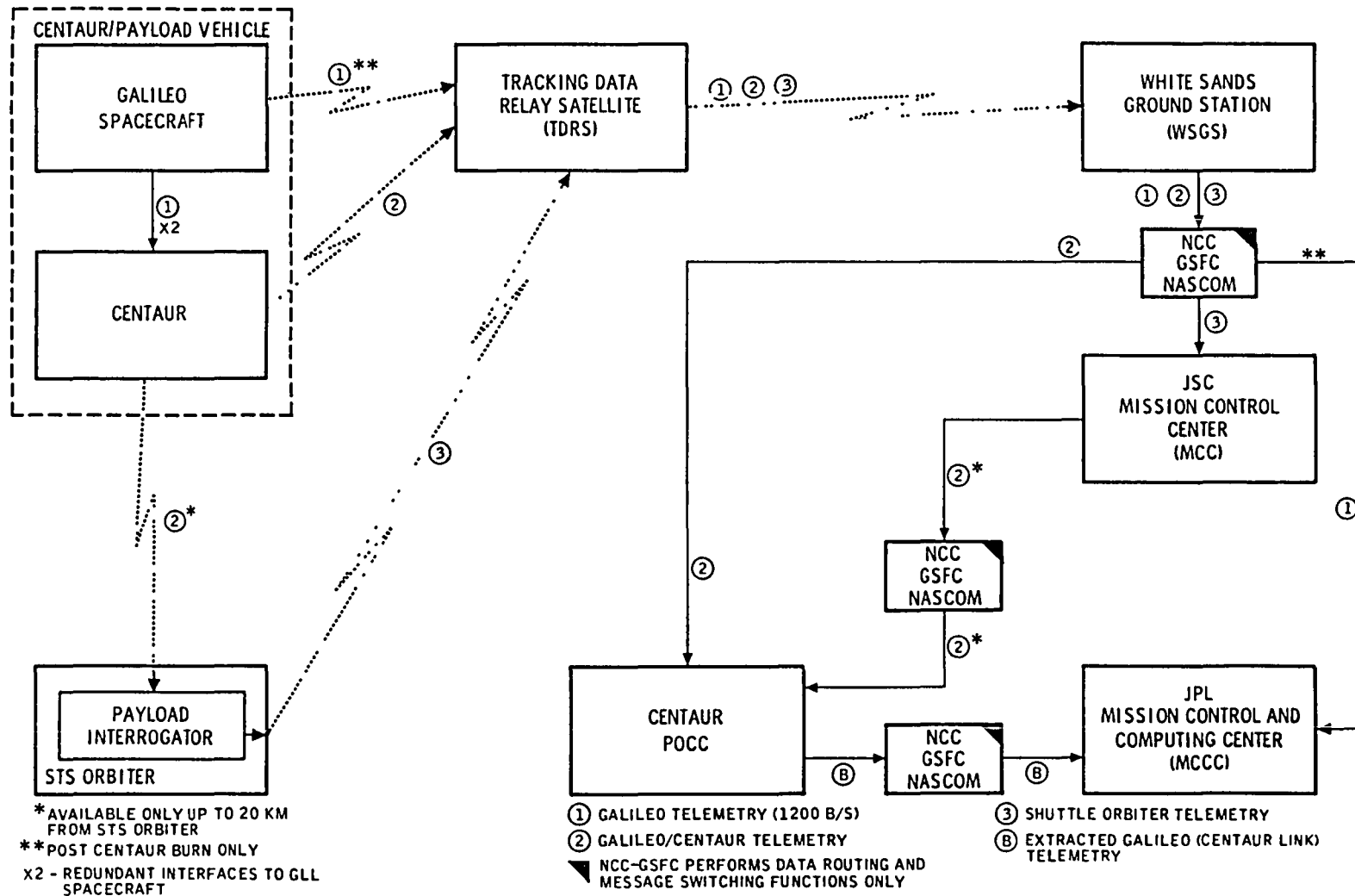


Fig 7 Telemetry data flow in the Shuttle detached phase

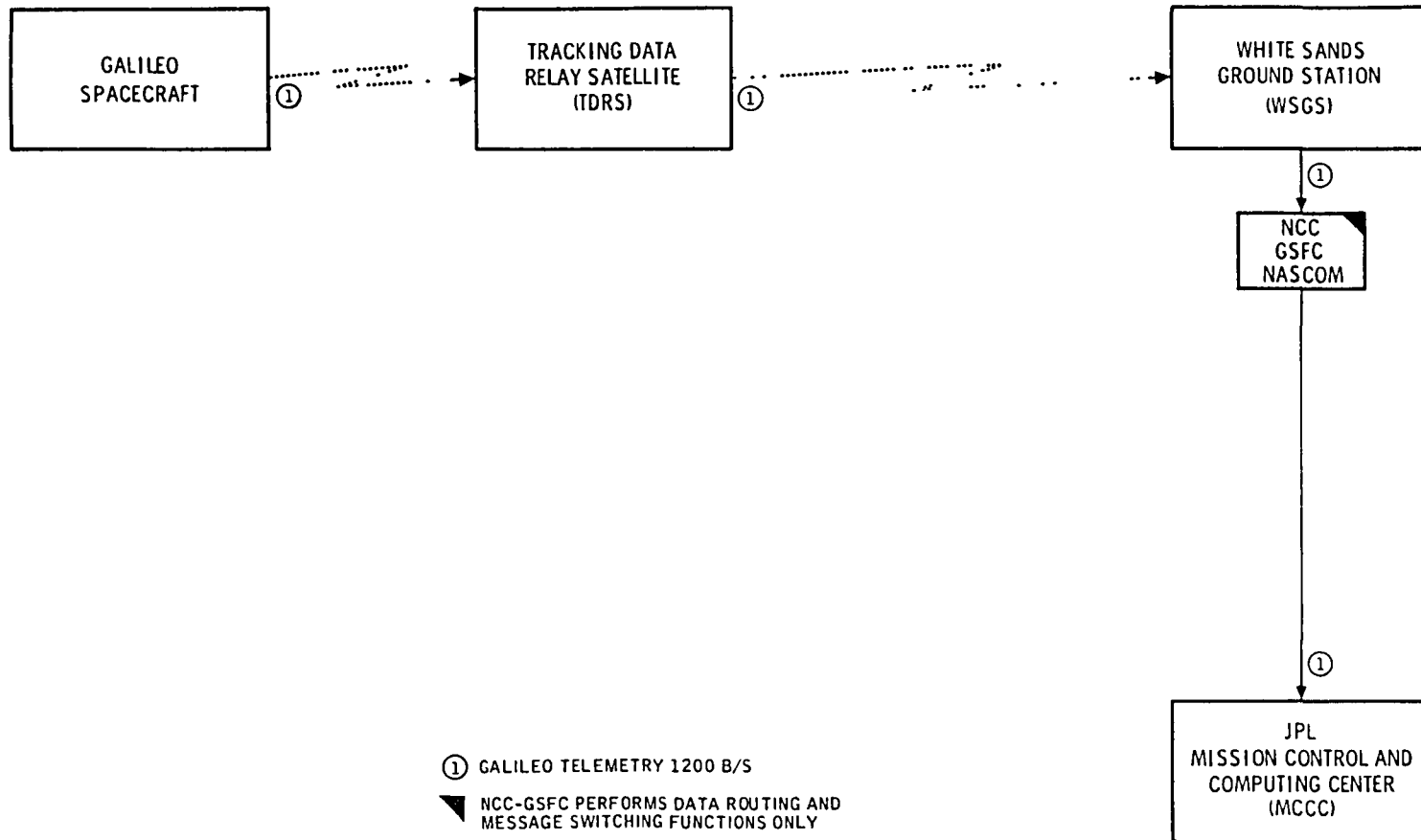


Fig. 8. Telemetry data flow in the Shuttle detached phase -- post Centaur separation phase

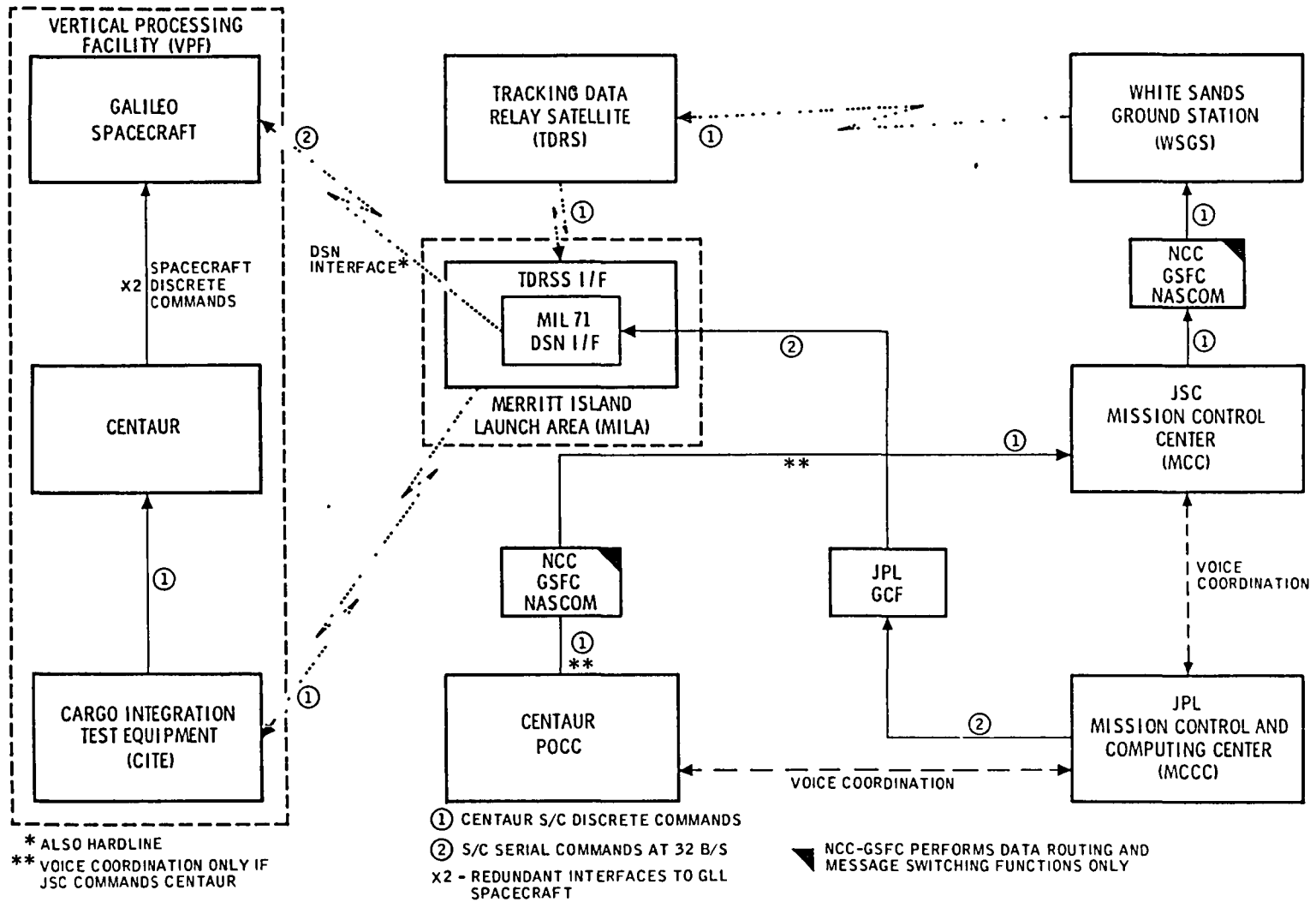


Fig. 9. Command data flow in the prelaunch phase – Vertical Processing Facility subphase

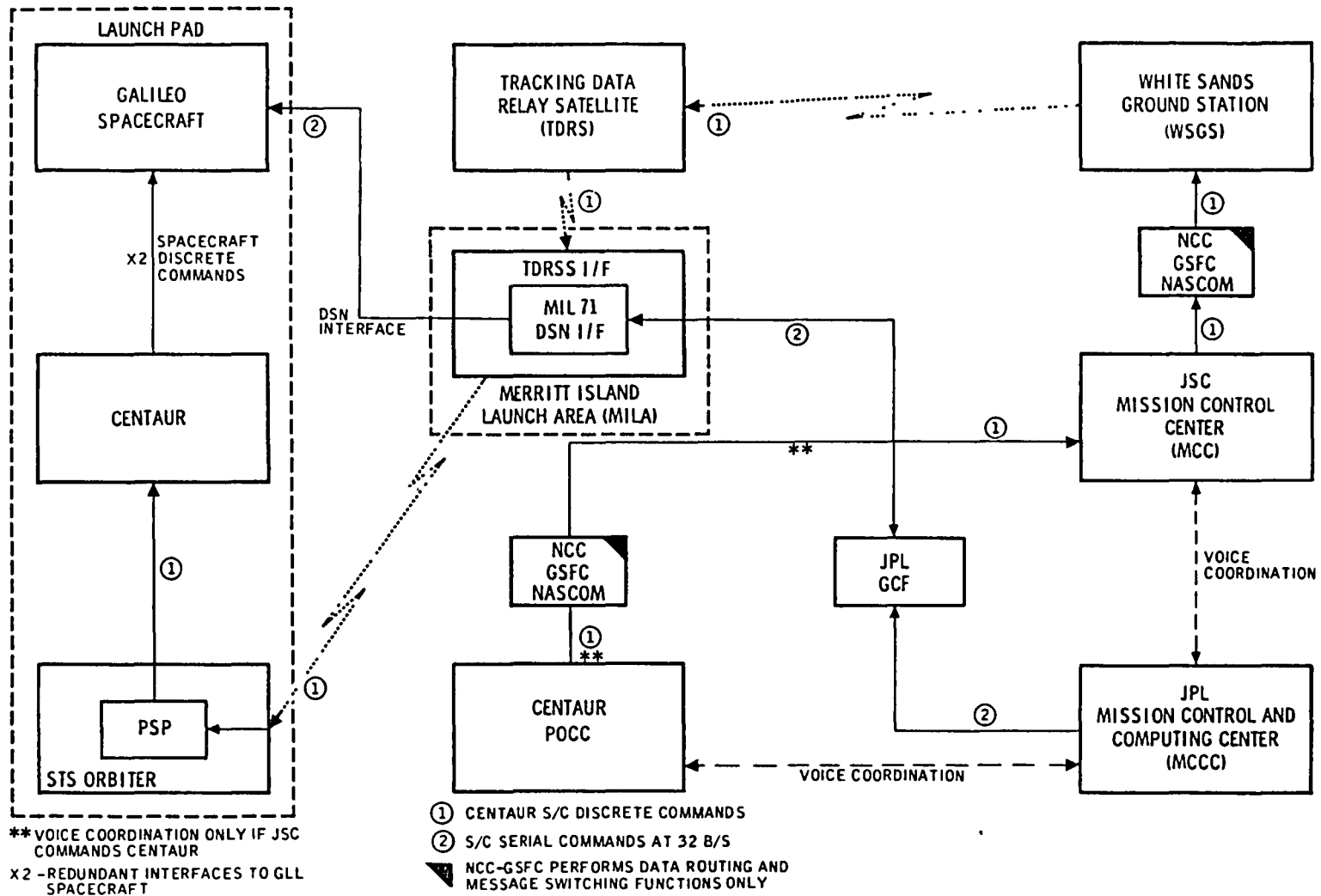


Fig 10. Command data flow in the prelaunch phase -- launch pad subphase

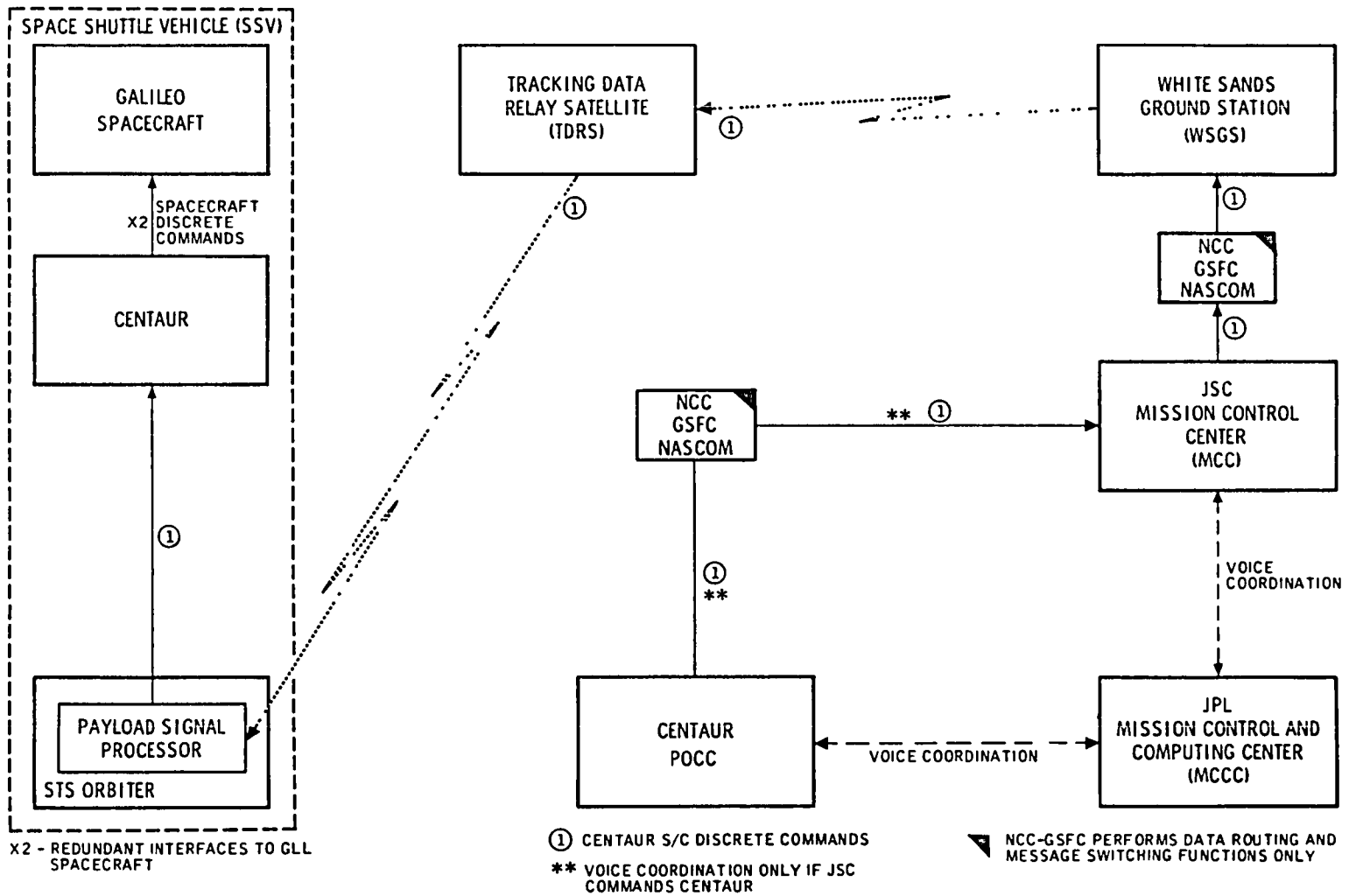


Fig 11. Command data flow in the Shuttle attached phase – Shuttle on-orbit subphase

# DSN Ground Communications Facility

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*A functional description of the GCF and its relationships with other elements of the DSN and NASCOM is presented together with development objectives and goals and comments on implementation activities in support of flight projects*

## I. Introduction

The Ground Communications Facility (GCF) is one of the three elements of the DSN. The GCF provides for transmission, reception, and monitoring of earth-based point-to-point communications between the Deep Space Stations (DSSs, one of the DSN elements), the Network Operations Control Center (NOCC, the other element) located at JPL, Pasadena, and the Mission Operations Control Center (MOC) at JPL. Voice, teletype, high-speed data, and wideband circuits of the world wide NASA Communications Network (NASCOM) are utilized for all long-distance circuits, except those between JPL and the Goldstone Deep Space Communications Complex (GDSCC) Goddard Space Flight Center (GSFC). NASCOM Engineering has delegated the responsibilities for planning, budgeting, design, implementation, operation, and maintenance of the communications requirements between Goldstone and JPL to the DSN GCF. Additionally, the GCF provides communications services between the DSSs at the communications complex at Goldstone, California, via intersite microwave systems and between separated areas of the NOCC at JPL via 230 kbit/s wideband data channels. Also, voice communications are provided within the complexes, and within the NOCC. The GCF is comprised of seven subsystems: Voice, Teletype, High-Speed Data, Wideband Data, GCF Monitor and Control, Data Records, and

Network Communications Equipment. The DSN Telecommunications and Data Acquisition Engineering Office of JPL provides the technical direction and systems management of the GCF and acts as the representative of NASCOM for switching and interconnect functions on the west coast.

## II. GCF-NASCOM Interrelationships

The interrelationships at the programmatic level between JPL's DSN GCF and the NASCOM Network, which is managed, engineered, and controlled at GSFC, are characterized as follows:

### NASCOM

- (1) Provides long-haul operational ground communications in support of all NASA projects and mission activities including those supported by the DSN.
- (2) Accepts and supports communications requirements established by the DSN and validated through continuing consultation and review.
- (3) Establishes in consultation with the users the basic characteristics of the NASCOM systems, such as teletype line rate and block header formats for switching, and the user electrical interfaces.

## GCF

- (1) Provides ground communications for all DSN missions and uses the services of NASCOM.
- (2) Establishes additional characteristics of all GCF subsystems on an end-to-end basis such as block multiplexing, error correction, GCF monitoring and control, and data records capabilities

### III. Objectives and Goals

The primary objectives of the GCF are to provide highest quality point-to-point transfer of operational data within the DSN and to provide simple user and NASCOM electrical and operational interfaces. These objectives are being met by:

- (1) Providing automatic message switching and routing
- (2) Providing data transmission subsystems that are as transparent to the user as possible
- (3) Minimizing project-dependent equipment within the GCF.
- (4) Providing a centralized common user data records capability.

The goals of the GCF are to provide highly reliable and cost-effective data transmission while continuing an adequate capability balance for multiple mission users. These goals include the following:

- (1) Equipment and routing redundancy to minimize single-point-of-failure impact
- (2) End-to-end performance which provides essentially block-error-free throughput
- (3) Design coordinated and consistent with the NASCOM Development Program.

### IV. Configuration and Functional Subsystem

The current GCF configuration, including the related NASCOM interfaces and functions, is illustrated in Fig 1. This configuration illustrates the long-haul communication circuit services external to JPL and Deep Space Communications Complexes which are the responsibility of NASCOM (except circuits between the Goldstone Complex and JPL). The point-to-point communications for Voice, Teletype, High-Speed Data, Wideband Data, and GCF Monitor and Control Subsystems are serviced by this Fig 1 configuration.

#### A. High-Speed Data Subsystem

This subsystem consists of the GCF assemblies that switch, transmit, record, process, distribute, test, and monitor digital data and is used for transmission of:

- (1) All digital data for the DSN Command, Tracking, and Monitor and Control Systems.
- (2) All low or medium rate data of the DSN Telemetry, Radio Science, Very Long Baseline Interferometry (VLBI), and the DSN Test Support System

The High-Speed Data Subsystem provides a capability for transmitting and receiving the serial bit stream block formatted data over a properly conditioned full duplex alternate voice/data channel having a 30-kHz bandwidth or over a time division multiplexed 56-kb/s satellite circuit. This serial bit stream is impressed on communication circuits at a continuous line bit rate divided into message segments referred to as high-speed data blocks. Two types of data blocks are used:

- (1) Data blocks containing user data bits to be transmitted
- (2) Filler blocks containing filler data bits provided by the GCF when the user data rate is insufficient to maintain contiguous blocks on line required for continuous line monitoring and error control.

Current capabilities for the GCF Mark III period provide the functional capabilities illustrated in Fig 2. The GCF High-Speed Data Subsystem is standardized on a 1200-bit block size (message segment) and a line bit rate of 7200 b/s. The subsystem capabilities include a 22-bit error detection encoding/decoding polynomial code, two error status bits, and error control using two eight-bit fields. The error control field facilitates numerical serialization and acknowledgement numbers for error correction by selective retransmission. (The error correction capability significantly reduces the post-pass time required for non-real-time replay of blocks received in error.) Figure 3 illustrates the High-Speed Data Subsystem and the GCF Subsystems located at the Central Communications Terminal.

#### B. Wideband Data Subsystem

The Wideband Data Subsystem consists of assemblies that switch, transmit, receive, process, distribute, test and monitor data requiring the use of bandwidths greater than those provided by standard high-speed data channels. The GCF Wideband Data Subsystem functionally illustrated in Fig. 4, together with a listing of functional capabilities provided, includes standard wideband circuits as well as all intersite



microwave (area microwave) capabilities. The Wideband Data Subsystem is used for the transmission of

- (1) All DSN Telemetry System high-rate data that exceed High-Speed Data Subsystem capabilities
- (2) Data interchange between the NOCC and GCF Communications Terminal at JPL
- (3) Data interchange between DSSs within a complex via intersite microwave, including critical timing signals and receiver baseband signals for antenna arraying and signal combining systems support
- (4) Simulation System Data from the Mission Control and Computing Center/Mission Operations Center to the DSSs
- (5) DSN Test Support System data from the Network Operations Control Center to the DSSs.

The wideband data circuits for interchange of data between the DSSs and JPL are impressed with serial bit streams at a continuous line rate, typically 56 or 230.4 kb/s, divided into 4800-bit message segments (data blocks). Similar to the High-Speed Data Subsystem, the blocks are either data blocks or filler blocks inserted when the user data load is insufficient to maintain contiguous data blocks.

### C. Voice Subsystem

The Voice Subsystem consists of GCF assemblies that switch, transmit, receive, distribute, test, and monitor transmissions originally generated in vocal form, and includes internal voice communications within the Deep Space Station Communications Complexes, DSSs, and the NOCC. The subsystem service provides capabilities between those areas and to non-DSN area interfaces as follows:

- (1) NOCC and DSS
- (2) NOCC and MCCC/MOC (or remote MOC).
- (3) MOC and DSS for Command System backup

The Voice Subsystem functional capabilities and key characteristics include:

- (1) Standard voice-data grade circuits for all traffic
- (2) Conferencing capability on one intercontinental circuit during noncritical periods for all Deep Space Stations supporting a single project (individual circuits for each DSS during critical periods, resources permitting)
- (3) User-controlled intercomm switching
- (4) Circuits used for high-speed data transmission (backup) if required

- (5) Voice traffic recording in the Central Communications Terminal upon request

### D. Teletype Subsystem

The teletype (TTY) subsystem uses an eight-level ASCII (American Standard Code for Information Interchange) national standard. The subsystem consists of assemblies that switch, transmit, receive, distribute, test and monitor digital signals at a TTY line rate of 100 words per minute. The operational use of teletype continues to be deemphasized and is used primarily for emergency backup operational transmissions and administrative communications. Service functions and key characteristics include:

- (1) Handling Eastern Space and Missile Center (ESMC) generated predicts for DSN initial acquisition
- (2) Transmitting nonoperational messages between the JPL Message Center and other locations
- (3) Use of standard NASCOM format and the NASCOM communications processor for message switching
- (4) Employment of time division multiplexing techniques to reduce trunk circuit costs

### E. GCF Monitor and Control Subsystem

The GCF Monitor and Control Subsystem consists of assemblies that collect, process, and display the status and performance of the GCF Subsystems in real-time. The GCF Monitor and Control Subsystem functional capabilities are illustrated in Fig. 5. Functions are implemented in minor subassemblies located at each DSS to interface station GCF status and performance indicators to the CMF for monitor block formatting and transfer to the Central Communications Monitor (CCM) Processor at JPL. The CCM also receives real-time status and performance information from local GCF subsystems. All real-time status and performance information received by the CCM is processed and displayed relative to preset standards and limits to facilitate operations monitoring and technical control. Information and alarms are displayed on continuous line performance and data flow throughput including error control.

### F. Data Records Subsystem

The DSN requirements for the data record processing and production functions are implemented in the GCF Data Records Subsystem. The Data Records Subsystem consists of assemblies in the CCT that log in real-time, monitor, identify gaps, provide for processing and editing of data gap lists, control data gap recalls from the DSSs and generate and account for Intermediate Data Records (IDRs) and selected fill-data records.

The Data Record Subsystem maintains accountability of high-speed and wideband data, performs automatic recall (under operator control) of missing data, and generates near-real-time Intermediate Data Records for delivery to the Mission Control and Computing Center

### **G. Network Communications Equipment Subsystem**

The Network Communications Equipment (NCE) Subsystem consists of GCF minicomputers and peripheral I/O assemblies that switch, transmit, and receive data. The NCE Subsystem assemblies are located in the Network Data Processing Area (NDPA). The NCE assemblies comprise a GCF Area Communications Terminal located in the NDPA.

The NCE interchanges multiplexed block formatted data with the ECS assembly located in the CCT over full-duplex 230-kb/s wideband data channels (see Fig 6). This GCF-NDPA interface function provides for

- (1) Processing data for transmission to and accepting data from the GCF CCT
- (2) Multiplexing/demultiplexing and buffering data for all NDPA processors
- (3) Routing data to and from NDPA processors.

### **V. Typical Configuration**

The DSN GCF is designed for multiple mission support. Improvements and additions are integrated to meet new era and project requirements. Support for the Voyager, Pioneer 6 through 12, Viking and Helios missions is continued. The capability for West Coast Switching Center support of the

IRAS Project has been completed. Requirements for the AMPTE, ISPM and Galileo projects are being defined and included in ongoing plans. Figure 7 illustrates, in general, the GCF configuration or support of these projects. Additionally, remote information centers and other non-DSN NASCOM-serviced installations on the west coast are serviced through the NASCOM West Coast Switching Center, an integral part of the GCF 20/Central Communications Terminal at JPL.

### **VI. Implementation Activities**

Implementation planning and design to meet the requirements of the DSN's Mark-IV long range plans were begun in late 1979. The priority design and implementation effort has been to consolidate the data communications interfaces to a single communications processor interface at each of the three Deep Space Communications Complexes (DSCCs). This single interface processor will accommodate the DSN's Networks Consolidation Program plans to simultaneously support multiple spacecraft from a single signal processing center located at each DSCC. Other less significant GCF changes and additions required within the DSCCs along with changes at the CCT are included in this effort to be completed in 1985. With the consolidation of the networks, the GCF will see a significant increase in data communications bandwidth requirements, as both deep space and high elliptical near-earth orbiting spacecraft are tracked, and acquired data will pass through the single communications processor at the DSCC. Modifications will be made in the CCT to accommodate the new DSCC interfaces and the new type of remote operations control centers.

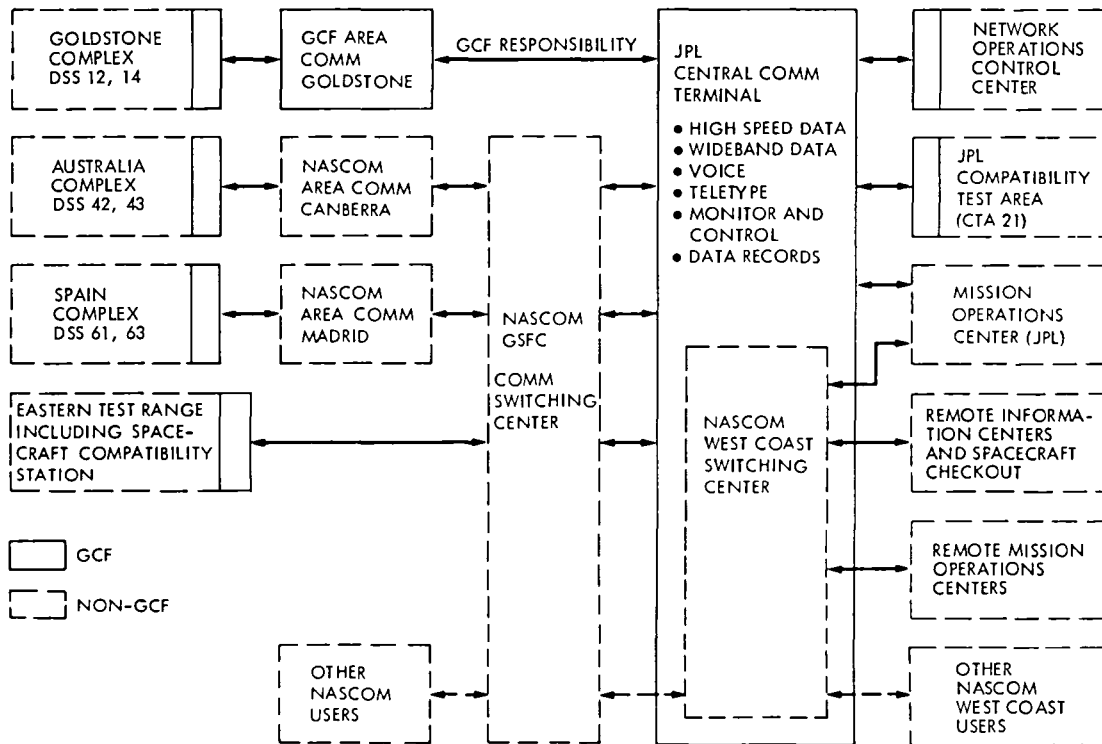


Fig 1. GCF configuration

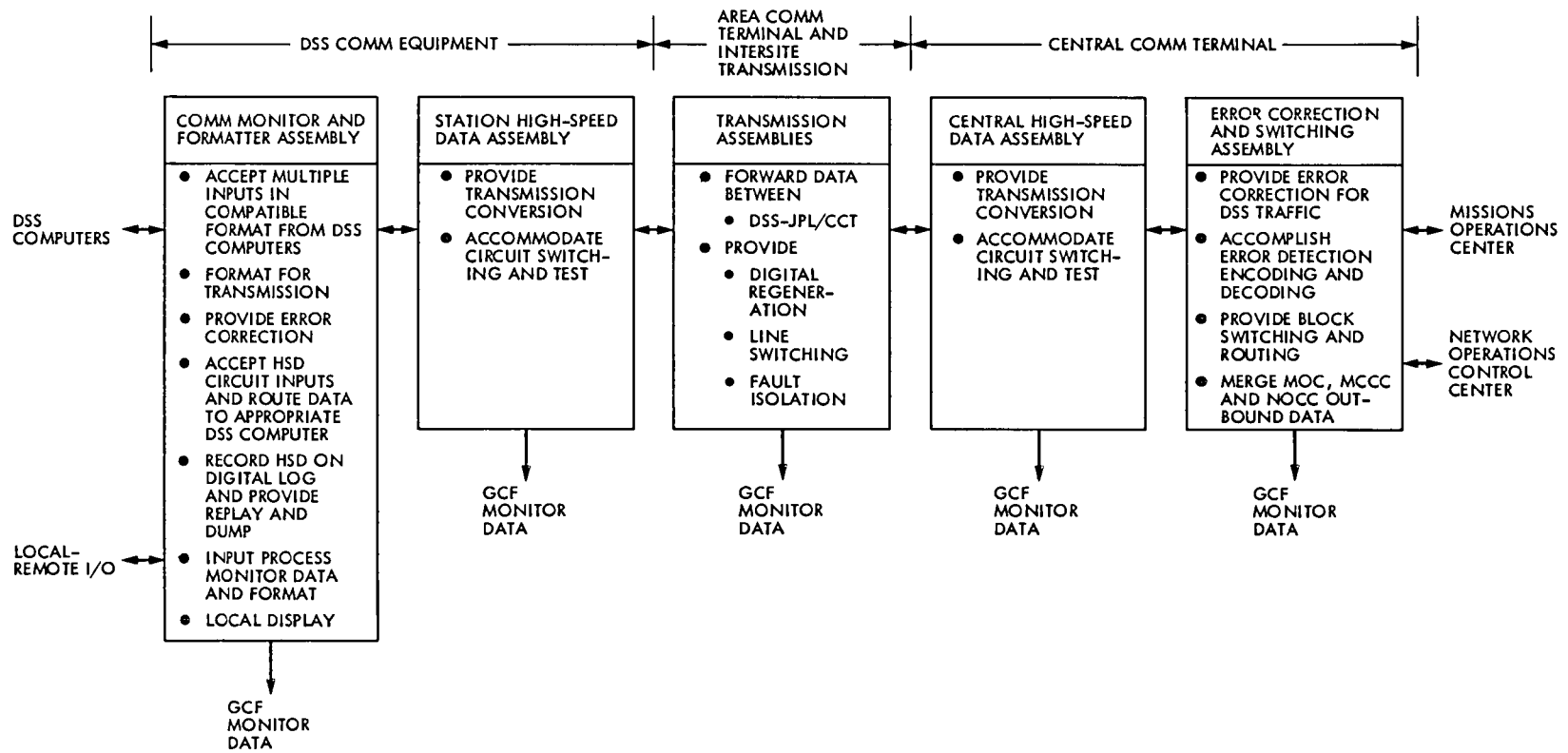


Fig. 2. GCF High-Speed Data Subsystem functional capabilities

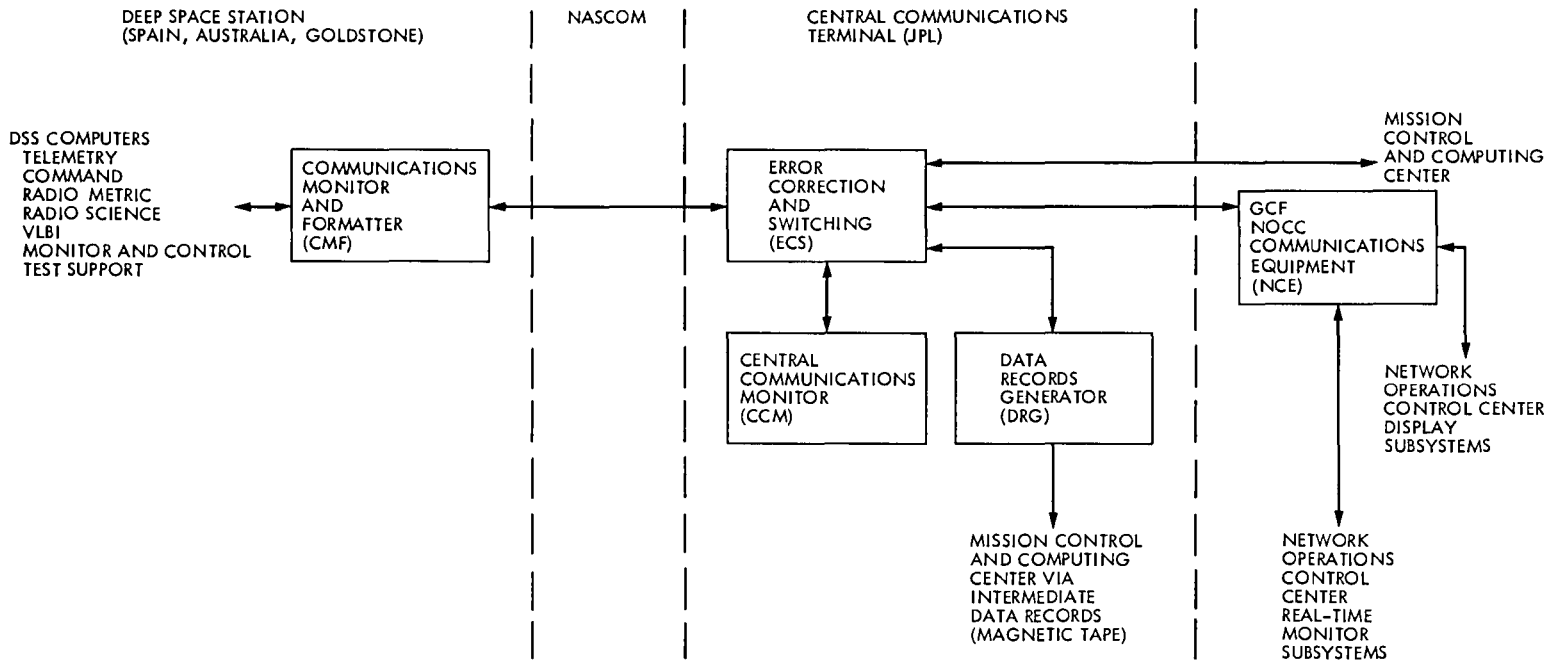


Fig. 3. GCF High-Speed Data Subsystem and Central Communications Terminal Subsystems and interfaces

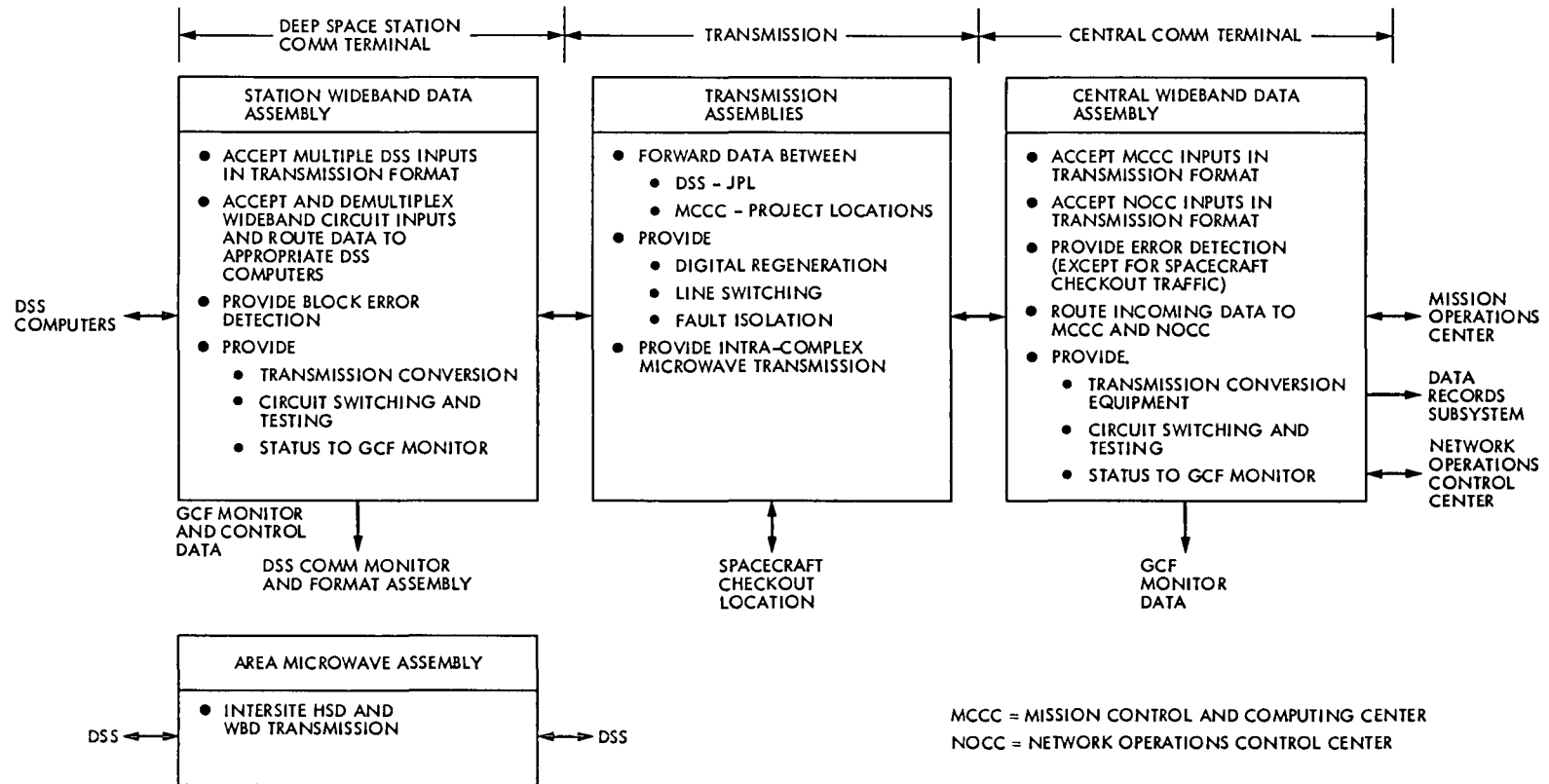


Fig 4 GCF wideband subsystem

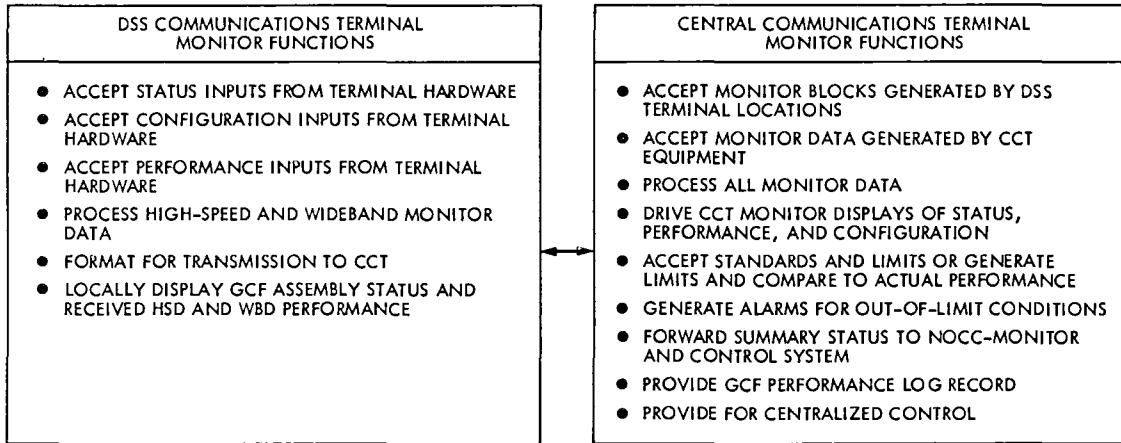


Fig. 5 GCF Monitor and Control Subsystem functional requirements

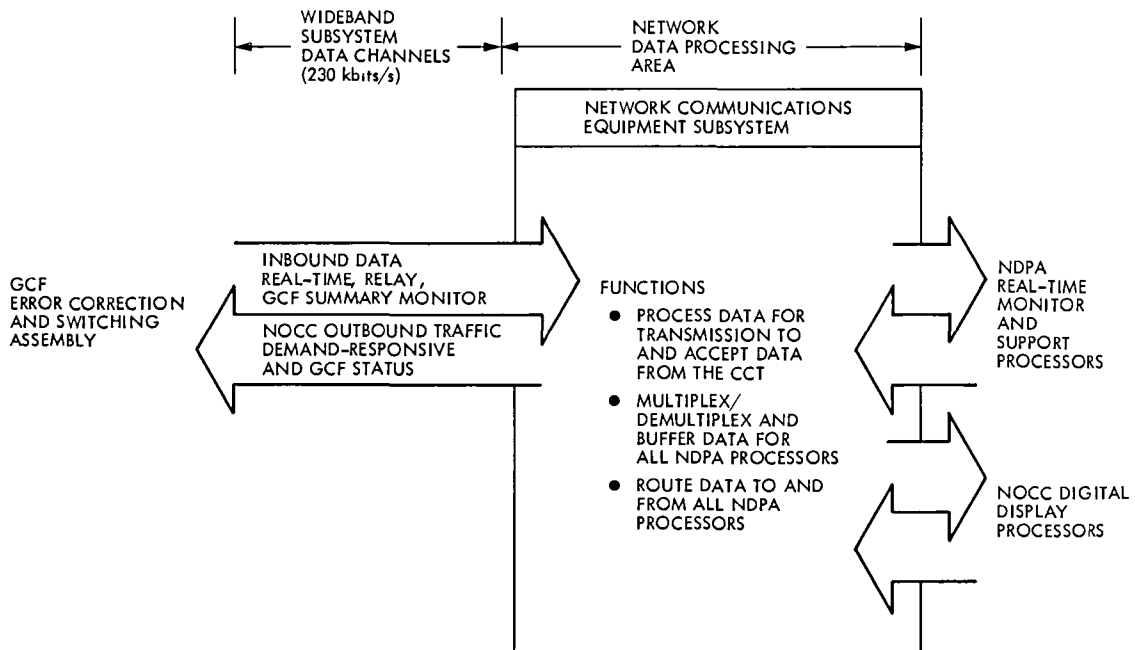


Fig 6 GCF Network Operation and Control Communications Equipment Subsystems

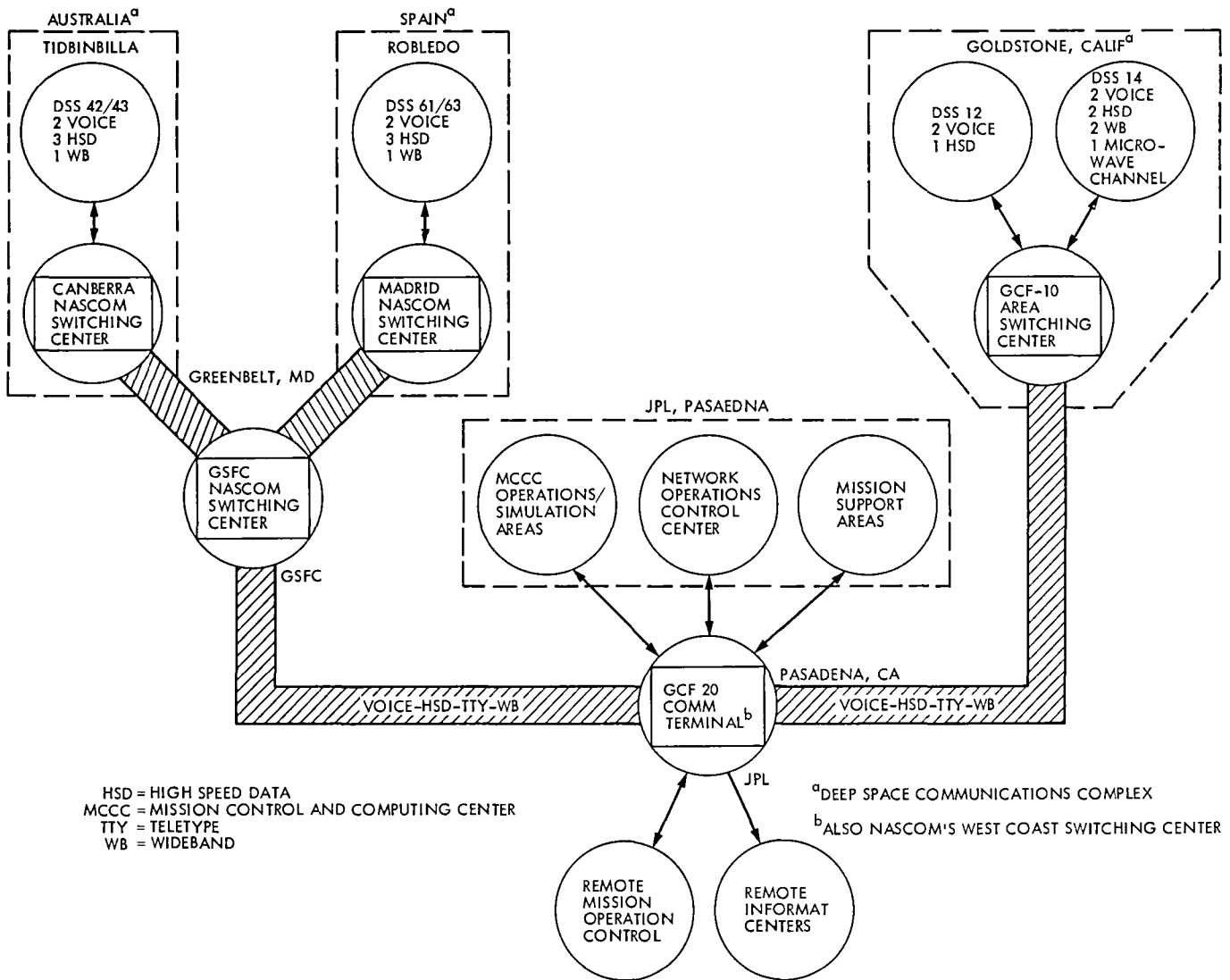


Fig. 7. DSN support locations and GCF-NASCOM circuit requirements



# MV-3 Communication Subsystem

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*The Operational Mobile VLBI Data Acquisition System (MV-3) being designed by JPL requires that a voice link be established between the mobile data system and a fixed site. A communications subsystem was incorporated in the MV-3 design which consists of HF radio, VHF mobile radio telephone and conventional land line telephone. The HF antenna design was optimized for short and long range transmission using both inverted V and yagi antennas mounted on a self-supporting telescoping mast.*

## I. Introduction

The Operational Mobile VLBI Data Acquisition System (MV-3) is being implemented by JPL in support of the NASA Crustal Dynamics Project. MV-3 is also known as Operational Radio Interferometry Observing Network (ORION) and consists of an electronics van and 5-meter dish antenna with transporter. Operating in conjunction with one or more VLBI data acquisition systems, it will provide a means of measuring length and orientation of vectors between sites in areas of geophysical interest. The technology used to accomplish this is Very Long Baseline Interferometry (VLBI, Ref 1).

A communications subsystem (COM) was incorporated to establish a voice link between the mobile data system and a fixed site. This is necessary to coordinate experiment start times and help solve problems as they arise during setup or during data acquisition. Reliable communications are also necessary for safety reasons while MV-3 is operating in remote locations.

A single control station will be used to coordinate all mobile and fixed data acquisition systems during an experi-

ment. Therefore, MV-3 need only communicate with this one station which can act as a relay for messages to other stations.

## II. Requirements

The MV-3 Design Requirements Document (an internal JPL document) forms the basis for COM requirements. Generally, COM must include those transceivers and antennas necessary to provide 90% reliable voice contact communications to the control station from any place in the continental United States. Practical reasons dictated the limitation to the continental U S in that it would be difficult to design a system around all possible foreign regulations and systems available. Foreign operation may therefore require tailoring the COM capability to meet the requirements.

Because MV-3 is a mobile system, COM must also meet the following requirements:

- (1) COM components must survive road vibrations and stress encountered while in transit.

- (2) Equipment mounted to the exterior of the electronics van must meet provisions set down by state vehicle codes. The design requirements specifically prohibit the use of equipment which would require that MV-3 obtain special road permits.
- (3) COM must meet its reliability requirement under a wide range of terrain conditions. Mountains and valleys must be considered in the design.
- (4) The setup and teardown of the COM subsystem should be held to a practical minimum. Ease of operations can considerably lower cost to operate MV-3.

### III. Tradeoff Studies

The following seven modes of communication were considered for their merits in optimizing the MV-3 COM design.

- (1) Land lines.
- (2) Low and medium frequency radio
- (3) High frequency radio
- (4) Repeater communications
- (5) VHF radio telephone.
- (6) Satellite microwave
- (7) ATS satellite.

#### A. Land Lines

The land line system considered here is that which is commercially available from the telephone company. Once linked, it achieves better than the 90% reliability imposed on COM and is rarely affected by unpredictable environmental conditions.

The land line system has two drawbacks. First, it depends on the availability of the lines. Some remote locations would require a special hookup, raising the cost of the system. Second, it poses a logistics problem. Arrangements must be made weeks in advance for the connection. Also, a representative of the user must be available at the site to coordinate the placement of the lines. The land line system is therefore practical only in cases where other systems cannot be used, or when the length of stay at a particular site warrants its installation.

#### B. Low and Medium Frequency Radio

Radio communication is generally a practical alternative to land lines. Each carrier frequency subdivision has its own main characteristics, as summarized in Table 1. Low frequency (LF) radio is generally good for medium and long range transmission, either by traveling close to the ground (ground wave)

or reflecting off the ionosphere (sky wave). Medium frequency (MF) transmission is good for short range communications such as local AM broadcasting. At night, MF can sometimes bounce off the ionosphere, which increases its range. Picking up a faraway AM broadcasting station at night is a typical example of this.

One serious disadvantage of LF or MF is that the antenna is extremely large. The length of an efficient antenna is typically either  $1/4$  or  $1/2$  wavelength, which implies that the antenna would be tens or hundreds of meters long. This is practical only in fixed locations where such an antenna could be permanently installed.

#### C. High Frequencies

High frequency (HF) transmission is generally good for short distance ground wave and long distance sky wave communication. Figure 1 shows the typical range for HF, which has a wavelength from 10 to 160 meters. With multiple hops, HF will also transmit over intercontinental distances. Ground wave transmission is usually useable up to 320 km, but depends greatly on terrain, while sky wave transmission works at distances greater than 160 km.

HF has two drawbacks. First, it still requires a large antenna. Typical HF antennas are from 3 to 60 meters in length depending on the frequency. Second, proper ionospheric conditions are required for reliable transmission. HF alone would not meet the reliability requirement imposed on COM.

#### D. Repeater Communications

Frequencies above the HF band are generally good only for ground wave transmission and require repeaters to extend its range. Frequencies in this range include very high frequencies (VHF), ultra high frequencies (UHF), and microwave. Repeater stations are usually on mountain tops, and receive the incoming signal while retransmitting the same signal to a receiver that would otherwise be out of range of the first transmitter. The advantages of this system are: First, these frequencies are only negligibly affected by atmospheric or environmental conditions. Repeater communications are therefore very reliable as long as the user's transceiver is in range of a repeater. Second, the antennas for this mode of transmission are small, typically a few feet, which simplifies design and operations. The disadvantage of this scheme is the requirement for costly repeater systems throughout the area of anticipated use. This system would only be practical for COM if existing repeater systems are used.

#### E. VHF Radio Telephone

Mobile VHF radio telephones have all of the characteristics of a repeater communications system. A critical advantage is

that radio telephones use a repeater network owned and operated by the telephone company and its contractors. Through mobile telephone charges, repeater time is rented and shared by multiple users, bringing the cost down to an affordable value. This system is designed around multiple channels shared by mobile telephone users in the immediate area. Direct dial and automatic scanning features will be available this year, which will eliminate the requirement that channel number be coordinated before the call is placed.

The disadvantages of radio telephone are: First, the system is not designed to reliably cover remote areas, particularly off-highway. Also, mountains in the radio path to the repeater can reduce its reliability considerably. Second, the radio telephone channels tend to be overcrowded in populous areas. Some waiting period could be encountered before contact is made.

#### F. Satellite Microwaves

As the frequency spectrum has become crowded over the years, microwaves have become popular for transmitting information and messages. Microwaves are negligibly affected by atmospheric conditions, but because of their short wavelength ( $< 0.3$  m), transmission is in a strict line of sight path and thus requires repeaters to extend the range. Satellite repeaters are practical for this purpose because of the large area of the earth covered by each satellite. Numerous microwave transceivers can beam signals to a satellite repeater for reliable short or long range transmission.

There are two considerable disadvantages of satellite microwave systems for MV-3. First, a large dish antenna (3 to 5 meters) must be pointed at the satellite for transmission. MV-3 already incorporates a dish antenna, and to add another would raise the costs significantly and decrease its mobility and versatility. Second, renting satellite time is costly.

#### G. ATS Satellite

NASA-Goddard Space Flight Center is currently coordinating the use of two VHF communications satellites (ATS) which are available to government experimenters on a priority basis as determined by NASA. When available, these satellites would provide reliable communication for MV-3 operations. There are, however, three serious disadvantages. First, use of the satellites must be coordinated well in advance of anticipated use. This generates manpower requirements to coordinate the logistics and also disqualifies the system for emergency use. Second, since ATS is shared by other users, there may be times when the system is unavailable to MV-3. Third, ATS is an experimental satellite with no guarantees that it will continue to function for an extended period of time.

### IV. Systems Chosen

From the characteristics of each system, it was decided that COM incorporate the following communication modes: (1) land lines, (2) VHF mobile radio telephones, (3) HF radio. Normally the VHF mobile radio telephone will be used where available because of its ease of use. The HF radio will back up the radio telephone as necessary. If a long stay at a site is anticipated, MV-3 will be configured with standard telephones so that a direct land line connection is possible.

Both the mobile radio and land line telephones are standard telephone company installations. A 1-m whip antenna will be installed on the van roof with a cable leading to the VHF transceiver mounted in an equipment rack. MV-3 is equipped with a 12-Vdc battery system which will power the mobile telephone. This system will therefore be useable before the MV-3 generators are started.

The HF radio requires multiple frequencies on which it will operate to ensure reliable transmission over a wide range of distances (Fig. 2). The lowest frequency (channel) is determined for its ability to transmit over short distances. Higher frequency channels should progress in octaves to a channel with long range transmission characteristics. COM will use the frequencies issued to NASA (namely 3 395, 6 9825 and 14 455 MHz) which have the necessary qualifications.

### V. HF Antenna Tradeoffs

A well designed antenna is critical for reliable HF communication. Because of its size, it also warrants the most attention to optimize ease of operations. The following antenna types were considered for tradeoffs for the HF system (Fig. 3): (1) whip, (2) horizontal dipole, (3) inverted V dipole, (4) yagi.

#### A. Whip

The whip antenna has two advantages. First, as can be seen from its radiation pattern (Fig. 4), the whip transmits best on horizontal paths. This quality is excellent for transmitting ground wave and long range ( $> 750$  km) sky wave. In both cases the radiation angle (angle between the main transmission lobe and horizon) is small. Second, the whip transmits in all horizontal directions (omnidirectional), eliminating a need for antenna pointing.

The whip has two disadvantages. First, it does not transmit well at high angles required for short distance sky wave transmission (150-750 km). This mode is usually necessary in mountainous areas when ground wave communication is not possible. Second, the whip requires a resonant ground plane for optimum performance which would be difficult to set up.

The electronics van roof might be used as a substitute but would not provide optimum performance

## B. Horizontal Dipole

The horizontal dipole transmits well both horizontally and vertically at all but very low radiation angles. Normally one thinks of the horizontal dipole as having a figure-eight shaped radiation pattern, which is only true in free space. A computer-generated mathematical model (Ref 2) simulated the radiation patterns as a function of antenna height. These are shown in Fig 5 and were confirmed by antenna range tests. This antenna is generally good for short and medium range (0-1800 km) transmission using both ground and sky waves. Another advantage of the dipole is that no ground plane is required for good performance.

A disadvantage of the horizontal dipole is that it requires two long support towers, which would be difficult to erect. In order to achieve the desired radiation angle, both towers must be more than 10 km high. (This antenna height also helps match the antenna impedance to the 50 ohm transceiver as shown in Fig 6). Further difficulties arise in positioning the towers for cases in which the electronics van is parked near a hill.

## C. Inverted V Dipole

The inverted V dipole is different from the standard dipole only in that the antenna ends are brought down close to the ground. A similar computer-generated antenna pattern (Fig 7) shows that the inverted V, like the horizontal dipole, transmits well horizontally and vertically. Like the horizontal dipole, the inverted V does not require a ground plane and will transmit reliably up to distances of about 1800 km.

An important advantage of the inverted V over the horizontal dipole is that it requires only one high support tower. This tower can be mounted on the electronics van with the dipole ends brought down to near ground level. Because the inverted V is nearly omnidirectional, any convenient orientation can be used. The antenna ends can be held in place by tripods or sandbags, depending on terrain. Also, because of the antenna shape, the inverted V requires less space than the horizontal dipole.

## D. Yagi

The yagi antenna is a very directional antenna which focuses almost all of the radiation energy in one direction (Fig 8). Because of the large gain in signal strength in the one direction, the yagi will exceed the performance of other antennas when aligned properly with the point of contact. Unfortunately, the yagi is very large for HF frequencies.

Fixed-installation yagi antennas are rarely constructed for frequencies less than 6 MHz. Because of the mobility requirements of MV-3, the yagi will only be practical with the 14 455-MHz channel. Even at this frequency the boom length is about 25 feet, which is almost the limit of what the MV-3 crew could easily erect. Fortunately, the 14 455-MHz channel will usually only be used for long distance communication when the extra antenna gain is needed. Therefore the yagi is only practical in this case for long range transmission (> 1800 km).

## VI. Antenna Selection and Implementation

Using the previous information it was decided that COM incorporate both the inverted V dipole and the yagi antennas. For most applications of short to medium range contact (0-1800 km), the omnidirectional inverted V will be used. The yagi, which will be more difficult to erect, will be used only for long range contact (> 1800 km), using the 14 455 MHz channel. If conditions require that other channels be used for long range contact, the inverted V will be available.

Tuned resonant traps in the inverted V will eliminate the need for separate antennas for each channel. These traps are resonant L-C circuits which essentially "divorce" the wires beyond the traps at the appropriate frequency. The traps are positioned in the antenna so that each operating frequency will be matched properly to the antenna. The overall length of the antenna is 43.2 m, which is approximately 1/2 wavelength at 3.395 MHz.

Both the inverted V and the yagi antennas are supported by a pneumatic telescoping mast as shown in Fig 9. The mast is available as a catalog component from Wil-Bert Company and is self supporting and easy to erect. Mobile television systems are the prime user of the mast for transmitting microwave video information locally. The mast is to be mounted to the van rear door with a separate air compressor mounted near the van suspension. Activating the compressor extends the mast from 2.2 to 14.8 meters. The entire mast is rotatable by hand for yagi antenna alignment.

As can be seen by Fig 8, the yagi favors a high mounting position for long distance communication. As the height of the yagi is increased, the radiation angle decreases, which is desirable to help overcome the barrier of the earth's curvature. The 14.8 m mast plus 1.2 m of trailer suspension holds the yagi approximately 3/4 of a wavelength above ground.

The inverted V antenna is supported on the ends by two tripods to reduce the capacitive effect of the ground near the antenna ends. If the terrain makes this impractical, the antenna ends can be supported by rope tied to sandbags.

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**Table 1 Main transmission characteristics of frequency classes**

Classes	Frequency subdivisions	Main characteristics	Principal uses
Very long waves and long waves	VLF and LF	Wave travels to considerable distances over earth's surface and to great distances by reflection from lower edge of ionosphere	Medium and long distance point-to-point communication Long-wave medium distance broadcasting
Medium waves	MF	Wave travels over earth's surface to relatively short distances during day, at night some energy comes from ionosphere and range increases	Local broadcasting, marine and aircraft communication, direction finding
Short waves	HF	Wave travels up to ionosphere whence it is reflected back to earth. Conditions vary greatly with time of day and season, but great ranges obtained if conditions favorable	Long distance broadcasting, point-to-point communication. Amateur communication, etc
Very short, ultra short, and microwaves	VHF, UHF, microwaves	No ionospheric reflection. Wave travels directly through lower atmosphere from transmitter to receiver	Short distance communication, FM broadcasting, television, radar, aircraft guidance systems. Amateur communication

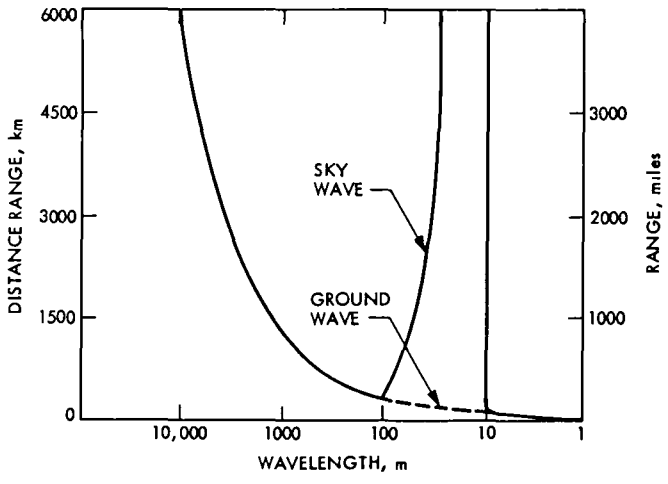


Fig. 1 Variation in range with wavelength

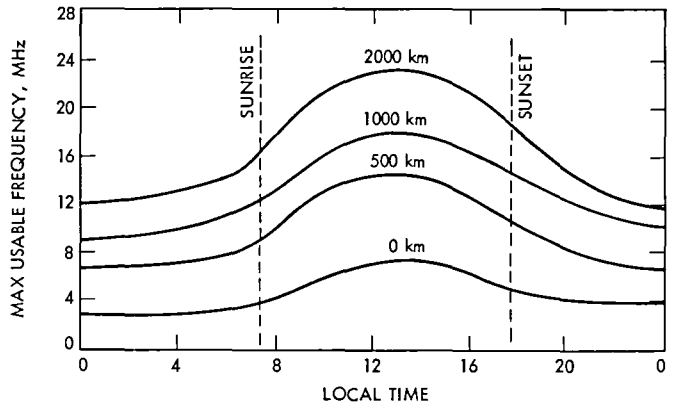


Fig 2 Maximum useable frequency chart

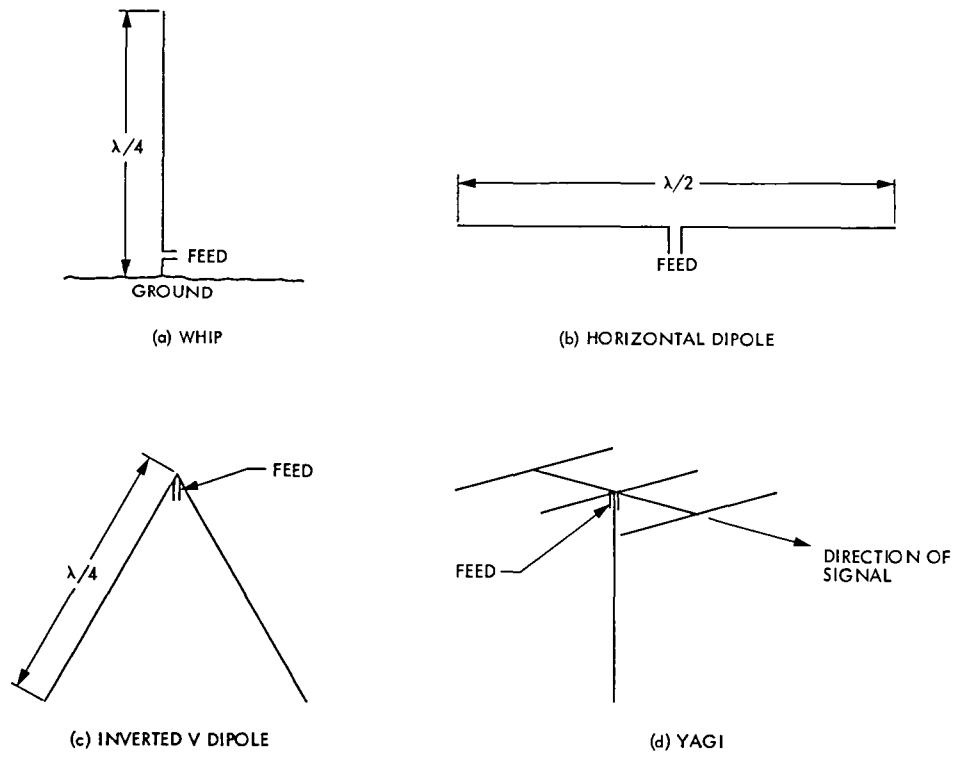


Fig 3. Antenna configurations

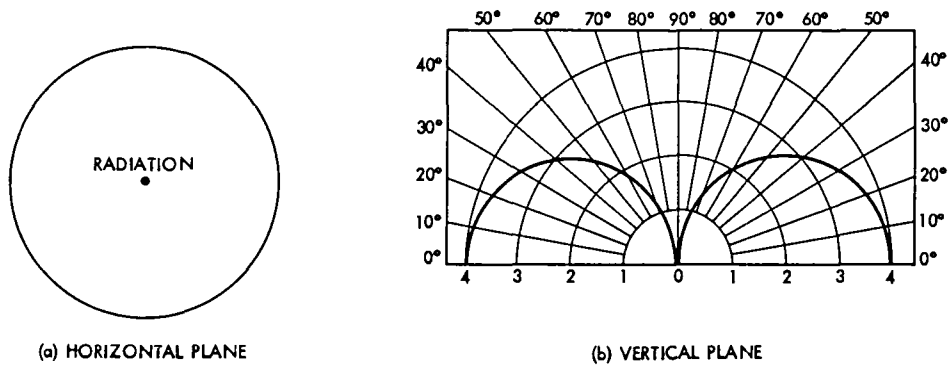


Fig. 4 Whip radiation patterns (assuming ideal ground plane)

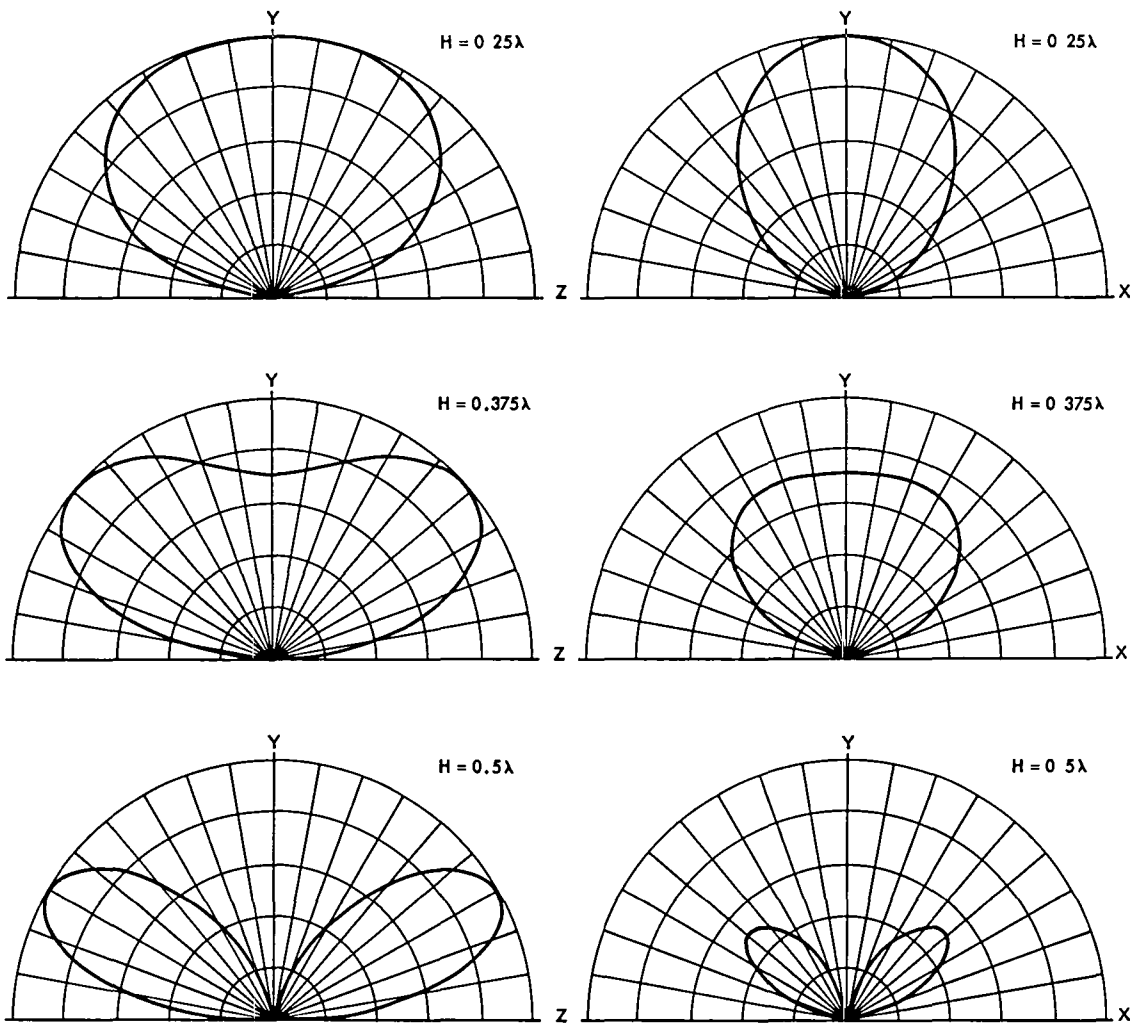
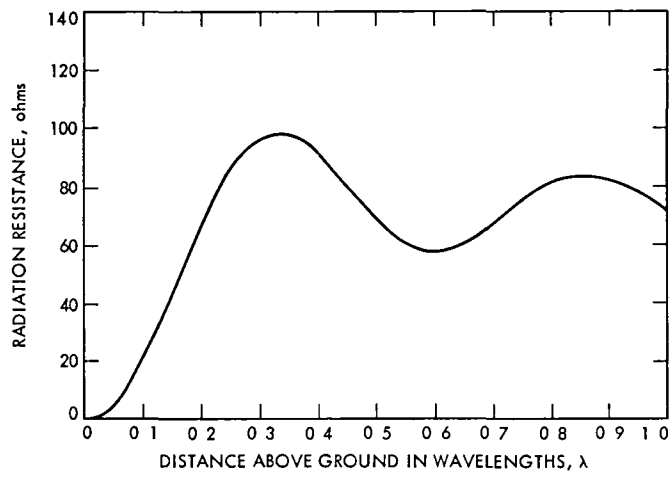
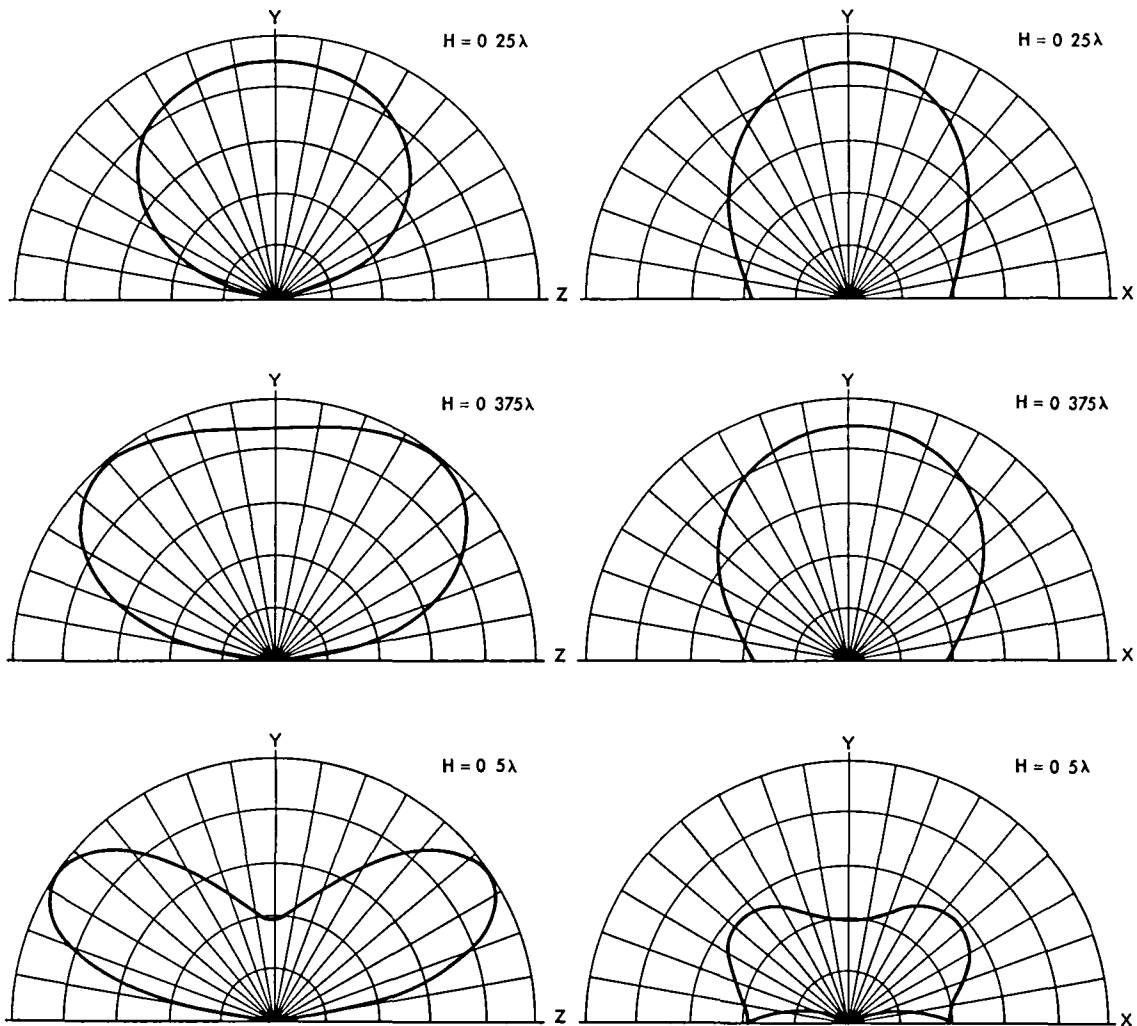


Fig 5 Antenna patterns off the end of (left) and broadside to (right) a horizontal dipole as a function of antenna height above ground (in wavelengths  $\lambda$ )





**Fig. 6. Input resistance of a horizontal half-wavelength dipole antenna at a height  $\lambda$  wavelengths above a ground plane**



**Fig. 7. Antenna patterns off the end of (left) and broadside to (right) an inverted V dipole as a function of antenna height above ground (In wavelengths  $\lambda$ )**

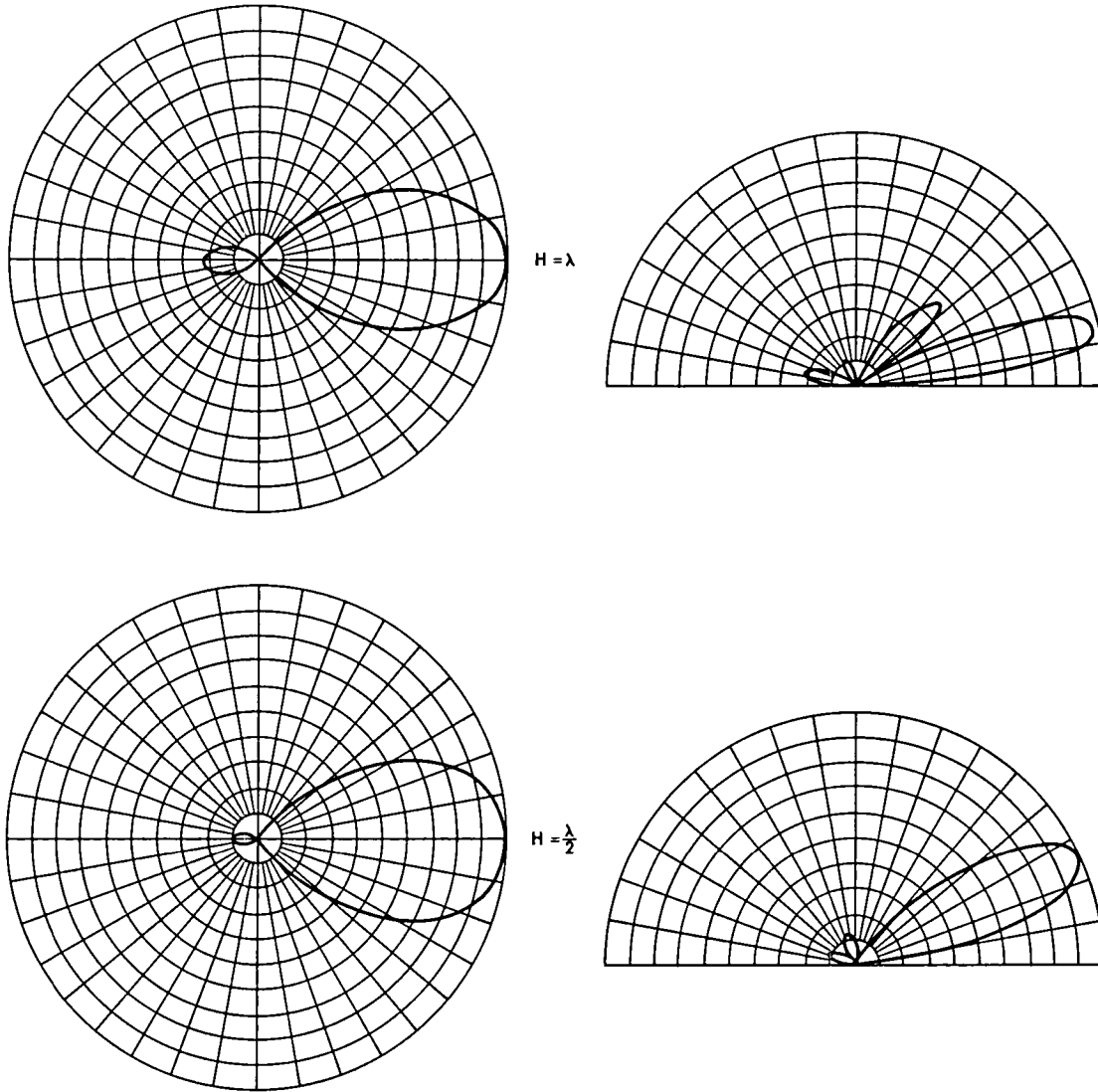


Fig 8 Yagi radiation patterns for horizontal plane (left) and vertical plane (right) as a function of antenna height (in wavelengths  $\lambda$ )

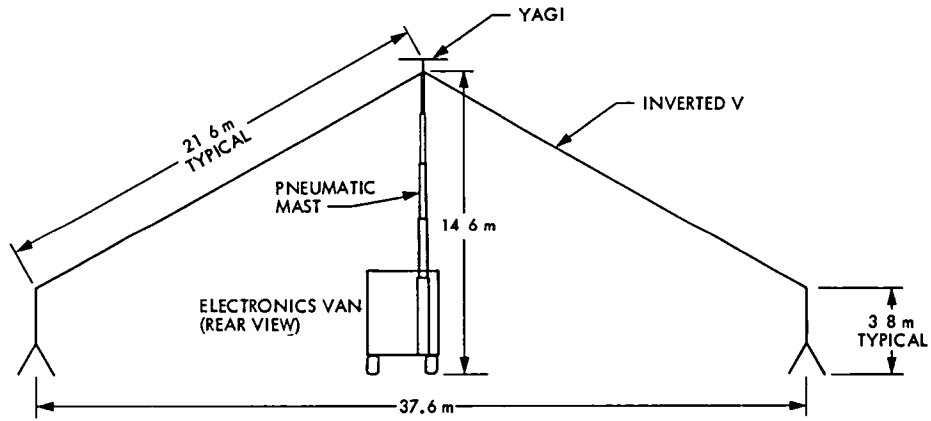


Fig 9. Antenna implementation

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