(NASA-CR-170160) SATELLITE TIME AND FREQUENCY TRANSFER (STIFT) Final Report, 1 Jun. 1980 - 31 Mar. 1983 (Smithsonian Astrophysical Observatory) 22 p HC A02/MF A01 CSCL 22B G3/15
N83-20990
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SATELLITE TIME AND FREQUENCY TRANSFER (STIFT)

GRANT NAG-8006

Final Report
For the period from 1 June 1980 to 31 March 1983

Principal Investigator
Dr. Robert F.C. Vessot

Prepared for
National Aeronautics and Space Administration
George C. Marshall Space Flight Center
Alabama 35812

MARCH 1983

Smithsonian Institution
Astrophysical Observatory
Cambridge, Massachusetts 02138

The Smithsonian Astrophysical Observatory and the Harvard College Observatory are members of the Center for Astrophysics

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SATELLITE TIME AND FREQUENCY TRANSFER (STIFT)

FINAL REPORT

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1. INTRODUCTION

In March, 1980 the Smithsonian Astrophysical Observatory (SAO) submitted a proposal to study the concept of placing a hydrogen maser high stability clock in earth orbit to provide accurate time and frequency comparisons worldwide to major timing centers and to a large number of radio observatory antenna sites involved in VLBI measurements. The proposal was chiefly directed toward studies and initial hardware designs for time comparisons between hydrogen maser frequency standards and to modifications of the hydrogen maser for long-term use in space.

Initial work began in June, 1980 in two general areas:

1. A Study of the modulation requirements and general design parameters for a 3-link time and frequency comparison system to achieve timing accuracy $\Delta t < 1 \times 10^{-9}$ seconds and frequency comparison accuracy $\Delta f/f\ (T) < 1 \times 10^{-14}$ at $T \cong 100$ seconds.

   The chief goal was to arrive at a design concept suitable for a low cost ground station for multiple production to ensure availability of the orbiting clock technique to a reasonably large number of laboratories around the world. Computational and hardware simulation work would be done along with the evaluation of some critical hardware components once they are specified.

2. <u>A study of the 4-link system as an extension of the 3-link system design concept to evaluate its effectiveness in time and frequency transfer for long light times (1000 seconds or longer).</u>

Computational simulations would be done using a system model based on the best available information about the general architecture of deep space tracking systems to be used in the near future. Critical components in the signal flow paths would be identified and hardware tests would be made of their phase and frequency stability to guide the computer simulation of the system performance.

2. HISTORY OF WORK ON THE STIFT PROJECT

During the first 6 months, work was begun on the three-link microwave system and a conceptual design of the space maser microwave cavity assembly was completed. The maser cavity design (Figure 2-1) shows a method of kinematically positioning the resonant cavity within four layers of magnetic shielding and two stages of thermal control, each having three separate zones. The concept is based on our experience with the gravitational probe rocket experiment and further development work with the Naval Research Laboratory. A preliminary thermal analysis was conducted to calculate the expected thermal resistances and heat flow patterns. In the next step of the design these data will be used as input for a computer model to optimize the system and to estimate its thermal performance.

In September 1980, Dr. Rudolph Decher, NASA-MSFC Technical Officer, held a review at SAO of the work being done on the STIFT Program by SAO. Other attendees were Dr. R. Vessot (SAO), Dr. G. Winkler (U.S. Naval Observatory), Dr. D. Allan (National Bureau of Standards) and Professor C.O. Alley (University of Maryland). The committee was satisfied by the good progress on the microwave and maser parts of the system. The current status of the experiment was presented at the Precision Time and Time Interval (PTTI) conference in December 1980.

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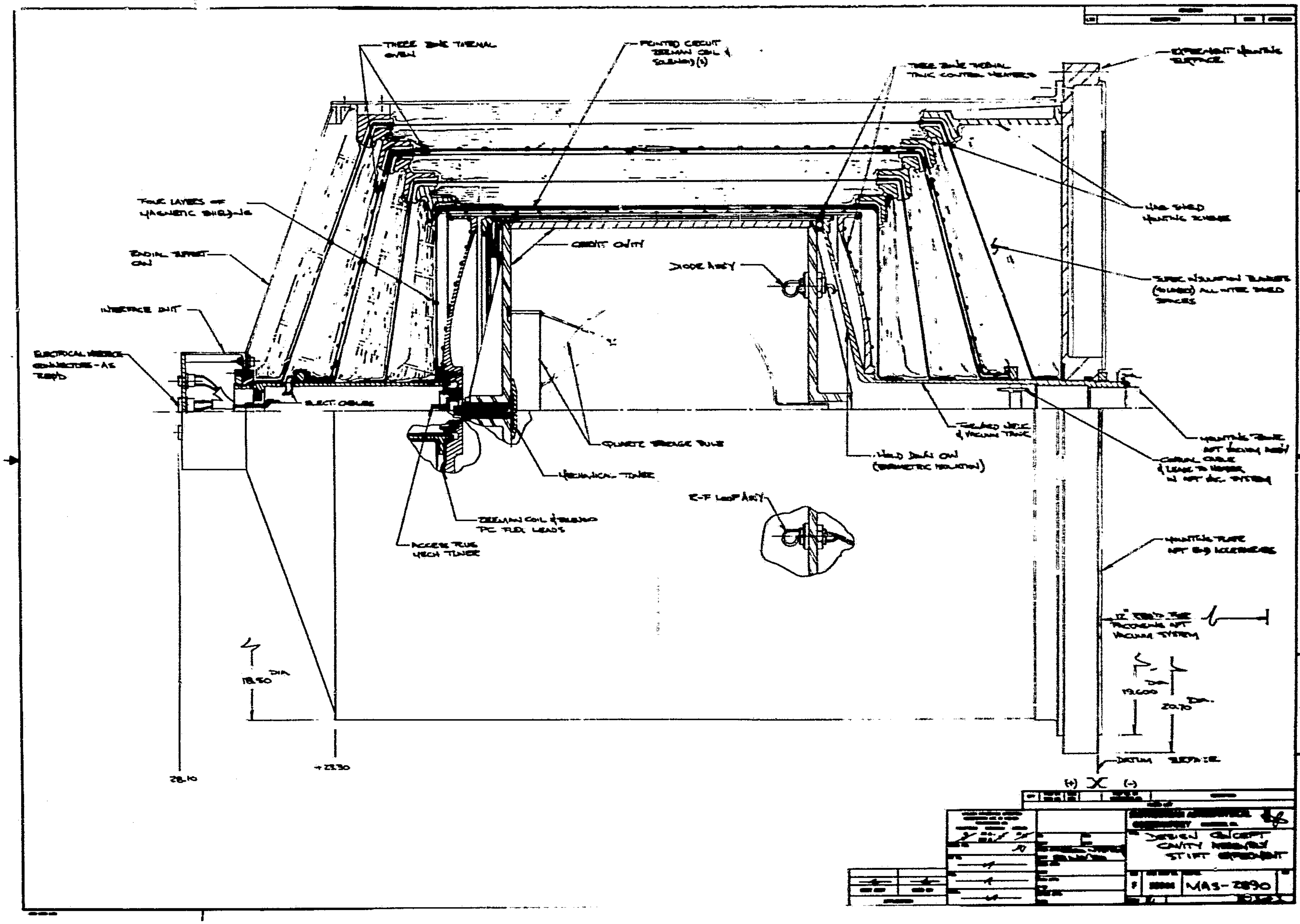

FIGURE 2-1 - STIFT Maser Cavity Assembly

In January 1981, a program plan and budget was jointly generated by MSFC and SAO for presentation to NASA in March 1981 to seek support to build an operational system and perform the STIFT experiment. A major goal of the plan is to provide microwave ground terminals costing no more than $100,000 to make the experiment accessible to the VLBI and national timing laboratories worldwide.

In December 1980, a STIFT Program review was held at MSFC. This review covered the total system concept as well as the probe maser preliminary mechanical, thermal and magnetic designs. The program plan outline and assignments of the MSFC/SAO contribution to the plan were established. In late January 1981, the SAO contributions to the STIFT Program Plan were submitted to MSFC along with a budgetary estimate of cost for the SAO portion of an assumed 4-year program.

In February 1981, a letter proposal (P1062-2-81) was submitted to MSFC to continue the STIFT study at the overall system level and to develop a preliminary sub-system design for the microwave ground terminal (including breadboard construction and evaluation). This proposal was accepted and implemented by MSFC, extending the program at SAO until 15 November 1981.

In early Spring 1981, SAO lost its microwave engineer and began a search for a replacement. At SAO's request, a no-cost extension was granted to extend the program to May

1982. In February 1982, Mr. Hays Penfield, a microwave engineer from Harvard Observatory, continued the work on the STIFT Program concentrating on the design of the microwave ground terminal.

On June 9-11, 1982, a review of the STIFT ground terminal design was made at MSFC. The SAO design was found to be acceptable and MSFC instructed SAO to continue its investigation of the ground terminal design by breadboarding crucial circuits and testing them. The testing program was directed toward establishing the performance of critical parts of the system to assure that the design goals of the ground terminal were met.

Additional funding was received in early September 1982, to continue the work through 31 March 1983. During this time a proposal was prepared and submitted to NASA-MSFC to continue the STIFT work, concentrating on the microwave ground terminal during FY 1983. However, to date, no further support has been received by SAO and work on the project was concluded on 31 March 1983. This report describes the progress of the work at SAO up to this date.

3. CURRENT STATUS OF THE MICROWAVE GROUND TERMINAL DESIGN

The microwave ground terminal design has progressed from the conceptual stage to a more detailed investigation of critical circuit requirements. Development work during calendar 1982 resulted in a hardware oriented system block diagram with unit designations for each subsystem.

(Appendix A describes the microwave ground terminal.)

Particular emphasis during the design study has been placed on the time difference measurement subsystems of the microwave ground terminal. A computer program was written to simulate the shift register circuit for generating the PRN modulation used to "time-tag" the microwave link signals. This program was then used to study various circuit configurations that produce "Maximal Linear" code sequences and to investigate the characteristics of truncated maximal linear codes. The proposed code for the STIFT Ground Terminal is generated by an eight stage shift register which is configured to produce a truncated code length of 250 rather than its natural code length of 255. This code length in combination with a 2 microsecond code element yields a 500 microsecond period for the PRN modulation to provide a convenient "time-tag" for the microwave links.

A technical paper titled "Design of the STIFT Time and Frequency Transfer microwave ground terminal" was presented at the PTTI Conference in early December 1982, at Goddard Space Flight Center. This paper describes the design approach for achieving simultaneous time difference measurements via microwave links between a ground terminal and a terminal in the orbiting shuttle spacecraft.

Breadboarding and testing of circuits critical to the operation of the time difference measurement portion of the microwave ground terminal was initiated in July 1982.

At present the five circuit boards, which constitute the coded pseudo-random noise (PRN) phase modulator and the time-discriminator for the receiving systems, have been completed. These systems (two of which are needed in the ground terminal) each perform the major part of the locking-on and tracking of the PRN signal. When these circuit boards are interconnected they will form a complete closed-loop time discriminator providing a digital output that is a measure of the time difference between the input PRN and modulation and a local PRN reference. Figures 3-1, 3-2 and 3-3 show these circuit boards in breadboard test setup in the SAO Electronics Laboratory. In Figure 3-1, the oscilloscope is monitoring the ramp output from one of the DC integrators for the condition of the reference PRN code being aligned with the modulation code on the IF signal. (See Figure 3, page 8 of Appendix A.) In Figure 3-2, the spectrum analyzer is monitoring the 85 MHz IF signal with the 90 degree PRN phase modulation. The vertical displacement on the analyzer display is logarithmic and clearly shows the sine x/x distribution of sidebands centered about the carrier frequency. Figure 3-3 is a close-up view of the five breadboard circuits with the order of boards from left to right being No. 1, No. 2, No. 4, No. 5 and No. 3. A brief summary of the function of each of the

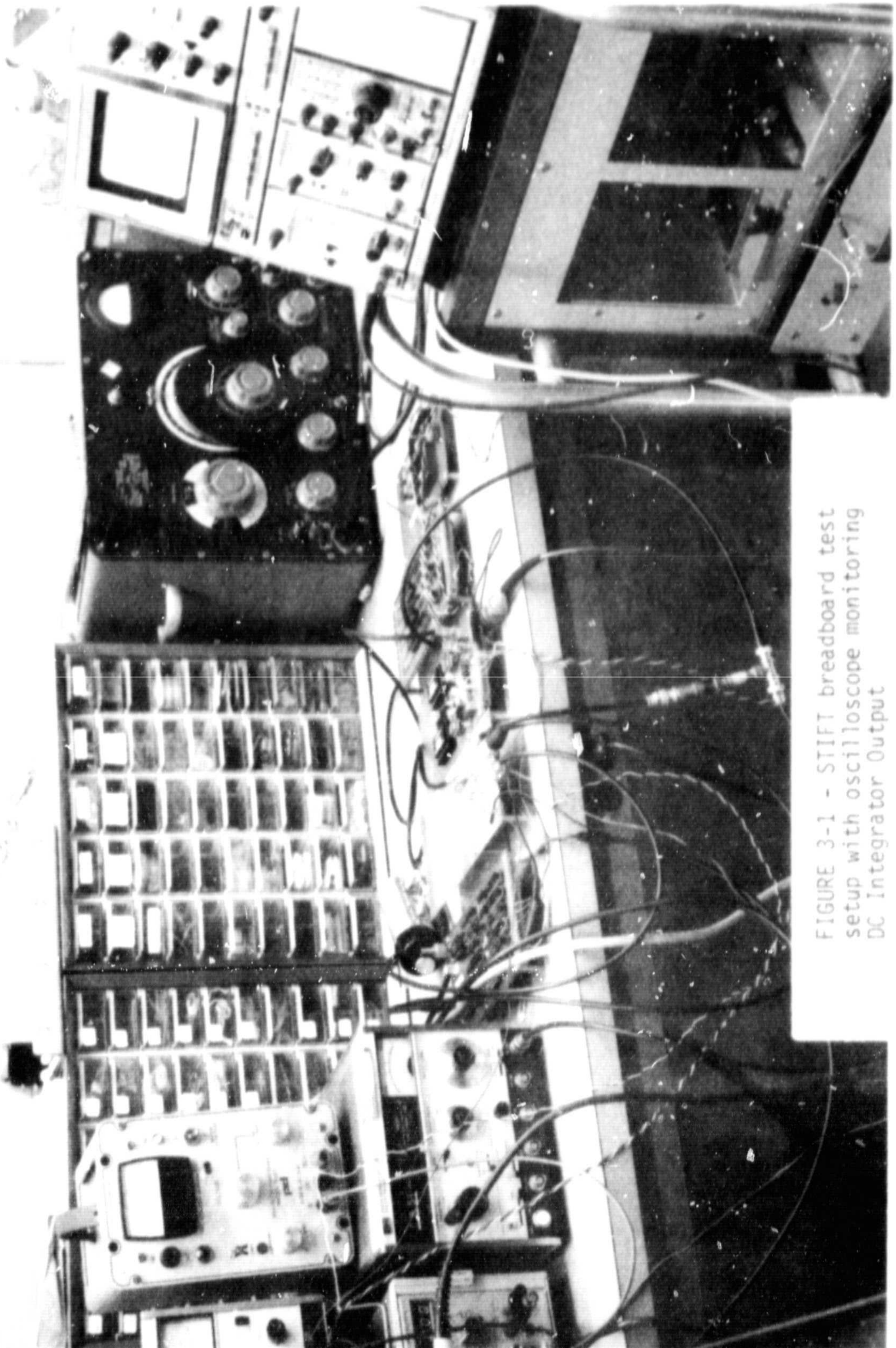

FIGURE 3-1 - STIFT breadboard test setup with oscilloscope monitoring DC Integrator Output

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FIGURE 3-2 - STIFT breadboard test setup with spectrum analyzer monitoring PRN phase modulated 85 MHz signal.

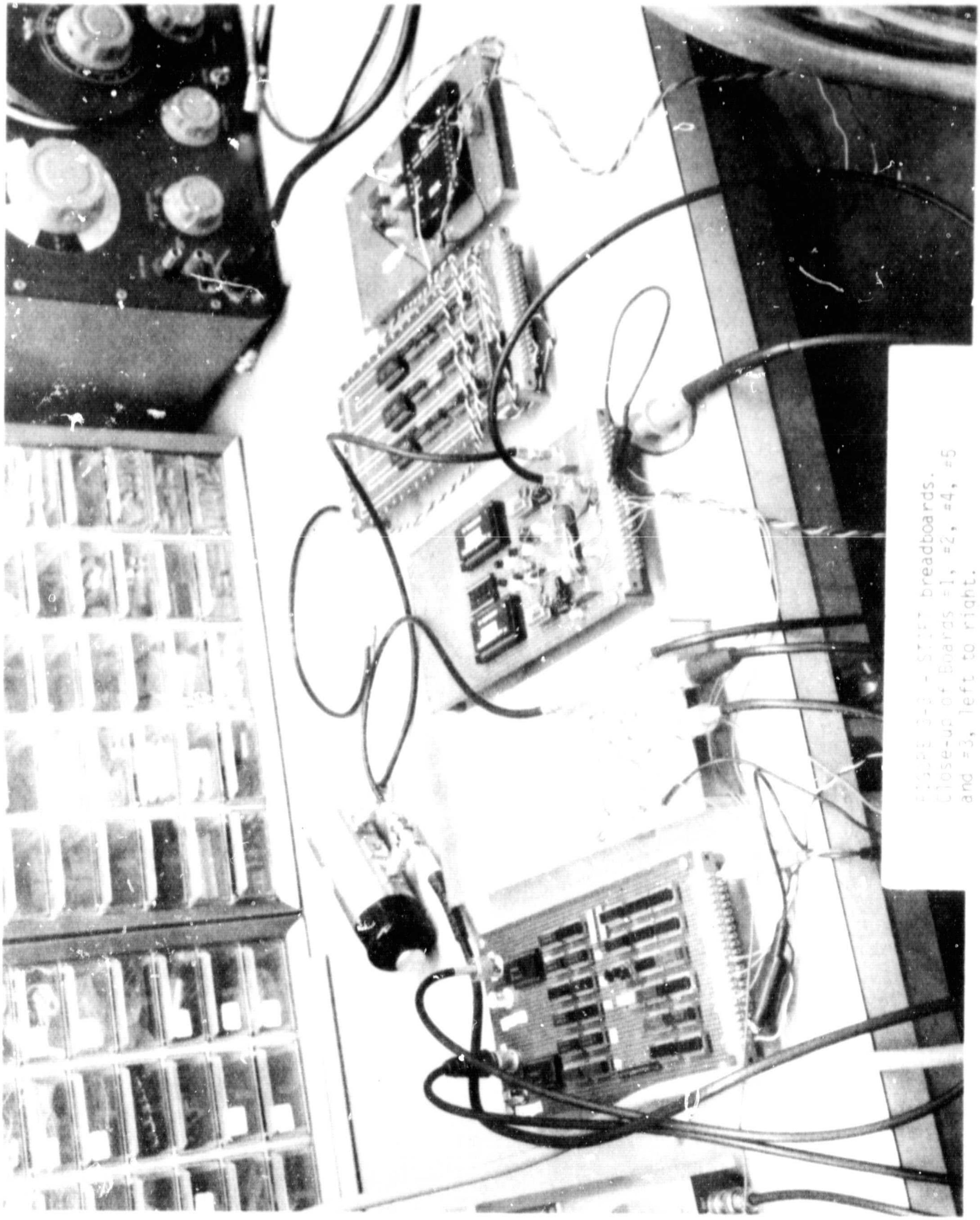

FIGURE 3-3 - STIFF breadboards.
Close-up of Boards #1, #2, #4, #5
and #3, left to right.

five breadboard circuits and their status is given below:

<u>BOARD</u> <u>No. 1</u> - PRN generators "A" and "B" and timing circuit to generate "A" and "B" clock pulses. Also includes control pulses for DC Integrator, Sample/Hold, Analog-to-Digital converter and Sense Reversal.

This board has been tested to check that the PRN code generators are producing the proper code and it has been verified that the code agrees with the code produced by the computer program. In addition, it has been verified that the logic for truncating the code is correct. The control pulses have also been checked to determine that they are of proper polarity and occur in the desired time relation.

<u>BOARD</u> <u>No. 2</u> - Dual Channel Signal Mixer (PRN Demodulator) and IF Amplifier

This board heterodynes an IF signal at 85 MHz with the PRN video modulation signal. If the IF input is an 85 MHz signal with no modulation, the output is a PRN phase demodulated signal with 90 degree phase excursion. If the IF input is a PRN phase modulated signal with 90 degree phase excursion, the output, when the PRN codes are matched in time, is a demodulated IF signal. The performance of this board has been checked by using one channel to modulate the cw 85 MHz carrier and the other channel to demodulate the output from the first channel. A spectrum analyzer has been used to observe the signal at each point in this process.

<u>BOARD No. 3</u> - Sin/Cos Digital-to-Analog Converter and 16 MHz Phase Shifter

This board accepts a 10 bit natural binary input and produces two analog outputs. One output is proportional to the sine of the digital input and the other is proportional to the cosine. These sine and cosine signals are combined in a single sideband modulation circuit with a 16 MHz input to produce a 16 MHz output whose phase (relative to the input) is controlled by the 10 bit digital input. The sin/cos Digital-to-Analog portion of this breadboard has been tested and is operating properly. The single sideband mixer circuit has been tested; however, a large unbalance in the 16 MHz quadrature circuit requires further investigation.

<u>BOARD No. 4 - Dual Channel Carrier Mixer, DC Integrator, Sample/Hold and Analog-to-Digital Converter</u>

This board heterodynes the 85 MHz demodulated output from the dual channel signal mixer and IF Amplifier Board (Board No. 3) with the 85 MHz carrier signal. The dc component from each of the carrier mixers is integrated for one period of the PRN code sequence and converted to a 12 bit digital signal. This board is being tested at the present time.

BOARD No. 5 - Combiner, Sum/Difference Logic and Digital Loop Filter

This board accepts the 12 bit digital outputs from the dual channel A/D converters on Board No. 4 and computes the sum and difference. Also planned for this board is the Digital Loop Filter. At the date of this report, the digital loop filter design for this breadboard is not finalized and only the Combiner and Sum/Difference Logic has been wired on the board. No testing has been done on this board.

In addition to the five breadboards, a number of other components have been procured for use in the STIFT Design Study Program. These include the following:

1. Low-Noise Amplifier, 2.15-2.35 GHz
2. S-Band Circulator
3. Two (2) S-Band Double Balanced Mixers
4. 85 MHz VCXO
5. 182 MHz VCXO

These components are to be used in future tests involving the modified USB Transponder and Translator on loan from Marshall Space Flight Center.

4. CONCLUSION

The overall system level study of STIFT and the subsequent development of a preliminary sub-system design for the microwave ground terminal has progressed according to plan. Functional units within the ground terminal have been identified and specified. Time coding, decoding and time difference measurement have been studied analytically with experimental circuit breadboards constructed to test several areas of the time difference measurement portion of the Microwave Ground Terminal.

A total of five circuit breadboards was constructed and testing of individual boards has verified the performance of the PRN code generator circuit, the overall timing synchronization of the time discriminator and the demodulation, integration and digitizing functions in the time discriminator circuit.

Our plans for future work include completion of the time discriminator testing and a demonstration of closed-loop tracking of the time difference measurement portion of the ground terminal.

APPENDIX A

DESIGN OF THE STIFT TIME AND FREQUENCY TRANSFER MICROWAVE GROUND TERMINAL

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ABSTRACT

The Satellite Time and Frequency Transfer System (STIFT) is intended to provide, simultaneously, global time comparisons at the subnanosecond level and frequency comparisons to better than 1 part in 10^{14}. It utilizes an orbiting hydrogen maser clock and frequency standard that communicates, via microwave links, time and frequency information to earth terminals operated by hydrogen masers controlling local clocks. A two-way microwave link, to and from the space vehicle provides Doppler information used to cancel the Doppler shifts in a one-way link from the spaceborne oscillator. Pseudo-random noise (PRN) modulation in the two-way link also provides range information to cancel the range delay in the PRN time transfer between space and earth. The pseudo-random noise modulation system for time difference measurement and its incorporation in the Doppler cancellation system for frequency comparison is explained. The particular PRN code sequence selected and an analysis of the system is discussed.

BACKGROUND

The ongoing development of atomic frequency standards presently provides us with stability better than 1 part in 10^{15} over hourly averaging intervals and ever increasing accuracy. This now poses a serious challenge to current techniques that transfer time and frequency on a global scale so that the precision of transfer is commensurate with the performance of these clocks. Today, the most commonly used time transfer technique is the transportable clock. However, this method has significant disadvantages. It is prohibitively costly, if carried out on a continuous basis, and the process is generally limited to an uncertainty of the order of 100 nanoseconds owing to environmental conditions during transport. The Global Positioning System (GPS) offers an alternative with an accuracy level of the order of 10 nanoseconds, but it does not have the capability of transferring frequency. However, the concept of a transportable clock, moving from one site to another as a means for coordinating time and frequency on a global scale can be logically extended to a clock orbiting the earth in a satellite. The hydrogen maser, with its frequency stability of the order of 6 parts in 10^{16}, for 1 hour averaging interval, is ideal for such an orbiting time transfer clock and oscillator to provide time and frequency difference measurements of 1 nanosecond and 1 part in 10^{14} respectively with stations anywhere in sight of the orbiting system.

Relation to GP-A

Experience already exists in the practical application of hydrogen maser technology to the space environment. In June 1976, SAO participated with NASA in the Gravitational Probe A (GP-A) experiment. This experiment was designed to probe the Earth's gravity field in a nearly vertical trajectory to an altitude of10,000 KM. The hydrogen maser used in GP-A required highly specialized design to cope with the traumatic changes in thermal, magnetic and gravitational environment without allowing time for thermal stabilization or magnetic readjustment. Even though the maser operating life in space on GP-A was limited to about two hours, the maser was designed for continuous operation throughout many months of testing. The experience gained in designing and operating the space maser on GP-A is directly applicable to the orbiting transfer clock. In addition to the space maser technology, GP-A also provided important experience in the use of microwave links to compare space and ground clocks. The feasibility of cancelling propagation effects in the troposphere and ionosphere and of removing Doppler shifts was demonstrated successfully by the GP-A experiment.

Choice of Frequencies

The microwave frequencies used in the GP-A experiment were chosen to be compatible with the Unified S Band (USB) System and this same choice is to be carried through the Satellite Time and Frequency Transfer Experiment. The STIFT system requires both a 2-way up/down link and a 1-way down link between the orbiting clock and the ground clock. The frequencies in this system are selected to cancel the first order ionospheric dispersion. Thus, the 2-way link utilizes 2,117 MHz for the up transmission to the orbiting clock and 2,299 MHz for the transponded down transmission to the ground clock. The 1-way link transmits down from the orbiting clock on a frequency of 2,203 MHz. With this selection of frequencies, the combined ionospheric dispersion for the 2-way link is just twice that of the 1-way link and may be cancelled in subsequent frequency difference processing in the STIFT ground terminal.

SYSTEM DESCRIPTION

The STIFT system is designed to provide simultaneous precision measurement of time difference and frequency difference between a ground clock and the orbiting space clock. Figure 1 illustrates the overall STIFT system design including the Space Terminal, Microwave Ground Terminal and Laser Ground Terminal. The discussion that follows deals primarily with the Microwave Ground Terminal and the other sections of the system are shown in Figure 1 only to give some perspective to the role of the Microwave Ground Terminal.

2-way Link

The 2-way microwave link between the space terminal and the ground terminal serves three major functions. First, it measures the path delay between the two terminals; second, it provides the reference for first order Doppler cancellation and third, it compensates for ionospheric dispersion through selection of link operating frequencies relative to the 1-way link frequency.

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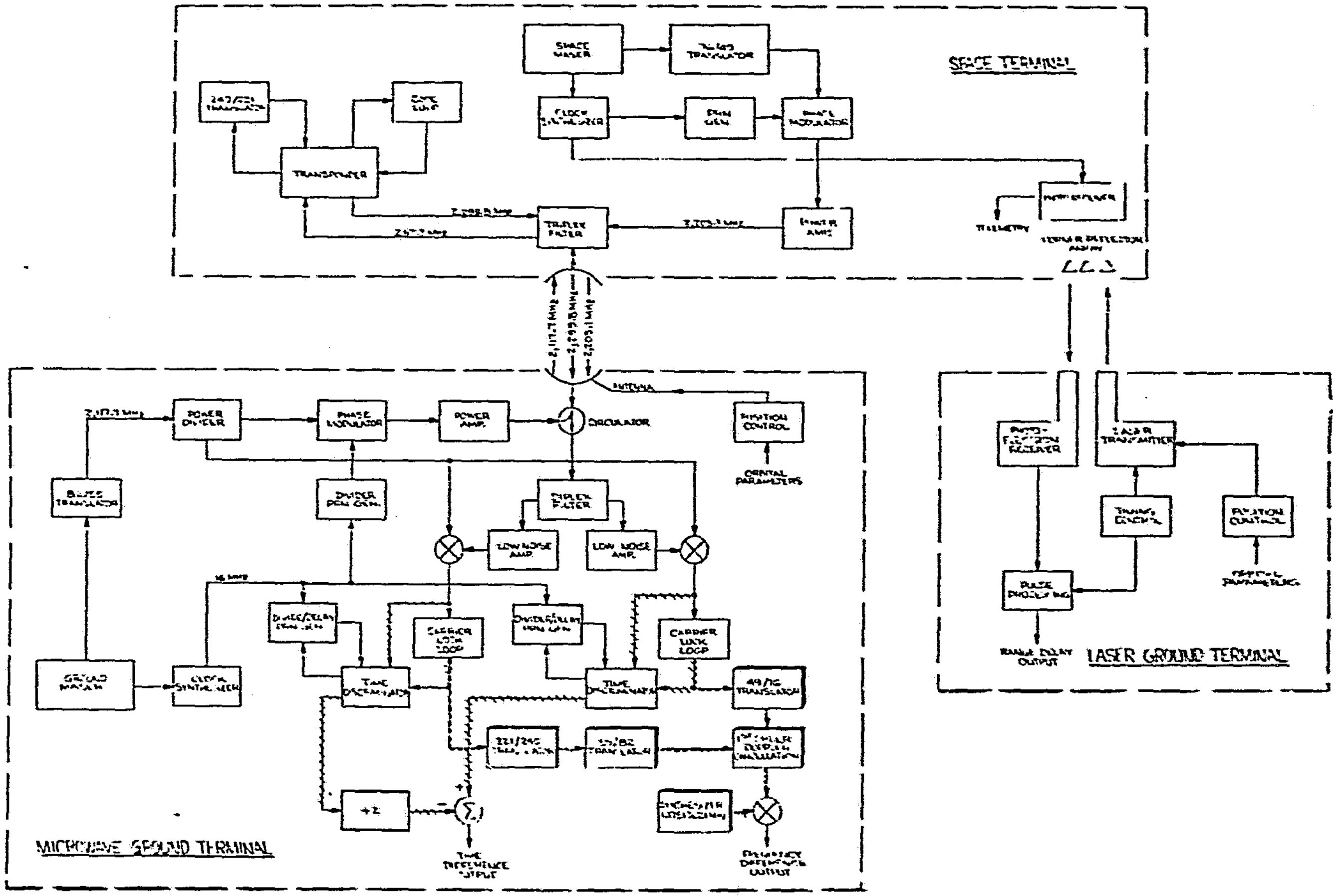

Figure 1. Block Diagram of STIFT (Satellite Time and Frequency Transfer) System

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The up-link transmit frequency is derived from the atomic hydrogen maser oscillator by a coherent frequency translation process. In Figure 1, the atomic hydrogen maser oscillator is shown as the ground clock and the 2,117 MHz up-link transmit carrier frequency is obtained by an 82/55 translation of the L-Band maser output. A part of the 2,117 MHz carrier is utilized as coherent local oscillator drive in the first heterodyne mixers of both the 2-way and 1-way receivers in the ground terminal.

Prior to being transmitted, the 2,117 MHz signal passes through a phase modulator which impresses a 90 degree phase shift on the carrier under the control of a pseudo-random noise (PRN) generator. This PRN phase modulated signal is then amplified to approximately ten watts and coupled through a ferrite circulator to the ground terminal antenna. This single antenna is a common element for both the 2-way and 1-way links and consists of a small (1 meter) steerable parabolic dish with a gain of about 25db and a half-power beamwidth of 9.5 degrees at the S-band operating frequencies.

The 2,117 MHz ground terminal transmit signal is received by a broad beam, circularly polarized antenna at the space terminal and is coupled through a triplex filter to the input of a phase coherent transponder. The transponder strips the phase modulation from the received signal, coherently translates the 2,117 MHz carrier by the ratio 240/221 and reapplies the phase modulation to form a 2,299 MHz transponder output signal. This 2,299 MHz signal is coupled through the triplex filter to the space terminal antenna and transmitted toward the ground terminal.

The received 2,299 MHz signal, at the ground terminal, is picked up by the parabolic antenna and coupled through the ferrite circulator to a diplex filter that separates the 2-way and 1-way received signals and provides high rejection to the 2,117 MHz transmit signal. The low-noise amplifier, at the output of the diplex filter, feeds the 2,299 MHz signal to a mixer where it is heterodyned with the 2,117 MHz local oscillator signal to form a 182 MHz IF signal. The carrier component of the IF signal is extracted by a carrier phase-lock loop and is used in subsequent processing to cancel first order Doppler in the frequency difference determination.

The 182 MHz IF signal (full band) and extracted IF carrier component also are fundamental input signals for the 2-way Time Discriminator. A digitally delayed PRN code generator (with the identical code sequence used to modulate the 2,117 MHZ transmit signal) is coupled in a closed-loop tracking configuration with the Time Discriminator to automatically lock-on and track the path delay in the 2-way link. This 2-way path delay, in digital form, is divided by two and used in conjunction with the 1-way receiver output to determine time difference between the space clock and ground clock.

1-way Link

The 1-way microwave link between the space terminal and the ground terminal operates on a carrier frequency of 2,203 MHz. This signal is derived from the L-Band hydrogen maser coupled to a 76/49 frequency translator. A PRN

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code generator, with the identical code sequence used by the PRN generators in the ground terminal, controls the phase modulation (90 degrees) that is impressed on the 2,203 MHz carrier prior to transmission from the space terminal. The triplex filter couples the 1-way link transmission to the common space terminal antenna used for both the 2-way and 1-way links.

The received 2,203 MHz signal, at the ground terminal, is coupled from the parabolic antenna through the ferrite circulator to the diplex filter in the same fashion as the 2,299 MHz 2-way link signal. The 2,203 MHz signal is split off separately by the diplex filter and fed through a low-noise amplifier to a mixer where it is heterodyned with the 2,117 MHz local oscillator to form an 85 MHz IF signal. The carrier component of this IF signal is extracted by a carrier phase lock loop and after frequency translation, is compared with the processed 2-way IF carrier to establish the frequency difference between the space clock and the ground clock.

The 85 MHz IF signal (full band) and extracted IF carrier component are coupled to the 1-way Time Discriminator. A PRN code sequence (identical to the other PRN code sequences used in the ground and space terminals) is digitally delayed and automatically locks-on and tracks the 1-way path delay plus the time difference between the ground clock and the space clock. Path delay is cancelled by subtracting half of the 2-way path delay leaving a direct, real time output of the apparent time difference between the two clocks. True clock difference is obtained from post-real time data reduction in which relativistic and gravitational effects are removed.

TIME DIFFERENCE MEASUREMENT

The STIFT microwave ground terminal is designed to provide time difference measurements between a space clock and a ground clodk to an accuracy of 1 nanosecond. This time difference measurement is implemented through the use of a periodic pseudo-random noise (PRN) code that is phase modulated on the 2-way and 1-way microwave links. Simultaneous measurement of time delay in both the up/down (2-way)and down (1-way) receivers in the ground terminal allows cancellation of propagation delay between the space terminal and ground terminal.

Figure 2 is a detailed block diagram of the time delay tracking loops for both the 2-way and 1-way receivers.

PRN Time Discriminator Tracking Loop

The operation of the delay tracking loops is based on a time discriminator circuit that senses time coincidence between the PRN code modulation on the received signal and a PRN code that is precisely shifted relative to the received code.

When two similar periodic time functions, such as the two PRN code sequences, are shifted in time relative to each other and then multiplied together and averaged over one period, the result has the form of the autocorrelation

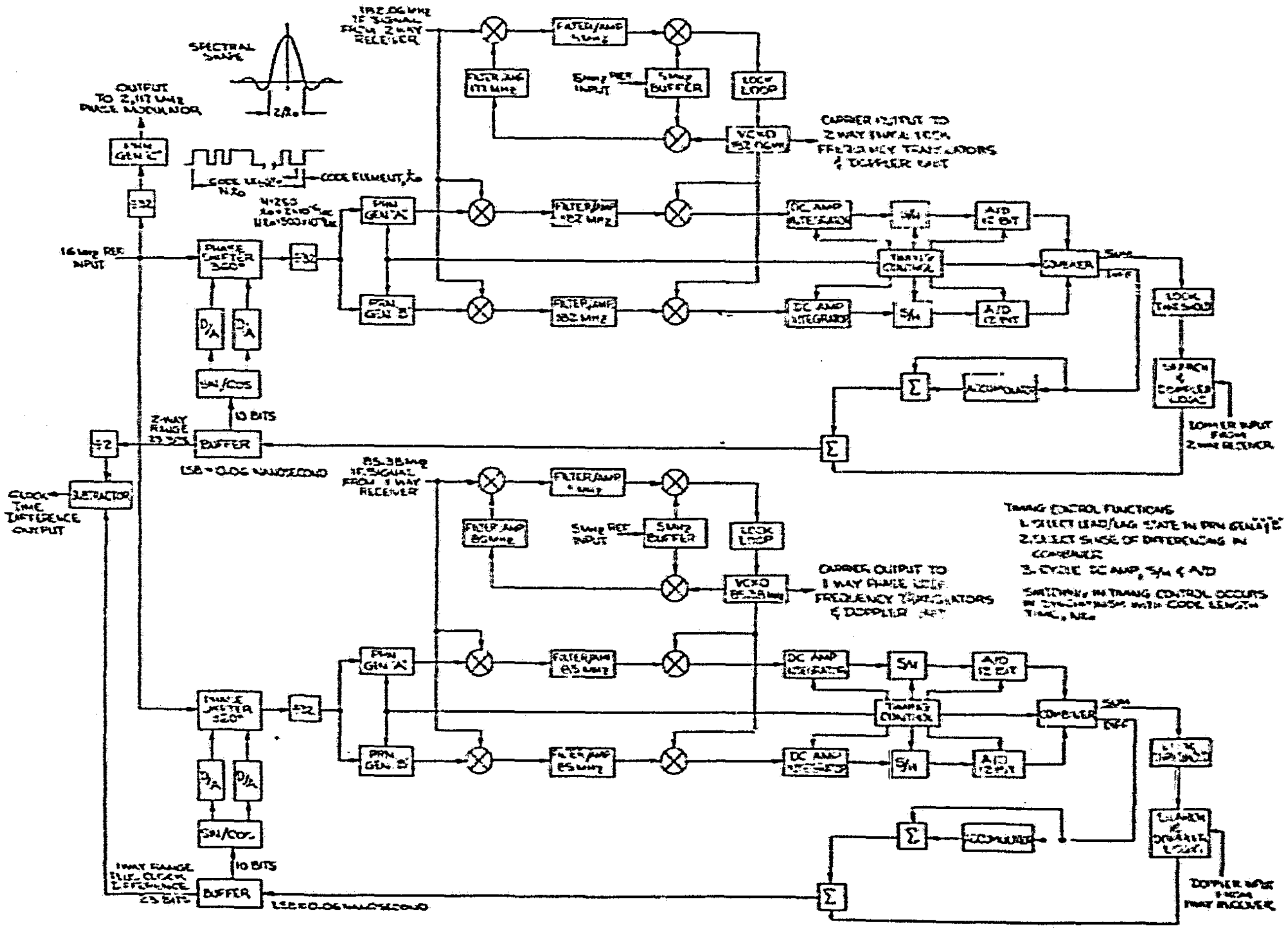

Figure 2. STIFT Microwave Ground Terminal Time Delay Tracking Loops

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function of the original time function. Periodic autocorrelation functions are characterized by a maximum value that occurs when the relative time shift is zero or some integer multiple of the time function period. As the relative time shift increases (or decreases), the autocorrelation function drops to a lower value. The rate at which the autocorrelation function decreases either side of maximum is proportional to the spectral width (bandwidth) of the time function. The magnitude of the "hash" level between maximum points is determined by several factors including the product of the spectral width and time function period (time-bandwidth product) and the detailed shape of the time function. The PRN code sequence represents a time function that can be optimized to have an "ideal" autocorrelation function in which the "hash" level is uniform with a magnitude, relative to the maximum, that is equal to the reciprocal of one half the time-bandwidth product.

The PRN code is a binary sequence composed of "N" code elements in each period. Individual code elements have a duration of "t_0" and may be either "1"(+) or "0"(-) as determined by the code sequence. Figure 2 gives a pictorial representation of the autocorrelation for a 31 element PRN code. Two different cases are illustrated; the first with an offset or delay (i.e., non-aligned codes) and the second without any offset (i.e., aligned codes). In each case, the output obtained by integrating the product of the codes over a complete period (Nt_0) is shown below the codes. Note that when there is no delay between codes, the integrated output builds up linearly over the period whereas in the case of an offset, the integrated output fluctuates back and forth about zero during the period.

The time discriminator utilizes two correlation circuits that are driven by separate PRN code generators operating with a fixed offset in time, relative to each other, equal to one code element, t_0. Figure 4 is an expanded picture of the correlation process as a received code moves in delay relative to the two PRN codes, A and B. The code sequence in this figure is the same as that used in Figure 3. The A and B codes are displaced relative to each other by t_0 and are symetrically displaced about the nominal zero delay point by $t_0/2$. Thus, the maxima in the A and B correlation outputs are displaced symetrically either side of zero delay. When the A and B outputs are added together, the result is a flat topped signal with a half amplitude duration of 2 t_0. When the B output is subtracted from the A output, the result is a time discriminator signal with a linear slope of $2/t_0$ times the amplitude of the maxima and passing through zero at the zero delay point. This time discriminator signal is utilized as the time error signal in the closed loop time tracking sections of both the 2-way and 1-way receivers of the STIFT microwave ground terminal.

Figure 5 shows a block diagram of the time discriminator and delay tracking loop. The circuit requires three inputs:

1. Signal(full band receiver IF),
2. Carrier (receiver IF carrier), and,
3. Clock drive for PRN generators,

and provides a digital output representing the delay between the received signal modulation code and the nominal zero delay point of the A and B PRN generators.

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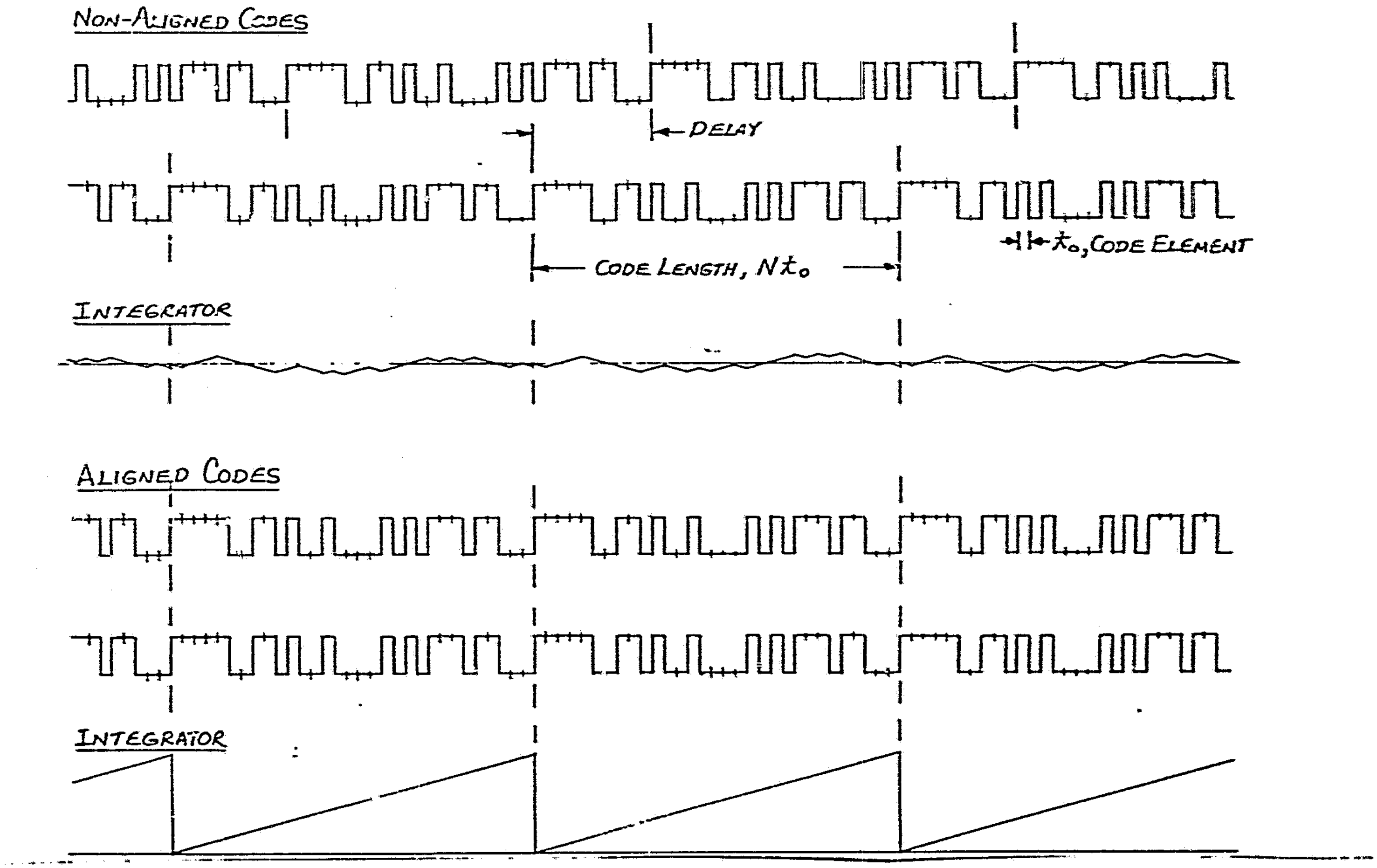

Figure 3. Non-aligned and Aligned Correlation of 31 Element Code

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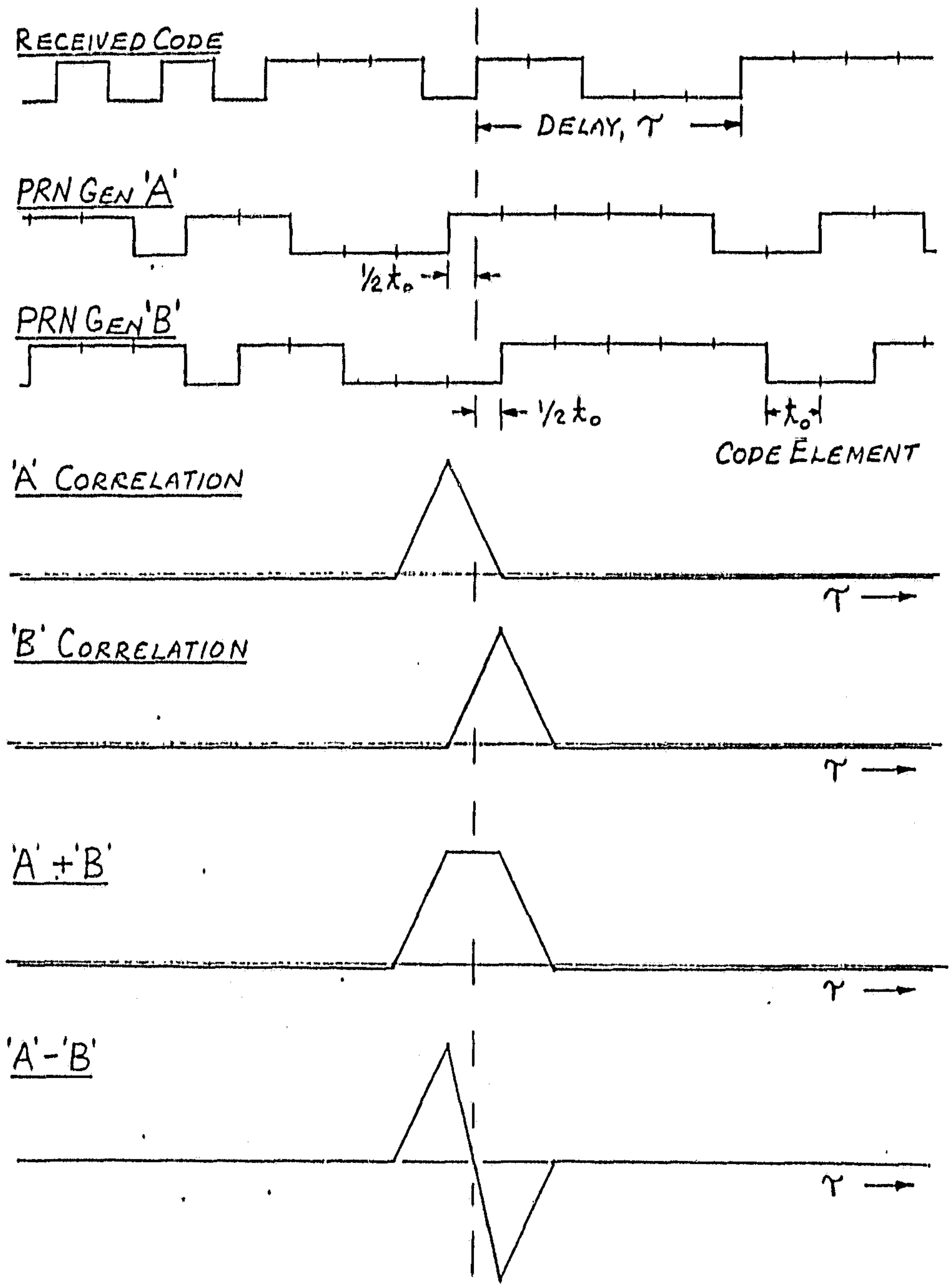

Figure 4. Time Discriminator Autocorrelation and Sum and Difference Outputs for 31 Element Code

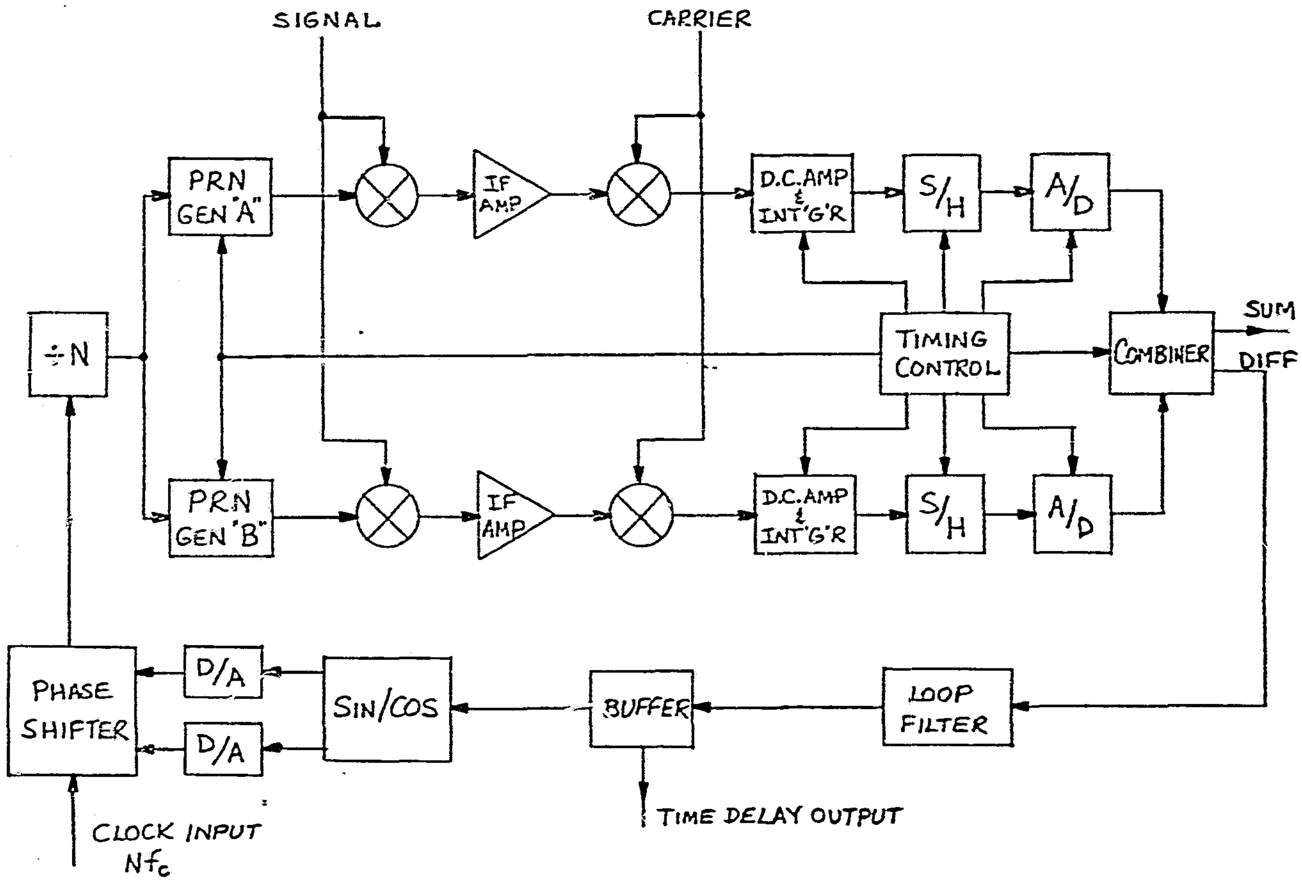

Figure 5. Block Diagram of Time Discriminator and Delay Tracking Loop

Precise control of the delay of the A and B PRN generators is obtained by passing the clock input through a phase shifter followed by a divider. The phase shifter is essentially a single sideband modulator that is driven by the digital error output from the time discriminator. The lower ten bits of the error output are converted to sine and cosine terms and these are then converted to quadrature analog signals by digital-to-analog (D/A) converters that drive the single sideband modulator. The full range of ten bits (1024) represents one complete cycle of phase shift at the clock input frequency. The phase shifted clock frequency drives a digital divider circuit to reduce the clock frequency to the $1/t_0$ rate required to drive the A and B PRN generators.

In the STIFT system, the value of t_0 is 2 microseconds (500 KHz clock rate) and the digital divider factor is 32. Hence, the clock input frequency to the phase shifter is 16 MHz. The time resolution of the time discriminator in terms of the time delay represented by 1 bit in the error signal is $1/(1024 \times 16 \times 10^6)$ or 0.06 nanoseconds. Relating this to the slope of the time discriminator response yields a total of 32,768 bits (32 complete cycles of the phase shifter) over the linear slope region between the maximum and minimum points. The period of the PRN code is 500 microseconds (250 code elements) thus, the total number of bits for one period is 8,192,000 (8,000 complete cycles of the phase shifter). In order to accommodate this full range of delays, the digital output from the tracking loop will be a 23 bit binary number. Since the period in terms of bits is less than the maximum numerical value of the 23 bit binary number (8,388,607) the loop output will be reset to zero when it tries to exceed 8,191,999.

The operation of the time discriminator and delay tracking loop illustrated in Figure 5 starts with the injection of the full band receiver IF signal into balanced demodulators (mixers) in both the A and B channels of the time discriminator. The A and B PRN generators provide the other input to these demodulators. The output from each demodulator (still at the IF level) is amplified and coupled to a second mixer where it is heterodyned with the IF carrier. The output from each of the second mixers is a fluctuating dc level representing the correlation voltage. This dc level is then amplified and integrated over the duration of one period of the PRN code. At the end of the integration cycle, the integrator output is sampled and held and converted to a 12 bit binary number. The separate digital outputs from the A and B channels are combined to form both sum and difference output numbers. The cycle timing for the integrators, sample/hold circuits and Analog-to-Digital converters is generated by a timing control circuit. This circuit also reverses the relative delay "sense" of A and B PRN generators and combiner differencing polarity at the end of each period in order to cancel drifts and voltage or current offsets in the A and B channels of the time discriminator. Thus, two complete periods of the PRN generator (1 millisecond) is the time required to obtain a valid error signal (difference output at the combiner). The digital error signal is coupled to a loop filter (accumulator) and then through a buffer to the sine/cosine unit to complete the tracking feedback loop.

The time delay output in the form of a 23 bit binary number is taken from the buffer. The update rate on this output is 1KHz.

PRN Code Sequence

There is no unique PRN code sequence that is optimum for the time difference measurement in STIFT. Several guidelines that were followed in establishing the parameters for the code are listed below:

1. The code length, or period, should be sufficient to allow easy resolution of any ambiguity.
2. The code length should be such that it can be conveniently used to time-tag the clocks.
3. The code element duration should be selected to give reasonable time discrimination characteristics and also to be compatible with the USB transponder bandwidth of approximately 800 KHz.
4. The code element duration should be compatible with the code length in the sense that the code length must be an exact integer multiple of the code element duration.
5. The code element duration should allow a clock rate that is easily derived from standard frequencies.

After consideration of the various tradeoffs involved in satisfying these guidelines, the choice of 500 microsecond code length with 250 code elements of 2 microseconds duration was made.

The STIFT system requires a total of six PRN generators, all of which may* provide identical code sequences. A convenient means of generating these codes is to utilize a digital shift register with feedback from two or more stages through Exclusive Or (XOR) logic gates to the input stage. With proper feedback connections, these shift register code generators provide a code that repeats every 2^n-1 shifts (elements) where "n" is the number of stages in the shift register. This special class of codes, known as "Maximal Linear Codes", have the characteristic of containing all possible sequences of length "n" except for the sequence of all zeros. These Maximal Linear Codes also exhibit a very special autocorrelation function in which the hash level between maxima remains "flat" with a magnitude of $1/(2^n-1)$ relative to the maximum value.

A PRN code generator utilizing an eight stage shift register will yield a code sequence that contains 255 elements. The required length of 250 may be obtained by truncating this 255 element code. The truncation is implemented either by counting clock pulses and parallel loading the initial state in the shift register each time 250 pulses are counted or by sensing the particular state corresponding to the 250th shift and parallel loading at that point.

The truncation of a maximal linear code sequence destroys the flat hash level and replaces it with a noise-like fluctuation.

The behavior of the hash level in truncated maximal linear codes for eight stage shift registers has been investigated to determine if there is any particular feedback configuration that gives the lowest hash level. The results of this investigation are summarized in Table 1.

*The 2-way and 1-way link codes may use different sequences.

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Table 1. Truncated Maximal Linear Code Hash Level

Stages Tapped for Feedback	Hash Level Peak	Ave.	Rms
2,3,4,8	22	7.4	5.1
2,5,6,8	22	8.0	6.0
1,3,5,8	22	7.2	5.5
3,5,6,8	26	7.8	6.0
1,6,7,8	30	7.2	5.4

There is no great difference among the five feedback configurations that were studied, except for the higher peak hash values found with the last two configurations. In general, the hash level for all configurations remained low (peak less than 6) for at least 16 code element delays either side of the zero delay maximum. The 1,3,5,8 feedback configuration showed the lowest hash level for the greatest delay either side of zero delay so, for this reason plus its good performance in terms of overall average and rms hash level, it has been selected for the STIFT system. Figure 6 shows one-half cycle of the autocorrelation function for the PRN code sequence generated by the 8-stage shift register with feedback taps from stages 1,3,5 and 8.

FREQUENCY DIFFERENCE MEASUREMENT

The STIFT microwave ground terminal is designed to provide frequency difference measurement between a space clock and a ground clock to an accuracy of 1 part in 10^{14}. The realization of this level of accuracy is made possible through a first order Doppler cancellation scheme that was utilized in the GP-A experiment in 1976. The key to the first order Doppler cancellation is the use of phase-locked frequency translators in the 2-way and 1-way receivers to reference the 2-way IF carrier to the ground maser frequency and the 1-way IF carrier to the space maser frequency.

Frequency Translation

The ground maser frequency undergoes two coherent frequency translations prior to being received by the ground terminal 2-way receiver. A translation of 82/55 is applied prior to transmission toward the space terminal and a 240/221 translation occurs in the space terminal transponder. Both of these translations are then counteracted by 221/240 and 55/82 translations in the 2-way receiver. The space maser is translated in frequency by the ratio 76/49 prior to being transmitted toward the ground terminal and this translation is counteracted by a 49/76 translation in the 1-way receiver.

The effects of frequency translations and Doppler on the 2-way link are analyzed as follows:

Carrier transmitted to space terminal, f_1

$$f_1 = f_0 \ (82/55) \qquad (1)$$

where f_0 = ground maser frequency.

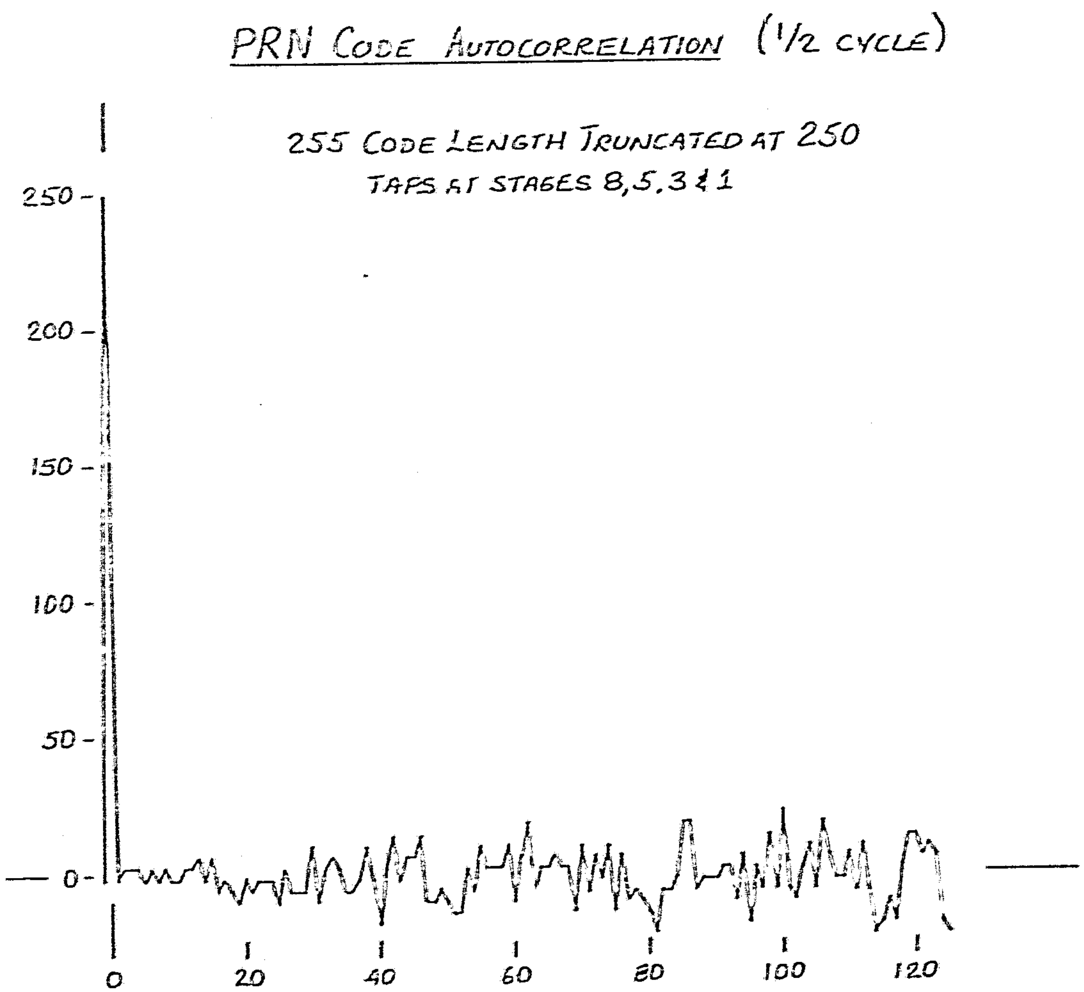

Figure 6. One-half Cycle of Autocorrelation for Truncated Eight Stage Shift Register PRN Code

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Carrier received at space terminal,

$$f_1 \ (1 + v/c) \quad (2)$$

where v = radial velocity of space terminal relative to the ground terminal
c = velocity of propagation

Carrier transponded from space terminal,

$$f_1 \ (1 + v/c) \ (240/221) \quad (3)$$

Carrier received at ground terminal

$$f_1(1 + 2 \ v/c) \ (240/221) \quad (4)$$

First IF carrier in 2-way receiver

$$f_1 \ (1 + 2 \ v/c) \ (240/221) \ -f_1 \quad (5)$$

2-way IF carrier after 221/240 and 55/82 translation

$$f_1 \ (1 + 2 \ v/c)(55/82)-f_1(221/240)(55/82)$$

or by substituting f_o (82/55) for f_1

$$f_o \ (1 + 2 \ v/c)-f_o(221/240) \quad (6)$$

This last expression, (6), is the 2-way IF carrier output reference to the the ground maser frequency. Note that only first order Doppler is shown in the expression above. Relativistic and gravitational effects are discussed in the section below on Doppler cancellation.

The effects of frequency translations and Doppler on the 1-way link are analyzed as follows:

Carrier transmitted from space terminal, f_2

$$f_2 = f'_o \ (76/49) \quad (7)$$

where f'_o = space maser frequency.

Carrier received at ground terminal,

$$f_2 (1 + v/c) \tag{8}$$

First IF carrier in 1-way receiver,

$$f_2 (1 + v/c) - f_1 \tag{9}$$

1-way IF carrier after 49/76 translation

$$f_2 (1 + v/c) (49/76) - f_1 (49/76)$$

or by substituting for f_2 and f_1

$$f'_o (1 + v/c) - f_o (49/76)(82/55) \tag{10}$$

Expression (10) is the 1-way IF carrier output referenced to the ground maser frequency f_o, and the space maser frequency, f'_o.

Doppler Cancellation

First order Doppler cancellation is achieved by dividing the 2-way IF carrier output, (6), by 2 and taking the difference between that and the 1-way IF carrier output (10).

The desired end result in this process is to obtain a direct measure of Δf, the frequency difference between the ground maser and the space maser.

$$\Delta f = f_o - f'_o \tag{11}$$

Carrying through the Doppler cancellation and the substitution of Δf for $f_o - f'_o$, the following result is obtained:

$$\Delta f + f_o (83/100320) + \left[(f_o/c^2) (\vec{r}_{sp} \cdot \vec{\alpha}_e) - 1/2 (\vec{v}_{sp} - \vec{v}_e)^2 + \Delta\phi\right] \tag{12}$$

In addition to the desired frequency difference. Δf, the output of the Doppler cancellation process contains; 1) a constant term, $f_o(83/100320)$, 2) an uncancelled Doppler term, $(f_o/c^2)(\vec{r}_{sp} \cdot \vec{\alpha}_e)$ where $\vec{r}_{sp}$ is the line-of-sight vector and $\vec{\alpha}_e$ is the ground terminal acceleration due to Earth rotation, 3) a second

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order Doppler term $(f\ /2C^2)(\vec{v}_{sp}-\vec{v}_e)^2$ where $\vec{v}_{sp}$ and $\vec{v}_e$ are the space and ground terminal velocity vectors and 4) a redshift term $(f_0/c^2)\Delta\phi$, where $\Delta\phi$ is the Newtonian potential difference between Earth and space. The constant term is removed by heterodyning the Doppler cancellation output with a synthesized frequency equal to f_0 (83/100320), approximately 1.1751621 MHz. The uncancelled Doppler term, second order Doppler and redshift terms are removed by post-real time processing based on the position and velocity information computed from the satellite flight profile relative to the ground terminal.

ACCURACY ANALYSIS

The accuracy expected from the STIFT microwave ground terminal is calculated using the following system parameters:

Microwave Ground Terminal	
Transmit Power 2-way uplink)	10 watts
Peak-to-Peak Phase Excursion	90 degrees
Receiver Noise Figure (includes input losses)	5 db
Antenna Gain (1 meter, 55% efficiency)	25 db
Polarization	Linear
PRN Code Element Duration	2 microseconds
PRN Code Sequence Length	250 elements
Space Terminal	
Transmit Power (2-way down link)	250 milliwatt
Transmit Power (1-way link)	250 milliwatt
Peak-to-Peak Phase Excursion	90 degrees
Receiver Noise Figure (2-way uplink)	13 db
Antenna Gain	3 db
Polarization	Circular
PRN Code Element Duration	2 microseconds
PRN Code Sequence Length	250 elements

Propagation Path (2 cases)		
Zenith Angle (ground terminal)	0 degrees	80 degrees
Range	437 km	1536 km
Path Loss	125 db	163 db
Atmospheric Loss	0 db	0.4 db

Link Noise Margins

Utilizing the system parameters listed above, the signal-to-noise ratios in a 1 MHz bandwidth at the receiver inputs are as follows

Ground Terminal Zenith Angle	0 degrees	80 degrees
S/N 2-way up link	13.8 db	2.8 db
S/N 2-way down link	6.1 db	-4.9 db
S/N 1-way down link	6.1 db	-4.9 db

If the space terminal transmit channels are increased to a 10 watt power output, the situation is improved as follows:

Ground Terminal Zenith Angle	0 degrees	80 degrees
S/N 2-way up link	13.8 db	2.8 db
S/N 2-way down link	22.1 db	11.1 db
S/N 1-way down link	22.1 db	11.1 db

When the carrier is phase modulated by a square wave type signal such as the PRN code, the fraction of carrier component in the output is given by:

$$(2/\Pi\beta) \sin (\Pi\beta/2) \tag{13}$$

where, β is the modulation index. For the case of 90 degree modulation (i.e., $\pm$ 45 degrees) β is unity and the carrier component, following phase modulation, has an amplitude of $2/\Pi$ relative to the unmodulated carrier. Thus, in the 90 degree PRN phase modulated signal, 40.5 percent of the energy is in the carrier and 59.5 percent is in the modulation sidebands.

Time Difference Uncertainty

The uncertainty in time difference measurement is computed by proceeding stepwise through the time discriminator and delay tracking loop. For the case of a 250 milliwatt space terminal transmit power and the 80 degree zenith angle, the full band signal-to-noise ratio in 1 MHz is -4.9 db. Since 59.5 percent of the signal power is in the modulation sidebands, the effective signal-to-noise ratio at the signal input of the time discriminator is -7.2 db. The demodulation and integration process in the time discriminator provides coherent enhancement to the signal-to-noise ratio, under locked conditions, that is equal to the number of code elements in one period of the code sequence. Thus, the signal-to-noise ratio at the output of the integrator for the STIFT code sequence of 250 elements is 40.8 db. When the A and B channels of the time discriminator are combined to form the difference (loop error) output, this signal-to-noise ratio is reduced to 37.8 db. Further reduction in noise occurs in the loop filter where 64 of the difference outputs are accumulated to provide a resultant signal-to-noise ratio of about 46.8 db. In terms of time uncertainty, this represents an rms value of about 4.6 nanoseconds as determined from the slope of the time discriminator response.

When the 2-way and 1-way time tracking loop outputs are subtracted to get the time difference, the overall uncertainty is about 6.4 nanoseconds. Further improvement of this uncetainty to a subnanosecond level will occur through averaging of the difference output and improved IF signal-to-noise obtained at lesser zenith angles and/or increased transmit power.

Frequency Difference Uncertainty

The frequency difference uncertainty is computed for the case of a phase-lock tracking bandwidth of 50 Hz. Taking the 40.5 percent of signal power in the carrier and the case of 250 milliwatt space terminal transit power and 80 degree zenith angle, the effective signal-to-noise ratio for the carrier in the full IF band is -8.8 db. Applying the reduction in bandwidth to this ratio yields a signal-to-noise of 34.2 in the 50 Hz tracking bandwidth. Combining the 2-way and 1-way receiver signals in the Doppler cancellation process reduces the signal-to-noise ratio to 31.2 db.

When the Doppler cancellation output is converted to a zero frequency base band, through heterodyning with a 1.17517621 MHz signal followed by low-pass filtering, the maximum anticipated frequency is the order of 1 Hz. Assuming a 100 second averaging time to measure this frequency difference, the uncertainty is about 5 parts in 10^{14} for the 250 milliwatt transmit power at 80 degree zenith angle. When the transmit power is increased to 10 watts in the space terminal, the calculated uncertainty improves to well below 1 part in 10^{14}.

SUMMARY

The design study of the STIFT system shows that, with the current state of the art in atomic hydrogen maser oscillators and with existing time and frequency measurement techniques, it is reasonable to expect accuracy levels of 1 nanosecond or less in time difference and 1 part in 10^{14} or better in frequency difference. The application of the STIFT system to time metrology on a global scale and to potential users such as the Deep Space Network, the Orbiting Space Station and the Very Long Baseline Interferometer stations would provide 1 to 2 orders of magnitude improvement over the present accuracy level.

ACKNOWLEDGMENT

The STIFT microwave ground terminal design has been supported through NASA grant NAG-8006.

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