N88-18785

511-35
128227

TDA Progress Report 42-92

October–December 1987

16P

Phase Calibration Generator

E. H. Sigman
Tracking Systems and Applications Section

A phase calibration system has been developed for the Deep Space Stations to generate reference microwave comb tones which are mixed in with signals received by the antenna. These reference tones are used to remove drifts of the station's receiving system from the detected data. This phase calibration system includes a cable stabilizer which transfers a 20 MHz reference signal from the control room to the antenna cone. The cable stabilizer compensates for delay changes in the long cable which connects its control room subassembly to its antenna cone subassembly in such a way that the 20 MHz is transferred to the cone with no significant degradation of the hydrogen maser atomic clock stability. The 20 MHz reference is used by the comb generator and is also available for use as a reference for receiver LOs in the cone.

I. Introduction

A Phase Calibration Generator (PCG) has been developed for the Deep Space Stations to provide phase calibration of the station's receiving system. This subsystem generates high stability microwave comb tones in the antenna cone which are referenced to the hydrogen maser in the control room. These comb tones are mixed with the signal received by the antenna. When the data from the received signal is processed, these tones are extracted from the data stream and their phase is used to determine the phase error which has been added to the data stream by the station's receiving system.

The PCG consists of a Transmitter unit in the control room and a Receiver unit in the antenna cone area. The Receiver unit contains the comb generator which generates tones at every integer multiple of $5/N$ MHz where N is an integer from 5 to 99. The PCG also contains a cable stabilizer which is split between the transmitting and receiving units. This cable stabilizer compensates for phase changes in the long cable connecting the Transmitter unit to the Receiver unit. The output of the cable stabilizer consists of stabilized signals at $5/N$ MHz and at 20 MHz. The $5/N$ MHz is used by the comb generator to generate the comb tones. The stabilized 20 MHz is used as a frequency reference in the cone for generation of receiver LOs. A 5 to 20 MHz converter in the Transmitter unit provides the 20 MHz for the cable stabilizer. The $5/N$ MHz is supplied by a divide-by-N circuit in the Transmitter which is synchronized to the station's 1 pps time reference. This synchronization assures that the delay from the station's 1 pps to the comb generator pulse will always be the same constant value. A detailed description of the PCG design and performance follows. First the cable stabilizer will be discussed, and then the comb generator will be described.

II. Cable Stabilizer

A. Description

A block diagram of the cable stabilizer is presented in Fig. 1. More detailed block diagrams of the transmitting and receiving

PRECEDING PAGE BLANK NOT FILMED

units are shown in Figs. 2 and 3. The Transmitter generates 20 MHz and $5/N$ MHz signals from the 5 MHz station reference. These 20 and $5/N$ MHz signals are sent up the cable to the Receiver where they are separated by filters and mixed together in a double balanced mixer. The resultant $20 + 5/N$ MHz and $20 - 5/N$ MHz components are sent back down the cable to the Transmitter for phase comparison with the 20 MHz signal being sent up the cable. The error signal from this phase comparison is used to drive a voltage controlled phase shifter (VCPS) which compensates for phase changes in the cable. The result of this process is stabilized 20 MHz at the Receiver unit. The $5/N$ MHz into the Receiver unit is not cable stabilized but is used by the GATE circuit to gate individual cycles of the stabilized 20 MHz in order to obtain stabilized $5/N$ MHz pulses to drive the comb generator.

The power divider through which the $20 + 5/N$ MHz and $20 - 5/N$ MHz components are sent down the cable does not have perfect isolation. Thus some $20 + 5/N$ MHz and $20 - 5/N$ MHz components will appear on the cable-stabilized 20 MHz in the Receiver. Since the $5/N$ MHz is not cable stabilized, these components could cause phase instabilities in the $5/N$ MHz output of the GATE circuit. Therefore, before the cable-stabilized 20 MHz is sent to the GATE, it is filtered by a crystal filter which attenuates the $20 + 5/N$ MHz and $20 - 5/N$ MHz components down to a level at which they will not significantly affect the phase stability of the GATE circuit's output. The 20 MHz also passes through a second crystal filter before being made available as an output for generation of the Receiver LOs. This second crystal filter is necessary because of the stringent limits on the maximum allowable $20 + 5/N$ and $20 - 5/N$ components for the 20 MHz frequency reference.

The phase comparison process in the Transmitter requires a $5/N$ MHz signal which has approximately the same time delay with respect to the Receiver as do the $20 + 5/N$ and $20 - 5/N$ signals. This is provided by the Receiver, which has a high input impedance at $5/N$ MHz. Since this input impedance does not match the 50 ohm line, the $5/N$ MHz signal is reflected back down the cable. A hybrid in the Transmitter separates this reflected $5/N$ MHz signal from the $5/N$ MHz being sent up the cable.

The operation of the cable stabilizer may be described in more detail as follows. The signal sent up the cable has the form:

$$\sin 2\pi 20t$$

where t is in μsec.

This is multiplied by a $5/N$ MHz signal in the Receiver's double balanced mixer so that the signal sent down the cable to the Transmitter is

$$\sin (2\pi 20t + \phi_1) \sin 2\pi \frac{5}{N} t$$

where ϕ_1 = two way cable delay.

The first mixer in the Transmitter multiplies this by 20 MHz to give

$$\sin 2\pi 20t \sin (2\pi 20t + \phi_1) \sin 2\pi \frac{5}{N} t$$

The product of the first two sinusoids generates a 40 MHz term and a dc term. The 3 MHz low-pass filter (LPF) filters out the higher frequency terms leaving a term of the form

$$\cos \phi_1 \sin 2\pi \frac{5}{N} t$$

The second mixer in the Transmitter multiplies this by the reflected $5/N$ MHz to give

$$\cos \phi_1 \sin 2\pi \frac{5}{N} t \sin \left(2\pi \frac{5}{N} t + \delta\right)$$

where δ = error between the phase of the reflected $5/N$ MHz and the phase of the $5/N$ MHz modulation on the 20 MHz, and -45 degrees $< \delta < 45$ degrees.

The $5/N$ MHz phase error, δ, is due to the cable dispersion and hybrid imperfections. The product of the sinusoids generates a $10/N$ MHz term and a dc term. After low-pass filtering in the loop filter, the remaining term is

$$\cos \phi_1 \cos \delta$$

This signal is integrated by the loop filter to generate a control signal for the VCPS. The VCPS controls ϕ_1 thus completing the feedback loop. The action of the feedback loop is to keep

$$\cos \phi_1 = 0 \tag{1}$$

or

$$\phi_1 = \frac{\pi}{2} + 2n\pi \tag{2}$$

where n is an integer. As a result, the total delay through the VCPS plus the cable is held constant.

Note that the $5/N$ MHz phase error, δ, only affects the amplitude of the detected signal and does not affect the phase at which lockup occurs as long as δ does not get close to ± 90 degrees. The use of the reflected $5/N$ MHz for demodulation assures that δ stays close to zero.

It can be seen that the feedback also has a potential lockup point for

$$\phi_1 = \frac{\pi}{2} + (2n+1)\pi \qquad (3)$$

However, this is an unstable lockup point because the feedback is positive. The type of feedback is determined by the slope of $\cos\phi_1 \cos\delta$. Thus, for $0 < \phi_1 < \pi$ the feedback is negative, whereas for $\pi < \phi_1 < 2\pi$ the feedback is positive. In the positive feedback region, the feedback loop will be unstable, which will cause the VCPS control voltage to swing to either its positive or its negative saturation limit. To avoid this problem, a search circuit is used. Whenever the VCPS control voltage gets too close to either its positive or its negative saturation limit, the search circuit takes over and sweeps the VCPS control voltage over its range until a stable lockup point is found. Since the VCPS is designed to have a range of slightly greater than 180 degrees (which makes its two way range slightly greater than 360 degrees), a stable lockup point will always be found. However, one must not get too close to the limits of the VCPS's range. If the VCPS control voltage gets too close to one of its limits, the search circuit will take over and find another lockup point near the other extreme of the VCPS's range. This will cause the output of the cable stabilizer to slip by one cycle. (The search circuit is described in more detail in a NASA technical brief which will be published soon.)

B. Error Sources

There are many error sources which can degrade the performance of the cable stabilizer. The most significant error sources which have been considered in the design of the cable stabilizer are discussed in the following sections.

1. Spurious signals. Spurious signals at any point in the cable stabilizer can alter the phase of the signal at that point in the circuit. The worst case is when the spurious signal is 90 degrees out of phase with the desired signal. Consider this case as shown in Fig. 4. The desired signal is the **D** vector and **S** is the spurious signal's vector. The resultant vector, **R**, is displaced in phase from the true value by the angle α. The phase error, α, is given by

$$\alpha = \tan^{-1}\left(\frac{\mathbf{S}}{\mathbf{D}}\right) \qquad (4)$$

where **S** and **D** represent the vector magnitudes.

For **S** small with respect to **D**, this may be approximated by

$$\alpha = \frac{\mathbf{S}}{\mathbf{D}} \qquad (5)$$

Let τ_α = the delay error.

Then, if f is the frequency of the desired signal,

$$\alpha = 2\pi f \tau_\alpha \qquad (6)$$

so that

$$\tau_\alpha = \frac{\mathbf{S}}{2\pi f \mathbf{D}} \qquad (7)$$

The design goal has been to keep the cumulative effect of all such errors less than 1 ps. From the above equation, any single such error will cause less than a 1 ps error at 20 MHz if $(\mathbf{S}/\mathbf{D}) < -78$ dB. Since there could easily be on the order of 10 such error sources in the cable stabilizer, if these error sources add in a root sum square manner, then each error source should have $(\mathbf{S}/\mathbf{D}) < -88$ dB. With this in mind, the cable stabilizer has been designed with the goal of keeping all spurious signals at least 90 dB down, and if possible 100 dB down.

Numerous spurious signal sources have been considered in the design of the cable stabilizer. For example, the 5 to 20 MHz converter in the Transmitter which generates the 20 MHz must have its spurious output components adequately suppressed. The 15 and 25 MHz components must be especially well suppressed since they are well within the bandpass of the 20 MHz bandpass filters (BPFs) of the cable stabilizer. Also, spurious signals from amplifiers in the 20 MHz paths of the cable stabilizer are kept within acceptable levels by using amplifiers with reasonably low distortion. These amplifiers are operated well below their maximum output level to assure that the harmonics they generate are kept down to an acceptable level.

In addition, interactions of components in the cable stabilizer have been considered. For example, spurious signals from one component which could generate unwanted intermodulation products in another component must be adequately suppressed. In particular, the signal voltages on the varicaps in the VCPS cause phase modulation of the signals in the VCPS. This generates intermodulation products of the 20 MHz and $20 + 5/N$ or $20 - 5/N$ MHz signals in the VCPS. These intermodulation products could generate intermodulation products in the mixer which might degrade the performance of the cable stabilizer. To avoid this problem, an extra 20 MHz BPF is used (shown on the left side of the VCPS in Fig. 2) to suppress intermodulation products generated by the VCPS. In addition, the amplitude levels of the signals in the VCPS have been chosen to keep errors from intermodulation products down to acceptable levels.

2. **Amplifier nonlinearities**. An additional effect which can be caused by amplifier nonlinearities is a shift of the zero crossing of a signal. This is of concern since the GATE circuit in the Receiver converts the 20 MHz sine wave to a square wave by using a zero crossing detector. If the gain of an amplifier is asymmetric about zero volts, the amplifier will generate even harmonics and shift the zero crossing time of a sine wave. The amount of zero crossing shift will be dependent on the amplitude of the sine wave. If the harmonic distortion is low, the amount of zero crossing phase shift for a given amplitude change is, to a good approximation, proportional to the second harmonic distortion. Thus the zero crossing shift can be determined from the second harmonic distortion. The amplitude sensitivity of the zero crossing as a function of second harmonic distortion is given in Table 1 for a 20 MHz sine wave. Note that the phase error is independent of frequency. Thus the time error is inversely proportional to frequency.

To keep the zero crossing shift errors low, the 10 dB amplifier which drives the GATE has a 20 MHz BPF on its output to attenuate the amplifier harmonics. This filter keeps the second harmonic at least 75 dB down. From Table 1, the amplitude dependence of zero crossing time will then be kept down to 0.14 ps/dB. The main source of amplitude changes in the cable stabilizer is the VCPS, which has a loss which varies as its phase setting is changed. For normal operation, the loss variation is well under 1 dB so that zero crossing shifts due to amplitude changes are under 0.1 ps.

3. **VSWR induced error**

a. Analysis. Standing waves on the cable between the Transmitter and Receiver caused by a mismatch between the cable and the components connecting to the cable can reduce the accuracy with which the cable stabilizer can measure a change in cable length and correct for it. Consider the case shown in Fig. 5 where a signal with a voltage magnitude V is transmitted from the left end of the cable to the right end. A mismatch at the right end reflects the signal back down the cable with magnitude V'. The magnitude of the reflection coefficient is

$$\rho_2 = \frac{V'}{V} \tag{8}$$

Similarly, at the left end the signal V' is re-reflected back up the cable with a magnitude of V''. The magnitude of the reflection coefficient at the left end is

$$\rho_1 = \frac{V''}{V'} \tag{9}$$

The signal received at the right end of the cable is the vector sum of V and V''

When the cable changes in length, the magnitude of the change can be determined by measuring the phase change of the signal received at the right end of the cable. The presence of the re-reflected signal, V'', will corrupt this measurement. The error will be largest when V'' is in phase with V or 180 degrees out of phase with V. Consider the case where V'' is initially in phase with V at the right end of the cable. Then let the cable stretch slightly so that the signal V at the right end of the cable increases in phase by $\Delta\phi$. We wish to measure this phase change $\Delta\phi$ to determine the change in cable length and correct for it. Let us calculate the error in measuring $\Delta\phi$ caused by the re-reflected signal V''. The signal V'' has traversed the cable three times by the time it reaches the right end so it will increase in phase by $3\Delta\phi$. Thus, after the cable stretch, the phase of V'' with respect to V will be $2\Delta\phi$. The phase relationship between these vectors is depicted in Fig. 6. The vector $\mathbf{V_R}$, which is the vector sum of V and V'', is the signal which will be detected at the right end of the cable. The phase angle, ϵ, between $\mathbf{V_R}$ and V is the error in measuring the phase of V and is thus the error in measuring $\Delta\phi$ from which the change in cable length is determined. From Fig. 6, the error angle ϵ is given by

$$\epsilon = \tan^{-1}\left(\frac{V'' \sin 2\Delta\phi}{V + V'' \cos 2\Delta\phi}\right) \tag{10}$$

From Eqs. (8) and (9)

$$V'' = \rho_1\rho_2 V \tag{11}$$

so

$$\epsilon = \tan^{-1}\left(\frac{\rho_1\rho_2 \sin 2\Delta\phi}{1 + \rho_1\rho_2 \cos 2\Delta\phi}\right) \tag{12}$$

For small changes in cable length,

$$2\Delta\phi << 1$$

so

$$\sin 2\Delta\phi \approx 2\Delta\phi$$

$$\cos 2\Delta\phi \approx 1$$

We then get

$$\epsilon = \tan^{-1}\left(\frac{2\rho_1\rho_2\Delta\phi}{1+\rho_1\rho_2}\right) \tag{13}$$

In the typical case

$$\rho_1\rho_2 << 1$$

so the above expression becomes

$$\epsilon = 2\rho_1\rho_2\Delta\phi \tag{14}$$

Let

$\Delta\tau$ = the change in cable time delay

τ_ϵ = error in measuring $\Delta\tau$

f = frequency of signal in cable

then

$$\Delta\phi = f\Delta\tau \tag{15}$$

$$\epsilon = f\tau_\epsilon \tag{16}$$

If we substitute Eqs. (15) and (16) into Eq. (14) we get

$$\tau_\epsilon = 2\rho_1\rho_2\Delta\tau \tag{17}$$

It is seen that the error in time delay measurement caused by a mismatch is independent of frequency.

Consider the case where

$$\rho_1 = \rho_2 = \rho \tag{18}$$

The voltage standing wave ratio (VSWR), S, is given by

$$S = \frac{1+\rho}{1-\rho} \tag{19}$$

so that

$$\rho = \frac{S-1}{S+1} \tag{20}$$

From the above expressions we can determine the worst case errors for measurement of cable delay change as a function of the VSWR. This worst case error is tabulated in Table 2 for a cable length change of 1 ns. In the Deep Space Stations the maximum cable length change over a 24 hour period is normally less than 1 ns so this table provides upper limits for VSWR induced errors.

b. Design. From the foregoing analysis it is seen that it is important to keep the VSWR of the cable from the Transmitter to the Receiver as low as possible along with all components which connect to the cable. Thus the VSWR of every component which connects to the cable has been carefully considered in designing the cable stabilizer. The 3 MHz Low-Pass Filters (LPFs) have tuned traps which isolate their circuitry from the cable at 20 MHz. If no other mismatches exist, the 3 MHz LPFs will degrade the VSWR to no worse than 1.02.

The 20 MHz BPFs use a circuit configuration which minimizes the VSWR at the center frequency and keeps the VSWR low in a symmetric manner about the center frequency. The physical layout of the circuit has been carefully determined for minimum VSWR, and each filter is individually tuned for minimum VSWR. Typical performance for these filters is a VSWR of <1.02 for 19.5 to 20.5 MHz and a VSWR of <1.06 for 19.0 to 21.0 MHz.

The VCPS has also been designed for low VSWR. The typical VCPS has a VSWR which drops to a minimum value of about 1.02 somewhere near the center of its phase correcting range. At the edges of its range the VSWR is <1.22. For each VCPS the control voltage at which it has minimum VSWR is determined in the lab. When the PCG is installed in the field, the length of the cable from the Transmitter to the Receiver is adjusted so that the cable stabilizer is operating with a control voltage close to this value. As a result of this procedure, the VCPS is normally in a range where its VSWR is <1.04. Then for the overall Transmitter the VSWR is normally <1.09.

For the Receiver, the input VSWR is normally <1.19. Using this number for the Receiver and 1.09 for the Transmitter, the reflection coefficients for the Transmitter and Receiver are:

$$\rho_T < 0.043$$

$$\rho_R < 0.087$$

If we put these numbers into Eq. (17) we get

$$\frac{\tau_\epsilon}{\Delta\tau} < 7.5 \text{ ps/ns}$$

for the worst case VSWR induced error.

The reciprocal of this number gives:

$$\text{cable correction factor} > 134$$

This number indicates the amount by which the cable stabilizer improves the cable performance. Both laboratory and field tests of the cable stabilizer have indicated that the typical cable correction factor obtained is 100 to 1000. The close agreement between measured performance and calculated performance seems to indicate that VSWR is the main limiting error source.

4. **Dispersion.** Dispersion in the cable or cable stabilizer components can generate errors in the cable compensation. The cable stabilizer feedback loop holds the two way delay in the cable constant. If the delay through the cable in each direction is equal, then the one way delay will be held constant. However, a typical cable will have some dispersion so that the delay at $20 + 5/N$ MHz and $20 - 5/N$ MHz will not be the same as the 20 MHz delay. But the $20 + 5/N$ and $20 - 5/N$ delays will differ from the 20 MHz delay by about the same amount except in opposite directions. Since the cable stabilizer uses both the $20 + 5/N$ and $20 - 5/N$ signals, the dispersion errors will almost cancel if the cable stabilizer has equal gain at these two frequencies and symmetric phase response about 20 MHz. The components of the cable stabilizer which limit its ability to meet these criteria are the 20 MHz BPF and the VCPS.

As mentioned previously, the 20 MHz BPF uses a circuit configuration which gives virtually symmetric behavior about the center frequency. The flatness of this filter is ±0.075 dB from 19 to 21 MHz. The attenuation at $20 + 5/N$ MHz matches the attenuation at $20 - 5/N$ MHz within 0.1 dB. The VCPS is flat to better than ±0.1 dB from 19 to 21 MHz. These performance specifications keep the gain of the cable stabilizer at $20 + 5/N$ MHz within 4 percent of the gain at $20 - 5/N$ MHz, thus minimizing dispersion effects.

5. **Choice of cable stabilizer frequency**. Since 5 MHz is the reference frequency used for the cable stabilizer, an integer multiple of 5 MHz is the logical choice for the cable stabilizer frequency. For a good crystal filter 20 MHz is about the highest practical frequency, so this was one of the driving factors in the choice of frequency. In addition, 20 MHz is a low enough frequency for the cable stabilizer to handle any expected cable length changes without a cycle slip. In practice, for typical cable length changes the VCPS stays well within its range of optimal VSWR performance when 20 MHz is used.

The effect of cable stabilizer frequency on the cable stabilizer's ability to correct changes in cable length is also of interest. As indicated in previous sections, errors due to spurious signals and amplifier nonlinearities decrease with increasing frequency. Also, phase noise tends to be constant with frequency so the time jitter caused by phase noise will be lower for higher cable stabilizer frequencies. However, the VSWR induced error is a function only of VSWR and is independent of frequency. Since the VSWR of most electronic circuits tends to increase with frequency, the VSWR induced error tends to increase with frequency. As indicated previously, in this cable stabilizer the VSWR induced error appears to be the dominant error source. Thus, if cable correction ability is the primary criterion used to select the cable stabilizer frequency, one would probably not want to choose a higher frequency than 20 MHz. In fact, it may be possible to improve the cable correction ability by a factor of 2 or so by going to a lower frequency. However, use of a lower frequency would increase phase noise effects. Thus, one would have to weigh phase noise effects against cable correction ability if another operating frequency for the cable stabilizer were to be considered.

6. **Thermal drifts**. The long term stability of the PCG is primarily limited by thermal drifts of the electronic components. To minimize the effects of environmental temperature changes, all critical electronic circuitry is housed in ovens which maintain a relatively constant temperature.

a. Ovens. All of the ovens used in the PCG have a copper or aluminum baseplate which acts as a temperature controlled surface. All electronic modules are built in aluminum boxes which are mounted on the baseplate with thermal grease to assure good thermal conduction. A thermistor at the center of the baseplate senses the baseplate temperature and connects to a proportional controller which controls the oven temperature. An aluminum box encloses the electronic modules and the baseplate. This box is surrounded by a layer of insulation and then by an outer box.

The Transmitter oven has its inner box completely covered with pad heaters which provide uniform heat. A layer of air between the inner and outer boxes provides insulation. A fan circulates air around the outer box to remove excess heat. The oven holds a nominal internal temperature of 50°C for external temperatures of 0°C to 25°C. Any external temperature change is reduced by a factor of about 100 on the inside of the oven.

The Receiver oven uses thermoelectric heat pumps to control the interior temperature by pumping heat into or out of the baseplate depending upon the outside temperature. The aluminum baseplate is mounted on six copper rods. Each copper rod connects to a heat dissipator plate on the outside box through a thermoelectric heat pump. The space between the inner and outer boxes is filled with urethane foam and Styrofoam insulation. The oven maintains a nominal inside temperature of 60°C for outside temperatures in the range of -55°C to +55°C. External temperature changes are reduced by a factor of about 100 on the inside of the oven.

In some installations the Receiver is too large to be mounted close to the comb generator's microwave injection point on the antenna assembly. In such cases, the comb generator is mounted in its own small oven called a CGA (Comb Generator Assembly). The CGA also uses thermoelectric heat pumps. Its copper baseplate is mounted on four thermoelectric heat pumps which in turn are mounted on the external heat dissipation plate. Styrofoam insulation is used between the inner and outer boxes. The oven holds a nominal internal temperature of 60°C for external temperatures of -55°C to +65°C. The oven reduces external temperature changes by a factor of about 150.

b. Temperature coefficients of electronics. The temperature coefficients of the various modules in the PCG have been measured by changing the set point of the oven and measuring the resultant phase change. The phase change was determined by using the long term stability test setup described in Section II.C.2.b. The comb generator phase change was measured using the comb generator test setup described in Section III.C. For the LEVELING-AMP-GATE the temperature coefficient was not measured but was determined from known characteristics of the chips in the circuit. The typical temperature coefficients of delay for the major modules in the PCG are shown in Table 3.

The overall typical temperature coefficients for the Transmitter and Receiver electronics are:

Transmitter electronics	13 ps/°C
Receiver electronics	
To comb generator output	48 ps/°C
To 20 MHz output	80 ps/°C

The ovens reduce the sensitivity of the electronics to the environment. For each of the PCG components the typical sensitivity to the environment is:

Transmitter	0.6 ps/°C
Receiver	
To comb generator output	0.5 ps/°C
To 20 MHz output	0.8 ps/°C
CGA	0.01 ps/°C

Note that the Transmitter oven reduces the environmental sensitivity of the electronics to about 0.1 ps/°C. However, the cable in the Transmitter chassis which brings the 5 MHz input from the back panel to the oven is not a phase stable cable and is not protected from the environment. This cable will add about 0.5 ps/°C to the environmental sensitivity of the Transmitter.

C. Performance

1. Cable correction ability. The cable correction ability of the cable stabilizer has been tested in the laboratory by using the long term stability test setup which is described in Section II.C.2.b. A piece of cable with a known delay of about 2 ns was added to the cable between the Transmitter and Receiver, and the delay change in the Receiver's output was measured. This test has been performed for various values of N as part of the acceptance testing for each unit which has been installed in the field. The results of these tests indicate that the normal range of cable correction ability is 100 to 1000.

The first two PCGs which were installed at the Goldstone Deep Space Station were also tested in the field for cable correction ability. The phase of the phase calibrator tone was monitored through the station's receiving system while a cable with a 4.3 ns delay was added to the cable from the Transmitter to the Receiver. The results of this test are shown in Table 4. This test also showed the cable correction ability to be in the range of 100 to 1000.

2. Long term stability

a. Expected performance. The following estimates for the overall performance of the PCG subsystem in a Deep Space Station are based upon the data presented in the previous sections. The cable from the Transmitter to the Receiver is taken to be a 1000-foot-long hard line. Such a cable would typically have a temperature coefficient of 25 ppm/°C and a delay of 1.5 microseconds. This would give a temperature coefficient for the cable delay of 37.5 ps/°C. Since the cable stabilizer corrects the cable by a factor of at least 100, the corrected cable would have a temperature coefficient of less than 0.375 ps/°C.

A setup is considered in which the comb generator is separate from the Receiver in its own CGA oven. It is assumed that the CGA is connected to the Receiver through a 10-foot-long phase stable cable with a temperature coefficient of 14 ppm/°C. The delay through such a cable would be 15 ns, which would give a delay temperature coefficient of 0.2 ps/°C.

The coupling of the comb generator into the front end of the station's microwave receiving system is also considered. The present coupler is a loop coupler which does not have adequate performance for phase calibration applications. This coupler is going to be replaced with a Bethe hole coupler on the antenna's feed horn. Aside from the improved performance which will be obtained, an important advantage of the Bethe hole coupler is that it will inject the phase calibrator tone as far forward in the receiving system as is possible.

The estimated environmental temperature changes for the PCG subsystem over a 24 hour period are taken as:

Transmitter	±1°C
Cable (from Transmitter to Receiver)	±5°C
Cone (includes Receiver, cable from Receiver to CGA, CGA, and Coupler)	±15°C (±5°C)

For the cone the estimates indicated are for poor environmental control. Numbers in parentheses indicate performance for more reasonable control of cone temperature. With these temperature variations, the expected delay variations over 24 hours can be calculated from the temperature coefficients and are shown in Table 5.

It is seen that the main limitations on performance are probably the loop coupler and temperature control of the cone area. When the Bethe hole coupler is installed, the PCG should meet its specification of 10 ps for delay variations over 24 hours. Reasonable control of the cone temperature will assure that the PCG is well within this specification.

Over a 1000 second period the estimated environmental temperature changes for the PCG subsystem are taken as:

Transmitter	±0.5°C
Cable (from Transmitter to Receiver)	±0.16°C
Cone (includes Receiver, cable from Receiver to CGA, CGA, and Coupler)	±0.5°C (±0.16°C)

For the cable and the cone, these estimates are derived by assuming that the variations shown above for 24 hours occur linearly over an 8 hour period. From these temperature variation estimates the expected delay variations over 1000 seconds can be calculated and are shown in Table 6. In Table 6 the calculated delay variations are divided by 1000 sec to obtain the expected Allan variance.

We can also obtain expected Allan variance for the 20 MHz output of the cable stabilizer which is used as a reference for receiver LOs. From the temperature variations shown above and the previously quoted temperature coefficients the results shown in Table 7 were obtained.

Note that the cone temperature variation is probably the primary limiting factor for cable stabilizer performance. If temperature variations in the Transmitter are significant, an improvement in performance may be attainable if the cable bringing the 5 MHz into the Transmitter oven is replaced with a shorter cable or a phase stable cable.

b. Measured performance. The long term stability of cable stabilizers was tested in the laboratory with the setup shown in Fig. 7. This setup measures the phase of the 20 MHz output with respect to the 5 MHz input. The delay resolution on the output is about 1 ps. As part of the acceptance testing for each PCG which has been built, the cable stabilizer is left running for at least 24 hours and its performance is recorded. The delay at different points in time is read off the strip chart and the Allan variance is calculated from these delay points. The typical Allan variance which has been seen for the 20 MHz output of the cable stabilizer is shown in Table 8.

Allan variance tests have also been performed on the cable stabilizer at the Maser Test Facility of the Frequency and Timing Subsystem Group. For these tests the output of a cable stabilizer running off one hydrogen maser was compared with another maser whose output was down-converted to 20 MHz. The results showed no measurable degradation of the maser stability. In these tests the Allan variance for each of the masers alone was 1×10^{-15} over 1000 sec.

The long term stability tests in our laboratory showed that the long term behavior was dominated by a drift which was almost linear with time and had a drift rate which decreased with time. In one typical case, the drift was 41 ps for the first 24 hours, 19.5 ps for the second 24 hours, and 17.5 ps for the third 24 hours. This drift is caused by the crystal filter. The crystal filter manufacturer does burn in the crystal filters for a few months to reduce aging effects. However, each time a crystal filter is brought up to operating temperature it will take some time to stabilize. No long term stability tests have been performed in our laboratory after more than a 3 day warm-up. However, in normal operation, the oven will be kept at operating temperature on a continuous basis. This will minimize crystal filter drifts.

3. Phase noise. The phase noise of the cable stabilizer was measured by mixing the output of two cable stabilizers together in a mixer. The phase of one cable stabilizer was delayed in order to zero out the dc component in the mixer output. The remaining output of the mixer, which is the phase noise, was amplified and measured. The typical total phase noise measured in a 3 MHz bandwidth was:

20 MHz output	0.4 ps
20 MHz to Leveling-Amp-Gate	0.6 ps

The phase noise spectrum was measured at the Maser Test Facility of the Frequency and Timing Subsystem Group. A similar procedure was followed except that a digital spectrum analyzer was used as the measuring device. The typical results that were obtained are shown in Table 9.

III. Comb Generator

A. Description

The microwave comb generator generates a comb spectrum from 2 to 10 GHz. The input is the $5/N$ MHz pulses from the cable stabilizer which have a repetition rate of up to 1 MHz. The comb generator uses a step recovery diode (SRD) which is pulsed at the input repetition rate to generate the comb tones. Whenever the diode current switches from forward bias to reverse bias, the diode continues conducting for a time period equal to τ, the minority carrier lifetime, and then very quickly switches to the nonconducting state in 50 ps. This rapid change in diode current induces a 50 ps voltage pulse in an inductor which is in series with the diode. This voltage pulse, which is the output of the comb generator, has harmonics up through 10 GHz which are coherent with one another in phase.

The time of occurrence of the comb generator's output pulse depends upon the τ of the SRD. The τ is dependent on temperature and on the ratio of forward SRD current to reverse SRD current. The forward and reverse currents of the SRD are set by a high stability dc bias circuit which precisely controls their ratio in order to keep τ at a value of about 1.5 ns. The temperature coefficient of τ is 1 percent/°C so the temperature dependence of the SRD would be 15 ps/°C. A thermistor in the dc bias circuit changes the ratio of the forward and reverse currents in a manner which compensates for the SRD's temperature dependence at the operating temperature of 60°C in order to minimize temperature dependence. Tests of the comb generator have shown that the overall temperature coefficient at 60°C is about 1 ps/°C.

The magnitude of the SRD output can be set by a digital control word. The forward and reverse SRD currents are determined by the dc bias circuit which maintains them at values proportional to the digital control word. The ratio of the forward and reverse currents is always maintained at the same constant value except for temperature compensation.

B. Dispersion

The output of the comb generator must have very low dispersion, that is, very small deviation from a linear phase versus frequency characteristic. For VLBI (Very Long Baseline Interferometry) the performance goal for the comb generator is to limit dispersion to ±0.88 degree over a 400 MHz bandwidth.

The main cause of dispersion is VSWR in the microwave circuitry. Consider the cable in Fig. 5 where a signal V is being sent from the left end to the right end. Any mismatch at the right end will reflect a signal V' back down the cable with a reflection coefficient of ρ_2. Similarly, at the left end a signal V'' is reflected back up the cable with a reflection coefficient of ρ_1. At the right end of the cable the re-reflected signal, V'', will add vectorially to V and alter its phase by an amount ϵ. This situation is shown in Fig. 6 where the two way cable delay, $2\Delta\phi$, is the phase difference between V and V'' at the right end of the cable. Since the cable delay, $\Delta\phi$, is a function of frequency, the phase deviation of the signal, ϵ, will be a function of frequency. The maximum positive value of ϵ will occur when $2\Delta\phi = 90$ degrees, and the maximum negative value will occur when $2\Delta\phi = -90$ degrees. The frequency span over which ϵ changes from a maximum positive value to a maximum negative value will be the frequency for which the cable is one quarter wavelength long.

Since it is desirable to make the rate of change of ϵ as small as possible with respect to frequency, it is best to make the cable as short as possible. This will be the case when the Bethe hole couplers are installed in the Deep Space Stations. The CGA unit will be physically mounted right next to the Bethe hole coupler. Its output connector will connect directly to the coupler's connector with no intervening cable. The effective cable from the comb generator output to the coupler will then be about 15 cm long. The resultant frequency span from maximum positive ϵ to maximum negative ϵ will be about 300 MHz. Thus, ϵ will always have at least one maximum in a 400 MHz frequency span. In order to meet VLBI requirements, the maximum ϵ should be less than 0.88 degree. The maximum value of ϵ can be determined by putting $2\Delta\phi = 90$ degrees into Eq. (12) to get

$$\epsilon = \tan^{-1}(\rho_1\rho_2) \tag{21}$$

or

$$\rho_1\rho_2 = \tan\epsilon \tag{22}$$

For $\epsilon < 0.88$ degree

$$\rho_1\rho_2 < 0.0154$$

Consider

$$\rho_1 = \rho_2 = \rho$$

Then

$$\rho < 0.124$$

So from Eq. (19) the requirement on the VSWR is

$$S < 1.28$$

The coupler which injects the comb tones into the microwave receiving system must meet this requirement. The comb generator output must also meet this requirement. Since the SRD is a very poor match to a 50 ohm line, the comb generator has a 10 dB attenuator on its output to isolate the SRD from the line. The output VSWR of a 10 dB attenuator is always <1.25 no matter how poor its input match. The line from the SRD to the 10 dB attenuator input has a poor match at the SRD end, but its length is kept very short (<3 mm) so that the phase variation with frequency is minimized.

C. Performance

The long term stability of the comb generator has been measured with the setup shown in Fig. 8. A PCG Transmitter and Receiver were set up with a Comb Generator inside the Receiver. A second comb generator inside a separate oven was also connected to the Receiver's cable stabilizer output. A power switch operating off a 1 pulse per second (1 pps) signal alternated power between the two comb generators so that one comb generator operated on even seconds and the other comb generator operated on odd seconds. A power combiner connected the outputs of the two comb generators to a receiving setup consisting of a microwave receiver, IF Converter, and Formatter to down-convert the microwave signal and digitize it. The digital data stream went to a Digital Tone Extractor (DTE) which determined the phase of the phase calibration tone. The DTE was synchronized with the 1 pps and determined the phase every second by integrating over the last half of each second. This allowed the phase of each comb generator to be tracked separately over time.

Two tests were performed with the test setup. In the first test the temperature coefficient of the comb generator was determined by varying the set point of the second oven and observing the change in relative phase of the two comb generators. In the second test the oven temperatures were held constant and the phase behavior of the two comb generators was recorded over time. The Allan variance of the phase difference between the two comb generators was determined for various time intervals. This Allan variance was divided by the square root of 2 to determine the Allan variance for a single comb generator. The results of these tests are shown in Table 10. It can be seen that the stability of the comb generator is significantly better than the cable stabilizer. Thus, for the entire PCG the stability is limited by the cable stabilizer.

IV. Very Long Baseline Interferometry Results

In field usage the PCG has proven to be very useful and in some cases invaluable for improving the quality of VLBI data. Analysis of VLBI experiments in which the prototype PCG unit was used at Deep Space Station 13 has been performed by Chris Jacobs at JPL. For phase delay tests in the 8.4 GHz (X) band, he found that application of phase calibrator correction to the data reduced noise, thus making it easier to resolve ambiguities and connect phase points. In another experiment, the Traveling Wave Maser (TWM) was drifting badly and had to be retuned often. Application of the phase calibrator correction saved this experiment.

Some results of another experiment which was saved by the PCG are shown in Figs. 9 and 10. In this experiment the differential group delay between Deep Space Station 13 and 45 was being determined using 40 MHz Bandwidth Synthesis (BWS). The TWM at station 13 was erroneously set to 20 MHz bandwidth so that all the BWS channels being recorded were outside the TWM bandwidth. The differential group delay over time for the two outermost BWS channels is shown in Fig. 9. At each point the error bar due to system noise is shown. For the first part of the experiment it can be seen that the drift over time is large compared to the error bars, thus degrading the data. The same data after phase calibrator correction is shown in Fig. 10. It is seen that the drift is virtually gone. Without the phase calibrator correction there were many points for which the ratio of group delay error to system noise was on the order of 30. Phase calibrator correction dropped this ratio to less than 1, thus making the data usable.

Table 1. Amplifier non-linearity: f = 20 MHz

Second harmonic attenuation from fundamental (dB)	$\frac{\text{Zero crossing shift}}{\text{Amplitude change}}$ (ps/dB)
40	8
50	2.5
60	0.8
70	0.25
75	0.14
80	0.08

Table 2. Cable change measurement error due to VSWR

VSWR	Error (ps)
1.05	1.18
1.10	4.52
1.15	9.74
1.20	16.5
1.25	24.7

Note: Worst case error in measurement of 1 ns change in cable delay.

Table 3. Temperature coefficients for modules

Module	Temperature coefficient (ps/°C)
5 to 20 MHz converter	1
10 dB amplifier	3
20 dB amplifier	6
10 dB amp-filter-coupler	25
20 dB amp-filter	29
1st crystal filter	10–14
2nd crystal filter	3–5
Leveling-amp-gate	2
Comb generator	1

Table 4. Field test of cable correction ability conducted at Goldstone Deep Space Station (cable delay increased by 4.3 ns)

Frequency band	N	Cable correction error (ps)	Correction ability
2.2 GHz (S)	5	4.4	977
2.2 GHz (S)	10	34.9	123
8.4 GHz (X)	5	11.9	361
8.4 GHz (X)	10	27.3	158

Table 5. Expected 24 hour stability for comb tone output

Component		Delay variation (ps)				
Transmitter		±0.6				
Stabilized cable		±1.9				
Receiver		±7.5	(±2.5)			
Cable to CGA		±3.0	(±1.0)			
CGA		±0.15	(±0.05)			
Coupler	Loop	±15.0	(±5.0)	Bethe	±1.5	(±0.5)
RSS		17.2 ps	(6.0 ps)		8.5 ps	(3.4 ps)

Note: Numbers in parentheses indicate performance with moderate temperature control of cone.

Table 6. Expected 1000 second stability for comb tone output

Component	Delay variation (ps)					
Transmitter		±0.3				
Stabilized cable		±0.06				
Receiver		±0.25	(±0.08)			
Cable to CGA		±0.1	(±0.03)			
CGA		±0.005	(±0.002)			
Coupler	Loop	±0.5	(±0.16)	Bethe	±0.05	(±0.016)
RSS		0.65 ps	(0.36 ps)		0.41 ps	(0.32 ps)
Expected Allan variance (1000 sec)	6.5×10^{-16}		(3.6×10^{-16})	4.1×10^{-16}		(3.2×10^{-16})

Note: Numbers in parentheses indicate performance with moderate temperature control of cone.

Table 7. Expected 1000 second stability for 20 MHz output

Component	Delay variation (ps)	
Transmitter	±0.3	
Stabilized cable	±0.06	
Receiver	±0.4	(±0.13)
RSS	0.50 ps	(0.33 ps)
Expected Allan variance (1000 sec)	5×10^{-16}	(3.3×10^{-16})

Note: Numbers in parentheses indicate performance with moderate temperature control of cone.

Table 9. Measured phase noise spectrum: typical values for 20 MHz output

Frequency offset from carrier (Hz)	Phase noise in 1 Hz bandwidth (dB from carrier)
1	−107
10	−118
100	−125
1000	−125
10,000	−130

Table 8. Measured stability of 20 MHz output

Time interval (sec)	Allan variance
1	1.5×10^{-13}
10	2×10^{-14}
100	3×10^{-15}
1000	4×10^{-16}
10,000	1×10^{-16}

Table 10. Measured comb generator stability

Temperature coefficient:	1 ps/°C	
Allan variance:	3×10^{-16}	over 100 sec
	7×10^{-17}	over 1000 sec
	1×10^{-17}	over 10,000 sec

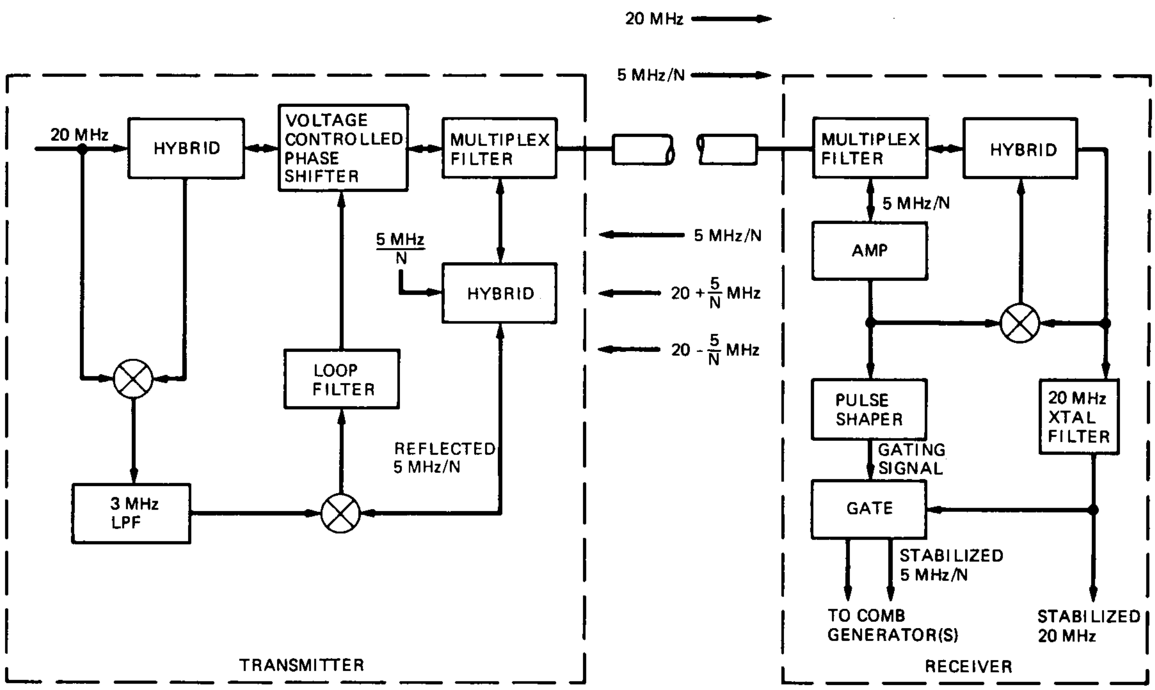

Fig. 1. Cable stabilizer

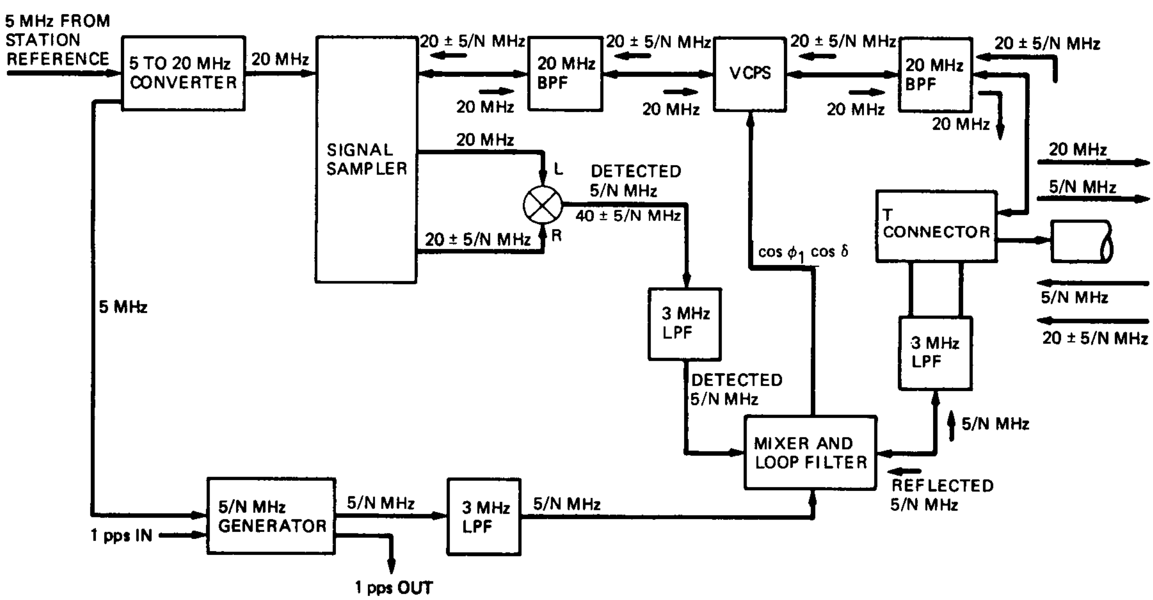

Fig. 2. PCG transmitter

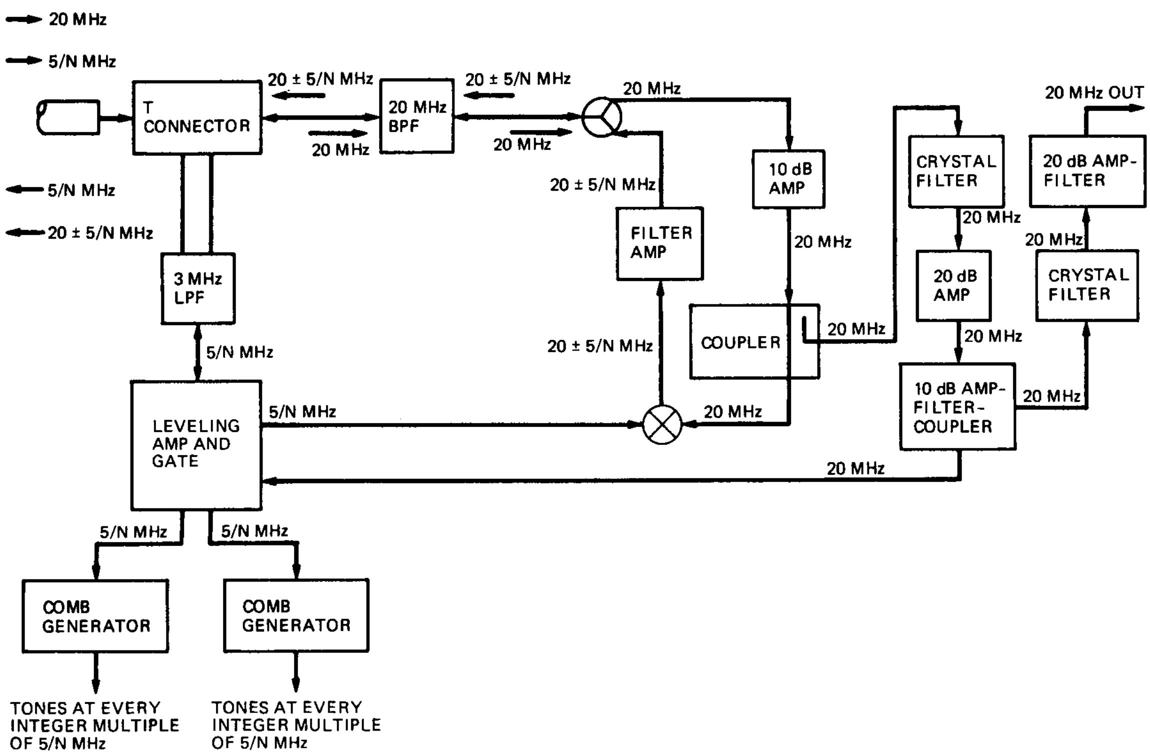

Fig. 3. PCG receiver

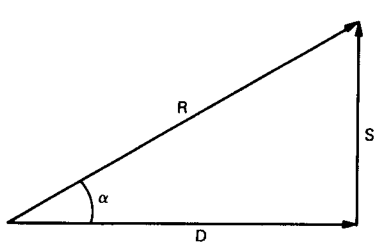

Fig. 4. Phase error due to spurious signals

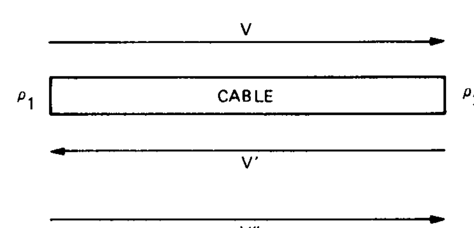

Fig. 5. Cable reflections

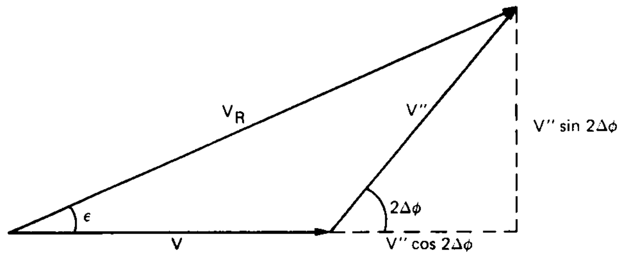

Fig. 6. Phase error due to cable reflections

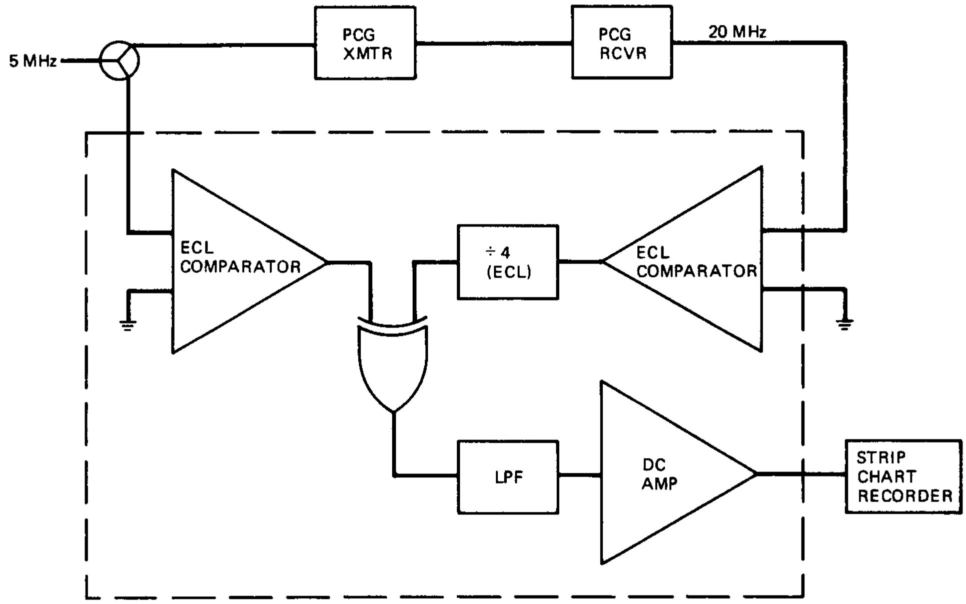

Fig. 7. Test setup for long term stability measurement

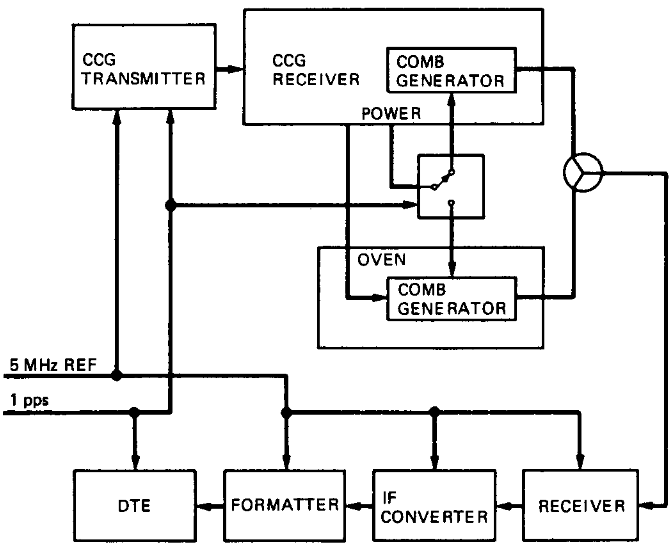

Fig. 8. Comb generator test setup

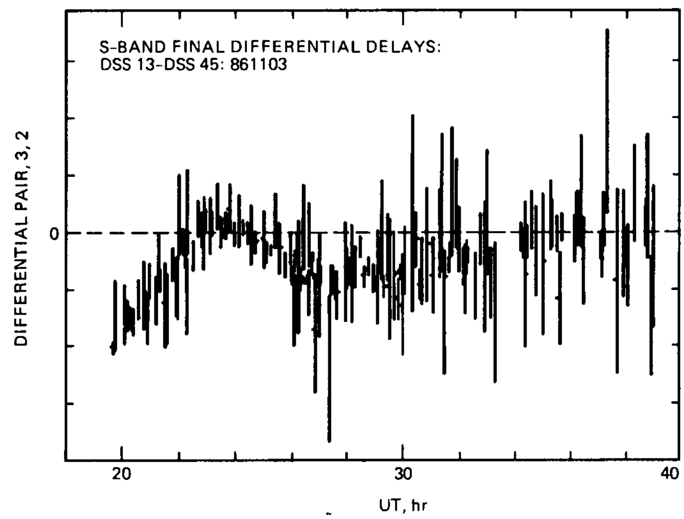

Fig. 9. VLBI results without phase calibration correction

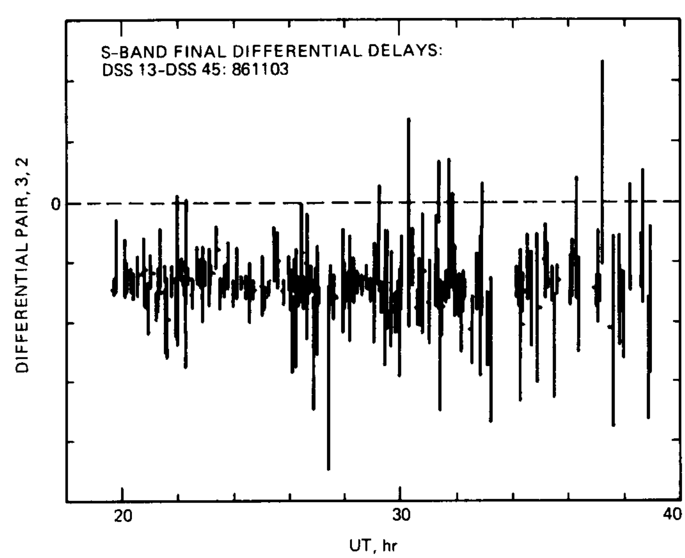

Fig. 10. VLBI results with phase calibration correction