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A 10.6-μm LASER-RECEIVER RF SUBSYSTEM

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16. Abstract

Both government and industry are developing optical communications systems to provide greater communications capability. One of the problems inherent in the use of such a system is the exceedingly high doppler shift which can be expected. Until now, receivers capable of tracking frequency deviations as large as 1 GHz and more were unheard of. With the recent development of stable, linear, octave-bandwidth, voltage-tunable oscillators, this problem has been solved. A laboratory tracking receiver which is capable of tracking a 1-GHz frequency shift at rates up to 12 MHz/s with subsequent coherent data demodulation has been designed and tested. This work demonstrates that the design and fabrication of an RF subsystem, utilizing coherent demodulation of data, for wideband communications systems is now both feasible and practical.

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INTRODUCTION

A good deal of work has been done in recent years on the feasibility of using lasers as communications links between deep-space satellites and earth-based stations and between two orbiting spacecraft.* However, considerably more work remains to be done. Spacecraft optical equipment must be developed and satisfactorily tested, research must be continued to develop better IR detectors, the life expectancy of laser tubes must be improved, a better understanding of the effects of the atmosphere on laser transmissions must be obtained, and the problems associated with acquisition and tracking of the narrow laser beam must be solved. These and other problem areas must be eliminated before a serious assessment of optical communication systems can be made. The laser communications experiments on the ATS F and G are designed to demonstrate most of this technology.

Associated with any coherent optical communications system is the RF support equipment which follows the first mixer. Perhaps the most difficult problem encountered in an RF subsystem is the accommodation of the exceedingly high Doppler offsets that generally occur in systems using heterodyne detection techniques for satellites in nonsynchronous orbits.

EXPERIMENT DESCRIPTION

Since 1965, Goddard Space Flight Center has been active in an investigation of the feasibility of a 10.6-μm laser communications system. Work began with a preliminary analysis** and has progressed to a prototype system which is about to be field tested. The GSFC 10.6-μm prototype communications system consists of an optical ground-station transmitter and receiver with supporting RF equipment. The experiment will be carried out using reradiation of modulated and unmodulated 10.6-μm

*McAvoy, N., "10.6 Micron Communication System", NASA-Goddard Space Flight Center Document X-524-65-641, Goddard Space Flight Center, Greenbelt, Md., November, 1965.
McElroy, J. H., et al., "An Advanced 10.6 Micron Laser Communications Experiment", NASA-Goddard Space Flight Center Document X-524-68-478, Goddard Space Flight Center, Greenbelt, Md., November, 1968.
Richards, W. E., "10.6 Micron Transceiver", NASA-Goddard Space Flight Center unpublished report.
**McAvoy, N., op cit.

energy from the satellite Pageos, in an effort to evaluate optical and RF components developed for laser communications systems and to determine the problems inherent in the use of such systems.

In conjunction with the passive reradiation experiment, it was determined* that the maximum Doppler shift and maximum rate of change of Doppler would approach 1 GHz at rates up to 12 MHz/s. In order to extract information from the received signal, it was necessary to build a receiver whose predetection bandwidth would be 1 GHz. Since data bandwidths are typically much smaller than the maximum Doppler shift, the receiver had to be a coherent phase-lock receiver so that the signal-to-noise ratio would not be degraded by the detector. State-of-the-art tracking receivers had not been developed which were capable of tracking carrier shifts of 1 GHz. The objective of the RF program was to determine if a phase-lock receiver could be developed using conventional carrier tracking-loop components.

LASER COMMUNICATIONS RECEIVER RF SUBSYSTEM

General Description

This report will not describe the optical transmitter or receiver, but will concern itself only with the laser-receiver RF subsystem. Figure 1 is a simplified block diagram of a laser receiver for the 10.6-μm laser communications experiment; the RF subsystem for this experiment has already been designed and tested. The laser carrier is optically mixed with a laser local oscillator to translate the received signal to an IF signal which varies between dc and 1 GHz. This signal is fed to an RF subsystem that is functionally not unlike the tracking receivers NASA currently uses. The receiver subsystem consists of a carrier tracking loop for removing the Doppler shift, a discriminator loop to aid in carrier acquisition, and a subcarrier demodulator to obtain data. The RF subsystem is a coherent system; i.e., the subcarrier and all reference frequencies are generated from the same 5-MHz standard.

Carrier Tracking Loop

The carrier tracking loop in Figure 1 is a conventional second-order phase-lock loop (Reference 1) consisting of a broadband preamplifier and mixer for obtaining a fixed IF frequency, a narrowband filter, a phase detector, and a voltage-controlled oscillator (VCO) for tracking the carrier signal. The VCO produces a 25-MHz to 1.025-GHz local-oscillator signal for the signal mixer. When the loop is locked, i.e., when the VCO is tracking the input signal, the loop IF signal is fixed at 25 MHz. The narrowband IF filter passes the 25-MHz IF signal to the loop phase detector, eliminating the sidebands due to the 10-MHz subcarrier. The phase-detector output is a measure of the phase difference between the 25-MHz IF signal and the 25-MHz reference signal. This error voltage is filtered by a low-pass filter which suppresses noise and high-frequency signal components. The filter also helps to establish the dynamic performance of the loop. The filtered voltage is then applied to the loop VCO, thereby changing the frequency in a direction that reduces the phase difference between input and reference signals.

*Chu, Y., "Applications of MIMIC Language at GSFC", NASA-Goddard Space Flight Center Document X-700-67-96, Goddard Space Flight Center, Greenbelt, Md., February, 1967.

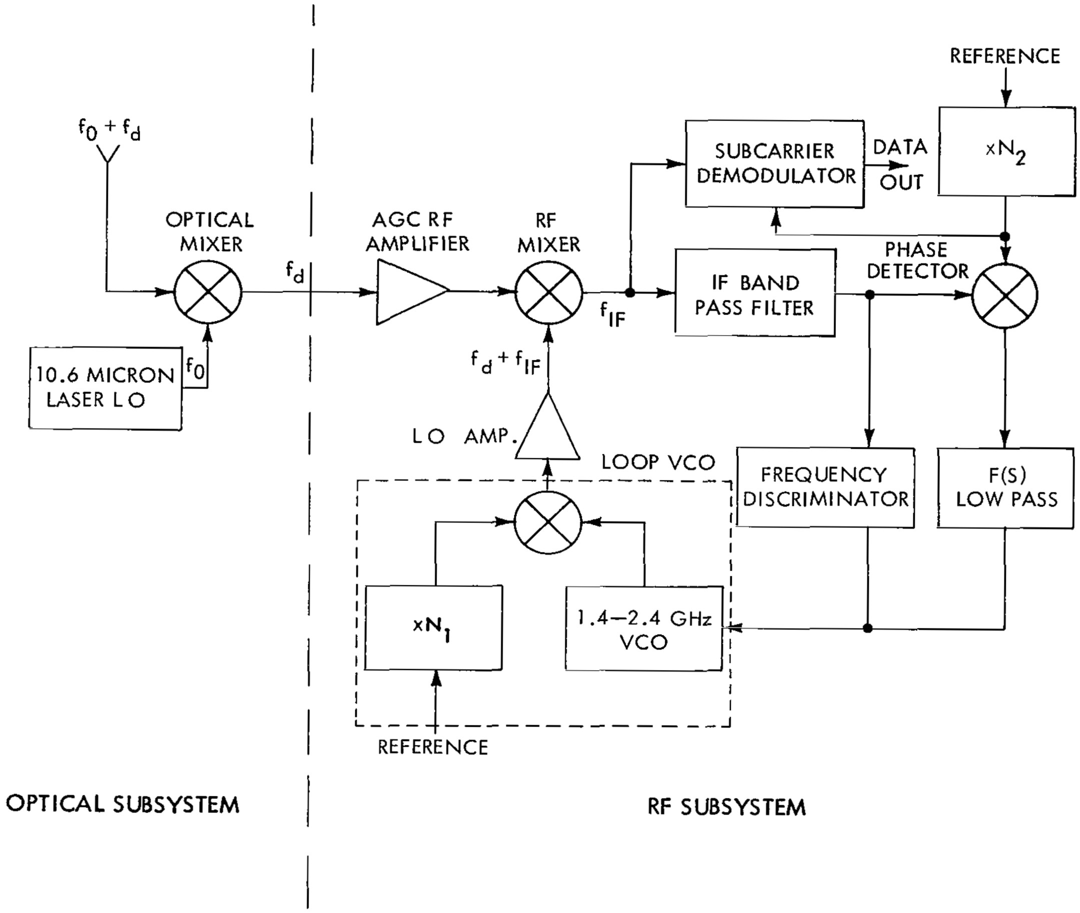

Figure 1—A simplified block diagram of a laser receiver.

Acquisition Loop

In order to facilitate rapid acquisition during instances where a good signal-to-noise ratio input to the loop exists, an acquisition loop having a wider bandwidth than the carrier tracking loop is built within the tracking loop. For acquisition, the composite loops would have a wide bandwidth, whereas for tracking, the loop bandwidth would be considerably narrowed. The acquisition loop, also shown in Figure 1, employs a frequency discriminator in a conventional AFC arrangement. If the frequency difference is large, the discriminator pulls the VCO toward the direction of lock; when the difference comes within range of the tracking loop, the phase detector takes over and locks the loop.

Phase Demodulator

In order to obtain the transmitted PM data, the prefiltered IF spectrum, which contains the 10-MHz subcarrier, is heterodyned with the 25-MHz reference signal to produce the subcarrier signal. This signal is then phase-locked to a 10-MHz signal in the subcarrier demodulator loop shown in Figure 2.

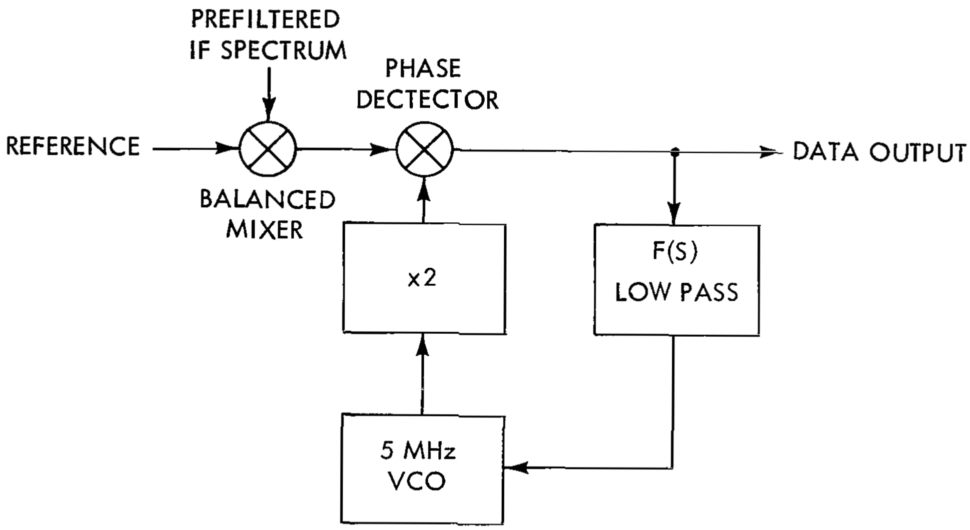

Figure 2—Subcarrier demodulator.

The demodulator is a narrowband second-order phase-lock loop, which produces the detected data at the loop phase-detector output. The 10-MHz phase-modulated subcarrier is obtained by phase modulating a 5-MHz signal from the reference oscillator. The modulation index of the 5-MHz signal is kept small (approximately 0.08 radians) so that good linearity and low distortion will be maintained. The signal is then multiplied to 90 MHz, where it is coherently down converted to 10 MHz to serve as the system subcarrier. The resultant modulation index is approximately 1.4 radians with linearity and distortion less than 3 percent. The modulating-signal input has a response from dc to 30 kHz. The data bandwidth is therefore approximately 200 kHz.

KEY RECEIVER COMPONENTS AND THEIR DEVELOPMENT

Key components of this receiver are the broadband variable-gain preamplifier used to maintain a fixed signal level at the input to the phase detector, a voltage-controlled oscillator capable of tracking a Doppler shift of 1 GHz, a 1-GHz-bandwidth amplifier for the carrier loop VCO signal, and broadband high-frequency balanced mixers for generating the loop IF and carrier-loop VCO signals.

The beginning of the wideband receiver development effort early in 1969 coincided with the introduction of a commercial wideband high-frequency balanced mixer. Soon after, a GSFC-sponsored development contract produced amplifiers with a 1-GHz bandwidth. Sweeper manufacturers were in the process of developing solid-state, octave-bandwidth, voltage-tunable transistor oscillators for a new line of solid-state sweepers. These oscillators are conventional LC oscillators which use a varactor in the tank-circuit for tuning 1 GHz in frequency. Normally, the tuning range is limited by an upper frequency bound which is determined by the transistor output capacitance; however, by the use of the transmission line techniques reported by Herbert and Chernega (Reference 2) the shunt capacitance can be used advantageously to greatly increase the upper frequency bound and hence the tuning range of the oscillator.

The first generation of these wideband oscillators were limited in their performance. The transfer characteristics of these oscillators were very non-linear, causing variations in tracking-loop bandwidth of as much as 30 percent. Voltages of 60 V and greater were required to tune a full 1 GHz in frequency. Kruse Stock Electronics, under contract to GSFC, produced the first linearized, solid-state, octave-bandwidth, voltage-controlled oscillator that was tunable over a 1-GHz range. This oscillator maintained near-linear transfer characteristics over its entire range. Worst-case slope deviations were approximately 5 percent. However, it was not without its drawbacks. Because of the oscillator sensitivity (25 MHz/V to 100 MHz/V), noise and 60-cycle hum frequency modulated the carrier so that it was difficult to maintain receiver lock. In addition, the modulation bandwidth varied over the tuning range and was the determining factor for the loop bandwidth. Continued efforts to improve the Kruse Stock oscillators produced the present carrier-loop VCO. To reduce incidental frequency modulation of the carrier VCO signal, circuits were redesigned so that the oscillator could be operated entirely on batteries. An active loop filter utilizing new low-noise operational amplifiers further reduced incidental FM of the carrier-loop VCO. Modulation bandwidth compensating networks were used to compensate for bandwidth narrowing of the oscillator over its tuning range, so that the effect of narrowing on the noise bandwidth of the tracking loop would be minimized.

LABORATORY TEST RESULTS

Dynamic Measurements

So that the expected Doppler rates of 12 MHz/s can be accommodated, a theoretical minimum tracking bandwidth of 5 kHz is required. This assumes no noise contributions from the VCO; however, preliminary laboratory tests indicated that a 20-kHz bandwidth was necessary in order to maintain a high lock-confidence level.

Since the phase-lock loop frequency response $H(j\omega)$ fully describes the noise bandwidth of the loop, two different test methods, both of which yield $H(j\omega)$, were used to determine if actual noise bandwidth agreed with a 20-kHz design goal. The phase-locked VCO was frequency modulated while the VCO output, which is $H(j\omega)$, was monitored. In addition, a carrier test signal was phase modulated and the loop phase-detector output, which is $1 - H(j\omega)$, was monitored. Each measured response was in close agreement with the corresponding theoretical response for a 20-kHz loop.

Viterbi (Reference 3) has determined that the theoretical maximum permissible rate of change of input frequency before loss of lock is

$$\Delta\dot{\omega} = \omega_n^2 , \tag{1}$$

where ω_n, the loop "natural frequency" (Reference 1), is related to the loop noise bandwidth B_L and the loop damping factor δ by the equation

$$\omega_n = \frac{2B_L}{\delta + 1/4\delta} . \tag{2}$$

For a 20-kHz loop with 0.707 damping, the theoretical maximum permissible rate of change of input frequency that the receiver can track is approximately 220 MHz/s. As an additional check on the noise bandwidth, loop input signals centered throughout a 1-GHz band were frequency modulated with a triangular voltage. The receiver maintained lock for rates that were in close agreement with the theoretical maximum.

In order to determine if the receiver would track a signal whose frequency continuously varied by approximately 1 GHz, a sweep generator capable of being swept from 2-4 GHz was heterodyned with a fixed 3-GHz microwave signal. The test signal served as the simulated Doppler signal. The modulating source was selected so that a rate of change of frequency in accordance with the loop bandwidth could be maintained. Observation of the lock indicator and phase-detector monitoring scope showed the loop to be constantly in lock, with no skipping of cycles. The level of the test signal was made to vary in order to simulate 15 dB of signal fading; the tracking loop automatic gain control (AGC) maintained an IF level tolerance of ±0.2 dB. The theoretical acquisition or lock-in range for a 20-kHz loop bandwidth is ±8 kHz. The discriminator-loop acquisition range was determined to be in close agreement with the design goal of ±350 kHz.

Threshold Measurements

Determining the unlock threshold of a phase-lock receiver is a simple matter, although variations in consecutive measurements are unavoidable because loss of lock is a random statistical phenomenon. However, variations will not exceed several decibels. To measure the threshold (phase-lock sensitivity), the receiver is first synchronized with a strong stable signal and the level is then reduced until loss of lock occurs; this level is recorded as the receiver's threshold.

Gardner (Reference 1) has shown that the signal-to-noise ratio for the loop is

$$(\mathrm{SNR})_L = \frac{P_s}{2B_L W_i}, \tag{3}$$

where

P_s = loop input signal power in watts,

B_L = one sided noise bandwidth in Hz,

W_i = loop input noise spectral density in W/Hz.

But

$$W_i = FkT_0 , \tag{4}$$

where

F = system noise figure at 290 K,

k = Boltzmann's constant = 1.38×10^{-23} J/K,

T_0 = 290 K.

By substituting Equation 4 into Equation 3 and solving for P_s, we obtain

$$P_s = 2B_L FkT_0(\mathrm{SNR})_L \ . \tag{5}$$

Assume the tracking receiver noise figure is determined by the input-signal amplifier. Argon gas and hot-cold noise figure measurements indicate an average noise figure of the receiver to be about 6 dB across the entire 1-GHz band. Charles and Lindsey (Reference 4) have shown experimentally that the transition between loop lock and unlock occurs when $(\mathrm{SNR})_L$ lies between 5 and 6 dB. For a value of $(\mathrm{SNR})_L$ equal to 6 dB, a receiver noise bandwidth equal to 20 kHz, and a receiver noise figure equal to 6 dB, the theoretical minimum receiver sensitivity is −116 dBm.

Threshold sensitivity measurements were made at several fixed signal frequencies within the 1-GHz band. As expected, readings of threshold sensitivity varied between −108 dBm and −111 dBm for repeated measurements with each signal frequency.

The authors feel that differences between theoretical estimates and measured values of the threshold sensitivity can be attributed to several factors. Equation 3 is a somewhat arbitrary definition based on a linear consideration that the mean-square loop-output phase jitter is

$$\overline{\theta^2_{no}} = \frac{1}{2(\mathrm{SNR})_L} = \frac{W_i B_L}{P_s} \ . \tag{6}$$

This relationship is valid for an rms phase jitter $(\overline{\theta^2_{no}})^{1/2}$ of less than 13 deg and a value of $(\mathrm{SNR})_L$ equal to or greater than 10 dB. Measured phase jitter was approximately 30 deg rms and a value of $(\mathrm{SNR})_L$ equal to ±6 dB was assumed in the calculation of loop threshold sensitivity. Variations in VCO sensitivity and linearity at various center frequencies within the oscillator tuning range also account for part of the discrepancy.

CONCLUSIONS

The object of this work has been to determine if conventional phase-lock receivers using standard loop components are capable of operating in a high Doppler-shift tracking situation where the limiting factor on the threshold is not the stability of the VCO. On the basis of the results obtained, it is felt that the design and fabrication of an operational system should not pose any major difficulties.

It should be noted that the wideband tracking method described in this report is by no means the only technique that will allow coherent demodulation in a high-Doppler environment. An automatic frequency control (AFC) technique has been employed by Arams (Reference 5), and a digital VCO technique has been utilized by the authors (Reference 6). Arams' application was similar to the one described in this report. The digital VCO technique is applicable to a tracking situation such as a deep-space probe for which a high Doppler shift can be expected at small Doppler rates.

ACKNOWLEDGMENT

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Goddard Space Flight Center
National Aeronautics and Space Administration
Greenbelt, Md., April 9, 1971
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