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Short Papers

The Aerospace Low-Noise Millimeter-Wave Spectral Line Receiver

WILLIAM J. WILSON

Abstract—**A multichannel millimeter-wave receiver has been designed and built to study narrow-band signals from the natural background. This receiver is used with the Aerospace 4.6-m millimeter-wave antenna at El Segundo, CA. It is a superheterodyne mixer receiver which will tune over the 70–120-GHz frequency range with a system noise temperature of ~500 K (DSB).**

The spectral line receiver includes the RF system, the IF system, the local oscillator (LO) phase-locking system, the multichannel filter receivers, and the data processing system. This short paper describes the design and operation of the spectral line receiver and provides a sample of the results obtained.

I. Introduction

A multichannel millimeter-wave receiver has been designed and built to study narrow-band signals from molecules in the interstellar medium and in planetary (including the earth's) atmospheres. These weak noise signals are concentrated in bandwidths of 100 kHz–10 MHz with received antenna temperatures ranging from 10^{-3} to 10^{2} K. This receiver is used with the Aerospace 4.6-m millimeter-wave radio telescope located in El Segundo, CA [1]. The receiver is a superheterodyne type which will tune over the 70–120-GHz frequency range with a system noise temperature of ~500 K (DSB), making it one of the world's lowest noise room-temperature receivers at millimeter wavelengths.

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The spectral line receiver includes the RF system, the IF system, the local oscillator (LO) phase-locking system, the multichannel filter receivers, and a computer-controlled data processing system. A block diagram of the receiver is shown in Fig. 1 and a picture of the control room in Fig. 2.

The ~100-GHz signal is received by a feed horn located at the antenna's Cassegrain focus, where it passes through the LO injection filter to the room-temperature single-ended mixer. The mixer downconverts the signal to 1.38 GHz, where it is amplified by an uncooled parametric amplifier which has a noise temperature of ~45 K and a bandwidth of 100 MHz. The signal is further amplified by transistor amplifiers and then sent to the receiver control room. In the control room, the signal is filtered, amplified, and converted to 150 MHz, and distributed to the filter receivers and a total power/synchronous detector receiver. All receiver tuning adjustments are servo controlled at the radiometer control panel. The millimeter-wave local oscillator klystron (located in the antenna equipment box) is phase locked to a ~100-MHz frequency synthesizer (in the control room), which sets the precise operating frequency. Using this phase-locking system, the klystron can be switched over a 100-MHz frequency range at a 10-Hz rate. The LO signal is sent to the mixer through a LO injection filter, which serves to minimize signal and LO losses and also to filter the LO signal.

Two multichannel filter receivers are used to analyze the received signal. A 64-channel 1-MHz bandwidth-per-channel filter receiver is used to analyze wide bandwidth signals, and a 128-channel 250-kHz bandwidth-per-channel receiver is used to obtain narrow frequency resolution of the incoming signals. The outputs of the filter banks are sent to a 192-channel multiplexer

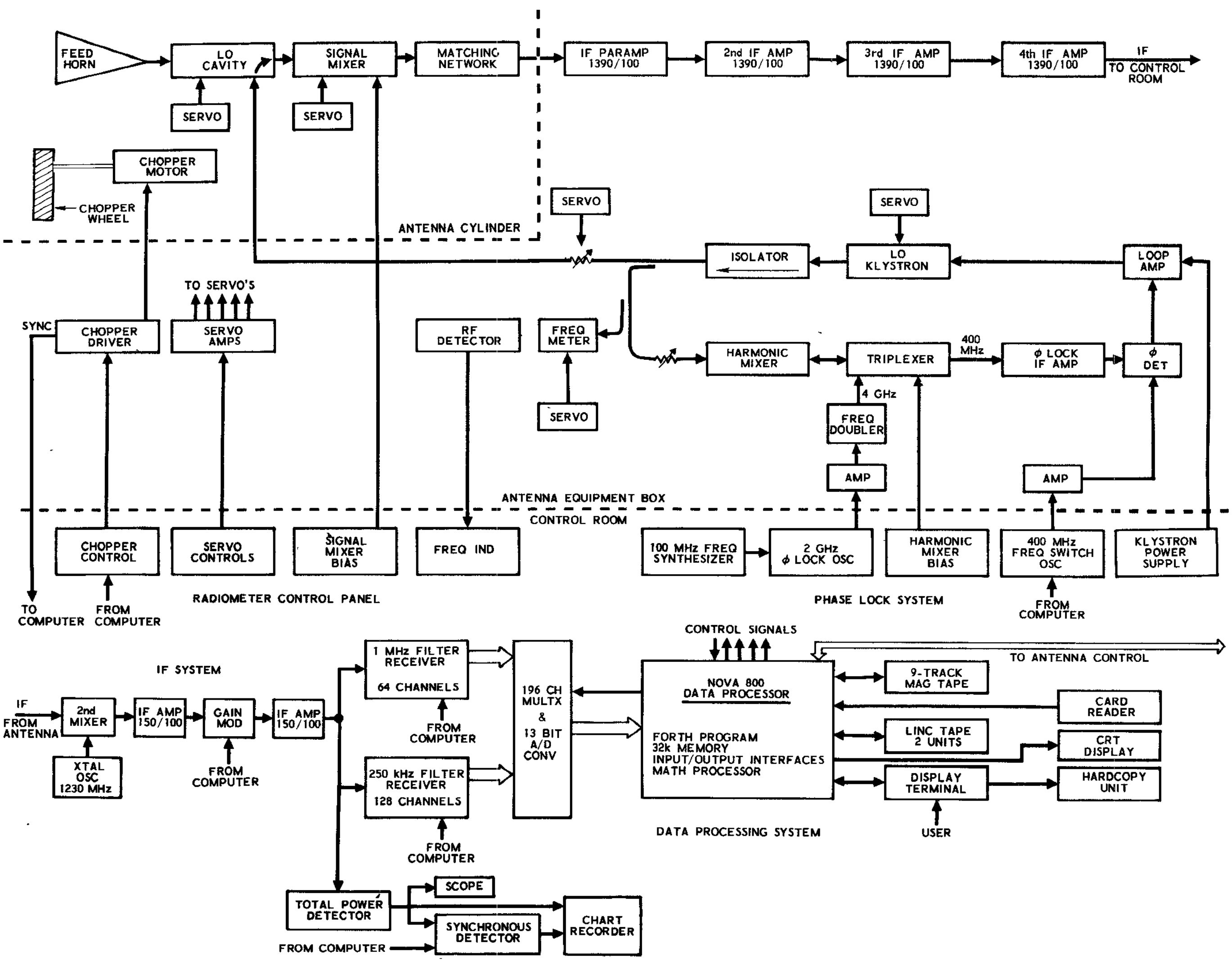

Fig. 1. Block diagram of the Aerospace low-noise millimeter-wave spectralline receiver. This receiver has a tuning range of 70–120 GHz and a system noise temperature ~500 K (DSB).

and then to a 13-bit A/D converter. These digital data are read by a NOVA 800 minicomputer which also controls receiver operations, antenna positioning, and data processing and display. The computer is programmed in the FORTH language, which provides a flexible and easy-to-use real-time system.

In the following sections each subsystem will be described and samples of the results will be presented.

II. RF System

The RF system consists of the feed horn, LO injection cavity, and the signal mixer and matching network as shown in Fig. 1. The goal of this system is to minimize the signal loss into the mixer to produce the lowest noise figure receiver possible. A chopper-wheel system with an ambient load is used for system tuning, calibration, and determination of atmospheric attenuation.

The heart of this low-noise millimeter-wave receiver is the signal mixer which was built at Aerospace and is a single-ended mixer using a GaAs diode mounted in one-fourth-height WR-10 waveguide. The mixer design was based on a design by Kerr [2] and consists of three main parts: 1) a stepped waveguide transformer designed to have a low reflection over the 80–120-GHz frequency range; 2) the one-fourth-height diode mount with a stycast mounted IF choke; and 3) the rectangular backshort with micrometer drive. The assembled mixer is shown in Fig. 3 and a cross section is shown in Fig. 4. The GaAs Schottky-barrier diodes were made by Dr. R. Mattauch at the University of Virginia and have ~2-μm-diam contacts with series resistance of ~10 Ω.

At 90 GHz, the best mixer was measured to have a conversion loss ~5.4 dB and a mixer noise temperature of ~600 K. The measured total system noise was ~500 K (DSB). At 115 GHz, corresponding figures are a conversion loss ~6.2 dB and a mixer noise temperature of ~700 K; total system noise temperature was ~600 K. All measurements were made using thermal sources and radiometric techniques. The mixer IF impedance at 1390 MHz was ~150 Ω.

III. IF System

The RF input signal is downconverted in the signal mixer to the first IF center frequency of 1380 MHz. This mixer output signal passes through a quarter-wave matching network and then through a low-loss semirigid cable (~0.5 m) to the IF system assembly located in the antenna equipment box. The first IF amplifier is a Micromega nondegenerate parametric amplifier. This device has a gain of ~17 dB, a noise temperature of ~45 K,

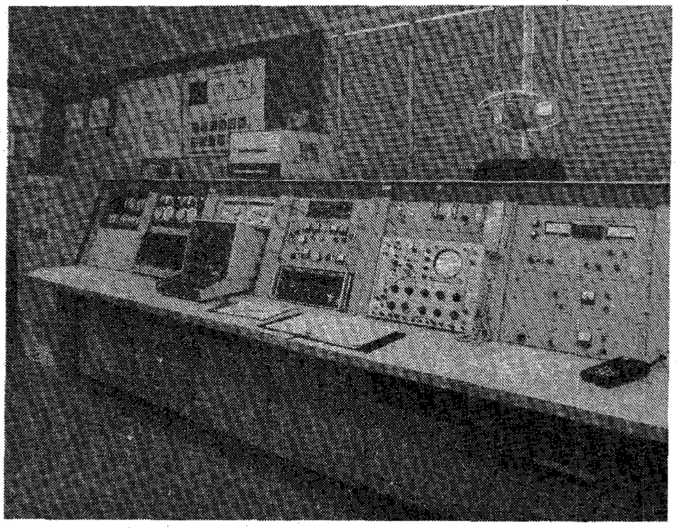

Fig. 2. The control room for the Aerospace spectral line receiver showing the radiometer control panel, the LO klystron phase-locking system, the multichannel filter receivers, and the computer control and data acquisition system.

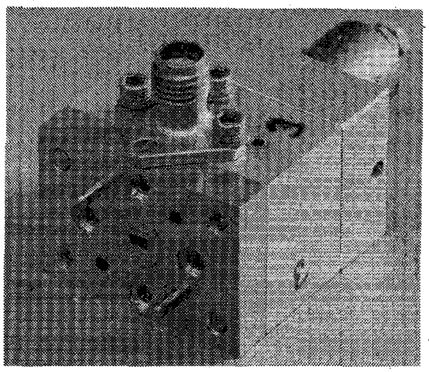

Fig. 3. The 80–120-GHz millimeter-wave mixer.

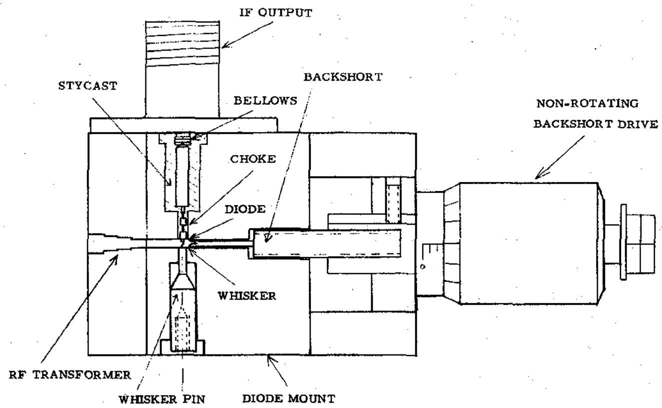

Fig. 4. Cross section of the 80–120-GHz millimeter-wave mixer.

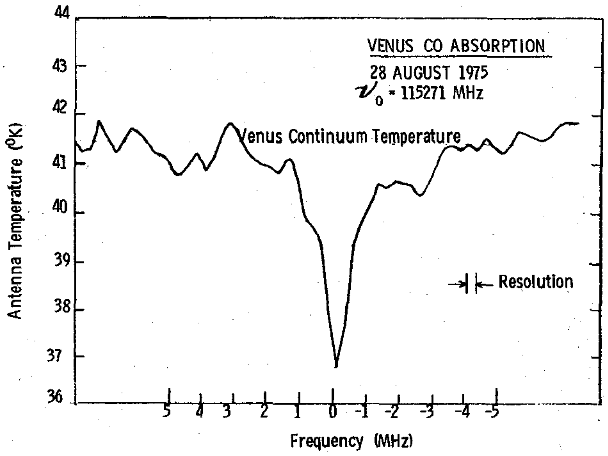

Fig. 5. Absorption spectrum of carbon monoxide in the upper atmosphere of Venus measured with the Aerospace spectral line receiver and 4.6-m radio telescope.

and a 1-dB bandpass of ~100 MHz centered at a frequency of 1390 MHz. Gain flatness over the central 80 MHz of bandpass is ±0.6 dB when the Gunn pump oscillator is maintained within about ±1°C of its normal operating temperature. To minimize gain variations, the paramp and other IF components are housed in a separate enclosure with an active temperature controller.

Following the paramp is a transistor amplifier which has a gain of 28 dB, a noise temperature of ~200 K, and a bandpass of 200 MHz centered at 1390 MHz. The next amplification of the IF signal is with transistor amplifiers, which produce 54 dB of gain over a 1000–1500-MHz bandpass with a noise temperature of ~300 K. This signal is then sent to the control room via coax cable.

In the control room the IF signal is filtered and mixed with the output of a 1230-MHz crystal-controlled oscillator to yield an output signal centered at the second IF frequency of 150 MHz. The second LO at 1230 MHz was chosen to center the first IF band at 1380 MHz since paramp gain variations are less pronounced at the low-frequency end of the passband.

This downconverted signal is attenuated to the desired level, then filtered and amplified and passed through a gain modulator. The signal is then further amplified and is sent to the multichannel filter banks. A portion of the signal is sent to the IF spectrum monitor, and to the total power and synchronous detectors.

IV. LO Phase-Locking System

To accurately control the frequency of the receiver, the millimeter-wave LO klystron is phase locked to a 100-MHz frequency synthesizer located in the control room. The basic design of the phase-locking system is similar to one designed by Weinreb of the National Radio Astronomy Observatory [3], and the block diagram is shown in Fig. 1. The output of the synthesizer is sent to a phase-locked oscillator which locks onto a harmonic of the synthesizer, to produce an output in the range 1.7–2.0 GHz. This 2-GHz output is sent to the antenna equipment box where it is amplified and doubled to the 3.4–4.0-GHz frequency range. This 4-GHz output is then applied to a millimeter-wave harmonic mixer where it is multiplied up to ~100 GHz. The multiplied output is then mixed with the ~100-GHz klystron output to produce a difference frequency near 400 MHz. This difference frequency is coupled from the harmonic mixer, filtered, and amplified in the phase-lock IF, and then applied to a phase detector. Another output near 400 MHz, from an oscillator in the control room, is also applied to the phase detector, and the low-frequency output from the phase detector is sent to the loop amplifier. The phase-lock loop amplifier filters and amplifies the phase detector output producing a control voltage of ±50 V, which is added to the klystron reflector voltage to control the phase (and thus frequency) of the klystron. Frequency switching is accomplished by using a computer-controlled solid-state switch to alternate between two 400-MHz oscillators located in the control room. A search signal is provided to the loop amplifier to aid in the phase locking.

V. Multichannel Filter Receivers

The multichannel filter receivers measure the relative spectral power distribution of the incoming signal. Spectral resolution is achieved by using an array of adjacent bandpass filters to cover a section of the IF signal bandpass. The input signal is filtered, detected, amplified, and integrated for 45 ms. At the end of this integration period, the output of each integrator is digitized with a multiplexer-A/D converter and stored in the NOVA 800 computer. When the computer finishes storing the data, the computer sends a reset pulse to each integrator to begin another integration period. The Aerospace spectral line system provides for simultaneous use of two multichannel filter spectrometers, the first having 64 channels with 1-MHz filter resolution, the other having 128 filters each 250 kHz wide.

VI. Data Processing and Control System

The data acquisition and processing is performed by a NOVA 800 minicomputer which has 32 768 16-bit words of memory. Computer peripherals include two LINC magnetic tape units, a CRT display terminal with hard copy output, a card reader, a 9-track magnetic tape unit, and a CRT display as shown in Fig. 1. The computer controls the observations by providing control signals to the receiver and position information to the antenna servos. The computer reads the multichannel filter data from the A/D converter and performs the synchronous detection and calibration of the data in real time. At the end of each observation the averaged spectrum is written on magnetic tape for later processing. The computer is programmed in the FORTH language which provides a flexible and easy to use system. The operator can display the received spectra and can also do a significant amount of data reduction during the observations.

VII. Results

The Aerospace spectral line receiver has been operating since December 1974, and many observations of interstellar molecules and planetary atmospheric molecules have been made. A sample spectrum of the 2.6-mm absorption line due to carbon monoxide in the upper Venus atmosphere is shown in Fig. 5 [4]. This result was the first measurement of a microwave planetary spectral line and introduces a new method for studying the upper atmospheres of the planets. Many outside scientists are also using the facility for research under a program partially supported by the National Science Foundation.

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Error Minimization in Network Analyzer Measurement of Varactor Quality

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***Abstract*—Errors are minimized in the four-bias measurement of varactor quality by a holder designed so that $R_s/Z_0 \approx 1.5/\sqrt{Q}$. The optimum intermediate biases depend upon the expected error in reflection coefficient measurements.**

Introduction

The measurement of varactor cutoff frequency [1] or dynamic quality factor [2] has been both useful and difficult for the low-loss varactors used in parametric amplifiers. Theoretical criteria for optimum holders for such measurements are given here, along with principles and suggestions for the conduct of measurements.

Low-noise amplification enhances the value of expensive communication and radar systems, so that severe selection of a premium fraction out of a varactor population is economically advantageous, even if many sound diodes are discarded. Quality measurements conducted by tedious slotted-line measurements and graphical data reduction have thus been justified. A quality measurement is usually more reproducible and informative than circuit tests in parametric amplifiers. It is possible to design holders for quality measurement that permit relatively easy and nondestructive insertion of diodes. Such easy insertion is not always the case for amplifiers designed for the ultimate in low-noise broad-band performance. The advantages of a precise initial selection are especially apparent when the amplifier requires matched pairs of diodes.

Single-frequency measurement techniques for evaluating parametric amplifier varactors [2]–[5] have the advantage that the quality factor can be deduced without an explicit model of the parasitic reactances of the diode cartridge and test holder. The four-bias single-frequency procedure to be analyzed here has been satisfactorily implemented on a computer-controlled automatic network analyzer.

Transmission measurements on varactor diodes mounted in shunt with a transmission line have been used to evaluate varactor quality, by single-frequency [6] and swept-frequency [7] methods. In both cases, the accuracy of the result is dependent upon correct modeling of the test holder (including diode cartridge) and the possible errors are not limited in either direction. In fact, some notorious overestimates of varactor quality have resulted from misapplication of these techniques, in that the quality factor of passive metallic capacitance was averaged in with the quality of the varactor.

Measurements based on the bandwidth and pulling range of high-Q varactor-loaded cavities [8] are relatively insensitive to the assumed circuit model and seem dependable for rapid approximate evaluation of varactor quality through visual inspection of resonance curves. However, for present automatic network analyzers, the determination of bandwidth is an iterative process [9] slowed by the settling time for frequency changing.

The equivalent circuit of the varactor contains frequency-dependent loss elements because of skin effect. Therefore, the

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