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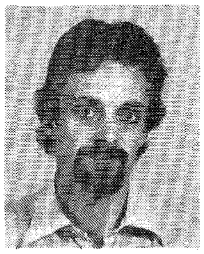

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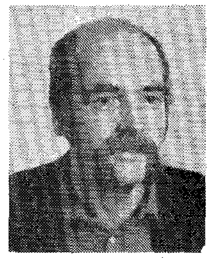

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V-Band Low-Noise Integrated Circuit Receiver

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***Abstract*—A compact low-noise *V*-band integrated circuit receiver has been developed for space communication systems. The receiver accepts an RF input of 60–63 GHz and generates an IF output of 3–6 GHz. A Gunn oscillator at 57 GHz is phaselocked to a low-frequency reference source to achieve high stability and low FM noise. The receiver has an overall single sideband noise figure of less than 10.5 dB and an RF to IF gain of 40 dB over a 3-GHz RF bandwidth. All RF circuits are fabricated in integrated circuits on a Duroid substrate.**

I. Introduction

THE RAPID ACCELERATION of millimeter-wave activities has led to a strong demand for low-noise integrated circuit receivers for both civil and military applications. These include communications, instrumentation, radiometry, radars, missile seekers, and electronic warfare. Conventional millimeter-wave receiver technology is well established based on waveguide components. Integrated circuit technology, on the other hand, has been trailing waveguide technology because of the absence of good beam-lead diodes, the difficulty in achieving low-loss and wide bandwidth, and the radiation loss involved in mounting active devices in integrated circuit media. Resolving these problems is important because integrated circuit receivers provide significant advantages.

a) Beam-lead mixer diodes eliminate the need for mechanical contacts; they are mechanically rugged and able to withstand high vibration and shock.

b) Planar circuits can be precisely controlled through high resolution processing techniques to achieve a high degree of reproducibility.

c) When a design is established, the complicated circuit can be fabricated at low cost through the use of photolithographic methods.

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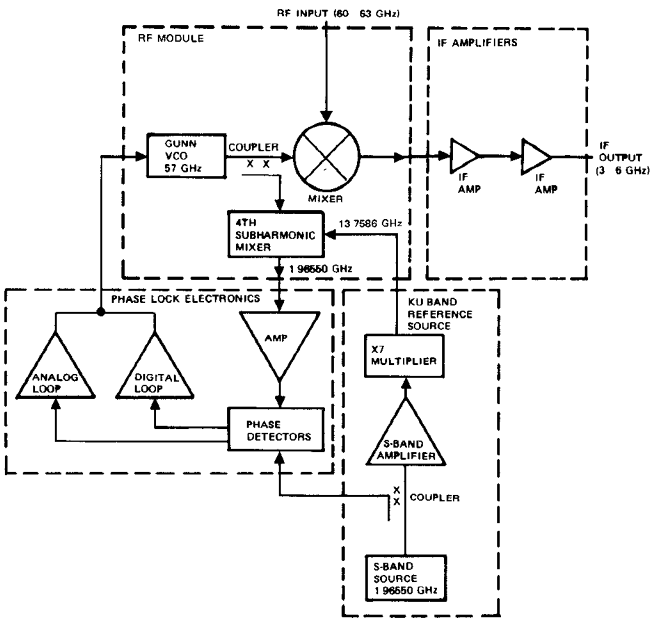


Fig. 1. Receiver functional block diagram.

d) Planar circuits are easy to assemble and thus have enhanced reliability and good module-to-module repeatability.

e) Planar circuit techniques allow a tremendous size and weight reduction as compared to conventional components using waveguides.

f) Planar circuit designs can eventually be translated into monolithic circuits to further reduce the size and weight and improve performance.

Because of these advantages, several planar receivers have been built in the lower millimeter-wave frequency range in finline [1], microstrip [2]–[4], and dielectric waveguide [5]. This paper describes a *V*-band integrated circuit receiver system (Fig. 1) using microstrip and suspended stripline as the transmission media.

This receiver accepts an RF input of 60–63 GHz and generates an IF output of 3–6 GHz. A Gunn VCO is phaselocked to a low-frequency stable reference source to achieve high stability and low FM noise. State-of-the-art integrated circuit components have been fabricated as building blocks of the receiver. The receiver has an overall single sideband noise figure of less than 10.5 dB (including 4-dB IF amplifier noise figure) and an RF to IF gain of 40 dB over a 3-GHz RF bandwidth. The receiver design can be scaled to *W*-band and should have many applications in high data rate space communication systems.

II. Phaselocked Gunn VCO Development

To achieve high stability and low FM noise, the Gunn VCO must be phaselocked to a low-frequency stable reference source. The phaselocked Gunn VCO circuitry consists of an MIC Gunn VCO, subharmonic mixer, *Ku*-band reference source, and phaselocked electronics.

A. MIC Gunn VCO

It is well known for a negative resistance oscillator that oscillation can occur whenever the real part of the Gunn

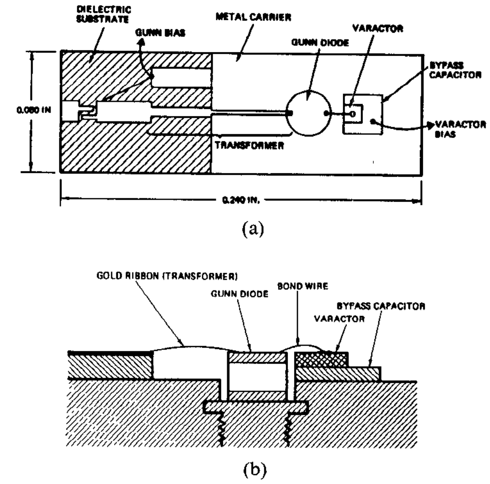
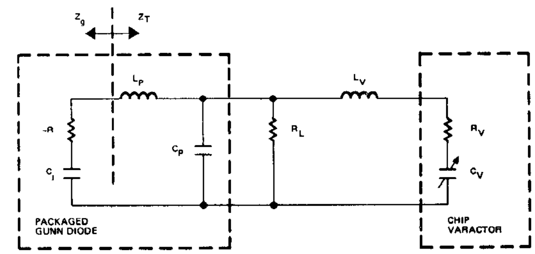


Fig. 2. *V*-band MIC Gunn VCO. (a) Top view. (b) Cross section.



WHERE $-R$ IS THE NEGATIVE RESISTANCE OF GUNN DIODE
C_j IS THE JUNCTION CAPACITANCE OF GUNN DIODE
L_p IS THE PACKAGE INDUCTANCE OF GUNN DIODE ($\approx$0.15 nH)
C_p IS THE PACKAGE CAPACITANCE OF GUNN DIODE ($\approx$0 12 pF)
R_L IS THE TRANSFORMED LOAD IMPEDANCE
L_V IS THE BONDING WIRE INDUCTANCE
R_V IS THE SERIES RESISTANCE OF VARACTOR
C_V IS THE CAPACITANCE OF VARACTOR

Fig. 3. Gunn VCO equivalent circuit.

device impedance is greater than the external load resistance and the imaginary part of the circuit impedance is a conjugate match to that of the device. Aside from the basic frequency range over which the Gunn device may be operated, the tuning range of a varactor-tuned Gunn oscillator is also governed by the circuit configuration, the varactor capacitance ratio, and circuit parasitics. To minimize circuit parasitics, chip varactors should be employed.

The Gunn VCO circuit is similar to those reported in the lower frequencies [6]–[7]. As shown in Fig. 2, the varactor chip was mounted in shunt with the Gunn diode. A two-section transformer was used to accomplish impedance matching. The equivalent circuit of this Gunn VCO is shown in Fig. 3.

With the varactor chip disconnected, Z_g and Z_T can be given by

$$Z_g = -R - j\frac{1}{\omega C_j} \tag{1}$$

$$Z_T = \frac{R_L}{1+\omega^2 C_p^2 R_L^2} - j\frac{\omega}{1+\omega^2 C_p^2 R_L^2}\left[L_p - R_L^2 C_p + \omega^2 C_p^2 R_L^2 L_p\right]. \tag{2}$$

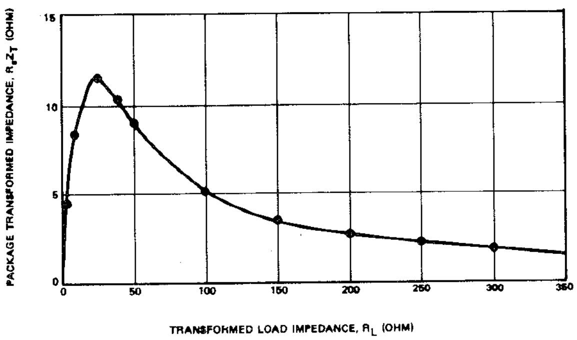


Fig. 4. Transformed impedance versus transformed load impedance.

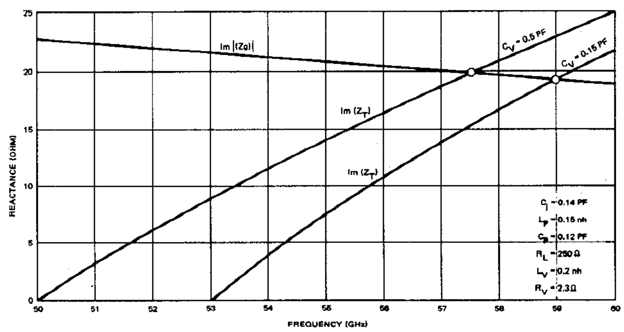


Fig. 5. Theoretical VCO tuning range.

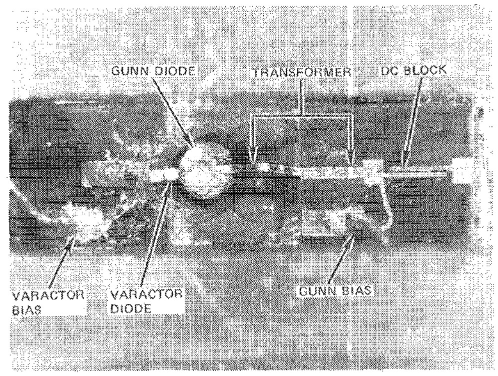


Fig. 6. V-band MIC Gunn Oscillator with transition.

The oscillation occurs in the following conditions:

$$I_m(Z_g) = -I_m(Z_T) \tag{3}$$

$$|R_e(Z_g)| > R_e(Z_T). \tag{4}$$

Since R is generally small (around 8 Ω in this case), the package transformed impedance (R_eZ_T) must be smaller than R to satisfy (4). The package transformed impedance (R_eZ_T) as a function of transformed load impedance (R_L) is shown in Fig. 4. It is seen that a transformed load impedance of 250 Ω can be transformed through the package parasitics into a small value of less than 3 Ω. After the transformed load impedance of 250 Ω has been selected, a two-section transformer, as shown in Fig. 2, was designed to match this 250-Ω impedance level to the 50-Ω output microstrip line.

By observing the oscillation frequency with the varactor disconnected, C_j can be calculated by inserting (1) and (2) into (3), and is given by

$$C_j = \frac{1+\omega^2 R_L^2 C_p^2}{\omega^2\left(L_p - R_L^2 C_p + \omega^2 C_p^2 R_L^2 L_p\right)}. \tag{5}$$

Using this C_j and the equivalent circuit of Fig. 3, we can calculate Z_g and Z_T

$$Z_g = -R - j\frac{1}{\omega C_j} \tag{6}$$

$$Z_T = j\omega L_p + \frac{R_L\left[R_V + j\left(\omega L_V - \dfrac{1}{\omega C_V}\right)\right]}{R_L + R_V - \omega^2 R_L C_p L_V + \dfrac{R_L C_p}{C_V} + j\left(\omega L_V - \dfrac{1}{\omega L_V} + \omega R_L R_V C_p\right)}. \tag{7}$$

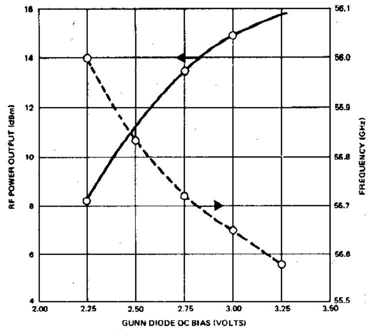
Using (6) and (7), the varactor tuning range can be computed by solving (3) for different values of C_V. A theoretical tuning range of 1.5 GHz is possible at a center frequency of 58.25 GHz, as shown in Fig. 5. This tuning range is more than sufficient for the phaselock application.

A V-band MIC Gunn VCO operating at 57 GHz was developed, based on the theoretical analysis described above using GaAs Gunn diodes. The circuit was built on 5-mil Duroid 5880 material. This material was chosen for its low-loss characteristics and ease of fabrication. Using the equivalent circuit configuration of Fig. 3, the circuit layout shown in Fig. 6 was fabricated. A 5-mil wide gold ribbon was connected between the Gunn diode and the microstrip circuit. The Gunn diode used was a packaged device with a diameter of 30 mils. The tuning varactor was a chip device. The performance of this Gunn VCO is summarized in Fig. 7. An optimum power output of 37.1 mW was achieved at 56.08 GHz. A bias tuning range of 400 MHz was achieved. With the varactor incorporated, the electronic tuning range from the varactor was observed to be 1.1 GHz with 10-mW output power. The power variation over the tuning range is less than 1 dB. This tuning range agrees very closely with the theoretical prediction shown in Fig. 5. These results represent the highest operating frequency to date from an MIC Gunn VCO.

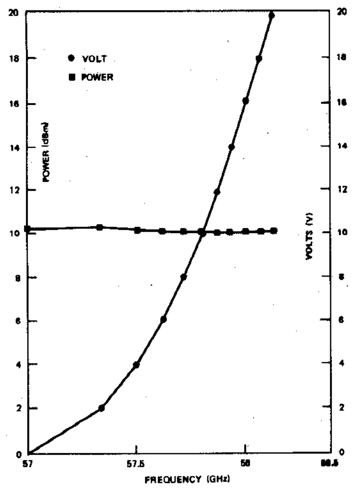
B. MIC Subharmonic Mixer

A subharmonic mixer offers a simple and direct way of downconverting millimeter waves with a low-frequency local oscillator. Downconversion is accomplished by mixing the millimeter-wave signal with the appropriate harmonic of the LO generated in the mixer itself. This eliminates the need for multiplier chains. The subharmonic mixers are particularly well suited for phaselocked loop applications where millimeter-wave sources need to be referred to a low-frequency crystal controlled oscillator.

In addition to being able to use a low LO frequency, the subharmonic mixer has the advantage of suppressing the



(a)



(b)

Fig. 7. Performance of Gunn VCO. (a) Bias tuning. (b) Varactor tuning.

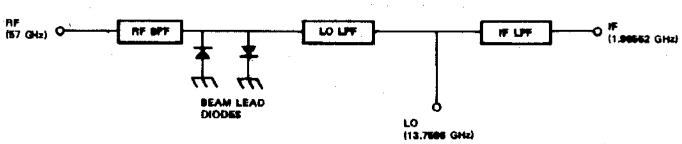


Fig. 8. Subharmonic mixer schematic.

fundamental and other odd harmonic mixing products as well as the even harmonics of the LO. The suppression of these products depends on the balance of the two diodes.

A basic schematic diagram of a subharmonic mixer is shown in Fig. 8. In this circuit, the RF bandpass filter is used to suppress all the higher order mixing products and reject the LO and IF frequencies. An LO low-pass filter is used to transmit the LO frequency while preventing the RF from reaching the LO port. An IF filter is used as a low-pass filter to extract the IF while rejecting the LO and any residual unwanted signals from reaching the IF output port. As with the Gunn VCO, the subharmonic mixer circuit was realized using 5-mil thick Duroid 5880 material as shown in Fig. 9. Two beam-lead diodes are used in an antiparallel configuration.

The RF bandpass filter is not critical, and it is simply a quarterwave line which gives sufficient rejection at the LO and IF frequencies. Since the diodes are in an antiparallel configuration, no dc return path is needed. With minimum circuit optimization, a conversion loss of 20 dB at 60 GHz

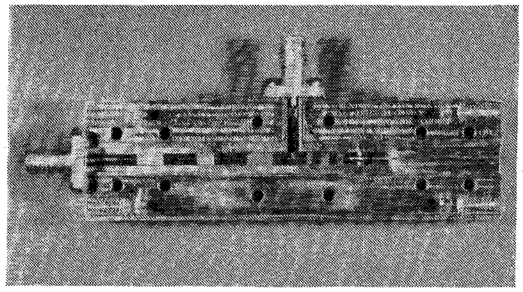


Fig. 9. *V*-band 4th subharmonic mixer and its test fixture.

was achieved. This conversion efficiency is adequate for proper phaselocked operation.

C. *Ku*-Band Source

The *Ku*-band source consists of two major components: 1) *S*-band reference oscillator, and 2) $\times 7$ frequency multiplier (with *S*-band input amplifier). The *S*-band reference oscillator is a crystal-referenced phaselocked source. It provides a stable 1.9655-GHz signal to drive the $\times 7$ oscillator. The performance of this unit is:

Frequency	1.9655 GHz
Power output	+21.4 dBm
Spurious rejection	> 70 dB relative to the carrier
Stability	± 30 ppm (-25–$60°$C).

The $\times 7$ frequency multiplier/amplifier multiplies the *S*-band reference oscillator frequency output by a factor of 7 and provides a stable 13.7586-GHz signal to drive the subharmonic mixer. The $\times 7$ is a microstrip design. The circuit consists of a driver amplifier, input isolator, and low-pass network to provide input matching and idler paths. Multiplication is accomplished with a step-recovery diode and the output frequency is selected by a three-pole combline bandpass filter. A microstrip isolator provides isolation at the output interface. All harmonics at the output are suppressed to over 53 dB below the desired signal (13.7586 GHz) and the conversion loss is about 5 dB. The performance is summarized in Fig. 10.

D. Phaselock Electronics

The MIC Gunn oscillator is phaselocked to achieve high stability with minimal phase noise. Phaselocking consists of comparing the IF output from the 4th subharmonic mixer to a crystal frequency source. Control is obtained by driving a varactor that controls the Gunn oscillating frequency. As shown in Fig. 1, the actual phaselock electronics consist of the phase detector, digital loop, and analog loop/integrator.

The phase detector compares the phase of the downconverted RF signal with that of the crystal reference source to produce two output signals (sine and cosine) to determine the phase error without ambiguity. At the heart of the phase detector electronics are two *S*-band balanced mixers (phase comparator). The IF outputs provide the sine and cosine waveforms from the amplifier subharmonic mixer 0° and 90° outputs, respectively.

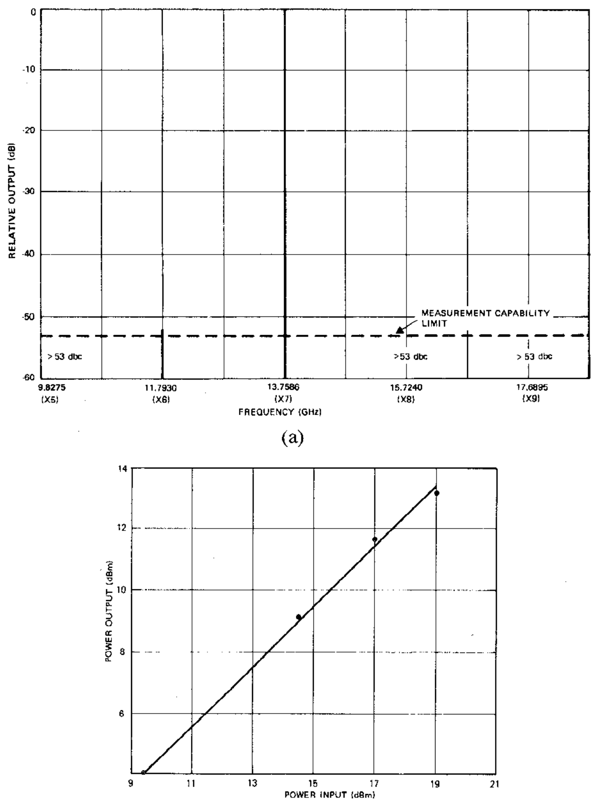


Fig. 10. Performance of $\times 7$ multiplier/amplifier. (a) Output spectrum. (b) Output power versus input power.

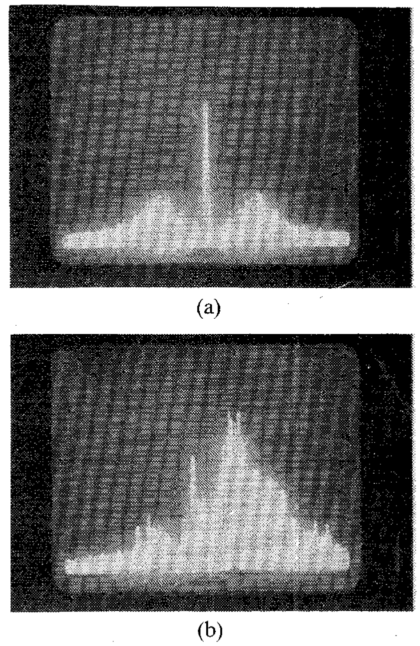


Fig. 11. IF signal from Gunn VCO (Vertical: 10 dB/div.; Horizontal: 0.5 MHz/div.; Lock is at center). (a) Using phaselock loop. (b) Using highly stable dc power supply.

The phaselock acquisition loop (or digital loop) is used during cold start to acquire the signal. In addition, large drifts in the Gunn which would normally cause the analog loop to lose lock are compensated by this acquisition loop. Thus, low stress offsets are achieved which are especially desirable since the oscillator tuning characteristics are nonlinear and temperature- and time-dependent. The digital loop provides a forcing function to allow the loop to lock up from outside its locking range. This is particularly important for a cold start. It also provides a sense mechanism to determine when a critical loop stress level is exceeded, and what is the polarity of the error. This is useful in tracking drifts due to oscillator aging or change in its thermal environment.

The purpose of the analog is to maintain phaselock with the desired loop bandwidth and damping. The phase error signal is converted to a dc voltage that controls the varactor tuning of the Gunn oscillator, thus providing the desired frequency correction.

In this application, the digital input signal is amplified and inverted to sweep the entire negative voltage VCO range. During acquisition, the high-frequency input from the analog channel has little effect on the loop and the digital channel controls the VCO. Upon acquisition, the fixed dc level of the digital input is essentially eliminated by a 100-K resistor. The VCO is now steered by the analog loop.

The optimum loop bandwidth of the control loop depends on the relation of the free-running Gunn phase noise and the phase noise of the reference oscillator. The ideal design would allow for a loop bandwidth to be equal to the crossover frequency. Simultaneously, two other criteria would have to apply.

1) Gain and phase margin must be adequate at the crossover frequency.

2) The loop group delay (as measured from the instant the RF signal leaves the oscillator to the instant an error voltage it produces reaches the frequency controlling element) must not exceed $1/4f_{cr}$, the crossover frequency.

The bandwidth measured on the final unit is approximately ± 700 kHz. The integrator/filter design has a natural frequency of 500 kHz with an 0.7071 damping factor. A picture of the locked signal is shown in Fig. 11(a). To demonstrate the impressive noise reduction capability, the phaselock electronics were replaced by a highly stable dc power supply. The poor performance of this dc power supplied VCO input is shown for comparison in Fig. 11(b).

III. Crossbar Stripline Balanced Mixer

The circuit configuration of our crossbar stripline mixer is shown in Fig. 12. The mixer was built in suspended stripline for low loss. The RF signal is applied to the mixer diodes from a waveguide perpendicular to the circuit board. The crossbar configuration is formed by two mixer diodes with opposite polarity connected in series across the broadwall of the waveguide. The mixer diodes are thus in series with respect to the RF signal and in parallel with the IF circuit. The IF signal is extracted via a low-pass filter and the LO signal is injected from the other side through a broadside coupler.

A. Broadside Coupler Development

The RF power propagating down a suspended stripline couples to the stripline on the other side of the substrate

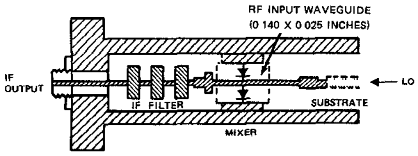


Fig. 12. Crossbar stripline mixer.

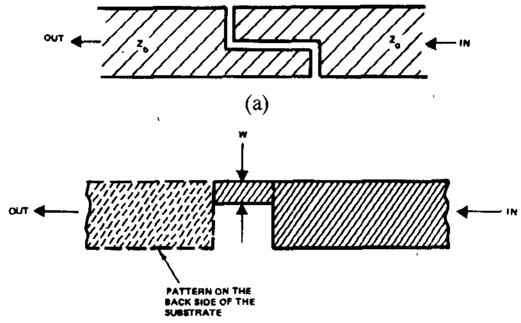


Fig. 13. Broadside couplers.

through a broadside coupler. The broadside coupler serves as a dc block and presents an open circuit to the IF frequencies.

At lower frequencies, the conventional broadside coupler shown in Fig. 13(a) has been used quite successfully. At *V*-band, to achieve minimum insertion loss, it is necessary to have very tight coupling with a gap too small to realize practically. In this case, thin lines are superimposed and separated by a substrate material of low dielectric constant, with thickness of 0.005 in, as shown in Fig. 13(b). The design of this coupler is straightforward and can be found in Matthaei [8].

The length of the coupling section is about a quarter of a wavelength corrected for the compensation due to discontinuities. The insertion loss of the coupler was measured to be less than 0.2 dB.

B. Mixer Design Considerations

Theoretical studies have considered the case of an *H*-plane slab of dielectric, centrally located within the waveguide as shown in Fig. 14 [9]. The dominant waveguide mode can be either the first longitudinal section magnetic (LSM_{11} or quasi-TE_{10}) mode or the distorted TE_{01} mode, depending on the dielectric permittivity and guide dimensions. For this case, we are interested in where $h/b \leqslant 0.25$ and $\epsilon_r \leqslant 10$. The cutoff frequency of the dominant waveguide mode LSM_{11} can be approximated by [10]

$$f_c = \frac{c}{2a}\sqrt{1-\frac{h(\epsilon_r - 1)}{b\epsilon_r}} \tag{8}$$

where a is the width of channel, b is the height of channel, h is the thickness of the substrate, ϵ_r is the relative dielectric constant, and c is the velocity of light in a vacuum.

The a dimension of suspended stripline is chosen so that the first higher order mode is much higher than the frequency band of interest. In the case of a channel with cross section of 0.080×0.025 in^2 and a substrate thickness of 0.005 in, the cutoff frequency for Duroid is 70 GHz,

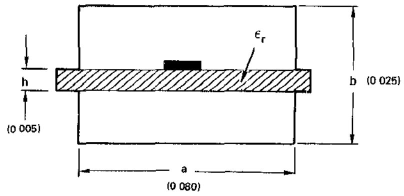


Fig. 14. Suspended stripline (all dimensions in inches).

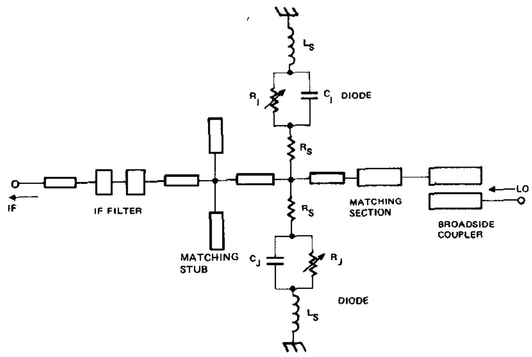


Fig. 15. Crossbar stripline mixer equivalent circuit.

which guarantees a simple quasi-TEM mode of operation at 60 GHz.

In a mixer design [11]–[13], the performance parameters of primary concern are the operating bandwidth and the conversion loss. To treat this analytically, we have developed a circuit model (Fig. 15) which deals quantitatively with the mixer performance.

In general, the conversion loss of a mixer is considered to consist of three parts:

- L_1 mismatch loss due to impedance mismatch at the RF and IF ports;
- L_2 diode parasitic loss due to its junction capacitance and series resistance; and
- L_3 intrinsic junction loss of the ideal diode.

L_1 is minimized by optimizing the RF and IF circuit matching. The junction resistance R_j of the mixer diode is varied with the LO voltage and its value can be as low as 100–150 Ω under fully turned-on conditions. Waveguide impedance is in the range of 400–600 Ω and can be matched to the diode impedance by a reduced-height taper transformer. The sliding short on the opposite side of the RF port should tune out the reactance part of the diode.

IF and LO matching are achieved by a computer analysis of the equivalent circuit shown in Fig. 15. An IF filter passes the IF frequency band and rejects the LO and RF signals. The connecting transmission line between the IF and LO ports can be optimized to provide matched conditions at the LO and IF frequencies.

A broadside coupler was designed to present an open circuit to the IF frequencies to prevent dissipation of IF power in the LO port. A double open matching stub was used to facilitate the LO matching.

The diode parasitic loss L_2 depends on the values of R_s, R_j, and C_j of the diode. The effect of series resistance and junction capacitance can be analyzed in terms of the diode

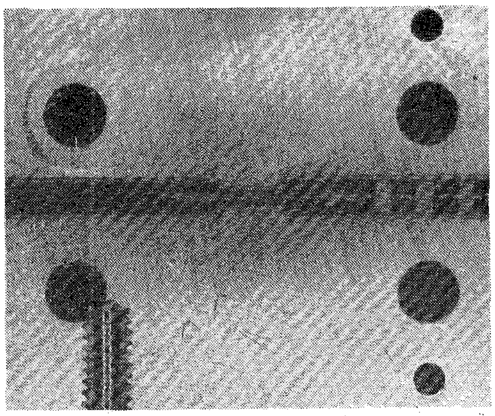


Fig. 16. Circuit layout of V-band crossbar stripline mixer.

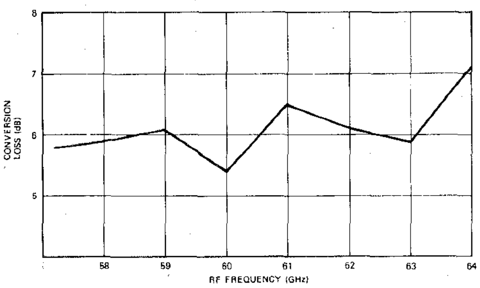


Fig. 17. Crossbar mixer performance (LO = 57 GHz).

cutoff frequency

$$f_c = \frac{1}{2\pi R_s C_j}. \tag{9}$$

As a rule of thumb, the mixer design mandates high quality diodes with a cutoff frequency at least 10 times the operating frequency. Nevertheless, the return diminishes once the limit is exceeded. Selecting the diode pair with matched characteristics will also have a pronounced effect on LO noise suppression and LO to RF isolation. Experimental data indicates that LO noise suppression on the order of 20–30 dB can be achieved for matched diode pairs. It can be shown that, with its noise suppression in this range, the contribution of LO noise to the mixer noise figure is negligible.

C. Mixer Performance

A photograph of the crossbar stripline circuit layout is shown in Fig. 16. With the LO at 57 GHz, a conversion loss of less than 6.5 dB for over 6-GHz instantaneous IF bandwidth was achieved with two beam-lead diodes as the RF is swept from 57 to 64 GHz (Fig. 17). The LO/RF isolation is over 20 dB and the LO pumping power requirement is about 10–12 dBm. The beam-lead diodes are commercially available diodes with a C_j of approximately 0.04–0.05 PF and R_s of 5–7 Ω.

The intercept point, measured in dBm, is a figure of merit for intermodulation product suppression. A higher intercept point means higher suppression of undesired intermodulation products. The intercept point measurement for our mixer is shown in Fig. 18. It can be seen that the 1-dB compression point occurs at the input power of

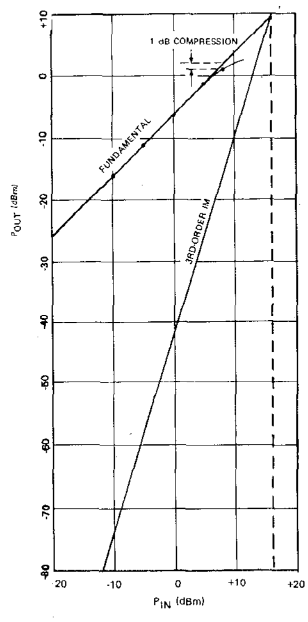


Fig. 18. Third-order intercept point measurement.

+8 dBm. The third order intercept point occurs at the input power of +16 dBm and the mixer will suppress third-order products over 55 dB with both signals at −10 dBm.

IV. IF Amplifier

Two commercially available IF amplifiers in cascade are needed to achieve high RF to IF gain. The first-stage amplifier has a noise figure of 3.5–3.9 dB over the IF frequency of 3–6 GHz with an associated gain of 19–20 dB. The second-stage amplifier has a noise figure of 4 dB and provides 25-dB gain over the 3–6-GHz frequency range. The combined gain of the cascaded amplifier is about 43–45 dB, the noise figure measurement is about 4 dB, and the third-order output intercept point is at 31 dBm.

V. Receiver Integration and Testing

The receiver consists of four modules as shown in Fig. 1: RF module, IF amplifier module, Ku-band reference source, and phaselock electronics. The modular approach facilitates testing, measurement, and integration. The module interfaces chosen provide convenient noncritical test points and allow each module to be optimized separately. All modules are integrated on an aluminum baseplate as shown in Fig. 19.

The RF module consists of a V-band crossbar stripline mixer, Gunn VCO, coupler, and subharmonic mixer integrated into a housing measuring $0.4 \times 3.6 \times 0.7$ in. The crossbar stripline mixer assembly protrudes from the housing by about 0.5 in. Details of this module are shown in Fig. 20.

The receiver was thoroughly characterized. Fig. 21 shows the measured single-sideband noise figure of the receiver as a function of RF input frequency swept from 60 to 63 GHz with the LO at 57 GHz. The noise figure varies from 10 to 10.5 dB which includes approximately 6-dB conversion loss

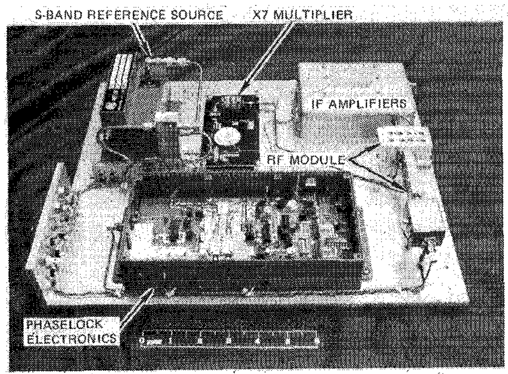


Fig. 19. Photographs of integrated circuit receiver.

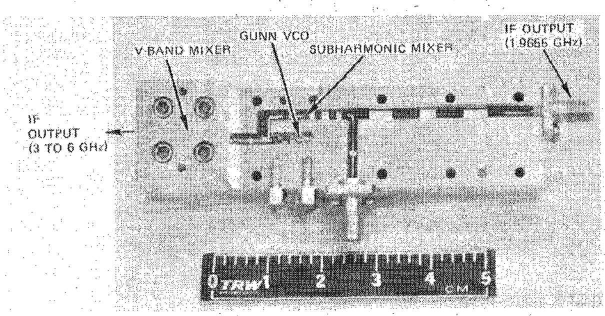


Fig. 20. Photograph of the RF module.

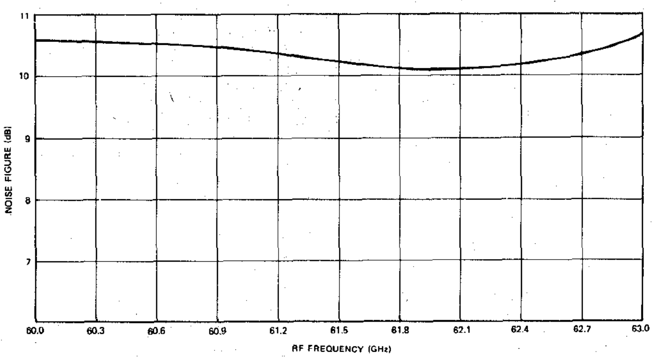


Fig. 21. Receiver single-sideband noise figure as a function of RF frequency (LO at 57 GHz).

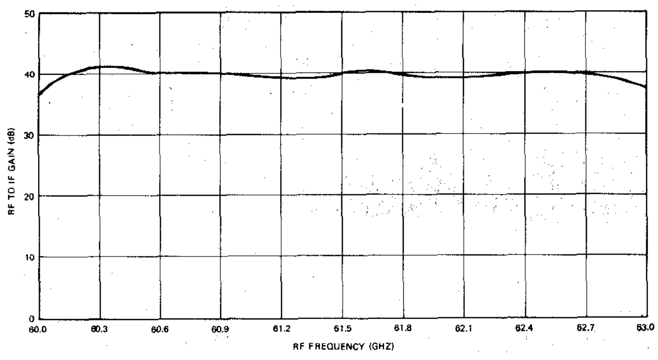


Fig. 22. Receiver RF–IF gain as a function of RF frequency (LO at 57 GHz).

from the mixer and 4-dB noise figure from the IF amplifier. Fig. 22 shows the measured RF to IF gain as a function of RF input frequency swept from 60 to 63 GHz. A gain of 40 dB was achieved and the gain flatness is 1 dB except at points near the band edges.

VI. Conclusions

A *V*-band integrated circuit receiver has been developed using microstrip and suspended stripline technology. The receiver consists of a mixer, Gunn VCO, subharmonic mixer, phaselock electronics, low-frequency reference source, and IF amplifiers. The receiver exhibits an overall single-sideband noise figure of less than 10.5 dB and RF to IF gain of 40 dB over the 3-GHz bandwidth. This performance provides a firm confirmation that integrated circuit technology has been advanced to the point where it can be applied in many system applications.

Acknowledgment

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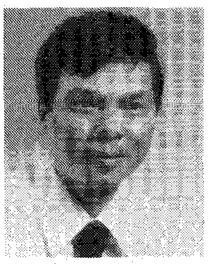

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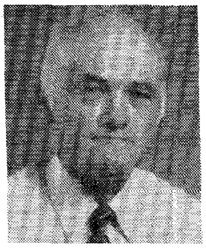

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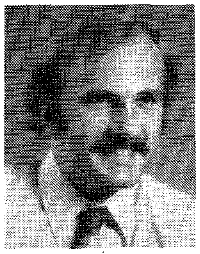

Mitchell J. Mlinar was born in 1957 in Milwaukee, WI. He received the B.S. and M.S. degrees in electrical engineering from Marquette University in 1978 and 1979, respectively. Presently he is a Ph.D. student at USC.

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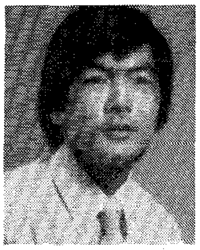

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