Low-Noise Millimeter-Wave Receivers

JAMES J. WHELEHAN, MEMBER, IEEE

Invited Paper

***Abstract*—Over the past several years, significant improvements have been made in solid-state devices (that is, avalanche diodes, Gunn diodes, varactors, mixer diodes, etc.) that have enhanced the overall capability and low-noise performance of millimeter-wave receivers. With these improved devices, it is now possible to configure completely solid-state low-noise millimeter-wave receivers. As is similar in the microwave region, low-noise parametric amplifiers, broad-band low-conversion-loss mixers, and solid-state local oscillators are now available. Furthermore, cryogenically cooled parametric amplifiers and mixers are also being developed that will result in achieving the ultimate in system sensitivity. With the flexibility offered by these completely solid-state millimeter-wave components, it is now possible to design the optimum system configuration for the intended application whether it be an advanced communication system, a sophisticated EW application, a radar system, a radiometric system, or satisfying any of the numerous receiver requirements that are being evolved.**

This paper explores the trends that are being developed in the millimeter-wave region and their application to system design. The performance criterion of various receiver systems and their sensitivity requirements are presented. A review of the system operating noise temperature concept and the method by which it can be determined and its applicability to low-noise components is demonstrated. A review of the state-of-the-art of low-noise systems and experimental data obtained in the millimeter-wave region is also presented.

I. INTRODUCTION

OVER the past several years, numerous laboratories (both military and commercial) have been investigating the potential application of the millimeter-wave spectrum to provide solutions for some very specific problem areas. As is very well known, some of the more pertinent millimeter-wave applications are as follows:

—Covert ground communications over moderate distances ($\approx$ 35 GHz);

—Secure satellite-to-satellite communication systems in the 60-GHz region;

—High data-rate communication systems;

—High-resolution radar systems for both ground and airborne applications;

—Active and passive radiometric applications.

While these applications of millimeter-waves will be very useful, the advancements made in evolving practical solutions have been relatively slow when compared to the rate of developments previously accomplished in the microwave region. One reason has been the reduction in research and development efforts over the past few years. Another has been the relatively poor predicted system performance that can be achieved based on the state-of-the-art performance of presently available components. This is particularly apparent when examining the sensitivity requirements for mission compatibility. With reasonable antenna sizes, the limited transmitter power that is presently available, and the relatively high signal attenuation over moderate distances ($\approx$ 150 dB/10 km at 35 GHz in a moderate rainfall); it becomes evident that receivers with extremely high sensitivity (low-noise figure) are urgently required. However, low-noise receivers are generally not compatible with low cost and, as a result, developments in this field have been relatively slow. This paper reviews the present state-of-the-art of millimeter-wave receivers, and indicates the needed developments for achieving high performance at minimum cost.

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The author is with AIL, a division of Cutler-Hammer, Melville, NY 11746.

II. BACKGROUND

The basic millimeter-wave component and system design have generally been extensions of the fundamental design techniques that were developed for the microwave region. An example of the extrapolation from microwaves to millimeter waves is the selection of a transmission line medium for a particular application. The transmission lines that are actively being investigated for the millimeter-wave region are the following:

—stripline;
—fin line;
—microstrip;
—image guide;
—waveguide;
—coaxial structures.

Some of these transmission lines are proven approaches when applied to designs in the microwave region. However, these same techniques when used in the millimeter-wave region demand tighter control of some of the key elements of the design (i.e., control of spurious modes, radiation, losses, tolerances, etc.) that are considered to be second- and third-order effects in the microwave region. This has produced more costly designs than previously encountered as well as poor reproducibility. The important criterion is the attainment of satisfactory system performance within reasonable economic constraints if the millimeter-wave region is to be truly exploited in the near future.

By the same token, as the microwave community extends its proven techniques and new ideas into the millimeter-wave region, developments that have been brought to fruition in the optical infrared and submillimeter regions will be looking to extend their techniques to lower frequencies. As a result, cross fertilization and competition between the two developing disciplines for application in

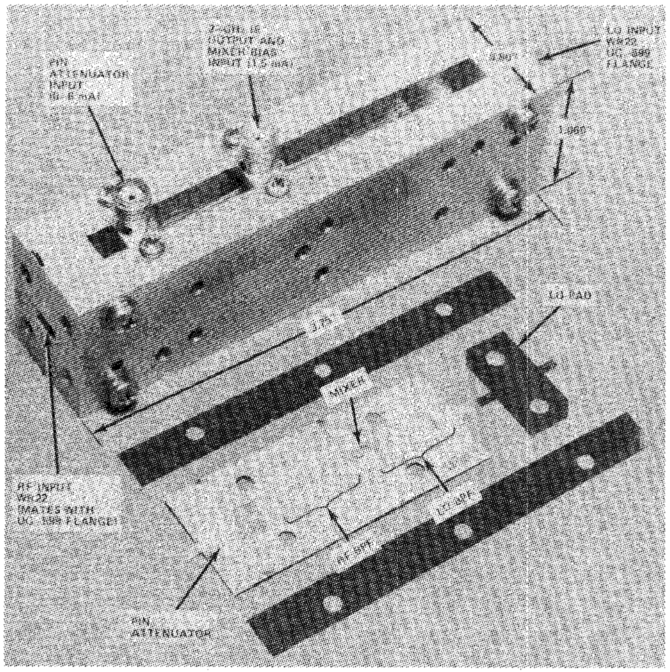

Fig. 1. Fin-line RF receiver module.

the new system designs will naturally develop. It will, however, be necessary not only to evolve viable approaches that can be readily utilized in the millimeter-wave region, but also to become familiar with other emerging disciplines to configure optimum systems approaches for the problems to be solved.

An extension of these ideas can be applied to the design of present-day low-noise millimeter-wave receivers. All of the necessary tools that are available must be optimally coupled to evolve system approaches (including cost) to solve the specific problems that will be considered in the millimeter-wave region. The designer of millimeter-wave components and systems must be flexible and be able to properly evaluate all of the viable alternatives.

Another consideration that is a natural extension of the techniques that have been developed in the microwave region is the discrete interconnection of the major building blocks (i.e., antennas, multicouplers, RF amplifiers, mixer/IF amplifiers, etc.) by standard transmission lines. This might not be the optimum approach in the millimeter-wave region for all applications. It certainly might not be cost effective. An optimum cost-effective approach for some receiver applications is the combining of the discrete components on a common substrate with noncritical tolerances. An example of such an RF receiver module at 35 GHz (using fin-line techniques) is shown in Fig. 1. This receiver contains a p-i-n-diode attenuator, an RF bandpass filter, a single-ended mixer, a local oscillator bandpass filter, and a local oscillator attenuator, all on a single substrate. This compact receiver, which is likely suited for radiometric applications, has a sensitivity that is comparable to those built with standard waveguide techniques, but has the added advantage that in quantities of 100 or more can be produced for less than $1000. As a further improvement,

RECEIVER TYPE	PERFORMANCE CRITERIA
• COMMUNICATIONS	$\frac{G_{ANT}}{T_A + T_R}$
• RADAR	$r \propto \left[\frac{P_T}{T_A + T_R}\right]^{1/4}$
• RADIOMETRIC	$\Delta T = K \frac{T_A + T_R}{\sqrt{B_{IF}\,\tau}}$

$T_{op} = T_A + T_R$

T_A = EQUIVALENT ANTENNA TEMPERATURE

T_R = RECEIVER NOISE TEMPERATURE

Fig. 2. Performance criteria.

it would be relatively simple to incorporate both a printed circuit antenna with moderate gain and a fin-line local oscillator (the development of which has already been completed) on the same substrate. The choice of fin line as the transmission medium for this system was optimum for this particular application. However, examples of receivers using a combination of transmission lines are presented later.

A discussion of some basic receiver designs and component performance in the 20–100-GHz frequency ranges, which is presently the state-of-the-art for millimeter-wave receivers, is presented. Some of the components developed at AIL and at other laboratories, as evidenced by published data, both theoretical and experimental, illustrate the applicability of the design criteria.

III. Receivers

Although it is not possible to discuss all of the varied configurations of receivers, three classifications are considered and their degrees of commonality are highlighted. The sensitivity of each of the receiver classifications is the primary area of investigation that is considered. The classifications of receivers are those that are used in communication, radar, and radiometric systems. While the transmitting and signal processing techniques for each of these receivers are different, their commonality with regard to system sensitivity (which is extremely important in the millimeter-wave region) is similar. The sensitivity of each of these receivers can be compared by a performance criterion that is directly related to their specific type of operation. Communication ground receivers used for satellite communication systems can be evaluated by the ratio of the antenna gain to the system operating noise temperature G/T_{op}. A radar system can be judged by the maximum range at which a finite target can be detected (assuming all other parameters remain constant) and is proportional to $(P_t/T_{op})^{1/4}$ where P_t is the transmitter effective power and T_{op} is the system noise temperature. For a radiometric receiver, whether it be active or passive, the important criterion is the detectability of very low-level noiselike signals. The minimum detectable signal (ΔT) from an extended target is again proportional to the system operating noise temperature (T_{op}). A summary of the system performance criteria for each of these different systems is shown in Fig. 2. It is apparent that the system

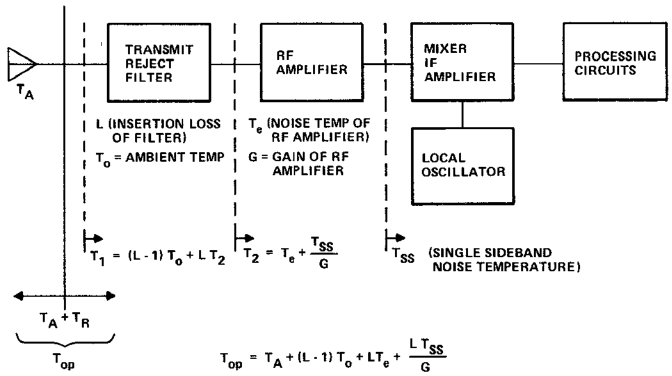

Fig. 3. Communications receiver, block diagram.

operating noise temperature (T_{op}), which is the sum of the equivalent antenna temperature (T_A) and the receiver noise temperature (T_R), must be minimized to improve the sensitivity and hence operating performance of each of the systems. The elements that contribute in determining T_{op} are considered. First, a brief review of each of the receiver systems and an evaluation of their system performance are presented.

A. Communication Systems

Numerous configurations can be evolved for communication systems depending on the overall application. This can vary from single-channel point-to-point ground communication systems to the multichannel sophisticated satellite communication ground-based receiving equipment. Since the basic concepts for any communication equipment are inherently similar, a consideration of the communication receiver used for a satellite ground terminal is considered. This is directly applicable to the millimeter-wave region since space-to-earth frequency allocations have been made for numerous frequency bands in the millimeter-wave region. A simplified block diagram of one type of communication receiver is shown in Fig. 3. When a signal is received at the antenna, it is passed through a low-loss transmit–reject filter, amplified by a low-noise RF amplifier and then downconverted to an appropriate IF by a mixer. The incoming signal is then processed with post detection circuitry.

For a satellite communication receiver, the critical down-link performance criterion is G/T_{op} as previously indicated. This can be easily seen from the following [1]:

$$C/N = \text{EIRP} + \frac{G}{T_{op}} - L_1 - L_2 - K - B$$

where

- C/N receiver carrier power-to-noise ratio;
- EIRP satellite effective radiated power;
- G/T_{op} antenna gain to operating system noise temperature;
- L_1 propagation and other down-link loss;
- L_2 atmospheric loss;
- K Boltzmann's constant $= -228.6$ dBW/K/Hz;
- B noise bandwidth.

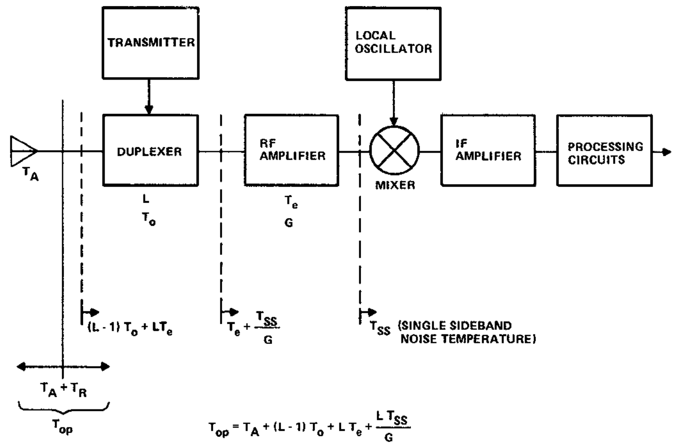

Fig. 4. Radar receiver, block diagram.

The primary element that directly affects the quality of the received signal is G/T_{op} since all of the other parameters are fixed for a specific system. Therefore, the performance criterion by which different receiver configurations can be evaluated is by comparing their G/T_{op} ratio, and hence their signal-to-noise ratio. The higher the ratio (generally expressed in decibels), the better the quality of the system. The gain of the antenna over an isotropic radiator (G) is a function of the antenna design and is one of the more pertinent system decisions that are made. However, for a particular antenna gain (G), the only means to improve G/T_{op} is to decrease the system operating noise temperature. The procedure for determining T_{op} by knowing the noise temperature of the individual components is also shown in Fig. 3.

B. Radar Systems

A block diagram of a simplified radar system is shown in Fig. 4. A duplexer is used at the input after the antenna (rather than a transmit–reject filter) for directing the high-power transmitted pulse power to the antenna and for providing protection for the receiver from both the high-energy-spike leakage and CW power leakage during the transmit period. The insertion loss of the duplexer is designed to be as low as possible in order not to degrade overall system performance. An RF amplifier is also utilized to establish the system sensitivity requirements.

For a particular set of system operating parameters, the maximum range at which a target can be detected is proportional to $(P_t/T_{op})^{1/4}$. Since it is generally difficult to significantly increase the transmitter power, particularly in the millimeter-wave region where high power is very difficult to achieve, the only way to increase the range is to decrease the system operating noise temperature. Since the antenna noise temperature is known and not readily changed, the improvement must be made in the receiver. This is accomplished by providing either a low-noise RF amplifier or a low-noise single-sideband mixer. The method by which the system noise temperature can be determined is also shown in Fig. 4.

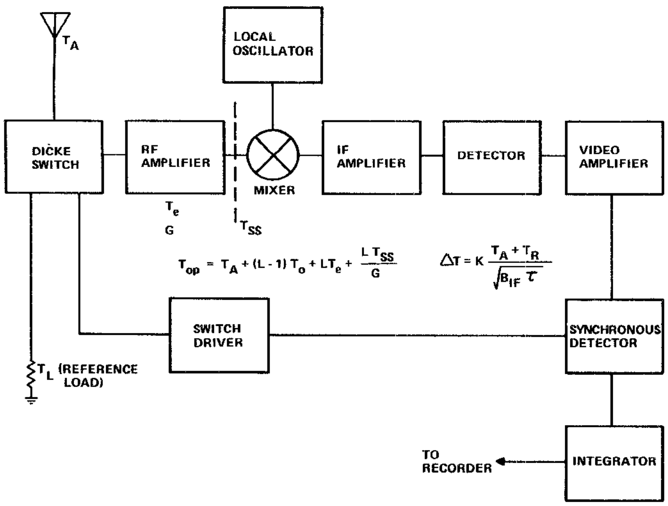

Fig. 5. Radiometric system, block diagram.

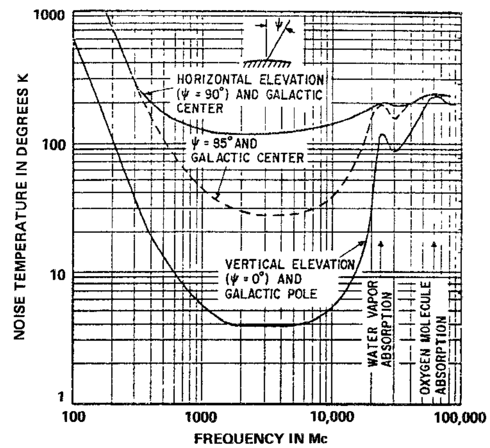

Fig. 6. Sky noise temperature as a function of frequency and elevation.

C. Radiometric Systems

A simplified block diagram of a typical switched radiometric receiver is shown in Fig. 5. A Dicke switch [2] is used at the input of the receiver after the antenna to compare the noiselike signals received from the antenna with a signal observed from a load that is connected to the third port of the Dicke switch. The RF output of the switch is amplified by an RF amplifier prior to being detected, processed, and recorded. The minimum detectable signal (ΔT) from an extended target is again proportional to the system operating noise temperature and, to improve system performance, T_{op} must be minimized. Again, this is accomplished by providing the most sensitive receiver. The determination of T and T_{op} is shown in Fig. 5.

A primary difference with radiometric systems as compared to communication and radar systems is that their operation may be either single or double sideband. A single-sideband receiver is used if the signals of interest are discrete lines (e.g., hydrogen line at 1.40 GHz). For this type of operation, single-sideband mixers and nondegenerate parametric amplifiers are the primary components used in determining receiver noise performance.

However, of the measurement of interest is the determination of the total power across a broad spectrum as received, from a source, a double-sideband system is the more appropriate choice. For this type of operation, a double-sideband mixer of a degenerate parametric amplifier that is used in conjunction with a zero-frequency IF amplifier [3] may be used to provide a lower receiver noise temperature than could be achieved with a single-sideband system. As an example, the single-sideband noise figure of a mixer/IF amplifier is 3 dB higher (minimum) than that of the same unit used for double-sideband operation. As a result, the double-sideband system would have the better sensitivity and the choice of system approach would depend on the application.

The common element that determines the system sensitivity in each of these systems is the operating noise temperature (T_{op}), which is comprised of the equivalent antenna temperature and the receiver noise temperature. The equivalent antenna temperature is determined primarily by sky noise, while the receiver noise temperature is dependent on the receiver frequency, operating bandwidth requirements, environmental restrictions, and is normally determined by the RF amplifier and/or the mixer IF amplifier. Since these elements are critical in determining system sensitivity, this paper explores these components in detail.

IV. Antenna Noise Temperature

The importance of the antenna noise temperature in determining system sensitivity has been shown for the three classifications of receivers previously presented. Antenna noise temperature is a measure of the noise power available from the antenna terminals. It is caused by galactic noise and the absorption in the antenna environment that interacts with the antenna gain pattern. The noise temperature of an ideal antenna can be readily determined if the antenna pointing angle, operating frequency, and environmental conditions are specified. A nonideal antenna with losses, spillover, and minor side lobes will generally have reduced gain and a higher noise temperature. A plot of sky noise for an ideal ground-station antenna from 100 MHz to 100 GHz [4] is presented in Fig. 6 as a function of elevation angle. Sky noise in the microwave region is extremely low when looking at the galactic pole and increases when looking at the galactic center. In the millimeter-wave region, the sky noise contribution is quite significant as the signal frequency is increased from 20 to 100 GHz. The sky noise temperature converges to approximately 250 K when the antenna is looking at either the galactic pole or galactic center at 100 GHz.

In the design of millimeter-wave systems, the receiver noise temperature that is required for a specific application should be chosen judiciously. Low-noise receivers in the millimeter-wave region are a critical component due to the limited power available from available transmitters. Receiver cost rises rapidly with improvements in sensitivity. For this reason, the total system operating noise temperature

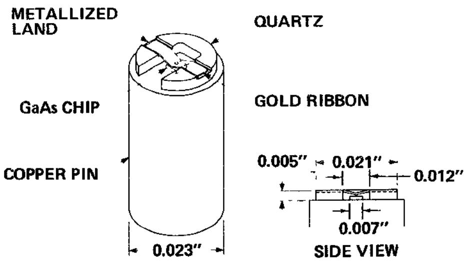

Fig. 7. Varactor holder assembly.

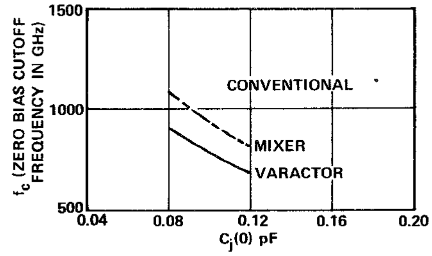

Fig. 8. Cutoff frequency versus junction capacitance for present AIL varactors and mixers.

should be considered prior to making the final decision. As an example, a change in receiver sensitivity from 600 to 300 K would result in only a 3-dB improvement in receiver sensitivity. When viewed from the system operating noise temperature, the improvement can be expressed as follows:

$$\Delta S\ (\text{dB}) = 10 \log \frac{T_{e1} + T_A}{T_{e2} + T_A}$$

where

T_A antenna noise temperature = 300 K;
T_{e1} higher receiver noise temperature = 600 K;
T_{e2} improved receiver noise temperature = 300 K.

Assuming an antenna noise temperature of 300 K, the actual improvement in system sensitivity is 1.8 dB for a 3-dB improvement in receiver performance. For this reason, all facets of the system design should be carefully reviewed.

Now that the antenna noise temperature has been considered, the critical components in determining receiver sensitivity (T_R) are evaluated.

V. Receiver Components

The key components that primarily determine the receiver noise temperature for millimeter-wave receivers are as follows:

—parametric amplifiers (both ambient and cryogenic);
—mixers (conventional IF as well as paramp IF);
—mixers (cryogenic operation);
—masers (requiring cryogenic refrigerator).

Another new element that is showing promise is the Josephson junction mixer (requiring cryogenic operation), and the performance that can be obtained with this advanced component is briefly discussed.

A. Varactor/Mixer Diode

In the majority of millimeter-wave systems that are being configured, mixers or parametric amplifiers are the main candidates chosen in determining receiver sensitivity. For these devices, the key element is the varactor [5] for the parametric amplifier and the mixer diode for the millimeter-wave mixer. While many organizations have reported significant advances for both of these devices, the approach that AIL has chosen is to use the same form factor for both of these critical elements. Fig. 7 shows the form factor for this essential building block. This is a field-proven device with many million hours of successful operating life over varied and diverse environments. The key to its success in the millimeter-wave region is its very high cutoff frequency and low parasitics. When coupled to an optimum embedding network, low-noise performance and extremely broad bandwidths can be achieved. The series self-resonant frequency of the varactor with a junction capacitance of 0.1 pF is 65 GHz. This varactor has provided very efficient multiplier performance up to 300 GHz. The theoretical cutoff frequency, defined to be

$$f_c = \frac{1}{2\pi R C_0}$$

where R is the spreading resistance and C_0 is the zero bias capacitance at 0-V bias versus junction capacitance for both the AIL varactor and mixer that are being used for deliverable millimeter receivers, is shown in Fig. 8. With a 0.1-pF junction capacitance, the 0-V bias cutoff frequency is 800 GHz for the varactor diodes and 900 GHz for the mixer diode. The reason for the higher mixer cutoff frequency is the different processing technique that is used.

With these two essential blocks, it is now possible to predict the capability of present-day parametric amplifiers and mixers.

B. Parametric Amplifiers

The noise temperature and nondegenerate parametric amplifier (single sideband) can be shown to be the following:

$$T_e = \frac{\dfrac{f_1}{f_2} + \dfrac{f_1 f_2}{M^2}}{1 - \dfrac{f_1 f_2}{M^2}}\, T_D$$

where

f_1 signal frequency;
f_2 idler frequency (pump frequency minus signal frequency);
M varactor figure of merit = $(C_1/C_0)f_c$;
$\dfrac{C_1}{C_0}$ varactor nonlinearity ratio;
f_c varactor cutoff frequency;
T_D junction temperature.

Using the previous equation for the noise temperature of the parametric amplifier, and by assuming that the amplifier is designed for minimum noise temperature (single

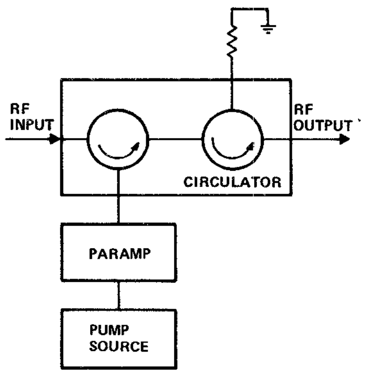

Fig. 9. Single-stage paramp.

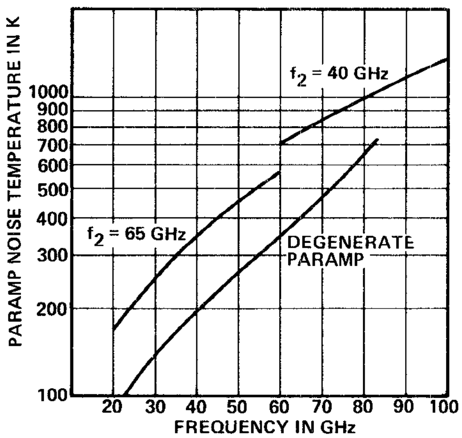

Fig. 10. Theoretical paramp noise temperature (20–100 GHz) versus frequency.

pass of circulator between antenna input and paramp port as shown in Fig. 9), a plot of theoretical single-stage parametric-amplifier noise temperature (including circulator loss) versus center frequency for available nondegenerate parametric amplifiers when operated at room ambient is shown in Fig. 10.

Two basic types of nondegenerate parametric amplifiers are illustrated. Parametric amplifiers that are designed to operate at a center frequency of 20–60 GHz will have an idler frequency of 65 GHz that utilizes the self-resonant frequency of the varactor. This is the conventional parametric-amplifier design in which the idler frequency is higher than the signal frequency.

Further improvements in noise temperature can be obtained by operating the parametric amplifiers at a cryogenic temperature of 20 K. With this type of operation, noise temperature of under 75 K can be readily obtained from 20 to 60 GHz.

By utilizing the self-resonant frequency of the varactor, it is possible to use the balanced parametric-amplifier design.[1] This type of design has the potential of achieving a much greater bandwidth than its single-ended counterpart. Utilizing this design approach, bandwidths of 5–10 percent are achievable.

[1] AIL patent 3105941.

As the signal frequency is changed from 20 to 60 GHz, the pump frequency required to ensure proper amplifier performance will vary from 85 to 160 GHz. The parametric amplifiers are operated with a solid-state pump source that is generated by a fundamental Gunn-effect oscillator that is followed by an efficient varactor multiplier. Pump sources have been successfully developed with sufficient power to pump a parametric amplifier as high as 200 GHz [6].

An alternate parametric-amplifier design that is very practical in the millimeter-wave region is one whose idler frequency is lower than the signal frequency. This type of design is used in the frequency range of 60–100 GHz with an idler frequency of 40 GHz. This type of design approach is also compatible with available solid-state sources and can be designed for spaceborne operations. The theoretical parametric-amplifier stage noise temperature (single sideband) for this advanced design with an idler frequency of 40 GHz is also shown in Fig. 10. This parametric-amplifier design is a single-ended design, and bandwidths of 0.5–1 percent are readily achievable.

As previously indicated, a design that has widespread use for radio astronomy applications is the degenerate parametric amplifier. In this type of design, the pump frequency is twice the signal frequency. The noise temperature of a degenerate parametric amplifier can be shown to be the following:

$$T_e = \frac{T_D}{\frac{M}{f_1} - 1}$$

where

- T_e double-sideband noise temperature;
- T_D junction operating temperature;
- M varactor figure of merit;
- f_1 signal frequency.

A plot of the theoretical stage double-sideband noise temperature is also shown in Fig. 10.

Parametric amplifiers incorporating all of the techniques described have been fully developed for many applications. These subsystems are completely solid state and have been fully tested for both the several environmental constraints associated with a unit qualified for an airborne as well as ground-based requirements. These parametric-amplifier subsystems can also be designed and are capable of being qualified for space application. These highly successful developments have shown that completely military-qualified low-noise amplifiers and their associated receivers can be designed to satisfy overall system requirements over a wide range of environments. Some of the parametric amplifiers that meet these conditions are presented here.

Fig. 11 shows a 35-GHz parametric amplifier [7] with a midband gain of 17 dB and a 1-dB bandwidth of 100 MHz minimum. A single-sideband noise figure of 3.8 dB has been measured using an argon noise source. The parametric amplifier is a completely solid-state device and is pumped at 101 GHz (idler frequency is 66 GHz) which is derived from a fundamental 50.5-GHz Gunn-effect oscillator and

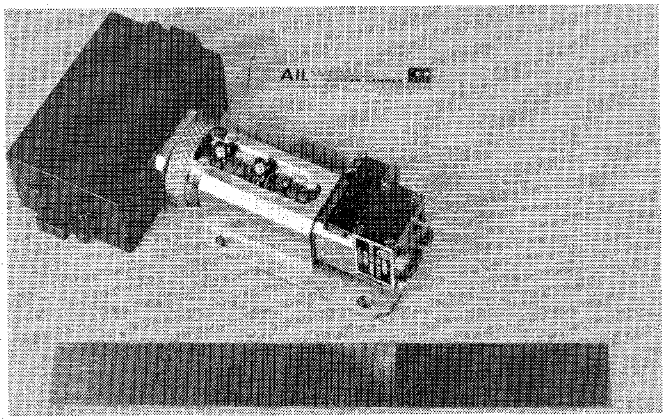

Fig. 11. 35-GHz paramp module.

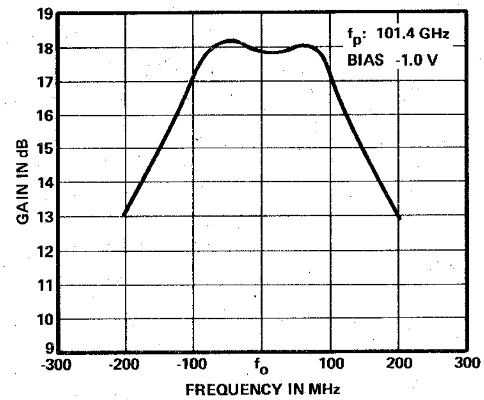

Fig. 12. Gain versus frequency for *Ka*-band parametric amplifier.

is multiplied to 101 GHz by an efficient varactor doubler (efficiency greater than 30 percent). The varactor used in the doubler is identical with the high-cutoff-frequency varactor used in the parametric amplifier. The parametric amplifier is unconditionally stable due to the high isolation provided at the input from the antenna by two passes of the circulator. The gain-bandwidth performance is shown in Fig. 12. The parametric amplifier has successfully passed a 1-W CW leakage test without any degradation in performance. A gain stability of ± 1.0 dB over an environmental temperature range of $-10°C–+50°C$ has been obtained. The system has been qualified to full military environmental extremes for both airborne operation (MIL-E-5400) and for ground-base operation. Further developments have resulted in a parametric amplifier that is now electronically tunable over a 1200-MHz band around the same center frequency and could also be broad banded to cover a full 1000-MHz instantaneous bandwidth.

Fig. 13 shows a cryogenically cooled parametric amplifier that is tunable over a minimum 2-GHz band in the 20–30-GHz frequency band.[2] The potential application of this unit was for radiometric observations of discrete lines in the galaxy. The first-stage parametric amplifier is cryogenically cooled to 20 K in a closed-cycle refrigerator. The second stage is operated at room-temperature ambient. The mid-

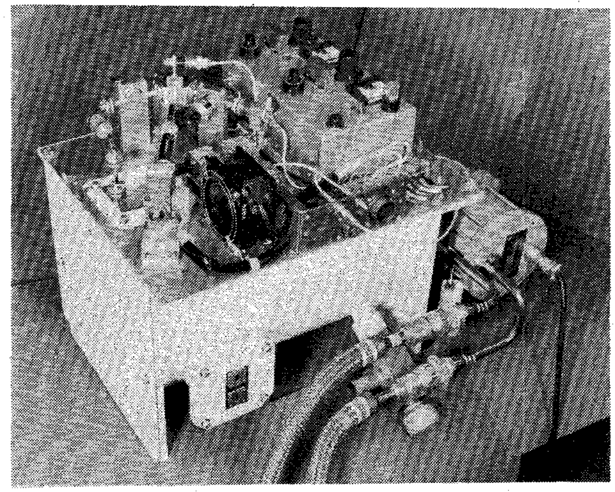

Fig. 13. 22–24-GHz tunable cryogenically cooled system.

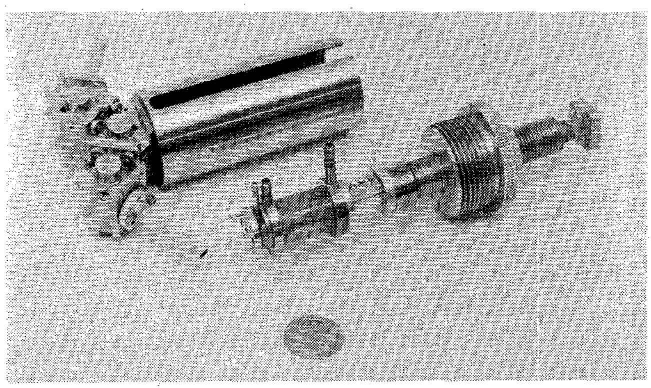

Fig. 14. 60-GHz parametric amplifier.

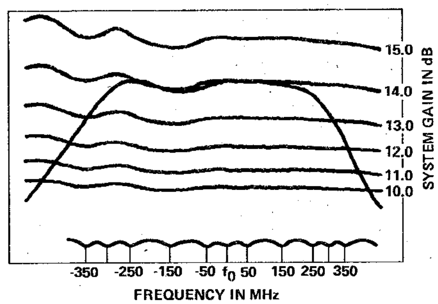

Fig. 15. Gain versus frequency of millimeter-wave (60-GHz) paramp.

band gain is 27 dB across the full tunable 2-GHz band with a minimum instantaneous bandwidth of 100 MHz and a maximum instantaneous bandwidth of 900 MHz. The measured single-sideband noise temperature is 100 K (maximum) when measured with a liquid-nitrogen/room-temperature hot/cold load.

Fig. 14 is a photograph of a nondegenerate completely solid-state 60-GHz parametric amplifier [8] that was double tuned to provide a maximum gain of 14 dB with a 1-dB bandwidth of 760 MHz (minimum) as shown in Fig. 15. The measured noise figure was 5.9 dB, measured with a hot load ($\approx$ 1000°C) [9] and a room ambient load. The paramp was pumped at 105 GHz and the 15 mW of pump power was generated by a Gunn-effect oscillator operating at 35

[2] System was developed for MPI, Bonn, Germany.

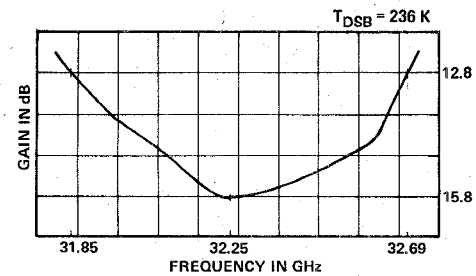

Fig. 16. Gain versus frequency of 32.25-GHz degenerate parametric amplifier.

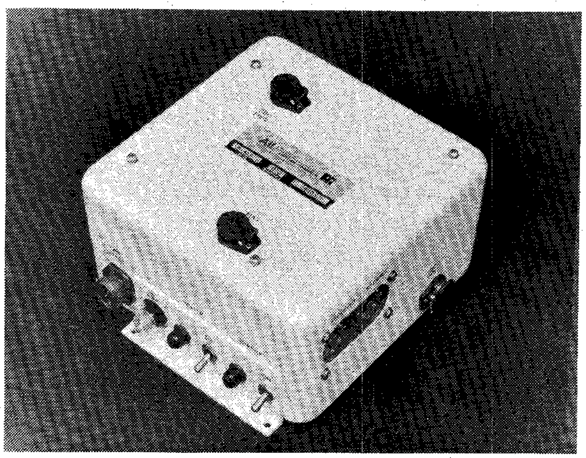

Fig. 17. 33-GHz degenerate parametric amplifier.

GHz with an efficient varactor tripler (≈ 20-percent efficiency). The unit was temperature stabilized and had a gain stability of ±1 dB over a temperature range of 0–50°C.

A degenerate parametric amplifier for an airborne radiometric application at 33 GHz has also been developed, and the measured gain/bandwidth performance is shown in Fig. 16. A double-sideband gain of 15.8 dB and a 3-dB bandwidth of 840 MHz were obtained. The measured double-sideband noise figure as measured with an argon noise source was 2.6 dB. The pump power was derived from a fundamental Gunn-effect oscillator operating at 33 GHz, some of which was decoupled by a directional coupler to provide the local oscillator power for the mixer while the remaining power was efficiently doubled to 66 GHz with an efficient varactor multiplier. The unit is shown in Fig. 17.

A degenerate parametric amplifier at 46 GHz [10] that uses a Schottky-barrier varactor in a modified Sharpless wafer has been successfully developed. With a varactor cutoff frequency of 600 GHz at a bias voltage of 0.5 V, a gain of 15 dB was achieved with a 3-dB bandwidth of 300 MHz. The double-sideband noise temperature (including a 1-dB circulator loss) was 398 K.

A degenerate parametric amplifier at 46 GHz [11] has also been successfully operated at 20 K. The varactor was a platinum–gallium arsenide Schottky-barrier junction that was mounted in a modified Sharpless wafer. A gain of 23 dB with a 3-dB bandwidth of 390 MHz was obtained with the paramp/circulator operated at 20 K. The measured double-sideband noise temperature was nominally 40 K.

Further developments to fully exploit the technology that has been developed will be greatly expanded in the very near future. The emphasis will be placed on fully exploiting the varactor figure of merit that will result in even lower noise temperature and broader instantaneous bandwidths. Further use of cryogenic cooling as well as thermoelectric cooling to −30°C to improve noise performance will be further developed. It is expected that bandwidths up to 5 GHz should be readily available in the near future.

C. Mixers

One of the key elements that has attracted much attention in the millimeter-wave region has been the development of low-noise mixers (both double and single sideband). Noise figures well under 8 dB are now possible. The attainment of low-noise performance has been accelerated by the availability of high-cutoff-frequency Schottky-barrier gallium arsenide mixer diodes that have been successfully developed by numerous laboratories. One major achievement being exploited in the millimeter-wave region is the cryogenic cooling of mixer diodes to obtain low-noise performance. System configurations that are actively using mixer front ends include, but are not limited to, the following:

—Mixer–transistor–amplifier IF (double and single sideband);
—Mixer–paramp IF (cooled and room ambient).

All of these configurations will be utilized in the millimeter-wave region.

The conversion loss in a mixer is a function of the diode cutoff frequency. A diode that has a very high cutoff frequency implies lower series parasitic resistance resulting in greater voltage developed across the junction nonlinear resistance and, therefore, lower conversion loss. Fig. 18 is a plot of the conversion loss versus pulse duty ratio with the ratio of a single frequency to cutoff frequency as a parametric as reported by Dickens and Maki [12] using an extension of a mixer theory due to Barber [13]. Using this theory and with further refinements developed by AIL, it is theoretically possible to predict the achievable noise figure based on the characteristics of the AIL mixer diodes.

A variety of circuit configurations and construction techniques are available for millimeter-wave mixers. A balanced mixer is sometimes the preferred approach when compared to a single-ended one since 20–30 dB of local oscillator noise suppression is provided. In radiometric receivers, where double-sideband mixers can be used, this suppression allows a lower IF band to be chosen, which generally reduces the system costs without degrading overall system performance. However, as previously shown for both communications and radar receivers, a preselector is required for image rejection and, therefore, a high IF frequency is normally selected to ease the design constraints that would be placed on this preselector. For this reason, single-ended mixers are very acceptable for the design of millimeter-wave systems.

The noise figure of a millimeter-wave mixer depends upon the diode characteristics, circuit loss, sum frequency

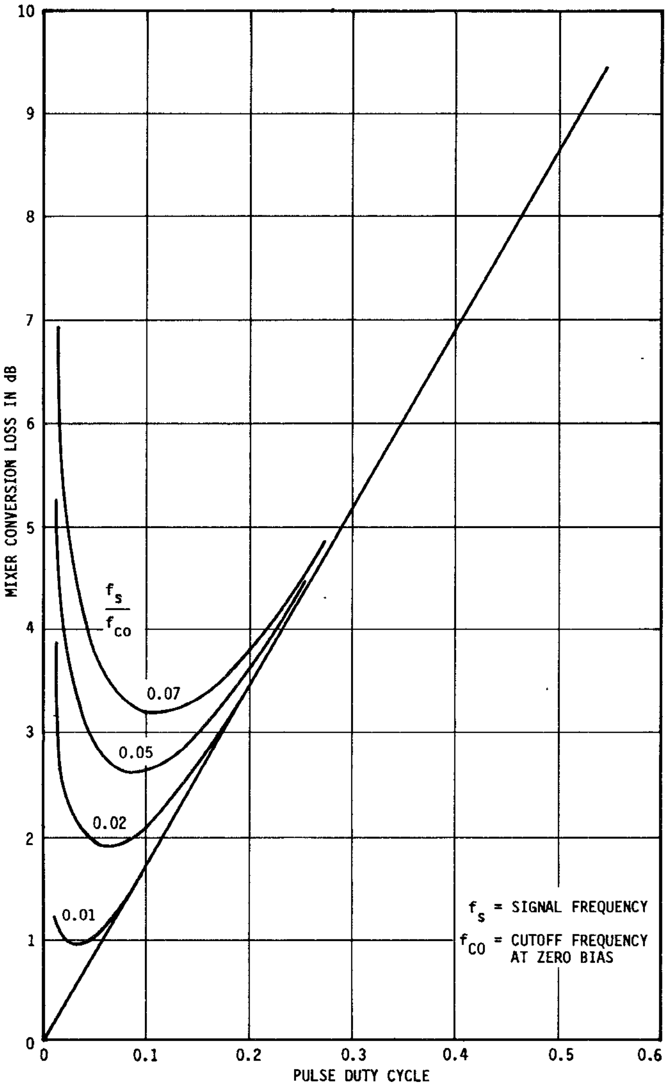

Fig. 18. Mixer conversion loss versus pulse duty ratio.

termination, local oscillator characteristics, and operating temperature. The diode characteristics are determined by the semiconductor material, junction capacitance, series resistance, package parasitics, and noise temperature ratio. With the gallium arsenide mixer diode previously described, the high cutoff frequency (800 GHz at 0-V bias), coupled with very low parasitics (0.03-pF stray capacitance), satisfies all the conditions required for good mixer action. The noise figure of a single-sideband image-enhanced mixer can be shown to be

$$F_{0v} = L_f L_m (t + F_{\mathrm{IF}} - 1)$$

where

- F_{0v} single-sideband noise figure of image-enhanced mixer;
- L_f insertion loss of filter;
- L_m conversion loss of mixer;
- t noise ratio;
- F_{IF} IF noise figure.

The high-quality mixer diodes that have been developed minimize the conversion loss (L_m) that results in extremely

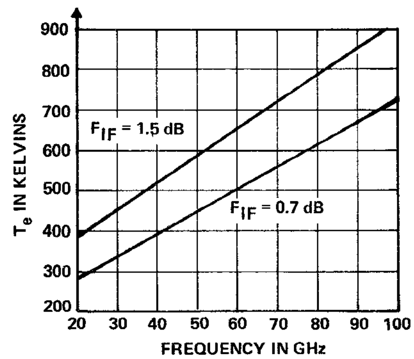

Fig. 19. Theoretical noise figure versus frequency of image-enhanced mixer.

low-noise-figure image-enhanced mixers that can be readily utilized for both communication and radar receivers.

For radiometric receivers in which a double-sideband receiver (therefore, a double-sideband mixer) is used, an improvement in sensitivity can be obtained. The noise figure for a double-sideband receiver with a mixer front end can be shown to be

$$F_{0v} = \frac{L_m}{2}(t + F_{\mathrm{IF}} - 1)$$

where

- F_{0v} double-sideband noise figure of image-enhanced mixer;
- L_m conversion loss of mixer;
- t noise ratio;
- F_{IF} IF noise figure.

As is obvious, for the same mixer-diode characteristics, the conversion loss of a double-sideband receiver is reduced by a factor of 2 and the loss of the filter that was used for both image rejection and enhancement is eliminated. For those radiometric receivers that can use this approach, a definite improvement in sensitivity can be obtained.

Fig. 19 shows the theoretical noise figure of an image-enhanced mixer/IF amplifier from 20 to 100 GHz using available mixer diodes (Fig. 9) with 0-V bias cutoff frequencies of 800 GHz. The theoretical single-sideband noise figure varies from 4.6 dB at 20 GHz to 8.2 dB at 95 GHz, and assumes an IF amplifier noise figure of 1.5 dB and a waveguide filter loss (with increasing loss due to increasing frequency taken into account). The waveguide filter provides both the image rejection and the correct image termination for improved noise performance. Further improvements in system sensitivity can be made by cryogenically cooling the mixer and also by using a cryogenically cooled paramp as the IF amplifier.

Fig. 19 shows that with an IF amplifier noise figure of 1.5 dB, the theoretical single-sideband noise figure of an image-enhanced mixer is 5.6 dB. Using the same mixer-diode parameters, the theoretical double-sideband noise figure with a 1.5-dB IF amplifier noise figure is 2.26 dB.

Fig. 20 is a block diagram of a 35-GHz image-enhanced mixer using laminant mode waveguide technology that has been recently developed. With an argon noise source

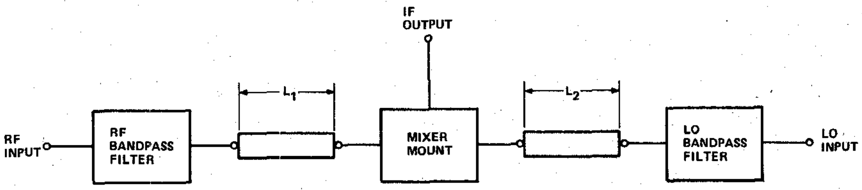

Fig. 20. Block diagram of image-enhanced mixer.

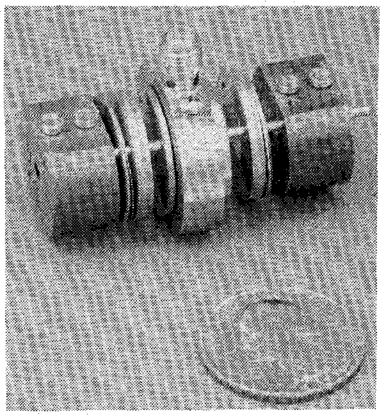

Fig. 21. 35-GHz image-enhanced mixer.

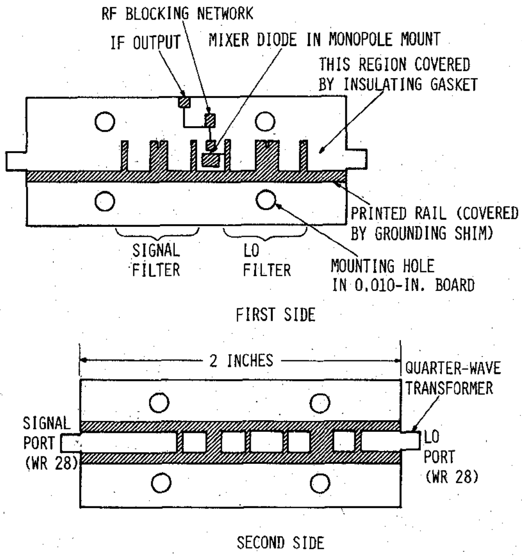

Fig. 22. Integrated receiver front end.

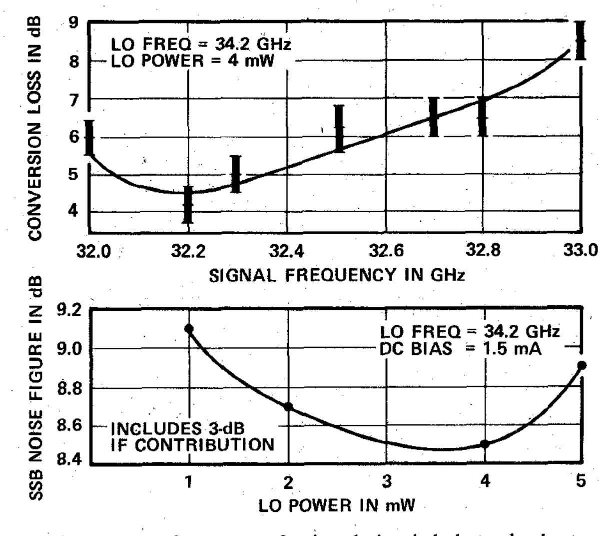

Fig. 23. Performance of printed-circuit balanced mixer.

used for the noise figure measurement, a single-sideband noise figure of 5.7 dB has been measured. The measured conversion loss of the mixer from the 35-GHz RF input frequency to the IF output frequency of 1.2 GHz was 3.1 dB. This includes the effect of a bandpass filter that was used for image enhancement that has a 0.7-dB insertion loss. The measured noise figure of 5.8 dB includes an IF amplifier at 1.2 GHz with a noise figure of 2 dB. This advanced mixer is shown in Fig. 21.

A printed-circuit version of an image-enhanced mixer for the 9-mm band has also been developed [14]. This printed-circuit design contains a monopole coupled mixer and two bandpass filters that diplex the signal and local oscillator ports. Inductive filter elements are printed on both sides of the board to achieve the high susceptance required for typical diplexing applications. The schematic representation of this advanced integrated receiver front end is given in Fig. 22. A conversion loss of 4.5 dB has been obtained at 32.2 GHz with a commercially available gallium arsenide beam lead diode. The measured performance (conversion loss and single-sideband noise figure) of the printed-circuit image-enhanced mixer across a 1-GHz band from 32 to 33 GHz that includes a 3-dB IF contribution is shown in Fig. 23. Further improvements in conversion loss (hence noise figure) can be obtained by using the high-quality mixer diodes previously shown in Fig. 8.

Another printed-circuit technique that has been successfully developed for a low-noise mixer in the 5-mm band utilizes oversize microstrip. Fig. 24 illustrates the essential features of the mixer. Both the local oscillator and the signal are coupled from the waveguide by a monopole, whose length and shape are selected to provide a broadband impedance match to the mixer diode.

Fig. 25 shows this 60-GHz mixer constructed in oversized microstrip. The monopole, diode-mounting loads, and RF blocking network are all printed on 0.005-in mylia gasket. A plot of the measured noise figure versus local oscillator power with GaAs mixer diode is shown in Fig. 26. The measured single-sideband noise figure was 9–10 dB for local oscillator frequencies in the 60–61-GHz range for drive levels of 0.9–1.8 mW. The noise figure of the 1.5-GHz IF amplifier was 5 dB.

A millimeter integrated circuit (IC) balanced mixer that

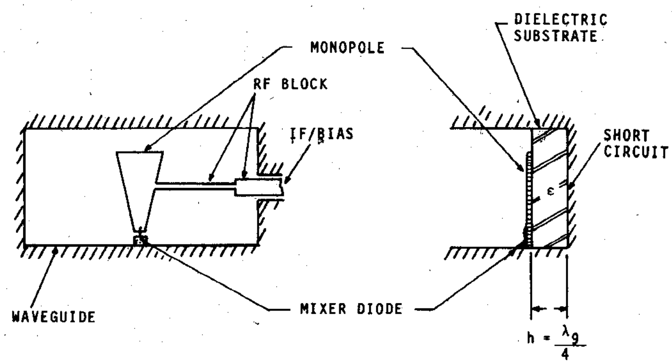

Fig. 24. 60-GHz mixer design.

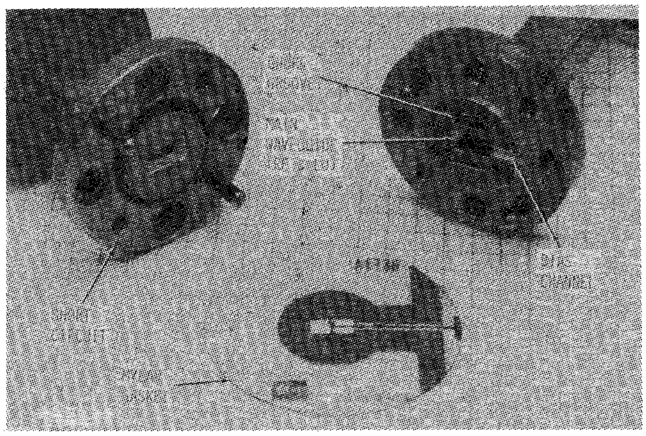

Fig. 25. 60-GHz mixer.

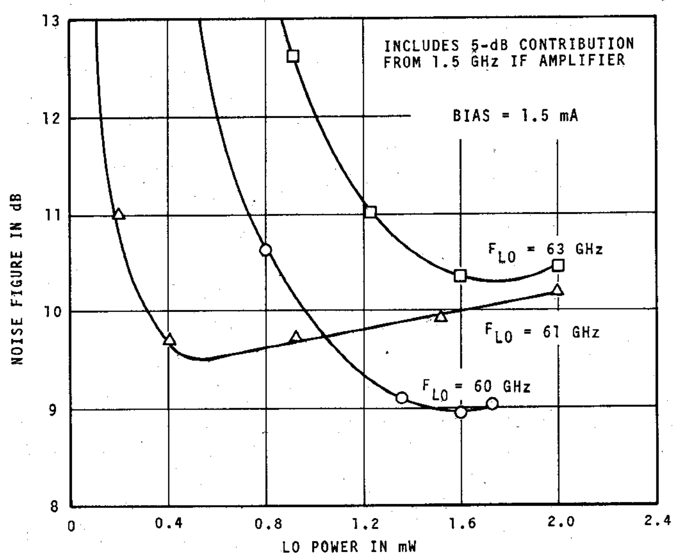

Fig. 26. Noise figure of mixer with GaAs diode.

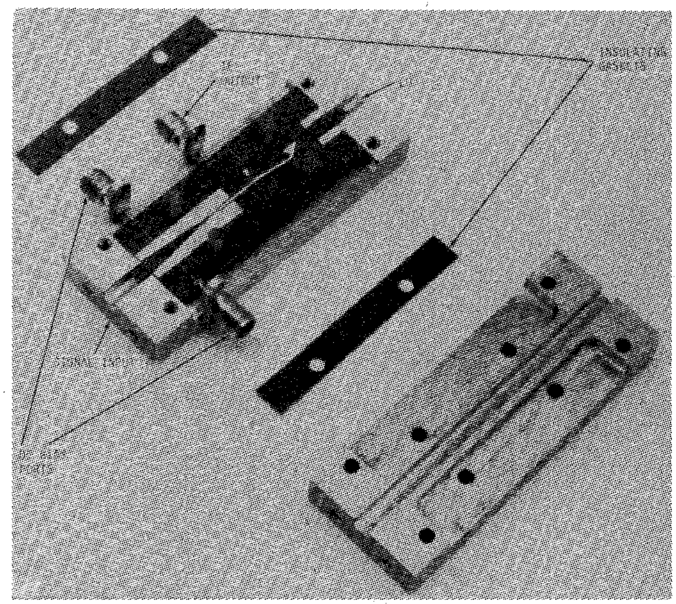

Fig. 27. Balanced mixer.

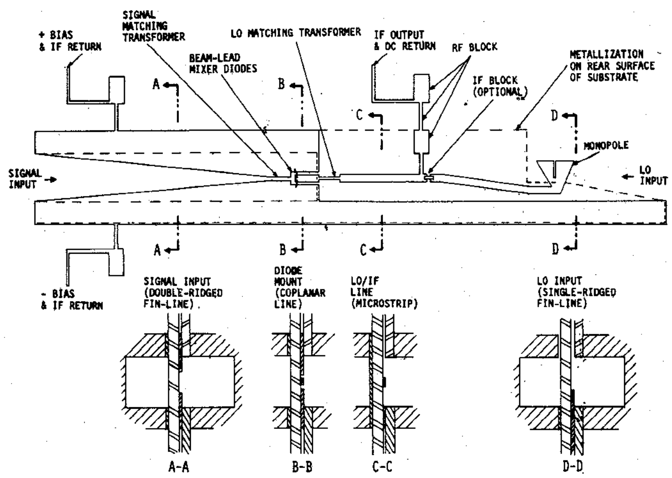

Fig. 28. Balanced mixer cross section.

has the capability of low-noise performance coupled with very low cost is shown in Fig. 27. As shown in the photograph, the aluminum housing is fabricated in two parts to accommodate the printed-circuit board and all of its associated components. The mixer was designed for the 26–40-GHz band. The construction details of this balanced mixer are presented in Fig. 28. One of the features of this design is the multiplicity of transmission lines used in this unique design. The signal line is designed in fin line while the local oscillator is coupled to the mixer diodes by microstrip. The IF output is also in the microstrip medium. The mixer diodes are mounted in coplanar line. The measured performance of this existing IC balanced mixer with commercially available diodes is shown in Fig. 29. Across the 30-percent band from 26.5 to 36 GHz, the RF to local oscillator port isolation is 20–26 dB and the measured double-sideband noise figure is 5.5 $\pm$ 1.2 dB including a 1.5-dB contribution from a 30-MHz IF amplifier. The design approach used could be readily extended to 94 GHz with an expected noise figure of 7.5 dB with higher quality mixer diodes.

With the development of high-cutoff-frequency GaAs Schottky-barrier diodes having low conversion loss in the millimeter-wave region, an alternate approach to obtain a low-noise figure has been to cryogenically cool the mixers to 20 K. The single-sideband noise temperature of a mixer receiver can be shown to be

$$T_s = T_m + LT_{IF}$$

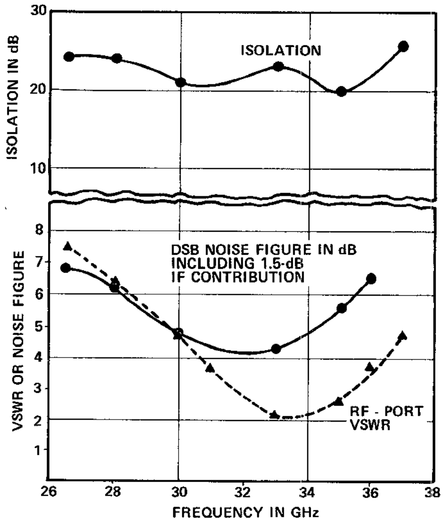

Fig. 29. Performance of printed-circuit image-enhanced mixer.

where

T_s single-sideband noise temperature;
T_m mixer noise;
L mixer conversion loss;
T_{IF} noise temperature of IF amplifier.

Mixer noise (T_m) is related to the noise temperature ratio [15] of the mixer by

$$T_m = T_0(L - 1 - L/L_i)$$

where

T_0 diode physical temperature;
I_m noise temperature ratio;
L conversion loss of mixer;
L_i conversion loss of mixer at image frequency;
L/L_i = 0 for image rejection mixer.

The usefulness of cooling a mixer is based on the fact that the conversion loss will not change, but that T_m will be reduced as the physical temperature of the mixer diodes is decreased. As is obvious, it is necessary that the IF noise temperature (T_{IF}) be sufficiently low to prevent LT_{IF} from being the dominant term. The use of a cryogenically cooled parametric amplifier with a noise temperature of approximately 20 K to be used as the IF amplifier to achieve very-low-noise performance becomes quite apparent.

Significant results have been obtained using this approach [16]. Measured values of T_m, the mixer contribution to the single-sideband noise temperature of 280 K at 85 GHz and 200 K at 33 GHz with good conversion loss, have been reported. Using these mixers with a cryogenically cooled parametric amplifier, it is possible to achieve receiver noise temperatures of 350 K at 85 GHz and 260 K at 33 GHz [16]. It has also been reported [17] that the single-sideband conversion loss mixer noise temperatures at 115 GHz are 5.8 dB and 300 K when cryogenically cooled to 77 or 18 K with a 4.75-GHz IF frequency. However, further investigation to correlate the theory with the experimental data is still necessary since the measured value of mixer noise (T_m) is higher than predicted.

D. Josephson Mixers

A recent advancement in the millimeter-wave region has been the use of Josephson mixers to obtain low-noise performance. The Josephson junction is a superconducting device and must be cooled to 4 K for proper operation. Although cooling the junction to 4 K might be a drawback, it requires a very minimum amount of local oscillator power for operation. It has been reported [18] that at 47 GHz, a single-sideband noise temperature (T_m) of 71 K has been measured with a corresponding conversion loss of 12.3 dB. The developments in this area should be followed for its use as a potential candidate for future low-noise receivers. This approach also appears to have application into the submillimeter-wave region.

The foregoing examples are but a few of the advanced developments that are being exploited in the millimeter-wave region. As the technology in this region matures, various configurations will be available from which future systems can be adequately designed.

VI. Conclusions

A review of some of the fundamental parameters that determine overall system sensitivity (antenna noise temperature as well as the receiver noise temperature) has been explored. The importance of developing receivers with very-low-noise performance has been demonstrated. The capability of those elements and the various techniques to achieve low-noise performance have been presented. It has been shown that it is now possible to configure numerous systems for many applications with the developments that have been achieved to date. Those developments have advanced beyond the laboratory curiosity stage and the receiver elements have been fully qualified to stringent military requirements. However, in order to fully exploit the millimeter-wave region, it will be necessary to achieve the low-noise performance consistent with the system economic constraints. Unless these two elements are closely coupled together, further developments will be hindered. It is necessary for the microwave/millimeter-wave community to configure systems based on the technology that has been developed and to provide the stimuli for new ideas for future systems.

Acknowledgment

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A Cryogenically Cooled Two-Channel Paramp Radiometer for 47 GHz

JOCHEN EDRICH, SENIOR MEMBER, IEEE

***Abstract*—A field operational radiometer for 47 GHz with 18 K cooled degenerate paramps is described which has parallel channels for two ortholinear or circular polarizations. It exhibits an instantaneous RF bandwidth of 300 MHz and a tuning range up to 2 GHz. Its double-sideband system noise temperature of 100 K at midband and its minimum rms noise fluctuation of 0.16 K are several times lower than those of existing millimeter wave receivers.**

I. Introduction

During the last decade uncooled parametric amplifiers have been developed for millimeter waves and found some applications in radiometry, radar, communications, and radio astronomy up to about 38 GHz [1]–[3]. Above this frequency a few laboratory experiments at room temperature have been reported at 46 GHz [4], 60 GHz [5], [6], and 94 GHz [7]. Paramps with even lower noise which are cryogenically cooled have been reported up to 24 GHz [2], [8]. This report describes the first field operational and cryogenically cooled millimeter-wave paramp system for 47 GHz.

II. Design Considerations

A. System Layout

The main aim of this program was to develop a broadband radiometer with two channels for two linear or circular polarization. In order to obtain a small rms noise fluctuation, a low ratio $T_{op}/\sqrt{B}$ of system noise temperature T_{op} and RF bandwidth B is required [9]. A secondary goal was to tune this system for spectral line observations over a band of several gigahertz around 47 GHz. This made a low value of T_{op} even more important. A system with two cooled and synchronously pumped degenerate paramps can best meet these requirements [10]. Such a system was therefore chosen and is shown schematically in Fig. 1.

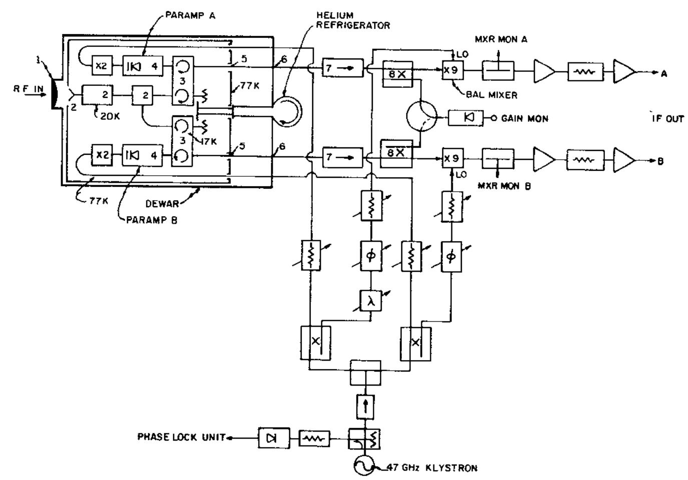

Fig. 1. Block diagram of the cooled 47-GHz receiver (numbers refer to Table I).

It requires only one source at a relatively low frequency, i.e., a klystron at 47 GHz, because it can simultaneously drive two paramps via doublers and also supply the phased LO signal to the following mixers. Another advantage of

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The author is with the Denver Research Institute, University of Denver, Denver, CO 80210.