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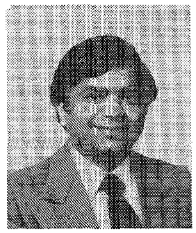

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Wide-Band Subharmonically Pumped *W*-Band Mixer in Single-Ridge Fin-Line

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Abstract —**A subharmonic mixer is described that has an instantaneous bandwidth of 11 to 14 GHz centered near 95 GHz. A wide bandwidth is achieved by the close integration of a low-capacitance diode mount, printed circuit matching elements, and simple yet effective filters which are uniquely suited to realization in a single-ridge fin-line. The mixer also has a two-terminal shunt mount that will accept two conventional beam-lead diodes or a single dual-junction device. With the latter, a minimum conversion loss of 8.5 dB has been achieved with a drive level of only 4 dBm at 45 GHz.**

I. Introduction

SUBHARMONICALLY pumped mixers have been constructed in various transmission media including stripline [1], microstrip, waveguide [2], and double-ridged fin-line [3]. Although excellent results have been obtained at millimeter wavelengths [4], tunable waveguide back-shorts have so far limited the instantaneous bandwidth. The existing shunt mounts accommodate two diodes rather than the latest dual-junction antiparallel devices [5], [6]. Since these mounts are susceptible to asymmetric bonding resulting in degradation [7], a two-terminal mount would be ultimately preferable. A wide-band fin-line mixer with a two-terminal shunt mount and unique RF/LO filters is described.

II. Circuit Description

Fig. 1 shows the construction features of the wide-band single-ridge fin-line mixer. The fin-line is printed on 5-mil Duroid which is suspended in the E-plane of a rectangular housing. The housing, in the indicated region, has the same inner dimensions as WR-10 waveguide. The gap (d) between the edge of the fin and the upper housing wall is chosen to be small enough to propagate the subharmonic local oscillator (LO), yet large enough to accommodate the antiparallel diodes. By selecting a gap ratio (d/b) equal to 0.3, the cutoff frequency of the fin-line is placed at 37 GHz which is well below the intended LO band.

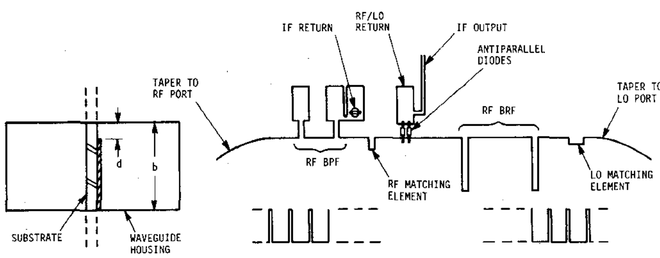

Fig. 1. Subharmonic mixer in single-ridge fin-line.

The RF signal enters from the left and passes through a cosine taper, matching the fin-line to a standard WR-10 port. The signal then passes through a single-section bandpass filter (BPF). The filter is formed by two inductive strips each joining the edge of the fin to a ground return. Grounding in the RF and LO bands is achieved with low-impedance stubs within the choke region [8] of the housing. Grounding in the IF band is provided by a feedthrough that links the illustrated pattern to a rear-surface metalization. After exiting from the BPF the signal passes through the RF matching element and reaches the antiparallel diodes which are reactively terminated by the

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RF band-reject filter (BRF). The diodes are mounted between the edge of the fin and a low-impedance stub (RF/LO ground return). The IF leaves the low-capacitance diode mount through a microstrip line and an SMA connector (not shown).

The LO enters the circuit from the right in Fig. 1. The LO-matching elements include a notch in the fin-line, a cosine taper, and a step in the width of the housing. The latter allows the subharmonic pump to propagate without the aid of the fin-line, and provides a direct interface with a standard WR-19 port. After emerging from the matching elements the LO passes through the RF band-reject filter and reaches the diode mount. It is then reactively terminated by the RF bandpass filter.

The fin-line filters, which diplex the RF and LO, are key components of the subject mixer.

III. RF Bandpass Filter

The fin-line bandpass filter was analyzed and optimized by computer-aided techniques, based on the equivalent circuit shown in Fig. 2. The circuit includes the normalized shunt susceptances (b_3 and b_5) separated by a line length (l_4) that controls the resonant frequency. The circuit also includes input and output lines (l_2 and l_6) that model the dissipative loss in the cosine tapers. All the transmission lines are assigned a normalized characteristic impedance of unity. (The tapers are assumed to be reflectionless.)

To utilize the equivalent circuit, information is required on the fin-line properties including the equivalent dielectric constant (k_e), cutoff wavelength (λ_c), and unloaded Q. Because this information was unavailable for the preferred single-ridge configuration, appropriate measurements were conducted. A sliding-short technique [8] provided measurements of the guide wavelength (λ_g) at widely spaced frequencies. The fin-line properties were then found from

$$\lambda_c = \sqrt{\frac{\lambda_2^2 - \lambda_1^2}{(\lambda_1/\lambda_{g1})^2 - (\lambda_2/\lambda_{g2})^2}} \tag{1}$$

$$k_e = (\lambda_1/\lambda_{g1})^2 + (\lambda_1/\lambda_c)^2 \tag{2}$$

where λ_{g1} is the measured guide wavelength at the free-space wavelength λ_1 and λ_{g2} is the guide wavelength at λ_2. For the preferred configuration ($d/b = 0.3$ in a WR-10 housing), k_e was found to be 1.22 and λ_c was determined to be 0.292 in. Based on other measurements of insertion loss the unloaded Q was estimated to be approximately 500.

The final parameter needed to analyze the circuit of Fig. 2 is the normalized shunt susceptance. This parameter was determined by embedding a single inductive strip between matched transitions and measuring the insertion loss across the bands of interest (40 to 60 GHz and 80 to 110 GHz). At each frequency, the normalized susceptance was calculated from

$$b_n = 2\sqrt{\log^{-1}(L/10) - 1} \tag{3}$$

where L is the measured loss (less the transition loss) in decibels. After investigating a variety of analytical expressions, it was demonstrated that the susceptance could be accurately modeled by

$$b_n = k(f - f_c)^n \tag{4}$$

where f and f_c are the operating and cutoff frequencies, respectively. The constants k and n were determined from a log/log plot of b versus $f - f_c$.

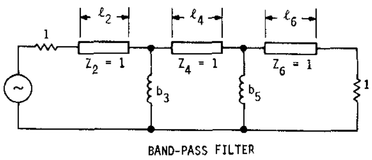

Fig. 2. Equivalent circuit of fin-line bandpass filter.

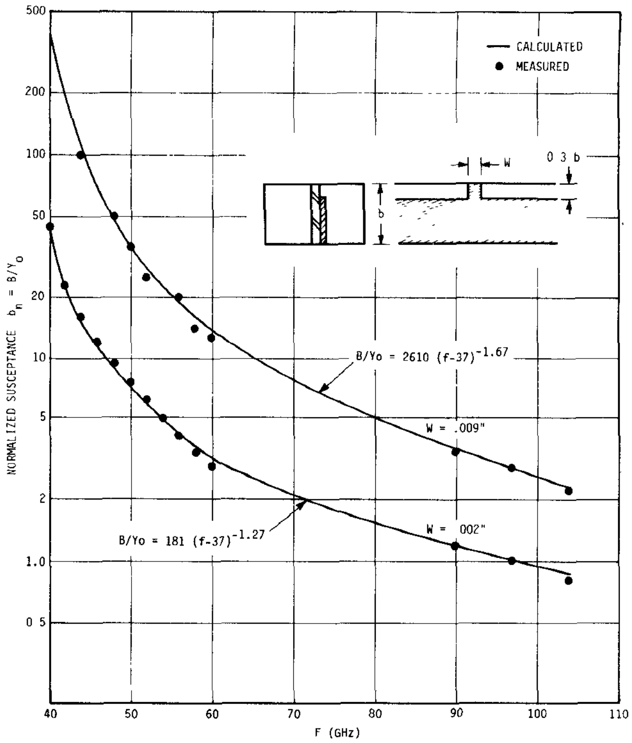

Fig. 3 Susceptance of inductive strips.

Fig. 3 shows how the susceptance varies with frequency for typical inductive strips in single-ridge fin-line. Good agreement between (4) and the measurements has been obtained for the values of k and n appropriate to a particular strip width w. Because of the exponential variation, the susceptance is an order of magnitude larger in the LO band than it is in the RF band. This condition is highly favorable to the design of the RF bandpass filter. The exponential variation assures that high rejection in the LO band and a wide RF bandwidth can be simultaneously obtained with a simple one-section filter.

Through a computer-aided analysis of the circuit of Fig. 2, it was demonstrated that adequate rejection could be obtained in the LO band with a small (2-mil) strip width. The preliminary analysis also predicted a suitable value for the resonator length l_4. Based on the assumption that the electrical and mechanical values of l_4 were equal, a bread-

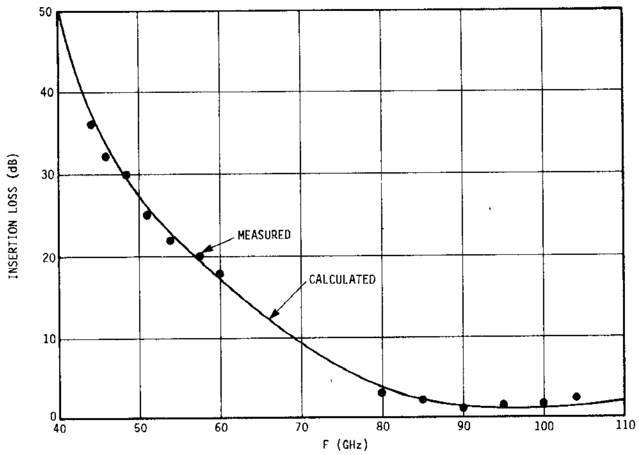

Fig. 4. Response of bandpass filter.

board model of the RF bandpass filter was constructed and tested. The measurements showed that the passband was centered at a frequency 7 percent higher than that desired. Using this information the design was refined to include an offset between l_4 and the centerline spacing of the inductive strips.

Fig. 4 shows the measured performance of the revised RF filter together with the calculated response. The measured rejection is better than 25 dB in the LO band of 42 to 51 GHz and the insertion loss is 1.4 ± 0.4 dB across the RF band of 86 to 102 GHz. The loss, measured from flange to flange in an oversized fixture, includes superfluous line lengths which were eliminated in the subharmonic mixer. The RF input loss in the final mixer design is estimated to be 0.5 dB.

It should be emphasized that the favorable asymmetric response of Fig. 4 was obtained with a single-section fin-line filter. To duplicate this performance with quasi-lumped elements in a TEM medium, at least six elements would be required [9].

IV. RF Band-Reject Filter

A second filter is required in order to complete the RF/LO diplexer. Although this filter could be a bandpass filter tuned to the subharmonic LO frequency, it would have an unsuitable susceptance variation. (In Fig. 3, the high susceptance in the passband would result in an impractically narrow bandwidth, and the low susceptance in the RF band would afford little rejection.) Consequently, a band-reject filter was chosen to block the RF from the LO port.

Fig. 5 shows the structure and the equivalent circuit of the RF band-reject filter. The filter consists of two notches that are etched into the fin-line. It has been shown that such a structure can be modeled by shortcircuited stubs that appear in a series with the main line [10]. The circuit contains the series lines (l_3 and l_5) which are each a quarter-wave long at the center of the rejection band. The stubs are spaced along the main line by l_4, which is the length that controls the match in the desired passband, as well as the center frequency of the second passband. (The latter falls above the rejection band at a frequency where l_4

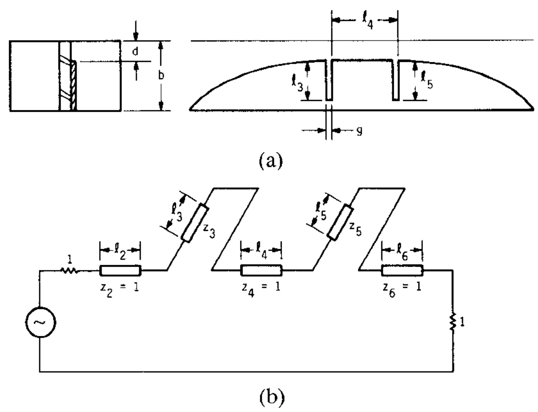

Fig. 5. RF band-reject filter. (a) Structure. (b) Equivalent Circuit.

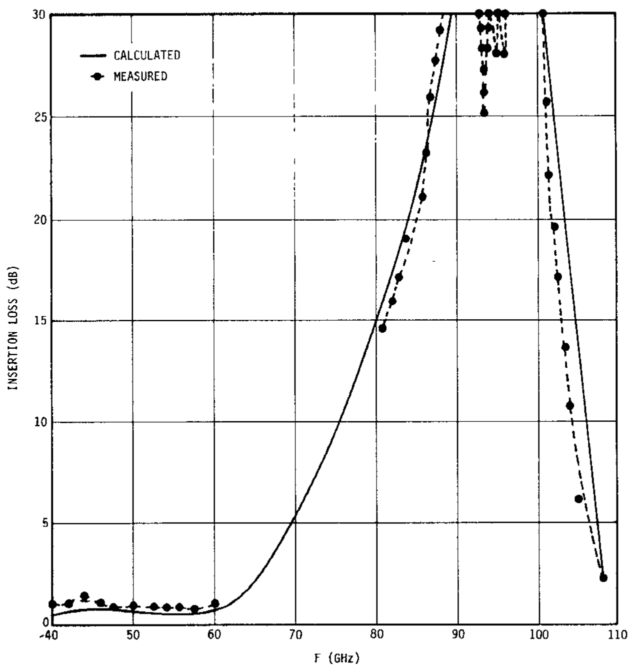

Fig. 6. Response of two-slot band-reject filter.

is approximately a half wavelength.) The equivalent circuit includes input and output lines (l_2 and l_6) to model dissipative loss in the cosine-taper transitions to standard waveguide ports.

The paper design of the filter was started with this equivalent circuit and the aforementioned fin-line parameters. Based on earlier work [10], plausible magnitudes were assigned to z_3 and z_5, and the remaining variables (l_3 through l_5) were optimized by computer-aided calculations. After obtaining a satisfactory response, a breadboard filter was constructed and tested. The model was then refined to obtain the best fit with the measurements. Good agreement was obtained with

$z_3 = z_5 = 0.95$

$l_4 =$ physical spacing between the slots

$l_3 = l_5$: 17 percent longer than the physical slot length.

With these parameters the computer model was utilized to optimize the filter performance for the intended application. Calculations showed that l_4 should be a half wavelength just above the stopband. This prevents a spurious response within the RF band and yet keeps l_4 as large as possible, thereby obtaining a favorable (quarterwave) spacing of the slots near the LO band. Fig. 6 compares the measured and calculated response of the optimized band-reject filter. The halfwave response has been fixed at 108 GHz which provides ample rejection at the upper edge of the RF band. Across the RF band of 86 to 102 GHz, the measured and calculated rejection is greater than 20 dB. In the LO band the measured insertion loss, from flange to flange, is typically 1 dB. Since the measurement includes superfluous line lengths and a pair of WR-19/WR-10 transitions, the LO input loss in the final mixer is estimated to be 0.5 dB.

V. Mixer Integration and Performance

The previously described filters were next integrated with antiparallel diodes and other circuit elements to form a mixer similar to that shown in Fig. 1. In the first phase of integration, however, the circuit did not include RF- or LO-matching elements. In keeping with the wide-band objective, low-parasitic beam-lead devices [5] were chosen for the antiparallel devices. Separate diodes were utilized in the preliminary breadboard [11]. They were later replaced by a dual-junction device, to be described.

The RF bandpass filter was located at a quarter LO wavelength from the diodes to provide a wide-band backshort termination. The band-reject filter, which appears as an open-circuit to the RF, was located as close to the diodes as possible. The LO- and RF-matching elements were then added, respectively, and optimized by an iterative technique.

Fig. 7 shows the breadboard mixer, with a duplicate of the circuit board which is suspended in the *E*-plane of the housing. As in recent designs [12], the board is held in a recessed portion of the housing and is aligned by pins external to the waveguide channel. The overall assembly measures 1.1 by 1.5 by 2.8 in.

With separate beam-lead diodes, the measured conversion loss was 12 ± 2 dB across a 12-GHz band centered at 95.5 GHz. These results were obtained with a fixed LO at 45 GHz at an input level of 16 dBm [11]. The moderate conversion loss and the high drive requirement are attributed to unequal bonding inductance [7] and the finite separation between the diodes.

The performance of the fin-line mixer was improved by replacing the separate diodes with an advanced dual-junction device [5]. This device, whose structure is shown in Fig. 8, has the following advantages:

1) The junctions are formed at almost the same location on a given wafer, thereby promoting uniformity and providing a better lumped-element circuit approximation.
2) The circulating current within the antiparallel loop is not affected by bonds external to the device.
3) Fewer bonding operations are required and the alignment during bonding is less critical.
4) The n and n^+ layers can be etched away beneath the junctions, thereby reducing the parasitic package capacitance without sacrificing mechanical strength. (Each end of the device has a wide ohmic contact.)

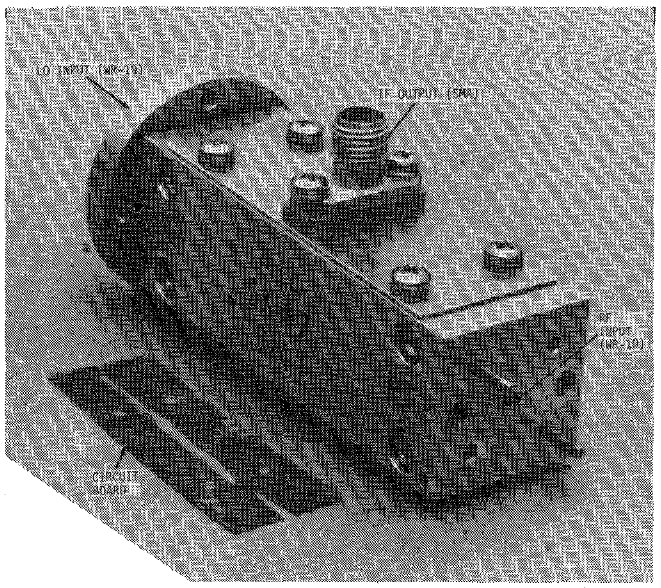

Fig. 7. Fin-line subharmonic mixer.

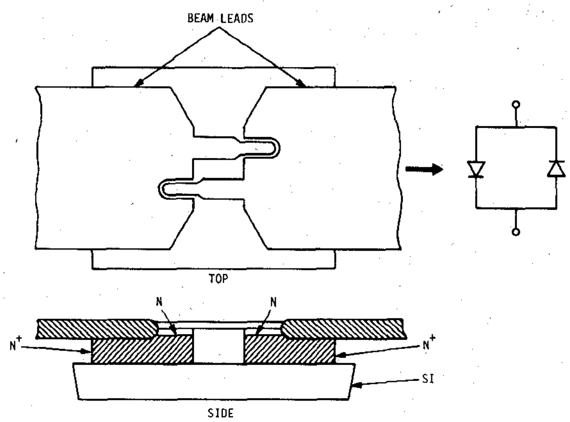

Fig. 8. Beam-lead dual-junction device structure.

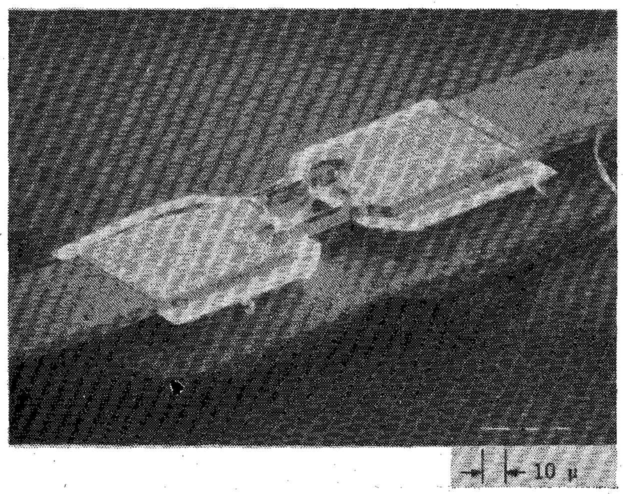

Fig. 9. SEM photograph of dual-junction device.

Fig. 9 is an SEM photograph of the dual-junction device. The device has a package capacitance of 10 fF, which is at least three times lower than that achievable with a pair of

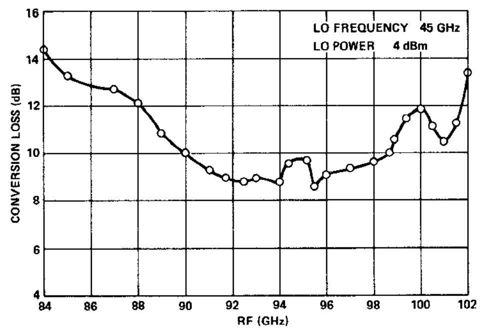

Fig. 10. Conversion loss of fin-line mixer with dual-junction device.

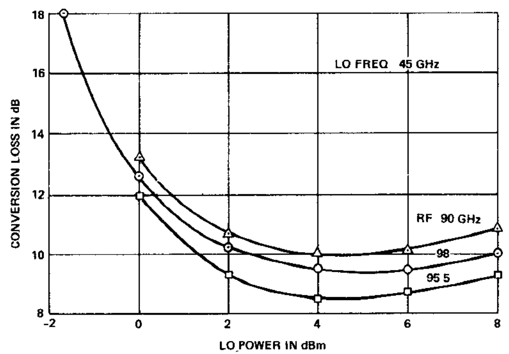

Fig. 11. Conversion loss versus LO power (dual-junction device).

the best available beam-lead diodes. Each junction has a zero-bias capacitance of 4 fF and a series resistance of 8 to 10 Ω, resulting in a minimum zero-bias cutoff frequency of 4 THz. The measured ideality factor is in the range of 1.05 to 1.07.

Fig. 10 shows the measured conversion loss of the fin-line mixer with a dual-junction device. To assure the measurement accuracy, the RF power was measured with a wide-band thermistor, calibrated against a wet calorimeter. The minimum conversion loss is 8.5 dB at 95.5 GHz and the 3- and 4-dB bandwidths are 11 and 14 GHz, respectively. As shown in Fig. 11, the optimum LO drive level is 4 to 5 dBm. This drive requirement is unusually low for a pair of self-biased high-barrier junctions. This effect is attributed to efficient LO coupling, reduced diode parasitics, and favorable input/output impedances at moderate drive levels.

The useful LO bandwidth of the existing design is approximately 3 GHz. Although a wider bandwidth could be achieved by moving the LO-matching element closer to the diodes, further design iterations would be necessary to simultaneously optimize the RF and LO match.

Because the subharmonic LO is cutoff at the WR-10 RF port, the LO/RF isolation is extremely high. Of concern in some applications, however, is the radiation from the RF port of the virtual LO, which is generated at the second harmonic of the LO. Although such even-order products can be suppressed by over 30 dB at centimeter wavelengths [2], unequal diode parasitics generally limit this suppression at millimeter wavelengths. Fig 12 shows the measured

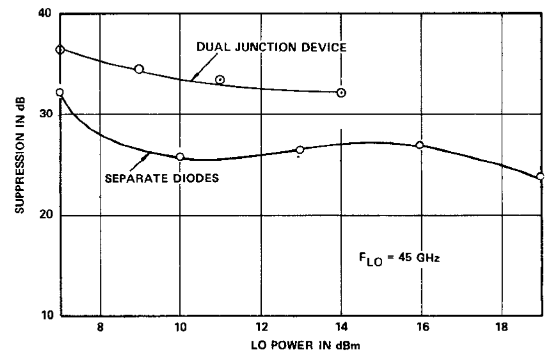

Fig. 12. Suppression of LO second harmonic.

suppression of the second LO harmonic for the fin-line mixer. The output level at the RF port, relative to the input level at the LO port, is plotted versus the LO input power. At normal drive levels, the suppression is typically 25 dB for separate diodes and 35 dB for the dual-junction device.

VI. Conclusion

A subharmonically pumped mixer has been developed with a useful instantaneous bandwidth of 11 to 14 GHz centered near 95 GHz. The wide bandwidth is achieved by closely integrating a low-capacitance diode mount, printed-circuit matching elements, and simple yet effective filters which are uniquely suited to realization in single-ridge fin-line.

The mixer includes a single-section fin-line filter which performs as well as a six-element filter in a TEM medium such as stripline. Also included is a unique shunt mount for conventional or dual-junction devices.

Although a wide bandwidth has been achieved with separate antiparallel devices, superior performance has been demonstrated with a dual-junction beam-lead device. The lower parasitics and better balance afforded by the dual-junction device result in lower conversion loss (8.5 versus 10 dB, minimum), reduced LO power requirements (4–5 versus 16 dBm), and better suppression of the LO second harmonic (35 versus 25 dB).

Because the performance of the subject mixer is comparable to that obtained with commercial fundamental mixers, the new design should be applicable to a wide range of advanced systems. The mixer is well suited to low-cost receivers—particularly where a simple yet stable (phase-locked) LO is required. The approach is also applicable to those RF bands (above 120 GHz) where fundamental low-noise solid-state oscillators are not currently available.

Acknowledgment

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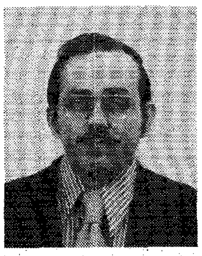

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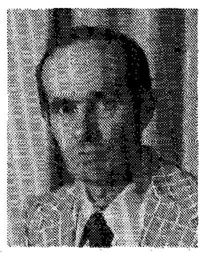
In 1961 he joined the Central Research Group of Eaton Corporation, AIL Division. In June 1973 he was promoted to Research Scientist and in April 1978 he was promoted to the rank of Senior Research Scientist. Since 1969 he has directed all the semiconductor research and development of state-of-the-art devices such as varactors, beam lead mixers, hyperabrupt tuning diodes, GaAs MESFET's, as well as studies of epitaxial growth technologies such as molecular beam epitaxy and ion implantation. He has also developed and fabricated high-performance detectors for the submillimeter and optical region. He is presently engaged in the development of high-performance GaAs monolithic microwave integrated circuits. He currently holds 8 patents, and several other patents are pending. He has published over 20 technical papers, and he has made several technical presentations at various conferences and symposiums.

Mr. Calviello is a member of Sigma Xi, and of the professional group on Electron Devices and Microwave Theory and Techniques. He is listed in "Who's Who" in the East, "Who's Who" in Technology Today and in the "Directory of World Researchers 1980's."



Paul R. Bie (S'67–M'72) was born in Brooklyn, NY, in 1930. He received his B.S. in engineering science from Long Island University in June 1971.

Since joining the Central Research Laboratories of Eaton Corporation, AIL Division in 1962 he has worked on the development and fabrication of several III–V compound semiconductors. Since 1966 he has worked exclusively on the development of Gallium Arsenide devices. This includes PN and Schottky varactor diodes, metalized and beam lead mixer devices. He is currently working on MESFET devices and monolithic circuit applications.