Fixed-Tuned Submillimeter Wavelength Waveguide Mixers Using Planar Schottky-Barrier Diodes

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***Abstract*— The design, construction, and evaluation of fixed-tuned submillimeter wavelength waveguide mixers using planar Schottky diodes are presented in this paper. Electromagnetic fields within the planar diode package were analyzed using the finite-element method (FEM). Mixers using the University of Virginia SC1T5 planar diode were designed at both 585 and 690 GHz. A double sideband (DSB) system noise temperature of 2380 K was measured at 585 GHz using 1.16 mW of local oscillator (LO) power, and a system noise temperature of 2970 K DSB was measured at 690 GHz using 1.04 mW of LO power. In addition, the 585 GHz mixer was cooled to both 77 K and 4.2 K, with measured system noise temperatures of 1240 and 880-K DSB using LO powers of 0.47 and 0.14 mW, respectively. The modeling techniques were found to predict the measured conversion loss to within 1 dB. The performance of planar diode mixers is now within a factor of 1.5 of the best whisker-contacted Schottky diode mixers in this frequency range.**

***Index Terms*— Cryogenic noise measurement, finite-element method, fixed-tuned, planar Schottky diode, submillimeter mixer, waveguide.**

I. Introduction

THE BEST Schottky diode mixers at submillimeter wavelengths use whisker-contacted diodes, which make the receiver design and assembly quite expensive and complicate the space qualification process [1]. Also, many of the best mixers rely on variable tuning elements integrated with the mixer, which introduce loss and complicate mixer construction. The main goal of this research is to develop a fixed-tuned broadband mixer using planar Schottky diodes with performance comparable to the best whisker-contacted Schottky diode mixers.

In particular, this paper describes the design, fabrication, and evaluation of waveguide mixers at 585 and 690 GHz using state-of-the-art planar Schottky diodes [2], [3]. The basic design procedure for the mixer, which includes numerical modeling of the diode and mixer circuitry using Hewlett Packard's high-frequency structure simulator (HFSS) and microwave design system (MDS), is described. Room-temperature receiver results at both 585 and 690 GHz and cryogenic-receiver results at 585 GHz are presented. In order to gauge the accuracy of the modeling techniques used, the mixer losses are estimated and the modeling is compared with the measured results.

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II. Basic Mixer Configuration

The mixer block design, shown schematically in Fig. 1, was originally developed for use with superconductor–insulator–superconductor junctions. The block that was used during this research was fabricated at the Rutherford–Appleton Laboratory [4] using direct machining techniques. The block was made in two pieces and was assembled with dowel pins. The split in the block was made in the E-plane of the waveguide to minimize the losses at the discontinuity. A diagonal feedhorn [5] integrated into the mixer block was used to couple the local oscillator (LO) and RF power into a 200-by-400-μm waveguide. The transition from waveguide to microstrip was designed using a scale model at 3.3–4.9 GHz, and exhibited a return loss of greater than 25 dB over the full waveguide band. The fixed waveguide backshort is formed by packing indium into the guide at a set distance from the transition. An IF and dc ground return is provided by a 25-μm-diameter gold wire, which has one end bonded to the microstrip and the other end contacted in indium at the end of a quarter-wave side channel. The diode, a University of Virginia SC1T5 planar diode with $2 \cdot 10^{17}$ cm^{-3} epitaxial layer doping and 1.2-μm anode diameter, is mounted across a gap in the microstrip. The distance between the gap and the low-pass filter was the main circuit element used to set the diode's embedding impedance.

III. Diode Modeling and Circuit Design

The equivalent circuit of Schottky junctions has been extensively investigated and is rather well understood. The SC1T5 diodes used during this research have diode parameters of $R_s = 14\ \Omega$, $\eta = 1.17$, $I_{\rm sat} = 3 \cdot 10^{-17}$ A, and $C_{jo} = 2$ fF. The values for R_s, η, and $I_{\rm sat}$, were determined by a least-squares fit of the measured diode I–V to the standard exponential diode equation. The harmonic-balance routines in MDS were used to determine the variation of

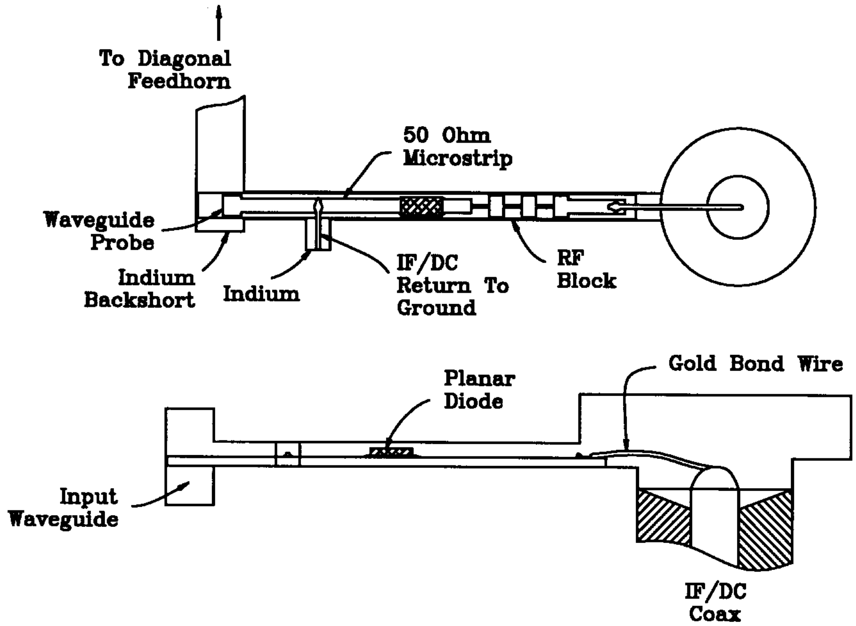

Fig. 1. Schematic of the interior of the mixer block, showing the quartz circuit and diode chip mounted in the block.

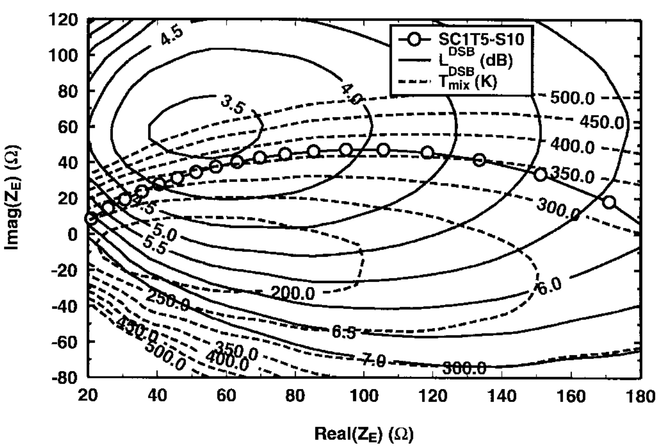

Fig. 2. Predicted contours of constant conversion loss and noise temperature for the SC1T5 diode with 0.5 mW LO power and an IF impedance of 150 Ω. The circles (◯) show the predicted embedding impedance for the SC1T5-S10 planar diode chip as the distance between the diode and the low-pass filter (l_{match}) is varied.

the mixer performance versus diode embedding impedance. The simulations include plasma resonance and skin effect by the addition of a complex series resistance [6], but do not include hot electron noise in the diode, which could be a significant source of excess noise [7]. There was an uncertainty in the embedding impedances at the higher harmonics, but simulations performed with these harmonics matched, open circuited, and short circuited changed the conversion loss no more than 0.5 dB, and a nominal value of $10 + j0\ \Omega$ was used in this research. Fig. 2 shows the predicted contours of constant noise temperature and conversion loss at 585 GHz for the SC1T5 diode with an available LO power of 0.5 mW and an IF port impedance of 150 Ω. The mixer simulations predict a conversion loss minimum of 3.4-dB double sideband (DSB) for an RF embedding impedance of $50 + j60\ \Omega$, and a mixer noise temperature minimum of 155-K DSB at $40 - j10\ \Omega$ for

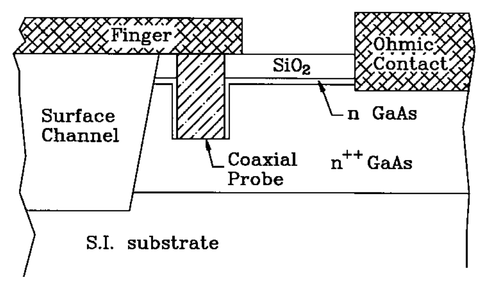

Fig. 3. Schematic of the planar diode chip near the anode with a coaxial probe inserted near the anode. This was used during the finite-element modeling to determine the diode embedding impedance.

the intrinsic diode without mixer circuit and coupling losses. Additional simulations at 690 GHz indicated similar mixer performance and optimum embedding impedances.

Once the region of optimum embedding impedance had been determined using the harmonic-balance routines in MDS, HFSS and the linear circuit simulator of MDS were used to design the mixer circuitry for the proper embedding impedance. The effect of the planar diode chip structure on the diode's embedding impedance was modeled by using HFSS to solve for the fields within the diode chip when mounted in the microstrip channel. By adding a small coaxial probe near the anode in the HFSS model, as shown in Fig. 3, the diode's embedding impedance can be determined without the use of an equivalent circuit model for the diode. The S-parameters generated by the HFSS simulation were used with the linear circuit simulator of MDS to determine the effect of the mixer circuitry external to the diode packaging. A schematic of the basic mixer microstrip configuration is shown in Fig. 4. On one side of the diode is a length of transmission line, l_{match}, between the diode and the low-pass filter. The low-pass filter presents an open circuit at its input to the LO and RF. On the other side of the diode is a length of microstrip

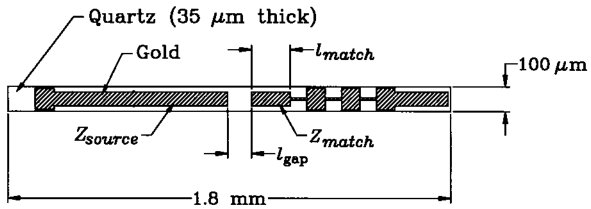

Fig. 4. Schematic of the basic mixer circuit configuration used during this research.

line running to the waveguide transition. Fig. 2 shows the predicted embedding impedance for the SC1T5-S10 planar diode (where S10 indicates a 10-μm finger length) as $l_{\rm match}$ is varied. Other important factors in determining the embedding impedance are the width of the gap across which the diode is mounted ($l_{\rm gap}$), and the diode chip geometry (e.g., finger length). The mixer circuit chosen at 585 GHz for the SC1T5-S10 diode had $Z_{\rm source} = Z_{\rm match} = 50\ \Omega$, $l_{\rm gap} = 60\ \mu$m, and $l_{\rm match} = 150\ \mu$m, yielding a predicted embedding impedance for the intrinsic diode of $45 + j30\ \Omega$, a mixer conversion loss of 3.8-dB DSB, and a mixer noise temperature of 350-K DSB. The simulations indicated that this mixer design has a 3-dB conversion-loss bandwidth of approximately 110 GHz full width. Finally, the simulations predicted that this mixer will be relatively insensitive to small changes in the mixer circuitry and assembly.

IV. Mixer Assembly and Evaluation

The microstrip circuits were fabricated on 35-μm-thick quartz substrates. In order to support the thin quartz, the wafer was mounted with wax on a silicon support wafer. The quartz was sputtered with a metal seed layer of approximately 5 nm of chromium followed by 200 nm of gold (the chromium layer aids the adhesion of the gold to the quartz). A layer of positive photoresist was then patterned onto the surface, and a 2-μm thickness of gold was plated onto the clear regions to form the microstrip circuitry. The photoresist was removed and the seed layer of gold and chromium was sputtered away from the unplated areas, leaving behind the desired circuit pattern. The quartz wafer was diced into individual circuits before removal from the silicon carrier. The bottom surface of the quartz was left uncoated, and the microstrip ground was provided by the channel bottom. Once the circuit fabrication was completed, the IF/dc connection wires were bonded onto the circuit and the diode was soldered across the gap. Finally, the quartz circuit was mounted into the mixer block and held in place by the wires, which were pressed into indium. A photograph of a quartz circuit mounted in a mixer block is shown in Fig. 5.

A schematic of the quasi-optical test-set used to measure the mixer performance is shown in Fig. 6. A Martin–Puplett diplexer [8] and an off-axis parabolic mirror with a focal length of 60 mm are used to couple the LO and RF power into the feed horn. The LO power is provided by a CO_2-pumped far-infrared (FIR) gas laser [9], [10]. The FIR laser emits a Gaussian beam with a beam waist of 8.3 mm. The parabolic mirror is placed approximately 800 mm from the

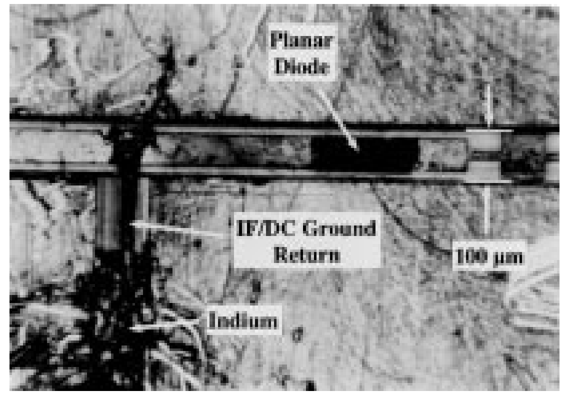

Fig. 5. Photograph of quartz circuit and diode mounted in the mixer block.

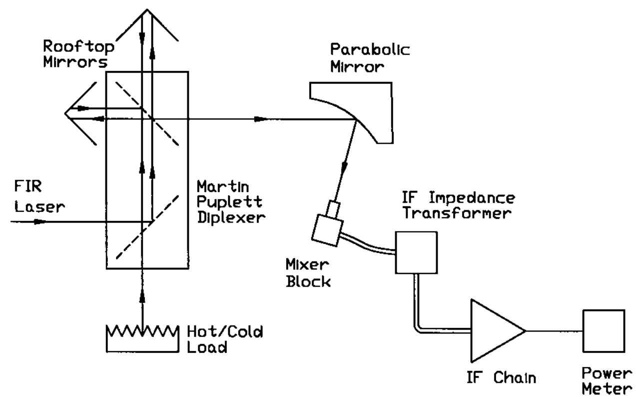

Fig. 6. Schematic of quasi-optical receiver test setup.

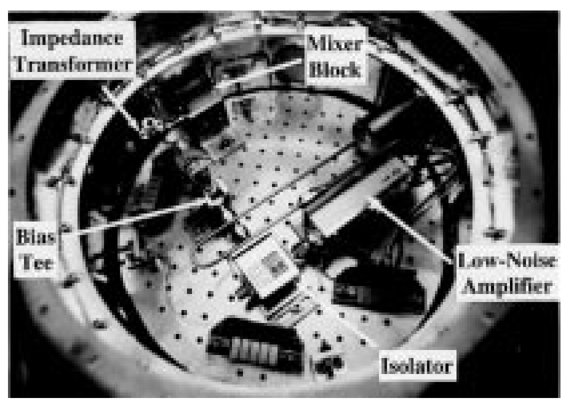

Fig. 7. Photograph of the Infrared Laboratories HD-3(8) dewar used during the mixer testing at 77 and 4.2 K.

waist, yielding an output waist of 0.58 mm, which matches well to the predicted horn waist of 0.62 mm.

The system noise temperature is measured using the Y-factor method [11], alternating between room temperature and 77-K absorber. The IF power at 1.8 GHz is amplified and then fed into a square law detector. During testing of the quasi-optical coupling to the mixer, the LO power level and the diode bias were adjusted for minimum system-noise temperature. The IF section has a variable attenuator which can be used to vary the IF noise temperature, thus allowing calculation of the mixer noise temperature and conversion loss [12]. In order to match the diode's IF impedance (typically about 150 Ω) to

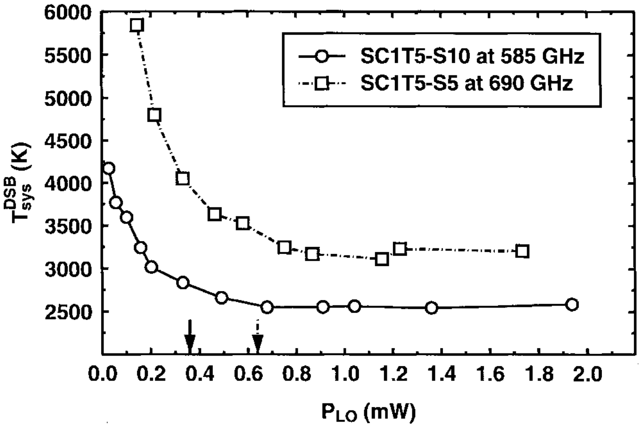

Fig. 8. 585 and 690 GHz receiver results using the University of Virginia SC1T5 planar diode at room temperature. The arrows indicate the power at which the system noise temperature has risen by 10% from its minimum value.

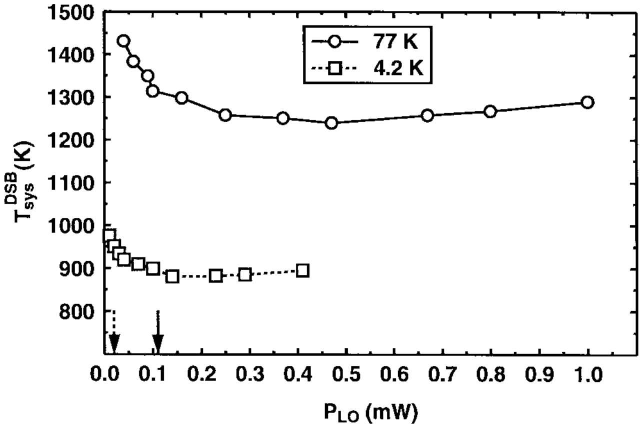

Fig. 9. 585-GHz results for the UVa SC1T5-S10 planar diode at 77 K and 4.2 K. The arrows indicate the power at which the system noise temperature has risen by 10% from its minimum value.

50 Ω, a quarter-wave microstrip IF impedance transformer was designed. The transformer was measured to have an IF return loss of greater than 20 dB and a room temperature conductor loss of 0.1 dB at 1.8 GHz.

Mixer performance was also measured at cryogenic temperatures in an Infrared Laboratories HD-3(8) dewar, shown in Fig. 7. The mixer block, IF impedance transformer, bias tee, isolator, and a low-noise amplifier (with a gain of 38.4 dB and a noise temperature of 4.1 K when cooled to 15.3 K) are mounted on the cold work surface, which can be cooled to liquid nitrogen and liquid helium temperatures. The LO and RF enter the dewar through a Teflon window (not shown), and the IF power is output through a stainless-steel semirigid coaxial cable for further amplification.

V. Room Temperature Results at 585 and 690 GHz

The best mixer results achieved at 585 GHz were obtained using the SC1T5-S10 diode, although similar results were obtained with the 5- and 20-μm finger-length diodes of the SC1T5 series. A receiver noise temperature of 2380-K DSB and mixer conversion loss of 7.6-dB DSB were measured using 1.16 mW of LO power. A plot of the system noise temperature versus LO power for a typical mixer is shown in Fig. 8. The arrow on the horizontal axis marks the power at which the system noise temperature has risen 10% from its minimum value. The power was measured using a Scientech Power-Energy Meter [13].

Performance was also measured at 690 GHz using the same mixer block but with a new circuit designed specifically for this frequency. The SC1T5-S5 planar diode yielded a system noise temperature of 2970-K DSB and mixer conversion loss of 8.8-dB DSB using 1.04 mW of LO power. A plot of the system noise temperature versus LO power at 690 GHz for a representative mixer is shown in Fig. 8.

VI. Cryogenic Results at 585 GHz

The 585-GHz mixer with an SClT5-S10 diode was evaluated at both 77 K and 4.2 K. The receiver noise temperature dropped to 1240-K DSB when the mixer was cooled to 77 K.

TABLE I
Summary of Receiver Test Results at 585 and 690 GHz for the SC1T5 Planar Schottky Diode, Including Representative Measurements of the System Noise Temperature at Low-Power Levels

ν_{RF} (GHz)	Temp. (K)	P_{LO} (mW)	T_{sys}^{DSB} (K)	T_{mix}^{DSB} (K)	L^{DSB} (dB)
585	300	1.16	2380	1800	7.6
		0.34	2660		
585	77	0.47	1240	1110	9.0
		0.09	1350		
585	4.2	0.14	880	840	9.0
		0.02	950		
690	300	1.04	2970	2240	8.8
		0.52	3250		

Further cooling to 4.2 K reduced the system noise temperature to 880-K DSB. No corrections were made for losses in the Teflon Dewar window. In addition to the improvement in system performance, the LO power requirement for the mixer dropped significantly upon cooling, as shown in Fig. 9.

The best room temperature and cryogenic receiver results are summarized in Table I. Table I also shows the significant reduction in the LO power requirement when the system noise temperature is allowed to increase by approximately 10%.

VII. Comparison of Simulations with Measured Results

The estimated losses for the 585- and 690-GHz mixers at room temperature are given in Table II. The losses in the microstrip and in the planar diode chip were estimated using the two-dimensional (2-D) port solve routine in HFSS. Conductor losses are difficult to estimate for transmission lines with significant surface roughness. However, the loss for a microstrip line with a surface roughness much larger than the skin depth (0.1 μm for gold at 585 GHz) has been shown to be approximately double that of a smooth line [14]. The microstrip and planar diode chip conductor losses in Table II have, therefore, been doubled from the value predicted for a smooth conductor. The conductor loss in the feedhorn was estimated by assuming that the feedhorn has a loss similar to

TABLE II
COMPARISON OF THE MEASURED RESULTS WITH THE MODELED RESULTS, INCLUDING ESTIMATED SYSTEM LOSSES

	585 GHz	690 GHz
Modeled		
L_{HB}^{DSB} (no loss) (dB)	3.8	5.5
Microstrip Losses (dB)	1.0	1.0
Losses in Diode Chip (dB)	0.7	0.7
Horn Losses (dB)	0.6	0.6
Diplexer and Mirror Losses (dB)	0.7	0.7
L_{HB}^{DSB} (with loss) (dB)	6.8	8.5
Measured		
L^{DSB} (dB)	7.6	9.2

that of the input waveguide, which was calculated to be 0.05 dB/mm. For a horn length of 12 mm this yields a horn loss of 0.6 dB. The losses in the quasi-optical system consist of losses in the off-axis parabolic mirror and the Martin–Puplett diplexer. At submillimeter wavelengths, the off-axis mirror has a loss of approximately 0.22 dB, while the diplexer mirrors have losses of about 0.07 dB per reflection [15]. Each wire grid is estimated to cause 0.1 dB of loss. Power passing through the diplexer is affected twice by the mirrors, and three times by the grids, leading to an estimate of 0.7 dB for the total quasi-optical system loss.

Using these estimates of the system losses, the modeling can be compared with the measured results, and, as seen in Table II, the two agree to within 1 dB. In general, it was found that the modeling techniques used in this research are useful for designing a mixer to near the optimum operating point.

VIII. CONCLUSION

For the first time, a planar diode mixer exhibited performance that is within a factor of 1.5 of the best whisker-contacted diode mixer in the 500–700-GHz frequency range [1]. Furthermore, it is important to note that this performance was obtained with no variable tuning elements in the mixer, in contrast to the best whisker-contacted mixers, which used tunable backshorts. Also, the planar diode used for this research was not optimized for operation at 600 GHz. By making slight changes to the mixer block and using higher doped, smaller anode diameter diodes, one expects planar diode mixers to perform as well or better than the best whisker-contacted diode mixers in this frequency range. Finally, this was the first time that the performance of a submillimeter wavelength planar diode mixer had been measured at cryogenic temperatures.

This research demonstrated that through the use of modern high-frequency (HF) simulation tools, it is now possible to design and fabricate optimized submillimeter wavelength mixers based on planar diodes. The modeling was used to design a robust and broad-band mixer that was insensitive to small changes in the mixer construction, thus allowing the results of this research to be readily reproduced. These new fixed-tuned planar diode mixers are expected to replace whisker-contacted mixers at most submillimeter wavelengths, thus providing a simple, rugged, room-temperature receiver technology with excellent sensitivity.

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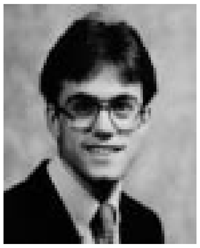

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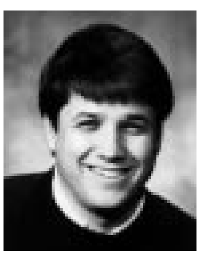

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Dr. Bradley is a member of the American Institute of Physics, American Astronomical Society, International Union of Radio Science (Commissions D and J), Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

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