A New Portable 449 MHz Spaced Antenna Wind Profiler Radar

Brad Lindseth¹,², Senior Member, IEEE, William O.J. Brown¹, Jim Jordan³,
Daniel Law³, Terry Hock¹, Member, IEEE, Stephen A. Cohn¹, Zoya Popovic²,

¹National Center for Atmospheric Research,
²University of Colorado at Boulder
³National Oceanic and Atmospheric Administration
Boulder, Colorado

Abstract— This paper presents the design of a 449 MHz radar for wind profiling, with a focus on modularity and solid-state transmitter design. It is one of the first wind profiler radars to use low-cost LDMOS power amplifiers combined with spaced antennas. The system is portable and designed for 2-3 month deployments. The transmitter power amplifier consists of three 1-kW peak power modules which feed 54 antenna elements arranged in a hexagonal array, scalable directly to 126 elements. The power amplifier is operated in pulsed mode with a 10% duty cycle at 54% power added efficiency (PAE). The antenna array is designed to have low sidelobes, confirmed by measurements. The radar was operated in Boulder, Colorado and Salt Lake City, Utah. Atmospheric wind vertical and horizontal components at altitudes between 200m and 4km were calculated from the collected atmospheric return signals.

Index Terms— Wind Profiler Radar, High Power Amplifier, Push-Pull Amplifier

I. INTRODUCTION

The spaced antenna wind profiling method [1-2] is a way to determine the horizontal velocity of the wind without steering the antenna beam as in the Doppler Beam Steering (DBS) method [3]. Advantages of this method include a simpler RF network for the antenna feed and improved time resolution of the wind velocities. Existing wind profiler systems at the National Center for Atmospheric Research (NCAR) operate at 915 MHz. Other wind profiling systems such as the National Profiling Network [4], the MU radar [5], the OQNet [6], the Lindenberg radar [7], the Gadanki MST radar [8] and other commercially built systems [9] use 50, 449, and 482, and 915 MHz frequencies. Further review of the current systems, networks, and techniques can be found in [10] and [11]. Higher frequencies, e.g. 915 MHz are more sensitive to Rayleigh scatter from precipitation while lower frequencies, e.g. 50 MHz are more sensitive to clear-air echoes from temperature and humidity fluctuations [12].

While vertical wind velocities are found from measured Doppler shift, the horizontal winds are computed using the spaced antenna method. Historically the spaced antenna technique has been used with large dipole antennas at HF wavelengths to study the ionosphere [13]. More recently the spaced antenna technique has been developed to measure atmospheric winds with a simpler antenna topology, while adding some processing and receiver complexity. It has been successfully used, e.g. at Jicamarca for wind measurements at 50MHz using dipole arrays [14], and also at the Adelaide radar [15], and MU radar [16]. This method
determines the velocity by computing the cross correlation between three or more different receivers using a method called Full Correlation Analysis (FCA) [17]. As the wind and turbulence move over the receivers, the velocity can be computed from the time lag measured between adjacent receivers. Another disadvantage of spaced antenna (SA) is that the signal to noise ratio is typically lower than with Doppler Beam Steering (empirical observations suggests SA has a 10 dB lower SNR), which uses the first moment. FCA uses higher order moments.

The spaced antenna technique offers improved time resolution over Doppler Beam Steering because the beam is pointed in one direction, while DBS systems require averaging over at least 3 different beam directions. In DBS the sampling volumes are widely separated (up to 500 m apart at 1 km range), so to satisfy continuity among the sampling volumes, long time averages (> 10 minutes) are used. The continuous vertical beam used for SA can also allow measurement of boundary layer fluxes of momentum, sensible heat, and latent heat [2]. Another technique that can be used with a spaced antenna system is post beam steering, where the phase of the spaced antenna receiver data is altered in post processing so that the receive beam is steered.

This new radar system is similar to the NCAR Multiple Antenna Profiler Radar (MAPR) which is a deployable 915 MHz spaced antenna radar [2]. Because of high antenna sidelobe levels, the MAPR antenna requires a ground clutter fence, which consists of metal panels placed around the perimeter of the antenna and is difficult to deploy because of its large size (about 3m x 3m square) and mass (about 100kg). The 449-MHz radar presented here uses an antenna design with lower sidelobe levels that eliminate the need for a clutter fence. While 50-MHz systems are most sensitive to clear-air echoes from temperature and humidity fluctuations, the size of the antenna arrays at this frequency is quite large and not as suitable for portable systems. The 449-MHz wind frequency is a good compromise between antenna size and clear air wind sensitivity, and is chosen for the system described in this work and shown in Figure 1(a).

The requirements for wind profilers are different than those for other radars, e.g. precipitation radars [18], because wind profilers receive very little of the transmitted signal. The desired target is Bragg scatter from turbulence [19], which occurs from irregularities in the index of refraction produced by temperature and humidity fluctuations. The radar is designed to measure boundary layer winds from an altitude of 100m to 5km.

A general block diagram of the 449 MHz wind profiler is shown in Figure 1(b), while the basic radar parameters are given in Table I. A pulse is generated by a D/A converter within the computer system. The system uses a pulse with a 4-bit complementary code [20] and an inter-pulse period of 50μs. This pulse is amplified by the transmitter and transmitted on all three antennas. Each antenna in the block diagram represents a 18-element array, as described in Section II. Each of the three antennas, receives the radar return signal separately. The signal from each of the three receivers, described in detail in Section III, is processed
separately to compute horizontal winds, as discussed in Section V. The LDMOS power amplifier (PA) design and characterization is presented in Section IV.

II. MODULAR ANTENNA ARRAY DESIGN

The antenna used in this system is a modular design using three 18-element linearly-polarized circular patch antenna hexagonal arrays (54 elements total). Circular patch antennas in hexagonal arrays have been considered by simulations earlier [21]. This type of antenna array is chosen for the wind-profiler due to its low sidelobe levels which minimize ground clutter and allow operation without a clutter fence. The basic unit array is the 18-element hexagonal array, shown in Figure 2(a). This array can be used as a sub-array and lends itself easily to modular and scalable arrays. Figure 2(b) shows a simulated antenna pattern for the 18-element hexagonal array, using Ansys HFSS [22] for the single element and a standard array pattern calculation for the array.

The circular patch antennas are probe-fed for linear polarization [23] and implemented using single sided copper FR4 circuit board material separated from the ground plane by a 13-mm thick Nida-Core H8PP honeycomb material which has a relative permittivity of ε_r=1.12. The ground plane is a 3-mm thick aluminum sheet chosen for light weight and mechanical stability. Figure 3(a) shows the construction of one of the 6-element parallelograms that make up the hexagonal array. The patch feed points are 5.8 cm offset from the center for a good match to 50-Ω SMA connectors. After adding a 25-mm thick polystyrene flat radome above the patches, the return loss of the antenna was measured using a vector network analyzer calibrated to the SMA connector. The match is better than -15dB at the 449-MHz design frequency, as shown in Figure 3(b).

Due to the size of the array and the large wavelength, it was not possible to measure the antenna pattern in an anechoic chamber. To confirm the important low sidelobe levels, measurements were made at an outdoor antenna range as illustrated in Figure 4(a). The 18-element array was driven by a calibrated 18-way splitter network. The sidelobe levels were measured at horizon (0°), 12.6° and 20°, limited by the height of the telephone pole used for the probe antenna in the far field of the array. The measurement data is shown in Figure 4(b), confirming the low sidelobe levels at the horizon. The measured -30 dB level is sufficient to eliminate the need for a ground clutter fence.

After measurements confirming low sidelobe levels in the 18-element array, two additional 18-element arrays were built to complete the 3-receiver system with a total of 54 elements. The configuration and simulated antenna patterns for the 54 element array factor is shown in Figure 5(a) and (b). The size of the arrays were chosen to allow easy transport between field projects. After construction of the three arrays and their splitter networks and feed cables, phase and amplitude were measured at each element to confirm proper phasing of the antenna. This test was accomplished by using a VNA and a near field test patch positioned above each element in a repeatable manner. Amplitude data from one of the hexagons is shown in Figure 6.
The receive section is shown in Figure 7. The receive section consists of a limiter to protect the LNA, a 5 MHz filter, and a mixer / IF amplifier stage to downconvert the 449 MHz to 60 MHz. The 5 MHz filter blocks out RFI and also keeps out of band noise from aliasing into band during the mixing process. The IF amplifier drives the 30m cable back to the data system and A/D converter. The gain of the receive section is designed so that the receiver noise will be amplified enough that the 2 least significant bits of the A/D converter will always be active.

A receiver noise temperature of 250K is measured after the LNA, this includes sky noise, LNA noise figure, and cable losses. To calculate required system gain, first the receiver noise power is calculated using the front end filter bandwidth of 5 MHz and the receiver noise temperature of 250K.

\[ P_{\text{Noise}} = 10 \log \frac{kTB}{.001} = -107.6 \text{ dBm} \]

Because winds and atmosphere have radar returns in the -140 to -150 dBm range we need additional processing to see below -107 dBm. The SNR after coherent integration of the radar signal is given by:

\[ \text{SNR} = N \cdot \text{SNR}_{\text{Single Pulse}} \]

where \( N \geq 128 \) is the number of coherent integrations. For this value, the coherent averaging gain is about 21 dB. The minimum detectible signal to noise ratio due to spectral averaging was determined empirically by [24,25]:

\[ \text{SNR}_{\text{Avg}} = 10 \log \left( \frac{25 \sqrt{\text{NSP} - 2.3125} + \frac{170}{\text{NFFT}}}{(\text{NFFT})(\text{NSP})} \right) \text{ [dB]} \]

where NFFT=256 is the number of points in the FFT and NSP=17 is the number of spectral averages. For these values, the spectral averaging gain is about 16.5 dB. This gives a total processing gain of 37.5 dB, resulting in a minimal detectable signal of \(-107+37.5\) = -144.5 dBm.

Full scale input for the Pentek 7642 A/D converter is +10 dBm. The SNR in the LTC2255 datasheet is given as 71 dBFS at 60 MHz [26]. Thus the required input power to be above the A/D noise is +10 dBm-71 dB = -61 dBm. This results in a required IF gain of at least 107.6 dBm-61 dBm = 46.6 dB.

During the prototype phase, a number of sensitivity tests were conducted. This test involves checking each receiver for sensitivity to a -150 dBm test signal. This test is important because it verifies the performance of the system from the antenna.
terminal through the receiver to the signal processing software. All of the receivers had a detection threshold near -150 dBm. The receive signals are processed with spaced antenna software based on the NCAR Maprdisplay package [27], and further processed for horizontal winds using Briggs’ Full Correlation Analysis (FCA) method [17].

IV. TRANSMITTER ARCHITECTURE

The transmit section of the system is shown in Figure 8. It consists of a mixer to convert the 60 MHz transmit pulse from the D/A converter to 449 MHz. A 10W driver and 80W amplifier stage drive the final amplifier. The core of the transmit section is the three 1 kW peak power amplifiers. These three amplifiers are combined using a reactive combiner, the output is transmitted through a 30m heliax cable to a 3-way splitter located underneath the outdoor antenna. The output is split to the three hexagonal antennas and then split 18 ways using phase matched cables to each circular patch antenna.

The MRF5S9070N is a low cost (~$35) LDMOS transistor capable of 80W CW. To design a high efficiency amplifier using this transistor, Class-E amplifier theory was used [28-29]. The transistor is modeled as a switch with an output capacitance. For class-E operation, a network is added to the output of the transistor that forms a low pass filter so that only a sinusoidal waveform is seen across the output load. Using the Class-E theory derived in [30], the ideal output impedance to achieve a sinusoidal waveform is given by:

$$Z = \frac{0.0446}{C_s f} e^{j49.05^\circ} [\Omega]$$

$C_s$ can be calculated from the given S-parameters for a transistor. In this case the manufacturer provided a value for $C_s$ in the datasheet, 34 pF. Using this capacitance value, the calculated value of the output impedance is $Z = 2.8 + 3.3j$ Ω. Because of the high output capacitance of this device, a low output circuit impedance is required.

The substrate used for the 80W amplifier is Rogers 4350B. The output network was designed for the target impedance and then a shunt capacitor was used on the output to tune the amplifier for best power added efficiency (PAE) and output power, Figure 9. A PAE of 68% was measured with an output power of about 49 dBm, Figure 10. Additional tests were conducted to confirm the phase stability of multiple amplifiers over temperature. The measured phase variation was a maximum of 8 degrees of phase over the -15C to 40C temperature range. Because of the low cost, this transistor was initially considered for use in an active array design with an amplifier located behind each antenna. As higher power final stages were considered, this amplifier became a low cost driver amplifier.
Improvements in LDMOS device technology have enabled kW-level amplifiers. The Freescale MRF6VP41KHR6 [31] was evaluated for use as a 1 kW peak power pulsed 449 MHz high power amplifier. It is packaged in a push-pull configuration, so that two devices are easily combined. A manufacturer test circuit was modified for best gain, efficiency, and output power at 449 MHz. The goal for this application is to have a 1 kW, 10% duty cycle, 449 MHz pulse amplifier.

The 1-kW peak pulse amplifier uses lumped element components. The overall schematic including bias tees is shown in Figure 11(a). The amplifier layout was fabricated on Rogers 4350 substrate. Some of the benefits of a push-pull amplifier are a doubling of the input and output impedances and reduced even harmonics. Coaxial baluns made of $25\Omega$ coax are used to transform the unbalanced $50\Omega$ input and output impedances into lower impedances that are easier to match to the transistor. The input return loss of the balun at 449 MHz is -12 dB (VSWR 1.7:1). The insertion loss of both balun networks measured back-to-back is 0.1 dB. The amplifier is tuned for best gain, PAE, and output power at 449 MHz by changing the value and position of the capacitors in the input and output matching networks, however the efficiency at 1 kW output is limited by the 110V maximum rating for Vds. The output matching network is shown in Figure 11(b). The input matching network has a similar topology.

A photograph of the amplifier with a light-weight aluminum heat sink is shown in Figure 12. A copper insert was installed below the transistor to allow more heat transfer between the transistor and the aluminum heatsink. A transistor clamp made of Teflon allows for easy test and replacement of the transistor if needed.

Output power, efficiency, and gain of the amplifier as a function of input power are shown in Figure 13. The best amplifier operating point is $P_{in}$ (peak) = 41.5 dBm, $P_{out}$ (peak) = 60.1 dBm, PAE = 53.8%, gain = 18.5 dB. Since the final PA stage has 18.5 dB gain, the overall efficiency of the amplifier chain is dominated by that stage. The PAE at 1 kW output is limited by the 110V maximum rating for Vds. Figure 14 shows the amplifier performance vs. frequency. The amplifier performance is frequency dependent because of the narrowband nature of the input and output matching networks. Because the only modulation of the radar pulse are phase shifts, the bandwidth needed for the amplifier is less than 5 MHz.

Another amplifier requirement is to produce low noise between pulses. To accomplish this simply, the transistor is biased below cutoff with $V_{gs}$=0.9V. This bias voltage turns the transistor off between pulses and allows for sufficient gain during the pulses.

V. SYSTEM MEASUREMENTS

Final tests involved the whole system including both the transmitter and receiver. The radar was tested for compliance with ITU requirements [32] and the United States requirements [33]. Pulse shaping is used to limit the bandwidth of the transmit pulse to the requirement of a -20 dB bandwidth of 2 MHz. There are a number of frequency-dependent components that aid in filtering the
449 MHz pulse bandwidth to the -20 and -40 dB levels and its harmonics down to the -60 dB levels (e.g. the circulators and splitters). All other requirements such as side-lobe suppression, frequency tolerance, and peak EIRP are also satisfied.

Another significant test is a blanker delay test. In a pulsed radar, the blanker switches off the receive signal to the A/D converter during the transmit pulse. Connecting the A/D converter to the receiver during the radar pulse will saturate the A/D converter. This saturation takes a longer time to recover from than the time it takes to switch the blanker. For this test, the system is run normally as a radar during a period with good atmospheric SNR in the range below 1 km and little or no RFI. Every 2 minutes, the blanker turn-off time is adjusted using the profiler control software and a 0 dB SNR level is collected for each blanker turn-off time. Figure 15 shows the data from this test. Note that 0 dB SNR was a level chosen for this test. The system is able to compute winds below 0 dB SNR using additional averaging. The goal of this test is to find the optimum blanker turn-off time. As shown in Figure 15, the SNR at the lower range gates can be significantly improved with the optimum blanker turn-off time.

A. Wind Data from Boulder, Colorado

As a confirmation of the performance of the new radar, the new wind profiler radar was first operated during prototype tests at the NCAR Foothills Laboratory in Boulder, CO. Figure 16 shows data from October 23rd, 2010. Precipitation can be seen in the data at 9UT and between 15 and 18UT as indicated by the downward vertical velocities.

The top plot in Figure 16 shows Signal to Noise ratio. Note that SNR is decreased below 1 km because of ground clutter and antenna ringing. SNR is higher during the precipitation events at 9UT and 15-18UT. The middle plot shows the vertical velocity. The vertical velocity is computed directly from the Doppler shift of the return signal. The bottom plot shows horizontal wind barb data. These horizontal winds are computed by cross-correlations between the receivers using the spaced antenna method. The wind direction is indicated by position of the barb (e.g. a barb pointed downwards with the tail straight up is a wind from the North). The wind velocity is indicated by the number of lines at the tail of the barb and also by the color code.

B. Wind Data from West Jordan, Utah

After initial prototype tests in Boulder, the radar was transported and deployed as part of the Persistent Cold Air Pool Study (PCAPS) in West Jordan, Utah, south of Salt Lake City. A commercially available 915 MHz wind profiler was also located at this site to allow comparison of the data between the two profilers. A nearby site contained a SODAR and radiosonde launch site for additional comparisons. Data for all systems at PCAPS was collected from 15 November 2010 until 15 February 2011.

Figures 17 and 18 show a comparison of data from both Wind Profilers. The SNR and wind data are quite similar. One issue that can be seen in the raw signal data is the presence of RFI. Because wind profiler radars typically have low SNR, RFI is a common problem [8]. RFI is seen in the data as a constant signal return from all range gates. Because of the position of the
antennas, each antenna sees a different level of RFI. As an example, RFI is seen in Figure 17, Signal Channel 2 from 12-21UT. The FCA processing algorithm is normally able to detect winds in the presence of RFI. Because snow is the only expected precipitation during this project, a modified Butterworth filter rejects all vertical Doppler velocities greater than 5 m/s. This filter can be modified for other precipitation. RFI can still affect the SNR as seen at 06UT. The RFI sources are usually communication and pager sources at frequencies such as 450 and 451 MHz.

A precipitation event on 9 January 2011 is seen in the data in Figures 17 and 18 between 05 and 10UT. The precipitation is indicated by the high SNR signals. A comparison of the 449 MHz and 915 MHz profiler data shows that the 449 MHz radar is able to sense winds at a higher altitude than the 915 MHz profiler. Higher height coverage can be attributed to the higher transmit power (500W peak for the 915 MHz system vs. 3000W peak for the 449 MHz system). Increased sensitivity can also be explained by the wavelength dependence of the effective aperture, which is directly related to SNR. The effective aperture of the 449 MHz system was calculated using an antenna gain simulation to be 10.5 m\(^2\), and for the 915 MHz system it is 3.4 m\(^2\). The difference in aperture provides about 5 dB of difference in SNR. These calculations assume no losses from the antenna feed networks and 100% antenna efficiency. The clear air backscatter cross section is only weakly dependent on wavelength \((\lambda^{1/3})\), so this provides about 1 dB of SNR difference between 449 and 915 MHz. total SNR improvement is 13 dB. Given that the spaced antenna method has about a 10 dB decrease in SNR, then there is a 3 dB net improvement, this is consistent with the data. The data also shows that the 449 MHz system is able to sense winds down to the 500m level, while the 915 MHz system can sense down to the 200-300m level. The reason for this difference is that short transmit pulses were not yet implemented in the profiler control software, this is planned in future work.

VI. CONCLUSIONS

A new 449 MHz radar wind profiler has been demonstrated. This system will continue to be deployed in support of NCAR scientific field projects. A transmit section and three receivers were designed and tested. With low noise figure and adequate gain these receivers were tested to detect return signals with power levels down to -150 dBm.

An 80W LDMOS amplifier with PAE of 65% and gain of 13 dB was demonstrated at 449 MHz and used as a driver amplifier. Three 1 kW amplifiers based on LDMOS transistor technology were successfully combined to operate as a high power transmitter for this application. The 1 kW amplifiers have a PAE of 53.8% and a gain of 18.5 dB for operation up to 10% duty cycle.

Future plans include the integration of a power amplifier with phase/amplitude adjustment behind each antenna element. This will require evaluation of individual 1 kW amplifiers for phase stability vs. temperature. Higher power versions of this system with larger 7 and 19 hexagon antennas (126 and 342 elements) are also planned for the future.
ACKNOWLEDGEMENTS

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REFERENCES


Fig 1. (a) Photograph of the 449 MHz radar wind profiler at Salt Lake City, Utah in November 2010. The receivers are located under the antenna panels. The transmitter and data system are located inside the trailer. (b) Block diagram of the 449 MHz radar wind profiler.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
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<tbody>
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<td>Radar Frequency</td>
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<tr>
<td>Transmit Power</td>
<td>3 1 kW modules</td>
</tr>
<tr>
<td>Inter-pulse Period</td>
<td>50 µs</td>
</tr>
<tr>
<td>Pulse coding</td>
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<td>Maximum Range</td>
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Minimum Range 200 m
Range Resolution 150 m
Transmit Array Gain 24.8 dBi
Receiver Array Gain 19 dBi
Receiver Noise Figure 2 dB

Fig 2. (a) 18-Element hexagonal circular patch antenna array for 449 MHz. (b) Antenna pattern simulation. Scale is in dB with reference to main beam. Edge of circle is horizon, center is zenith. Sidelobes at horizon are more than 35 dB less than main beam.
Fig 3. (a) Photograph of a 6-element subarray from Figure 2(a), showing the patch probe feed position. A 25-mm thick polystyrene layer is epoxied on top of the panel to protect against moisture. (b) Measured and simulated $|S_{11}|$ of one of the circular patch antennas.
Fig 4. (a) Procedure for measuring 18-element antenna array sidelobes. Antenna is rotated about its axis for each elevation angle. (b) Single hexagon, 18-element array measured antenna sidelobe levels (receiver polarization oriented horizontally). Main beam is pointing at zenith.
**Fig 5.** (a) 54-Element hexagonal circular patch antenna array for 449 MHz with connections to the transmitter and receivers as shown. The center to center spacing is $D = 231\text{cm}$. (b) Antenna pattern simulation. Scale is in dB with reference to main beam. Edge of circle is horizon, center is zenith. Sidelobes at horizon are more than 35 dB less than main beam.

![Antenna Pattern Simulation](image)

**Fig 6.** Amplitude check of each element of an 18-Element array using a VNA. A theoretical 54-way split is -17 dB.

![Amplitude Check](image)

**Fig 7.** Block diagram illustrating one of the three receiver channels.

![Block Diagram](image)

**Fig 8.** Diagram illustrating the transmit section of the radar. The 1 kW High Power Amplifiers are driven by 80W and 10W stages.

![Transmit Section](image)
Fig 9. Photograph of 80W amplifier based on a Freescale MRF5S9070N transistor. The transistor is mounted underneath a clamp in the center of the circuit.

Fig 10. Measured 80W LDMOS Pout, gain, and PAE vs. input power. The operating point is Pin = 30 dBm, Pout = 49.1 dBm CW, PAE = 68%, gain = 19 dB.
Fig 11. (a) Schematic of the 1 kW (peak) LDMOS pulse amplifier. Coaxial baluns drive the push-pull transistor pair and combine the output.
(b) Output matching network of the 1 kW LDMOS amplifier using low impedance microstrip transmission lines.

Fig 12. Photograph of the 1-kW (peak) pulse amplifier module based on the Freescale MRF6VP41KHR6 LDMOS transistor in push-pull configuration.
Fig 13. 1 kW LDMOS Amplifier performance vs Input Power. Pulsed operation, 10% duty cycle. Best amplifier operating point is Pin (peak) = 41.5 dBm, Pout (peak) = 60.1 dBm, PAE = 53.8%, gain = 18.5 dB.

Fig 14. Measured 1 kW LDMOS Amplifier performance vs. frequency. Pulsed operation, 10% duty cycle. Amplifier has output power above 59 dBm and efficiency above 44% from 440 to 455 MHz. Input power was reduced to 40.38 dBm for this sweep.

Fig 15. Measurement of the lowest range gate SNR while varying the A/D converter blanker turn-off time. The SNR of the lowest range gates is affected by the time that the blanker deactivates.
Fig 16. Boulder, Colorado wind profiler data on 23 October 2010 from 449 MHz system using 3 combined 1kW amplifiers. The plots show altitude vs. time with the color code indicating SNR for the top plot and velocity in m/s for the bottom two plots. Precipitation is indicated by strong SNR and negative (downward) vertical velocities.
Fig 17. 449 MHz spaced antenna wind profiler data at West Jordan, Utah on 9 January 2011. (a) Three channel raw signal level vs. altitude and time in dB. (b) SNR (dB) and horizontal winds (m/s).

Fig 18. Data from commercial 915 MHz Doppler beam steering wind profiler located at the same site. SNR (top) and winds (bottom) during the same time period.
Brad Lindseth (S’95–M’06–SM’10) received the B.S. degree in electrical engineering from Washington University in St. Louis. After working as an engineer on MRI and NMR systems from 1999-2002, he joined the Vestibular Neuroscience Laboratory at Washington University and received an M.S. in electrical engineering in 2005 with a thesis on biological magnetic field sensors in pigeons. From 2005 to 2007 he was with the Atomic Devices and Instrumentation Group at the National Institute for Standards and Technology (NIST) in Boulder, Colorado where he worked on chip scale magnetometers. In 2007 he joined the Earth Observing Laboratory at the National Center for Atmospheric Research (NCAR) in Boulder, Colorado where he is currently an Electrical Engineer working on 449 and 915 MHz wind profiler radars. He is currently working toward a Ph.D. degree in electrical engineering from the University of Colorado at Boulder. He was awarded first place in the 2011 NXP High Power RF Design Challenge for a 2 kW radar pulse amplifier.

William O.J. Brown studied at the University of Canterbury in Christchurch, New Zealand, receiving B.Sc. (1984), M.Sc. (Hons, 1986), and Ph.D. (1993) degrees in the Department of Physics and Astrophysics. He did post-doctoral fellowships at Kyoto University’s Radio Atmospheric Science Center in Uji, Japan and at the Department of Atmospheric and Oceanic Sciences at McGill University in Montreal, Quebec, Canada. Since 1998 he has been at the National Center for Atmospheric Research (NCAR) in Boulder, Colorado. He is a Project Scientist in the Earth Observing Laboratory (EOL) and leads the Atmospheric Profiling Group in the In-situ-Sensing Facility (ISF). This group deploys ISS (Integrated Sounding System) and GAUS (GPS Advanced Upper-air Sounding) facilities including wind profilers, radiosondes, and other sensors for the meteorological research community at locations all around the world. His research focus is on development and applications of wind profiler radars.

Zoya Popović (S’86–M’90–SM’99–F’02) received the Dipl.Ing. degree from the University of Belgrade, Serbia, Yugoslavia, in 1985, and the Ph.D. degree from the California Institute of Technology, Pasadena, in 1990. Since 1990, she has been with the University of Colorado at Boulder, where she is currently a Distinguished Professor and holds the Hudson Moore Jr. Chair in the department of Electrical, Computer and Energy Engineering. In 2001, she was a Visiting Professor with the Technical University of Munich, Munich, Germany. Since 1991, she has graduated 40 Ph.D. students. Her research interests include high-efficiency, low-noise, and broadband microwave and millimeter-wave circuits, quasi-optical millimeter-wave techniques for imaging, smart and multibeam antenna arrays, intelligent RF front ends, and wireless powering for batteryless sensors.

Prof. Popovic was the recipient of the 1993 and 2006 Microwave Prizes presented by the IEEE Microwave Theory and Techniques Society (IEEE MTT-S) for the best journal papers, and received the 1996 URSI Issac Koga Gold Medal. In 1997, Eta Kappa Nu students chose her as a Professor of the Year. She was the recipient of a 2000 Humboldt Research Award for Senior U.S. Scientists from the German Alexander von Humboldt Stiftung. She was elected a Foreign Member of the Serbian Academy of Sciences and Arts in 2006. She was also the recipient of the 2001 Hewlett-Packard (HP)/American Society for Engineering Education(ASEE) Terman Medal for combined teaching and research excellence.