Dual-Polarized Array for Signal-Processing Applications in Wireless Communications

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***Abstract*—A novel dual-polarized antenna array designed for a spatial division multiple access (SDMA) system working in the 1850–1990-MHz band is designed and built. The antenna is designed to have similar element patterns, and measurements of S-parameters and radiation patterns are presented. The array signal processing performance of the array using all elements and with no compensation for mutual coupling or differences in element patterns is studied through direction-of-arrival (DOA) estimation using total least squares estimation of signal parameters via rotational invariance techniques (TLS-ESPRIT). The results show that the accuracy of the DOA estimates is quite acceptable for wireless communication applications.**

***Index Terms*—Antenna arrays, signal processing antennas.**

I. Introduction

During the last few years there has been an enormous expansion in wireless communication systems and the number of people using their services. This expansion has resulted in a demand for increased system capacity, but since there is only a limited bandwidth available for commercial communications, the capacity issue is not easily solved. Another problem is the radio channel. In mobile communications, fading and interference limits both the quality and the capacity of the systems.

In order to fulfill these increasing demands on capacity and coverage it has recently been suggested to employ smart or intelligent antennas at the base station [1]. Such an antenna consists of an array of horizontally distributed sensors that can transmit or receive energy independently from one another. By using signal processing algorithms to continuously distinguish between desired signals, multipath and interfering signals it is possible to smoothly track users with main lobes and interfering signals with nulls and thereby constantly maximizing the carrier-to-interference (C/I) ratio. One may describe this as multiplexing channels in the spatial dimension similar to what is done in the frequency and time dimensions in frequency division multiple access (FDMA) and time DMA (TDMA) systems. The method is often referred to as spatial division multiple access (SDMA) and may, in high-mobility systems, significantly increase system capacity [2].

If dual-polarized sensors are used so that two orthogonal polarizations are received, this may enhance the array signal processing in two different ways. One is that the number of signals available is increased which may further improve C/I after signal processing [3]. Another simpler alternative is to use a post-detection maximum-ratio combination of the signals from the two sensors. This method is similar to the one frequently used in today's polarization diversity systems for mobile communication base stations [4]. Because of the nonoptimal combining used in these commercial systems, it is often preferable to receive equal mean power from the two sensors [5]. Since the incident signals from a mobile station exhibit a vertical polarization bias (e.g., [6]–[8]), this is one reason why polarizations of slant $\pm 45°$ to vertical most often are used. Another reason is that slanted $\pm 45°$ elements will provide identical azimuth patterns. Vertical and horizontal polarizations would give different beamwidths. In the antenna array to be presented the choice of $\pm 45°$ polarizations also simplifies the feed-network layout.

A problem associated with antenna arrays is mutual coupling, which changes the element patterns and, thus, degrades the array signal processing [9], [10]. The coupling can be compensated for by an analog low-loss network [11] or by matrix multiplication in the digital processing of the signals [12]. An alternative approach is to minimize the mutual coupling when designing the antenna.

In this paper, we present a novel dual-polarized antenna array designed for a SDMA system working in the personal communication system (PCS) band (1850–1990 MHz). The antenna is designed to have similar element patterns and measurements of S-parameters and radiation patterns are presented. We study the signal processing performance of an actual array with no compensation for mutual coupling and differences in element patterns. More specifically, we study the error in direction-of-arrival (DOA) estimation using total least squares estimation of signal parameters via rotational invariance techniques (TLS-ESPRIT). Previous work has analyzed the effect of using dummy elements at the edge of the array on the signal processing performance [9]. Here we choose to use all the array sensors, which is the most effective if the total array width is limited.

II. Antenna Design

Fig. 1 shows the antenna array. The dimensions are $1500 \times 1000 \times 40$ mm. It consists of 12 identical sensors for the PCS band 1850–1990 MHz, forming an array along the horizontal x axis. Each of the sensors consists of a 12–element array along the z axis. These elongated antenna sensors are referred

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Fig. 1. The 12 × 12 dual-polarized array.

TABLE I
NOMINAL EXCITATIONS OF THE COLUMN ELEMENTS LISTED FROM TOP TO BOTTOM

Element	1	2	3	4	5	6	7	8	9	10	11	12
Amplitude	1.00	0.86	1.19	1.25	1.11	1.23	0.94	1.25	0.95	0.65	0.63	1.05
Phase °	0	-25.4	-46.9	-74.4	-85.3	-82.1	-90.3	-88.2	-125	-93	-135.7	-142.1

to as columns and they are designed to provide high-element gain and a suitable vertical pattern. The column spacing is 80 mm (0.512λ at the center frequency 1920 MHz). This spacing needs to be small to reduce nonambiguity in the signal processing; a 0.5λ spacing at 1990 MHz would, of course, have been ideal, but this would not have accommodated the feed network. The array of the 12 columns provides the opportunity of beamforming in the horizontal plane.

The antenna is designed for dual-linear polarization and each column has two channels, one for $+45°$ and one for $-45°$ slant to vertical. We have a total of 24 channels, which sample orthogonal components of the incident field at 12 different locations. To reduce the mutual coupling, there are corrugations between the columns.

A. Design Considerations

In the design of the antenna columns we have the following goals.

Horizontal Pattern: The major concern with the horizontal patterns of the sensors is to keep them as similar as possible and preferably to have a fairly large half-power beamwidth in

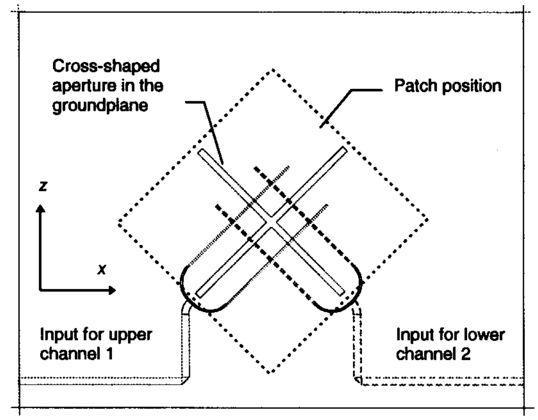

Fig. 2. Feed layout for a single dual-polarized element.

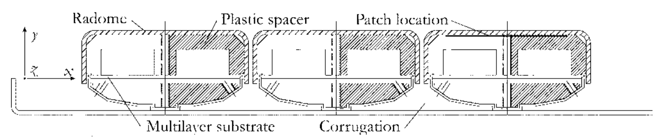

Fig. 3. Cross section of three antenna columns mounted on a common backplane.

order to cover the typical 120° wide cells used in cellular systems. To achieve similar horizontal patterns for interior and edge sensors in the array, the mutual coupling between the columns must be reduced or dummy sensors placed at the edges. Since dummy sensors increase cost and size, it is preferable to try to reduce the coupling.

Elevation Pattern: The array has 12 elements with a spacing of 125 mm or 0.8λ at 1920 MHz. The elevation pattern is designed for beam peak at $-2°$ elevation. We also want to achieve some null-fill and suppression of the first upper sidelobe while maintaining a high directivity. The choosen element excitations are given in Table I. Since high gain is important, there is a tradeoff between how much amplitude taper we can allow and the pattern characteristics. In the pattern synthesis we allowed the beamwidth to increase from the minimum 4.7–5°.

B. Sensor Geometry

The sensors are dual-polarized aperture-coupled patch sensors similar to the designs in [13] and [14]. A multilayer circuit board is used where the feed network for channel 1 is located above the ground plane and consequently the feed network for the orthogonal polarization of channel 2 below the ground plane. Fig. 2 shows part of the two feed networks, the cross-shaped aperture and the patch. The ground plane with the aperture is located between the substrates. The main advantage of this configuration is that we use the limited width of the columns in an effective manner; the possibility of coupling between the feed networks of the two channels is also eliminated.

C. Corrugations

Fig. 3 shows a cross section of three antenna columns. The multilayer board is mounted in a aluminum reflector.

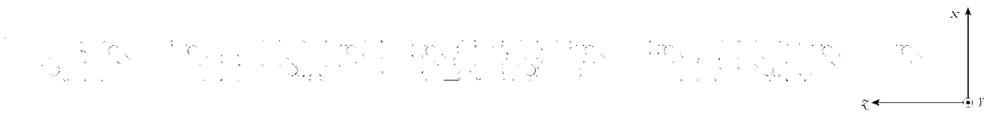

Fig. 4. The feed network of one channel in one column. Also shown are the cross-shaped apertures in the common ground plane of the two channels.

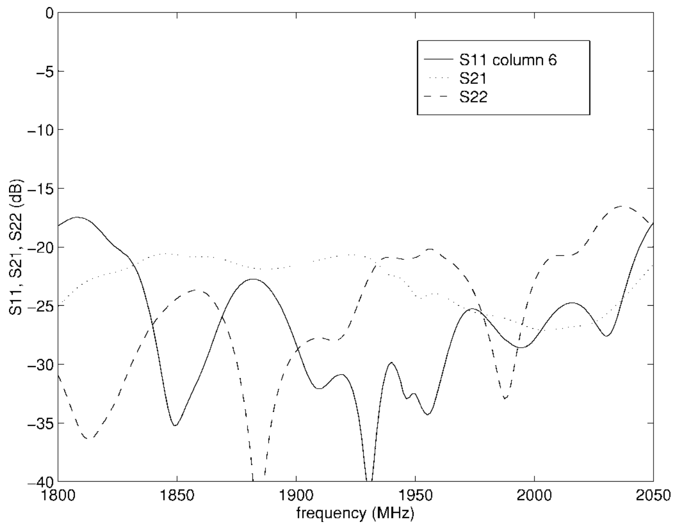

Fig. 5. S-parameters for column 6 in the array environment.

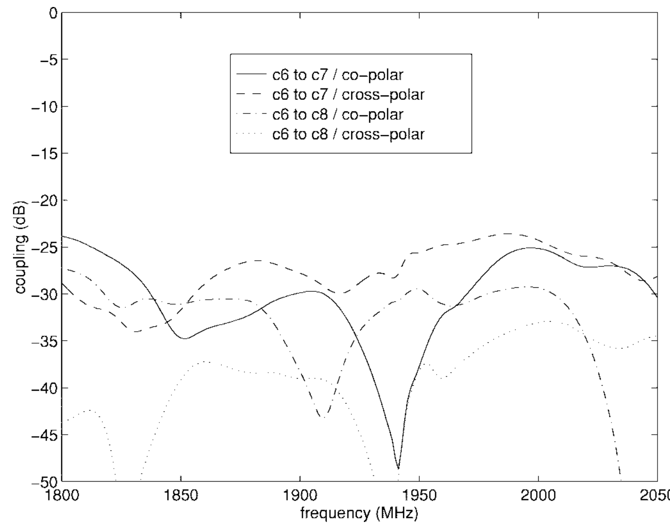

Fig. 6. Mutual coupling between column 6 and columns 7 and 8.

A radome with patches on the inside is attached to each reflector, thus forming an antenna column. Each column is then mounted on a larger back plane where the spacing between the columns provide an approximately $\lambda/4$ deep corrugation. The corrugation prevents TM_y waves from propagating across the array and, therefore, reduces the coupling. In theory, the TE_y waves are not affected by such corrugations, but the such waves do not propagate along the solid ground plane and are, therefore, of less concern. The corrugation suppresses a TM_y surface wave propagating normal to the corrugation, but not a wave at oblique angles. In general, the corrugation should be $\lambda/(4\cos\theta)$ deep, but we expect the coupling to be strongest between adjacent elements and, therefore, we chose to make the corrugation approximately $\lambda/4$. The exact shape of the corrugation is determined using a full-wave moment method [15] calculating the coupling between two longitudinal slots on an conducting surface of constant cross section along the z axis. The calculations predict a 5-dB reduction due to the corrugation of the E-plane coupling between the patches.

D. Feed Network

The feed network is illustrated in Fig. 4. It is identical for both polarizations and consists of microstrip lines etched on an Arlon DiClad 527 substrate with reactive power dividers. The entire feed network is designed and optimized in HP-EEsof. The feed point is at the center of the column and at the first level the power is divided into three branches. At the next level there are three two-way power dividers and at the final stage we use six two-way power dividers to yield the desired element excitation. Due to the limited space on the board, five of the 12 slots are fed from the opposite compared to the other. This causes a 180° phase shift that is used to reduce the length for the lines and the circuit crowding. This results in an increased frequency dependence of the elevation pattern, but the limited PCS bandwidth of only 7.3% makes this a minor concern. The microstrip feed network connects to an output type-N connector via a RG-400 coaxial cable.

III. Antenna Performance

A. VSWR and Isolation

The aperture coupled patch elements used in the array are possible to match to a 23–25 dB return loss. In the array environment, however, mutual coupling between the sensors within a column changes the input impedance. The limited width of each antenna combined with the multilayer structure makes experiments and tuning cumbersome and the patch elements are matched only to a 18–20 dB return loss. The transition between the input coaxial cable and the microstrip network also gives a 28–30 dB reflection at 1.9 GHz in addition to the mismatch caused by possible error in the impedance of the RG-400 cable, rated $50 \pm 1.5\Omega$. The resulting typical return loss for a column in the complete array environment is 20 dB for channel 1 (upper layer) and 15 dB for channel 2 (lower layer). Fig. 5 shows the measured S-parameters for columns 6 and 7.

The measured isolation within a single column not mounted in the array is more than 25 dB, but in the array environment the isolation decreases to 20 dB as seen in Fig. 6. A possible source of the increased coupling between the polarization channels is the corrugation aligned vertically, i.e., 45° to the polarization directions.

Fig. 6 shows the measured mutual coupling between column 6 and two of its closest neighbors, columns 7 and 8. Notice

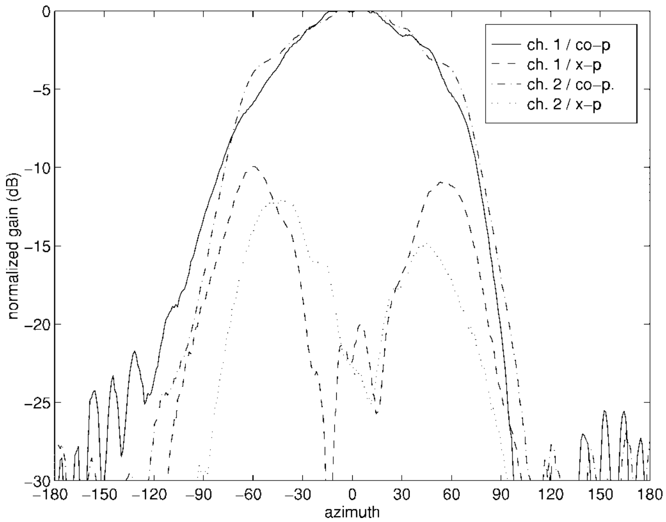

Fig. 7. Element patterns of the two channels of column 6.

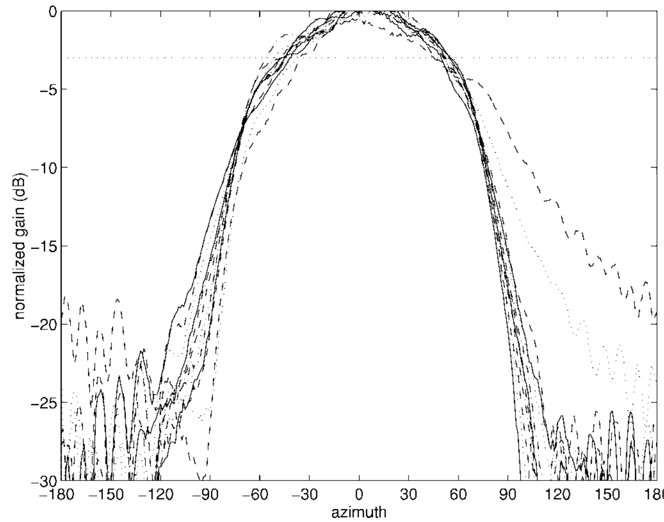

Fig. 8. Copolar element patterns of channel 1 of all 12 columns.

that the copolar coupling (meaning coupling from one $+45°$ channel to another $+45°$ channel) is the weaker between column 6 and 7, whereas it is the stronger coupling between columns 6 and 8. The coupling is less than -23 dB, but we still expect disturbed element patterns.

B. Radiation Patterns

Fig. 7 shows the measured copolar and cross-polar horizontal patterns of column 6 in the array environment.[1] The cross-polar levels are quite high and this is probably caused primarily by the corrugations, but also by the feed network of channel 1, which is located above the ground plane. Off boresight, we have an additional geometrical error since the copolarization is $\pm 45°$ to vertical.

In Fig. 8, we see the horizontal plane element patterns of channel 1 of the 12 columns. The element gain varies some 2 dB from element to element over the $\pm 60°$ sector.[2] There are no blind spots.

As discussed above, we did not design the vertical array for maximum directivity. Additional errors in the feed network resulted in a vertical half-power beamwidth of 5.6°. Fig. 9 shows the ideal pattern and the measured vertical co-polar patterns of the two channels of column 6. Just as for the input voltage standing wave ratio (VSWR), we see higher with sidelobes for channel 2. The null fill still extends to the first two nulls for both channels. The gain varies between different columns and frequencies due to the mutual coupling and the average gain for column 6 is 16 dB. The performances of the other columns are similar.

[1] The copolar and cross-polar patterns are measured according to the third definition of Ludwig [16], i.e., a fixed transmit antenna is used and the antenna under test is rotated. The radiated field of the transmit antenna matches the antenna under test at boresight (or beam peak) and the response at different angles of rotation yields the copolar pattern. The cross-polar pattern is measured with the transmit antenna shifted so that a field orthogonal to the previous is radiated toward the antenna under test.

[2] Fig. 8 shows a peculiar behavior between 75° and 180° azimuth for two columns; these are the edge columns 11 and 12. A corresponding behavior is not observed for columns 1 and 2 at the other edge and, therefore, we attribute it to mechanical defects or measurement errors.

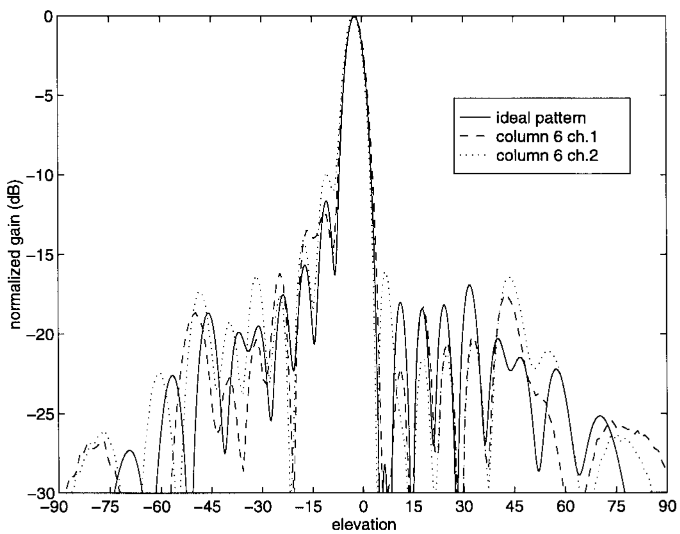

Fig. 9. Calculated vertical pattern from the ideal excitation and measured patterns of channel 1 and channel 2 of column 6.

IV. A SIGNAL-PROCESSING APPLICATION

The antenna is intended to be used as a base-station antenna in a wireless communication system—namely PCS. We see in Fig. 9 that the element patterns differ significantly from channel to channel and this will degrade the performance of the signal processing algorithms and, consequently, the overall system performance. As a figure of merit, we choose to characterize the antenna performance in terms of error in the estimate of the DOA. The DOA estimator used in the simulations is TLS-ESPRIT [17]. We generate signal data for 12 copolar channels using the measured far-field response in amplitude and phase.

Three equally strong uncorrelated signals centered[3] at 1850 MHz are impinging on the array. The true DOA's are $-30°$, $-10°$, and 20°, respectively. A batch of $N = 128$ snapshots

[3] We assume narrow-band signals in the sense that their wavelength is equal and practically constant over time, although the actual signal values at a given time are uncorrelated. This relates the time between the snapshots and the bandwidth of the channel.

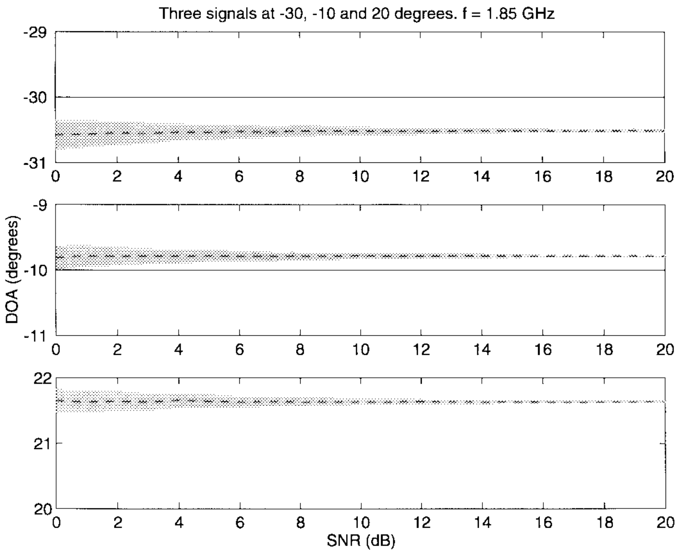

Fig. 10. DOA estimates at 1850 MHz for three signals showing biased error (dashed line) and the rms error at different SNR; 128 snapshots and 256 independent trials.

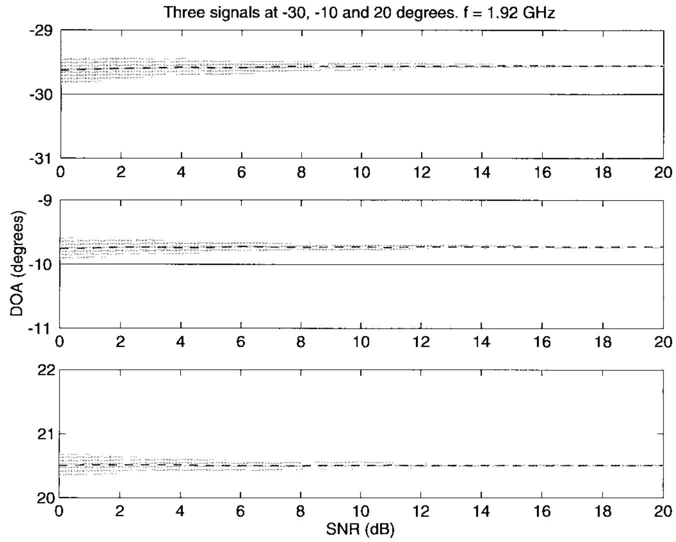

Fig. 11. DOA estimates at 1920 MHz for three signals showing biased error (dashed line) and the rms error at a different SNR; 128 snapshots and 256 independent trials.

is generated for these three signals. Each snapshot contains the signal output of each of the 12 sensors based on the measured far-field pattern and additive white Gaussian noise (AWGN). Once the data is generated, we use TLS-ESPRIT to estimate the DOA's. Fig. 10 shows the results of a Monte Carlo simulation involving 256 independent trials for each signal-to-noise ratio (SNR). The SNR is normalized to the gain at boresight of channel 1. The dashed line is the mean of the estimated DOA and the root mean square (rms) error is indicated by the gray area. We conclude that the imperfections in the array induce a biased error in the estimates which dominates the rms error at all SNR > 0 dB. The biased error is less than 1.7° for all three signals.

In a mobile communication system, we generally only have information of the location of the mobile station available at the up-link frequency. Since the antenna array has frequency dependent patterns, an error is induced if we use the up-link data for the down-link beamforming. We can expect this error to depend on the specific algorithm used. Therefore, here we seek to characterize the antenna array deficiency in terms of the difference in the DOA estimates at the two frequencies. In Fig. 11 we see the result of the same simulation as discussed above but now the frequency is 1920 MHz. The difference in the DOA estimates at the two frequencies is less than 1.2°.

V. Discussion and Conclusions

A dual-polarized array with an element spacing of 0.512λ has been presented. Although the separation between the aperture fed patches is only 0.27λ, the coupling is typically below -25 dB. Comparing with the numerical result of typically -20 dB coupling for a similar aperture couple patch geometry in [18], we conclude that the corrugations in the array works to our advantage. A disadvantage is an increased cross polarization. The thickness of the array is only 40 mm, less than $\lambda/4$. Using a two-substrate feed network, we have managed to fit the whole feed network of 12 dual-polarized sensors onto a board only 70 $\times$ 1500 mm in size.

In spite of the variation seen in the element patterns, quite acceptable accuracy is found for direction of arrival estimates using TLS-ESPRIT. In a three-signal scenario, the sum of biased and RMS error is less than 2° for a sensor SNR $= 0$ dB. More importantly, the DOA estimates differ by less than 1.2° when we shift the frequency 70 MHz. This indicates that the antenna is suitable for use in a mobile communication system.

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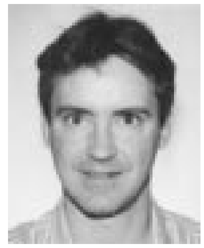

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