

Results: The simulated and measured VSWR of the proposed antenna are shown in Fig. 2. By carefully designing the matching network, very good impedance performance is obtained. The 2:1 VSWR is obtained from 3.9 to 6.9 GHz. The bandwidth is 3 GHz, or 55.6%. The measured axial ratio and gain of the proposed antenna are shown in Fig. 3. The results show that from 4.8 to 6.0 GHz, less than 3 dB axial ratio at boresight for the forward direction (from the ground plane) is achievable. The bandwidth is 1.2 GHz or 22.2%. The gain is -0.5 to 1.4 dBic in the above frequency range. The measured axial ratio in the backward direction (from the opposite plane which includes the feed network) is 0.5 to 1.0 dB poorer because of the radiation of the feed network strip-line.

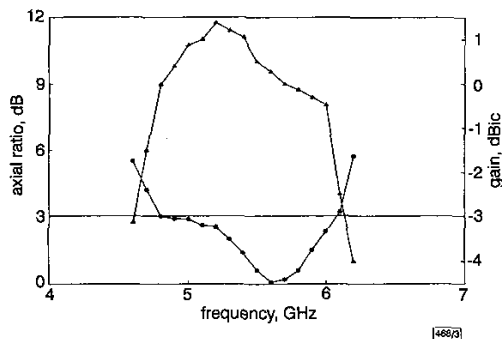


Fig. 3 Measured axial ratio and gain

▲ gain
● axial ratio

Conclusions: Our proposed microstrip three-stub hybrid coupler feed slot loop antenna shows an extremely broadband impedance match and axial ratio performance. This antenna also has bi-directional radiation which is very useful in some wireless communications and RF tagging applications. In addition, it is a very low cost design suitable for commercial implementation.

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Propagation measurement using antenna array

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The design and construction of a 2.4 GHz antenna array suitable for measurement of the time, angle, and complex amplitude of path arrivals in an indoor radio channel are described. Calibration of the array is facilitated with the aid of an anechoic chamber. An optimal least-squares processor is derived, which compensates for systematic calibration errors. Early measurement results are presented, and future directions of the research are indicated.

Introduction: Emerging applications are exploiting the spatial properties of wireless communications to deliver new features and enhanced performance. These applications are perhaps best studied and optimised on the basis of a sound understanding of the spatial and temporal characteristics of the channel. Until recently, few procedures have been available which are capable of providing simultaneous estimates of the times, angles and amplitudes of arriving paths. The procedure in [1] may be viewed as an extension to the frequency-domain channel sounder presented in [3]. This new system uses a 60 cm parabolic dish, mounted on top of a computer-controlled turntable. A complete measurement consists of 180 frequency sweeps, between which the antenna is rotated by 2°. As compared with the rotating dish concept, our array approach features a simplified setup, rapid (< 1 s) measurement, and the option of employing super-resolution post-processing.

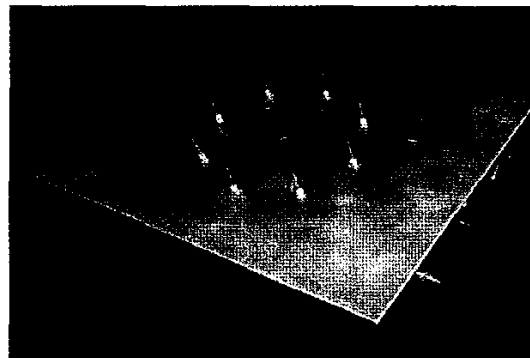


Fig. 1 Eight-element circular array

Array construction: The authors constructed the antenna array shown in Fig. 1. The array consists of eight nominally identical quarter-wave monopole elements, separated by one-third of a wavelength around a circle. The signal received at each element is fed to an eight-channel switch by means of a short run of semi-rigid coaxial cable. As in [1, 3], measurements are conducted in the frequency domain, using a standard vector network analyser. When configured for 101 points and with a predetection bandwidth of 3 kHz, a total of 750 ms is required to measure and store all eight channels.

Array calibration: A series of calibration measurements was conducted using a compact indoor range. Between each measurement, the array was rotated by 5.625°, relative to its previous orientation. Each measurement produced a 101 × 8 matrix of the form

$$U_m = \begin{bmatrix} u_{1,1,m} & u_{1,2,m} & \dots & u_{1,8,m} \\ u_{2,1,m} & u_{2,2,m} & \dots & u_{2,8,m} \\ \vdots & \vdots & \ddots & \vdots \\ u_{101,1,m} & u_{101,2,m} & \dots & u_{101,8,m} \end{bmatrix} \quad (1)$$

where $m = 1, 2, \dots, 64$. The indices are chosen such that $m = 1$ corresponds to 0°, $m = 2$ corresponds to 5.625°, and so forth.

Spatial-temporal signal processing: Numerous algorithms have been proposed for the angle-of-arrival estimation of signals contained in an array [2]. These algorithms are based on the assumptions that (i) the number of arrivals is known precisely and is strictly less than the number of elements, (ii) the arrivals are uncorrelated or may be made so by averaging techniques, and (iii) the array factor is ideal, such as that of a perfect ULA. The present case violates all of these conditions. In addition, the existing techniques do not exploit *a priori* knowledge of the signal structure, which in our case is provided in the form of the calibration matrices of eqn. 1. For these reasons, we have developed a new algorithm which may be viewed as an extension of conventional beamforming techniques. Though inferior to subspace and model-based approaches [2], at least so far as angular resolution is concerned, beamforming is far simpler, and adequate for the task at hand. We shall consider algorithms for super-resolution processing in a forthcoming publication.

One possible approach to beamformer design involves the estimation of the array factor directly from the measurements of eqn. 1, followed by the use of a standard beamformer

configuration [2]. We have developed an alternative approach, wherein the problem is interpreted as the design of a least-squares spatial filter. This approach has numerous benefits, including improved resolution, lower sidelobes, and the automatic removal of systematic errors in the array response. In our procedure, we form the cost function

$$f(\mathbf{w}_m) = \sum_{j \neq m} \sum_{l=1}^{101} \left(\sum_{n=1}^8 u_{l,n,j} w_{n,m} \right) \left(\sum_{n=1}^8 u_{l,n,j}^* w_{n,m}^* \right) \quad (2)$$

which we minimise, subject to the additional constraint

$$c(\mathbf{w}_m) = \sum_{l=1}^{101} \left(\sum_{n=1}^8 u_{l,n,m} w_{n,m} \right) \left(\sum_{n=1}^8 u_{l,n,m}^* w_{n,m}^* \right) - 1 = 0 \quad (3)$$

When applied to the frequency-domain calibration data, satisfaction of eqn. 2 results in a set of taps which derive minimal energy from all directions except the desired. The constraint of eqn. 3 ensures that the energy from the desired direction is exactly normalised. Using the method of complex Lagrange multipliers the cost and constraint equations are combined to form the *adjoint equation*:

$$\frac{\partial f(\mathbf{w}_m)}{\partial \mathbf{w}_m^*} + \lambda \frac{\partial c(\mathbf{w}_m)}{\partial \mathbf{w}_m^*} = 0 \quad (4)$$

Performing the partial differentiation over each element of the complex-valued tap weight vector in eqns. 3 and 4 produces a system of eight equations:

$$F_k(\mathbf{w}_m, \lambda) = \sum_{j \neq m} \sum_{l=1}^{101} u_{l,j,k} \sum_{n=1}^8 u_{l,j,n}^* w_{n,m} + \lambda \sum_{l=1}^{101} u_{l,m,k}^* \sum_{n=1}^8 u_{l,m,n} w_{n,m} = 0 \quad (5)$$

where $k = 1, 2, \dots, 8$. Since the constraint equation must also be satisfied, we have

$$F_9(\mathbf{w}_m, \lambda) = \sum_{l=1}^{101} \left(\sum_{n=1}^8 u_{l,n,m} w_{n,m} \right) \left(\sum_{n=1}^8 u_{l,n,m}^* w_{n,m}^* \right) - 1 = 0 \quad (6)$$

Taken together, a solution to eqns. 5 and 6 provides the optimal taps for the array processor, as well as the Lagrange multiplier, λ . Although the eqn. 5 equations are linear in the independent variables, eqn. 6 is quadratic in the tap weights. As a result, standard linear algebraic techniques, such as Gaussian elimination, cannot be used to compute the solution immediately. Instead, we employ the complex Newton's method to solve for the variables iteratively. The residual error at a given iteration is taken as $f^H f$, where f is the 9×1 function vector $[F_1 F_2 \dots F_9]^T$. Starting with all taps initially set to 1, 6–8 iterations are generally required to reduce the residual below 10^{-14} . This process is repeated in order to develop 64 sets of optimal taps.

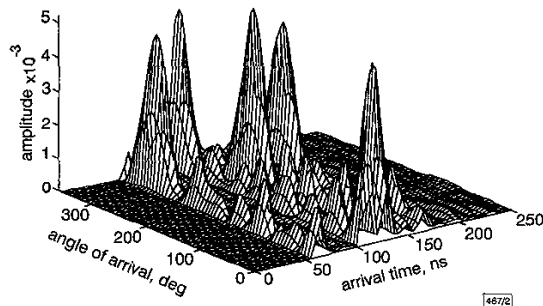


Fig. 2 Measured response for five discrete arrivals

Given an eight-channel measurement U , we form the spatial matrix

$$\mathbf{X} = [\mathbf{U}\mathbf{w}_1 \quad \mathbf{U}\mathbf{w}_2 \quad \dots \quad \mathbf{U}\mathbf{w}_{64}] \quad (7)$$

where each column of $\mathbf{X}$ gives the frequency response in a particular look direction. Each column of $\mathbf{X}$ is modified with a 101 point Hamming window, prior to inverse Fourier transformation. The result is the spatial-temporal matrix $\mathbf{Y}$, where the column index indicates the angle of arrival in 5.625° increments, and the row index indicates the time of arrival in 5ns increments. In Fig. 2, we show the result of a measurement, after spatial-temporal processing, for a case of five arrivals of identical amplitude. The path parameters are then found by identifying the local maxima of $\mathbf{Y}$, and taking the row and column indices as the time and angle of arrival, respectively.

Conclusions: We have presented an approach to the use of an antenna array for the joint spatial/temporal characterisation of an indoor radio channel. As compared with the rotating antenna concept, the array is far more compact, and delivers measurements much more rapidly. We have also presented a least-squares post-processing algorithm, which compensates for systematic errors while estimating the angle, time, and amplitude of path arrivals. In forthcoming work we shall discuss strategies for super-resolution path identification, as well as the results of a recent indoor measurement study.

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Reducing the influence of feed cables on small antenna measurements

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It is shown that in measurements of mobile handset antennas a cap on the coaxial feed cable inserted relatively close to the handset acts practically like an open termination to the induced currents on the surface of the phone chassis. Therefore, the measurement results with the cap are much closer to those of an isolated handset than without any countermeasures or with ferrite chokes.

Introduction: Antennas used in portable handsets are fed with the chassis of the handset acting as a ground reference. The handset though does not resemble an infinite ground plane forming a perfect mirror image of the antenna. Therefore, a phone with a monopole antenna resembles an asymmetric dipole driven by the transmitter unit inside the handset. Standard methods for antenna calibration often require the measurement of the complex transmission and reflection coefficients with the antenna under test (AUT) and a well known measurement antenna. Therefore, the AUT must be connected to a vector network analyser with a coaxial cable. This situation does not resemble the standard operation situation of a portable handset.

Two distinct phenomena can be observed that will affect the radiation characteristics of the AUT. First, the feed cable is placed in the near field of the antenna thus changing the resistive and reactive parts of the input impedance, and hereby also the radiation