MEASURING ANGLE-OF-ARRIVAL IN OVER OCEAN PROPAGATION EXPERIMENTS

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ABSTRACT

Radio propagation experiments are a useful tool when verifying propagation models and identifying anomalies in a practical link. This paper details a method to measurement amplitude profiles and angle-of-arrival in an over-ocean propagation experiment. A new technique for measuring the phase of radio signals received by incoherent detectors is described. Frequency drift and offsets, arising because the different receivers are incoherent, are compensated for by way of a common reference signal injected into each channel. The phase of the unknown signals are obtained by processing the down-converted and digitised waveforms. A novel technique based on an adaptive, discrete-time quadrature delay estimator (QDE) algorithm is used for this purpose. This algorithm is insensitive to variations in the amplitudes of the input signals, and does not require an accurate prior estimate of the frequency of the input sinusoids. This approach is shown to be an accurate, low cost alternative to conventional vector measurement techniques when used with large antenna arrays and is therefore well suited to fixed link, angle-of-arrival measurements.

INTRODUCTION

The ability to measure the arrival direction of a radio wave can be useful in various radio engineering applications. Finding the direction to an opposing force from intercepting radio transmissions from radar or communication equipment is clearly a handy tactic in military situations. Measuring the directions at which signals arrive in a mobile radio environment is also of interest to studies involving smart antennas [1], mobile radio propagation modelling [2] and location based servicing [3]. Long term, Angle-of-Arrival (AoA) measurements have also been employed to determine the multipath environment experienced in fixed terrestrial radio links [4]. In this application, the AoA measurements provide information on the reflectivity of the earth and the vertical structure of the atmosphere. However, to date, very little work has been reported on angle-of-arrival measurements in the maritime environment. Over-ocean amplitude profile measurements have been taken by numerous researches [5], [6] but no long-term angle-of-arrival measurements have ever been reported. This is despite the fact that this region should provide interesting results from the point of view that atmospheric structures like the evaporation duct are very common in this environment.

A fundamental requirement in all angle-of-arrival measurement systems is equipment with which to measure the direction of arrival of the incident radio signal. Two types of measurement system are commonly employed. One approach uses a simple amplitude measurement at the receiver. In this technique, the radiation pattern of the receiving antenna(s) provides the required angular discrimination [7]. Better directional resolution is usually possible by performing a vector measurement (both amplitude and phase) at the receiver. In this case, a multi-element antenna array is commonly used at the receiver in order to detect multiple signals with different receive angles simultaneously. Different receiver array configurations are possible [8].

The focus of this paper is angle-of-arrival measurement techniques suitable for use on fixed, over-ocean, radio links. In this situation, a large array using vertically stacked elements is best suited to separating the tightly grouped elevation angles at which rays are normally received. Conventional vector measurement systems for this situation usually use a microwave frequency Local Oscillator (LO) signal that is distributed to mixers attached to each antenna in the receiver array. Each signal is therefore synchronously detected for later amplitude and phase measurement at a lower Intermediate Frequency (IF). A problem with this measurement system though, is the cost and complexity of the electronics needed to ensure signal coherence. The magnitude of the problem is only fully appreciated when it is considered that a practical receiver array may utilize 10-20 elements spread over a vertical height of 20 meters or more [9].

This paper describes a new, vector measurement technique suitable for use with large receiver arrays used in angle-of-arrival measurements. In this method, signals received on each element of the array are downconverted to an IF frequency using separate, low cost, Low Noise Converters (LNC's). Offsets and drift in the different LNC's are compensated for using a reference signal that is externally injected into the array. The IF signals are digitized and processed to determine the received signal amplitude profile and AoA readings. The processing technique employed is based on a novel quadrature phase detection technique, which, besides giving excellent accuracy, is also insensitive to variations in the amplitudes of the input signals. This approach also does not require an accurate prior estimate of the frequency of the input sinusoids. The theoretical basis for this technique and some practical measurements are reported.

ANGLE-OF-ARRIVAL MEASUREMENT TECHNIQUE

The usual angle-of-arrival measurement technique employed for microwave terrestrial links consists of a vertical array of uniformly spaced antennas. Consider a linear array of N elements aligned along the z-axis as shown in Fig. 1. If the incoming wave is composed of M rays, each with a distinct angle-of-arrival θ_i, the vector signal received at each antenna may be written as

$$r(z) = a(z).e^{j\phi(z)} = \sum_{i=1}^{M} a_i.e^{j\phi_i} \tag{1}$$

where a_i are the amplitudes and ϕ_i the phases of the M components making up the incoming wave. The values a(z) and $\phi(z)$ are the total amplitude and phase respectively, resulting from the vector addition of all M incoming rays.

The amplitudes of each of the incident rays are assumed to be constant over the length of the array L, but the phase term ϕ, depends on the relative position along the array. Measuring phase with respect to the antenna at position z=0, the phase of a particular ray will be related to the AoA of the same ray by,

$$\phi_i = \frac{2\pi\, z \sin\theta_i}{\lambda} \tag{2}$$

where λ is the wavelength and θ_i the angle-of-arrival of the i-th ray.

The angle-of-arrival spectrum can be obtained from the Fourier transform of the complex amplitudes r(z), measured along the array, that is

$$R(\gamma) = \int_{-L/2}^{L/2} r(z).e^{-j2\pi\gamma z}\, dz \tag{3}$$

where L is the length of the array and γ is the Fourier transform variable.

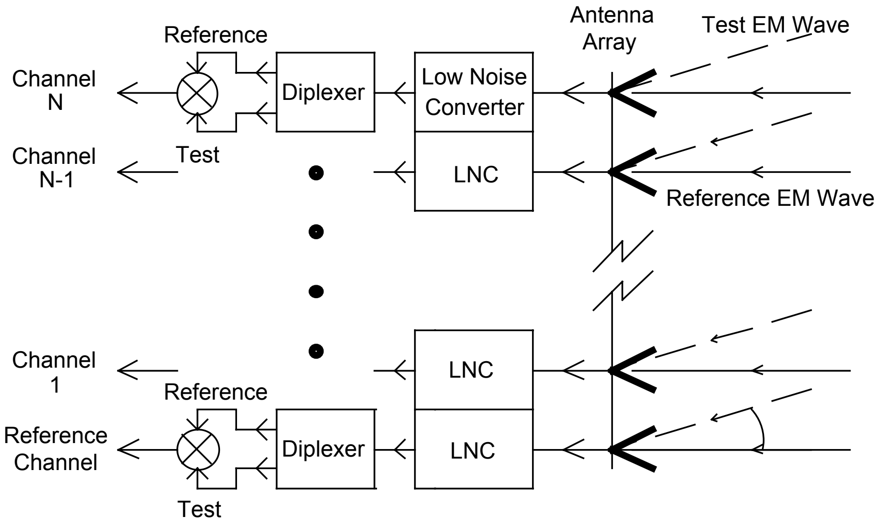

Fig. 1. M-channel angle of arrival (AoA) receiver array.

Since the arrival angles θ_i are distinct, (2) shows that each ray produces a phase term with a unique rate of change with z. The function described by (1) is therefore analogous to a time varying signal composed of a sum of M sinusoidal terms, each with a different frequency. Using this analogy, the angle-of-arrival spectrum described by (3) will produce a spectrum with peaks at points corresponding to

$$\gamma_{peak} = \sum_{i=1}^{M} \frac{\sin\theta_i}{\lambda} \tag{4}$$

Equation (4) shows the AoA of the different rays, θ_i, can be found from calculating the location of the spectral peaks, γ_{peak}. In any real measurement system the theory described previously is influenced by the practical constraint of having only a finite number of noisy samples of the function r(z). In this case, alaising and windowing effects will restrict the practical limits on range and resolution obtainable in the AoA measurements [4]. However, these effects are not studied in this particular paper.

From an implementation perspective, the only measurement needed to build an AoA sensor is the ability to measure the vector amplitudes of the signals received at the various antennas making up the array. That is, sample the value of $r(z) = a(z) \angle\phi(z)$ in (1). Practically, there can be some problems in making these measurements though. The spread in the AoA values to be measured on a typically radio link are normally very small (often <1°) and therefore precise raw measurements are important. Another problem is that the number and physical spacing of the antennas used in a typical array makes conventional vector measurement electronics expensive to build. The standard technique for making vector measurements at high frequency, synchronous detection, relies on mixing of the incoming signals with a common Local Oscillator signal [9]. The low frequency output from the mixer contains the required amplitude and phase information of the incoming RF signal. Unfortunately, this scheme may be expensive to implement. The discrete mixers attached to each antenna in the array can be a large cost. The cost of the high frequency cabling needed to distribute the common Local Oscillator (LO) signal to each of the antennas, is also a significant expense that needs to be considered. Finally, another potential problem with this approach is phase errors due to uncalibrated phase changes in the LO signals applied to the different mixers caused by heating and cooling of the high frequency cables and the electronics.

To overcome the high cost and potential problems of conventional vector measurement schemes when applied to an AoA measurement system using a large array, the authors have developed a new technique for this application. In this technique, antennas in the AoA array are attached to separate, low cost, LNC converter modules. Normally this type of measurement system would not be suitable for measuring phase since the LNC outputs are incoherent. This problem is overcome in the new technique by injecting a reference radio signal into the front of the array as shown in Fig. 1. The injected reference signal has a slightly different frequency to the test signal. The signal captured by an element of the array situated at $z = z_n$ can therefore be written as,

$$V_{RF}(z_n) = a(z_n)\cos(\omega_t t + \phi(z_n)) + \cos(\omega_r t + \varphi) \tag{5}$$

where $a(z_n)$, $\phi(z_n)$ are the time varying magnitude and phase terms representing the sum of the M rays propagating via different paths from the other end of the radio link at a frequency ω_t, while φ and ω_r are the phase and frequency of the injected reference signal. If the reference signal is employed, a conventional, incoherent Low Noise Converter (LNC) can be used at each antenna element. The signal at each element after downconversion is given by:

$$V_{IF}(z_n) = \frac{a(z_n)}{2}\cos\{(\omega_t - \omega_L)t + \phi(z_n) - \varepsilon_L(t)\} + \frac{1}{2}\cos\{(\omega_r - \omega_L)t + \varphi - \varepsilon_L(t)\} \tag{6}$$

where ω_L is the LO frequency of the LNC attached to the antenna element at $z = z_n$ and $\varepsilon_L(t)$ is the LO phase.

Remembering the LO frequency and phase will vary with time, both effects are accommodated in (6) by writing the LO phase as a time varying parameter, $\varepsilon_L(t)$. It should be noted though that these terms cancel in the mixing process described below and therefore do not influence the accuracy of the overall measurement. One approach to making the necessary measurement is to split the signals using a diplexer and mix the reference and test signal in each channel, Fig. 1. The output of this mixing process (after filtering) is a difference signal with a phase equal to the difference between the test and reference signals, given as:

$$V_e(z_n) = \frac{a(z_n)}{8}\cos\{(\omega_t - \omega_r)t + \phi(z_n) - \varphi\} \tag{7}$$

The immediate objective is to find $a(z_n)$ and $\phi(z_n)$. These values are samples of the general amplitude and phase variation over the length of the array, $r(z) = a(z) \angle\phi(z)$. Hence, once the amplitude and phase is measured at each element in the array, the AoA can be found using the method described in (1)-(4).

Equation (7) shows that the output obtained in each channel is a sinusoidal signal with a constant, yet unknown frequency corresponding to the difference in frequency between the incoming test and reference signals, and amplitude proportional to the size of the incident test signal. The required magnitude term, $a(z_n)$, can be found by determining the amplitude of this signal. The proposed method of doing this in this case is to sample the signal and then find its root mean squared amplitude. Finally, the measurement of the phase term in (7) reduces to a problem of determining the relative phase difference between this output at $z = z_n$ and a reference antenna output (typically the output at $z = 0$), which has an output given by:

$$V_r(z) = \frac{a(0)}{8}\cos\{(\omega_t - \omega_r)t + \phi(0) - \varphi\} \tag{8}$$

Measuring the phases of the signals produced in this type of system presents some problems though. Firstly, the frequencies of the signals to be measured may not be known exactly as standard sources like a YIG or a DRO will not be frequency locked. Furthermore, the measurement frequency will fluctuate with time as the various sources heat and cool. In the array built for this project, the difference signals will be centered at about 5MHz but it is estimated that the actual frequencies may drift by up to a 100kHz about this center frequency. Secondly, the received signal magnitudes will vary as the individual rays with different angles of incidence sum at the antenna input. The phase detector therefore needs to adapt to these changing conditions. Finally, the measurement needs to be performed quickly if we wish to measure phases over the full length of the array at a rate sufficient to monitor short-term variations in the atmospheric conditions.

The phase estimation problem can be formulated by the need to determine the phase between two versions of the signal, $x(t)$, in the presence of measurement noise, $n(t)$. In a discrete-time sampled data system, with sampling period T, the estimation problem becomes one of determining an estimate of the true phase delay using the measured samples, $x_1(kT)$ and $x_2(kT)$, of the signals $x_1(t)$ and $x_2(t)$. For simplicity and without loss of generality, we normalize the sampling period (i.e. let $T = 1$), which results in the sampled signals as given in (9):

$$\begin{aligned} x_1(k) &= A_1 \cos(2\pi f_0 k) + n_1(k) \\ x_2(k) &= A_2 \cos(2\pi f_0 k + \phi) + n_2(k) \end{aligned} \tag{9}$$

The problem is to estimate the phase ϕ from the current and past samples of $x_1(k)$ and $x_2(k)$. A common processing technique to find this phase is based on a quadrature phase detector [10]. In this approach it is first necessary to generate quadrature reference signals. If the input frequency is known, the quadrature-phase signal can be generated from the in-phase signal by a simple delay [10]. In this application though, the input frequency is not accurately known, and the quadrature-phase signal must be generated from the in-phase signal, $x_1(k)$ using the Hilbert transformer [11]. That is;

$$\begin{aligned} x_{1I}(k) &= x_1(k) = A_1 \cos(2\pi f_0 k) + n_1(k) \\ x_{1Q}(k) &= \mathcal{H}[x_1(k)] = A_1 \sin(2\pi f_0 k) + n_3(k) \end{aligned} \tag{11}$$

where $\mathcal{H}[\ ]$ is the Hilbert transform operator.

The input signal $x_2(k)$ from (9) is then multiplied by $x_{1I}(k)$ and $x_{1Q}(k)$ to generate the modulated terms $Qm_1(k)$ and $Qm_2(k)$. Finally, once we lowpass filter these outputs, the phase term is obtain as in (12):

$$\phi = \tan^{-1}\left[\frac{LPF[\mathrm{Qm}_2(kT)]}{LPF[\mathrm{Qm}_1(kT)]}\right] \approx \tan^{-1}\left[\frac{\sin\phi}{\cos\phi}\right] \tag{12}$$

MEASUREMENT RESULTS

To test the performance of the new incoherent AoA measurement technique, the system shown in Fig. 2 was assembled. The signals detected on each antenna in the receiver array are amplified and frequency down-shifted using commercial LNCs. Each LNC has a separate internal oscillator so the received signals will be incoherent in this system. To synchronize the received signals so phase delay measurements can be performed, a reference source is injected along with the test signal into the front of the receiver array. Two, 10mW DRO sources at 10.728GHz and 10.73GHz are used as the test and reference sources in this case. After a second stage of downconversion, the received signals are finally digitized and sent to a PC for storage and analysis. The signals were sampled at a rate of 50MS/s using an 8-bit converter.

The AoA system built was primarily designed for testing indoors, rather than as a fully-fledged AoA measurement sensor and as such only incorporates a simple two-element receiver array. The test system was also different to the proposed AoA sensor in that the “split and mix” operations were not implemented in hardware but were done off-line by processing the digitized signals. In a practical implementation, this type of approach might have speed or memory storage limitations but this design was adequate to demonstrate the principle.

To simulate variations in the angle-of-arrival of the test signal, the position of the 10.728GHz test source was varied linearly in 25mm steps. At each position, the raw data sequences were stored on the PC, where they were later processed. The data was processed using MatLab and the procedure described above to determine the time delay difference in the tests signals measured in the two array channels. The measured delay (in units of the sample period) versus position is graphed in Fig. 3. Using only 2 elements in the receiver means it will only be possible to measure the AoA of a single incident ray. Receiving more than one signal, as is highly likely in the indoor environment used for this experiment, will perturb the expected delay reading. The measurements seen in Fig. 3 therefore appear to be reasonable. The response is essentially linear but with some perturbations about the predicted straight line. This type of response is characteristic of a dominant direct ray and a smaller reflected signal.

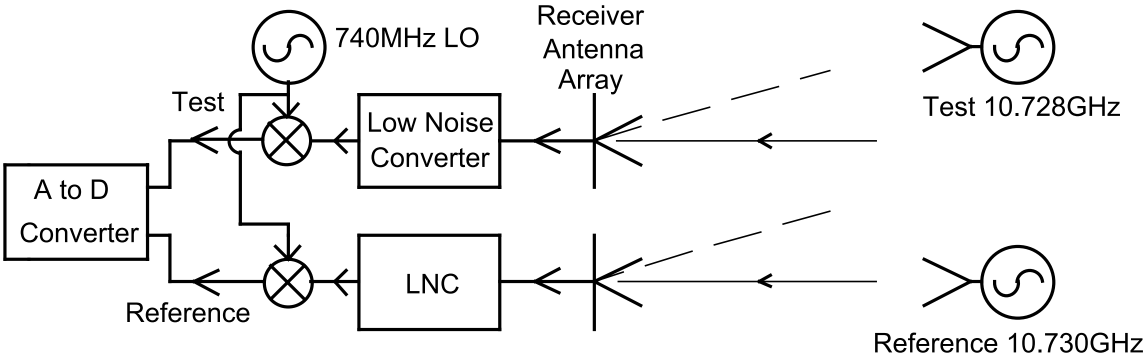

Fig. 2. Two element AoA measurement system used in testing.

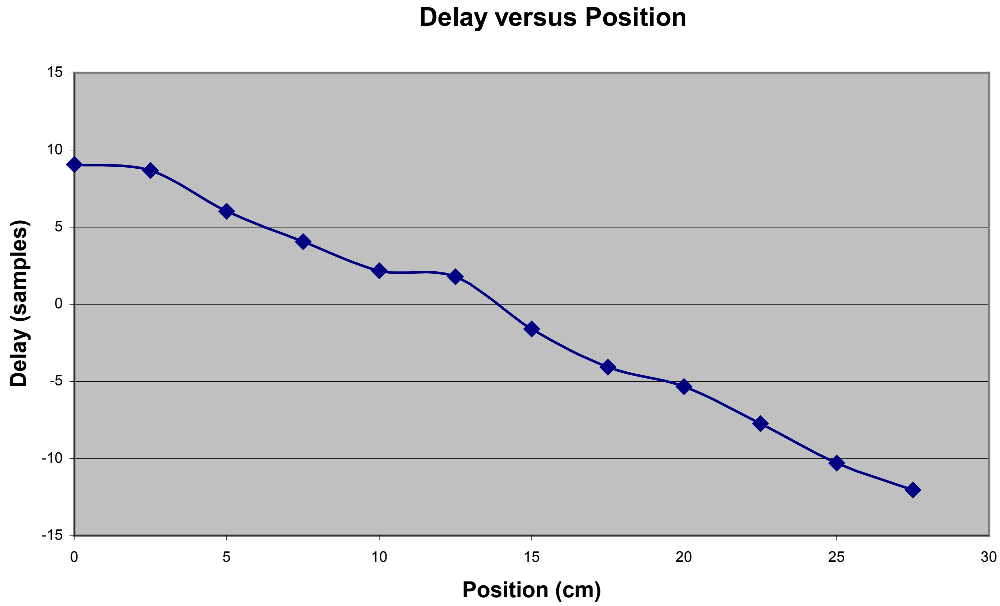

Fig. 3. Phase delay versus position of test antenna.

CONCLUSION

A new angle-of-arrival measurement technique designed for over-ocean propagation studies was described. The advantage of this approach was that common "off-the-self", incoherent receivers could be employed in the receiver array to reduce costs while still providing very good accuracy. The various channels in the receiver are synchronised by injecting a reference signal into each of the receiver elements. A new technique for processing the received signal to determine the phase delay was also described. This approach was based on a quadrature phase detector technique using the Hilbert transform to generate the quadrature reference component. This technique was attractive in this situation as the frequency of the detected signals does not need to be known accurately and it was therefore possible to accommodate frequency drift within the transmitter and receiver equipment. Laboratory measurements using a two-element receiver array were presented to demonstrate this method. Results consistent with an indoor propagation path were recorded giving confidence in the proposed techniques. A 10GHz angle-of-arrival receiver which uses a 20-element array and this measurement technique for over ocean propagation studies is current being built.

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