A 5.5 GHz CHANNEL SOUNDER FOR FIXED WIRELESS CHANNELS

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Abstract: This paper describes a channel sounder based on correlation of a maximum-length sequence. The sounder is designed for the evaluation of the absolute path loss and multipath delay spread of a fixed wireless local loop channel at 5.5 GHz. The sounder uses two stage down conversion with the second stage and subsequent processing being done in software, resulting in relatively simple hardware. Following explanation of the motivation for the work, a brief review of channel sounding techniques is given. The maximum-length sequence method is explained and the sounding system described. The method used for amplitude distortion correction. is summarised. Finally, some example results are presented.

1. MOTIVATION FOR THE WORK

Channel characterisation is an essential part of performance prediction for wireless communications systems. Work currently in progress aims to assess the suitability of the High Performance Radio Local Area Network Type 2 (HIPERLAN/2) physical layer for Wireless Local Loop (WLL) applications. This is to be done by means of computer simulation [1]. Simulation models of the 5.5 GHz WLL channel are required for various scenarios. The channel is characterised in terms of absolute path loss, signal angle of arrival and multipath delay spread profile (effectively the Impulse Response (IR) of the channel), as no data is available in the literature.

2. REVIEW OF WIDEBAND CHANNEL SOUNDING TECHNIQUES

Many techniques have been developed for the wide-band measurement of radio channel characteristics [2-4]. Indoor wireless Local Area Network (LAN) systems have received attention in the literature.

One way of performing such a characterisation is to repeatedly transmit a narrow periodic pulse [5]. The received signal is then the convolution of the transmitted pulse with the channel impulse response. However, this method suffers from the disadvantage the high power must be used because of the very short duration of the transmitted pulse. This causes difficulties with hardware design as well as possible regulatory problems.

Single-tone methods operate in the frequency domain. A single transmitted tone is stepped over a frequency range. For each step, the magnitude and phase of the received signal is recorded, resulting in a series of samples of the channel frequency response. The time period between frequency steps must be long enough to allow the channel to settle before a measurement is taken. In addition, the step size must be fine to give good resolution. Hence, measurement times for this method are relatively long and information about time variance of the channel is poor

Chirp sounding techniques use a continuous transmitted signal with a linear frequency modulation [6]. A matched filter or a heterodyne detector at the receiver compresses the chirp pulse. Such methods are capable of giving good results, but the hardware is complex.

3. THE MAXIMUM-LENGTH SEQUENCE CORRELATION METHOD

This method uses a maximum-length binary sequence as the probing signal [7]. Cross-correlating the received signal with the transmitted signal to give the periodic impulse response of the channel performs pulse compression. Maximum-Length Sequence (MLS) methods are relatively simple to implement. The correlation can be performed in hardware using (a) A Surface Acoustic Wave (SAW) convolver, or (b) a sliding correlator or (c) software DSP techniques [8,9].

The SAW convolver has the advantage that it allows measurement of the channel in real time. The primary disadvantage of the method is the poor dynamic range (typically 22-25 dB) of the SAW filter, caused by reflections from the ends of the piezoelectric substrate.

The sliding correlator method involves correlating the received sequence with a replica of the transmitted sequence that is clocked at a slightly slower rate. The difference in clock rates between the two sequences causes them to slide past one another as time passes. The time scaling inherent to this method causes averaging of the impulse response over the duration of the measurement and limitation of the maximum sounding rate.

The MLS method as been used in studies at 900 MHz [8] and 2 GHz [10,11]. The work of [12] and [9] used the software-correlation approach, which results in the simplest possible hardware. A software approach also allows the distortion due to the measurement chain (Amplifiers, filters etc.) to be easily and accurately compensated for. The high-speed sampling and digitisation of the signals that is required is not the problem that it once was.

4. SOUNDING SYSTEM DESIGN

An initial design for the channel sounding system was produced, based on reference [12]. The transmitter consists of a 5.5 GHz local carrier directly modulated by a 511 chip bipolar MLS (Binary Phase Shift Keying (BPSK) modulation) with a chip rate of 19.668 MHz. The receiver uses direct quadrature down conversion after bandpass filtering and amplification. The Local Oscillators (LO) are implemented by means of RF signal generators. No carrier tracking mechanism is employed. A digital storage oscilloscope then captures the I and Q channel data and software then performs the remaining processing. An inverse transfer function equaliser removes the effect of the measurement chain distortion and cross correlation of the resulting complex sequence is the performed with a stored copy of the receiver signal.

The hardware components are all commercially available parts with the exception of the receiver baseband amplifiers and low-pass filters. The MLS generator is constructed using standard high-speed logic integrated circuits. The transmitted Effective Isotropic Radiated Power (EIRP) is designed to be 11.3 dBm and the calculated input voltage to the Digital Storage Oscilloscope (DSO) is 0.8 mV RMS into 50Ω with 55 dB path loss.

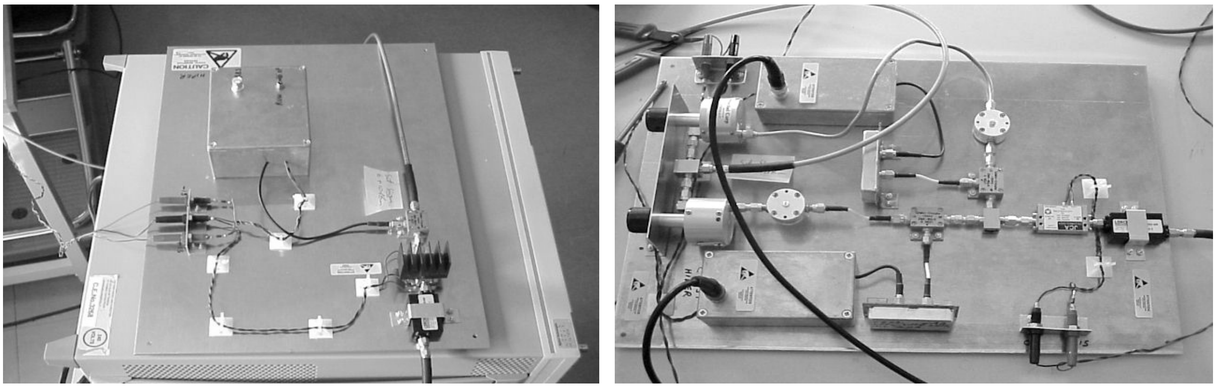

Figure 1: Photographs of the Hardware: Transmitter (Left) and Receiver (Right)

Absolute path loss is calibrated by measuring the mean complex (I+jQ) magnitude of the received signal with fixed attenuators of various values inserted in a back-to-back link cable. A least-squares fit is then performed on the data, giving a relationship between received mean complex magnitude and absolute path loss. This can then be used in subsequent measurements. Only the relative magnitudes of the multipath components are required in the final IR result so further calibration of the absolute levels of the individual multipath components is not required. It is important to realise that the antenna gain must be allowed for when assessing the absolute path loss.

Two types of antenna are used: narrow beamwidth for transmitter and receiver, or an alternative omnidirectional one for the receiver. This allows assessment of signal angle of arrival by rotating the narrow beamwidth antenna. Since the antennas can be considered as a part of the channel, the effect of antenna type on the multipath delay spread can also be assessed. The omnidirectional antennas are simple monopoles and the narrow band antenna is a corner reflector with a horizontal plane -3dB beamwidth of about ±10°.

During initial testing it was found that the LO frequency offset due to the lack of a carrier tracking mechanism caused the sidebands of the signal to fail to centre exactly on DC after demodulation. This results in some mixing of the sidebands and subsequent Double Sideband Suppressed Carrier (DSB-SC) - like Amplitude Modulation (AM) envelope distortion on the baseband signal. The resulting phase inversions on alternate half cycles of the AM envelope destroyed the correlation. Various post-processing algorithms were tried in order to directly remove the phase reversals in the time domain but without success.

A dual conversion approach was then employed to reduce the significance of the LO offset. The modified hardware down converts the signal to a 50MHz Intermediate Frequency (IF). Digitisation is then performed by the DSO and the second stage demodulation and subsequent processing performed in software. The LO signal for the second stage is provided by recording the IF signal with the MLS generator disabled. This eliminates the majority of the LO offset.

A further refinement is to correlate not with the transmitted baseband signal but with the received baseband signal with a back-to-back cable link. The measurement chain distortion is then included in the reference signal. Hence, the correlation forms a matched filter that recovers the impulse response of the radio channel without the distortion due to the measurement system. This approach is simpler than the inverse transfer function equaliser and it was adopted here. Figure 2 shows the final receiver architecture with dual conversion and the matched filter.

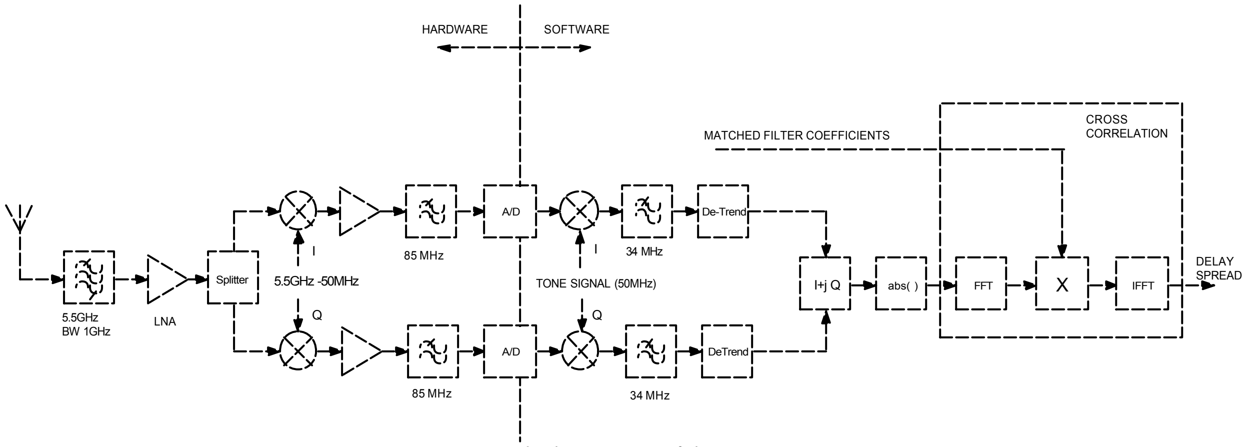

Figure 2: Block Diagram of the Receiver

The basic matched filter is simply the complex conjugate of the received signal spectrum with a back-to-back cable link. In practice, however, the basic matched filter will correct all of the phase distortion caused by the measurement chain but not all of the amplitude distortion. This is due to the fact that the absolute magnitude of the measurement chain transfer function will not be flat over the frequency range of interest. A method of modifying the matched filter in order to correct this amplitude distortion is employed, as proposed by Fannin *et al.* [9]. A brief summary of the method is given here.

The following notation is adopted: Transmitted sequence spectrum – $PN_c(f)$, Frequency domain form of matched filter – $M_c(f)$, Transfer function of the measurement chain (back-to-back link) – $L_c(f)$, Received Signal spectrum (back-to-back link) – $N_c(f)$, Received signal spectrum (with multipath radio link to be measured) – $S_c(f)$, Transfer function of the channel to be measured – $HM_c(f)$. The basic matched filter described above is then expressed as:

$$HM_c(f) = M_c(f) S_c(f) \qquad (1)$$
$$where: \quad M_c(f) = N_c^*(f)$$

Leading to the overall transfer function:

$$L_c(f) M_c(f) = PN_c^*(f) L_c^*(f) L_c(f) \qquad (2)$$
$$L_c(f) M_c(f) = PN_c^*(f) |L_c(f)|^2$$

Since $|L_c(f)|^2$ is not flat over the measured frequency range, we must find $|L_c(f)|^2$ by:

$$N_c(f) N_c^*(f) = |PN_c(f)|^2 |L_c(f)|^2 \qquad (3)$$
$$\therefore |L_c(f)|^2 = \frac{N_c(f) N_c^*(f)}{|PN_c(f)|^2}$$

This expression is not defined at $n=M$ (sequence length M, sample number n) or $n=0$ since $|PN_c(f)|^2$ is not defined at these points. Hence, amplitude distortion correction is not performed at these points. Finally, the modified matched filter is given by:

$$Mc(f) = \frac{N_c^*(f)}{|L_c(f)|^2} \qquad (4)$$

5. EXAMPLE RESULTS

Figure 3 shows an example delay spread profile form the sounder. This profile is taken from initial outdoor measurements in the area around Cambridge Consultants offices. A significant multipath component can be seen at 11 µs. The dynamic range can be seen to be about 40 dB. Sounders using DSP techniques are able to achieve dynamic ranges of this order, whereas hardware SAW convolver systems typically have a dynamic range of around 25 dB [9].

The performance of the LO offset correction method relies on the LO frequencies being unchanged in the time that elapses between the tone and MLS signals being recorded at the receiver. Any frequency drift between the two measurements will not be allowed for and AM envelope distortion will result. Ovenised references are employed in the signal generators. The LO drift will not be significant providing the signal generators are allowed sufficient time to stabilise before measurements are performed.

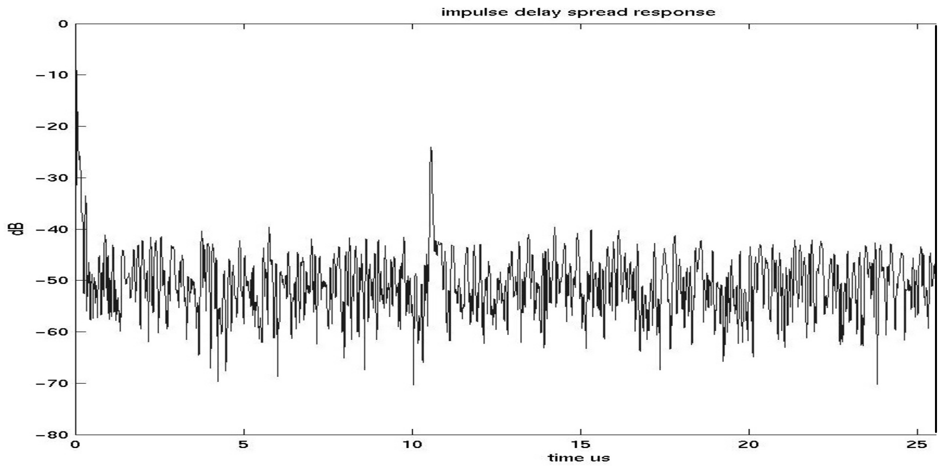

Figure 3: Example Delay Spread Profile

The IR obtained from the channel sounder can then be used to calculate channel parameters such as Ricean K-factor, average delay, delay spread, coherence bandwidth and peak delay. The data can be used in various ways to develop a simulation channel model based on a statistically varying tapped delay-line approach.

6. CONCLUSIONS

This paper has presented the design of a 5.5 GHz channel sounder. Channel sounding techniques have been briefly reviewed. Example results were presented. Further work now focuses on alternative methods of LO offset compensation, without the need to record the tone signal and the attendant problems of LO drift between the two measurements. This problem has now largely been resolved. A comprehensive measurement campaign will be undertaken. The 5.5 GHz WLL channel will be characterised in both suburban and dense urban environments, with a view to developing simulation channel models for various scenarios. Absolute path loss will also be measured and the effect of issues such as spatial diversity assessed.

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