SESSION 4C

WIRELESS COMMUNICATIONS THEORY
A Novel Blind Spectrum Sensing Approach for Cognitive Radios

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Abstract—Cognitive radio is believed to facilitate an efficient utilization of radio spectrum. Blind spectrum sensing is a key requirement enabling a cognitive radio technology by creating an awareness of other users’ activities. This report presents a novel blind spectrum sensing technique with a computationally efficient signal processing scheme which can achieve a good trade-off between latency and reliability. This approach is based on the statistical characteristics which differentiate between the transmitted signal and noise. It is shown that the proposed method is suitable for detecting other users’ activities. Computer simulations are carried out for a typical IEEE 802.11g scenario and it can be observed that this method outperforms the conventional energy detection technique.

I. INTRODUCTION

Traditional static spectrum allocation policies lead to situations where some radio bands are congested while other bands remain moderately or rarely occupied [1]. This implies that the inefficient radio spectrum handling has initiated a bottleneck for larger density wireless systems. Cognitive radio has been proposed as a potential wireless communication bottleneck for larger density wireless systems. Cognitive radio is the technology which utilizes the spectrum dynamically in an opportunistic manner without introducing harmful interference to the licensed users. The most fundamental step is to detect the presence of primary users’ signal and furthermore changing the radio parameters to exploit the unutilized segment of the spectrum. In cognitive radio terminology, users who have legacy rights on the usage of spectrum bands are called primary users while secondary users are the user who have lower priority and have to exploit the spectrum without introducing interference to the primary users [2]. This implies all secondary users have cognitive radio capabilities. There are several causes that prevent the spectrum sensing from functioning in satisfactory manner. One of the key factors effecting the performance is the multipath and shadowing loss which can attenuate primary users’ signal by the time it reaches the secondary user. Hence the secondary user has to employ a sensitive detection. The other concern in spectrum sensing is the sensing duration. Primary users can claim their frequency band at anytime. In order to prevent secondary user to induce interference to the primary users, they have to be able to identify the presence of the primary user as fast as possible and vacant the frequency band. Hence the spectrum sensing technique should not bring large latency in the system.

Spectrum sensing techniques can be classified in to two categories, i.e. blind spectrum sensing and non-blind spectrum sensing. The advantage of blind spectrum sensing is that it does not require any information of the source air interface. Examples of blind spectrum sensing techniques would be energy detection, wavelet based detection, and eigenvalue based detection and for the non-blind detection matched filtering and cyclostationary based detection can be pointed out. In category of non-blind sensing matched filtering is known to be the optimum method for detection. Hence it needs a perfect knowledge of the primary users’ air interface. Since cognitive radio needs receivers for all signal types, the implementation complexity of the sensing unit would be impractically large [3]. Cyclostationarity detection is another non-blind method used for detection of primary user. This technique makes use of the periodicity in the signals’ second order statistics [4],[5]. In this technique, the correlation function is used for detecting signals in a given spectrum band. However, this model is computationally complex and requires a long observation time which makes it impractical. It has also been shown that model uncertainties cause an SNR wall for cyclostationary based detection similar to energy detection [6]. Energy detector is the most common blind spectrum sensing technique being used. This is due to its implementation simplicity [7],[8],[4]. The shortcoming of energy detection is that it requires perfect information of the noise variance in order to perform satisfactorily. In practice it is not feasible to obtain an accurate value of the noise variance therefore this would introduce noise uncertainty [8]. Noise uncertainty makes energy detection performance quite unreliable which has been proven in [8],[6] and [7]. Wavelet based detection has been recently introduced in literature for spectrum sensing purposes [9]. In this method the wavelets are used for detecting the edges in the power spectral density (PSD) of a wideband channel. Once the edges have been detected, it can be decided whether the frequency band is vacant or not. This method is proven to be effective for wideband sensing [9]. Eigenvalue decomposition is another approach introduced which is based on the fact the received signal is usually correlated [10]. This is achieved by utilisation of multiple receive antennas. This paper proposes a novel blind spectrum sensing technique. The objective of the proposed method is to present a technique
which is not sensitive to primary users’ air interface and is able to perform satisfactorily in scenarios where more than one secondary users exist. At the same time it is computationally efficient and performs in reasonably low SNR without introducing high latency. This method provides a signal processing apparatus which enables an accurate discrimination of noise from received signal. This approach is based on second order statistical properties. The received information is divided into successive sets and power spectral of each set is derived. By sorting the power spectral densities in order of their magnitudes there exists a point where there is a sudden change in the magnitude. The sets which have PSD with magnitude less than this point are considered to be vacant band. In order to test the performance of the proposed approach, computer simulations are carried out for a typical IEEE 802.11g scenario.

II. SYSTEM MODEL AND PROBLEM FORMULATION

![Fig. 1. An Example of a system for studying blind spectrum sensing](image)

Fig(1) illustrates a scenario where there coexists a primary and several secondary systems. This is one of the scenarios where the conventional non-blind sensing techniques fail to operate in a satisfactory manner. This is due to lack of information about the other secondary users signalling formats. In this situation the secondary users should not only determine if a spectrum band is utilized by a primary user, it should also check if the band is used by another secondary user. The received signal at the secondary receiver is considered as

\[ y(t) = h(t) * s(t) + v(t), \quad 0 < t < T \]

(1)

where all the parameters in (1) are sequences in time domain and \( y(t) \) is the received signal, \( h(t) \) is a possibly random linear time-varying filter representing fading, \( s(t) \) is the primary users’ or other secondary users’ signal, and \( v(t) \) is the additive white Gaussian noise (AWGN) with zero mean and \( \sigma_v^2 \) variance and finally * is the convolution operator. Background noise is an aggregation of various sources like thermal noise, leakage of signals from imperfect front end filters, interference, etc. Therefore a stationary white Gaussian assumption is only an approximation. Throughout this paper the primary signal is assumed to be independent of noise and channel. In order to use the proposed signal processing algorithm for spectrum sensing, the signal is considered in the frequency band with central frequency \( f_c \) and bandwidth \( W \). The received signal is then sampled at a sampling rate \( f_s \) where \( f_s \geq 2W \). Therefore we can write the sampled received signal as

\[ y(nT_s) = h(nT_s) * s(nT_s) + v(nT_s) \]  

(2)

where \( T_s \) is the sampling period. Two hypotheses are considered for detecting the primary users’ presence which can be written as

\[ y(nT_s) = \begin{cases} v(nT_s), & H_0 \\ h(nT_s) * s(nT_s) + v(nT_s), & H_1 \end{cases} \]  

(3)

where \( H_0 \) and \( H_1 \) denote the absence and presence of the primary signal, respectively. The objective of this approach is to make a decision on presence or absence of the primary signal or any other secondary users on specific spectrum band.

III. BLIND SPECTRUM SENSING ALGORITHM

The block diagram of the proposed spectrum sensing technique is shown in Fig.2. To enable the detector the digitised received signal \( y(nT_s) \) is divided into several windows. Size of these windows are to be chosen cautiously since large data blocks will result in larger buffer size thus making the process more expensive. Also the larger the blocks are the more latency is introduced in the system. Each window has to be then represented in the frequency domain using one of the many methods such as Discrete or Fast Fourier Transform (FFT), Hartley, Cosine or Walsh. In this paper FFT has been used throughout. Size of the FFT determines the quality and accuracy of the process. After calculating the frequency equivalent the PSD of each window is calculated. In the next stage the PSDs have to be sorted in either ascending or descending order. At this point the algorithm looks for the point where the maximum change of gradient exists. If this value is smaller than the determined threshold then it is decided that the frequency band is vacant.

![Fig. 2. Block diagram of the proposed approach](image)

When the received \( y(t) \) only consists of AWGN:

\[ R_{yy} (\tau) = E \{ v(t) v^*(t + \tau) \} = N_0 \delta (\tau) \]  

(4)
where $R_{yy}(\tau)$ is the autocorrelation function and $N_0$ is the noise power. As it is known the autocorrelation at $\tau = 0$ and the PSD are Fourier transform pairs i.e $S_{yy}(f) = \mathcal{F}\{R_{yy}(0)\}$. Therefore, $S_{yy}$ of the received signal would be:

$$S_{yy}(f) = N_0, \quad \forall f$$  

(5)

In the case where received signal consists of both signal and AWGN i.e. $y(t) = h(t) \ast s(t) + v(t)$, we would have:

$$R_{yy}(\tau) = E\{(h(t) \ast s(t) + v(t))[h(t + \tau) \ast s(t + \tau) + v(t + \tau)]^*\}$$  

(6)

Since the propagation does not vary considerably within the observation time and substituting $\tau = 0$, (6) can be reduced to:

$$R_{yy}(0) = |h(t)|^2 E\{(s(t)s^*(t)) + E\{v(t)v^*(t)\}\}$$  

(7)

given that the primary/secondary signal $s(t)$ is uncorrelated and finally replacing (5) in (7) we would have:

$$R_{yy}(0) = |h(t)|^2 P_s \delta(0) + N_0 \delta(0)$$  

(8)

$$S_{yy}(f) = |H(f)|^2 P_s + N_0$$  

(9)

where $P_s$ indicates signal power and $H(f)$ is the frequency response of the propagation channel. It can be observed from (9) that when a primary/secondary signal exists the PSD of the received signal will no longer be flat and will be shaped according to the propagation channel, based on this theory this blind spectrum sensing technique was proposed. One of the most important parameter in every spectrum sensing technique is the probability of false alarm ($P_{FA}$) and probability of missed detection ($P_{MD}$) associated with it, where:

$$P_{FA} = \text{Prob}(U > \lambda | H_0)$$  

(10)

$$P_{MD} = \text{Prob}(U < \lambda | H_1)$$  

(11)

where $\lambda$ is the detector threshold. Therefore it would be appropriate to derive these two parameters for the proposed spectrum sensing technique. As it can be observed in (2) the signal $T(f)$ can be written as:

$$T(f) = \text{SORT} \left( \frac{1}{N} \sum_{n=1}^{N} |y_n(f)|^2 \right)$$  

(12)

where $y_n(f) = \mathcal{F}y_n(t)$ and $y_n(t)$ is the windowed version the input signal $y(t)$. $\text{SORT}(a_n)$ indicates the elements in $a_n$ being placed in the order of their magnitude in ascending order and finally $N$ represents the window length. Therefore the parameter $U$ can be expressed as:

$$U = \max \left( \frac{\partial}{\partial f} T(f) \right)$$  

(13)

where the function $\max(a_n)$ returns the element in $a_n$ with the maximum value and $rac{\partial}{\partial f}$ represents the differentiation operation with respect to parameter $f$. Based on central limit theorem we would have:

$$U | (H_0) \approx \mathcal{N}(0, \xi_0)$$  

(14)

$$U | (H_1) \approx \mathcal{N}(\phi, \xi_1)$$  

(15)

where $\mathcal{N}(\mu, \sigma^2)$ represents the normal distribution with mean $\mu$ and variance $\sigma^2$ and $\sigma^2$ represents the signal signals variance. Mean in $H_0$ scenario will be approximately zero as $N \to \infty$ also mean and the variance in $H_1$ are both function of the channels’ frequency response. Replacing (14) and (15) in (10) and (11) respectively we will have the following approximations:

$$P_{FA} \approx Q \left( \frac{\lambda}{\sqrt{\frac{2}{N} \xi_0}} \right)$$  

(16)

$$P_{MD} \approx 1 - Q \left( \frac{\lambda - \phi}{\sqrt{\frac{2}{N} \xi_1}} \right)$$  

(17)

$Q(.)$ is the standard Gaussian complementary cumulative distribution function (CDF).

IV. SIMULATION RESULTS

The proposed blind spectrum sensing algorithm has been evaluated though computer simulations. An orthogonal frequency-division multiplexing (OFDM) based air interface IEEE 802.11g system was chosen as the primary system. According to the IEEE 802.11g protocol each OFDM block contains 64 sub-carriers with cyclic prefix (CP) of length 16. One OFDM (including CP) duration is 4.0 $\mu$s and each OFDM frame contains 10 OFDM blocks. The channel adopted in the simulations are the frequency selective channel. System parameters are summarised in Fig(3).

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
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<tr>
<td>$N_s$</td>
<td>64</td>
</tr>
<tr>
<td>$N_s$</td>
<td>16</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>20MHz</td>
</tr>
<tr>
<td>Number of OFDM blocks per frame</td>
<td>10</td>
</tr>
</tbody>
</table>

Fig. 3. System parameters set-up
The results are obtained by averaging over 10000 independent realisations. In Fig(4) and Fig(5) the performance comparison of conventional energy detection with different uncertainty values and the proposed method have been shown. As it can be observed the proposed technique outperforms the energy detection even when the noise uncertainty (U) parameter is set to zero in other words when the energy detector has a perfect knowledge of the noise variance. The proposed method offers a better performance in lower SNR regions (\(\leq -13\) dB) and also degrades in a more steady fashion while energy detections’ performance tend to fall steeply after a specific SNR value. In this paper, we claim the secondary/primary is present if probability of detection \(P_D\) is equal or larger than 0.9. as it can be observed from Fig(5) the proposed technique offers a very high reliability when it come to \(P_{FA}\).

This implies that better spectrum efficiency can be achieved with use of this method. Fig(6) and Fig(7) demonstrate the relationship between the window length and \(P_D\) and \(P_{FA}\). As expected \(P_D\) is directly proportional to window length. This is due to the fact that the randomness of the noise would be more apparent as the window length become larger. Therefore the PSD of the sections which contain AWGN will experience less fluctuation and closer to the ideal constant magnitude which we are aiming for. It is worth mentioning that the sample size used for simulation results shown in Fig(4), Fig(5), Fig(6), Fig(7) is equal to 1024. Fig(8) provides the effect of observation sample size on \(P_D\) for a fixed SNR value. It can be seen that the in order to have a acceptable performance i.e \(P_D > 0.9\) a large sample size is not required. For example at SNR of -15 dB less than 2000 samples are required. This implies that FCC requirement [1] which states that the sensing
time of less 0.2 ms can be meet.

V. Conclusion

Even though blind spectrum sensing plays a significant role in the concept of cognitive radio, up to now, only a few algorithms have been proposed which are able to perform satisfactorily in low SNR regime. This paper has introduced a new blind spectrum sensing technique for cognitive radio. The proposed algorithm is based on second order statistical analysis which differentiate noise and transmitted signal. A decision on whether a sub-band is utilized by another user or not is made based on PSD estimation which is incorporated in the proposed technique. The proposed method can overcome the effect of noise uncertainty. Most importantly it does not require any information of the source air-interface while it can achieve a good trade-off between latency and reliability. Therefore this proposed method is suitable for the spectrum sensing in cognitive radio. The obtained performance results are promising, as it has been shown through simulations.

VI. Acknowledgement

This work was performed in the framework of EU-ICT WHERE2 project.

References

Experimental Investigation of the Performance of OOK-NRZ and RZ Modulation Techniques under Controlled Turbulence Channel in FSO Systems


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Abstract—A number of modulation schemes have been proposed and thoroughly studied for the successful operation of the optical wireless communication (OWC). Each modulation scheme has its advantages and disadvantages for the particular channel conditions. In this paper the comparative studies of on-off keying (OOK) return-to-zero (RZ) and non-return-to-zero (NRZ) modulation techniques have been carried out experimentally under controlled turbulence channel for free space optics (FSO) communication link. Weak turbulence effect is generated within a controlled chamber of 5.5m length. The experiment is performed with a temperature gradient of 4°C at wind velocity of 1m/s. A laser source is modulated using the selective modulation schemes and optical signal is allowed to propagate through the turbulence simulated chamber. The received signal is analyzed to measure the true effect of turbulence on the optical beam carrying information. The eye-diagram and signal distributions are analyzed for comparative studies. This research work is carried out under the EUCOST ACTION IC0802 project.

Index Terms—FSO, atmospheric turbulence, OOK

I. INTRODUCTION

FSO communications or better still, laser communications is an age long technology that entails the transmission of information laden optical radiation as the carrier signal through the atmospheric channel. FSO communication offers an increased information capacity compared to the radio frequency (RF) based communication systems [1]. The very narrow optical signal provides a secure link with adequate spatial isolation from its potential interferers. The electromagnetic spectrum used in FSO is license free, its initial set-up cost is lower and the deployment time is shorter [2]. FSO can deliver the same bandwidth as optical fibre but without the extra cost of right of way and trenching, without the electromagnetic interference due to the nature of information carrier photons unlike the RF based system, it has light weight and is very compact and consumes low power [3]. The performance of terrestrial FSO is highly dependent on the atmospheric conditions. A big challenge in FSO communications is the scintillation induced by atmospheric turbulence [3, 4]. The scintillation is induced due to the random changes in the refractive index of the atmosphere results from the combination of randomly varying temperature cells, wind speed and pressure in the intended path of the optical signal propagating in the atmosphere [5,6].

To improve the BER performance of a link due to scintillations, selection of appropriate modulation schemes is an important factor which determines the overall system performance and cost [7,8]. OOK is the simple and widely adopted modulation scheme used in commercial FSO communication system because of ease in implementation, simple receiver design, bandwidth efficiency and cost effectiveness [9]. However, the performance of OOK is very susceptible to environmental conditions.

From the viewpoint of the receiver’s sensitivity, RZ has been reported in [8,9] to offer better performance over NRZ in FSO links. However in turbulent-induced atmosphere, the demodulation of the received signal becomes less optimum using a fixed threshold [10,11]. Increasing the transmission power, which is limited due to eye safety regulations to combat turbulence fading results in high cost while employing optimal adaptive threshold for OOK brings complexity into the system [11].

This paper illustrates the experimental study for the performance of OOK-NRZ and OOK-RZ using different peak amplitude levels to demonstrate the effect of weak turbulence. The turbulence induced scintillations in the intensity of the optical signal decreases the performance of link significantly in a fixed threshold demodulation. The aim of the experiment was to optimize the link performance in the weak turbulence by increasing the transmission power of OOK-NRZ and OOK-RZ modulation and to compare the performance of both modulation schemes at the same transmission power. The paper is organized as follows: the Section II outlines the experimental set-up detailing the experimental parameters, transceiver design and chamber description. The experimental results are presented in Section III, with eye-diagram, signal
distribution and BER calculation. The conclusions based on the experimental results are drawn on the final section.

II. EXPERIMENTAL DESCRIPTION

A typical FSO link consists of a transmitter and receiver separated by the channel. The experimental set-up for the controlled study of scintillation effect on the FSO link for different modulation scheme is shown in the Fig.1. The transmitter uses a laser source with a maximum optical output power of 10 mW and a wavelength of 830 nm. The intensity of the output of a laser varied according to the modulating data format. To ensure the linearity of the system, the laser is properly biased and the peak-to-peak voltage of the input signal cannot exceed the specified values.

The receiver front-end consists of an optical telescope (or lens) and a photodetector. The electrical signal at the output of the photodetector is amplified using a transimpedance amplifier followed by circuitry for clock and timing recovery and regeneration of the transmitted data. The motivation of developing the chamber is to simulate the atmospheric channel affects on the optical signal traversing it at a controlled environment. Hence in the experimental set-up, data recovery circuitry is not utilized; rather the raw data at the receiver are analyzed. The complete simulation parameters used in the experiment are given in Table 1. The aim of the experiment is to demonstrate the effect of the scintillation for OOK-NRZ and OOK-RZ. The measurements are taken in similar channel conditions for the modulation schemes to achieve the optimized comparison. A periodic pseudorandom binary sequence (PRBS) of 1000 bit length is generated and converted to different signalling format as necessary to modulate the laser. The optical intensity at laser output is directly depended on the electrical input.

The data rates, amplitude level, wind velocity and temperature levels within the chamber are given in Table 2. The emphasis of the experiment is on simulating effect of the scintillation, and hence less focus is given on data rates, though higher data rate can be achieved with present experimental set-up. The OOK-NRZ pulse and OOK-RZ (with 50% duty cycle) pulse signalling format are shown in Fig. 2. The duty cycle for RZ pulse is half of the NRZ pulse. For a constant average optical power, amplitude of OOK-RZ can be γ times that of OOK-NRZ.

Higher amplitude peak power can be beneficial in the presence of turbulence. The experiment was carried out for different peak level (see Table. 2) under the same data rate. Three temperature readings were taken at the transmitting and receiving ends and at the centre of the chamber.

The table 1 and 2 are as follows:

Table 1: Main Parameters of FSO used in the experiment

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Laser diode</td>
<td></td>
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<tr>
<td>Peak wavelength</td>
<td>830nm</td>
</tr>
<tr>
<td>Maximum optical power</td>
<td>10mW</td>
</tr>
<tr>
<td>Class</td>
<td>Class 3b</td>
</tr>
<tr>
<td>Beam size at aperture</td>
<td>5mm×2mm</td>
</tr>
<tr>
<td>Maximum modulation frequency</td>
<td>150 MHz</td>
</tr>
<tr>
<td>PIN Photodetector</td>
<td></td>
</tr>
<tr>
<td>Wavelength of maximum sensitivity</td>
<td>900nm</td>
</tr>
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<td>Spectral range of sensitivity</td>
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<td>Active area</td>
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</tr>
<tr>
<td>Half angle</td>
<td>±75Deg</td>
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<tr>
<td>Spectral sensitivity</td>
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<tr>
<td>Rise and fall time of the photocurrent</td>
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</tr>
<tr>
<td>Rate</td>
<td>5Mbps</td>
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<tr>
<td>Chamber</td>
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</tr>
<tr>
<td>Dimension</td>
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</tr>
<tr>
<td>Temperature range</td>
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<tr>
<td>Wind speed</td>
<td>1m/s</td>
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</tbody>
</table>

Table 2: System parameters used for different modulations

<table>
<thead>
<tr>
<th>Modulation Type</th>
<th>Modulation Amplitude (mV)</th>
<th>Data Rate (Mbps)</th>
<th>Temperature °C</th>
<th>Wind speed (m/s)</th>
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</thead>
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<td>OOK-NRZ</td>
<td>50</td>
<td>5</td>
<td>24,28,27</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>100</td>
<td>5</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>150</td>
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<td>OOK-RZ</td>
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<td></td>
<td>200</td>
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<tr>
<td></td>
<td>300</td>
<td>5</td>
<td></td>
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</tbody>
</table>
III. EXPERIMENTAL RESULTS AND DISCUSSION

Experimental data for OOK-NRZ and OOK-RZ modulation schemes has been recorded for different modulation amplitudes under controlled turbulence environment and analysed using the eye-diagram and the received signal distributions. The eye diagram gives a quick examination of the quality of the optical signal before and after the turbulence.

The received signal distributions for bit ‘1’ and bit ‘0’ has been measured for OOK-NRZ and OOK-RZ modulation schemes and plotted as an eye diagram before and after the turbulence. In Fig. 3, the eye-diagrams of the received signal with and without turbulence for OOK-NRZ at peak transmitted amplitude of 50mV, 100mV and 150mV are presented.

The eye-diagrams clearly illustrate the adverse effect of turbulence on the optical signal as the width of eye-diagram is reduced significantly for channel with turbulence compare to the channel without turbulence. Significant distortion in eye-diagram in the presence of turbulence can be observed and the error probability is expected to be very high. It is also noticeable that there is progressive reduction of eye-opening as the peak amplitude reduces form 150 mV to 50 mV meaning higher peak amplitude can be used to reduce the effect of turbulence. However, power increment to mitigate turbulence level is simple but expensive solution and also the peak power that can be transmitted is limited by eye-safety regulation. Hence a number of alternative solutions like spatial diversity, temporal diversity have been proposed details of which can be found in [3, 11] and reference therein.

To further demonstrate the turbulence effect on the received signal, we also plotted signal histogram without turbulence and with turbulence (see appendix) for OOK-NRZ. The histogram showed that there is optimum threshold level between the received signal distribution of bit ‘1’ and ‘0’
without the induced irradiance fluctuations while with the induced irradiance fluctuations the threshold level is going to be perturbed as the variance of the distribution is going to be increase. This disturbance of the optimum threshold level due to the induced irradiance fluctuation is the worst case scenario results in the high error probability and link failure for the OOK-NRZ modulation scheme. Further the analysis for OOK-NRZ using curve fitting of two Gaussian distributions for 50mV gives average variance $\sigma = 3.49E-3$ without turbulence and $\sigma = 0.011$ with turbulence for signal amplitude respectively (see Fig. 4) increasing the error probability of the link.

This study showed that OOK-RZ is less sensitive to the turbulence, which can be verified by comparing the eye-diagrams of OOK-RZ for the same power level compared to NRZ pulse. Unlike the case of OOK-NRZ in the presence of weak turbulence, the eye-opening is significantly wider for OOK-RZ. For like-to-like comparison the peak amplitude of OOK-RZ is made twice that of OOK-RZ. The eye-diagrams of the received signal with and without turbulence for OOK-RZ at peak transmitted amplitude of 100mV, 200mV and 300mV are presented in Fig. 5.

![Eye-diagram](image)

Fig 5. Eye diagram for OOK-RZ with modulation amplitude 100mV, 200mV and 300mV respectively (a) without turbulence and (b) with turbulence.

![Eye-diagram](image)

Fig 6. The received signal distribution for OOK-RZ (a) without scintillation (b) with scintillation (Red line: experimental data; Blue lines: Theoretical fit)

The higher peak power adopted in OOK-RZ reduces the effect of turbulence which can be noticed by wider eye opening. In fact, OOK-RZ shows significantly improved performance compared to OOK-NRZ at higher turbulence level. The advantage of the OOK-RZ schemes comes at the expense of the bandwidth efficiency. The analysis using histogram and curve fitting of two Gaussian distributions for 100mV OOK-RZ pulse returns the average variance $\sigma = 5.58E-3$ without turbulence and $\sigma = 9.125E-3$ with turbulence, respectively (see Fig. 6). Notice that the variance for OOK-RZ without turbulence is double that of OOK-NRZ as bandwidth of OOK-RZ is twice that of OOK-NRZ. However, the variance of the distribution for OOK-RZ in the presence of turbulence is significantly lower than that of OOK-NRZ leading reduced error probability.

IV. CONCLUSION

The performance of OOK-RZ and NRZ modulation schemes under controlled turbulence environment has been experimentally studied for FSO communication link. The motivation of the experiment was to understand the turbulence effect on optical signal and also to demonstrate the performance of different modulation schemes. The experimental results have shown that OOK-RZ offer significantly higher resilience to turbulence compared to OOK-NRZ. The advantage of OOK-RZ comes from the higher peak power compared to that of OOK-NRZ, leading
lower error probability. Therefore the selection of suitable modulation scheme can be an effective measure to mitigate the irradiance fluctuations produced by the temperature variations and to increase the BER performance as well as the QoS of the FSO link.

APPENDIX A

![Graphs showing received signal distributions for OOK-NRZ for peak amplitudes of 50mV, 100mV, 150mV, and 200mV without and with scintillation.]

Fig 7 Received signal distributions for OOK-NRZ for peak amplitudes of 50mV, 100mV, and 150mV (a) without scintillation and (b) with scintillation.

![Graphs showing received signal distributions for OOK-RZ for peak amplitudes of 100mV, 200mV, and 300mV without and with scintillation.]

Fig 8. Received signal distributions for OOK-RZ for peak amplitudes of 100mV, 200mV, and 300mV (a) without scintillation and (b) with scintillation.

V. ACKNOWLEDGMENT

The research is part of the EUCOST ACTION IC0802. One of the authors (M. Ijaz) would like to acknowledge the financial supports received from School of Computing, Engineering and Information Sciences (CEIS).

REFERENCES

An Analytic Generalisation of a Maximum Entropy Customer Impatience Queueing Solution and its Nonbalking G/M/1/N Equivalence

Neelkamal Shah, Demetres D Kouvatsos, Rodney J Fretwell
School of Computing, Informatics and Media
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Abstract—This paper presents an analytic generalisation of the customer impatience model of Morse [1] and in addition shows maximum entropy (ME) significance of the generalisation thereby extending the association made by Kemp [2]. A nonbalking G/M/1/N queueing equivalence of the generalised result is derived for the purpose of performance evaluation.

I. INTRODUCTION

Entropy is a measure of information, choice and uncertainty of a random variable (rv) [3, pp. 11]. It is defined as the expectation of the logarithms of the probabilities of the rv taking its respective values. Let the rv be represented by X, then the entropy of X, \( H(X) \) is defined analytically as:

\[
H(X) = -\sum_{n} p_n \ln p_n
\]  

(1)

It appears in the laws that are fundamental to several fields of science for example data compression and transmission, ergodic and probability theory, statistical mechanics (thermodynamics), hypothesis testing, complexity theory, gambling and financial investment, quantum mechanics, dynamic systems, information theory and statistical inference and it is within the latter context that the principle of maximum entropy (PME) arises [4].

Entropy maximisation has been used to provide solutions in the areas of economics, business and finance, nonlinear spectral analysis, pattern recognition, transportation, urban and regional planning, queueing theory, parameter estimation and linear and nonlinear programming [5] among others.

With respect to queueing theory PME has been applied to solving numerous systems including but not limited to M/G/1 and G/M/1 queues [6], [7], finite and infinite capacity G/G/1 queues [8], [9], multiserver queues [10], [11], multiple class queues with priorities [12], queues with vacation [13]–[15] and queueing networks [16]–[18].

This paper initially presents an outline of PME in Section II. Then one of a number of landmark ME queueing results is briefly noted and its analytic generalisation subsequently detailed in Section III. An equivalent nonbalking queue is proposed in Section IV and the paper concluded in Section V.

In this paper, all QLD’s are considered solely in their limiting case i.e. as stationary QLD’s and the FCFS service discipline is implied in all the queueing systems.

II. THE PRINCIPLE OF MAXIMUM ENTROPY

The principle of maximum entropy (PME) provides a solution to the age-old problem of assignment of a probability distribution to an rv, which avoids bias while satisfying given or known information about the rv. This information may be very little. Probabilities are assigned to integers or intervals of values realised by an rv modelling a discrete or continuous quantity respectively of a system or process. In this way, PME purports to be the best inference of the unknown distribution.

Jaynes is credited with having formalised PME and in [19] he considers the relation of information theory to statistical mechanics instigated likely by their possession of the same mathematical functional \( -\sum_{n} p_n \ln p_n \) and argues that information theory justifies statistical mechanics and that statistical mechanics falls within the wider context of information theory. In [20] he charts the development of probability theory and statistical inference from its infancy, embodied in Bernoulli’s ‘principle of insufficient reason’ then advocates that the philosophy of maximum entropy was subsequently thought about by Laplace and Jeffreys. Further he admits that the mathematical facts concerning maximisation of entropy were pointed out long ago by Boltzmann and Gibbs who worked on the foundations of statistical mechanics, humbly debunking the view that PME is a radical innovation.

PME is described in relation to the discrete domain as that is the context in which it is used in the results reported in this paper. In PME, testable information about the rv is required in the form of its mean values and these are satisfied inherently in the inferred solution which has the maximum entropy of all the distributions that satisfy the known information. This method of inference can be further understood through the following interpretations and descriptions of the solution [19], [20]:

- It exhibits the greatest uncertainty of knowledge of the states of the system after satisfying known information.
Maximally non-committal to any particular state after accounting for known information about the system.

Least biased inference of the probabilities of system states.

The distribution that is spread out as uniformly as possible without contradicting the given information.

Mathematically the principle can be presented as follows:

It is known that the rv $N$ modelling a discrete quantity of a system or process takes values $n, n \in \mathbb{Z}$ but the probabilities $p_n = P(N=n)$ are unknown and to be determined. Some testable information about the system is known in the form of $m$ expectations of the rv given by:

$$f_i(n) = \sum_n f_i(n)p_n, n \in \mathbb{Z}, i = 1, 2, \ldots, m$$

Maximising the entropy functional (1) subject to the mean value constraints (2) and the normalising condition (3) is a standard variational problem solvable by the Lagrange multiplier technique yielding the following general distribution:

$$p_n = e^{-\sum_{i=1}^{m} \beta_i f_i(n)}, n \in \mathbb{Z}, m \geq 1$$

Where $f_0$ corresponds to the normalising condition implying that $f_0(n) = 1 \forall n$ and $\beta_i = 0, 1, \ldots, m$ are the Lagrangian multipliers.

The probability distribution can then be represented in terms of its normalising constant in the following standard form:

$$p_n = \frac{1}{Z} e^{-\sum_{i=1}^{m} \beta_i f_i(n)}, n \in \mathbb{Z}, m \geq 1$$

Where $Z$, referred to as the partition function in statistical physics is given by:

$$Z = \sum_n e^{-\sum_{i=1}^{m} \beta_i f_i(n)}, n \in \mathbb{Z}, m \geq 1$$

Re-parameterising (5), the following general product-form maximum entropy distribution (MED) results:

$$p_n = \frac{1}{Z} \prod_{i=1}^{m} f_i(n), n \in \mathbb{Z}, m \geq 1$$

III. THE MAXIMUM ENTROPY BALKING PARADIGM

A. The Discrete Half Normal

The discrete normal (dN) distribution was derived by maximising Shannon’s entropy subject to the mean and variance constraints [21]. The justification for this distribution being the dN is the well-known characterisation of the continuous normal, $\mathcal{N}(\mu, \sigma^2)$ as the unique distribution having maximum entropy for given mean and variance [22, pp. 162-163]. The form of the dN was given as [21]:

$$p_n = \theta^n q^{n(n-1)/2} p_0, n \in \mathbb{Z}, \theta > 0, 0 < q < 1$$

Subsequently the discrete half normal (dHN) was characterised as that discrete distribution with positive support that maximises Shannon’s entropy subject to the mean and variance. It is the dN left-truncated below zero and therefore its form is identical to (8) restricted by the condition $n \geq 0$. The dHN was contextualised as the QLD of a single server, infinite-capacity queue with exponential service and a Poisson stream of prospective customers filtered by Morse arrival balking [2].

B. Morse Balking

Morse proposed an analytic model to capture impatience of arrivals to an infinite capacity single server queue where the time that customers spend in service is exponentially distributed with mean rate $\mu$. In this model, prospective customers arrive in a Poisson stream with rate $\lambda$ and join the queue with probability $e^{-\alpha t}$ where $t$ is the customer’s estimate of his/her waiting time and $\alpha$ is a measure of the average impatience of customers. Consequently they balk with probability $(1 - e^{-\alpha t})$. However, the customer’s estimate of his/her waiting time most certainly depends on the number of customers present at the instant of arrival to the queue, $n$ and is also likely to depend on the estimated service time of each of the preceding customers in the queue, $E[s]$. This waiting time estimate can then be related to the latter two quantities by the following relationship:

$$t = nE[s] = \frac{n}{\mu}$$

Then the probability with which each prospective customer joins the queue becomes $e^{-\alpha t}$ [1, pp. 24].

After making the substitution $q = e^{-\alpha t}$, prospective customers can be seen to join the queue with population-dependent probability $q^n, 0 < q < 1$ or consequently balk from the queue with probability $(1 - q^n)$.

The condition $q = 1$ corresponds to all prospective customers joining the queue and the standard $M/M/1$ queue with arrival rate $\lambda$ results. This coincides with the condition of no impatience represented analytically by $\alpha = 0$.

The queue described above with Morse arrival balking and $q < 1$ can be modelled by a Markov chain and its QLD derived by balancing local flow between neighbouring states of the chain [2]. It is effectively a specific case of the $M(n)/M/1$ queue.

C. Morse Balking Generalisation

Now consider a single server (finite or infinite-capacity) queue with exponential service with mean rate $\mu$ and prospective customers forming a Poisson arrival stream with rate $\lambda$. Suppose that the probability with which a prospective customer joins the queue is given by $q_1^n q_2^m, 0 < q_1, q_2 < 1$, $q_1 \neq q_2$ and where $n$ is the instant queue size as seen by the prospective arrival, then the resulting QLD is a discrete MED (dMED) of the form (7) (with finite or infinite support respectively) where $m = 3$.

Further, consider individual single server (finite or infinite-capacity) queues with exponential service with mean rate $\mu$, prospective customers forming a Poisson arrival stream with rate $\lambda$ and each queue distinguished by a unique value of $m, (m = 2, 3, \ldots)$. If prospective customers to each of
these queues join with the following respective population-dependent probabilities:

\[ q(n) = \prod_{j=1}^{m-1} q_j^n, \quad 0 < q_j < 1, \quad m > 1, \quad q_k \neq q_k \forall k \neq l \quad (10) \]

then the resulting QLD's can be shown to be dMED's of the form (7) (with finite or infinite support respectively) derived from knowledge of \( m \) successive queue length moments. The QLD's can be derived by solving their Markov chain models as described briefly in Section III-B eventually bearing the following general form:

\[ p_n = p_0 \left( \frac{\lambda}{\mu} \right)^n \prod_{j=1}^{m-1} q_j^{n-k} k^j, \quad n \geq 0, \quad m > 1 \quad (11) \]

In order to represent the parameters of the dMED's of (7) i.e. \( x_i, i = 1, 2, \ldots, m \) in terms of the respective queueing parameters \( \lambda, \mu \) and the constituents of the joining probabilities \( q_1, q_2, \ldots, q_{m-1} \), the following two equivalent relationships are known to be correct:

\[ \sum_{i=1}^{\infty} (1 - q_i) i \lambda = \mu \]

Therefore the effective inter-arrival time distribution is simple the random sum of iid exponential rv's bounded by the geometric distribution and it can be derived mathematically as follows.

Let the rv \( X \) model the exponential durations between the prospective arrivals and let the rv \( T \) represent the effective inter-arrival time durations. Then the LST of \( T \) can be given by:

\[ L_T(s) = \sum_{i=1}^{\infty} \left( \frac{\lambda}{\lambda + s} \right)^i (1 - q_j)^{i-1} q_j = \frac{q_\lambda}{q\lambda + s} \quad (15) \]

Inverting this LST yields the exponential CDF with revised rate \( q_\lambda \). Therefore the single server (finite or infinite-capacity) queue with exponential service and prospective arrivals forming a Poisson stream with rate \( \lambda \) filtered by probability \( q \) is a (finite or infinite-capacity) M/M/1 queue with the following ME QLD:

\[ p_n = p_0 \left( \frac{\lambda q^n}{\mu} \right), \quad n \geq 0, \quad m = 1 \quad (16) \]

The following condition must hold to avoid analytic insolubility in this case characterises overloading in the queue:\footnote{The condition \( \lambda q = \mu \) is permitted in finite-capacity queues only and results in the discrete uniform QLD. The discrete uniform is also a dMED, derived by maximising Shannon's entropy subject to the normalising constraint (3) [19, pp. 622–623].}

\[ \lambda q < \mu \quad (17) \]

Therefore the probabilities defining arrival balking at standard queues that possess ME QLD’s of the form (7) are given collectively by:

\[ q(n) = \left\{ \begin{array}{c} q \prod_{j=1}^{m-1} q_j^{n-j} m = 1 \\ 0 \end{array} \right. \]

\[ m > 1 \quad (18) \]

**E. Service Balking**

Consider a single server (finite or infinite-capacity) queue with uncensored Poisson arrivals with rate \( \lambda \) and exponential service with rate \( \mu \). Here, customers at the head of the waiting line proceed to service with probability \( q_\lambda(n), n \geq 1 \) or consequently balk from service with the complement of this probability. Let these probabilities be defined by the following:

\[ q_b(n) = \left\{ \begin{array}{c} q(n-1) \prod_{j=1}^{m-1} q_j^{n-j} m = 1 \\ q(n-1) \geq 0 \end{array} \right. \]

\[ m > 1 \quad (19) \]

Then for the cases \( m > 1 \), the QLD's depart from (7) (despite possessing the same form) due to the parameter corresponding to \( q_j \) in (11) becoming greater than one. Interestingly, these queues can be interpreted as offering adverse service caused by customers being less likely to proceed to service the more populated the queue, leading rapidly to overloading.
Service balking is considered only for the case $m = 1$. By the same arguments used to prove the effective inter-arrival time in the arrival balking queue with probability $q$ of joining the queue, the QLD of the (finite or infinite-capacity) service balking queue with $q(n) = q$ is given below:

$$p_n = p_0 \left( \frac{\lambda}{q \mu} \right)^n, \ n \geq 0, \ m = 1 \quad (20)$$

Similar to the arrival balking case, the following condition must hold to avoid analytic insolubility characterising over-loading in the queue:

$$\lambda < q \mu \quad (21)$$

IV. Nonbalking G/M/1/N Equivalence

A nonbalking equivalence of the queues with arrival balking, pertinent to the cases $m > 1$, is proposed for the purpose of performance evaluation. This is achieved by taking the expectation of the joining probability, which is defined as follows:

$$E[q(n)] = \sum_n q(n)p_n, \ n \geq 0 \quad (22)$$

Since closed form expressions for $E[q(n)]$ are currently unobtainable for the cases $m > 1$, the equivalence can only be applied to finite-capacity queues for these cases, solvable by computation of the expectation.

In the equivalent queue, the probability of joining the queue is replaced by the expectation of joining. Taking the expectation has the effect of decoupling the probability of joining from dependence on the queue length resulting in the following being geometrically-distributed: (number of customers that balk between any two that join + 1).

Therefore by the same reasoning presented in Section III-D, the QLD’s of the equivalent nonbalking queues modelling single server finite-capacity queues with exponential service with mean rate $\mu$ and prospective arrivals constituting a Poisson arrival process with rate $\lambda$ are those of M/M/1/N queues with respective effective arrival rates $\lambda E[q(n)]$. Therefore the QLD’s of the equivalent queues are coincidentally dMED’s and are defined by the following truncated modified geometric:

$$p_n = \frac{1 - \frac{\lambda E[q(n)]}{\mu}}{1 - \left( \frac{\lambda E[q(n)]}{\mu} \right)^{N+1}} \left( \frac{\lambda E[q(n)]}{\mu} \right)^n, \ 0 \leq n \leq N, \ m \geq 1 \quad (23)$$

(Note that (23) holds for the case $m = 1$ too.)

Performance measures can then be approximated using the known results for M/M/1/N queues.

V. Conclusions

In this paper an analytic generalisation of the landmark result proposed by Morse [1, pp. 24] is presented. Morse submitted an analytic model of customer impatience at a queue implemented using the principle of arrival balking. In addition, ME significance has been shown for the extended result thereby furthering the contextualisation of Morse’s model by Kemp [2].

A nonbalking G/M/1/N equivalent queue is derived to model the queues with generalised arrival balking in order to obtain their approximate performance metrics.

Future efforts are intended to be directed at detailed comparisons between the arrival balking queues and more general ME and non-ME G/M/1/N equivalences.

REFERENCES


Comparison of the Return Loss and Axial Ratio for Circular Polarised Square Patch Antenna using Two Branch and Ring Coupler Feeds

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Abstract- Circular polarised antennas are used in many wireless communication systems where it is necessary to have a wide bandwidth of the return loss and the axial ratio. This paper compares the bandwidth of the above parameters for a circular polarised square patch antenna fed by a two branch coupler and a ring coupler. The design of the two couplers and the patch is initially reviewed. Based on the equivalent circuits of the two couplers and that of the patch the complete antenna is modelled using Microwave Office (AWR) software. The results obtained show that a ring coupler produces a wider bandwidth for the return loss and axial ratio than that obtained by a two branch coupler.

I. INTRODUCTION

Circular polarisation is used in systems such as, global system of mobile communication (GSM), global positioning system (GPS), satellite communication, radio frequency identification (RFID), and intelligent transport system (ITS) [1]. In these applications, due to their thin profile, ease of manufacture and small size microstrip patch antennas are normally used. At Northumbria University a microwave system has been developed which can be used in automatic tolling on motorways or where restricted access is required for cars e.g. for taxis entry at airports [2]. This system uses a tag placed behind the windscreen of a car which communicates with a roadside beacon. In this system linear polarisation cannot be used as at the tag the reflected signal from the car bonnet interferes with the direct signal from the beacon. This interference causes gaps in the communication between the tag and the beacon which can be overcome by using circular polarisation. The direct signal is transmitted with right hand circular polarisation and the reflected changes to left hand polarisation. Consequently the effect of interference of the reflected signal is virtually eliminated.

II. DESIGN OF 3DB TWO BRANCH AND RING COUPLERS

A directional coupler is modelled as a four port network where all ports are matched to 50Ω. For an ideal coupler, when power is fed into one port, output power is obtained from two of the ports and no power is coupled into the fourth port. The two outputs for an ideal two branch 3dB coupler are equal in magnitude and 90° out of phase. However, for a 3dB ring coupler the two output signals are equal but 180° degrees out of phase. Consequently to obtain circular polarisation an extra quarter wavelength long microstrip line must be used at one of the ports.

The structure and the equivalent circuit of a two branch coupler are shown in Fig. 1 where all the transmission lines are λ/4 long at the design frequency. For a signal fed into port one, the output signals are obtained from ports two and three and no power is dissipated in port four. To obtain this condition there is an effective short circuit at port four and hence it can be shown that Z1=35.5 Ω and Z2=50Ω.

A ring coupler is shown in Fig. 2 below where for a feed at port 1, port 3 is isolated and output signals are obtained from ports 2 and 4 which are 180° out of phase.
The equivalent circuit of the ring coupler is similar to the two branch coupler shown Fig. 1(a) where the effective short circuit is now at port 3. Using this condition it can be shown that the characteristic impedance of all the lines for a 3dB coupler is 70.7 Ω. Further details of the analysis and design of both couplers for can be obtained in ref. [3].

### III. PATCH ANTENNA

A basic structure of a patch antenna is shown in Fig. 3 where the rectangular patch has a physical length ‘Lp,’ width ‘Wp’ and the thickness of the substrate is ‘h’.

![Fig. 3. Structure of a rectangular patch antenna](image)

It is also possible to use circular or triangular patches [4,5]. The radiation from the patch is caused by two fringing fields of approximate length ΔL as shown in Fig. 4.

![Fig. 4. Side view of the electric fringing field](image)

In equation 1 \( L_E = L_P + 2\Delta L = \lambda / 2 \) (1) The equivalent of the patch antenna is shown on Fig. 5. \( R_1 \) models the power radiated from each edge of the patch and \( R_{12} \) models the power lost due to the coupling between the two edges of the patch. Capacitor \( C_1 \) models the two edge fringing fields and \( Z_0 \) is the characteristic impedance of the microstrip line connecting the parallel edge impedances. The equations for all the above design parameters in Fig. 5 are given in [6].

![Fig. 5. Equivalent circuit of microstrip patch antenna](image)

The block diagrams of the two circular polarised antennas are shown in Fig. 6 for a two branch and ring coupler feeds. The feed network in each diagram is a \( \lambda / 4 \) transformer which transforms the input impedance of the antenna to 50Ω.

![Fig. 6. Block diagrams of the antenna.](image)

(a) Two branch coupler fed antenna
(b) Ring coupler fed antenna

The models of the two antennas used in the AWR software are shown in Fig. 7.
Fig. 7. Equivalent circuit of the circular polarized antenna (a) Two branch coupler fed antenna (b) Ring coupler fed antenna
IV. COMPARISON IN THE PERFORMANCE OF THE ANTENNA FOR THE TWO TYPES OF FEEDS

To assess the performance of the circular polarised antennas two parameters, return loss and axial ratio are considered. The return loss is defined as equal to 20log\(_{10}(S_{11})\) where \(S_{11}\) is the reflection coefficient at the feed port and is a measure of the lost power which is reflected to the feed source. The axial ratio defined in equation (2) below and is a measure how equal in magnitude \(E_x\) and \(E_y\) are and how close to 90\(^{0}\) is their phase difference [6].

\[
Axial\ Ratio = \frac{E_x^2 + E_y^2 + (E_x^4 + E_y^4 + 2\times E_x^2\times E_y^2\times \cos\theta)^{\frac{1}{2}}}{E_x^2 + E_y^2 - (E_x^4 + E_y^4 + 2\times E_x^2\times E_y^2\times \cos\theta)^{\frac{1}{2}}} \quad (2)
\]

If \(E_x\) and \(E_y\) are equal and they are 90\(^{0}\) degrees out of phase then from equation (2) the axial ratio is equal to one or 0 dBs.

The frequency response of the return loss using the two couplers is shown in Fig. 8.

The frequency range normally defined in industry is -10 dB return loss. For this definition the bandwidth of a normal directly edge fed microstrip patch antenna is about 3-5 \%. [4]. If the couplers are used the bandwidth increases to 21.8 \% for the two branch coupler, and 125 \% for the ring coupler.

The frequency response of the axial ratio (AR) is shown in Fig. 9. The bandwidth of the two branch coupler antenna is 8.5 \% (from 2.348 to 2.552 GHZ) and ring coupler antenna is 29.1 \% (from 2.794 to 2.105 GHz). From the above results the ring coupler produces a wider bandwidth for both the return loss and for the axial ratio.

![Fig. 8. Simulation results for the two branch coupler (triangle line) and ring coupler (rectangular line)](image)

![Fig. 9. AR of the two branch coupler antenna (triangle line) and ring coupler antenna (rectangular line)](image)
V. Conclusion

In this paper, the theory of a two branch coupler, a ring coupler and a microstrip patch antenna were briefly reviewed. For each coupler feed the parameters of the equivalent circuit were determined and then used in simulation to compare the bandwidth of the return loss and the axial ratio. For both parameters, it was shown that a wider bandwidth was obtained by using a ring coupler feed. The disadvantage of using the couplers to increase the bandwidth was that the gain of the antenna was reduced by 1dB. This reduction of gain was caused by some power being dissipated in the isolated ports. This loss in the gain can be overcome by using an amplifier to increase the radiated signal.

Acknowledgements

The authors would like to express the gratitude to all the colleagues in NCR lab of Northumbria University.

References