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THE EFFECTS OF MULTIPATH PROPAGATION ON FREQUENCY HOPPING RADIOS

R. J. Preston and P. J. Munday

Introduction

Frequency hopping in the VHF band will be an Anti-Jam system deployed to defeat enemy jamming in the event of hostilities. The Racal Jaguar V system employs this technique and this paper sets out to predict the probability of communicating under conditions of multipath represented by a Rayleigh model and large terrain variations in a rural environment. The effects of multi-path in an urban environment is also discussed.

Rayleigh Multipath Model

The Rayleigh probability distribution function can be used to describe "small terrain" variations in received signal strength. This variation is due to a number of signal paths existing between the transmitter and receiver. Other received signal variations exist. The major and obvious variation is due to range, and there is a lesser variation due to "large" terrain variations such as site elevation. This large terrain variation will be dealt with later and is based upon the work by Egli (Reference 1).

The Rayleigh probability distribution function is given in Figure 1. It shows that there is a 90% probability that the received signal will be greater than 8.5dB below the median and a 10% probability that it will be 6.2dB above the median. In order to convert these results into range the following assumptions are made:

Transmitter output power	:	50W
Received detected S/N	:	$\geqslant$ 8dB
Whip antennas	:	2.4m
Path loss proportionality	:	r^{-4}

Table 1 System Parameters

Range Performance

The probability of communicating can now be calculated and is plotted in Figure 2. It can be seen that the median range is 39km. Now to examine what happens in the frequency hopping mode a few more parameters have to be declared.

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Number of F.H. channels	:	256
Minimum spread bandwidth	:	6.4MHz
Maximum number of unavailable channels	:	51 (20%)

Table 2 F.H. Parameters

The communications capability of frequency hopping systems depends on the number of channels that are unavailable. At this point this unavailability is assumed to be due to the fact that a particular channel is in a frequency null. Later the effect of other users blocking a channel will be considered. Frequency hopping systems degrade gracefully as the unavailable channels increase . The following table gives a subjective assessment of a frequency hopping link versus the percentage of unavailable channels.

%	Assessment
5	Noticeable but not objectionable
10	Some degradation
20	Degraded but readable
30	Difficult to understand
50	Unreadable

Table 3 Link Quality

A value of 20% unavailability has been chosen for this evaluation as it leaves a margin for losing channels due to other reasons, but still provides a usable link.

Figure 2 shows, therefore, that for a communications probability of greater than 80% frequency hopping is better, although, of course, of a different quality. The probability of achieving a range greater than 30km is very low. The significant point from Figure 2 is that under multipath conditions, the range for frequency hopping is fairly well fixed, whereas for fixed frequency, it can vary quite considerably. On average however, the range for frequency hopping is 75% of that for fixed frequency.

When the "large" terrain variations described by Egli are also taken into account, the probabilities of communication are as shown in Figure 3 and it can be seen that for hopping, the range is always lower. This apparently surprising result is explained by the fact that Figure 2 gives the probability of communication at sites having a given figure for the large terrain variations (in fact the median figure), whereas Figure 3 represents the effect of averaging over all sites, and on average the fixed frequency range is slightly greater.

Battlefield Model

The battlefield model will now be described. First of all the results of Egli have been used to give the probability of communicating assuming that the effects of Rayleigh fading have been avoided by either marginal movement of site positions or by a judicious choice of frequency (A practice which actually happens). It can be seen in Figure 4 that the median range is the same as the previous example for Rayleigh fading.

The frequency management plan assumes that within a division all frequencies are planned, but adjacent divisions are not and neither are, obviously, the adversaries. The number of nets in a division is 256. The forward nets are more active than the rear nets and the reserve nets are at half the density per unit area.

Figure 4 shows that using these parameters the 90% probability of communicating has reduced, for fixed frequency, from 18km to 13km. Using the frequency hopping mode gives almost identical ranges for greater than 80% probability of communication, with lower ranges than fixed frequency below this level. The shapes of the curves are similar to those for the Rayleigh Multipath model.

Urban Environment

The urban environment presents a vastly different problem. Other workers have done work in this area, notably R.C. French (Reference 2). In this paper, the particular effect of flutter is to be addressed. Flutter is a function of frequency and relative velocity. The nature of the environment also has an effect. Flutter rate, in Hz, is plotted against velocity in Figure 5 for 30MHz and 80MHz. Taking a relative velocity of 15m/s the flutter occurs at a rate between 3Hz to 7Hz. This has a disruptive effect on speech where the typical syllabic rate is about 2 syllables per second. Frequency Hopping produced an additional amplitude modulation. If it is fast enough, its effect will predominate over the basic flutter effect. Therefore, a hop rate of greater than 50 hops per second will significantly mask the flutter and hence improve intelligibility. Frequency hopping will also provide virtually continuous coverage in an urban environment since the phenomenon of deep, frequency selective fades is avoided. Major shading due to large buildings cannot, of course, be avoided.

Data

Frequency Hopping can provide adequate performance for data traffic providing forward error correction techniques, coupled with time diversity, are employed. The choice of a forward error correcting code is critical. It is highly desirable that not only should the code improve in input error rate but should never give an output error rate worse than the input error rate.

Conclusions

This paper has demonstrated that under conditions of Rayleigh fading, the use of frequency hopping provides an increased range performance for a 90% probability of communicating. The quality of the link will be different and will provide an almost constant range system. The effect of flutter in urban environments is virtually eliminated and can provide continuous coverage, neglecting shadowing effects.

Acknowledgements

We wish to thank Mr. N. R. Massey for his enthusiastic participation and invaluable work on the computer modelling and the preparation of the figures; Racal Electronics Limited for their permission to present this paper.

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Fig. 1 Rayleigh Distribution

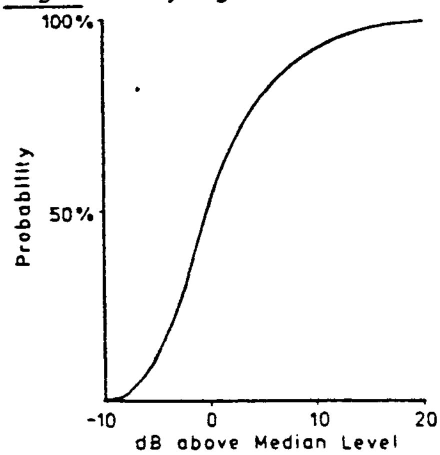
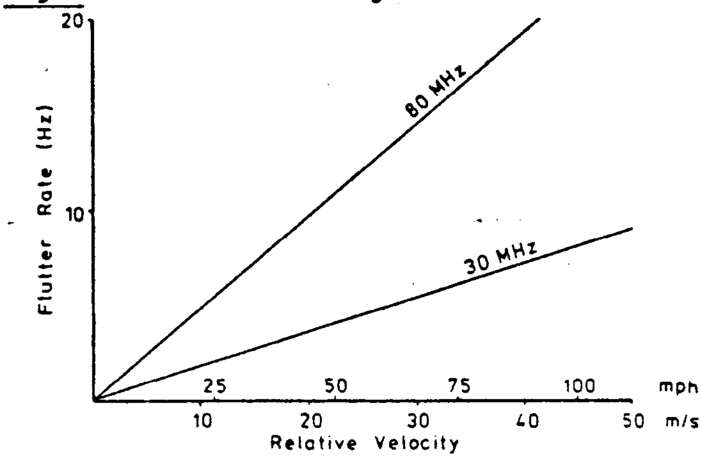


Fig. 5 Flutter in Moving Vehicles



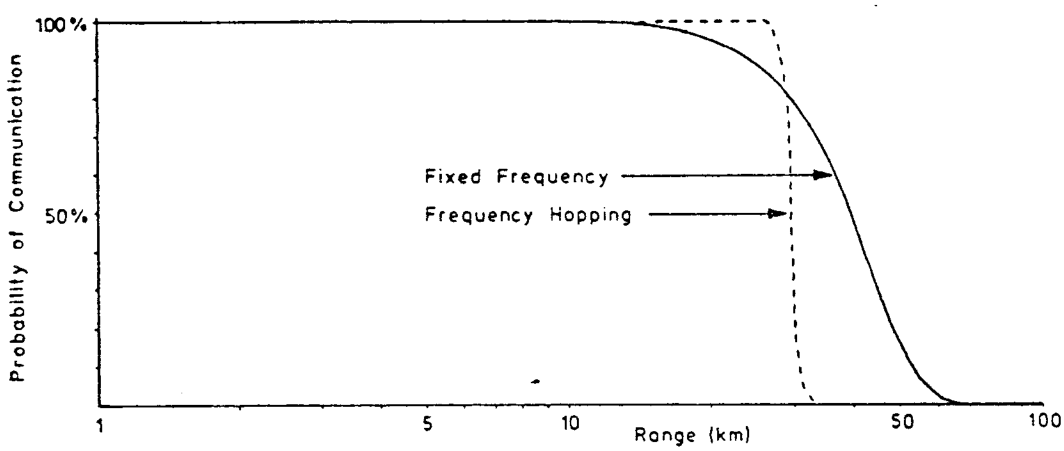


Fig. 2

Rayleigh Multipath Model

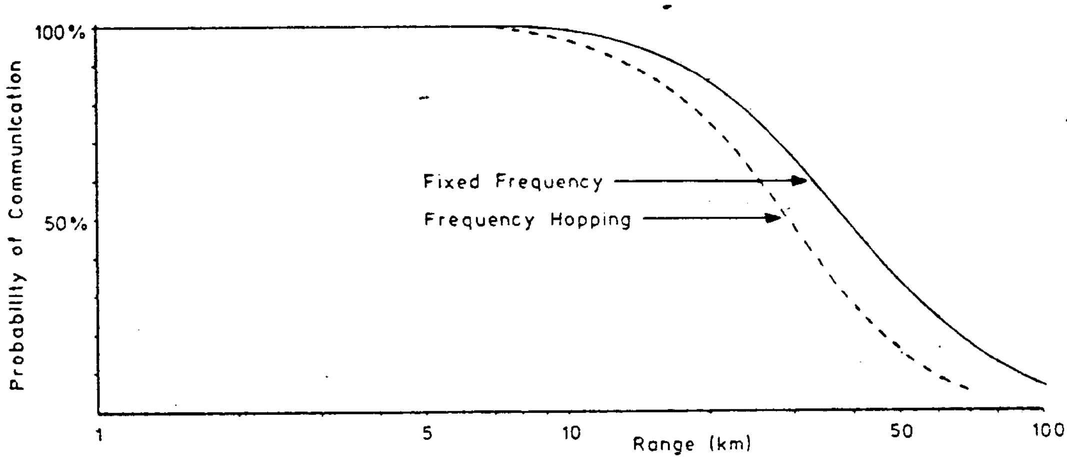


Fig. 3

Rayleigh and Large Terrain Variations

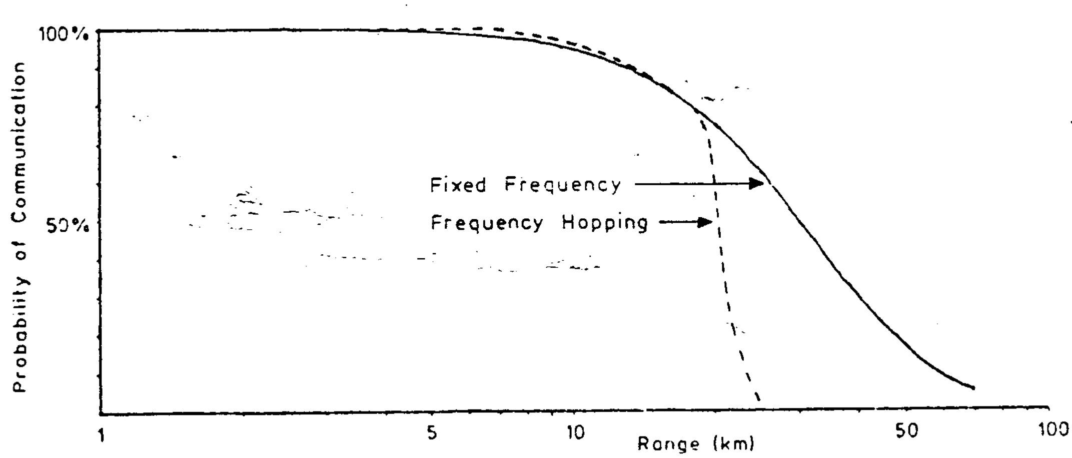


Fig. 4

Battlefield Model

ASPECTS OF CEPSTRUM ANALYSIS SIGNAL PROCESSING WITH REFERENCE TO THE MULTIPATH INTERFERENCE PROBLEM.

M.A. Jack, D. McKee, M. Flynn and M. Warden

Introduction.

Cepstrum analysis is, by definition, achieved by a serial arrangement of two Fourier transform processors. The first processor transforms from the time domain to the frequency domain yielding the spectrum of the input waveform. The second stage in cepstrum analysis involves taking the logarithm of the spectrum - a feature which allows deconvolution of echo waveforms - and the final operation is a second Fourier transform converting from the frequency domain to a pseudo-time (known as quefrency) domain. The output of this second Fourier transform yields the cepstrum of the input time domain waveform.

This paper discusses how recent developments in surface acoustic wave (SAW) device technology can be used to implement real time wideband Fourier transforms for incorporation in real time, wideband cepstrum analysers for application, in this case, to radar systems.

Cepstrum analysis.

Cepstrum analysis can be employed in two basic forms. The POWER CEPSTRUM is an effective analysis technique for determination of echo arrival times for a basic waveform distorted by echoes. The power cepstrum which is essentially a display processor which offers data in a more readily interpretable format for echo characterisation is defined[1] as the power spectrum of the logarithmic power spectrum of an input signal $\rho(t)$:

$$C(\tau) = |FT\{\log|FT\{\rho(t)\}|^2\}|^2$$

where FT denotes Fourier tranformation (spectrum analysis).

The COMPLEX CEPSTRUM is defined[1] as the complex inverse Fourier transform of the complex logarithm of the Fourier transform of a time signal. For complex cepstrum processing, signal phase information must be retained throughout. The complex cepstrum of a time waveform exhibits sharp discontinuities whose positions are determined by the echo delays. These can be removed by low pass filtering in the frequency domain. Subsequent inverse processing by taking the complex Fourier transform followed by amplitude exponentiation and complex inverse transformation yields the basic waveform recovered from its distorting echoes.

A simple description of the operation of the power cepstrum analyser is given in Figure 1 for a signal which consists of a pulse of duration T

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with an echo at delay (τ^1) relative to the main pulse. This signal may be interpreted as the time domain convolution of the basic pulse f(t) with two impulses separated by an interval τ^1. Such convolution in the time domain corresponds to multiplication in the frequency domain and taking the logarithm in the frequency domain reduces the frequency domain multiplication to a linear summation of two signals which are approximately periodic in the frequency domain. The second spectrum analysis effectively analysis the frequency periodic components of the log power spectrum to yield the cepstrum.

Surface Acoustic Wave Devices.

Surface acoustic wave devices permit efficient realisation of real time, wideband Fourier transform processors for use in cepstrum analysis[2] The Fourier transform relationship:

$$F(\omega) = \int f(t).\exp(-j\omega t)dt$$

is rewritten as an algorithm suited for surface acoustic wave implementation by means of the substitution

$$-2\omega t = -2\mu\tau t = \mu(t-\tau)^2 - \mu t^2 - \mu\tau^2$$

such that

$$F(\omega) = F(\mu t) = \exp(-j\tfrac{1}{2}\mu t^2)x$$

$$\int f(\tau).\ \exp(-j\tfrac{1}{2}\mu\tau^2).\ \exp(j\tfrac{1}{2}\mu(t-\tau)^2)d\tau$$

Thus an input f(t) is premultiplied by a chirp waveform, convolved with a second chirp waveform and finally post-multiplied with a third chirp to yield the Fourier transform by means of the CHIRP TRANSFORM ALGORITHM. For power spectrum analysis, where phase information is not required, the final chirp post-multiplier can be replaced by a square law detector.

SAW technology now permits, through the ready availability of high performance chirp filters, the realisation of Fourier transform processors using three SAW devices plus associated amplifiers and timing electronics. The effective parallel processing during convolution in the SAW chirp filter permits real time spectrum (and cepstrum) analysis with signals of duration up to 50 μs and bandwidths exceeding 10 MHz.

SAW power cepstrum analyser.

Power cepstrum analysis can be performed by coupling two SAW chirp transform spectrum analysers through a true logarithmic amplifier. The power cepstrum display of a pulse signal of duration T = 800 μs distorted by echoes is demonstrated in Figure 2 for echoes of amplitude equal to the basic pulse arriving at epochs of τ^1 = 1000 ns (Figure 2(a)) and τ^1 = 400 ns (Figure 2(b)). The power cepstrum display shows one response corresponding to the basic pulse duration at 800 ns with a further response at the echo epoch. The 400 ns echo epoch condition corresponds to a self-distortion of the waveform since the pulse and echo are partially time-coincident.

SAW power cepstrum analysis has also been demonstrated in the measurement of radar p.r.f., code lengths and bit rates[2] and in the discrimination of pulsed and CW signals[3].

SAW complex cepstrum analysis.

The chirp transform algorithm permits computation of a complex Fourier transform using SAW chirp filters[4]. The wideband, real time advantages of this Fourier transform method using SAW technology, are limited by the accuracy of the SAW realisation (6-7 bits). Preliminary results of a computer simulation of a complex cepstrum-analyser based on SAW devices are shown in Figure 3. Here a stylised radar pulse contaminated by an echo of relative amplitude 0.5 and relative delay 55 samples is shown in Figure 3(a) and the resulting complex cepstrum display is shown in Figure 3(b). Removal of the "spikes" in the complex cepstrum display, by means of interpolation, and subsequent inverse complex cepstrum processing yields the basic pulse free from the echo, Figure 3(c).

Conclusions.

SAW power cepstrum analysis has been shown to offer a method of echo arrival time characterisation in a multipath situation. SAW complex cepstrum analysis offers the potential for waveform recovery in a multipath situation.

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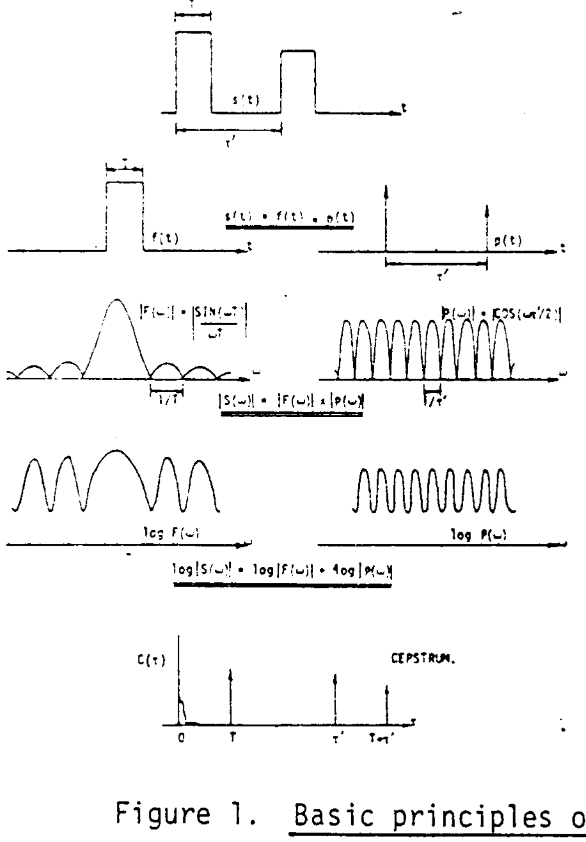


(a) Input signal consists of pulse and echo.

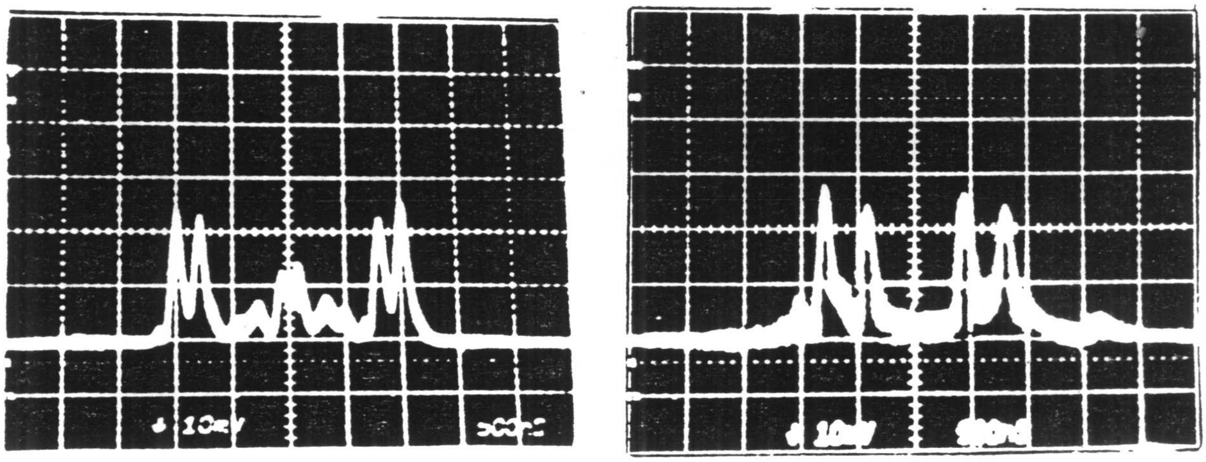
(b) Input can be considered as convolution of pulse and impulse functions.

(c) Spectrum is product of individual spectra.

(d) Log spectrum forms linear summation of waveforms.

(e) Power cepstrum displays pulse width and echo delay

Figure 1. Basic principles of power cepstrum analysis.



(a) Echo at 1000 ns delay. (b) Echo at 400 ns delay

Figure 2. Cepstrum deconvolution of pulse waveform. (pulse width 800 ns)

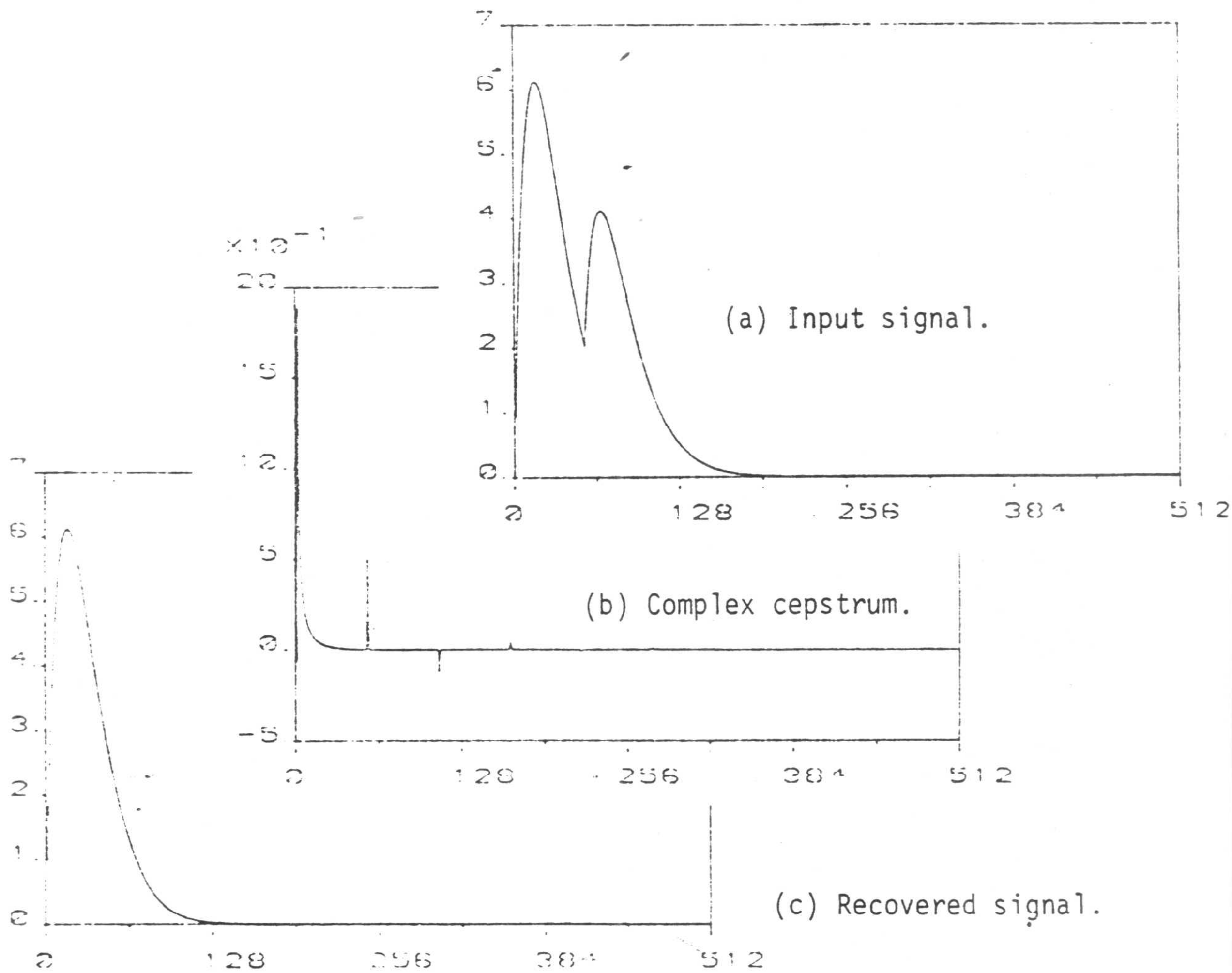


Figure 3. Waveform recovery using complex cepstrum analysis.

AM BROADCAST RECEPTION IN THE PRESENCE OF MULTIPATH INTERFERENCE AND IMPROVED DEMODULATION TECHNIQUES

E.C.Forster

Domestic AM broadcasting in most countries in the LF and MF bands is usually planned with the object of minmising multipath propagation. When, at night, significant sky wave propagation does occur it is usually considered a nuisance rather than a benefit.

The universal use of the envelope demodulator in AM radios makes reception under multipath propagation conditions very poor. This has to some extent been a deterrent to the use of the HF bands in providing extensive domestic coverage to the larger less developed countries, which could be achieved with low cost.

The use of different types of demodulator has in the past been discussed mainly with the object of improving receiver selectivity by post detection filtering or with the object of allowing a change in broadcasting from AM to some form of SSB in order to conserve the spectrum. Very little attention has been given to the performance of various demodulators under the conditions of ionospheric multipath propagation.

The general effect of multipath propagation can be understood best by considering just two possible signal paths, each having almost equal attenuation but with significantly different propagation times. The conditions which cause this are commonly the combination of ground wave and sky wave propagation, the reflection of the signal by two different ionospheric layers or the simultaneous existence of single and multi-hop propagation. But the resulting transmission characteristics are very similar and are shown by figure 1.
2 It is the result of additive and destructive interference of the two waves at the receiver. The depth of the nulls and their frequencies are time varying with the random fluctuations of the ionosphere. Significant attenuation and group delay distortion is produced in the process.

If ideal synchronous demodulation is used, the linear distortion of the received modulated signal is translated into linear distortion in the demodulated signal. An ideal synchronous demodulator is one operated by a local carrier having that frequency and phase which will yield optimum demodulation of the sidebands.

All practical AM demodulators including synchronous types convert the linear distortion of the received signal into an additional, non-linear, distortion in the demodulated output.

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Nearly all the demodulator types discussed here perform very well with undistorted AM signals but when the signal is linearly distorted some perform distinctly better than others.

Two principal results of this linear distortion of the AM signal are the production of quadrature modulation components with sideband suppression for parts of the modulation, and carrier suppression.

In the case of the envelope demodulator, by far the worst non-linear distortion is due to carrier suppression when the received carrier falls below the peak sideband envelope. This results in the familiar 'over modulation' effect and, owing to the need to maintain high transmitter efficiency by the use of high percentage modulation, it occurs at the slightest reduction of the carrier.

There are several different alternatives to the envelope demodulator and all are attempts at the ideal synchronous type. They differ in the way that the local carrier is formed.

The first regenerates the carrier by filtering the incoming carrier. This type can tolerate a lower carrier level than the envelope demodulator before the onset of gross distortion. In order to reduce the non-linear distortion in the demodulator resulting from linear transmission distortion, the carrier filter must have a bandwidth lower than twice the lowest modulation frequency. There must also be sufficient bandwidth in the filter to allow for overall frequency tolerances. High stability receivers are necessary and the results are only somewhat better than an envelope demodulator.

The second type of demodulator regenerates the carrier from information in both sidebands. A phase locked loop may be used but the error signal is dependent on having a modulated signal. Absence of modulation leads to some phase locking difficulties when the modulation reappears. This error signal contains double frequency components of the modulation which must be adequately suppressed by low loop bandwidth to prevent excessive frequency modulation of the regenerated carrier. This causes non-linear distortion in the demodulator output and it is not expected that this can be reduced to acceptable levels without causing unsatisfactory phase locking performance. There are other shortcomings. The error signal is derived from the product of the signal with itself and thus falls by 2 dB for every 1 dB fall in the sidebands, and there is also a 180 degree ambiguity in the regenerated carrier phase which may at times cause disturbing fluctuations in the demodulated output.

The third type of demodulator uses a phase locked loop to regenerate a local carrier directly from the received carrier. It has been the practice to use first or second order phase locked loops but with a type I control system. In such a system a finite error voltage is required to sustain the local oscillator control voltage. This is produced by the loop operating with a finite phase error. As the received carrier level falls the loop is forced to operate with a larger phase error until at a certain carrier level phase locking breaks down and the oscillator returns to its free running frequency. Demodulation is interrupted abruptly and with some noise. A higher gain loop, corresponding to larger loop bandwidth can tolerate lower carrier levels before phase locking breaks down and at the same time provide the pull-in and hold-in performance normally required. Unfortunately the linearly distorted AM signal is effectively phase modulated and this the high gain loop can successfully track. Non-linear distortion is caused in the demodulator output of a very similar kind to the envelope demodulator and it can only be reduced by restricting the tracking capability of the loop. This implies low loop gain which then compromises the pull-in and hold-in performance of the system. High stability receiver techniques become necessary and the system becomes unpractical.

The type of demodulator which, it is believed, is a practical success uses a type II phase locked loop having a near perfect integrator in the feedback loop, as shown in figure 2. Because it is a zero error control system in the steady state, lack of carrier is not unduly disturbing to the continuing demodulation of the sidebands providing that the carrier returns before the finite practical offsets have caused the local carrier phase to drift too far. In reality the linear transmission distortion is very dynamic and it is difficult to predict the behaviour of the system under actual conditions. Moreover, the multipath interference model suggests that at certain times there will be a complete rotation of the carrier phase relative to the sidebands. This occurs when the relative strengths of the two received components are reversed. It would be expected that when the local carrier reaches a phase error of 90 degrees from the ideal phase, that the demodulator output would fall to zero but tests show that the output hardly ever falls by more than 10 dB. This would seem to be due to the quadrature modulation components introduced by the linear transmission distortion.

The integrator performs the function of a memory, but it also allows the loop bandwidth to be reduced sufficiently to limit non-linear distortion to less than 1 percent and yet still allow the loop to track ionospheric doppler shifts. Practical tests show that the demodulator performs very well and it seems to be preferable for the carrier tracking to be fairly accurate.

The diversity advantage inherent in DSB transmission can be largely obtained, especially the reduction of linear distortion in the demodulated signal. This is shown in figure 3. The 3 dB noise advantage can only be obtained with low linear transmission distortion and low phase error from the ideal local carrier phase, but useful improvements are evident at times.

Certain aspects of this demodulator are covered by a recent UK patent application.

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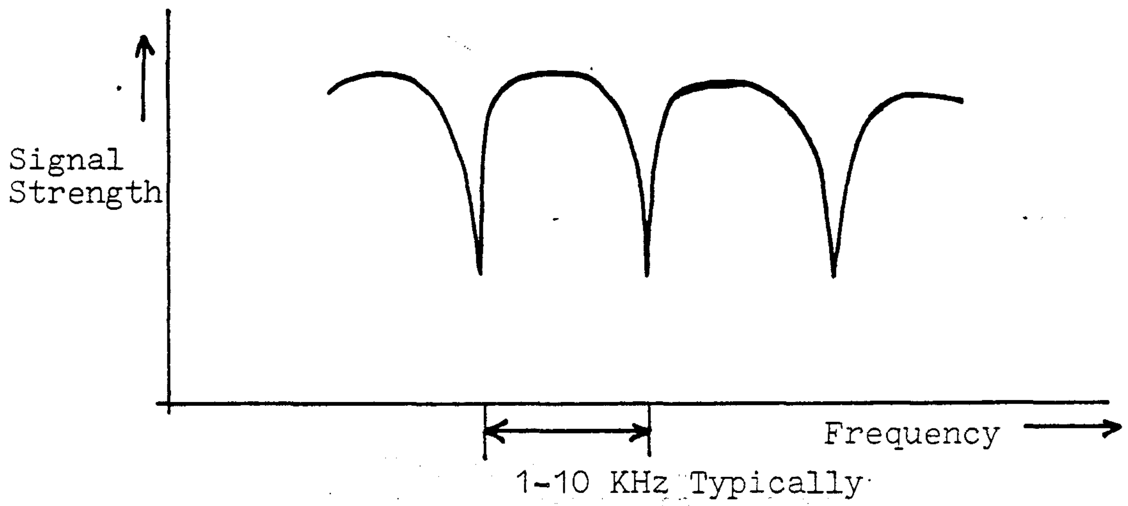


Fig.1 Ionospheric multipath transmission distortion

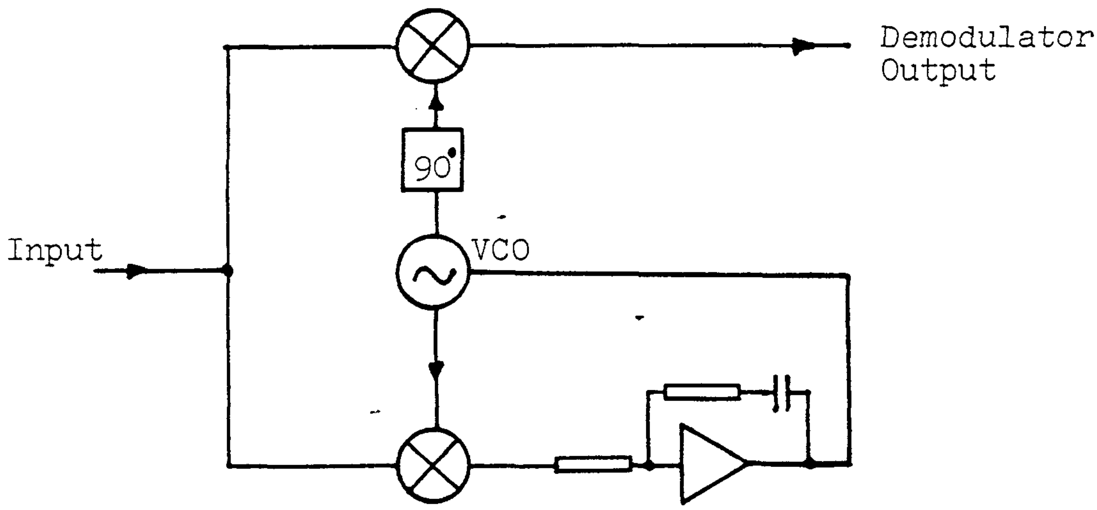
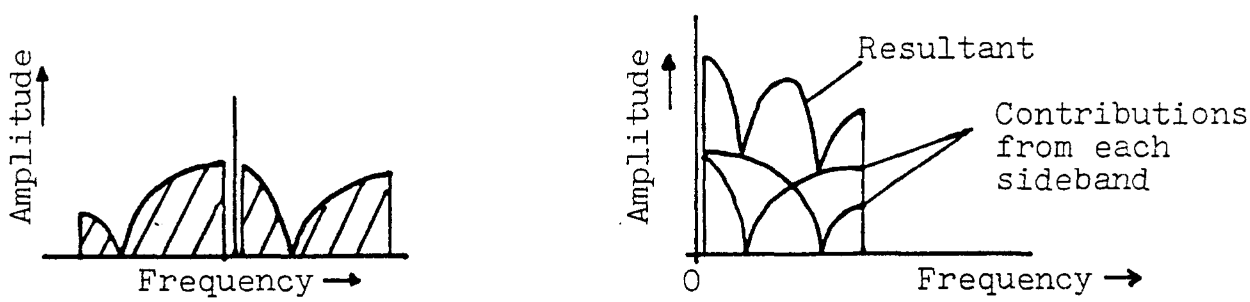


Fig.2 Phase locked demodulator with integrator



(a) Linearly distorted
AM signal spectrum

(b) Demodulator output

Fig.3 Diversity advantage in synchronous demodulation

Multipath propagation in the urban mobile radio channel

P.A.Matthews and G.N.Robinson.

1.0 INTRODUCTION

There is an extensive literature on multipath channels and fading in radio systems. Much of this is directed towards troposcatter or h.f. systems in which the propagation medium is time varying. However the propagation medium in l.m.r. systems is not normally time varying. For comparatively short paths for which the principal propagation mechanism is diffraction, reflection, or scattering from buildings or natural obstaclesthe propagation medium is static.

A static transmitter will lay down a static electromagnetic field over the ground. A vehicle moving through this field will receive a time varying signal. Components of this signal arriving from different scatterers will be subject to different Doppler shifts on reception at the vehicle. The total signal varies in amplitude as the phasor sum of the components varies. For a number of random components the probability distribution for samples of the field distributed in space will be Rayleigh. A vehicle moving through the field will receive a time varying signal with the same distribution. If there is one dominant component the distribution will become Rician.

A vehicle in the field will be a scatterer. If the vehicle moves in the field the scattered signal will be time varying and Doppler shifted. Locally the vehicle will produce a time varying field which may be received by an adjacent stationary or moving antenna.

If the transmitter is itself moving then the electromagnetic field will be time varying There will also be Doppler shifts of the received components because of the motion of the transmitter.

Thus in general the received signal is time varying, Doppler shifted, and comprised of components arriving with different time delays. This has been illustrated by Parsons and by Cox,[1].

2.0 SPATIAL MODELS

The simplest case is that of a static transmitter. The field pattern in space can be modelled by setting up a distribution of scatterers in space and calculating the resulting field distribution. The progress of a vehicle through this field can be modelled by sampling the field along a given path. In general the amplitude will be randomly distributed and there will be a non-uniform progression in phase. A spectral analysis of the sampled field will show the spatial frequency pattern along the path.

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This in turn can be related to the temporal frequency at the receiver.

To resolve the multipath time delay spread a wideband signal must be modelled. Then the samples along the path will vary with both position and frequency. A two dimensional distribution is then available the transformation of which will give the spatial frequency and time delay distributions.

A spatial models is expected to be of use in analysing the performance of different modulation systems as the model is of the field distribution, not of the response to a particular signal.

3.0 TEMPORAL MODELS

As a principal problem in a l.m.r. system is the time varying signal received at a moving antenna a temporal model of the path is useful. Fading dispersive media are well described in the literature,[2]. Of the different types of channel described the l.m.r. channel has received relatively little attention.

Three types of model may be set up, the delay line model, the diversity model, and the scattering model,[2]. The delay line model, fig1, models directly the relative time delays of scatterers and their scattering properties.

4.0 PRACTICAL MODELS

If the performance of real equipment is to be assessed and compared a practical model of the radio frequency path is required. A variety of path simulators have been described having more or less controllable characteristics.

The model should, whilst modifying a signal transmitted through it with the random characteristics of a real channel, be controlled in a reproducible way so that different trials can be carried out under set conditions.

We have under construction a time delay model, fig2, which will allow us to produce signal components with known time delays and known amplitude and phase probability distributions. This is achieved by controlling the in-phase and quadrature tap weightings by a microcomputer system. The microcomputer generates the control signals for the tap weightings. The control signals are modified to take account of the measured performance of the tap weighting circuits. In this model not only Rayleigh fading but also other fading distributions may be generated.

The model under construction will pass r.f. signals in the low v.h.f. band, 50MHz to 90MHz. To use the model at other frequencies down and up conversion of the r.f. signals would be necessary or other models could be constructed with the appropriate tap weighting networks.

5.0 PERFORMANCE OF THE MODEL

Tests have been carried out to show the range of operation of the model. It will be recognized that each tap weighting network is essentially a s.s.b. modulator. Such a modulator fed with in-phase and quadrature signals will give a frequency shifted output. This is required to simulate Doppler shifts. The output of the model for such a modulation is shown in fig3.

The time varying signal at the output of a single tap is shown in fig4. This tap is operating under computer control. The corresponding probability distribution is also shown.

Because the model operates under computer control the performance is reproducible. In principle by sampling the signals on a real system suitable control parameters can be measured which when fed into the control system of the model allow reproduction of the propagation channel of the real system.

6.0 COMPARISM WITH OTHER MODELS

Work on other models and simulations has been carried out, particularly by Turin,[3]. This work will be discussed further in the presentation.

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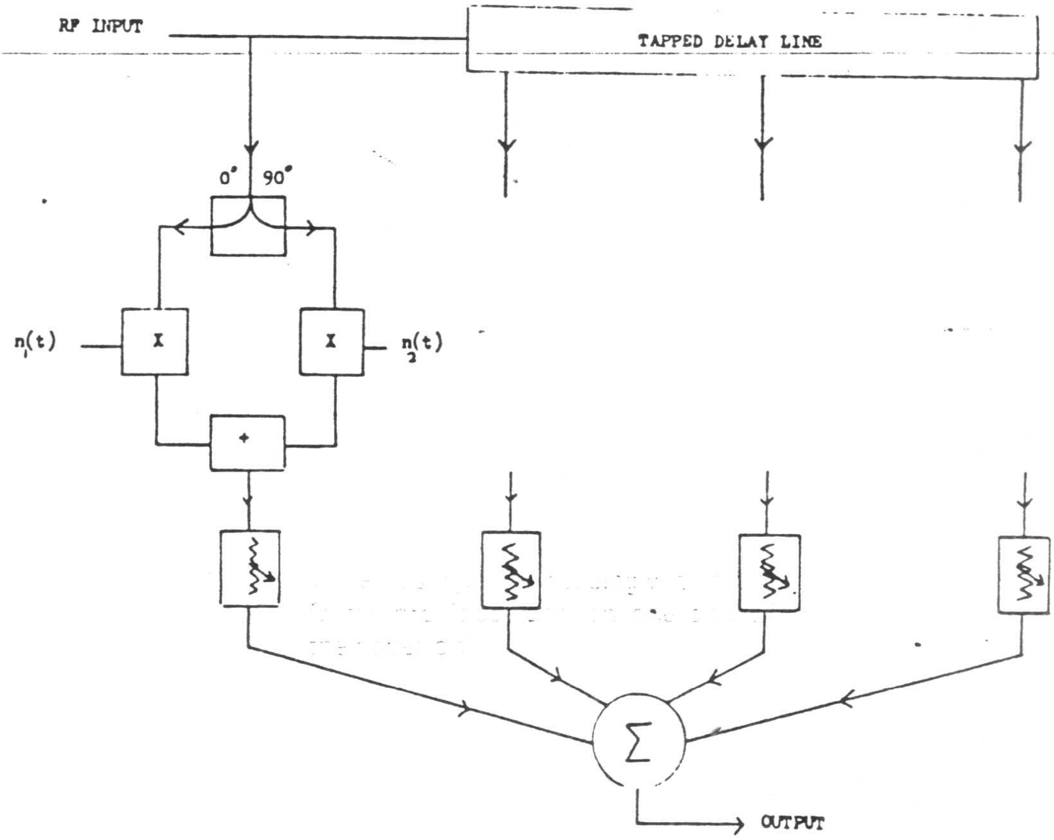

Fig 1. Tapped delay line model.

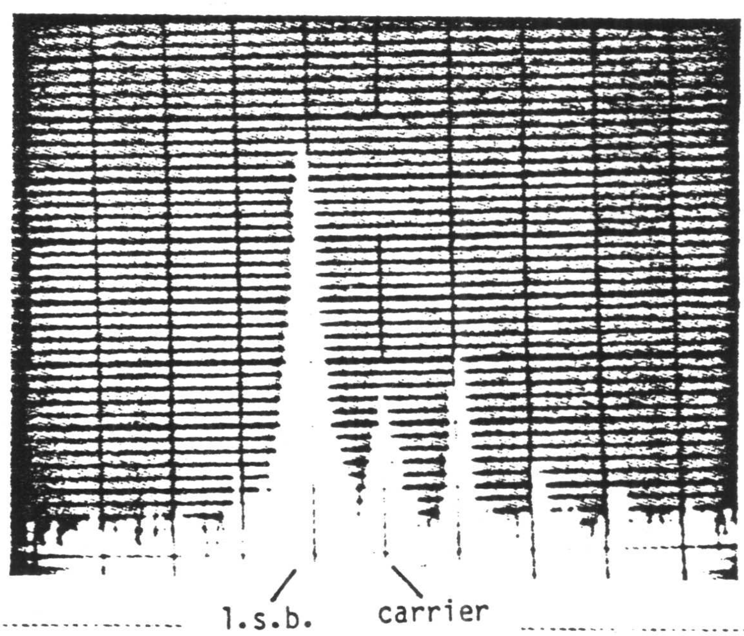

Fig 2. Tap output, frequency shifted.
Carrier 70MHz, modulation 500Hz.
Horizontal 500Hz/div.
Vertical 10dB/div.