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COUNTERACTION OF MULTIPATH INTERFERENCE BY A COMBINATION OF BEAMSTEERING AND ADAPTIVE EQUALIZATION

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INTRODUCTION

High speed hydroacoustic data transmission are strongly interfered by multipath propagation due to reflections from the sea surface, the bottom or obstacles from offshore installations. In order to achieve bandwidth effective transmission, the communication system has to adapt itself to the current hydroacoustic channel, thus rejecting the multipath interference or compensate for it. Adaptive beamforming is an example of the rejection technique while equalization is a compensation technique. By being different in nature these techniques offers partly complementary qualities thereby fitting well in a combined system. This paper presents simulation results showing the performance of a combined system and how the two described techniques supplement each other in combating the multipath propagation interference.

THE IMPACT OF ADAPTIVE BEAMFORMING AND ADAPTIVE EQUALIZATION ON THE CHANNEL IMPULSE RESPONSE

Beamforming/beamsteering

As described in the previous section, reflections from surface, bottom and in-sea obstacles may cause the transmitted signal to reach the receiver via different paths. These rays will usually approach the receiver from different angles. It is therefore possible to attenuate some of these unwanted signals by applying a narrow transducer beam directed towards the direct path.

When applying a simple tracking algorithm [1, 6] together with traditional beamforming [2] unwanted signals (including multipath signals and noise sources) outside the main lobe are attenuated according to the side-lobe level at the actual angle of incidence. Further suppression are possible by using Widrow's side-lobe canceller [3] or an optimum beamforming method [2, 4, 5].

An interesting aspect of the beamforming technique is revealed by observing a relation between incidence angle and propagation delay. This is illustrated in Fig. 1.

The longest possible path from transmitter to receiver via one single reflector within the joint area is the route via one of the area borders. Consequently the longest delay relative to the direct path depends heavily upon receiver and transmitter beamwidths. By simplifying the channel impulse response to consist only of the direct path together with single reflector paths within the joint area, this also gives a limit of impulse response duration.

From the figure we calculate the delay difference between the direct path and the maximum delayed path as

$$\Delta\tau_{max} = \frac{l_1 + l_2 - r}{c} = \frac{r}{c}\,\frac{\sin\frac{\varphi_1}{2} + \sin\frac{\varphi_2}{2} - \sin(\frac{\varphi_1+\varphi_2}{2})}{\sin(\frac{\varphi_1+\varphi_2}{2})} \qquad (1)$$

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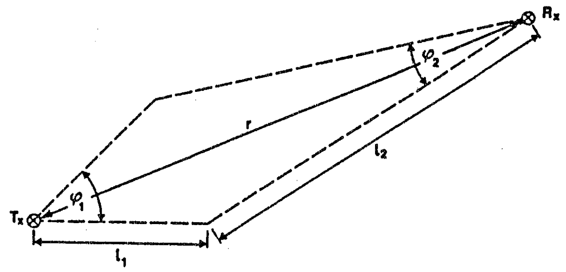

Figure 1. Geometry of area covered by transmitter and receiver mainlobes.

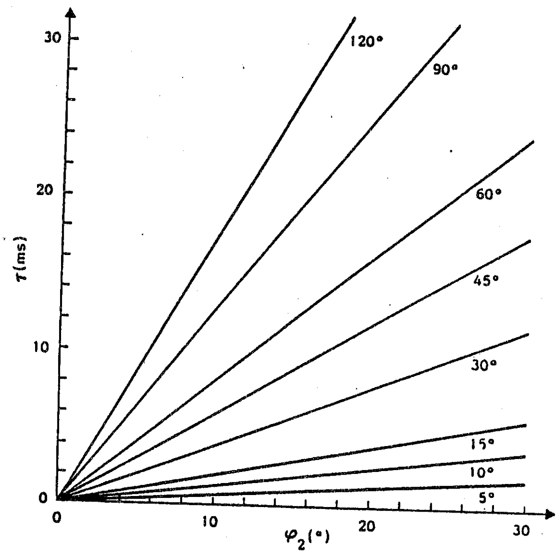

Figure 2. Delay difference as a function of transducer beamwidth.

where c is the sound speed in sea water.

Fig. 2 depicts the maximum delay difference as a function of beamwidths at 500 m distance between the transducers.

From this we deduce that the impulse response duration may be forced as small as wanted by making narrow transducers.

Now installation and alignment of the transducers may become difficult. The receiver transducer may be controlled automatically by some beam steering algorithm thus allowing fairly narrow beams. However, the tracking requirements increase with decreasing beamwidth and practical considerations also limits the possibility to manufacture transducers with mainlobe width less than about 5^{o}. Of course the transmitter transducer also allows for automatic alignment but this will require some kind of signalling from the receiver. In addition the transmission delays will limit the tracking capability of such a system. When using no kind of steering of the transmitting transducer, the width of the main lobe will limit the area of operation. A beamwidth of 60^{o} is considered as a practical choice. Now returning to Fig. 2 this gives a maximum impulse response duration of about 4 ms. That is 40 symbol intervals when transmitting symbols at rate 10 kHz (20 kbits/s with 4 PSK modulation). During operations near bottom or in between structures, reflectors may very well appear within the area covered by the mainlobes of both transducers. Thus some intersymbol interference is likely to be present in the received signal. This motivates the use of an adaptive equalizer together with the transducer system.

<u>Adaptive equalization</u>
While the narrow beam transducer prevents intersymbol interference by attenuation of auxiliary paths, the equalizer attempts to compensate for the resulting channel transfer function.

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Assume the following model of the received signal

$$y(t) \quad = \quad \underline{h}_N^H \, \underline{d}_N(t) + n(t) \tag{2}$$

where $y(t)$ is received signal sample at time t, $\underline{h}_N$ is the $N\text{x}1$ Channel impulse response vector and $\underline{d}_N(t)$ is the vector of the N last transmitted data symbols, $n(t)$ is an additive white noise term. Superscript H means Hermitian transpose.

$$\underline{h}_N \quad = \quad [h_0, \; h_1, \; \ldots \; h_{N-1}]^T \tag{3}$$

$$\underline{d}_N(t)[d(t), \; d(t-1), \; \ldots \; d(t-N+1)]^T \tag{4}$$

Now let

$$\underline{y}_L(t) \quad = \quad [y(t), \; y(t-1), \; \ldots \; y(t-L+1)]^T \tag{5}$$

Then

$$\underline{y}_L(t) \quad = \quad \underline{H}_{M,L}^H \, \underline{d}_M(t) \tag{6}$$

where

$$H_{M,L} \quad = \quad \begin{bmatrix} \underline{h}_N & 0 \; \ldots . & 0 \\ 0 & \underline{h}_N & ' \\ ' & 0 & 0 \\ ' & ' & \underline{h}_N \\ ' & ' & 0 \\ ' & ' & ' \\ 0 & 0 & 0 \end{bmatrix} \tag{7}$$

is the MxL matrix of Toeplitz structure constructed as described in Eq. (7).

With these signal definitions we turn to the well known normal equation [4] to calculate the optimum equalizer impulse response vector $\underline{W}_K$.

$$\underline{W}_K \quad = \quad [E(\underline{y}_K(t) \; \underline{y}_K^H(t))]^{-1} \; E(\underline{y}_K(t) \; d^*(t-T)) \tag{8}$$

$$\underline{W}_K \quad = \quad (\underline{H}_{M,K}^H \, \underline{H}_{M,K} + \sigma^2 I)^{-1} \; \underline{H}_{M,L}^H \begin{bmatrix} 0 \\ \vdots \\ 1 \\ \vdots \\ 0 \end{bmatrix} \tag{9}$$

where σ^2 is the noise variance, and $\underline{I}$ is the unity matrix.

The remaining interference is given by

$$\varepsilon = 1 - \underline{W}_K^H \; \underline{E}(y_K(t)\; d^*(t)) \tag{10}$$

$$\varepsilon = 1 - \underline{W}_K^H \; \underline{P}_K \tag{11}$$

where

$$\underline{P}_K = \underline{H}_{M,L}^H \begin{bmatrix} 0 \\ \vdots \\ 0 \\ 1 \\ 0 \\ \vdots \\ 0 \end{bmatrix} \tag{12}$$

When increasing the equalizer order K, ε asymptotocally approaches a value slightly greater than σ^2.

SIMULATION PRELIMINARIES

Channel model

The channel model for high frequency underwater data communication is based on a ray-tracing model, and includes the most important phenomena in underwater acoustic transmission. The modelled phenomena are sound refraction due to vertical sound velocity variation, reflections from the sea floor and from the surface, diffuse reflection due to surface waves and finally the surface motion. Various types of transducer diagrams can be modelled. Single element transducers as well as line arrays are available. The bottom is considered plane and horizontal, given by its sound speed and density. Thus, the reflection coefficient is complex, and depends on the angle of incidence. The channel model is designed to investigate point to point transmission conditions and utilizes a dedicated direction algorithm to find the possible ray paths from transmitter to receiver. The sound speed variation is described by a piecewise linear velocity gradient.

Surface reflection. The direction algorithm computes the possible ray paths from transmitter to receiver when the transmission channel is bounded by the plane bottom and a plane ocean surface. A rough surface, given by a surface function $S(x,t)$, changes the surface reflected sound characteristics. The surface function is split into two parts, the large scale wave function, and the corrugation function. The wave function is modelled as a sum of sinusoids with variable frequencies and amplitudes. This time dependent function makes specularly reflected ray paths possible via a number of surface segments, each path carrying limited energy.

The corrugation function is stochastic, and represents small capillary waves on the surface. In our model, the corrugation function is not time dependent. The mean wave amplitude of this function should be less than one sound wavelength. These capillary waves cause the occurance of diffuse surface reflections. The resulting pressure of the surface reflected sound at the receiver is found by a surface integration of incident sound pressure compensated in amplitude and phase due to variations in propagation range, surface angle, incident angle, reflection angle, transducer directivities, etc. The following algorithm describes how the integration is performed. The ray path is known through the specular reflection point of the plane surface case. Now, the surface is divided into a number of segments which are short enough to be considered plane. Ray paths are traced via some of these segments. The number of paths depends on the wavelengths of the surface wave. The first segments computed are those nearest the specular point. Surface segments further away are then successively included, until the pressure contribution is negligible.

Channel data

During our work, we have put some effort in investigating how the channel parameters influence the transmission conditions. The channel simulations done in this context represent typical transmission channels for ROV - surface communication. The horizontal range is varied to achieve a set of different channels.

Channel parameters:

Sea depth	:	50 m	Salinity	:	35 o/oo
Bottom sound velocity	:	2500 m/s	Bottom density	:	3000 kg/m^3
Wave function	:	A = 0 m	Capillary wave	:	σ = 1 mm
Carrier frequency	:	150 kHz	Transmitted effect	:	10 W
Transmitter depth	:	47 m	Transmitted beamwidth	:	60 deg.
Receiver depth	:	5 m	Receiver beamwidth	:	10 deg.
Bandwidth	:	1 kHz and 10 kHz			
Horizontal range	:	100 m/150 m/200 m			

Velocity gradient:

Winter gradient taken by M/K Simrad near Horten, Norway, 24.11.1986, 14:00, and approximated to a piecewise linear function. The gradient is shown in Fig. 3.

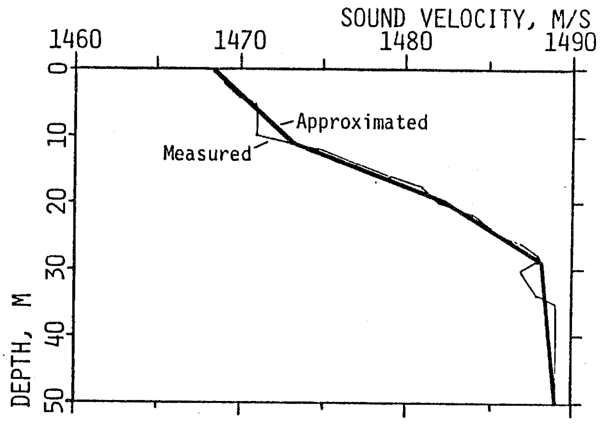


Figure 3. Sound velocity profile.

The receiver transducer with beamwidth 10 deg. points along the ray path from the trasmitter while the transmitter transducer is centered about the horizontal plane. The two transducer diagrams are quite simple. The main lobe level is constant as is the side lobe level with 20 dB attenuation. Because of this simplified model, the conditions stay constant for receiver beamwidths from 5 to 15 degrees, since the number of different ray paths inside the main lobes remains the same.

Resulting impulse response

An example of a simulated impulse response is showed in Fig. 5 a-b. This curve shows the module of the complex impulse response. The horizontal range of the case is 150 m. The bandwidths are 1 kHz (a) and 10 kHz (b). The shape of the curves is characterized by the different ray path arrivals. In the 150 m case, the delay of the directly transmitted signal is 105.5 ms. The bottom reflection arrives at 106.1 ms, the surface reflection at 107.2 ms, and the bottom/surface (BS) reflected ray at 108.6 ms. Because of the plane surface and the very small capillary wave, the surface reflections appear to be quite distinct. The next group of paths arrives at about 145 ms and contains the SB, SBS, BSB and BSBS rays. The SBSB, SBSBS, BSBSB and BSBSBS arrive at approx. 195-200 ms, and the last important group of rays, those that have crossed the sea depth seven times, give its contribution at 250-260 ms. Further multiple reflections are neglected by the model program because of low energy. The distinct peaks are followed by exponentially decaying tails which are typical for the diffuse reflections. The shorter the horizontal range, the larger is the time spacing between the ray groups, but the multiple reflections are more heavily attenuated relative to the direct sound.

SIMULATION RESULTS

We shall now examine the results from a simulation example. Block diagram of the simulations are given in Fig. 4.

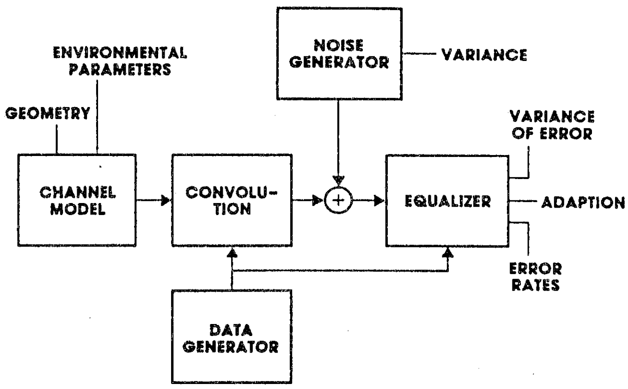
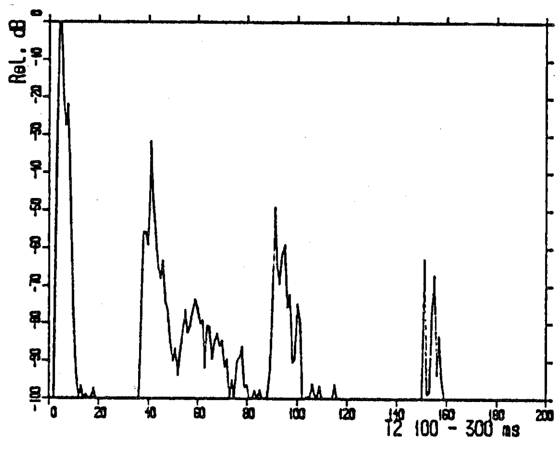


Figure 4. Block diagram of simulation system.

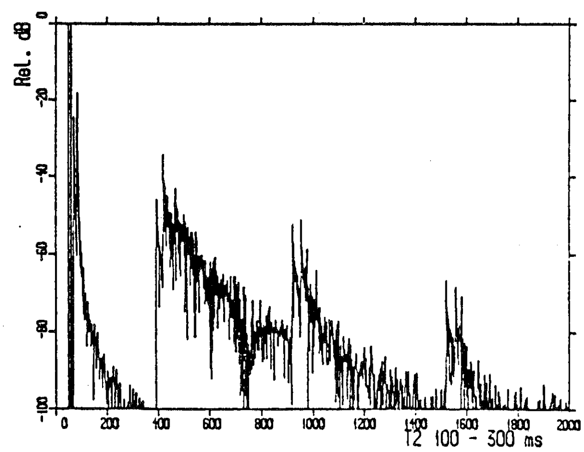
The horisontal distance is varied between 100 m and 200 m. This variation will only affect the impulse response and not the signal to noise ratio as the channel simulation produces noise free signals to which noise is added afterwards. Similarly the receiver beamwidth is varied between 5^{o} and 15^{o}. Also this variation is done without affecting the signal to noise ratio.

The transmitter beamwidth and center direction are chosen in a manner which sites both the direct path and the bottom reflection within the transmitter transducer main lobe. This is done to produce a "difficult" channel.

Fig. 5 shows the resulting impulse response at 150 m horisontal distance. The receiver beamwidth does not affect the channel very much. This is due to the special geometry. That is, there are no significant signal contributions approaching the receiver in the intervalls between $\pm 2.5^{o}$ to $\pm 7.5^{o}$ relative the beam axis. The strong bottom reflection thus reaches the receiver at an



a) Bandwidth 1 kHz.



b) Bandwidth 10 kHz

Figure 5. Resulting impulse response at 150 m horizontal distance and receiver beamwidth 15^{o}.

incident angle less than 2.5^{o} away from the direct signal path.

Now turning to equalizer performance. Fig. 6 shows the theoretical calculated mean square error at the output of an adaptive equalizer for orders between 50 and 300 and at different signal to noise ratios.

This is a result of applying Equations (8) - (12) to the simulated channel impulse response. We observe that the curve converges to a value slightly greater than the actual noise variance. The convergence appears for smaller order values when decreasing the signal to noise ratio. We also note that for S/N = 30 dB or poorer, an equalizer of order 80 is appropriate to reach the optimum performance.

Fig. 7 shows the learning characteristics of the stochastic gradient lattice equalizer [7, 8]. As noticed, the asymptotic mean square error is about 3 dB above the theoretical value. This is due to the noise caused by fluctuations of the equalizer coefficients.

Finally we present detection performance of the receiver with and without the adaptive equalizer. These results are presented in Table 1.

We notice that the error rates obtained from <u>noise-free</u> transmission through the channel are unacceptable for communication. On the contrary, by adding the adaptive equalizer to the system, the bit error rates turns considerably lower and allows for reliable transmission. This confirms that even with a narrow beam (5^{o}) receiver transducer multipath propagation may introduce severe intersymbol interference. Further it is shown that by including adaptive equalization reliable communication is still achievable.

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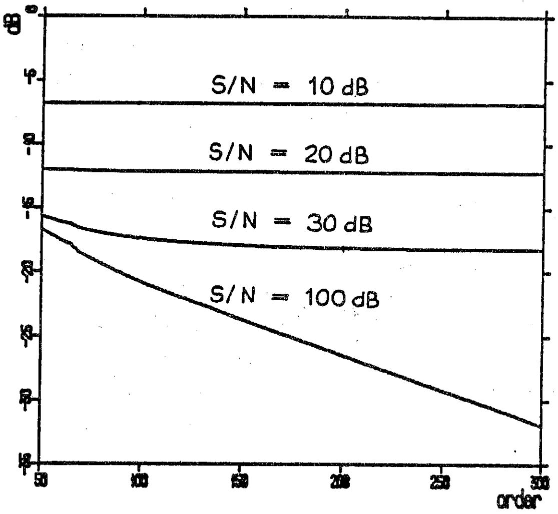

Figure 6. Minimum mean square error versus order.

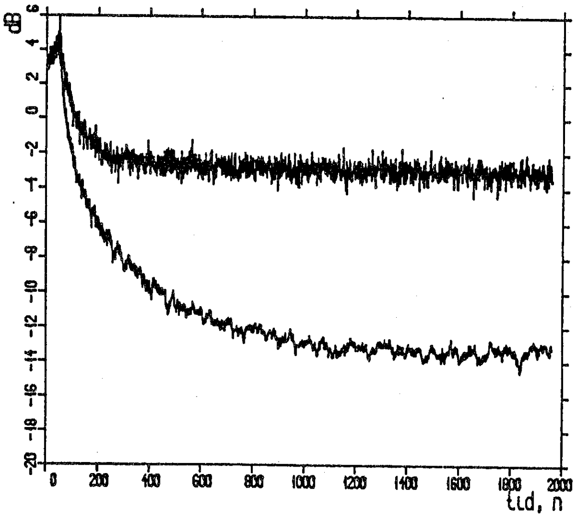

Figure 7. Learning characteristics of the stochastic gradient lattice equalizer.

Table 1. Error rate measured for different channels

a) Noise-free transmission without equalizer

Horizontal distance	Receiver beamwidth	Bit error rate
150 m	15^o	0,25
150 m	10^o	0,25
200 m	10^o	0,12

b) Simulations applying equalizer

Horisontal distance	Receiver beamwidth	S/N	Bit error rate
150 m	15^o	20 dB	$1{,}6\ 10^{-4}$
150 m	10^o	20 dB	$1{,}6\ 10^{-4}$
200 m	10^o	10 dB	0,0046
200 m	10^o	20 dB	$1{,}6\ 10^{-4}$
200 m	10^o	30 dB	0

CONCLUSION

This paper has pointed out potential features of a digital coherent receiver consisting of an adaptively steered antenna feeding an adaptive equalizer. Because a narrow beam transducer array causes shortening of the impulse response duration, an adaptive equalizer of moderate length will yield good receiver performance. It is also shown that even a receiver beamwidth of 5^o may allow severe multipath interference. Thus the combined system offers potential advantages for a high bitrate link between a ROV and a surface vessel. Especially during operations near the sea floor or in between offshore installations are situations likely to appear where simpler systems capitulate while the described system still renders a good connection.

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UNDERWATER ACOUSTIC COMMUNICATIONS: A REVIEW AND BIBLIOGRAPHY

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INTRODUCTION

The literature surrounding the subject of underwater acoustic communications is, in some respects, surprisingly scant. This is particularly the case if one concentrates only upon that material directly concerned with actual underwater communication systems. For example, covering that particular area and excluding general review papers whilst searching back over the past two decades, the authors have retrieved only some sixty titles.

In this paper we have deliberately chosen to take a broader view of underwater communications, to include consideration of the acoustic channel. In so doing, we have drawn upon material which, if not exactly remote from the topic of communications, is none the less of substantially wider scientific interest and application. The paper thus divides into four broad areas of interest, which are "Review and Fundamental Papers", "The Channel", "Engineering Aspects of Underwater Communications" and "Specific Communication Systems".

It must be said that, insofar as the first and second of these areas are concerned, there has been a distinct need to prune the available material, in order to fit within the space allocated to this paper. Hopefully the pruning and the organisation of the material as a whole, although idiosyncratic, will yet leave a useful collection of references for those who wish to pursue further the subject of underwater acoustic communication.

REVIEW AND FUNDAMENTAL PAPERS

In compiling this section, it has been necessary to eliminate a significant number of titles which only provide a low-level review for a general audience. That stated, one such paper by Anderson [1] is included, since it sets the scene in underwater communications as of two decades ago, and neatly reviews the major difficulties which, then as now, revolve around the problems of reverberation and multipath transmission and high attenuation at high acoustic frequencies. Quazi and Conrad [2] also contribute an interesting historical insight and make the suggestion that parametric transmission, because of its ability to establish pencil-beam transmission at relatively low frequencies, with physically small transducers, might have particular advantage in avoiding surface and sea-floor reflections and thus minimising or eliminating the corruptive effects of multipath transmission.

Parametric sonar might be less attractive than Quazi and Conrad suggest, since high directivity at a frequency approaching one of the parametric primaries is, in any case, readily achieved using conventional transmission. One of the remaining advantages of the parametric method is that transmissions using the lower, secondary frequency are less strongly attenuated, in water, than conventional transmissions at the primary frequency. For many applications this advantage would be offset by the poor power efficiency of parametric conversion. Another potential advantage is the possibility of making use of the extreme frequency agility of the secondary frequency. At least in sofar as bandwidth is concerned, the absolute width of sweep of the secondary cannot in any case exceed the primary bandwidth. Finally, the added complexity of a parametric projector would increase cost and could adversely affect robustness.