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INSTITUTE OF ACOUSTICS

UNDERWATER ACOUSTICS GROUP

PROCEEDINGS OF THE CONFERENCE

SIGNAL PROCESSING IN UNDERWATER ACOUSTICS

HELD AT:

LOUGHBOROUGH UNIVERSITY OF TECHNOLOGY

21st and 22nd MAY 1980

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Proceedings of The Institute of Acoustics

A HIGH RESOLUTION SUB-BOTTOM PROFILER FOR DEEP WATER USE

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Abstract

There is an assured place in marine geophysics for a profiling system capable of high resolution in the top 100 m of sediment in full ocean depths of 5 km or more, and it cannot be taken by the commercial systems designed for shallow water. This paper describes the approach taken at I.O.S.

Introduction

High resolution sediment profiling in very deep water with commercial profilers fails through poor signal/noise ratio for three reasons. First, the size of the transducers imposes a limit on the acoustic power which can be radiated, second, the transducer as a receiver provides poor discrimination against locally generated noise and third, the large distances involved incur heavy spreading losses. Attenuation in the water column is not significant, since fairly low frequencies are chosen to keep the attenuation in the sediment to acceptable levels. The literature refers almost exclusively to carrier frequencies of 3.5 kHz, but the good deep water results are without exception obtained with fairly large arrays of efficient transducers, and very little information is divulged about the details of the array. The opportunity for I.O.S. to enter the field arose when a small number of suitable transducers became available. The following sections describe the overall design process, the deployment of the transmitters, the receiving hydrophone design, and the signal handling arrangements.

The Basis of the System

The process of high resolution profiling below the seabed is subject to a number of conflicting restraints. On the one hand the minimum useful penetration (say 100 m) dictates a low carrier frequency due to the very high levels of acoustic attenuation, while on the other hand high resolution dictates a wide band of frequencies for the signal. The resultant low Q high efficiency requirement at frequencies in the low kHz region is difficult to meet with present transducer technology.

The time resolution requirement is about 2 msecs and the lowest carrier frequency at which a transducer can be efficiently matched over the necessary 500 Hz-1 kHz band is at least 2 kHz. The transducers available at I.O.S. were tuned at 2.25 kHz and can be driven over a band of several hundred Hz. Only three elements were available and because of their rarity and size it was clear that the cheapest and most effective way to use them would be as hull mounted transmitters, with a separate line hydrophone deployed as a receiving array. It is clear that the transmitters must be disposed athwartships in the hull to obtain a composite pencil beam directivity pattern. The peak power available from the transducers is limited by cavitation, and can be increased by an increase in hydrostatic pressure. To take advantage of this the transducer elements were mounted in the double bottom tanks of the NERC research vessel, RRS Discovery, and a head of water was maintained by a standpipe to a height of 5 metres. An added advantage was that the tanks contained freshwater ballast so that the corrosion problem was minimised. The outer bottom plates present no barrier to sound at this frequency, but the arrangement does have the domestic disadvantage of high coupling to the ship's internal spaces, and this problem is not yet satisfactorily solved. The prime need at this period was to avoid the delay and expense of developing a towed vehicle to house the fairly bulky transducers.

Each transducer was driven by a dedicated power amplifier capable of delivering 1 Kwatt. It was not possible to mount the transducers side by side, but the spacing was small enough to maintain the transverse sidelobes below a tolerable level.

The Acoustic Calculations

The calculations proceed on the assumption of a flat series of sedimentary layers which is a fairly close approximation to the situations in which high resolution SRP is most frequently used. One calculates the effective source level taking into account directivity gain, power level and processing gain, then one subtracts spreading loss (calculated at 6 dB for doubling for the two way journey as opposed to 12 dB for doubling for the single journey; this is because of the assumption of plane horizontal reflecting layers), water column attenuation, hydrophone noise (corrected for spectral width and array directivity) and reflection losses between the various layers (sea/sediment loss counted twice and sediment/sediment loss counted once). The final figure must be positive for a successful system and is the nett budget available for attenuation in the sediment. In the following example the conditions are: depth 4000 m, acoustic power 900 watts (= 300 per element), pulse length 28 ms, bandwidth 900 Hz. The ship noise spectrum level at 100 m is approximately -20 dB re 1 μBar/$\sqrt{Hz}$, and reduced by the 40 dB average discrimination of the hydrophone over the band the resultant level is -60 dB. The isotropic sea state spectrum level at sea state 5 is -42 dB and the hydrophone directivity is 12 dB, so the resultant level is -54 dB. The two incoherent noise sources differing in level by 6 dB combine to give a resultant level 1 dB above the higher level, i.e. -53 dB. As it happens the resultant noise level is little different at 10 knots in sea state 2. Flow noise is not a significant factor in the conditions specified with the tube diameter and elements used.

The on axis source level referred to 1 metre is in decibel notation

$$SL = +71.6 + 10\log_{10} I + 10\log_{10} A$$ where I = Intensity in watts/cm²

A = Total active face area.

To this must be added the processing gain $10\log_{10} BT = 14$ dB. Since $I = 0.5$ w/cm² and $A = 2200$ cm² we get

$$SL = +116 \text{ dB re } 1\ \mu\text{Bar at } 1 \text{ metre.}$$

The losses are:-

Spreading 20 $\log_{10}$ 8000	=	78 dB
Water path attenuation (0.13 dB/km)	=	1 dB
Sea/sediment loss (twice)	=	5 dB
Sediment/sediment loss	=	25 dB
	=	109 dB

The reflection losses are based on an impedance (PC) ratio of 0.6:1 between the sea and sediment and 0.9:1 between two sediments.

The effective noise level is given by

$$N = NL + 10 \log_{10}B$$
$$= -53 + 29.5$$
$$= -23.5 \text{ at 8 knots in SS 5.}$$

And the difference between this and the source level reduced by the transmission losses of 109 dB (= +7 dB) gives the budget available for loss by sediment attenuation while leaving a zero S/N ratio on the receiver output. At this value of S/N ratio a continuous reflecting horizon is clearly visible on a facsimile recorder. The budget figure is thus 30.5 dB, and the penetration allowed by this will depend upon the attenuation rate in the sediments. The primary mechanism of attenuation in sediments is friction between the particles leading to a constant loss per wavelength, and hence attenuation per unit distance directly proportional to frequency, at least in the range 1 to 10 kHz. The constant of proportionality will depend on the material, the packing pressure and to some extent the hydrostatic pressure. The consensus of measurements gives a value of 0.1 dB per wavelength leading to 0.15 dB/metre. This gives a sediment path of 203 metres for the two way journey, in other words a penetration of just over 100 metres, so that the figures conform fairly well with the design requirement.

The only other points of interest in this project concern the design and construction of the hydrophone and the realisation of the processing gain.

The Hydrophone

Clearly the suppression of endfire or near endfire lobes in the towed hydrophone is of paramount importance since the spectral level of isotropic sea noise even in a full gale fails by nearly 20 dB to match the directional noise level from the survey ship at over 8 knots. The technique used was, following Autrey (1972), to take a linear array of elements at uniform spacing and to sum the outputs through a weighting network to achieve the desired directional characteristic. The design parameter is:-

$$\psi = \pi f_{/f_D} \sin\theta$$

where f is the actual frequency, f_D the design frequency and θ the offset angle. This means that one can model the hydrophone in ψ and convert easily to real frequencies and angles. The design frequency corresponds to half wavelength spacing of the elements. ψ can be seen to be the progressive phase shift between elements and the array function $A(\psi)$ is

$$A(\psi) = \sum^{N} a_n \cos n\psi$$

$A(\psi)$ is clearly periodic in ψ with period 2π and is functionally identical to the Fourier Series expansion of a real even function. So if one wishes $A(\psi)$ to approximate some desired (real symmetric) array function $B(\psi)$ one can compute the required element weights by a Fourier Series analysis of $B(\psi)$ truncated at N terms. This approximation is optimum in the sense of having minimum mean squared error.

The design process is to model the desired function $B(\psi)$ at $f = f_D$ in the region $|\theta| < \pi/2$ corresponding to $|\psi| < \pi$, then

$$B(\psi) \approx A(\psi) = \sum^{N} a_n \cos n\psi$$

where $a_n = \frac{1}{\pi} \int_{-\pi}^{\pi} B(\psi) \cos n\psi d\psi$.

It is useful to remember when specifying $B(\psi)$, that if it has a first order discontinuity (step) then a_n falls off as $1/n$ and the truncation at finite n can lead to substantial errors. Similarly a second order discontinuity (abrupt change of slope) causes n to fall off as $1/n^2$. The ideal pattern for this application is one which is constant over a range of θ near zero then falls smoothly and rapidly to zero and remains as small as possible in the resultant cone about the axis. The pattern chosen for $B(\psi)$ was:-

$B(\psi) = 1$ for $0^\circ < |\theta| \leq 15^\circ$
$B(\psi) = (50 - \theta)/35 \quad 15^\circ < |\theta| < 50^\circ$
$B(\psi) \approx 0$ for $50^\circ < |\theta| < 90^\circ$

This pattern was approximated with an 11 element array with the following weights numbering the elements from -5 through 0 to +5.

$A(0) = 1$
$A(\pm1) = 0.581$
$A(\pm2) = -2.70 \times 10^{-2}$
$A(\pm3) = -0.120$
$A(\pm4) = +1.27 \times 10^{-2}$
$A(\pm5) = 2.29 \times 10^{-2}$

The $A(\pm5)$ values were changed later to 2.52×10^{-2} to smooth out a small hump in the pattern near $\theta = 90^\circ$.

The patterns of the uncorrected weights at the centre frequency and the band edges are shown in Fig. 1. Note the change of scale for the residual axial and near axial lobes.

The hydrophone construction is perfectly conventional, consisting of a semi-armoured quad cable towing a 5 cm polythene tube containing the eleven array elements. The weighting network is incorporated in the pre-amplifier at the head of the array, which is a charge amplifier (capacitive input elements and capacitive feedback around an operational amplifier). The charge inputs are scaled according to the weights by adjusting individual values and where necessary reversing the element polarity.

The Receiver

Apart from the realisation of the processing gain the receiver is perfectly straightforward, that is to say it consists of amplifiers, band-limiting filters, a detector and a display. The pulse, with a BT product of 25, is a linear frequency modulated sweep whose properties have been exhaustively covered in the literature (see for example Cook and Bernfeld, 1967). The pulse is clocked from a programmable read-only memory for stability. The processor is a dispersive analogue delay line based on a circuit originally suggested by Professor Berktay in a private communication. The design process follows O'Meara (1962) and consists essentially of concatenated groups of stagger-tuned two pole all pass networks.

The two pole all pass network has a transfer function as a function of complex frequency

$$T(s) = \frac{s^2 - 2\xi\omega_o s + \omega_o^2}{s^2 + 2\xi\omega_o s + \omega_o^2}$$

and the phase is

$$\beta(\omega) = 2\tan^{-1}\left(\frac{2\xi\omega\omega_o}{\omega_o^2 - \omega^2}\right)$$

The group delay T_D is

$$T_D = \frac{d\beta(\omega)}{d\omega} = 4\xi\omega_o\left(\frac{\omega^2 + \omega_o^2}{(\omega^2 - \omega_o^2) + 4\xi^2\omega_o^2\omega^2}\right)$$

This is a bell shaped curve peaking at $T_D(max) = \frac{2}{\xi\omega_o}$ when $\omega = \omega_o$. O'Meara shows how to choose the values of ω_o and ξ for each member of a group to realise any desired delay versus frequency curve. The present design uses six two pole networks in a group. The number of sections required cannot be less than $^{BT}/2$ where B is the bandwidth and T the dispersion. The advantage of using six or more sections in a group is that the total number of sections is not much more than $^{BT}/2$, in fact sextuplets have an efficiency of 55% compared with less than 20% for doublets (both for 1% delay error). In this example with $B = 10^3$ and $T = 28$ ms we find that 25 sections are needed if sextuplets are used or 78 for doublets. In the event four sextuplets were used with a slight reduction in B. In each sextuplet the centre frequencies and Q values were as follows:-

Section	f_o	Q
1	1605	25.4
2	1698	19.2
3	1811	17.5
4	1946	16.2
5	2109	14.3
6	2322	12.3

Though it is possible to interleave the critical frequencies in the separate sextuplets to reduce residual ripple the advantages are not very great, and the work is considerable.

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The two pole all pass network was chosen for its ease of adjustment and stability. It uses a single grounded inductor and a single differential input operational amplifier. Fig. 2a shows the circuit. Modelling the operational amplifier with a transfer function $A_o/1 + s\tau_o$ and the inductor losses as a parallel constant resistance r this circuit has the transfer function:-

$$T(s) = \frac{-\left(1 + \frac{sL}{R'}\left(1 - \frac{2R'}{R}\right) + s^2 Lc\right)}{\left(1 + \frac{sL}{R'} + s^2 Lc\right)\left(1 - \frac{2(1 + s\tau_o)}{A_o}\right)}$$

Provided one uses an amplifier for which in the band of interest $\frac{A_o}{2(1 + s\tau_o)} >> 1$ the effects of the amplifier can be ignored. This is equivalent to restricting the working range to well below half the unity gain frequency of the op-amp, a condition easily met for the 2.25 kHz frequency by a 741 op-amp. The effect of r is to displace both the poles and the zeros of T(s). Berktay's solution is to put a negative resistance across the inductor numerically equal to r. This makes R = R' and the poles and zeros are correctly placed. Because of the extra noise introduced by the amplifier needed to realise the negative resistance (approximately R/r time the equivalent noise input voltage of the amplifier) a different approach was used here as shown in Fig. 2b. The scaling of the resistors around the negative amplifier input terminal compensates the inband amplitude and unwanted phase distortion at the expense of a fixed attenuation, which is removed by the factor k of the pick-off potentiometer.

The transfer function of this network taking account of the amplifier response is:-

$$T(s) = \frac{\alpha}{k - \frac{(1 + s\tau_o)(1 + \alpha)}{A_o}} \times \frac{1 + sL/R'(1 - R'/R(1 + \alpha)) + s^2 Lc}{1 + sL/R' + s^2 Lc}$$

Once again the effect of the amplifier can be ignored and it is clear that the undistorted all pass response is achieved if

$$\alpha = 1 - \frac{2R}{r} \text{ and } k = \alpha$$

In this case the circuit Q is determined by the combined effects of R and r, which means that the originally calculated values of R have to allow for finite inductor Q.

The dynamic range of such a circuit is set by the noise level, and analysis shows that each amplifier contributes just its own noise voltage to the output. Each stage has a dynamic range of nearly 120 decibels and since the noises add incoherently the resultant output dynamic range is about 106 decibels. The input range needs to be reduced by processing gain (14 dB) to give just over 90 decibels of available input signal variation, which is ample for the purpose. In the case of the negative resistance compensator analysis shows that this figure is just less than 3 dB worse, so there is really little to choose for performance between the two. The main advantage of this realisation with the single grounded inductor is the independent adjustment of the real and imaginary parts of the natural frequency.

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Results

The first full scale trials in RRS Discovery were undertaken in December 1979 in highly adverse conditions when the ship was attempting to make up lost time on passage in a heavy seaway. Nonetheless results were encouraging and a second cruise in February 1980 saw the achievement of the design criteria. The sample record shows a small section of the Nares Abyssal Plain in the Western North Atlantic. The record was taken in nearly 6000 metres of water at 10.7 knots, under calm (Sea State 2-3) conditions. The doublet structure of the echoes due to surface reflection is clearly visible, but penetration of at least 100 metres of sediment (mostly red and grey clays) is visible.

Conclusions

No claims are made for originality in any particular parts of this paper but it is offered as an example of how careful attention to acoustic and electronic detail and the synthesis of diverse concepts into a single system can yield results which compare favourably with anything else available in the field.

Acknowledgements

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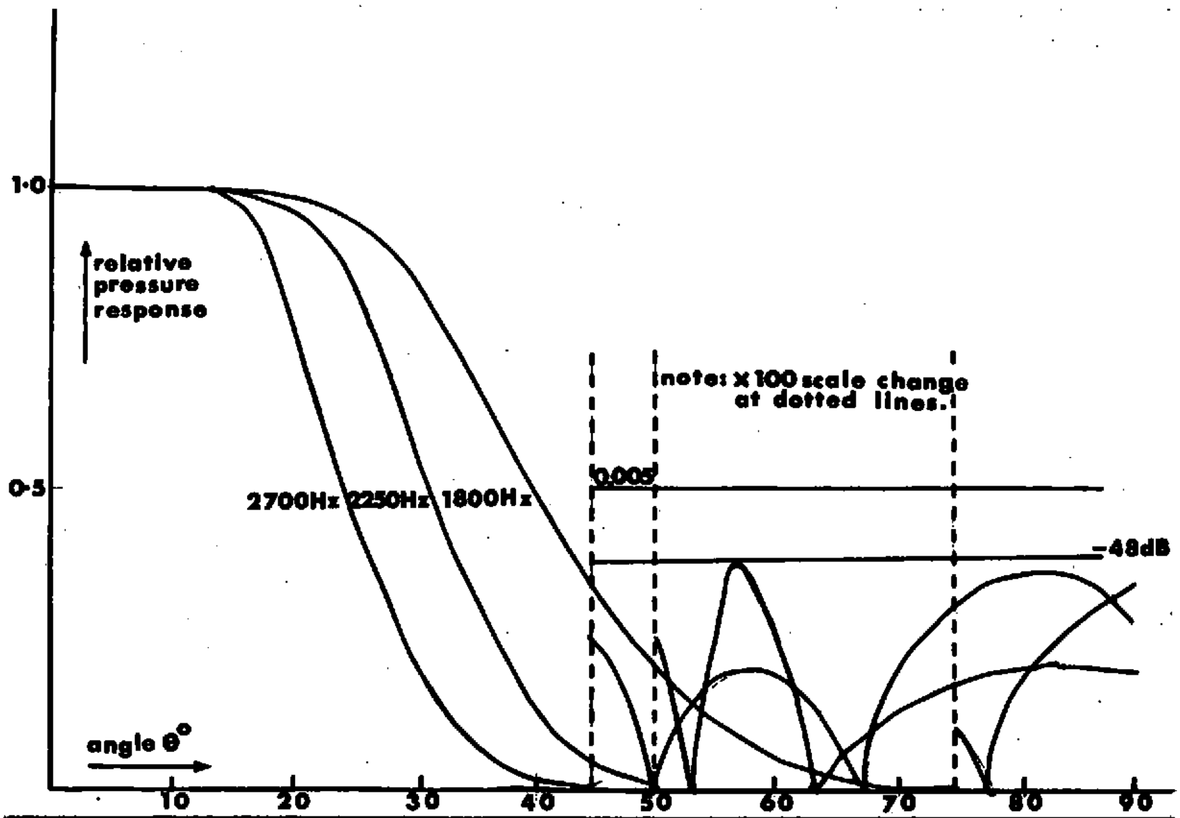

Fig. 1. Hydrophone Directivity Patterns.

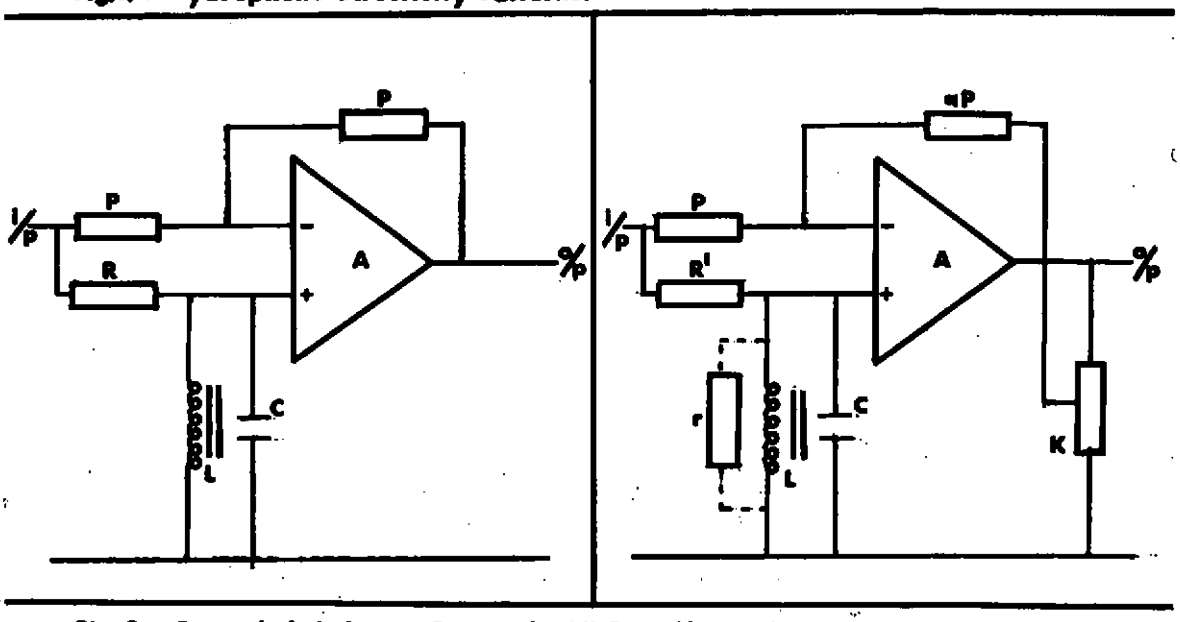

Fig. 2. Grounded Inductor Two Pole All Pass Network.
(a) basic network. (b) corrections for finite Q.

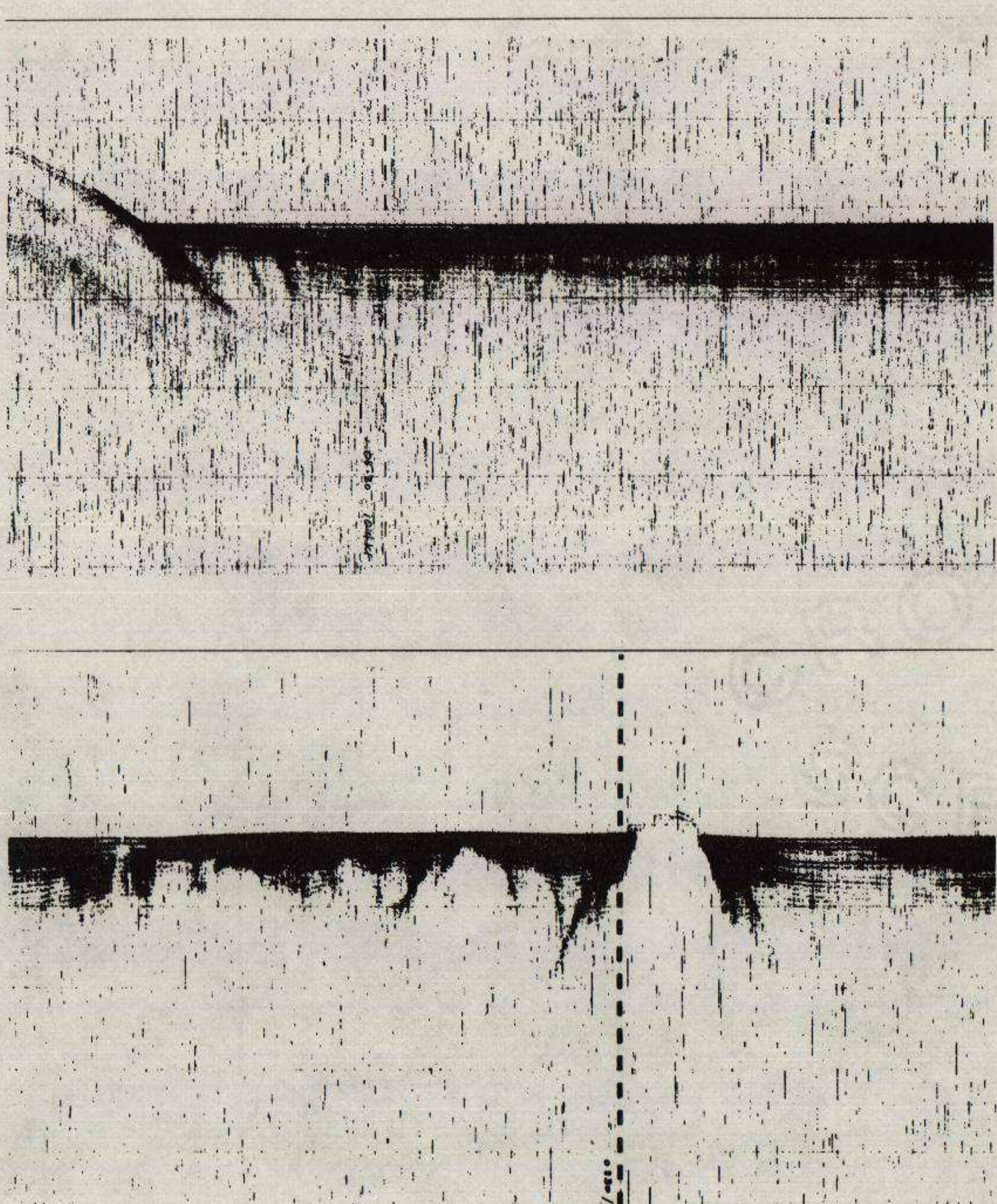

Fig. 3. Two sample records from the Nares Abyssal Plain showing the difference between 10 knots and 4. Both are at 5800 m depth.