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ADAPTIVE FILTERING AS A METHOD OF EQUALISING MICROPHONE RESPONSES

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1. INTRODUCTION

Techniques such as acoustic intensity measurement [1] and source location [2] often involve the use of multiple microphones. The signals from these microphones are combined so as to extract information which is not available from each microphone individually. Multiple microphone techniques such as those mentioned above normally require close matching of both amplitude and phase responses of the microphones over the frequency range of interest.

The present method of obtaining matched microphones is to measure the responses of a large batch during manufacture and pair those which have the closest responses. Once paired, similar microphones are relied upon to stay similar, despite operation under conditions of both temperature and humidity other than those present when the responses were measured. The relative microphone responses may also change due to differential ageing of the microphones or even minor mechanical abuse.

In practice not only the microphones but also any amplifiers and filters in their path must be matched. A simplification of this is to treat all the elements from microphone to filtered amplified voltage as a channel and to obtain matched channels rather than attempt to accurately match all the elements within individual channels. Considering channel error as one transfer function then allows the placing of an equalising filter in the path of one channel so that the resulting channel responses are more similar. Using this approach the task of obtaining matched channels is that of implementing a transfer function which is the difference in transfer functions between the channels. An approximation to this transfer function may be obtained by applying the Widrow-Hoff least mean squares adaption algorithm [3] to a digitally implemented finite impulse response filter. Suitably applying this algorithm will attempt to minimise the mean squared difference beween the filtered output of one channel and that of a second 'model' channel. The mean squared difference or error as it is normally referred to is a useful measure of both the phase and amplitude match of the channels as it requires both to be matched to produce a small mean square error. The FIR filter resulting from this adptive equalising may then be used as part of a channel to produce a matched pair of channels. The FIR filter may be adapted at any time so the adaptive equalisation technique does not need the long-term stability of response that matching by selection at manufacture requires. The lack of a requirement for long-term stability enables inexpensive electret microphones to be used as matched transducers when using adaptive equalisation to re-equalise these transducers as is necessary. A simple implementation of an adaptive equaliser along with its use for both instrumentation and commercial grade microphones is presented in this paper.

2. A BRIEF DESCRIPTION OF THE LMS ADAPTIVE FINITE IMPULSE RESPONSE FILTER

The finite impulse response filter has the property that its output sequence $y(n)$ is a linearly weighted summation of past and present input values, $x(n)$, but has no recursive component

$$y(n) = \sum_{I=0}^{I-1} w_i x(n - i) \qquad (1)$$

The output sequence is therefore linearly related to both the filter coefficients or weights w_i and the delayed input sequence $x(n - i)$. In the adaptive filter of Figure 1 it can be seen that the filter output $Y(n)$ is compared to some "desired" sequenced (n) by subtraction to produce a difference or error sequence $e(n)$.

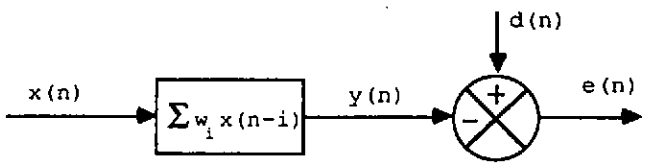


Figure 1

If we examine the mean squared error sequence as a function of the coefficients w_i it is clearly a quadratic function of w_i and so the set of filter coefficients which will produce a minimum mean square error will be identified by the gradients of the error with respect to the coefficients being zero. As there are no local minima in the gradient of $e(n)$, a gradient descent method such as the method of steepest descents will find the optimal set of coefficients w_i such that they cause the mean square error, $\overline{e^2(n)}$, to be minimised. The steepest descents algorithm causes each coefficient to be changed by subtracting a quantity proportional to the gradient of the error with respect to that coefficient. The gradient of the error may be evaluated thus

$$\frac{\overline{\partial e^2(n)}}{\partial w_i} = 2e(n)\,\frac{\overline{\partial e(n)}}{\partial w_i} = \overline{2e(n)x(n - i)} \qquad (2)$$

The averaging of gradient in equation (2) is computationally expensive and so in practice an instantaneous estimate of gradient is formed from the averaged error signal, $e(n)$, and the delayed input signal $x(n - i)$. The averaging of this gradient estimate takes place in the coefficient itself. We now have the Widrow-Hoff LMS adaption algorithm which updates each coefficient at every sample. The update of a coefficient w_i may be expressed as

$$w_i(n + 1) = w_i(n) + \alpha e(n)x(n - i). \qquad (3)$$

2.1 The LMS Adaptive Filter as a Signal Conditions Element

Although rigorous models for the behaviour of the LMS adaptive filter exist for certain special cases, a general model has yet to be found. Although no exact model exists, it may be found useful to view the filter of Figure 1 as an element which attempts to linearly filter the sequence $x(n)$ so as to make $y(n)$ match $d(n)$ as much as the limited filter length, 1, and the relationship between $x(n)$ and $d(n)$ will allow. It should be noted that by assuming linear filtering we are assuming that the filter coefficients are constant and unaffected by $x(n)$ or $d(n)$. This constraint is approximated if the time taken for coefficients to change appreciably during adaption is much longer than the period of the lowest frequency in both $x(n)$ and $d(n)$.

2.2 The LMS Adaptive Filter as Channel Equaliser

An adaptive filter may be used to equalise two channels driven by common input signals by arranging the adaptive filter such that in minimising its error sequence results in producing a filter which compensates for the difference in channel responses. A suitable arrangement is shown in Figure 2.

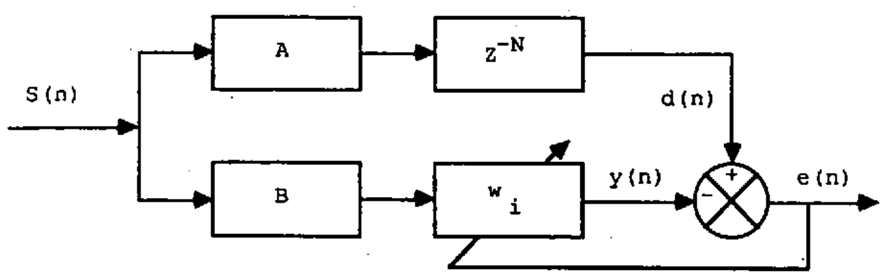


Figure 2. An arrangement of the LMS adaptive filter as an equaliser.

It should be noted that in order to minimise the mean squared error, the filter w must encompass a transfer function which is the inverse of channel B. Since the inverse of B need not be causal, a delay has been introduced into the A channel path so that the input to the filter, w, is advanced in time compared to the desired input $d(n)$. This results in w being able to produce a non-causal output relative to the desired input, $d(n)$. This delay technique is similar to the "modelling delay technique" discussed in [6].

The signal $s(n)$ should be chosen to contain frequency components over the entire range of interest since the effort placed on equalising at any one frequency will depend on the $s(n)$ amplitude at that frequency.

Once the filter w has adapted so as to minimise $\hat{e}^2(n)$ the adaption may be stopped and the equalising filter w used as shown in Figure 3.

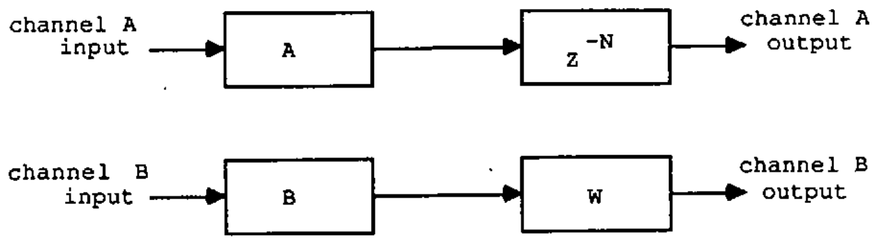


Figure 3. Use of the equalising filter.

The delay z^{-n} is still present because it was both A and this delay that the B and W transfer functions are made to match.

3. AN EXPERIMENTAL IMPLEMENTATION OF AN ADAPTIVE MICROPHONE EQUALISER

The digital portions of an adaptive equaliser were implemented using a multi-channel digital signal processor developed at the ISVR [6]. The block diagram of Figure 4 shows the arrangement of both electro-acoustic and digital elements of the equaliser.

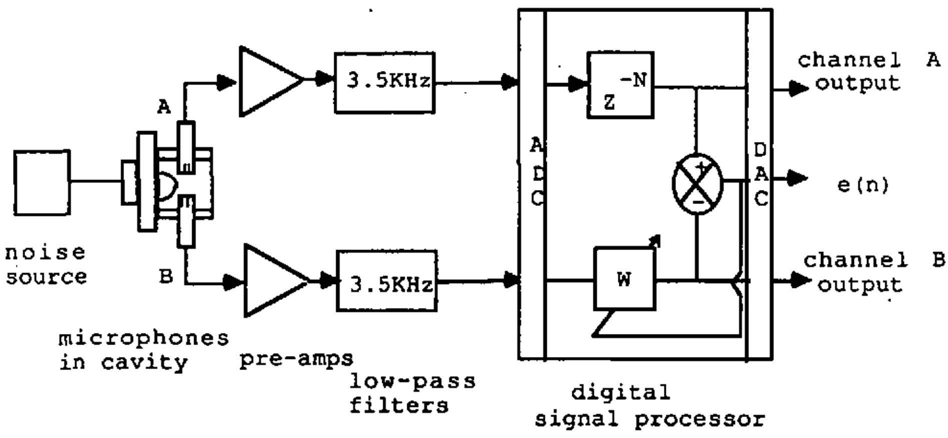


Figure 4

In an attempt to achieve equal acoustic pressure at both microphones they were placed in a small cylindrical cavity with an internal diameter of 40 mm and depth of 35 mm. The microphones were placed symmetrically near the centre of the cavity, which was driven by a 1" dome tweeter. Although the first longitudinal resonance of this cavity will occur at about 5 kHz and despite the presence of radial pressure variations [6], it was felt that the pressure at the adjacent microphones would be substantially equal up to a frequency of around 3 kHz. The sample rate of the digital signal processor was set to 8 kHz and low pass filter set at 3.5 kHz placed in the microphone signal paths to prevent aliasing.

The adaptive equaliser was programmed using the Texas Instruments TMS 32010 microprocessor present in the signal processing box. This microprocessor proved easily adequate to implement the 32 coefficient adaptive filter and 16 sample period delay at the 8 kHz sample rate used in this implementation.

Both the equalised channels and the residual error were made available as outputs from the processor box which enabled the performance of the equaliser to be estimated by comparison of the residual error after equalisation with the amplitude of one microphone channel. In order to assess how well the microphone channels could be equalised using only a simple gain adjustment a one coefficient LMS filter was also implemented. The one coefficient filter simply adjusts the gain of one channel so as to minimise the mean square difference between channels.

3.1 Optimising the Cavity Excitation

The adaptive equaliser attempts to minimise the total mean squared error and so the effort given to equalising the channels at any one frequency will depend on the amplitude at that frequency which is present in the cavity. It was desired that approximately equal effort was given to all frequencies up to 3 kHz and so the spectrum of the noise in the cavity should be white.

When the dome tweeter in the cavity was driven by a white noise source it was found that the dome tweeter and cavity had a resonant peak in its response of some 20 dB at 1 kHz as is illustrated in Figure 5.

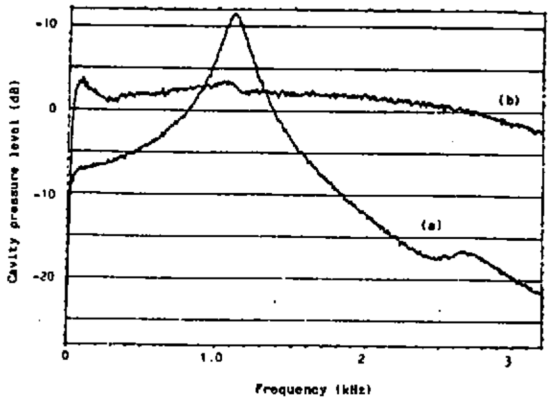
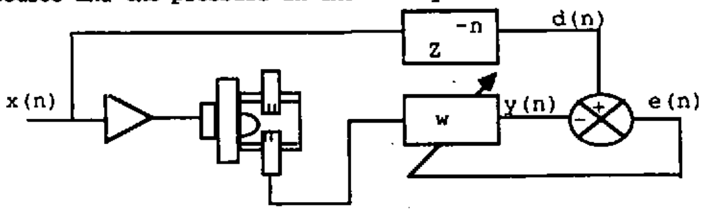
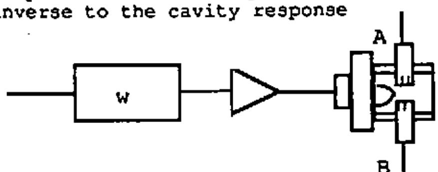


Figure 5

This response was flattened by using an adaptive filter to model the inverse of the response of the dome tweeter and cavity. The arrangement used, shown in Figure 6, attempts to minimise the difference between the white noise source and the pressure in the cavity.



adaptive filter arranged to find an approximate inverse to the cavity response



approximate inverse used to flatten cavity response

Figure 6

The results of this are given in Figure 4 and show a considerable improvement throughout the frequency range up to 3 kHz.

4. EXPERIMENTS

The first experiments were performed using two Brüel and Kjaer type 4134 instrumentation grade amplifiers followed by type 2619 and 2609 amplifiers. The spectrum of the channel, an output from the processor box, was then measured to give a level reference. The reference spectrum is shown in Figure 7. The 32 coefficient equaliser was then allowed to adapt and then, when there was no further change in the adaption, was stopped and the spectrum of the processor error output measured. This result, shown in Figure 7b, shows that the difference between the equalised microphone channels is around 40 dB below that of the reference channel.

The spectrum of the error signal produced by performing a simple gain matching of the channels using a 1 coefficient equaliser is shown in Figure 7c. The gain matching equalisation results in a residual error which is approximately 30 dB smaller than the reference spectrum and so the 32 coefficient equaliser has produced a residual error which is approximately 10 dB lower than achievable with a simple gain control.

The experiment was repeated with two B & K type 4133 microphones. The resulting spectra in Figure 8 show that although a lower residual error is produced by the 32 coefficient equaliser compared to simple gain equalisation, the improvement is not as marked as in the previous experiment. Comparing the gain equalised spectra for the 2 sets of microphones reveals that the 4133 were more closely matched before equalisation was applied and the residual errors of both sets of microphones drops to approximately the same level after use of the 32 coefficient equaliser.

A further set of experiments was performed using two pairs of electret microphones (Matsushita type WM-0637). These microphones were randomly chosen from a batch of 1000 microphones. The amplifiers used with these microphones consisted of a simple A.C. coupled operational amplifier circuit, which had a 3 dB bandwidth of approximately 20 Hz to 10 kHz. The results of these experiments can be seen in Figures 9 and 10. The result of using a 32 coefficient equaliser was to reduce the residual error to around 40 dB below the level of one microphone. The results of Figure 9 show an improvement of approximately 20 dB over the majority of the frequency range used.

The residual error after equalising the electret microphones can be seen to be similar in level to that of the equalised instrumentation microphones.

4.1 Some Limitations of the Equaliser Implementation

Some insight into the behaviour of the equaliser may be gained by examining the transfer function it implements in order to eualise a pair of microphones. As the filter in the equaliser is an FIR filter, the impulse response of the filter is given by its coefficients. The impulse response given in Figure 11 was obtained from filter coefficients of the first experiment (equalising 4134's). The time axis has been modified to take account of the 16 sample delay in the desired path of the equaliser. The impulse response consists of one large central peak and much smaller values elsewhere.

This is to be expected since the filter is attempting to model the difference between the two similar microphone responses. This form of impulse response is likely to introduce a problem of dynamic range in the coefficients of the filter. The coefficients on the TMS 32010 [7] are represented as 16 bit binary numbers so the presence of one coefficient of the order of a hundred times the magnitude of the others will restrict their effective representation to 8 or 9 bits.

This quantisation of coefficients combined with the limited length of the filter contribute to the ripples seen in the frequency response given in Figure 12. This response was obtained by Fourier transforming a zero padded version [4] of the impulse response of Figure 11.

There would appear to be some benefit to be gained from increasing the filter length and accuracy of coefficient representation in this application.

5. CONCLUSIONS

A method of equalising the response of one transducer to that of another has been presented. A 32 coefficient implementation of the equaliser has been shown to give significant improvements over simple gain adjustment when used to equalise microphones at frequencies of up to 3 kHz. In initial experiments the difference between two microphone responses was reduced to an extent such that error beween the microphone signals was found to be approximately 46 dB below the level of the signal from a single microphone. Further investigation into the form of the transfer function which is required to be implemented by the equaliser and the length and accuracy of filter required to adequately implement this form of transfer function is required.

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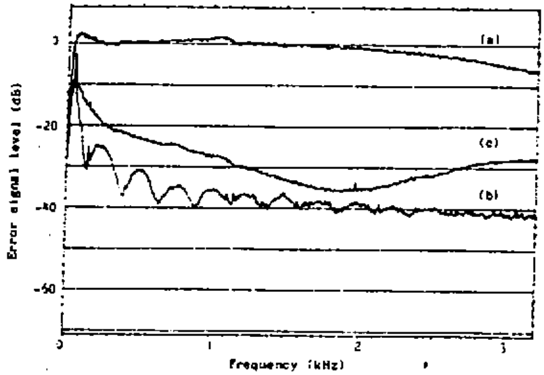

Figure 7. Results of equalising a pair of B & K type 4134 microphones.

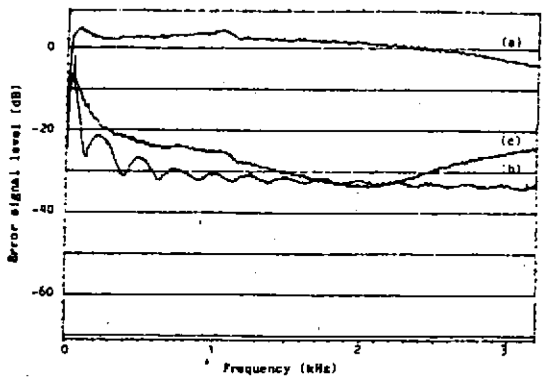

Figure 8. Results of equalising a pair of B & K type 4133 microphones.

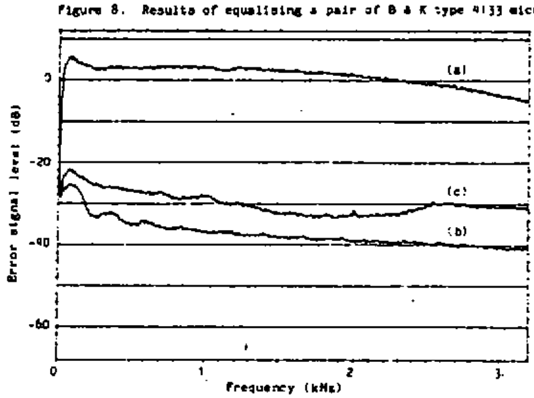

Figure 9. Results of equalising a pair of type WM-063T microphones.

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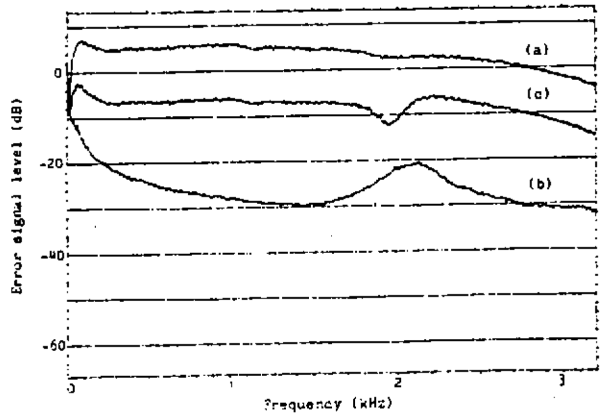

Figure 10. Results of equalising another pair of type WM-063T microphones.

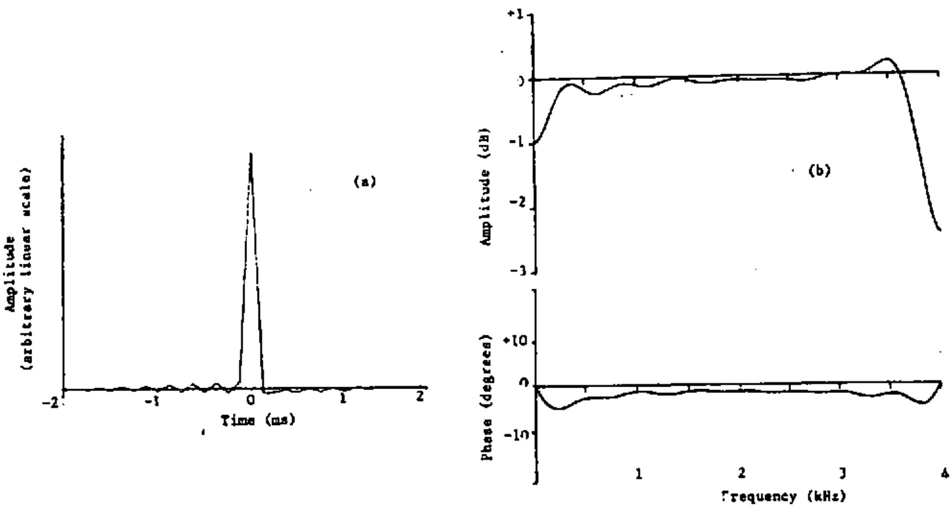

Figure 11. The impulse response (a) and frequency response (b) of the 32 coefficient filter which optimally equalises two type 4134 microphones.