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Adaptive Microphone Arrays for Noise Suppression in the Frequency Domain

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Abstract

Several methods for adaptive multichannel noise suppression of speech signals are compared. A modified weighting function is proposed to reduce linear distortions of the desired signal which may be introduced by the postprocessing system component. Experimental results using an array of four microphones placed at the corners of a PC monitor show a good agreement between theory and experiment.

Introduction

In this paper we discuss adaptive algorithms for multi-microphone noise reduction and echo cancellation in reverberant environments. The results may be applied to hands-free telephony systems and speech recognition applications under noisy conditions.

Conventional (i.e. non-adaptive) microphone arrays [1] require many sensors to obtain a high spatial selectivity in the direction of the desired speaker. Adaptive sensor arrays, as proposed by Frost [2], Griffiths-Jim [3] or Duvall [4] are able to steer the zeros of the space-transfer function into the direction of the noise signal. This approach can be very effective if the number of noise sources is smaller than the number of sensors. However, in reverberant rooms the number of unwanted noise sources is very high due to reflecting walls. Experiments under realistic conditions have shown that the performance of Frost´s algorithm (dashed line in fig. 3a) is not better than a non-adaptive beamformer that computes the average of the input signals (solid line in fig. 3a).

Allan [5] has developed a reverberation suppression system using the fact that the correlation of reverberation received at two-points is small compared with that of the direct source signal if the source is close to the array. Kaneda and Tohyama [6] have proposed a two sensor noise reduction system that is based on the same assumptions. Zelinski [7] has extended this method to a four sensor configuration and introduced an additional suppression factor to improve the noise reduction.

Their common pattern is the basic structure consisting of three steps. The delay differences between signal source and individual sensors are compensated in the first step (preprocessing). Thus the array beam is steered towards the desired signal source. In Allen´s system the compensation is performed in the frequency domain, whereas the preprocessing is implemented by Kaneda and Zelinski in the time domain with the aid of correlators. The estimation error of this correlator can be reduced by introducing a coherence detector [14]. In the second step (non-adaptive beamformer) the signals at the N sensors (2 used by Kaneda, 4 by Zelinski) are summed and weighted by 1/N. This step may also be implemented either in the time domain (Zelinski) or in the frequency domain (Allan, Kaneda). This already gives a noise reduction about 10log(N) dB. The remaining noise of the non-adaptive beamformer is further reduced in the third step (postprocessing) by a one dimensional noise suppression filter. The output of the non-adaptive beamformer is multiplied with the transfer function of the postprocessing filter to obtain the spectrum of the filtered signal. The final output signal is reconstructed by using inverse FFT and the overlap add (OLA) method [8].

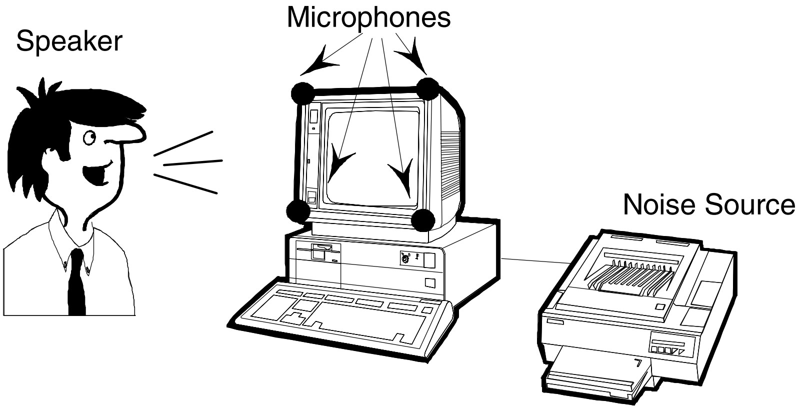

Fig. 1 Hardware configuration of the adaptive array

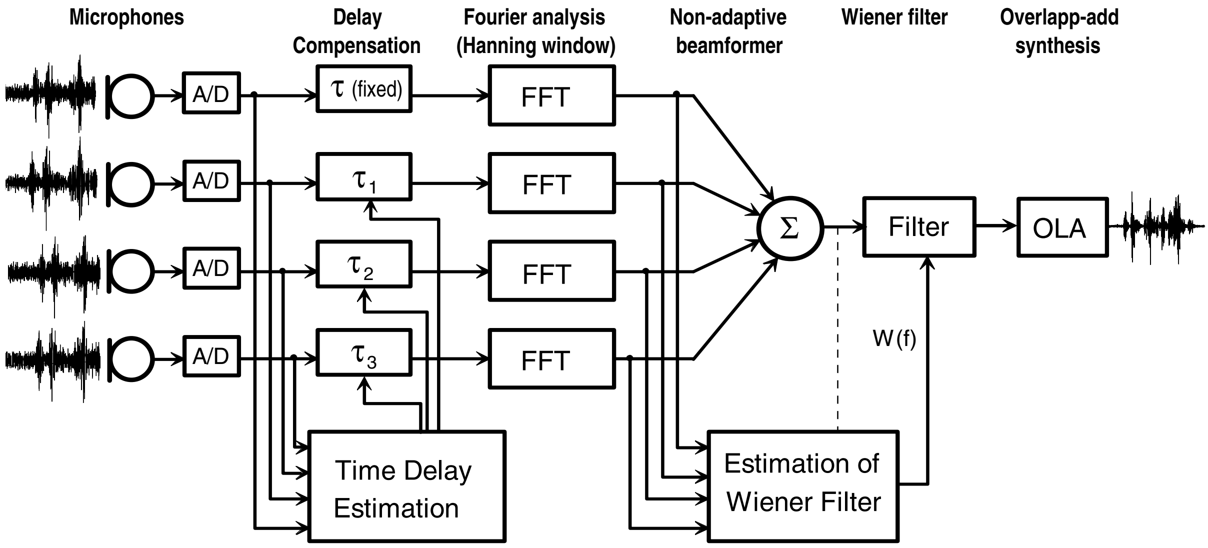

Fig. 2 Block diagram of the proposed system

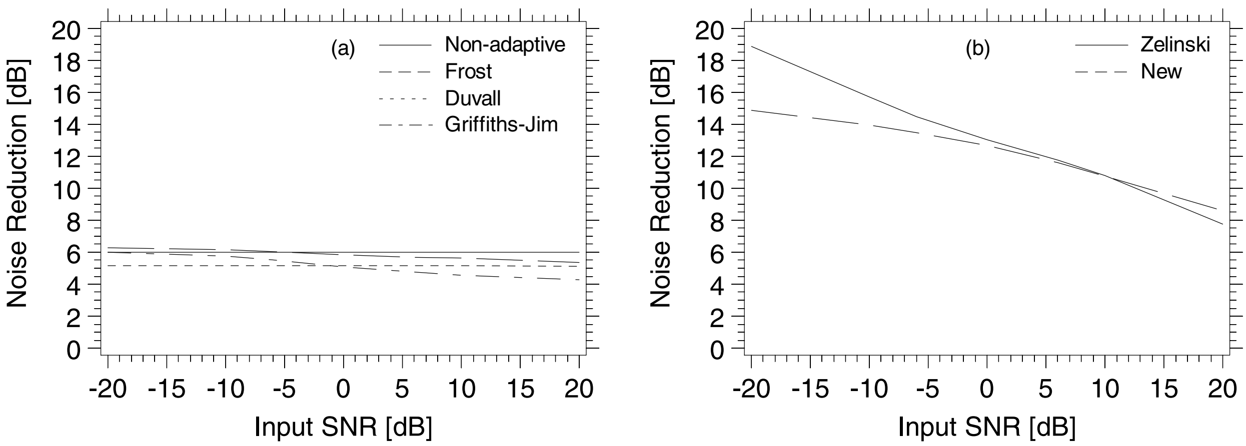

Fig. 3 Noise reduction of different algorithms as function of input SNR in an office room.

a) Noise cancellation algorithms: 4 sensor array 60x60cm, 33 taps, normalized stepsize μ=0.2

b) Noise suppression algorithms: 4 sensor array 60x60cm, window length 256, α=0.9 ([5] eq.15)

1. Placement of Sensors

The studied noise suppression systems are based on the assumption of a low correlation of the distorting signals at different sensors of the array. This supposition has been verified above a certain limiting frequency by measurements in reverberant rooms like usual office rooms [13]. Information about the spatial coherence (correlation) of the noise sources as function of frequency is required to determine the sensor distance of the microphone array. The reflecting walls of the room may be considered as acoustic sound sources which are almost statistically independent due to the large delays. Therefore a diffuse and isotropic sound field is found in reverberant rooms with strong reflecting walls, where the correlation is only a function of distance and does not depend on the local position. Such a sound field may be modeled by an infinite number of noise sources placed on the surface of a sphere with the microphones in the center ([10] p.16). Cron and Sherman [11] have derived on this condition the following relation for the cross-spectrum $\Phi_{ij}(f)$ between the sensor signals n_i n_j

$$\Phi_{ij}(d,\lambda) = \Phi_{nn} \frac{\sin(2\pi \frac{d}{\lambda})}{2\pi \frac{d}{\lambda}} = \Phi_{nn} \text{ sinc } (2\tfrac{d}{\lambda}) \tag{1}$$

d is the distance between the sensors, λ the wave length and $\Phi_{nn}(f)$ the power density spectrum of the noise signals. This model has been originally used to describe the volume noise in deep water sound propagation. Morrow has proved the validity of the same relation (1) for the diffuse sound field in a rectangular reverberation chamber [5].

Using the relation $\lambda = \frac{c}{f}$ between wavelength λ, frequency *f* and sound velocity c we may write:

$$\Phi_{ij}(d,f) = \Phi_{nn} \text{ sinc } (2\frac{d}{c} f) \,. \tag{2}$$

We get the complex frequency coherence function $\Gamma_{ij}(f)$

$$\Gamma_{ij}(f) = \frac{\Phi_{ij}(f)}{\sqrt{\Phi_{ii}(f)\,\Phi_{jj}(f)}} \,. \tag{3}$$

Since we have the same power density spectrum at all sensors $\Phi_{ii} = \Phi_{jj} = \Phi_{nn}$ in a diffuse, isotropic sound field the coherence of this sound field is given by

$$\Gamma_{ij}(f) = \text{sinc } (2\frac{d}{c} f) \,, \tag{4}$$

and the "magnitude-squared coherence function" (MSC) is computed as

$$\Gamma^2_{ij}(f) = \frac{|\Phi_{ij}(f)|^2}{\Phi_{ii}(f)\,\Phi_{jj}(f)} = \text{sinc}^2 (2\frac{d}{c} f) \,. \tag{5}$$

This equation gives the space frequency coherence of an isotropic sound field as function of the sensor distance d. The coherence function has its first zero at $f = \frac{c}{2d}$, the 3dB point $\Gamma_{ij} = 0.5$ is given at $f = \frac{c}{d}$.

Noise Cancellation algorithms yield only a noise reduction if there is a high coherence between the distorting signals. They fail if the coherence is low. The noise suppression systems discussed in this paper behave in an opposite way. The noise suppression works better with lower coherence of the distorting signals. Equation (4) and the figure 5 show a low coherence for an isotropic sound field only for frequencies

$f > \frac{c}{2d}$. This has to be considered for the positioning of the microphones. The distance of the sensors has to be determined by

$$d > \frac{c}{2 f_{\min}} \tag{6}$$

with f_{min} the lowest distortion frequency that has to be suppressed by the system.
The microphone distance however cannot be arbitrarily increased. With increasing microphone distance we eventually get an increasing distance between array and desired signal (speaker). Thus we get a decrease of its spatial coherence and hence a decrease of quality of the output signal [6]. An upper limit for the microphone distance is found from the necessity of "delay-compensation". The beam of the microphone array has to be steered into the direction of the speaker and has to follow his movements. The time delays between source and the particular microphones have to be estimated for this purpose. Due to the quasi-periodicity of the voiced portions of the speech signal (fig. 4b) a unambiguous estimation of the delay τ is only possible within the half pitch period $t_{pitch} = 1/f_{pitch}$

$$|\tau| < \frac{1}{2 f_{pitch}} \tag{7}$$

If one does not want to restrict the range of the speakers movements, one has to restrict the distance of the microphones. The maximum delay τ_{max} (with speaker 90° to plane of the microphones) is given as

$$\tau_{\max} = \pm \frac{d}{c_{sound}} \tag{8}$$

The maximum distance of microphones that allows an unambiguous location of the speaker is given therefore as

$$d < \frac{c_{sound}}{2 f_{pitch}} = \frac{\lambda_{pitch}}{2} \tag{9}$$

To fulfill the given conditions (6) and (9) of sensor distance, the lower cut-off frequency of the noise suppression system should not exceed the highest pitch frequency. The ideal sensor distance is therefore

$$d = \frac{c_{sound}}{2 f_{\min}} = \frac{c_{sound}}{2 f_{pitch_{\max}}} \tag{10}$$

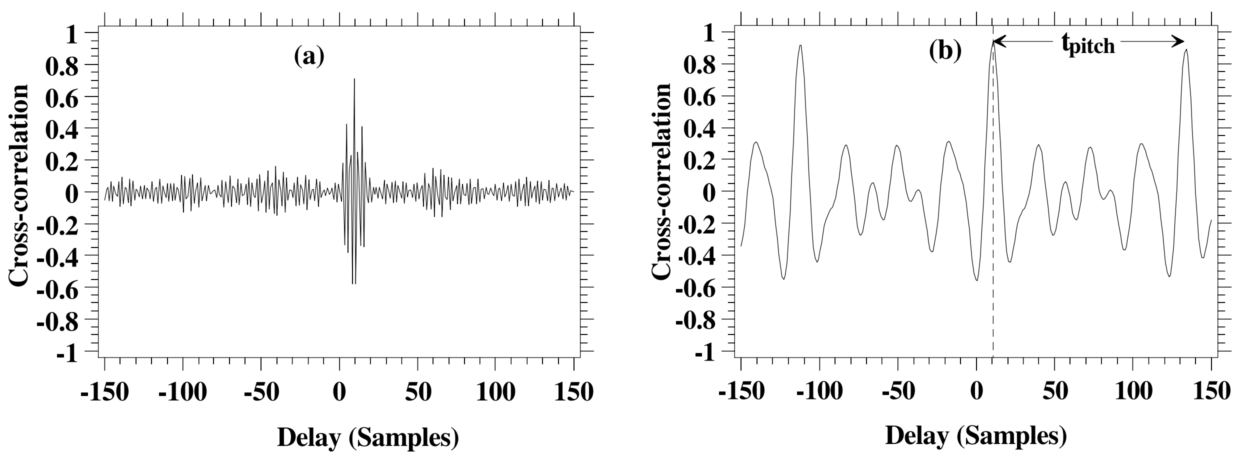

Fig. 4a Cross-correlation of unvoiced consonant [s] True delay 10 samples.

Fig. 4b Cross-correlation of voiced vowel [a] True delay 10 samples.

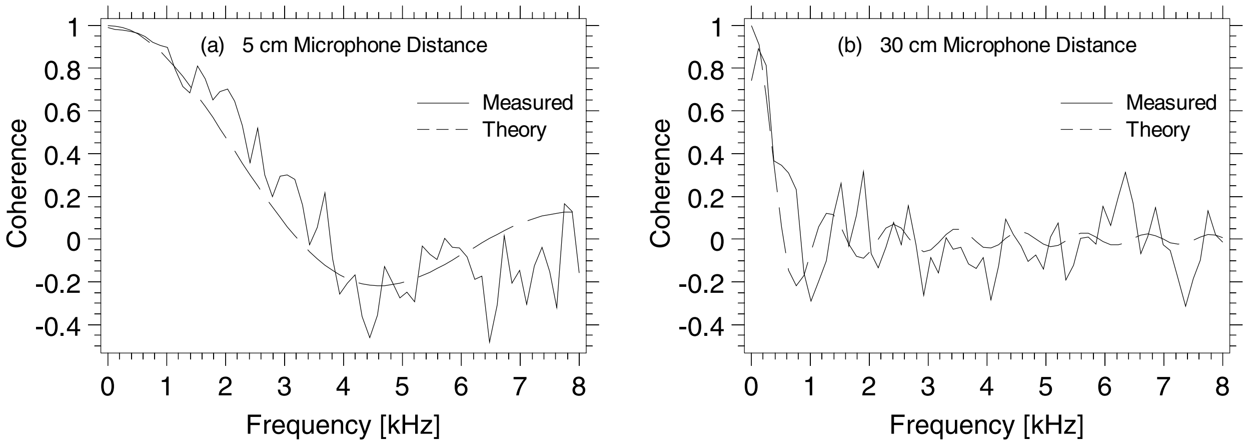

Fig. 5. Real part of the complex coherence function (eq. 3 and 4) measured in an office room.

2. Performance of the Non-adaptive Beamformer

To determine the efficiency and to develop the optimal filter function of the postprocessing filter (Wiener filter) we need to know the noise reduction of the non-adaptive beamformer as function of frequency. For that purpose we first determine the autocorrelation function $R_{\overline{n}\overline{n}}$ of the noise $\overline{n}$ at the output of the beamformer.

$$R_{\overline{n}\overline{n}}(\ell) = E\left[\frac{1}{N}\sum_{i=0}^{N-1} n_i[k] \cdot \frac{1}{N}\sum_{i=0}^{N-1} n_i[k+\ell]\right] \tag{11}$$

$$R_{\overline{n}\overline{n}} = \frac{1}{N^2}(R_{n_0 n_1} + R_{n_0 n_2} + \ldots + R_{n_0 n_N} + R_{n_1 n_0} + R_{n_1 n_1} + \ldots + R_{n_1 n_N}$$

$$+ R_{n_N n_0} + R_{n_N n_1} + \ldots + R_{n_N n_N})$$

$$= \frac{1}{N^2}\sum_{i=0}^{N-1}\sum_{j=0}^{N-1} R_{n_i n_j} \tag{12}$$

The power density spectrum $\Phi_{\overline{n}\overline{n}}$ at the output of the beamformer is given as

$$\Phi_{\overline{n}\overline{n}}(f) = \frac{1}{N^2}\sum_{i=0}^{N-1}\sum_{j=0}^{N-1}\Phi_{n_i n_j} \,. \tag{13}$$

If we assume identical autospectra of the particular noise signals, we can write, using the complex coherence function (equation 3)

$$\Phi_{\overline{n}\overline{n}}(f) = \frac{1}{N^2}\Phi_{nn}\sum_{i=0}^{N-1}\sum_{j=0}^{N-1}\Gamma_{n_i n_j} \,. \tag{14}$$

The coherence between N pairs of identical sensors is given as

$$\Gamma_{ii}(f) = 1$$

and for N^2-N pairs the following relation holds

$$\Gamma_{ij}(f) + \Gamma_{ji}(f) = \Gamma_{ij}(f) + \Gamma_{ij}^*(f) = 2\,\mathrm{Re}\{\Gamma_{ij}(f)\}\,.$$

Therefore we can write the power density spectrum at the output of the beamformer as

$$\Phi_{\bar{n}\bar{n}}(f) = \frac{1}{N^2}(N\Phi_{nn} + 2\Phi_{nn}\sum_{i=0}^{N-2}\sum_{j=i+1}^{N-1}\mathrm{Re}\{\Gamma_{n_i n_j}\})$$

$$\Phi_{\bar{n}\bar{n}}(f) = \frac{1}{N}\Phi_{nn} + \frac{2}{N^2}\Phi_{nn}\sum_{i=0}^{N-2}\sum_{j=i+1}^{N-1}\mathrm{Re}\{\Gamma_{n_i n_j}\} \qquad (15)$$

The noise reduction is given as

$$NR(f) = 10\cdot\log_{10}(\frac{\Phi_{nn}(f)}{\Phi_{\bar{n}\bar{n}}(f)}) = 10\cdot\log_{10}[N/(1+\frac{2}{N}\sum_{i=0}^{N-2}\sum_{j=i+1}^{N-1}\mathrm{Re}\{\Gamma_{n_i n_j}\})]$$

$$= 10\cdot\log_{10}(N) - 10\cdot\log_{10}(1+\frac{2}{N}\sum_{i=0}^{N-2}\sum_{j=i+1}^{N-1}\mathrm{Re}\{\Gamma_{n_i n_j}\}) \qquad (16)$$

Limiting cases: For very low frequencies $f << \frac{c}{2d}$ all N(N-1)/2 elements of the double sum converge against one ($\lim_{f\to 0}\mathrm{sinc}\,(2\frac{d}{c}f) = 1$), and there is no noise reduction. For large frequencies $f > \frac{c}{2d}$ there is a decrease of coherence, therefore we get approximately the spectrum of the averaged signal as

$$\Phi_{\bar{n}\bar{n}}(f) = \frac{1}{N}\Phi_{nn}\,. \qquad (17)$$

This corresponds to a noise reduction of $10\log_{10}(N)$ which is equivalent to 6 dB if we use four sensors. Fig. 6 shows that the noise reduction is close to this value above 500 Hz. If the coherence is positive then the noise reduction is a bit smaller and for negative coherence a bit larger than the average value.

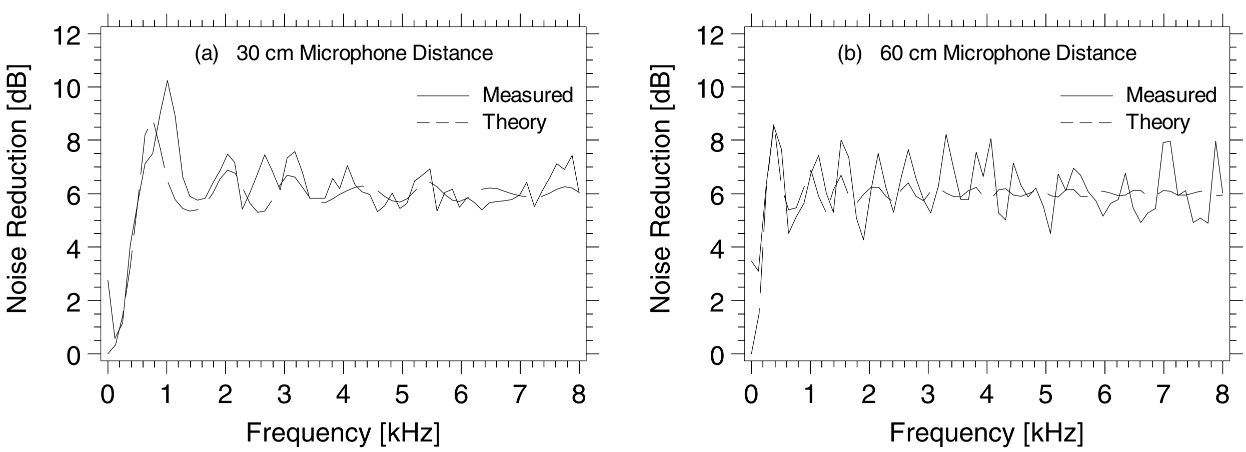

Fig 6 Noise reduction of the non-adaptive beamformer as a function of frequency (eq. 22)

In an isotropic sound field the coherence $\Gamma_{n_i n_j}$ depends only on the distance d_{ij} of sensor pairs. (eq. 4). The noise reduction in that case is given as

$$NR(f) = 10 \cdot \log_{10}(N) - 10 \cdot \log_{10}(1 + \frac{2}{N} \sum_{i=0}^{N-2} \sum_{j=i+1}^{N-1} \text{sinc}(2 \cdot f \cdot (\mathrm{d}_{ij}/c))) \qquad (18)$$

If we place four sensors at the corners of a d×d square, we get the lateral coherence between the four sensor pairs as

$$\Gamma_{i\,i+1}(f) = \text{sinc}\;(2\frac{d}{c}f) = \Gamma(d,f) . \qquad (19)$$

The coherence between the two diagonal pairs is given as

$$\Gamma_{i\,i+2}(d,f) = \text{sinc}\;(2\frac{\sqrt{2}d}{c}f) = \Gamma(\sqrt{2}d,f) \;. \qquad (20)$$

This yields the autospectrum of the average value of the sensor signals:

$$\Phi_{\overline{n}\overline{n}}(f) = \frac{1}{16}(4\Phi_{nn} + 8\Gamma(d,f)\Phi_{nn} + 4\Gamma(\sqrt{2}d,f)\Phi_{nn})$$

$$= \frac{1}{4}\Phi_{nn} + \frac{1}{2}\Gamma(d,f)\Phi_{nn} + \frac{1}{4}\Gamma(\sqrt{2}d,f)\Phi_{nn} . \qquad (21)$$

The noise reduction NR(f) of the non-adaptive beamformer of four sensors is given as

$$NR(f) = 10 \cdot \log_{10}(\frac{\Phi_{nn}(f)}{\Phi_{\overline{n}\overline{n}}(f)}) = 10 \cdot \log_{10}(4) - 10 \cdot \log_{10}(1 + 2\Gamma(d,f) + \Gamma(\sqrt{2}d,f)) \; (22)$$

3. Postprocessing

The remaining noise $\overline{n}$ of the averaged signal $\overline{x}$ may be further suppressed using a postprocessing filter. To derive such a Wiener filter we determine and minimize the power density spectrum Φ_{ee} of the error signal

$$\Phi_{ee}(f) = \Phi_{ss}(f) - \Phi_{s\hat{s}}(f) - \Phi_{\hat{s}s}(f) + \Phi_{\hat{s}\hat{s}}(f) \qquad (23)$$

The output power density spectrum of the filter is given by the Wiener-Lee formula

$$\Phi_{\hat{s}\hat{s}}(f) = \Phi_{\overline{x}\overline{x}}(f) | W(f) |^2 \qquad (24)$$

The cross-spectra between desired signal s and output signal of the filter $\hat{s}$ are given as

$$\Phi_{s\hat{s}}(f) = \Phi_{s\overline{x}}(f)W(f) \qquad (25)$$

and
$$\Phi_{\hat{s}s}(f) = \Phi_{\overline{x}s}(f)W^*(f) \qquad (26)$$

The power density spectrum of the error signal follows as

$$\Phi_{ee}(f) = \Phi_{ss}(f) - \Phi_{s\overline{x}}(f)W(f) - \Phi_{\overline{x}s}(f)W^*(f) + \Phi_{\overline{x}\overline{x}}(f) | W(f) |^2 \qquad (27)$$

if signal s and average noise $\overline{n}$ are uncorrelated we get

$$\Phi_{s\overline{x}} = \Phi_{\overline{x}s}(f) = \Phi_{ss}(f) \qquad (28)$$

and
$$\Phi_{\overline{x}\overline{x}}(f) = \Phi_{ss}(f) + \Phi_{\overline{n}\overline{n}}(f) \qquad (29)$$

The power density spectrum of the output error of the Wiener filter follows as

$$\Phi_{ee}(f) = \Phi_{ss}(f)\,(1 - W(f) - W^{*}(f) + |W(f)|^2) + \Phi_{\overline{nn}}(f)\,|W(f)|^2 \tag{30}$$

Using equations (15) we get the error power density of the combined system of beamformer and Wiener filter as

$$\Phi_{ee}(f) = \Phi_{ss}(f)\,(1 - 2\,\mathrm{Re}\{W(f)\} + |W(f)|^2) + \frac{1}{N}\,\Phi_{nn}\,(1 + \frac{2}{N}\sum_{i=0}^{N-2}\,\sum_{j=i+1}^{N-1}\mathrm{Re}\{\Gamma_{n_i n_j}\})|W(f)|^2 \tag{31}$$

The power density of the estimation error is minimized if the partial derivatives of real and imaginary part of W(f) are set to zero:

$$\frac{\partial\,\mathrm{Re}\{\Phi_{ee}(f)\}}{\partial\,\mathrm{Re}\{W(f)\}} = -2\Phi_{ss}(f) + 2\Phi_{ss}\,\mathrm{Re}\{W(f)\} + \frac{2}{N}\,\Phi_{nn}\,(1 + \frac{2}{N}\sum_{i=0}^{N-2}\,\sum_{j=i+1}^{N-1}\mathrm{Re}\{\Gamma_{n_i n_j}\})\,\mathrm{Re}\{W(f)\} = 0 \tag{32}$$

$$\frac{\partial\,\mathrm{Im}\{\Phi_{ee}(f)\}}{\partial\,\mathrm{Im}\{W(f)\}} = 2\,\Phi_{ss}(f)\,\mathrm{Im}\{W(f)\} + \frac{2}{N}\,\Phi_{nn}\,(1 + \frac{2}{N}\sum_{i=0}^{N-2}\,\sum_{j=i+1}^{N-1}\mathrm{Re}\{\Gamma_{n_i n_j}\})\,\mathrm{Im}\{W(f)\} = 0 \tag{33}$$

It follows that $W_{opt}(f)$ is real. This optimal weighting function in the sense of the least square error is given as

$$W_{opt}(f) = \frac{\Phi_{ss}(f)}{\Phi_{ss}(f) + \frac{1}{N}\,\Phi_{nn}\,(1 + \frac{2}{N}\sum_{i=0}^{N-2}\,\sum_{j=i+1}^{N-1}\mathrm{Re}\{\Gamma_{n_i n_j}\})} \tag{34}$$

Using (17) we get for frequencies $f > c/2d$ approximately

$$W_{opt}(f) \approx \frac{\Phi_{ss}(f)}{\Phi_{ss}(f) + \frac{1}{N}\,\Phi_{nn}(f)} \tag{35}$$

4. Estimation of the Weighting Function from Data

Zelinski´s estimation of the postfilter transfer function may be generalized to N sensors

$$W_z(f) = \frac{\frac{2}{N\cdot(N-1)}\sum_{i=0}^{N-2}\,\sum_{j=i+1}^{N-1}\mathrm{Re}\{\,X_i(f)\cdot X_j^{*}(f)\}}{\frac{1}{N}\sum_{i=0}^{N-1}\left|X_i(f)\right|^2} \tag{36}$$

were $X_i(f)$ is the Fourier transform of the input signal at sensor i. Assuming no correlation between signal and noise and the same power density of noise at all sensors we get

$$W_z(f) = \frac{\Phi_{ss} + \Phi_{nn}\,\frac{2}{N\cdot(N-1)}\sum_{i=0}^{N-2}\,\sum_{j=i+1}^{N-1}\mathrm{Re}\{\Gamma_{n_i n_j}\}}{\Phi_{ss} + \Phi_{nn}} \tag{37}$$

There is no noise reduction at low frequencies whereas for high frequencies $f > c/2d$ one has

$$W_z(f) \approx \frac{\Phi_{ss}}{\Phi_{ss} + \Phi_{nn}} \tag{38}$$

A comparison of the optimal solution (35) with Zelinski´s method shows an overestimation of factor N of noise power density given by his solution. As both signal and noise are weighted by W(f) we get a good noise suppression (solid line in fig. 3b) but on the other hand a linear distortion of the desired signal. This distortion increases with number N of sensors and with noise power Φ_{nn}. To avoid this effect we propose a modified version (dashed line in fig. 3b) of the postprocessing filter:

$$W_{sw}(f) = \frac{\frac{2}{N \cdot (N-1)} \sum_{i=0}^{N-2} \sum_{j=i+1}^{N-1} \mathrm{Re}\{ X_i(f) \cdot X_j^*(f) \}}{\left| \frac{1}{N} \sum_{i=0}^{N-1} X_i(f) \right|^2} \tag{39}$$

This alternative procedure estimates the power density spectrum of signal and noise by calculating the power spectrum of the sum of the output signals (dashed line in fig.2).Under the same assumptions as above we get

$$W_{sw}(f) = \frac{\Phi_{ss} + \Phi_{nn} \frac{2}{N \cdot (N-1)} \sum_{i=0}^{N-2} \sum_{j=i+1}^{N-1} \mathrm{Re}\{\Gamma_{n_i n_j}\}}{\Phi_{ss} + \frac{1}{N} \Phi_{nn} (1 + \frac{2}{N} \sum_{i=0}^{N-2} \sum_{j=i+1}^{N-1} \mathrm{Re}\{\Gamma_{n_i n_j}\})} \tag{40}$$

Although there is again no noise reduction for low frequencies, we get for $f > \frac{c}{2d}$ the approximation:

$$W_{sw}(f) \approx \frac{\Phi_{ss}(f)}{\Phi_{ss}(f) + \frac{1}{N} \Phi_{nn}(f)} \tag{41}$$

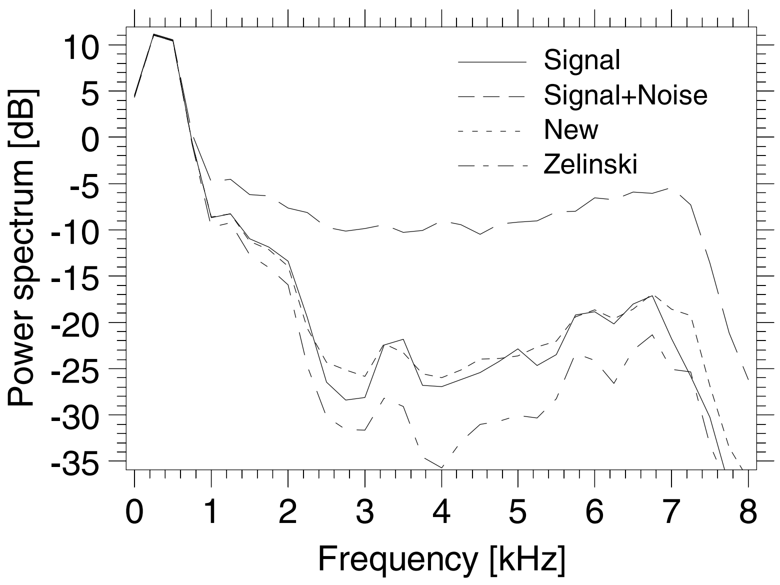

Fig. 7. Power spectra of processed signals as a function of frequency

5. Results

The figure shows the results of the compared methods. The solid line shows the power spectrum of the desired (original undistorted) signal. The dashed line gives the power spectrum of the distorted (original and noise) signal. The dashed dotted line shows the power spectrum processed by the method proposed in paper [7]. The dotted line proves the good agreement of the results of the procedure developed in this paper with the desired spectrum.

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