ACTIVE NOISE CONTROL
A New Active Sinusoidal Noise Control System

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This paper proposes a new method to reduce sinusoidal noise components whose frequencies are known. This new method bases the adjustment on the filtered-x algorithm, therefore requires correctly modeling a replica on the secondary path. However, the secondary path is continuously changing. The new method hence updates the replica at a specific interval by using the simultaneous equation technique that does not require feeding extra noise to the loudspeaker for the update. By slightly loosening the adjustment, the new method can compose the solvable simultaneous equations on the secondary path. The solutions give the gain and the phase rotation of the secondary path. This paper finally presents a simulation result to examine the performance of the new method.

INTRODUCTION

Active noise control system reduces primary noise by radiating secondary noise from a loudspeaker. In this system, the filtered-x algorithm is widely used as a means to adjust the amplitude and phase of the secondary noise. A problem is that the filtered-x algorithm requires correctly modeling a replica on the secondary path. In practical system, the secondary path is continuously changing. This paper proposes a new method to be able to reduce sinusoidal noise components under the changing. The new method updates the replica at a specific interval by using the simultaneous equation technique [1, 2] that does not require feeding extra noise to the loudspeaker for the update. This paper finally examines the performance of the new method with computer simulation.

SYSTEM CONFIGURATION

Figure 1 is a configuration of active sinusoidal noise control systems using the filtered-x LMS (least mean squares) algorithm. In this system, the primary sinusoidal noise $f(t)$ is reduced to $e(t)$, by the secondary sinusoidal noise $F(t)$ radiated from the loudspeaker. This paper here assumes that the primary noise is expressed as

$$ f(t) = \sum_{k=1}^{K} a_k \cos(\omega_k t - \phi_k) $$

$$ = \sum_{k=1}^{K} a_k \cos(\omega_k t) + \sum_{k=1}^{K} b_k \sin(\omega_k t), $$

where $T$ is the sampling period, and the radian frequency $\omega_k$ is known, while the magnitude $a_k$ and the phase $\phi_k$, i.e., $a_k = \alpha_k \cos \phi_k$ and $b_k = \alpha_k \sin \phi_k$, are unknown.

The filtered-x LMS algorithm updates the gains, $\hat{a}_k$ and $\hat{b}_k$, so as to minimize the error $e(t)$ by using the filtered reference signals,

$$ x_i(t) = (\gamma_{k} \cos \theta_{k} + \hat{\gamma}_{k} \sin \hat{\theta}_{k}) \cos(\omega_{k} t) $$

and

$$ x_i(t) = (-\hat{\gamma}_{k} \sin \hat{\theta}_{k} + \gamma_{k} \cos \theta_{k}) \sin(\omega_{k} t), $$

where $\gamma_{k}$ and $\theta_{k}$ are the previously estimated gain and phase rotation of the secondary path, respectively. A problem is that they are continuously changing.

SECONDARY PATH ESTIMATION

Figure 2 is an ordinary configuration for estimating the characteristics of the secondary path. A problem in this configuration is that the estimated gains, $c_k$ and $d_k$, provide only two equations having four unknowns, $a_k$, $b_k$, $\gamma_k$ and $\theta_k$,

$$ c_k = a_k + \hat{a}_k \gamma_k \cos \theta_k + b_k \gamma_k \sin \theta_k $$

$$ d_k = b_k - \hat{a}_k \gamma_k \sin \theta_k + \hat{b}_k \gamma_k \cos \theta_k. $$

From the above two equations, the characteristics, $\gamma_k \cos \theta_k$ and $\gamma_k \sin \theta_k$, cannot be derived.
The simultaneous equations technique [1, 2] states to be able to estimate the characteristics by using the adjustment error involved in the gains, $\hat{a}_1$ and $\hat{b}_1$, or adding small constants to the gains. Four different gains, $\hat{a}_1$, $\hat{a}_2$, $\hat{b}_1$, and $\hat{b}_2$, obtained by the addition, give the following equations:

\[
\begin{align*}
    c_1' &= a_1 + \hat{a}_1^2 \gamma_1 \cos \theta_1 + \hat{b}_1^2 \gamma_1 \sin \theta_1 \\
    d_1' &= b_1 - \hat{a}_1^2 \gamma_1 \sin \theta_1 + \hat{b}_1^2 \gamma_1 \cos \theta_1 \\
    c_2' &= a_1 + \hat{a}_2^2 \gamma_2 \cos \theta_1 + \hat{b}_2^2 \gamma_2 \sin \theta_1 \\
    d_2' &= b_1 - \hat{a}_2^2 \gamma_2 \sin \theta_1 + \hat{b}_2^2 \gamma_2 \cos \theta_1.
\end{align*}
\]

From the above equations, we have the solutions:

\[
\begin{align*}
    \gamma_1 \cos \theta_1 &= \frac{(c_1' - c_1^2)(\hat{a}_1^2 - \hat{a}_1^2) + (d_1' - d_1^2)(\hat{b}_1^2 - \hat{b}_1^2)}{(\hat{a}_1^2 - \hat{a}_1^2)^2 + (\hat{b}_1^2 - \hat{b}_1^2)^2} \\
    \gamma_1 \sin \theta_1 &= \frac{(c_1' - c_1^2)(\hat{b}_1^2 - \hat{b}_1^2) - (d_1' - d_1^2)(\hat{a}_1^2 - \hat{a}_1^2)}{(\hat{a}_1^2 - \hat{a}_1^2)^2 + (\hat{b}_1^2 - \hat{b}_1^2)^2},
\end{align*}
\]

where the solutions require the relation:

\[
(\hat{a}_1^2 - \hat{a}_1^2)^2 + (\hat{b}_1^2 - \hat{b}_1^2)^2 \neq 0
\]

**SIMULATION RESULT**

Figure 3 is an example to examine the performance of the new method using the simultaneous equations technique to estimate the secondary path. In this example, the primary noise consists of fifteen harmonics and fundamental of 200 Hz, and white noise is added to the primary noise as disturbance where their power ratio is –30 dB. In addition, the impulse response of the secondary path is provided as the sequence consisting of 32 normal random numbers.

Under the above conditions, the noise reduction effect is calculated as

\[
D_n = 10 \log_{10} \left( \frac{\sum_{j=a+1}^{(a+1)J} e^2(j)}{\sum_{j=a+1}^{(a+1)J} f^2(j)} \right)
\]

where $J = 320$. In addition, the adaptive algorithm to estimate the characteristics ($c_1$, $d_1$) and to update the gains ($\hat{a}_1$, $\hat{b}_1$) is both the block Implementation [3] of the filtered-x LMS algorithm whose block length and step size are 320 samples and 2.0/KJ, respectively.

In Figure 3, (a) is a result obtained by updating $\hat{a}_1$ and $\hat{b}_1$ after the characteristics has been estimated using the fixed gains $\hat{a}_1 = 0.6$, $\hat{b}_1 = -0.6$, $\hat{a}_2 = 0.59$, and $\hat{b}_2 = -0.59$. In the initial stage prior to starting the active noise control, we can fix the gains to any constants like this. Moreover, (b) is a noise reduction effect calculated by using the characteristics of the secondary path estimated with the gains, $\hat{a}_1 = \hat{a}_1^0 + 0.01 \hat{a}_1^0$, $\hat{b}_1 = \hat{b}_1^0 + 0.01 \hat{b}_1^0$, $\hat{a}_2 = \hat{a}_2^0 - 0.01 \hat{a}_2^0$, and $\hat{b}_2 = \hat{b}_2^0 - 0.01 \hat{b}_2^0$. which are fixed adding small constants to $\hat{a}_1^0$ and $\hat{b}_1^0$ obtained after the convergence shown in (a).

**REFERENCES**

Discrete-Time Parametric Modeling of Loudspeakers Used for Active Noise Control in Long Narrow Ducts

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The secondary path in active noise control systems is generally consisted of electromechanical (loudspeakers) and acoustic elements. The secondary path is substantial part of any active noise control system. For nonadaptive, it influences on its convergence properties. For nonadaptive, it makes an additional term in control chain that can cause instability in closed-loop control. If the duct is long enough, the system will cancel the propagating acoustic wave before it reaches the end. This fact allows us to use infinite acoustic waveguide model for the duct, and to estimate loudspeaker acoustic load in such a way. The loudspeaker itself is modeled using electro-mechanical analogies, and obtained continuous-time model is transferred in z-domain using bilinear transformation. Proposed method can be made to include finite-length waveguides.

INTRODUCTION

Active noise control (ANC) uses principle of destructive interference of the sound waves, e.g. in order to cancel undesired noise a sound wave with inverse sound pressure is generated [1]. The low frequency plane wave sound propagation in ducts makes of them the most promising area of application for ANC. In [2] the theoretical formulas for the sound field in a hard-walled duct with non-reflecting terminations have been derived. In [3] a theoretical, frequency domain model of the loudspeaker in a duct is introduced. In this paper a discrete-time model of dynamic loudspeakers in long narrow ducts will be developed.

FREQUENCY-DOMAIN MODELING

The relationship between terminal voltage and loudspeaker cone movement can be derived from [3,4]:

\[ H(s) = \frac{v(s)}{E(s)} = \frac{\rho L m}{B_l} s^3 + \frac{(L D + R n)}{B_l} s^2 + \left(\frac{L K + D R}{B_l} + B_l\right) s + \frac{K R}{B_l} \]  

where \( m \) denotes total effective mass of moving system (i.e. cone, voice coil and effective air mass), \( D \) denotes damping, including mechanical damping in moving system and radiation resistance, and \( K \) denotes effective stiffness of suspension and back enclosure, if any. Also \( L \) is inductance of the voice coil, \( I \) denotes wire length of voice coil, \( B \) is flux density in region of voice coil, and \( v \) is cone velocity amplitude. \( R \) is the resistance of the voice coil, and \( s \) denotes complex frequency.

Although perfectly anechoic termination of the duct is not realizable, for the reasonably effective attenuator the acoustic load contributed from the reflections from the duct termination can be ignored [3]. In this case, acoustic load over the face of the loudspeaker can be determined from the calculated sound pressure over the square piston being flush mounted on the wall \( x=0 \) wall in the rigid infinite duct of rectangular cross section \( 0<\alpha<\alpha, 0<\beta<\beta \) [2]:

\[ p(x, y, z, t) = \sum_{m} \rho c_{mn} v e^{i m(x - \delta_{in})(2 - \delta_{in})(2 - \delta_{in})} \cos \frac{m \pi x}{a} \times \left[ \cos \frac{n \pi y}{b} \cos \frac{n \pi y}{b} \right] \frac{1}{A} e^{-j \omega \delta_{mn} / c_{mn}} \frac{1}{\omega / c_{mn}} \times \left\{ \frac{e^{-j \omega \delta_{mn} / c_{mn}} - e^{-j \omega \delta_{mn} / c_{mn}}}{2 j} + \frac{1}{2} \frac{e^{-j \omega \delta_{mn} / c_{mn}} - e^{-j \omega \delta_{mn} / c_{mn}}}{2 j} \right\} \]  

where the piston is square in shape, of edge length \( d \). Hence, area of the piston is \( A = d^2 \). Its centre is at \((0,0,0)\) and \((x,y,z)\) are position coordinates (\( z \) is axial one) while \( a,b \) are duct cross-section dimensions. Also, \( \delta_{ij} \) is Kronecker delta function, \( j = \sqrt{-1} \), \( \rho \) is density of air, \( c_{mn} \) is mode axial phase speed and \( \Theta \) is the radian frequency. \( A \) is cross-sectional area of the duct.

Above the natural frequency of the mechanical system of the loudspeaker, the cone movement is controlled by its mass. Although acoustic load on the loudspeaker is consisted both from the propagating and evanescent modes, for low frequency range the imaginary part tends to be small compared to the mass of the loudspeaker cone, and real part tends to be well approximated with

\[ R_{M} = \frac{\rho c_{mn}^2}{2 A} \]
where $c$ is ambient speed of sound. As the matter of fact, the acoustic load influences mostly on the cone velocity in the vicinity of the natural frequency of mechanical system, and for other frequencies very different approximations give similar results.

The propagating pressure contribution for plane waves is [2,3]:

$$p(j\omega) = \frac{1}{2} \rho_c v(j\omega) \frac{A_p}{A} \sin(\omega d/2c)$$

(4)

The factor $\sin(\omega d/2c)/(\omega d/2c)$ can be neglected when the loudspeaker dimensions are small compared to the wavelength of sound.

**DISCRETE-TIME MODELING**

Applying the bilinear transformation [5] a $z$-domain transfer function can be derived:

$$H(z) = \frac{v(z)}{E(z)} = \frac{A_0 + A_1 z^{-1} + A_2 z^{-2} + A_3 z^{-3}}{1 + B_1 z^{-1} + B_2 z^{-2} + B_3 z^{-3}}$$

(5)

$$M = \frac{8}{Bl} \frac{Lm}{Bl} + 4T \frac{LD + Rm}{Bl} + 2T^2 \left( \frac{LK + DR}{Bl} + Bl \right) + T^3 \frac{KR}{Bl}$$

$$A_0 = A_1 = \frac{2T^2}{M}$$

$$A_2 = A_3 = -\frac{2T^2}{M}$$

$$B_1 = -\frac{24}{M} \frac{Lm}{Bl} + 4T \frac{LD + Rm}{Bl} + 2T^2 \left( \frac{LK + DR}{Bl} + Bl \right) + 3T^3 \frac{KR}{Bl}$$

$$B_2 = 24 \frac{Lm}{M} + 4T \frac{LD + Rm}{Bl} - 2T^2 \left( \frac{LK + DR}{Bl} + Bl \right) + 3T^3 \frac{KR}{Bl}$$

$$B_3 = -8 \frac{Lm}{M} + 4T \frac{LD + Rm}{Bl} - 2T^2 \left( \frac{LK + DR}{Bl} + Bl \right) + T^3 \frac{KR}{Bl}$$

where $T$ is sampling period.

Although it might seem to be advantageous to include acoustic load on the loudspeaker in a more accurate way, the differences are very small (below 1db), and , more over, approximation of the loudspeaker with the square piston is not accurate itself.

Since perfectly anechoic termination of the duct is not realizable, the exact experimental verification of the model is not possible. Beside numerical examinations, we made some experiments with a rigid finite 70 cm long duct with flush mounted broadband loudspeaker. The both ends of the duct were opened, and acoustic load on the loudspeaker and sound pressures in the duct were calculated using transmission lines theory [6] and loudspeaker and duct cross-sectional area ratio, similar as it is used for the reflex enclosure loudspeakers [4].

The radiation impedance on the ends was estimated as for pulsating sphere[6]. Figure 1. presents both calculated and measured response. The differences can be explained with losses in the system and directivity of the microphone.

**REFERENCES**


A Study of Active Noise Control in Ducts using Digital Signal Processing

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The present work describes some of the research effort on Active Noise Control (ANC) being jointly developed by the Catholic University of Minas Gerais (PUC-MINAS) and the Federal University of Minas Gerais (UFMG). Considerations about the implementation of Digital Signal Processing for noise control in ducts has been presented. The objective is to establish a study on Active Noise Control in ducts combining geometry and acoustic parameters modification together with adaptive digital filtering implementation. The main results are presented and considered according to their use in developing real applications. The idea is to provide an initial and useful insight for both designers and students concerned about Active Noise Control in ducts. The authors are continuing with this research, for other configurations and implementations and believe that the present studies should hopefully provide more favourable conditions to the implementation of real time ANC.

INTRODUCTION

In recent years Active Noise Control (ANC) has become a feasible reality for a variety of practical applications. The development of reasonably priced powerful processors and efficient software has enabled implementation of real time ANC [1].

ANC in ducts has deservedly earned a place of importance for a variety of reasons. Sound carrying ducts occur in a great number of important applications and the existing boundaries (the duct walls) may lead to simplifications in the wave propagation studies. However ANC in ducts still presents some difficulties, particularly when wide band noise is being analysed. In contrast to narrow band noise, wide band noise may occur in certain applications and is considerably more difficult to cope with, demanding efficient signal processing.

EXPERIMENT OVERVIEW

Testing was carried out in a set up represented in Figure 1, where one of the geometrical configurations used can be seen.

![Test set up](image)

FIGURE 1: Test set up

The primary source, a loudspeaker, was positioned at the left duct inlet, as shown and the control source, another loudspeaker at the inlet of the T joint. A microphone was positioned near the end of the right extremity of the duct, near the acoustic ending, which was varied during the test. Figure 1 shows a closed end termination. The microphones and the loudspeakers were connected to a Texas Instruments DSP-TMS 32C6211 board connected to a PC, providing the required signal processing.

CONTROL PROCEDURE

Adaptive filtering as initially proposed in the classical work of Wiener [2] has been the chosen solution. Figure 2 indicates a block diagram containing the main elements of the control system.

![Control System Diagram](image)

FIGURE 2: Control System Diagram

Noise at the input has been generated by means of a white noise generator. The white noise output was filtered before being input into the system, enabling appropriate choice of the input signal frequency range, which was kept within the lower frequency spectrum, for well known reasons [2].
Input signals for the adaptive FXLMS (Filtered-X Least Mean Square) algorithm were obtained from the data acquisition procedure. A FIR (Finite Impulse Response) digital filter was employed with the upper cut-off frequency set at 600 Hz. A single microphone was employed in the process, to provide the error signal. Plant identification compared with the (also) known noise input signal enabled the generation of the secondary source (control) signal, which was then input at the second loudspeaker.

The adaptive LMS algorithm may be conveniently described by equation (1), where $\mu$ was set to 0.014, which was found to be appropriate for this experiment [3],[4].

$$w(m+1) = w(m) + \mu (y(m)e(m)) \quad (1)$$

Implementation of the LMS algorithm has been carried out, according to the following sequence:
- Initial parameter choice
- Adaptive filter output evaluation
- Error signal evaluation
- Adaptive algorithm evaluates weighing
- System identification
- Noise cancelling signal generation

Equation 2 has been used to establish adaptive filter output.

$$y(n) = \sum_{k=0}^{L-1} w_k x(n-k) \quad (2)$$

**RESULTS AND CONCLUSIONS**

Using the previously described procedure the system was conveniently described by Figure 3, which was previously known, at the beginning of the test.

The duct characteristics have also been evaluated, producing results which can be seen in Figure 5.

**FIGURE 5: Duct Identification**

Once the plant had been characterized it was possible to generate the noise control signal (Figure 6), which was then input into the noise control source.

**FIGURE 6: Noise Control Signal**

The previously shown Figures indicate a typical set of results obtained from the test set up of Figure 1. Other tests have also been carried out, with a measure of success, for other geometric and signal conditions.

The previously described procedure is now being verified and refined in order to improve noise cancellation response, including the determination of optimal sensor and secondary source positioning for a variety of geometric configurations.

**ACKNOWLEDGEMENTS**

The authors would like to thank PUC-MG for the support given to the present research. They are also indebted to Rodrigo Costa Ivo, and Alexandre Augusto Moraes Nogueira for their vital participation in the development of the present research.

**REFERENCES**

Active Control of Double-glazing Windows: an Experimental Comparison between Feedback and Feedforward Controllers

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Double-glazing windows are distinguished by high acoustic transmission loss for the high- and mid-frequency range but weak in the low-frequency range especially around the mass-spring-mass resonance frequency of the panel-cavity-panel system. In the work presented the cavity sound field of a double-glazing window is actively minimized by means of secondary loudspeakers and error microphones inside the cavity. The influence of the minimization on the transmission loss of the double-glazing window was investigated experimentally. In this presentation will be reported the experiments performed with multichannel feedback control as well as those with multichannel feedforward control. A comparison will be given. In the case of the feedforward controllers the reference signal was obtained either by means of a microphone in the sending room of the window testing facility or directly from the signal generator. This leads to difficulties due to the properties of the sending room which will be discussed.

INTRODUCTION

In recent years some authors have investigated the active control of double-glazing windows by means of loudspeakers inside the cavity between the two panels (cf. [1, 2, 3, 4]). In this paper a comparison will be given between the use of two adaptive feedforward controllers and one adaptive feedback controller. The difference between the two feedforward controllers is the generation of the reference signal, which was obtained either directly from the signal generator as in [1, 3] or from a microphone in the sending room of the window testing facility. In all three cases a controller with 4 loudspeakers and 4 error microphones was used. The positions of the loudspeakers and microphones were near the corners of the window. All used filters were FIR filters and were adapted using the well known multiple error LMS algorithm.

FEEDFORWARD CONTROL

In Fig. 1 is depicted the experimental setup consisting of the double-glazing window built into the window testing facility and the adaptive feedforward controller. Six microphones in the receiving room were used to measure the mean sound pressure level with and without active control measures applied. The primary excitation was band limited white noise.

For a first test the reference signal of the feedforward controller was taken directly from the signal generator. This setup is somewhat “idealized” arrangement and in general will yield some kind of best case results. In the second test the reference signal is taken from an additional microphone in the sending room in front of the window. This is a more realistic case. In both test cases the controller hardware used allowed a maximal length of the (four) adaptive FIR filters of 256 coefficients.

The results of the tests, measured in third-octave bands, are shown in Fig. 2. The maximum level without control is found at 100Hz, because in this third-octave band the mass-spring-mass resonance frequency of the plate-cavity-plate system is included. In this region the sound insulation of the double-glazing window is comparatively low. The “idealized” feedforward control yielded a high level reduction in all measured third-octave bands. The reduction of the total sound pressure level was 9.5dB.

Upon employing a reference microphone, the improvements in the 100Hz third-octave band were...
FIGURE 2: Measured sound pressure level in the receiving room with and without active control. Excitation with band limited white noise.

nearly as high as in the case before but in the upper frequency range the improvements were much lower. In the 160Hz third-octave band was found even a higher sound pressure level than without control. The reduction of the total sound pressure level was only 4.7dB; the reasons being, firstly, that there is a feedback from the loudspeakers inside the cavity to the reference microphone that was not compensated for, and secondly, that the position of the reference microphone certainly is in one of the many nodes of the sending room modes. Out of these, the latter is the most important, which could be verified through the transfer function between primary and reference signals, showing a minimum around 160Hz. The results show that the relative position of the reference microphone with respect to the sending room field heavily influences the achievable level reduction by the active system. In a practical application however the reference microphone certainly would be built into the outer side of the window frame at a fixed position. This means that this position would mainly determine the measured transmission loss and of course would result in completely different transmission losses in other testing facilities. Free field conditions should be approximated in the sending room, in order to yield comparable results, but this is generally not the case in window testing facilities.

FEEDBACK CONTROL

The problems when using a reference microphone can be overcome employing a feedback controller instead of a feedforward. Here the adaptive feedback controller given in [5] has been investigated. This controller uses the error signals and the compensation signals for an estimate of the disturbance, which in turn serves as the reference signal for the adaptive algorithm. Fig. 3 illustrates the setup with the feedback controller. The block called ‘reference signal estimator’ includes an additional filtering with all the secondary paths. Due to the higher computational complexity compared with the feedforward controller only 128 filter coefficients could be realized with the existing hardware. The reduction of the number of filter coefficients negatively influences the performance of the controller (cf. Fig. 2). The reduction of the total sound pressure level was measured to only 3.3dB.

Nevertheless, some tests were performed with traffic noise, e.g. from highways, trains, jet aircrafts and helicopters, using the adaptive feedback controller (see also [4]). As always is the case in adaptive active noise control, the achievable performance highly depends on the nature of the signals. Especially for traffic noise signals, which consist of high harmonic components, reductions of the total sound pressure level of more than 8dB were obtained with the adaptive feedback controller.

ACKNOWLEDGEMENTS

This work was sponsored by the DFG, Deutsche Forschungsgemeinschaft, contract No. Mo390/6-1.

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An Improved Method for Off-line Secondary and Feedback Paths Estimation for Active Noise Control System

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In the Active Noise Control (ANC) system, the transfer function of feedback and secondary paths are always identified in off-line before adaptation of the ANC controller. In [1], an off-line method is presented to identify both feedback and secondary path simultaneously by inserting an additional white random noise to secondary loudspeaker. Two adaptive filters are used to identify both paths. However in case that it is impossible to turn off the primary noise, the adaptation of the two adaptive filters will be affected by the primary noise, especially the one to identify feedback path, since reference microphone is very close to the noise source. Therefore it is highly desirable to reduce the affect of primary noise to path identification. When the primary noise is periodic, its influence can be reduced by an adaptive line enhancer [2]. In this paper, two adaptive line enhancers are applied to remove the influence of primary noise to identification of both feedback and secondary path. Moreover, to save the computation power, two filters are not updated every sample. Simulation results show that more accurate transfer functions for both paths can be obtained, especially for the feedback path.

INTRODUCTION

Active Noise Control (ANC) system attempts to reduce acoustic noise by introducing a second sound source, which emits sound with same amplitude but inverse phase to the original sound. Feedforward ANC system is the most popular. A single input single output feedforward ANC system has two microphones and a secondary loudspeaker. A reference microphone senses the noise to be controlled, called primary noise, and provides prior knowledge about the primary noise to the controller. An error microphone measures the supposition result of primary and secondary noise, and also provides the adjustment information to the controller. To be adaptive to the changing environment, adaptive filters are always used as ANC controller. Based on the information provided by reference and error microphones, the ANC controller adjusts its weights and filters the reference signal to produce a proper signal so as to drive secondary loudspeaker. The loudspeaker emits an anti-noise signal, which destructs primary noise around the position of error microphone.

The most popular adaptive algorithm used in ANC system is the Least Mean Square (LMS) algorithm due to its simplicity and robustness. When the LMS algorithm is applied as the ANC controller, there exist two electro-acoustic paths, i.e., secondary path and feedback path. Secondary path starts from input of secondary loudspeaker to output of error microphone. It includes D/A converter, power amplifier, secondary loudspeaker, acoustic path from secondary loudspeaker to reference microphone, pre-amplifier, reference microphone and A/D converter. The existence of secondary and feedback paths causes potential instability of ANC system, so they have to be compensated. Although feedback path can be removed by using some physical methods, such as special arrangement of secondary loudspeaker or using directional reference microphone, etc. [1], sometimes the reference microphone unavoidably affected by the anti-noise from secondary loudspeaker. On the other hand, the secondary path always appears.

PROPOSED METHOD

To compensate the effect of secondary path, the filtered-x LMS (FxLMS) algorithm and its modified version are proposed [3][4][5][6]. The basic idea is to filter the reference signal \( x(n) \) by a estimated model of secondary path \( S(z) \), \( \hat{S}(z) \), and use the filtered reference signal to update the weights of ANC controller, \( \hat{W}(z) \) by the LMS algorithm. A way to compensate the effect of is feedback path neutralization [1], where the effects of feedback path \( F(z) \) is neutralized by a estimated model of this feedback path, \( \hat{F}(z) \).

The model of secondary and feedback path, i.e., \( \hat{S}(z) \) and \( \hat{F}(z) \) can be estimated by using adaptive filters without operation of the ANC controller \( \hat{W}(z) \).
This is called offline estimation of $\hat{S}(z)$ and $\hat{F}(z)$. A way to identify both paths at the same time by using adaptive filters is described in [1]. A white random noise $v(n)$ is used as excitation to modeling adaptive filters $\hat{S}(z)$ and $\hat{F}(z)$. The desired signal for secondary path $e(n)$ is obtained by error microphone, while that of feedback path $x(n)$ is sensed by reference microphone. When primary noise does not exists, $\hat{S}(z)$ and $\hat{F}(z)$ can model $S(z)$ and $F(z)$ quickly. However when primary noise exists, it becomes an interference to adaptation of both $\hat{S}(z)$ and $\hat{F}(z)$. In case the primary noise is periodic, it can be removed by using adaptive line enhancer [5]. This idea is proposed in [2] for online secondary path modeling. In this paper we extend it to offline secondary and feedback path modeling.

The proposed algorithm using line enhancer to reduce influence of narrowband primary noise to offline identification of secondary and feedback path is illustrated in Figure 1. Two adaptive line enhancers are applied. One is for removing effects of $x(n)$ to feedback path modeling, which is denoted as $\hat{H}_{f}(z)$, while another one is for reducing influence of $d(n)$ to secondary path modeling, which is expressed as $\hat{H}_{s}(z)$. Two delay units with taps $D_{f} > L$ and $D_{s} > M$ are used to decorrelate the signals due to $v(n)$.

**REFERENCES**


Figure 1. Block diagram of proposed method
The Problem of Choosing Sensor Configuration in Active Sound Control Systems: a Simple Numerical Illustration

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It has been shown recently that different sensor configurations can be made equally efficient in controlling an acoustic domain by weighting the primary pressure. There exists an optimal weighting which maximizes the robustness while still maintaining the efficiency. The very simple numerical application proposed allows us to verify some assertions that could seem surprising.

INTRODUCTION

Recent works have shown that all sensor configurations of an active sound control system could be made efficient by weighting the primary pressures, and also that robustness against spatial variations of the primary field could be maximized [1]. To illustrate these results, a very simple arbitrary numerical application is proposed, along with additional information obtained.

SENSOR EFFICIENCY THROUGH WEIGHTING

In auto-adaptive active sound attenuation systems, the secondary source driving signal(s), or control, is achieved by minimizing the total pressure at the control microphones. Ideally a large number N of them are located inside $\Omega$ where attenuation is sought. With $G$ the responses of the secondary source to the N sensors, and $I$ the identity matrix, the control $\phi_n$ that best minimizes the primary field $p_n$, called here 'of reference', solves the following problem:

$$\min_{\phi_n} \| G \phi_n + I \cdot p_n \| = \min J_{\text{res}}(p_n)$$

Whatever $p_n$, the optimal attenuation has the form

$$A^{\text{opt}}(p) = -10 \log_{10} \frac{p^* \cdot (I - A) \cdot p}{p^* \cdot p}$$

where $A = G^* \cdot H^* \cdot G$ and $H = H^* \cdot G$. It arises from the optimal control $\phi_{\text{opt}} = -H^* \cdot G^* \cdot p$. In particular, for $p_n$, we obtain $A^{\text{opt}}(p_n) = A_n$.

In reality, only a small number $N_c$ of sensors is available ($N_c < N$). Where should they be placed in order to achieve $A_n$ when dealing with $p_n$? Let $G_c$ be made up of the secondary source responses to the $N_c$ control microphones and $p_n$, the reference primary pressures at these sensors, and let $D$ be the weighting matrix that modifies the pressures. What $D$ leads to $A_n$ in $\Omega$ from an LMS control with $N_c$ sensors?

A first method consists in reducing to zero the functional $J_{\text{res}}(p_n) = \| G_n \cdot v + \tilde{D} \cdot p_n^* \|$ when $v = \phi_n$ leading to $\tilde{D} \cdot p_n^* = G_n \cdot H^* \cdot G^* \cdot p_n$. An infinity of full $\tilde{D}$ satisfy the equality while only one diagonal does.

Minimizing $J_{\text{res}}(p_n) = \| G_n \cdot v + \tilde{D} \cdot p_n^* \|$ at $\phi_n$ also leads to a solution. In this case, $G_n \cdot \tilde{D} \cdot p_n^* = H_n \cdot H^* \cdot G^* \cdot p_n$. An infinity of matrices $\tilde{D}$, diagonal or full, can satisfy the equality.

Having chosen a possible weighting $D$, the optimal attenuation of a primary field in $\Omega$, obtained from $N_c$ sensors, is expressed by

$$A_c(p) = -10 \log_{10} \frac{p^* \cdot (I - A + A_c) \cdot p}{p^* \cdot p}$$

where

$$A_c(D) = (H^{\dagger} \cdot G^* \cdot D \cdot P - H^{\dagger} \cdot G^{\dagger} \cdot H^{\dagger} \cdot (H^{\dagger} \cdot G^* \cdot D \cdot P - H^{\dagger} \cdot G^{\dagger}))$$

with $D$ such that $A_c(p_n) = A_n$.

Let us consider a very simple arbitrary situation. The domain $\Omega$ is made up of 3 microphones. The reference primary pressures and the secondary source responses are respectively 4,2,6. and 1,2,3. The configuration with both sensors $M_1$ and $M_3$ is chosen. Dealing with diagonal matrices $D$ only, we obtain

$$\hat{D} = \begin{bmatrix} 13 & 0. \\ 28 & 13 \\ 0. & 14 \end{bmatrix}$$

and

$$\tilde{D} = \begin{bmatrix} 65 & 28 \\ 65 & 9 \end{bmatrix}$$
Thus, \( J_m(p_n), \tilde{J}_m(p_n) \) and \( \tilde{\tilde{J}}_m(p_n) \), the latter for an arbitrary value of \( s \), reach their minimum at \( \phi_n \).

**MAXIMIZING ROBUSTNESS**

Let us work \( p = p_n + \delta p \) with \( \delta p \). Of all values of \( e \) leading to the same value of \( \delta A^{\text{opt}} \), one is minimum and called \( e_{\text{min}} \). Figure 1 shows all the possible values of \( A^{\text{opt}}(p) \) when all variations \( \delta p \) are admissible. The minimum attenuation \( A_{\infty}(e_{\text{min}}) \) decreases when \( e_{\text{min}} \) increases.

![FIGURE 1. Set of optimal attenuations with lower boundary \( A_{\infty}(e_{\text{min}}) \)](image)

The minimum attenuation is the solution of

\[
\begin{align*}
A_{\infty}(e_{\text{min}}) &= \min_{p \in E} A^{\text{opt}}(p) \\
E &= \{ p = p_n + \delta p, \|\delta p\|_{\text{min}} \leq e_{\text{min}} \}
\end{align*}
\]

With only a small number of sensors to control \( \Omega \), the minimum attenuation in \( \Omega \), \( A_{\infty}(e_{\text{min}}) \), still decreases when \( e_{\text{min}} \) increases, but now there exists a maximum value of \( e_{\text{min}} \) beyond which we cannot guarantee the absence of amplification. \( A_{\infty}(e_{\text{min}}) \) solves the same type of problem as above.

A sensor configuration is all the more robust that it guarantees a better minimum attenuation. Maximizing the robustness consists thus in finding an optimal weighting matrix \( \tilde{D} \) leading to the highest curve \( A_{s}^{\text{opt}}(e_{\text{min}}) \). At best, it is similar to \( A_{\infty}(e_{\text{min}}) \), i.e., \( \tilde{D}_{\text{opt}} \). Let us consider the configuration M1-M2 with \( \tilde{D} \) diagonal. Efficiency at \( p_n \) is assured by \( v = \left( \frac{65}{28} - s \right) \), and \( N(\tilde{D}) - A \) subsequently has the form

\[
\begin{bmatrix}
1. \\
2. \left( \frac{s}{5}, 1, \frac{13}{14}, -2, \frac{s}{5}, \frac{3}{14} \right) \\
3. \left( \frac{s}{5}, 1, \frac{13}{14}, -2, \frac{s}{5}, \frac{3}{14} \right)
\end{bmatrix}
\]

Thus \( \min (s) \left( \frac{s}{5}, 1, \frac{13}{14}, -2, \frac{s}{5}, \frac{3}{14} \right) \). By exhaustive calculation, curves \( A_{s}^{\text{opt}}(e_{\text{min}}) \), increased by optimization. Configuration M1-M3 already robust with \( \tilde{D} \) diagonal has not been improved, while configuration M2-M3 has reached \( A_{\infty}(e_{\text{min}}) \).

![FIGURE 2. Robustness maximization for configuration M1-M2](image)

**CONCLUSION**

Finally, additional information has been obtained: there exists a weak form \( \tilde{D} \) of the weighting matrix which makes it possible, when diagonal, to maximize robustness as defined in the text. In the particular example chosen for the illustration, the solution can almost be obtained by hand.

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A generalised Model of Sound Emission Transducers in active Noise Control and Arrays

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This paper deals with a generalised model of electroacoustical sound emission transducers for their design and measurements in the fields of ANC and arrays. This model, able to describe the behaviour of a transducer both in reception and emission modes, is based on the representation of the electromechanical (eg. electrodynamic, electromagnetic, etc.) and mechanoacoustical couplings by two-port networks, with associated two-terminal elements. In active noise control a major difficulty is to take into account the effects of the primary noise on the transducers, particularly regarding phase relationship. The same difficulty arises in loudspeaker arrays. A practical problem is to design the transducers in such a way that these effects are reduced to a minimum. The generalised model allows these problems to be solved by representing the primary field using a real source – eg. an ideal pressure source with an associated impedance – instead of modifying the radiation impedance as is usually done. Simplified diagrams, obtained by applying network theory, proved to be powerful and useful tools for the design and characterisation of sound actuators.

BASIC THEORY

Most of electroacoustic sound sources are made of an electromechanical transducer and a mechanoacoustical coupling involving a radiating face. The main assumptions for their study are: $kd < 1$, $k$, wave number, $d$, characteristic dimension (eg radius of a circular piston), quasi linear domain (small non-linearities). In the following, we use symbols, quantities, units, conventions and signs according to the IEC.

They can be described by lumped-constant circuits with two-port networks which represent the electromechanical and mechanoacoustical couplings [1]. Equivalent circuits can be drawn up that only include elements and quantities of the same nature by removing the coupling networks. Finally, very simple circuits are obtained using circuit theory. Figure 1 gives the basic circuit of a sound emission transducer. A two-port network connects an electrical system, represented by a real source ($U_g$, $Z_g$), and a mechanical/acoustical system or medium, represented by a mechanical impedance $Z_{ml}$ equivalent to the acoustical load. For example the acoustical radiation impedance becomes $Z_{ac} = S^2Z_w$, with an appropriate value of the radiating face equivalent area $S$. Figures 2 and 3 give the networks which correspond to the two fundamental linearized couplings, that is: a) force controlled by the current, - electrodynamic, electromagnetic and magnetostrictive conversions; b) force controlled by the electric charge, - electrostatic and piezoelectric conversions. These circuits can be reduced to a simpler one by Thevenins’ theorem, as given in the figure 4, where $E_g$ and $Z_{ml}$ represent the transducer. In ANC, the need for improved modelling of transducers is recognised, mainly in view of their design (definition of their requirements) and testing, taking into account the effects of external sound fields: a transducer can be subjected to the noise to be counteracted as well as many other neighbouring transducers. If we consider the resulting external sound pressure $p_e$, a net force $F_e$ is exerted on every transducer, which can always be expressed as $Sp_e$. Usually, as seen in the literature, the effect of external fields is taken into account by modifying the transducer radiation impedance, i.e. through the impedance $Z_{ml}'$. That means the “melting” of two different sources of forces: the reaction to the radiation (that is the fundamental meaning and “raison d’être” of the radiation impedance) and the external action. We think that this is not the best way of modelling and that it is preferable to maintain the dissociation.

DESIGN

Figure 5 shows how these effects can be simply represented, as previously, by a load impedance and a force source $F_e$. With ad hoc values of the elements, design can be carried out taking into account the external field effect. For example, if sound pressure cancellation is required on the radiating face, the transducer is short-circuited: $(v - v_e)$ is maximised and the displacement too. In consequence the effects of non-linearities are augmented. The ratio $Z_{load}/Z_{ac}$ and the value of $F_e$ determine the ability of a transducer to tackle this extreme condition. For example for an electrodynamic transducer, the most severe
perturbation occurs around the resonance frequency. These effects are minimised by current control instead of voltage control [2].

![Figure 1: Basic transducer circuit](image1)

To test the ability of a sound emission transducer to be used in a ANC system, a measurement set-up was built. The transducer under test is mounted on a baffle with an auxiliary source generating an external sound field. A sine wave signal is applied to the auxiliary source and a near-field microphone measures the sound pressure very close to the transducer. By acting on magnitude and phase of the transducer excitation, the minimum sound pressure level (or any pertinent value) at the microphone position is reached.

This test has been proposed within the framework of a BE European project and proved to be successful for transducer assessment. It was used to characterise some limitations of prototypes when interacting strongly with the an external field, in this case with almost no front acoustic pressure. Different transducer technologies (direct radiator, compliant structures, airflow modulation, electrodynamic, piezoelectric) have been assessed using this set-up.

Figure 6 presents typical spectra which are obtained with a transducer under such a test. The fundamental frequency has been reduced by 20dB, while the harmonic content increases. As has been theoretically described, this measurement shows clearly that in the presence of an external sound field the transducers exhibit increasing non-linear behaviour by getting closer to their working condition limits.

![Figure 6: Near-field SPL measurement with (bold) and without (fine) an external sound field (1900Hz sine wave).](image2)

ACKNOWLEDGMENTS

The work presented here has been supported by the Swiss Federal Education and Science Office (OFES), to whom we are grateful.

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A Practical, Fast and Cost-efficient Algorithm for Multiple Input, Multiple Output Active Noise Control Applications

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Active noise control has proven to be an efficient solution to low frequency noise problems in many different applications. A large part of these applications are concerned with harmonic noise control and typically require control systems with several inputs and outputs. As the systems grow, the demand for processor capacity increases rapidly, resulting in large and expensive hardware platforms. As multiple-input, multiple-output noise control is restricted to the control of periodic noise, a controller structure that is adapted to the signal type may significantly reduce the requirements on the hardware capacity. This paper discusses a complex, time-domain controller that is designed for the control of harmonic components. The structure of the controller is simple, easily implemented and can easily be extended to handle any number of noise references and any number of harmonics. The convergence properties of a multiple-input, multiple-output control system depends largely on the acoustic coupling between the active sources and the control sensors. To get accurate and stable control it is necessary to use normalization, i.e. a weighting function that optimizes the controller for each control source. With the presented controller structure, the use of normalization is straightforward and a number of different approaches for normalization is discussed as well as examples from practical implementions.

INTRODUCTION

The foundation for the discussion in this paper is a feed-forward, adaptive controller. Such controller structures have been extensively examined and discussed in the literature since the birth of the LMS algorithm in the late fifties. A feed-forward controller implies that there is a reference signal that is correlated to the disturbance. The control output is applied to the environment through an actuator. The influence on the plant by the actuator is measured by one or more control sensors. The discussion in this paper is directed towards the use of several control outputs and several control inputs (MIMO). The controller structure can also handle reference inputs from any number of noise sources.

The standard LMS algorithm is derived for random input signals. Since MIMO control implies the control of harmonic components, it makes sense to adapt the control scheme to the signal type. Periodic reference signals can be generated internally within the controller. There are huge advantages with this method compared to measuring the reference signal with some kind of sensor: The reference signal will contain only those harmonics that are to be controlled and the properties of these signals, i.e. frequency and signal power, are known. The reference signal is generated by using a synchronization signal that is taken from the noise generating mechanism, e.g. a rotating machinery. The synchronization signal can be used to control the sampling rate for the controller (order based control) or to generate reference signals with the appropriate frequencies in a controller with a fixed sampling rate.

THE COMPLEX LMS CONTROLLER

Traditionally, one composite reference signal is generated, containing all harmonics to be controlled. The reference signal is fed through an FIR filter with sufficient number of filter weights to adjust the magnitude and phase for each harmonic.

In the proposed controller structure, each harmonic is considered to be a separate reference signal and is treated equivalently to references from different sources\cite{1}. The reference generator produces one complex reference signal for each harmonic and each reference is controlled by one complex weight. Thus, the (real) signals at the L outputs $y(n) = [y_1(n) \cdots y_L(n)]^T$ are given by

$$y(n) = \sum_{r=1}^{R} \Re \{x_r(n)w_r(n)\}, \quad (1)$$

where $x_r$ is the complex reference signal for reference $r$, $w_r(n) = [w_r1(n) \cdots w_rL(n)]^T$ is the $L \times 1$ vector of complex weights for this reference and $R$ is the number of references. $\Re \{\cdot\}$ denotes the real part of the complex multiplication which implicates that in a practical implementation only the real part is evaluated. One may note that the output signal is simply obtained as the sum of the contributions from all references.

The LMS update algorithm is derived in the same manner as for the real LMS algorithm. For reference $r$ this results in

$$w_r(n + 1) = \beta w_r(n) - 2M_r x^*_r(n)\hat{F}_r e(n)$$

where $\hat{F}_r$ is a $M \times L$ matrix of complex gain elements, describing the change in magnitude and phase between
the elements in $\hat{F}$ are estimated (i.e. measured) values. $e(n) = [e_1(n) \ldots e_M(n)]^T$ is the $M \times 1$ vector of real control sensor inputs and $M_r$ is a square ($L \times L$) convergence factor matrix. The formulation of this matrix is further discussed below. The leakage factor $\hat{\rho}$ is close to one and is very useful in practical implementations, since it restrains the behavior of the weight vector $w$.

**NORMALIZATION**

The convergence characteristic for a multiple input, multiple output controller is determined by the acoustic (or structural-acoustic) conditions, i.e. the complex gain between each control output and every control sensor input. Variations in the complex gain between outputs results in different convergence rates for different outputs, which in turn effects the overall attenuation[3][4].

In general, the quantity that governs the convergence rate and the stability of the controller is given by the Hessian matrix $\mathbb{E} [x^r_r(n)\hat{F}^H_r \hat{F}_r x_r(n)]$ [2]. Under the given assumptions, the reference signal $x_r(n)$ is deterministic and for the ordinary MIMO LMS controller this expression reduces to[6]

$$M_r' = \mu_0 (\hat{\rho}, \text{trace} \{ \hat{F}^H_r \hat{F}_r \})^{-1} I$$  \hspace{1cm} (3)

where $I$ is an $L \times L$ identity matrix, $\mu_0$ is a conveniently chosen positive convergence factor, and $\hat{\rho}$ denotes the power in the reference signal (which is known). Equation (3) ensures a stable but slow controller, since the update algorithm is given the same normalization for all outputs. Another alternative is the Newton-like algorithm[2][6], which is obtained when the convergence matrix is chosen as

$$M_r'' = \mu_0 (\hat{\rho}, \hat{F}^H_r \hat{F}_r)^{-1}.$$  \hspace{1cm} (4)

By combining equations (2) and (4), it is obvious that the full matrix of complex gains is used in the update of each output controller. This is a very powerful algorithm, but the computational cost is fairly high since a full complex matrix multiplication must be performed at each sampling interval.

In the third alternative presented here, the update algorithm for a specific output is normalized with the complex gains that are related to that particular output[4]. The convergence factor matrix becomes

$$M_r'' = \mu_0 (\hat{\rho}, \text{diag} \{ \hat{F}^H_r \hat{F}_r \})^{-1}$$  \hspace{1cm} (5)

where the matrix $\text{diag} \{ \hat{F}^H_r \hat{F}_r \}$ is the diagonal matrix consisting of the diagonal elements of $\hat{F}^H_r \hat{F}_r$. This matrix has real elements, given by

$$\mu_{rI} = \frac{\mu_0}{\rho \sum_{m=1}^M |F_{rm}|^2}.$$  \hspace{1cm} (6)

These elements are calculated off-line and require only one (real) multiplication for each output controller (and reference).

In a practical application, the sum of the squares of the complex gains, as given by equation (6), may well be larger than the sum of the cross terms, i.e. complex gains between different outputs and control inputs. In this case, the matrix $\hat{F}^H_r \hat{F}_r$ is diagonally dominant, which leads to[3]

$$\text{diag} \{ \hat{F}^H_r \hat{F}_r \} \approx \hat{F}^H_r \hat{F}_r.$$  \hspace{1cm} (7)

Under these circumstances, the algorithm obtained by using the convergence matrix given by (5) has a performance that is comparable to the Newton algorithm.

**PRACTICAL APPLICATIONS**

The suggested normalization leads to a very simple algorithm with no matrix operations and with a minimum of complex operations. It can easily be extended to any number of inputs, outputs and references (harmonics). The algorithm has been implemented in C on a TMS320C32 floating point signal processor for use in a silent seat application and to reduce noise from the rotating cutter in lawn mowers. The algorithm has also been used to control propeller noise in a mock-up of a twin-prop aircraft[4].

**REFERENCES**

Fundamental Consideration of Active Noise Control Based on Division of Input Signal

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The Filtered-x LMS algorithm is very often adopted for the active noise control, but it requires the estimation of transfer function of error path \( \hat{C} \) in advance. In this study, differing from the Filtered-x LMS algorithm, a new method for ANC based on division of input signal is proposed, i.e. algorithm which identify the objective unknown system \( \hat{h} \) and transfer function of error path \( C \) simultaneously by dividing input signal into two components. The input vector is divided into two components by introducing an adequate coefficient matrix and signal vector. One component is used for estimating \( h \) and the other for \( C \). Finally, the effectiveness of the proposed method is confirmed through the simulations.

INTRODUCTION

Recently, the ANC system[1] which controls an objective noise mainly in a low frequency region by sound becomes very popular in many engineering fields according to the rapid development of the technique in Signal Processing, especially DSP (Digital Signal Processor). The Filtered-x adaptive algorithm is generally and very often adopted to update a single-channel Feed-forward ANC system on controlling the noise in a duct, etc. The Filtered-x algorithm usually needs to estimate in advance a system model of error (secondary) path (so-called \( \hat{C}(z) \) filter) between the loudspeaker and the error microphone, because the system \( C(z) \) is considered as a priori information. However, it is inevitable that this \( \hat{C}(z) \) filter has modeling error due to the temporal change of transfer function, etc.[2, 3] Therefore, the ANC system deteriorates its performance and sometime becomes unstable in the worst case.

In this study, differing from the usual Filtered-x LMS algorithm, a new method for ANC based on division of input signal is proposed, i.e. algorithm which identify the objective unknown system \( \hat{h} \) and transfer function of error path \( C \) simultaneously by dividing input signal into two components. The input vector is divided into two components by introducing an adequate coefficient matrix and signal vector. One component is used for estimating \( h \) and the other for \( C \). Furthermore, the effectiveness of the proposed method is confirmed through the simulations.

DIVISION OF INPUT SIGNAL

Let consider the ANC system as shown in Fig.1. Here, \( K(z) \) denotes the transfer function from the detection sensor to the error sensor and \( \hat{C}(z) \) from the loudspeaker to the error sensor.

Let \( x(k) \) be the input and \( d(k) \) the desired output, the error \( e(k) \) is given by

\[
e(k) = d(k) + H(z)C(z)x(k) = d(k) + \{X_{N,N}(k)C_N\}^T h_N
\]  

(1)

with

\[
X_{N,N}(k) = \begin{pmatrix}
  x(k) & x(k-1) & \cdots & x(k-N+1) \\
  x(k-1) & x(k-2) & \cdots & x(k-N) \\
  \vdots & \vdots & \ddots & \vdots \\
  x(k-N+1) & x(k-N) & \cdots & x(k-2N+2)
\end{pmatrix}
\]

(2)

By applying the LMS method, the successive estimation algorithm can be derived. However, since this LMS algorithm does not utilize directly the input noise \( x(k) \) but the filtered input \( X_{N,N}(k)C_N \), the coefficient vector \( \hat{C}_N \) must be estimated as \( \hat{C}_N \).

Now, by considering an arbitrary signal as the composition of random vectors[4], it can be represented in a linear combination of noise \( v_n \) as follows:

\[
x(k) = \sum_{n=0}^{K} a_n(k) v_n
\]  

(3)

Here, \( a_n(k) \) denotes a coefficient and \( K \) is the number of data points in each frame. We consider the input signal \( x(k) \) as the arbitrary signal. Equation (3) can be expressed in vector-matrix notation as follows:

\[
x(k) = a_1(k)v_1 + a_2(k)v_2 + \cdots + a_K(k)v_K = V a(k)
\]  

(4)

Then, the coefficient vector \( a(k) \) can be calculated (though is unknown) as follows:

\[
\hat{a}(k) = V^{-1}x(k)
\]  

(5)

When the random vector matrix \( V \) is assumed to be the sum of two random vector matrices \( V = V_1 + V_2 \), \( x_1(k) \) and \( x_2(k) \) can be set respectively as:

\[
x_1(k) = V_1 \hat{a}(k), \quad x_2(k) = V_2 \hat{a}(k)
\]  

(6)
On estimating \( \hat{h} \) and \( \hat{C} \), the effective information on the input can be divided into \( x_1 \) and \( x_2 \), though the error signal \( e(k) \) at the error sensor is entirely same as that in Eq.(1). This means that \( x_2 \) is an ambient noise when we regard \( x_1 \) as the effective information, while \( x_1 \) is ambient when regarding \( x_2 \) as the effective information.

Thus, the recurrence estimation algorithms for the unknown system \( \hat{h} \) and the filter \( \hat{C} \) can be obtained respectively as follows:

\[
\hat{h}_N(k+1) = \hat{h}_N(k) - \alpha \hat{e}(k)X_{N,N}^{(1)}(k)\hat{C}_N(k) \quad (7)
\]

\[
\hat{C}_N(k+1) = \hat{C}_N(k) - \alpha \hat{e}(k)X_{N,N}^{(2)}(k)\hat{h}_N(k) \quad (8)
\]

Here, \( \alpha \) is a constant related to the convergence speed. The matrices \( X_{N,N}^{(1)}(k) \) and \( X_{N,N}^{(2)}(k) \) correspond to the divided inputs \( x_1(k) \) and \( x_2(k) \), and the error signal can be constructed as follows:

\[
\hat{e}(k) = d(k) + \{X_{N,N}(k)C_N\}^T\hat{h}_N \quad (9)
\]

**DIGITAL SIMULATION**

![Figure 2](image2.jpg)

**FIGURE 2.** EA of \( \hat{h} \) (Top) and Reduction (Bottom).

The effectiveness of the proposed method is confirmed through digital simulation. The tap lengths of the systems are set as 128 and the number of data sample is 4 \( (K = 3) \). As input \( x(k) \), the time series with AR model \( x(k+1) = bx(k) + \varepsilon(k) \) of 1st order is adopted (where \( b = 0.5 \) and \( \varepsilon \) is the Gaussian random number with 0 mean and variance 1). To evaluate the performance of algorithm, the estimation accuracy\( (\text{EA}) \) and the noise reduction\( (\text{Reduction}) \) are introduced as:

\[
\text{EA} = 10 \log_{10} \frac{\sum_{i=0}^{N-1} h_i^2}{\sum_{i=0}^{N-1} (h_i + \hat{h}_i)^2}, \quad (10)
\]

\[
\text{Reduction} = 10 \log_{10} \frac{d_k^2}{(d_k + y_k)^2}. \quad (11)
\]

Figure 2 shows the estimation accuracy of parameters and the noise reduction \( (\alpha = 0.1 \times 10^{-4}) \). In this case, it is obvious that the good results in both EA and Reduction are obtained.

Next, at the \( 2 \times 10^7 \)-th iteration, the transfer function \( C(z) \) of error path is changed. That is, we increase all the coefficients of impulse response for \( C(z) \) with 10 % and give their sign at random. Figure 3 shows the estimation accuracy of parameters and the noise reduction. After the parameter change, we can see that both EA and Reduction are recovered up to the levels before changing.

![Figure 3](image3.jpg)

**FIGURE 3.** EA of \( \hat{h} \) (Top) and Reduction (Bottom).

Finally, the estimation procedure is also demonstrated when the S/N (Signal to Noise Ratio) is set to 20 (dB) as shown in Fig.4. It is found that the proposed method can also reduce the objective noise.

![Figure 4](image4.jpg)

**FIGURE 4.** EA of \( \hat{h} \) (Top) and Reduction (Bottom).

**REFERENCES**

Identification of Secondary and Feedback Paths in ANC
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The paper deals with the identification and realization of secondary and feedback paths in multi-channel active noise control systems. A feed-forward ANC system with microphones as reference and error sensors requires identification of the transfer function from the digital controller outputs to corresponding inputs to ensure convergence of the algorithm used; a filtered-x LMS algorithm was assumed. Three different off-line methods of identification were tested and compared: identification using white noise, identification using MLS signals and training by adaptive algorithm. The algorithms and arrangements were tested on an experimental duct of rectangular cross-section of dimensions similar to actual ventilating or air-conditioning ducts. The source of noise was realized by means of an axial flow fan with five blades, with the flow speed being regulated by regulation of fan revolutions. The tested algorithms were implemented on a digital control unit based on Texas Instrument floating point DSP.

INTRODUCTION

One of the first applications of active noise control (ANC) is the attenuation of fan noise in a duct. If the cross-section of the duct is large enough, higher modes have to be assumed in the frequency range of ANC, and a multi-channel system has to be assumed. If the feed-forward ANC system is used, a precise identification of all error and feedback paths is of crucial importance for the system performance. The off-line identification and modelling is reated in this paper.

IDENTIFICATION AND MODELLING

Off-line identification can be performed by three different methods: identification using white noise generated by a noise generator, identification using MLS signals produced by the controller, and identification based on training by adaptive algorithm. Identification process results in impulse responses or transfer functions of error and feed-back paths respectively. They can be realized by FIR of IIR filters in the controller. FIR filters are always stable and have a linear phase, but the length of the impulse response is determined by the order of the filter. Difficulty with the stability and design of IIR filters is compensated by a longer useful impulse response for a lower number of coefficients.

FIR filters modelling

FIR models of error and feed-back paths can be acquired by means of an adaptive training algorithm in which the white noise is sent to the secondary loudspeaker and picked up by the appropriate microphone. The adaptive algorithm minimizes the difference between the measured response of the chosen path and the response of the model filter.

The second possibility of identification is based on measurement of impulse response by MLS signals. MLS signals produced by the controller (DSP unit) pass through the chosen error of feed-back path and then the impulse response is calculated. First $N$ points of measured impulse response correspond to $N$ coefficients of FIR filter if measured on the same sampling frequency.

The length of the impulse response of FIR filter is determined by the number of coefficients $N$. However, impulse responses of both error and feed-back paths can be very long, because the transfer functions are affected by the reflections on the input end of the duct. To obtain good approximation of mentioned transfer functions, very long FIR filters are necessary.

IIR filter modelling

In the two-channel ANC system depicted in figure 2, six model filters must be used. Sufficiently long (large) FIR filters spend significant portion of controller memory and computational time. This can be solved by using IIR filters, which can be stable only if their poles lie
within the unit circle. The IIR filter coefficients can be obtained by solution of Prony equations to minimize a error function. As these equations are nonlinear with respect to both nominator and denominator coefficients the standard Prony method is used to obtain suboptimal solution. For precision of the filter coefficients, the Steiglitz-McBride method iterative method is used. In practical implementation, the filter poles can exceed the unit circle due to the approximation (rounding) error. This effect has to be considered in DSP realization. An example of error path poles is shown in figure 3. Comparison of measured and modelled feed-back path can be seen in figure 4.

rate up to 1.5 cubic meters per second was used. Flow speed was regulated by regulation of fan revolutions. A high pass filters of 100 Hz were implemented at all inputs of the controller [2]. For modelling of error paths, the IIR filters of 20th order were used. For feedback paths, the order of 40 was chosen. An example of attenuation of fan noise is shown in figure 5. In this article modelling of error and feedback paths using IIR filters for two-channel ANC in a duct was presented. Attenuation results obtained show the system to perform well, achieving attenuation up to 15 dB.

ACKNOWLEDGMENTS

This work has been partially supported by the research program of the CTU Prague J04/98:212300016 and partially by the GACR research project No.102/01/1370 Multichannel system of active noise control.

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