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Divers
General
Diverses
PROBLEMS OF THE DEVELOPMENT OF NOISE TEST AND LABELLING CODES IN RELATION TO ELECTRIC DRILLS

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1. Introduction

The determination of the sound power level of a noise source poses no problems from the acoustical point of view, since a series of basic international standards are available in that field. Experience, however, shows that the accuracy of the measurement of a particular machine depends on the operating rather than the acoustical conditions. Consequently, during the preparation of a Noise Test Code the mechanical and the acoustical requirements should be harmonically specified.

The paper discusses those questions arising at the development of the measuring method for electric drills. The aim of the work was to reveal the problems rather than to work out a definitive Noise Test and Labelling Code. The drills as such were chosen in our investigation mainly for their cheapness, easy operation under different conditions and the availability in sufficient number for statistics.

2. Measurements under normal working conditions

The sound power levels of four different drill types were measured under normal drilling operation. The drills were fixed in their own frame, equipped with new twist drills and a metal block was drilled with constant feed forces. The sound power levels were measured by means of a computer-controlled multichannel frequency analyzer with 1 sample/sec sampling speed. The rotation speed of the chuck was also measured and continuously recorded, enabling the sound power spectra to be measured as a function of the rpm. A number of measurements were done with different twist drill diameters, feed forces and materials.

The tests revealed that the sound power levels of the drilling operation are highly determined by the rotation speed, i.e., the load of the drill while the effect of the twist drill diameter, feed force or material are of little importance or negligible. A number of measurements for one drill are summarized in Fig. 1. In terms of A-weighted sound power levels. The results seemed to suggest that the sound power level of the drills can be measured under normal drilling operation. It was also established, however, that the sound emission of the drill can remarkably be influenced by radiation from the normal frames. This is visualized by sound intensity measurements made by means of a Brüel and Kjaer sound intensity analyzer,
see Fig. 2. On the other hand, the load seems to be an important operating parameter and therefore the sound emission of the drills has to be investigated and measured by means of a special loading equipment.

3. Measurements under simulated load conditions

The sound power level spectra were measured as a function of the load under stationary conditions in a 272 m² reverberation chamber by means of a loading equipment consisting of a car generator, an electronic voltage regulator and a loading resistor. Fig. 3 shows two typical curves for two drills of the same type. We found that the line of the load curve depends highly on the ratio of the tooth meshing and ventilation noise components. The tooth meshing components increase, while the ventilation components decrease by increasing load or decreasing rpm. The relative contribution of the tooth meshing components is influenced by a great number of parameters including the production quality and the alignment of the gears and the bearings, wear, temperature, fixation of the drill, etc.

To investigate the effect of the load on the average sound power level and the accuracy of the measurements, a round robin test was done with participation of four laboratories in Budapest. The sound power levels of ten drills of the same type were determined unloaded and at rated input electrical power according to [1]. In addition to the survey measurements, the sound power levels were also measured according to [2]. The results, including the average standard deviations of the repeatability and reproducibility computed according to [3], are summarized in the table below. The averages and the standard deviations of the differences between the survey and laboratory measurements as a function of octave band center frequency are depicted in Fig. 4.

<table>
<thead>
<tr>
<th>Operating conditions</th>
<th>Results of survey measurements</th>
<th>Differences between survey and lab. measurements</th>
<th>Characteristic standard deviations</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>Ave. S.d.</td>
<td>Ave. S.d.</td>
<td>Repeatability</td>
</tr>
<tr>
<td>Idling</td>
<td>94.3 1.2</td>
<td>2.8 1.4</td>
<td>0.4</td>
</tr>
<tr>
<td>Rated load</td>
<td>96.2 3.5</td>
<td>1.8 2.6</td>
<td>0.5</td>
</tr>
</tbody>
</table>

The table reveals that the measurements on the loaded machine may result in higher sound power levels than at idling. It means that the loaded operating conditions may also have to be considered or sometimes preferred, if one wants to specify the operating conditions of the highest noise emission. This is in conflict with the prescriptions of the present day existing national drafts or standards, see e.g. [4] and [5]. The higher standard deviation of the loaded measurements may also be advantageous since it makes the classification of the machines easier. The average differences between the survey and laboratory measurements is less for rated load than for idling, and can partly be explained by theoretical reasons [6]. On the other hand, the loaded measurements are connected with higher statistical uncertainty, although all these inaccuracies are significantly lower than the specified accuracy of the survey measuring method.
4. Methods of noise labelling

As opposed to the measurement of the sound power levels, very few experience is available about determining the noise emission values of machines. Some information can be gained from [7] and two procedures are described in [8]. Most of the noise labelling methods are of statistical nature and based on the assumption that the sound power levels of the machines are normally distributed. To approve or reject the hypothesis of normality, samples were taken from three drill and four other machine types. The size of the samples ranged from 10 to 152 items, and the sound power level data sets were tested for normality. Results of the tests will be detailed elsewhere [9]; all that we can note here is only that each distribution were found normal on 95 or 97.5% significance level.

According to the normal distribution, the noise label of a sample can be computed according to the formula

\[ L_C = \mu + k \cdot \sigma_f \]

where \( \mu \) is the mean value of the sample, \( \sigma_f \) is the total standard deviation of the sample, and \( k \) depends on the order of the quantile to be estimated [8]. The accuracy of the formula have been investigated theoretically and experimentally. We found that it results in an unbiased estimation but the confidence intervals of the estimation are relatively high, and decrease only slowly with increasing size of the sample. An estimation of the noise label with considerably smaller error than the total standard deviation can be gained if the size of the sample is higher than about 50.

5. Acknowledgements

This research was supported by the National Authority for Environmental Protection and Nature Conservation, Budapest. The author wishes to express his appreciation to Mrs. J. Angster, Mr. F. Hirka and Mr. F. Kvačka and their colleagues for their kind assistance in the round robin measurements and to Mr. L. Czabálly for his valuable remarks on the whole process of realizing our project. Thanks are also due to Mr. J. Braasch at Bruel and Kjær who was kind to make the intensity measurements possible for us.

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Problems of the development of noise test
HEAD AND TORSO SIMULATOR (SAMRAI) WITH SIMPLIFIED ARTIFICIAL EAR AND ITS APPLICATION TO SIMULATED IN SITU MEASUREMENT OF HEARING AID.

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Introduction

A simplified artificial ear terminated with resistance element has been developed. Results of measurement of hearing aid using Head and Torso Simulator (SAMRAI) with this artificial ear are shown. A predicting method of vent response of hearing aid is also discussed.

Concept of in situ measurement

For the evaluation of hearing aid performance, one of the most suitable positions is at the eardrum of a subject using the hearing aid. The measurement in that position is so-called in situ measurement. Among in situ measurement data, insertion gain is the most important one. Insertion gain is defined by

\[ 20 \log \frac{P_H}{P_0} \]  

(1)

where \( P_0 \) denotes sound pressure at the eardrum of a subject without hearing aid (i.e. open ear condition), and \( P_H \) sound pressure at the same position with hearing aid. Insertion gain represents effective amplification of the hearing aid. And it is the objective data to be compared with hearing level of the hearing impaired subject.

Head and Torso Simulator (HATS) for simulated in situ measurement

In practice, however, it is difficult to measure sound pressure at the eardrum of each subject. So it may be desirable to substitute a well-defined HATS for subject. Therefore, HATS should be designed with specified median adult dimensions, in order to simulate head diffraction, body baffle effect and pinna reflection of the median adult.

At first, artificial ear of HATS must simulate the average transfer function from ear canal entrance to eardrum of the real ear, because it must reproduce eardrum pressure under the
open ear condition. Secondly, artificial ear must simulate the average transfer impedance of the real ear so that it may reproduce eardrum pressure with an earphone inserted.

We have developed an artificial ear terminated with a simple acoustic resistance element of 320 cgs ohms. The resistance element was realized by an acoustic tube having a diameter of 4 millimeters and a length of about 5 meters.

Fig. 1 represents calculated transfer functions of the simplified artificial ear and the real ear model. The real ear model is defined as a theoretical model having a cylindrical ear canal and the eardrum impedance proposed by Shaw.

Fig. 2 represents calculated transfer impedances of the artificial ear and the real ear model. Below 0.7 kHz, the artificial ear shows disagreement in transfer impedance relative to the real ear model. The disagreement means that the earphone response measured with this artificial ear differs from the response on real ear at these frequencies. However, it is possible to compensate the response difference as far as the earphone produces a constant volume velocity.

Measurement results

Fig. 3 shows simulated open ear gain of HATS (SAMRAI) with the simplified artificial ear. Open ear gain represents transformation of sound pressure from the free field to the eardrum. The real ear data of Yamaguchi & Sushi and Shaw are shown in the same figure for comparison.

Fig. 4 shows measured insertion gain of a behind-the-ear hearing aid on SAMRAI. The other curves represent measurement results on 3 subjects using probe-tube microphone. All the subjects are Japanese male adults and have normal hearing. The response measured on SAMRAI shows good agreement with the real ear responses. From the theoretical point of view, SAMRAI's response must differ from real ear response below 0.7 kHz, because the simplified artificial ear of SAMRAI has different transfer impedance from real ear, as seen in Fig. 2. Nevertheless, there appears no obvious difference between the SAMRAI's response and the real ear responses below 0.7 kHz. The agreement may be due to natural leakage between the earmold and the ear canal of each subject.

Predicting method of vent response

Below 0.7 kHz, input impedance of the simplified artificial ear is mainly resistive. In the case of real ears, however, the input impedance consists of resistance and compliance at these frequencies. The compliance component is due to eardrum and cavity of the ear canal. So the measured vent response with the simplified artificial ear differs from the vent response on the real ear. The vent response is defined as the difference between the pressure response with vent open and that with vent closed.

We have developed a predicting method of vent response of
the real ear using the measured vent response with the simplified artificial ear. We assume that source impedance of insert earphone is very high and earphone produces a constant volume velocity. Under these assumptions, the vent response is almost independent of the earphone characteristics and can be expressed simply by impedance ratio as shown in Fig. 5. Then, parallel vent response H is described by the following equation:

\[ H = \frac{P_V}{P_U} = \frac{Z_V}{Z_V + Z_{IN}} \]  

(2)

where \( Z_V \) denotes impedance of vent opening, and \( Z_{IN} \) input impedance of the artificial ear (or real ear).

If input impedance of the artificial ear \( Z_{IN} \) and vent response on the artificial ear H are given, we can calculate vent impedance \( Z_V \) from eq. (2). Therefore, we can predict vent response on real ear, by substituting the calculated vent impedance \( Z_V \) and input impedance of real ear \( Z_{INR} \) into eq. (2).

Fig. 6 shows vent response on real ear predicted by this method. Input impedance of real ear is calculated from the ear-drum impedance proposed by Shaw. For comparison, measured vent response with Zwislocki type ear simulator is shown in the same figure.

Conclusion

The simplified artificial ear is sufficient for practical use. And the combination of SAMRAI and this artificial ear gives simulated insertion gain of real ear. A predicting method of vent response on real ear has been developed. This method gives correct result for parallel vent, but incorrect result for side branch vent. For precise prediction of side branch vent, an additional coupler is required.

References


Okabe, K. HATS (SAMRAI) with simplified artificial ear.

**Fig. 1** Calculated transfer functions of artificial ear and real ear model.

**Fig. 2** Calculated transfer impedances of artificial ear and real ear model.

**Fig. 3** Simulated open ear gain of SAMRAI and real ear data.

**Fig. 4** Measured insertion gain of a behind-the-ear hearing aid on SAMRAI and 3 real ears.

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RESEARCH AND DEVELOPMENT OF PRIMARY ACOUSTICAL STANDARDS AND TECHNIQUES FOR IMPLEMENTATION OF TRACEABILITY IN CANADA

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This paper describes the development of Primary Acoustical Standards at the National Research Council of Canada (NRC). Implementing traceability of standard and measurement microphones depends on precision comparison methods.

Introduction: Acoustical standards based on reciprocity have been in use for over forty years. During this period, numerous authors contributed various improvements, until the uncertainty of the microphone sensitivity was reduced to the order of 0.03 to 0.05 dB. Currently, the procedure of absolute reciprocity pressure calibration of condenser microphones is governed by national and international standards [1,2]. The aim of this paper is to describe the development of Primary Acoustical Standards at the NRC and our implementation of traceability of standard and measurement microphones.

Primary Acoustical Standards: Inability to achieve a higher accuracy in the reciprocity calibration of condenser microphones is mainly due to difficulties involved with the assessment of the equivalent volume of the calibration cavity, which includes the equivalent volume of the microphones. Three corrections, as required by the standards, for heat conduction, wave pattern effects and the capillaries are applied to the equivalent volume; and the uncertainty of each of these corrections is of the order of 0.01 dB.

The development of a precision A.C. nulling method, and the determination of the equivalent volume acoustically under the same conditions as used for reciprocity calibration [3] enables us to eliminate the above numerical corrections and hence substantially reduces the uncertainty of calibration. The schematic arrangement shown in fig. 1 consists of two microphones coupled by a common cavity. Microphone (A) is driven by a signal source (E). The signal $e_1$ that is derived from the driving current $I_A$ of microphone (A) is compared with the signal $e_2$ which is from the receiver microphone (B). With a lock-in amplifier as the null detector, and a precision ratio-transformer as the attenuator, the ratio of the two signals can be obtained very precisely. With the switching arrangement shown by $S$, $e_1$ and $e_2$ are selected in turn as the reference signal to obtain a null condition. It can be shown that with three microphones one has:
\[ M^{c}_{A}M^{c}_{B} = \hat{V}_{AB} k_{0}^{-1} \hat{N}_{AB^*}, \quad M^{c}_{B}M^{c}_{C} = \hat{V}_{BC} k_{0}^{-1} \hat{N}_{BC}, \quad M^{c}_{C}M^{c}_{A} = \hat{V}_{CA} k_{0}^{-1} \hat{N}_{CA} \]  

(1)

where \( M^{c}_{A}, M^{c}_{B}, \) and \( M^{c}_{C} \) are the sensitivity of the microphones; \( \hat{V}_{AB}, \hat{V}_{BC} \) and \( \hat{V}_{CA} \) are the equivalent volumes of the cavity and microphone arrangement; \( \hat{N}_{AB}, \hat{N}_{BC} \) and \( \hat{N}_{CA} \) are functions of the attenuator readings; and \( k_{0} \) is a constant derived from parameters of the circuit shown and the environmental conditions. In general, with the aid of the insert voltage method, and from eqn (1), the open-circuit sensitivity of a microphone is given by:

\[ M^{c}_{O} = [(\hat{V}_{1}\hat{V}_{3}/\hat{V}_{2}) (\hat{N}_{2}/\hat{N}_{1}\hat{N}_{3})] k_{0}^{-1} \alpha_{R} S_{h} S_{p} S_{t} \]  

(2)

where \( \hat{V} \) and \( \hat{N} \) are functions of equivalent volumes and attenuator readings respectively; \( \alpha_{R} \) is a parameter obtained with the insert voltage method for the assessment of the open-circuit sensitivity; \( S_{h}, S_{p}, \) and \( S_{t} \) are corrections for humidity, polarization voltage and temperature.

Precise Assessment of the Equivalent Volume: The equivalent volume \( \hat{V}_{AB} \) of the cavity, Fig. 2, 3 includes the equivalent volume of the microphones \( \hat{V}_{PA} \) and \( \hat{V}_{AB} \). The centre portion \( V_{0} \) can be modified by means of substitution of various optically flat spacers, and the net change in thickness (hence the cavity volume) is determined precisely by interferometry. With the nulling arrangement shown in Fig. 1, the relationship of \( \hat{V}_{AB} \) and \( \hat{N}_{AB} \) is derived from the readings of the attenuator, and is illustrated in Fig. 3. The equivalent volume is given by:

\[ \hat{V}_{AB} = \Delta V_{0} [\hat{N}_{AB}/(\hat{N}_{AB2} - \hat{N}_{AB1})] \]  

(3)

where \( \Delta V_{0} \) is the small change in volume due to the small changes \( \pm \Delta X/2 \) of the spacer thicknesses.

Implementation: The acoustical cavity which is a nominal 3 cc plane wave coupler [1,2] and the microphones are housed inside a calibration chamber with precise control of temperature, humidity and pressure. The small nominal change in volume of \( \pm 81.5 \text{ mm}^{3} \) is implemented with two different thicknesses of optical spacers. The exact change in volume is calculated from the calibrated geometrical shapes of the spacers. The effective driver circuit for microphone (A) is identical to the preamplifier (B & K 2627) for microphone (B). The signal source (Fig. 1) is a synthesizer (HP 3325A). The dual-phase lock-in amplifier (EG & G 5206) operating in conjunction with the seven decade attenuator (ESI DT72A) provides an equivalent resolution of better than 0.0001 dB. A precision substitution method based on a.c. bridge techniques enables the measurement of the shunt capacitor C, in situ, which includes capacitance of the connecting cables, to approximately 100 ppm. The measurements of pressure, temperature, humidity, voltage, frequency and capacitance are traceable to the Canadian National Standards maintained at NRC.

Error Analysis: The uncertainty of the open-circuit sensitivity given in eqn (2) is:
\[ M_{0}^{\text{T}} = \left( \begin{array}{cc} a_{V}^{0} & a_{V}^{n} \\ a_{V}^{n} & a_{V}^{n} \end{array} \right) \left( \begin{array}{cc} V'_{n} + V'_{n} & N'_{n} \\ N'_{n} & N'_{n} \end{array} \right) + \left( \begin{array}{cc} a_{k}^{0} & a_{k}^{n} \\ a_{k}^{n} & a_{k}^{n} \end{array} \right) \left( \begin{array}{c} k'_{0} \\ k'_{0} \end{array} \right) + \left( \begin{array}{cc} a_{q}^{0} & a_{q}^{n} \\ a_{q}^{n} & a_{q}^{n} \end{array} \right) \left( \begin{array}{c} a_{R} \\ a_{R} \end{array} \right) + \left( \begin{array}{cc} a_{s}^{0} & a_{s}^{n} \\ a_{s}^{n} & a_{s}^{n} \end{array} \right) \left( \begin{array}{c} S_{t} + S_{t} \end{array} \right) \] (4)

where \( V'_{n}, N'_{n}, k'_{0}, \ldots, S_{t} \) are small perturbations of the corresponding parameters. Based on eqn. (4), it can be shown that the uncertainties of the calibration are less than 0.005 dB.

Experimental Results: The experiments were performed under conditions as close as possible to the standard reference condition of 20°C, 101.325 kPa and 65% R.H. The maximum deviations are within ± 0.05°C, ± 10 Pa and ± 25% R.H.

The stability of the overall system was investigated by monitoring the null reading of the lock-in amplifier after obtaining a null-condition with the seven-decade attenuator. Over a period of 1 hour, the overall stability of the system is better than 50 ppm.

Comparison Method of Microphone Calibration: The absolute method of reciprocity pressure calibration of microphones is relatively time consuming. For some applications, the following comparison calibration method developed at NRC is very attractive economically:

(1) A reference microphone with a known pressure sensitivity, is coupled to a pistonphone calibrator. The signal from the microphone is monitored with a precision measuring amplifier [4]. The sensitivity of the reference microphone is entered into the measuring amplifier via a multiturn potentiometer dial, which is calibrated in the format of microphone sensitivity (mV/Pa). A level reading is made using an external digital voltmeter which has a resolution of better than 0.01 dB.

(2) The reference microphone is replaced with the test microphone. The calibrated sensitivity dial is adjusted until the same reading is obtained with the monitoring voltmeter. The dial reading gives the sensitivity of the test microphone in mV/Pa.

The total time required for the above comparison method is of the order of minutes.

The repeatability of the dial readings is better than 0.01 dB, and it is estimated that the error of the calibrated sensitivity of the test microphone is less than 0.05 dB plus the sensitivity uncertainty of the reference microphone.

References
FIG. 1 Reciprocity Calibration

FIG. 2 Cavity for Primary Acoustical Standards

FIG. 3 Equivalent volume based on attenuator readings.

(1) (2) microphone holders, (3) spacer with capillary tubes (9),
(4) Quartz spacer, (5) (6) holders, (7) support
(8) screws, (10) non-conductive screws,
(11) adjustment cap (12) isolating and adjustment sleeve
DETECTION DE DÉFautS DANS LES POSTES BLInDES

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Dans le cadre d'une recherche générale pour améliorer l'efficacité des essais sur le site des postes blindés à isolation gazeuse, le Département "Transport-Appareillage" d'EDF a souhaité l'introduction de méthodes nouvelles faisant appel à des techniques acoustiques. Actuellement, les essais sur site consistent à détecter un défaut par un amorçage. L'objectif était de se doter d'un moyen de détection, et de localisation, plus sensible et présentant moins de risques pour l'installation. Des essais effectués au Laboratoire d'Essais Haute Tension d'EDF, aux Renardières, ont mis en évidence l'existence de défauts diminuant nettement la tenue diélectrique d'un poste en choc de foudre, mais pas en 50 Hz. Le choix d'un défaut caractéristique s'est porté sur une pointe très fine au voisinage de laquelle un effet couronne stabilisant se développe. Le but des campagnes d'essais entreprises a donc été de détecter les manifestations de cet effet couronne, notamment dans le domaine acoustique.

Simulation du défaut-type en caisson blindé

Pour mieux comprendre le problème, et limiter les contraintes inhérentes aux essais de type industriel, différentes étapes ont été successivement franchies. L'étude a débuté sur un élément de poste blindé de tension nominale 123 kV de petite taille, puis s'est poursuivie sur un élément industriel de tension nominale 800 kV. Le gaz de remplissage a été de l'air puis du SF6 sous une pression de 1 à 4 bars absolu.
Deux configurations d'électrodes ont été retenues, pointe-sphère ou sphère-plan. L'électrode haute tension, initialement constituée d'un cône a été remplacée par une hémisphère munie d'une pointe très fine simulant le défaut-type choisi.

Capteurs acoustiques et méthodes

Les capteurs ont été placés en deux endroits :

- sur la paroi externe du caisson, collés ou montés sur couplant : accéléromètres à réponse linéaire jusqu'à 40 kHz ou capteurs d'émission acoustique, résonnant à 200 kHz ou à large bande jusqu'à 1 MHz.

- à l'intérieur du caisson, en contact direct avec la masse du gaz, en montage affleurant : microphones ayant une réponse linéaire jusqu'à 40 kHz, partiellement noyés dans l'araldite, supportant un passage par le vide absolu. Cette technique présente des difficultés d'installation mais demeure néanmoins le meilleur moyen d'investigation expérimentale pour la recherche des voies de propagation des ondes émises par les décharges électriques et de leurs caractéristiques d'atténuation.

![Diagramme](image)

Fig. 2 - Acquisition des données électriques et acoustiques

Les signaux captés sont analysés en bande étroite ou stockés sur magnétophone. La technique du moyennage est indispensable, elle seule permet de s'affranchir des parasites impulsionnels d'origines diverses propres à l'ambiance industrielle, et éventuellement d'améliorer le rapport signal/bruit :

- moyennage temporel avec déclenchement fourni par le signal de synchronisation d'un photomultiplicateur ou le signal à 50 Hz, pour tout signal de caractère impulsionnel, pseudo-répétitif.

- moyennage fréquentiel, pour tout signal quasi-permanent, pour rechercher une zone de fréquences caractéristique.
**Interprétation du signal acoustique**

Les caractéristiques de l'effet couronne changent avec la nature du gaz, sa pression, la distance inter-électrode, la tension appliquée, la forme des électrodes, tous paramètres contrôlables qui permettent de moduler l'intensité et le nombre des décharges. Par contre, est incontôrable l'effet du vieillissement soit du gaz, soit de l'électrode. Les décharges partielles apparaissent d'abord sur l'alternance positive du 50 Hz. Lorsque l'on augmente la tension, la valeur crête du courant de décharge augmente ainsi que la durée du paquet de décharges constitué d'un nombre important d'impulsions très brèves et séparées par des intervalles de temps aléatoires. Ces décharges acquièrent vite un caractère de répétitivité par rapport au 50 Hz. Le signal issu des capteurs suit très fidèlement le phénomène physique : on observe un phénomène impulsion très bref, isolé, qui s'étend rapidement, d'apparition aléatoire, d'où la nécessité d'un moyennage temporel, puis la répétitivité s'accentue et le signal d'amplitude croissante perd son caractère impulsion d'autant plus rapidement que la vitesse du son est plus faible dans le gaz (140 m/s dans le SF6 au lieu de 340 m/s dans l'air), le moyennage fréquentiel devient alors utile.

Des niveaux de décharges très faibles ont été mesurés, tant en ce qui concerne le courant (quelques micro-ampères) que le niveau de charge équivalente (10 à 100 pico-Coulombs) :
- tous les types de capteurs réagissent, certains sont d'une mise en oeuvre plus immédiate, ne portant pas atteinte à l'intégrité de la structure donc préférables sur site.

![Graph](image_url)

**Fig. 3** : Signaux temporales captés dans l'air et le SF6.

![Graph](image_url)

**Fig. 4** : Evolution du signal avec l'amplitude de la d.p. (0-2-25-50 pc).
- en accélérométrie, la zone de fréquences utile se situe entre 3 kHz et 20 kHz, les niveaux augmentent avec le niveau des décharges.

- en émission acoustique, le résultat est médiocre : la zone intéressante de 20 à 100 kHz en air, est considérablement réduite dans le SF6, de par l'atténuation dans le gaz ou l'inexistance de ces fréquences à la source. Rien de significatif n'a été observé dans la gamme 100 kHz - 1 MHz.

- l'effet de taille du support d'essai, (caisson 123 ou 800 kV), ne modifie pas les observations faites.

- pour un niveau de décharges partielles donné, on constate une diminution de niveau du signal acoustique en fonction de la distance, d'autant plus grand que la fréquence est plus élevée.

- le niveau du signal est considérablement affaibli au passage d'un cône d'étanchéité séparant deux compartiments voisins.

Conclusions et perspectives

La détection d'un défaut-type, matérialisée par un effet couronne de très faible amplitude sur un point de la barre haute tension d'un compartiment blindé 800 kV, à isolation gazeuse (SF6), peut se faire par des moyens acoustiques sensibles. Le capteur le plus simple de mise en œuvre et le plus robuste est l'accéléromètre. Une localisation du défaut est également possible.

Désormais l'étude doit se poursuivre dans deux voies différentes. D'une part, il faut valider la méthode par des essais sur site d'exploitation, et la comparer aux méthodes classiques. D'autre part, il faut envisager une recherche plus fondamentale concernant la modélisation de la propagation du phénomène physique, ainsi qu'une expérimentation portant sur la transmission dans le gaz et les structures associées.
6.1

Transducteurs. Haut-parleurs.
Microphones

Transducers. Loud-speakers.
Microphones

Wandler. Lautsprecher.
Mikrofone
AN APPLICATION OF NONLINEAR INTERACTION OF SOUND WAVES TO THE LOUDSPEAKER

FUNDAMENTAL AND IMPROVEMENT

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Introduction

The phenomenon that is known as nonlinear interaction of sound waves, has already been applied into sonar in the field of underwater acoustics\(^1\).

On the other hand, the authors have investigated it for the application to loudspeaker\(^2\). This is the first time for this phenomenon to be used practically in the air. This type of loudspeaker consists of ultrasonic transducer array which can radiate a finite amplitude ultrasound beam, amplitude modulated by audio signal, into the air. This AM ultrasound wave is self-demodulated in the air by the nonlinearity of the air. This type of loudspeaker is quite different from ordinary loudspeaker which radiate sound from a vibrating diaphragm, and shows very sharp directivity pattern.

Acoustic reproduction theory by nonlinear interaction of finite amplitude ultrasonic in air

When a finite amplitude sound wave having composite spectra (primary wave) is radiated into the air, new sound waves (secondary waves) having new spectrum construction may be produced as the result. This phenomenon is well known as the nonlinear interaction of sound waves\(^3\).

The sound pressure \(p_2\) of secondary wave produced by this phenomenon, was first derived by Westervelt as the solution of the following equation\(^4\).

\[
\begin{align*}
\rho \frac{\partial^2 p}{\partial t^2} - \frac{1}{c_o^2} \frac{\partial^2 p}{\partial x^2} &= -\rho \frac{\partial g}{\partial t} \\
\frac{\partial^2 g}{\partial t^2} &= \frac{\beta}{\rho^2 c_o^4} \frac{\partial}{\partial t} p^2
\end{align*}
\]

where, \(c_o\): small signal sound velocity, \(\rho\): density of fluid
\( \beta \): nonlinear parameter of fluid and \( p_1 \): primary wave sound pressure.

Now, a finite amplitude AM ultrasonic plane wave shown in eq. (3) is considered as the primary wave.

\[
  p = p_0 \left[ 1 + m \cdot g(t - \frac{r}{c_0}) \right] e^{-\alpha x} \sin \omega_b(t - \frac{r}{c_0})
\]  

(3)

Where, \( p_0 \): initial sound pressure of carrier sound  
\( m \): a parameter indicating modulation index  
\( \alpha \): absorption coefficient of carrier sound  
\( g(t) \): modulation signal (audio signal).

In this case, the carrier component and the side band component interact nonlinearly each other, and as the result, the AM ultrasound is self-demodulated in the air.

If a carrier plane wave with radius \( a \) is assumed to be radiated from the transducer array, the sound pressure of the secondary wave at the point \( r \) distant from the array on the axis, can be calculated approximately. In the case of sinusoidal modulation, we get,

\[
  \left\{ \begin{array}{l}
  p_s = -\frac{\beta p_0^2 a^2 m w^2}{8 c_0^4 c_0^2 \alpha r} \sin \omega(t - \frac{r}{c_0}) \\
  p_d = \frac{\beta p_0^2 a^2 m w^2}{8 c_0^4 c_0^2 \alpha r} \cos 2\omega(t - \frac{r}{c_0})
  \end{array} \right.
\]

(4)

(5)

In the above equation, \( p_s \) and \( p_d \) show the signal sound pressure and 2nd harmonic distortion sound pressure, respectively.

Accordingly, it is possible to construct a new type of loudspeaker if the modulation signal is selected as the programed audio signal. Figure 1 shows the signal sound pressure based on the calculation of eq(4).

By eq. (4) and (5), it is clear that the signal sound pressure is proportional to \( m \) and the second harmonic distortion sound pressure is proportional to \( m \). Therefore a good distortion ratio requires a very small \( m \) to prevent cross-interaction between the lower and the upper side band waves. The second harmonic distortion ratio \( \varepsilon \) can be expressed as

\[
  \varepsilon = \frac{|p_d|}{|p_s|} \times 100 \% = m \times 100 \%
\]

(6)

**Experiments**

The loudspeaker was developed by using 547 pieces of PZT bimorph transducers. The fundamental resonant frequency of a transducer is about 40kHz and the sound pressure level is 100 dB at the point 1m from it on axis.

A front view of the loudspeaker appears in Fig.2. The sound pressure frequency response characteristics of the primary wave and the directivity pattern at 40kHz of
the loudspeaker are shown in Fig.3 and 4, respectively.

Figures 5 and 6 show the frequency characteristics and the directivity pattern of the secondary wave, respectively. In this case, the input voltage to the loudspeaker was 10V r=4m and $m=0.5$. It is clear from Fig.6 that the secondary wave (signal wave) has very sharp directivity. The signal and 2nd harmonic distortion components of the secondary wave were analyzed by spectrum analyzer to check the distortion ratio. These results are shown in Fig.7 about three different $m$.

Discussion and Improvement

The most important feature of this type of loudspeaker is extreme sharpness of the directivity. No ordinary type of loudspeaker has ever had such a sharp directivity. However, to put to practical use there are many problems as follows,

1) Electrical power efficiency, 2) Sound pressure increase
3) Distortion ratio improvement 4) Flattening of frequency characteristics 5) Primary wave interception

One of the effective methods for decreasing 2nd harmonic distortion is to use single side band (SSB) modulation. As interaction between lower and upper side band is avoidable by using SSB, it is possible to perform low distortion reproduction. When SSB modulation is used, a deep modulation (large m) can be available without increasing harmonic distortions. This is excellent feature of SSB modulation. A experimental result is shown in Fig.8. The figure shows that the 2nd harmonic distortion sound pressure yielded by SSB modulation is about 10dB lower than that yielded by normal AM modulation under the same conditions.

References
2) M. Yoneyama et. al.: J. Acoust. Soc. Am. (will be published)

Figures

Fig.1

Fig.2

Fig.3
Fig. 4

Fig. 5

Fig. 6

2kHz

5kHz

Fig. 7

m = 1.0

m = 0.5

m = 0.3

Fig. 8

Signal

Distortion

Normal AM

35dB
ON THE DIGITIZATION OF MICROPHONE

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Introduction

The application of digital technology is now in progress in the wide range of audio engineering and telecommunication engineering fields [1,2]. The PCM taperecorder was firstly introduced to the audio engineering field in order to prevent signal deterioration which we often meet in analog taperecorders. In spite of the great merits in digital technique, all digitalization of audio systems including microphones and loudspeakers has not been put into developing stage. Among the various reasons, the dominant and straight one is that the analog-to-digital converter with 14 to 16 bits, which is necessary for high fidelity signal transmission, is very expensive today.

At the beginning of the developing the digital microphone, we considered the two methods of digitalization of the microphone. The first method is to try to realize a digital microphone whose transducer output is directly in digital form. The second method is to realize a digital microphone as a combination of electronic analog-to-digital converter and microphone of usual type. The first method has many problems in principle and is not in the stage of engineering development. Therefore we take the second method. The technical problems in the second method is to develop cheap and small analog-to-digital converter and acoustic antialiasing filter.

Purpose of Digital Microphone

The merits expected from the digitalization of microphones are:
(a) Reduction of wire cables connecting between many microphones and a mixing machine. This is possible by digital signal multiplexing technique.
(b) The addressing of microphones from mixing machine by digital technique. This makes mixing operations easy.
(c) Suppression of noise caused by interference signal or induction which may exist in the microphone cables.
(d) Digital signal processing applied to a microphone and microphone array. This brings us various kinds of special effect including sharp directivity.

A propos de la numérisation de microphone
Requirements for the Microphone Developed

(a) The required items to the digital microphone are as follows:
(b) The cost of the microphone should be less than sum of costs of both current professional microphone and analog-to-digital converter of high resolution type.
(c) The performances of digital microphone such as sensitivity, dynamic range, frequency response and directional characteristics should be of the same grade as those of the current professional microphones.

Design Policy of Digital Microphone

(a) An antialiasing low-pass filter indispensable for digital microphone should be of the order of more than nine. Such a high order filter will be large and expensive if we construct it with electronic elements only. Therefore we decide to develop it with tandem connection of an acoustic filter of order five and electronic active filter of order four.
(b) As the dynamic range of microphone is required 100dB, more than 16 bits resolution is necessary for analog-to-digital converter. On the other hand, 16 bit analog-to-digital converter is so expensive that it is not suitable to digital microphone. Therefore, we try to use two 12 bit analog-to-digital converter modules in parallel and in floating point type instead of a single 16 bit analog-to-digital converter module. Such a construction of analog-to-digital converter allows us to get digital microphone with wide dynamic range in cheaper price.
(c) An omni-directional electret condenser microphone is used as a transducer of the digital microphone.

The specifications of the analog-to-digital converter used for the digital microphone are as follows:

- Dynamic range : 96dB (This value corresponds to common 16 bit analog-to-digital converter.)
- Maximum value of signal-to-noise ratio : 79.8dB (this value corresponds to common 13 bit analog-to-digital converter.)
- Sampling frequency : 44.056 kHz
- Conversion time : 10 μsec.

The block diagram of the digital microphone developed is depicted in Fig. 1.

![Fig. 1 Block diagram of the digital microphone](image-url)
Experimental Results

The performance of the digital microphone developed are as follows:

(a) Directional characteristics: Omni-directional
(b) Frequency response: Shown in Fig. 2
(c) Sensitivity: -60 dB (0 dB = 1V/μbar)
(d) Dynamic range: 96 dB
(e) Quantization noise: -71 dB (0 dB = 1V)
(f) Distortion: 0.028%

Fig. 2 Frequency Characteristics of the Digital Microphone

As for the residual noise, the experimental results are shown in Fig. 3. The noise-to-signal ratio v.s. input level characteristics of the floating type analog-to-digital converter used in the digital microphone is nearly same as that of 13 bit linear analog-to-digital converter for the input level ranging from 0 to -24 dB and that of 15 bit linear analog-to-digital converter for the input level less than -24 dB, which corresponds to the switching level from upper 12 bit analog-to-digital converter to lower one.

Fig. 3 Measured noise (Distortion) characteristics of various A/D converters.

Fig. 4 Spectrum of recovered analog signal from digital output of the microphone.
The output digital signal of the floating type analog-to-digital converter is converted into analog signal through 16 bit digital-to-analog converter. An example of its spectrum for input signal of 1 kHz is shown in Fig. 4. The distortion which is inherently caused by switching the input is shown to be negligible.

Fig. 5 The outlook of the digital microphone

Conclusion

The specifications are satisfied in our experimentally developed model except the requirement of dynamic range. Namely, the dynamic range obtained is about 1.5 bit lower than the designed value for the low level range of input signal. This is caused by noise from sample and hold stage of analog-to-digital converter.

References


AN INTEGRATED SILICON-ELECTRET-CONDENSOR MICROPHONE

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I. INTRODUCTION

The increasing use of integrated electronics, in particular of micro-
computers, caused a growing demand for sensors to convert various physical
quantities into electrical signals. In the past few years several silicon
sensors have been developed which can be produced by means of modern semi-
conductor fabrication technologies [1,2], e.g. a gas chromatographic air
analyzer [3], pressure sensors [4,5,6,7] and accelerometers [8,9,10]. In
the present paper, we suggest the construction of an integrated electrct
condensor microphone. This transducer could be built on the same silicon
chip as the impedance-converter FET and the preamplifier. The diaphragm
of the microphone is a thin plate of silicon dioxide which also represents
the electret.

II. TRANSDUCER CONSTRUCTION

A cross sectional view of the proposed transducer is shown in Fig. 1. The
p-type silicon wafer is first heavily boron doped (surface concentration
10^{25} \text{ cm}^{-3}, diffusion depth 10 \text{ m}), excluding a small area of 20 \times 20 \text{ m}^2 to
be converted into a connecting hole. Then an epitactical layer (thickness
20 \text{ m}) is grown on the p\text{*} silicon. Next, the thermal oxidation of the epi-
silicon and the backside of the wafer (1 \text{ m oxide}), and the opening of the
windows in the oxide mask for the orientation dependent etching (ODE) are
carried out. Fig. 2 shows the resulting shape of the apertures. Etching with
a PEO-mixture (pyrocatechol, ethylene diamine and water)
is performed through these
apertures and from the back-
side of the wafer, resulting
in a 20 \text{ m} air gap and a con-
necting hole to the back chamber.

Assuming a transducer area
of 0.5 \times 0.5 \text{ m}^2, and with the
etch rates from ref. 11 to 13,
the etching procedure re-
quires more than 12 hours for
a complete underetching of
the silicon dioxide plate.
For such long etching times

Fig. 1: Cross sectional view of the proposed transducer.
the decrease of the thickness of the oxide may not be neglected. The reduced thickness of the plate is given by

$$d^* = d_{ox} - 2R_{SiO_x} \cdot t_e$$  \hspace{1cm} (1)$$

where $d$ is the thickness, $R$ is the etch rate and $t$ is the etching time. The term $2R\cdot t$ equals about 0.36 $\mu$m after 12 hours.

After the ODE procedure the etch apertures on the oxide plate have to be closed, e.g. with epoxy paste, and the surface of the plate is metallized with aluminum.

III. MECHANICAL AND ACOUSTICAL PROPERTIES

The partial differential equation for the free vibration of a rectangular plate is given by

$$\frac{\partial^2 u}{\partial x^2} + 2\frac{\partial^2 u}{\partial x \partial y} + \frac{\partial^2 u}{\partial y^2} + \frac{\rho d}{N} \frac{\partial^2 u}{\partial t^2} = 0,$$

where $x$ and $y$ are the normalized coordinates, $u$ is the normalized deflection, $\rho$ the density, $d$ the thickness, and $N$ the stiffness of the plate. $c$ is a normalization constant. The mechanical stress originating from the differing thermal-expansion coefficients of silicon and silicon dioxide is assumed to be negligible. The boundary conditions for a rectangular plate with all sides clamped are

$$u = 0 \text{ for } x = 0, x = 1, y = 0, y = 1$$
$$\frac{\partial u}{\partial x} = 0 \text{ for } x = 0, x = 1; \frac{\partial u}{\partial y} = 0 \text{ for } y = 0, y = 1$$  \hspace{1cm} (3)$$

There is no analytical solution of eq. 2 that can satisfy the boundary conditions (3). However, an approximate value of the first eigenfrequency of a square plate can be used\cite{15}:

$$f_{11} = \frac{c_{11}}{2\pi a} \sqrt{\frac{E \cdot d^2}{12 \rho (1-\sigma^2)}}$$  \hspace{1cm} (4)$$

where $c_{11} = 36$ \cite{15} is a constant, $E$ and $\sigma$ are Young's modulus and Poisson's ratio, respectively, and $a$ is the length of the plate. The thickness $d$ must be replaced by the reduced thickness $d^*$ (eq. 1). Since the thickness reduction is not uniform over the whole plate area, the remaining oxide is taken into consideration by an enlarged density $\rho^*$:

$$\rho^* = \rho(5 + d/d^*)/6$$  \hspace{1cm} (5)$$

Eq. 4 with $\rho = \rho^*$ and $d = d^*$ yields for the fundamental resonance frequency of a square silicon dioxide plate with $a = 0.5$ mm and $d = 1 \mu$m $f_{11} = 22.4$ kHz. The analogous electrical circuit of the transducer is shown in Fig. 3.
Inspection of the numerical solution of the one-dimensional problem (beam with two sides clamped) shows that below the first eigenfrequency the plate can approximately be treated as a simple mass-spring system, represented by $L_1$ and $C_1$. Capacitors $C_2$, $C_3$, and $C_4$ are due to the stiffness of the air in the air gap, the backchamber and the case, respectively. Inductance $L_2$ and resistance $R$ take the constriction into account. From the analogous circuit the effective deflection of the plate at frequencies from 20 Hz to 15 kHz and a sound pressure level of 1 Pa is found to be between 1.1 and 1.3 mm, and the resulting microphone sensitivity follows as approximately 11 mV/Pa (1000 Hz, assumed field: 10 MV/m).

**IV. OXIDE CHARGING**

For the investigations of the charge build-up in SiO$_2$ we used MOS structures. The silicon wafers (p-type, 3-5 μcm) were wet oxidized at 1100°C without applying a post-oxidation or post-metallization annealing step.

For high frequency avalanche injection [16] the aluminum layer on the oxide (oxide thickness 154 nm) is etched away leaving only small pads of 0.5 mm diameter. The flatband voltage shift of the MOS capacitors during the charging process is measured with a circuit similar to the one described in ref. 16. In order to avoid the so-called turn-around effect which is caused by positive charges located at the Si-SiO$_2$ interface [17], the capacitance vs. voltage curves [18] were measured at elevated temperatures (100°C). Fig. 4 shows the flatband voltage shift vs. current-time product $F_A$.

We reached voltage shifts of about 10.9 V, corresponding to projected charge densities of $2.4 \times 10^3$ C/m$^2$ at the Si-SiO$_2$ interface. The maximum values reported by Feigl et al. [19] for water diffused oxides are about $5 \times 10^3$ C/m$^2$.

Although high charge densities can be achieved, the avalanche-current charging method does not seem to be the most suitable one for charging SiO$_2$ electrets, since the rather long charging times (>2 h) and the need to charge each sample separately is not
suited for batch fabrication.

An alternative method is corona charging of electrets. This technique was already used to evaluate the break-down fields of silicon dioxide [20]. We charged thermally grown SiO₂ (thickness 1.1 μm) and measured the surface potential with a vibrating capacitor technique. Fig. 5 shows the temporal stability of the surface potential. The original potential of -165 V, corresponding to an equivalent surface charge density of $5.1 \times 10^{-9} \text{C/m}^2$, was stable for a time period of 1 year, no decay of the charge could be detected. This rather high charge density value well compares with values obtained on Teflon FEP foils [21] and suggests the use of SiO₂ as electret for conventional back plate electret condensor microphones. The corona charging method cannot be applied to the transducer of Fig. 1 because the bottom of the oxide plate is never open to the environment, not even during the fabrication process. Thus an electron beam charging method may be more suitable.

V. CONCLUSIONS

The present paper shows that it should be possible to build an integrated electret condensor microphone on a silicon wafer by use of integrated-circuit technologies. The microphone sensitivity that can be expected is sufficient for most purposes. The charge density achievable on silicon dioxide electrets by corona charging or by avalanche injection of electrons is sufficiently high and suggests applications other than in microphones.

VI. REFERENCES

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ON THE STUDY OF SIGNAL-NOISE RATIO OF MINIATURE ELECTRET
CONDENSER MICROPHONE

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INTRODUCTION
The miniature electret condenser microphone for hearing-aid which already developed in our factory, has been improved in signal-noise ratio for high grade professional and broadcasting use.

The results of manufacturing in trial of electret condenser microphone which is back-electret type are described in this paper.

(see Fig.1 Illustration of the construction of miniature electret condenser microphone and Fig.2 Acoustical equivalent network of this microphone.)
LIST OF SYMBOLS

\( m_0, r_0, s_0 \): acoustic effective mass, resistance and stiffness of the diaphragm

\( p \): sound pressure

\( S \): effective area of the diaphragm

\( s_i \): acoustic effective stiffness of back cavity

\( P_{nE(A)}, P_{nM(A)} \): equivalent sound pressure level of electrical noise and mechanical thermal noise

\( d_i \): gap between the diaphragm and the back electrode

\( \varepsilon \): dielectric constant of air

\( r \): ratio between the area of the back electrode and effective area of the diaphragm

\( R_{i1}, R_{i2} \): input and output equivalent noise resistance of circuit

\( B_i, B_i, B_e \): band width

\( \mu \): stability of the diaphragm

\( C_s \): stray capacitance of the gap

\( \omega_0, Q_0 \): angle-frequency and damping factor of the higher limit

THEORETICAL DESIGN

This miniature microphone was designed in non-directional, so, the open terminal voltage sensitivity \( E_o(\omega') \) are

\[
\frac{E_o(\omega')}{P} = \frac{d_o S}{(A_o + A_i) E M Y} 
\]  

(1)

If \( S, d_o, A_i, Y \) are constant, the maximum conditions of \( E_o(\omega') \) are

\( S_o = S_1 \)  

(2)

The equivalent noise sound pressure level \( P_{n(A)} \) are

\[
P_{n(A)} = \sqrt{P_{nE(A)}^2 + P_{nM(A)}^2} 
\]  

(3)

\[
P_{nE(A)} = \frac{4 \pi T M d_o^3 (d_o + d_i)^2}{\varepsilon Y S^3 A_o} \left[ \frac{1}{R_{i1} B_1} + R_{i2} B_2 \left( \frac{E Y S d_o}{\varepsilon M Y} + C_s \right) \right] 
\]  

(4)

\[
P_{nM(A)} = \frac{4 \pi T B_o \omega_1 (1 + \omega_1 \omega_0)}{S^2 \omega_0 Q_0} 
\]  

(5)

If \( s_1 \) is constant, the minimum conditions of \( P_{nE(A)} \) are

\( s_o = s_1 \)  

(6)
This condition equals to the one of the maximum conditions of $E_d(0^*)$. Accordingly,

$$P_{n,E(A)} = \frac{\frac{1}{2} \frac{\alpha T M}{eYS^3}}{1 + R_{in} \frac{R_1}{R_2}} \frac{1}{1 + \frac{E\gamma S}{d_b C_S}}$$

(see Fig.3 Relation between $P_{n,E(A)}$ and $s_0 / s_1$.)

On the other, the minimum conditions of $P_{n,H(A)}$ are

$$s_0 \ll s_1$$

(8)

In case of this microphone, this one's $P_{n,H(A)}$ is a little larger than $P_{n,E(A)}$, so, to improve signal-noise ratio, it must be minimize $s_0$ than $s_1$. But, from Fig.3, it is in substance reasonable to minimize $s_0$ 1/2 $\sim$ 1/4 than $s_1$.

**EXPERIMENTAL RESULT**

The structural points of this microphone as follows:

1. Back-electret type
2. Thin diaphragm of titanium (1μm)
3. Using field effect transistor with high Ioss

(see Fig.4 Circuit of this microphone)

The characteristics of the microphone are as follows:

1. Sensitivity ; $-54 \sim -52$ dB (0dB = 1V/μbar)
2. Equivalent noise ; $22 \sim 24$ dB SPL
3. Dynamic range ; 93 dB

![Fig.3 Relation between $P_{n,E(A)}$ and $s_0 / s_1$](chart1)

![Fig.4 Circuit of this microphone](chart2)

$V_{cc} = 9$ V

$R_L = 3.3$ kΩ
The frequency characteristics of this microphone is shown in Fig. 5.

The characteristics of this miniature microphone are very close to results of theoretical design.

Fig. 5 The frequency characteristics of this microphone

CONCLUSION

On the basis of these results, it has been cleared possibility of detail design and production of the electret condenser microphone for high grade professional and broadcasting use.
INFLUENCE OF DIFFERENT FORMS OF LOUDSPEAKER SYSTEM PHASE RESPONSE ON SUBJECTIVE EVALUATION OF REPRODUCTION

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Introduction

The audibility of the phase distortion has been the subject of discussion for many years. Some authors find it is not sufficient to be content with a flat amplitude response and low harmonic and intermodulation distortion but also with low phase distortion. They believe phase response is as important as amplitude response, while others neglect the importance of this response.

The testing methods used by previous experimenters are very different and there are numerous qualitative and quantitative interpretations [1-5]. But still it cannot be reliably said what is the amount of change of the loudspeaker system phase response which can be felt by man.

Today there are commercially available loudspeaker systems with axially shifted loudspeakers to obtain linear phase response. This results in a good phase response, transient response and consequently good quality of the reproduced sound. However, there remains a question whether all these improvements can be felt by man. This paper shows the influence of different forms of loudspeaker system phase response on subjective evaluation of speech and music.

Loudspeaker system for testing

In order to determine the ability of the subjects to discern modification phase responses, a two-way loudspeaker system with the third-order Butterworth crossover network is chosen. This crossover network has flat amplitude response whether the loudspeakers are connected in phase or out of phase, but group delay response is being very modified at the crossover frequency and round it [6]. This characteristic of the crossover network is employed to change system phase response with the same loudspeakers without changing the amplitude response.

The measured amplitude and phase responses of the designed loudspeaker systems when loudspeakers are connected out of phase (denoted with A) and in phase (denoted with B) are given in Fig. 1. The responses of another loudspeaker system (denoted with C) which is designed so that the tweeter is axially shifted in relation to the woofer in system A are also shown in the same figure. This system has similar amplitude response to the mentioned one but the phase response is quite changed. These
characteristics relate to the axial response, 1 m from the system, in an anechoic chamber.

The amplitude responses for all the three systems are similar so it is assumed that small changes will have no influence on reproduction. The phase responses for all of them are very different. The phase response of the system A gradually decreases, that of the system B very rapidly decreases around 2 kHz, and that of the system C has a rapid bent in the

Fig. 1 Measured amplitude and phase responses of the designed loudspeaker systems A, B and C
same narrow region as system B. These modifications of phase responses are within the relatively narrow frequency band where the listener's ear is very sensitive.

Subjective evaluation

The listeners chosen for subjective evaluation were selected based on the results of the previous tests in which they had showed high reliability. All the listeners were divided into two groups: musically educated listeners (I group) and musically perceptive listeners (II group). There were from 15 to 20 listeners in each group.

The evaluation began in an anechoic chamber where some differences in reproduction were noticed, and then it was continued in a sitting room (3.5 x 5 x 2.4 m, the reverberation time about 0.5 s).

The evaluation was always done with only one listener, his head in the axis in front of the loudspeaker system, at 1 m and 2.5 m distance from the system. In this way all the listeners were in the same position during the evaluation. The experimenter (the author) quickly changed from one system to the other, giving a visual sign that he was doing it, making the comparison possible at the most convinient passages in the music. The program material was speech by male speaker, drums, popular music (PM) and piano solo.

Before the evaluation began, each listener was informed about the method of the work and the demonstration was done. The listener listened to the program material successively about 5 s from one system and then from the other, several times. The listener had to point out which system gave a better reproduction. The expression "better" was not predefined, it was a subjective opinion of the listener. If the listener couldn't define which one was "better" he had to put a sign "=" (undefined reply). During the evaluation breaks were made for the listener to write down results, the impressions or comments on the reproduction.

Analysis of results

The influence of different forms of loudspeaker system phasis response on subjective evaluation of reproduction noticed in an anechoic chamber, was less noticeable in the sitting room, especially at greater distances from the loudspeaker systems (Table 1). The evaluation results showed too many undefined replies indicating that the perception of changing the program material was neglectable, and that those, slightest changes in sound quality could be heard only by a very sensitive ear, yet not always capable of deciding which system was better.

Nevertheless, the system A has proved to be slightly better comparing to other two systems. Some listeners commented that system A gave a slightly clearer reproduction which sounded better to them for certain kind of the program material, but not for all.

Conclusion

The results of the subjective evaluation show that different forms of loudspeaker system phase responses (being either phase response gradually decreased, or response very rapidly decreased in the narrow region, or response having rapid bent) produce neglecting changes in the program
Table 1 The results of the subjective evaluation (%)  

<table>
<thead>
<tr>
<th>Group</th>
<th>Speech</th>
<th>Drums</th>
<th>PM</th>
<th>Piano</th>
<th>Total</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>I  2.5</td>
<td>II  1.25</td>
<td>I  2.5</td>
<td>II  1.25</td>
<td>I  2.5</td>
</tr>
<tr>
<td>A</td>
<td>22 36 33 31</td>
<td>62 21 28 23</td>
<td>16  9  39  46</td>
<td>11  23 17  36</td>
<td>14  23 29  35</td>
</tr>
<tr>
<td>B</td>
<td>22 21 17 15</td>
<td>22 29 28 15</td>
<td>6  33 17 27</td>
<td>50  54 33 21</td>
<td>25  34 24 20</td>
</tr>
<tr>
<td>C</td>
<td>30 47 50 82</td>
<td>27 36 45 72</td>
<td>56  56 24 82</td>
<td>44  61 82 72</td>
<td>39  50 50 77</td>
</tr>
<tr>
<td>D</td>
<td>33 50 27 27</td>
<td>37 56 33 22</td>
<td>67  56 18 44</td>
<td>44  56 10 44</td>
<td>46  54 23 34</td>
</tr>
<tr>
<td>E</td>
<td>56 12 36 36</td>
<td>50 11 33 56</td>
<td>22  0  45  0</td>
<td>44  11 30 11</td>
<td>43  9  37 26</td>
</tr>
</tbody>
</table>

Material quality, especially in the sitting room. These slight changes in the sound quality can be heard only by the listeners with a very sensitive ear. The phase response is important only if a very high-fidelity reproduction is wanted.

Acknowledgment

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References

AN ANALYSIS OF LOUDSPEAKER TRANSIENT DISTORTION IN THE COMPLEX FREQUENCY DOMAIN

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Introduction

Any linear network or electro-acoustic transducer (assuming its nonlinear distortion is very low and can be ignored) can cause a kind of distortion called transient distortion (T.D.) when it is used to transfer some transient signals, such as speech and music signals whose amplitudes alter rapidly. More precisely speaking, any system whose transfer function $G(S)$ has one or more poles will cause T.D.

For instance, when a system is excited by an input signal

$$X(t) = u(t)\sin \omega_f t,$$

where

$$u(t) = \begin{cases} 0, & \text{when } t \leq 0, \\ 1, & \text{when } t > 0, \end{cases}$$

its response (output) will be

$$y(t) = y_f(t) + y_n(t),$$

where $y_f(t)$ is its forced response or steady-state response, and $y_n(t)$ is its natural response and

$$y_n(t) = \sum_{i=1}^{k} y_i(t),$$

where

$$y_i(t) = u(t)A_i \exp(-R_i t)\sin(\omega_i t + \phi_i),$$

and $A_i, R_i, \phi, \phi_i$ are constants. $R_i$ and $\omega_i$ are determined by $S_i$, the $i$'th pole pair of its $G(S)$ through $S_i = -R_i \pm j\omega_i$ and $k$ is the pole pair number. For the stable systems which will only be discussed here, $R_i > 0$, i.e. all $y_f(t)$ terms represent an attenuation. However, during the first period of time after the signal has been applied, they will superimpose the forced response $y_f(t)$. Thus the T.D. will affect the sound reproduction quality of loudspeakers, and its measurement and description have become an interesting topic today.
Review

For a long time the tone burst and square waveform signals have been used to examine T.D. of loudspeakers. But these can only give some qualitative knowledge about it. Nowadays, Frequency, Time and Amplitude Three Dimension Tone Burst Response and Cumulative Spectra have been used to describe T.D. in some detail. In this paper, another method is proposed for an analysis of T.D. in the complex frequency domain.

Analysis of T.D. in complex frequency domain

If G(S) of a system is given, its response y_f(t) and y_i(t) for the special input signal u(t)sin\( \omega_f t \) can be calculated by complex frequency transform.

Now, let
\[
\begin{align*}
  y_f(t) &= \frac{1}{A_1} y_i(t) = u(t) \sin(\omega_f t + \phi), \\
  y_i(t) &= \frac{A}{A_1} \exp(-R_i t) \sin(\omega_i t + \phi_i), \\
  y_n(t) &= \sum_{i=1}^{k} y_i(t) \quad \text{and} \\
  Y(t) &= Y_f(t) + Y_n(t).
\end{align*}
\]

We further define the quantities
\[
\begin{align*}
  P_i(\omega) &= |Y_i(\omega)| \quad \text{and} \\
  P_n(\omega) &= |Y_n(\omega)|,
\end{align*}
\]

where \(|Y_i(\omega)|\) and \(|Y_n(\omega)|\) are the spectrum modes of \(y_i(t)\) and \(y_n(t)\), respectively.

The \(y_f(t)\) and \(y_i(t)\) will be used to describe T.D. in time domain and the \(P_i(\omega)\) and \(P_n(\omega)\) can be used to indicate T.D. in frequency domain.

Analysis example

As an example, a loudspeaker mounted in an infinite baffle and operating at low frequencies is considered. Its G(S) can be deduced from its approximate analogous circuit, Fig. 1 and

\[
G(S) = \frac{\frac{M Ca}{\omega_o} S^2}{S^2 + \frac{\omega_o}{Q} S + \omega_o^2},
\]

where
\[
\omega_0 = \frac{1}{\sqrt{(M + M_a)C_a}} \quad \dot Q = \frac{\omega_0 (M + M_a)}{R_a},
\]

\(R, M, \) and \(C, \) are the effective acoustic resistance, mass and compliance of the loudspeaker. The free-field sound pressure \(p\) at a distance \(r\) from the loudspeaker can be determined as the pressure across an reactance \(M\) given by \(M = \rho/2\pi r = 0.188/r,\) \(\rho\) being the density of air. If \(r = 1(\text{m})\) is considered, \(M = 0.188.\) \(R\) is the equivalent value of the pressure on the blocked loudspeaker diaphragm resulting from the driving force. For large woofers, \(M << M_a\) can be assumed.

Obviously, this \(G(S)\) has only one pole pair \(S_1\) and

\[S_1 = -R_1 \pm j\omega_1,\]

where

\[\omega_1 = \sqrt{\omega_0^2 - \omega_0^2}, \quad R_1 = \frac{\omega_0}{2Q},\]

if let \(P_{ea} = X(t) = u(t)\sin \omega_0 t\) and \(p = y(t),\)

\[Y_f(t) = u(t)\sin(\omega_0 t+\phi),\]

\[Y_1(t) = u(t) A_1 \exp(-R_1 t)\sin(\omega_1 t+\phi_1),\]

\[P_1(\omega) = \frac{1}{A} \sqrt{(q/u-r/v)^2+(z/u+w/v)^2},\]

where

\[A_1 = 2a^2b^2, \quad A = 2c^2d^2, \quad \phi_1 = \text{arc sin} \frac{2a}{A_1}, \quad \phi = \text{arc sin} \frac{2c}{A},\]

\[q = b(\omega+\omega_1)+ar_1, \quad r = b(\omega-\omega_1)-ar_1, \quad z = a(\omega+\omega_1)-br_1,\]

\[w = a(\omega-\omega_1)+br_1, \quad u = (\omega+\omega_1)^2+r_1^2, \quad v = (\omega-\omega_1)^2+r_1^2,\]

where

\[a = \frac{ek-qh}{h^2+k^2}, \quad b = \frac{eh+tk}{h^2+k^2}, \quad c = \frac{em}{m^2+n^2}, \quad d = \frac{-1m}{m^2+n^2},\]

where

\[e = MC \sqrt{\omega^2_0 (R_2^2-\omega_2^2)}, \quad g = 2MC R_1 \omega_0^2 \omega_1, \quad h = 2\omega_0 (R_1^2+\omega_2^2-\omega_2^2),\]

\[k = 4R_1 \omega_0^2, \quad l = MC \sqrt{\omega_2^2}, \quad m = 4R_1 \omega_1, \quad n = 2(R_1^2-\omega_2^2+\omega_2^2).\]

Here, \(Y_n(t) = Y_1(t)\) and \(P_n(\omega) = P_1(\omega).\)

The T.D. for loudspeaker parameters \(F = \omega_0/2\pi = 54 (\text{Hz}),\)

\(C_e = 3.8 \times 10^{-17}(\text{m}^3/\text{Pa})\) and \(Q = 3\) and \(\dot Q\) expressed by its \(Y_1(t)\) and \(P_1(\omega)\)

curves have been calculated and are shown in Fig. 2.

Discussion

In this paper, a simple theoretical analysis method of loudspeaker T.D. in complex frequency domain is given and an example of low frequency T.D. of a loudspeaker is shown. From the results it can be seen that the amplitude of the interference spectrum \(P_n(\omega)\) depends very much on the \(Q\)-factor, and its peak frequency is \(F = \omega_0/2\pi\). When \(Q\) decreases, not only \(P_1(\omega)\) will be diminished rapidly, but also the peak frequency \(F\) will be lower. This means that the lower its \(Q,\) the less its T.D. This conclusion is quite in agree-
ment with the experience that a loudspeaker with a low $Q$-factor will produce high quality sound.

At high frequencies the loudspeaker cone will break up in different vibrating modes and phases, and an analogous circuit cannot be derived. Therefore, its $G(S)$ cannot be obtained by the above simple method. However, it may be calculated from its impulse response $g(t)$ which can be measured by digital technique, and it is also possible that $Y_1(t)$ and $P_1(\omega)$ can be directly calculated from its $g(t)$. Accordingly, some quantitative knowledge can be obtained about T.D. contributed by each pole, which, generally, is related to a certain part of the loudspeaker vibration system.

(a) $Q=3$

\begin{align*}
Y_1(t) & \quad F_f=30 \\
2 & \quad 50 \quad t(\text{ms}) \\
& 100 \\
2 & \quad 0 \\
P_1(\omega) & \quad 0.001 \\
& 0 \quad 50 \quad f \\
& 100
\end{align*}

\begin{align*}
Y_1(t) & \quad F_f=60 \\
2 & \quad 50 \quad t(\text{ms}) \\
& 100 \\
2 & \quad 0 \\
P_1(\omega) & \quad 0.001 \\
& 0 \quad 50 \quad f \\
& 100
\end{align*}

\begin{align*}
Y_1(t) & \quad F_f=90 \\
2 & \quad 50 \quad t(\text{ms}) \\
& 100 \\
2 & \quad 0 \\
P_1(\omega) & \quad 0.001 \\
& 0 \quad 50 \quad f \\
& 100
\end{align*}

(b) $Q=1$

\begin{align*}
Y_1(t) & \quad F_f=30 \\
2 & \quad 50 \quad t(\text{ms}) \\
& 100 \\
2 & \quad 0 \\
P_1(\omega) & \quad 0.001 \\
& 0 \quad 50 \quad f \\
& 100
\end{align*}

\begin{align*}
Y_1(t) & \quad F_f=60 \\
2 & \quad 50 \quad t(\text{ms}) \\
& 100 \\
2 & \quad 0 \\
P_1(\omega) & \quad 0.001 \\
& 0 \quad 50 \quad f \\
& 100
\end{align*}

\begin{align*}
Y_1(t) & \quad F_f=90 \\
2 & \quad 50 \quad t(\text{ms}) \\
& 100 \\
2 & \quad 0 \\
P_1(\omega) & \quad 0.001 \\
& 0 \quad 50 \quad f \\
& 100
\end{align*}

Fig.2
RECTANGLES AS A SHAPE OF LOUDSPEAKER BAFFLES

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Introduction

When a loudspeaker is mounted in a finite plane baffle, the aim is at extending the frequency response to lower frequencies and a desire for flat and smooth frequency responses is accompanied. It has long been thought good for the aim to lengthen the radii to the baffle-edge from the loudspeaker, and good for the desire to break the uniform distribution of their length. The idea of the length of radii and its distribution was put into a theory [1]. It served to design a flat-frequency-response baffle [2].

That baffle cost only two times of area of the circular baffle whose cut-off frequency was the same. It was not a rectangle in shape which was the most convenient to make and use. This paper searches many rectangles and many locations of loudspeaker for flat and smooth frequency responses. The search is made concerning the far-field axial response. More than twice the area is spend on the baffles.

Processing [1] the Shape of Rectangles

The edge of the baffle is described (Fig. 1) in polar co-ordinates $R(\theta)$ with the origin at the loudspeaker. The

\[
L \times B = 1.6 \times 0.63 \\
L/B = 2.6 \\
LB/S_0 = 2. \\
\bar{R} = .3996
\]

Fig. 1
length $R$ of the radii is averaged into $\bar{R}$ over the whole range of angle. A circular baffle is supposed whose radius is $R_0 = \bar{R}$, whose area is $S_0 = \pi \bar{R}^2$ and at the centre of which the loudspeaker is mounted in. These circular and rectangular baffles have the same cut-off frequency $f_L$.

The $R(\theta)$ of a rectangle is not monotonic. The radii are re-arranged around the origin into an increasing order of length. This changes the baffle-shape $R(\theta)$ into a spiral one $R(\theta)$ (Fig. 2). We take the distribution of $R$ and calculate its density (Fig. 3) by

$$p(R) = (d\theta/2\pi)/dR.$$

It is changed in sign and preceded by a unit impulse. Their argument $R$ is changed into the time $t$ by $R = ct$, $c$ being the sound velocity. Then their Fourier transform gives the far-field axial frequency response (Fig. 4) of the baffle $R(\theta)$.

Length-to-breadth Ratios and Loudspeaker-locations

Ratios of length $L$ to breadth $B$ of rectangles are given in four levels,

$$L/B = 1, 1.6, 2.5, 4.$$  

We give such $L$ and $B$ (Fig. 5) while keeping the area $S = LB$ unity. Ratios of the area of rectangular baffles to that of circular ones are given in four levels,

$$S/S_0 = 1.12, 2, 2.8, 4,$$

which leads to different values of $\bar{R}$. A set of $S$ and $\bar{R}$ gives a contour of loudspeaker locations (Fig. 6).
The \( p(R) \) and frequency response are calculated for many sets of \( L, B, \) and loudspeaker positions. All the sets are searched for a \( p(R) \) and response which will be as approximate as possible to those (Figs. 8-9) of the flat-frequency-responded baffle (Fig. 7) [2].

**Feature of Rectangles in \( p(R) \) and Response**

This extent of search has led to finding the following. Frequency response of rectangular baffles does not get rid of waviness (Fig. 4). Their \( p(R) \) does not get rid of four spikes (Fig. 3) and so remains different from that in Fig. 8. Symmetry in baffle-shape increases the waviness in frequency response making the four spikes degenerate into three (Fig. 10) or less. The spikes never vanish nor break into five or more. Increasing the area \( S \) relative to the \( S_0 \) does not let the \( p(R) \) nor the response approximate those in Figs. 8-9. Rather, this possibly decreases the sensitivity in the pass-band (Fig. 11) by locating the loudspeaker on the edge.
INOUÉ T.  Rectangular Baffles

The p(R) and Response of Triangular Elements

The result can be explained with one of four triangular elements (Fig. 12),

\[ R(\theta) = A \sec \theta, \quad (-\phi_2 < \theta < \phi_1), \]

into which a rectangular baffle is divided. It resembles a sector element of a circular baffle in p(R) (Fig. 13) and in response (Fig. 14). The frequency response has alternate peaks and dips. The p(R) has one spike resembling the

\[ p(R) = \delta(A) \]

of a circular baffle \( R(\theta) = A \). The non-vanishing, non-breakable four spikes show that a rectangle approximates a superposition of four sectors as a shape for loudspeaker baffles.

Fig. 12  Fig. 13  Fig. 14

Conclusion

It is yet to be done to evaluate and minimize the difference of levels between peaks and dips in frequency response. This paper has shown that rectangular baffles cannot, by nature, be the flat-frequency-responded baffle I proposed to the 10-ICA. They look nothing more than weighted superpositions of four approximately circular baffles.

References

A FIBER OPTIC LEVER HYDROPHONE

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Introduction

During the past few years two classes of fiber optic sensors developed. The first includes devices which utilize optical interferometry and rely on single mode fibers and phase detection methods, while the second consists of sensors based on intensity modulation and multimode technology. Within the second class it is possible to differentiate the sensor concepts into two sub-classes. The first basically comprises a conditioned light input signal passing through some form of acousto-optical interaction region where the light beam is amplitude-modulated by the acoustic field and then detected upon transmission. The second type, although similar in concept, detects its ensuing modulations upon reflection within the interaction region. This latter approach falls within the nomenclature of fiber optic levers and it is the subject of this paper. By definition a fiber optic lever consists of a light beam conveyed to a vibrating target (interaction region) by a method that causes the power in the reflected beam to be modulated in a manner proportional to the displacement of the target. This paper concentrates on the use of a fiber optic lever in the design of a hydrophone consisting of a flexible bifurcated bundle of optical fibers whose common end is placed in the vicinity of a reflective surface (Ref.1). In this manner any motion of the reflector caused by acoustic perturbations modulates the light intensity of the reflected beam entering the receive fibers and generates an electrical signal proportional to the displacement.

Theory

The fiber optic lever transducer basically consists of four components, as shown in Fig.1: a light source, a fiber optic light guide, a pressure sensing element such as a reflective surface sensitive to the acoustic field and a photodetector. The optical fibers consist of a light transmitting bundle and a light receiving bundle. Each bundle may contain very few fibers or as many as hundreds, depending on the configuration and light level requirements. Both bundles are arranged at the distal end of the transducer to obtain maximum sensitivity relative to the diaphragm displacement, while at the proximal end they are separated allowing the light source and photodetector to be connected.

Kobayashi et al. (Ref.2) evaluate the overall performance of the fiber optic lever in terms of a light transfer coefficient, η, which is defined
as the ratio of the total radiated power from the proximal tip of the receiving fiber to the total incident light power into a transmitting fiber. Their calculations are based on a two-fiber system as shown in Fig. 2. The quantity \( n \) is given by two factors, the first \( [\ ] \) representing the light transmission loss and the second \( [\ ] \) dependent on the mirror displacement:

\[
\eta = \frac{1}{\pi^2 a_c} \int_0^{\theta_2} \frac{4}{\pi} \tau(\theta)^2 \exp(-a/\cos\theta) \int_0^{\theta} (1-T)(\tan \theta' / 2a \sqrt{1-z^2}) \sqrt{1-z^2} dz \frac{x[\pi\phi_c - 2}{a} \int_0^{\phi_c} \{x_0 / 2a \sqrt{1-(x_0 / 2a)\beta} + \sin^{-1}(x_0 / 2a)\} \, d\phi \, d\theta \tag{1}
\]

where: \( \sin \theta = \sqrt{n_1^2 - n_2^2} \) = numerical aperture.
\( \theta' = \) refraction angle which satisfies \( \theta' = \sin^{-1}(\sin \theta / n_1) \).
\( \tau(\theta) = \) Fresnel reflection coefficient at both ends of glass core.
\( a = \) absorption coefficient of glass core.
\( t = \) fiber length.
\( T = \) loss coefficient between core and cladding.
\( a = \) fiber core radius.
\( z = \) distance from core center.
\( \phi_c = \cos^{-1}\left\{ [R_1^2 + (2d \tan \theta)^2 - (2a)^2] / 4R_1 d \tan \theta \right\}.
\( x_0 = [R_1^2 + (2d \tan \theta)^2 - 4R_1 d \tan \theta \cos \phi]^{1/2}.
\( R_1 = \) distance between transmitting and receiving fiber centers.
\( d = \) distance between reflector and distal end.

The integral limits \( \theta_1 \) and \( \theta_2 \) are obtained from the appropriate conditions set on \( \theta_c \).

Among the acoustic parameters, the signal sensitivity and the minimum detectable pressure level have been used to describe the performance of fiber optic hydrophones. The signal sensitivity in this case can be represented in terms of the change in optical power at the interaction region between the common fiber tip and the reflector as a function of the mechanical displacement due to acoustic pressure variations. This ratio can be experimentally determined under static conditions. Thus, the minimum expected displacement, \( \Delta d \), can be given by

\[
\Delta d = \Delta P / x_P \tag{2}
\]

where \( x \) is the sensitivity in \( m^{-1} \), \( \Delta P \) is the estimated minimum detectable power in watts, for a given optical power, \( P \), at the detector in watts. From this expression it is possible to obtain the noise-equivalent pressure, \( \Delta P \), assuming that the mechanical displacement follows the particle motion in water. This assumption ignores the effects on free-field conditions due to the diaphragm stiffness or other factors inherent in a practical hydrophone design. On this basis, the relation giving the acoustic pressure for plane waves is

\[
\Delta p = \rho c w d \tag{3}
\]
where $p_c$ is the characteristic impedance of the medium in SI-plates, and $\omega$ is the angular frequency in rad/s. It should be stressed that since operation below mechanical resonance is most useful for these devices Eq. (3) is not realistic and similar expressions including explicitly the effects of compliance due to loading should be used. This approach will predict a frequency response deviating markedly from the 6 dB/octave increase given by Eq. (3).

Experiments

We briefly discuss the performance of two fiber optic lever hydrophones labeled A and B. Unit A consists of a pressure device utilizing a commercially available bifurcated fiber optic bundle with random distribution at the distal end. The tip is held in place at the center of a cylindrical body and a front surface mirror is cemented on a silicone window installed in a ring with fine threads. A simple adjustment of the ring on the transducer body provides a suitable method to select the optimum working path between the fiber tip and the reflector. The transducer body is designed for air-backed and water-loaded operation. Unit B consists of a brass body containing a miniaturized bifurcated bundle, a silicon photodiode and a small incandescent lamp. The reflector assembly is of a spring-loaded design and can be installed to provide pressure and pressure gradient data. Fig. 3 gives the open circuit receiving pressure sensitivity of units A and B while the responses of a B&K type 8103 and a LC-32 hydrophone are included for comparison. Fig. 4 illustrates the omnidirectional characteristics of unit A, water-filled. In Fig. 5 the open circuit receiving pressure gradient sensitivity of unit B is shown. Units B1 and B2 differ in the reflector construction only, the first being spring-loaded while the second consisting of a front surface mirror edge-mounted on an aluminum ring. Unit C is a pressure gradient sensor using a metal-backed piezoceramic wafer edge-clamped between two rings. Fig. 6 shows the dipole pattern obtained with unit B2.

Conclusion

It has been shown that fiber optic levers are adaptable to hydrophone construction and provide good sensitivity and directivity. An in-depth analysis of the experimental results is given in Ref. 3.

References


Fig. 1. Fiber optic lever transducer.

Fig. 2. Geometry of two-fiber system.

Fig. 3. Open circuit receiving pressure sensitivity.

Fig. 4. Directivity pattern. Unit A - water loaded. Frequency - 240 Hz.

Fig. 5. Open circuit receiving pressure gradient sensitivity.

Fig. 6. Directivity pattern. Unit B2. Frequency - 202 Hz.
ELECTROSTATIC TRANSDUCERS WITH BACKPLATE ELECTRET

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INTRODUCTION
Electret microphones are operating on the same principle as normal condenser microphones but they do not need an external polarisation voltage because a charged electret material is used /1/. Present electret transducers utilize either an electret diaphragm or an electret-coated back electrode. In both cases FEP is generally used as electret material because of its excellent charge-storage properties.

In the backplate electret microphone described in this paper metallized polyethylene terephthalate (PETP) membranes were used but other materials, for example metal, are possible as well. The advantage of the PETP membrane is that it has a small mass per unit area and can be stretched to high tension resulting in a small vibration sensitivity and a wide frequency range. Fig. 1 shows the cross-sectional view of the main parts of such an electret microphone.

In the following, calculations and measurements on circular and annular ring backplate electret microphones are discussed.

CIRCULAR MICROPHONES
The amplitude and phase response of circular transducers have been calculated using the theory of membranes. The solution includes the radiation pressure and the reaction pressure due to the interaction between membrane and air gap /2/, /3/.
Measurements of the acoustic properties of the microphones show very good agreement between theoretical and experimental data. As an example, Fig. 2 depicts the sensitivity as function of frequency for a typical microphone with a membrane radius a = 3 mm. The 12 µm Teflon electret is charged to $1.5 \times 10^{-4} \text{C/m}^2$, i.e. equivalent to 100 V dc. This transducer can be used for audio applications.

Fig. 2: Sensitivity level relative to 1V/Pa and phase response of a circular electret microphone (pressure response).
Membrane radius $a = 3 \text{ mm}$, air gap $h = 30 \mu \text{m}$, membrane tension $T = 766 \text{ N/m}$, charge $S = 1.5 \times 10^{-4} \text{C/m}^2$.

COMBINATION OF CIRCULAR AND ANNULAR TRANSDUCERS
An interesting combination for microphone design is that of a circular and an annular element. This kind of transducer has the advantage of high microphone capacitance. Each element produces a separate microphone voltage. Therefore, the voltage sources in the equivalent electric circuit of the circular and annular transducer $U_c$ and $U_a$ are in parallel branches and in series to their element capacitance $C_c$ or $C_a$ respectively /4/. Including the stray capacitance and the amplifier input capacitance, one obtains for the whole open circuit output voltage $U_m$:

$$U_m = \frac{C}{C_c} U_c + \frac{C}{C_a} U_a \quad \text{(Eq. 1)},$$

where $C$ is the sum of all above-mentioned capacitances.

In order to achieve a smooth amplitude and phase response of the microphone, and a high sensitivity, we aim at equal resonance frequency and equal sensitivity for the elements. The voltages $U_c$ and $U_a$ in Eq. 1 are proportional to the average membrane deflections and the electric field in the air gap, while the capacitances $C_c$ and $C_a$ are proportional to the active transducer areas. The resonance frequencies can be calculated from the membrane parameters. Using these properties, and for reasonable membrane dimensions, we get the following conditions for the ratio of circular membrane and annular membrane radii $a$ and $a_1$, $a_2$ respectively.

same sensitivity: $a = (a_2 - a_1) \cdot (1.2 \ldots 1.3) \quad \text{(Eq. 2)}$

same resonance frequency: $a = (a_2 - a_1) / 1.3 \quad \text{(Eq. 3)}$

It is evident, that both conditions can not be fulfilled simultaneously.
Fig. 3: Sensitivity and phase response of a microphone consisting of a combination of circular area and annular ring.
\[ a = 3 \text{ mm}, \quad a_2 - a_1 = 2.6 \text{ mm}, \]
\[ T = 450 \text{ N/m}, \quad h = 39 \mu\text{m}, \]
\[ S = 1.5 \times 10^{-9} \text{ C/m}^2. \]

Fig. 4: Same as Fig. 3 but:
\[ a_2 - a_1 = 3.1 \text{ mm}, \quad T = 500 \text{ N/m}. \]

Fig. 5: Same as Fig. 3 but:
\[ a = 2.3 \text{ mm}, \quad a_2 - a_1 = 4 \text{ mm}, \]
\[ T = 393 \text{ N/m}. \]

Some experimental results are shown in Fig. 3, 4, and 5. For high sensitivity of both transducers (Fig. 3), we achieve separate resonance peaks of the elements at 12 kHz from the circular membrane and at 16.8 kHz from the annular element. For larger annular elements, or smaller circular elements respectively, the ring transducer dominates. In Fig. 5 the average membrane deflection and the capacitance of the ring element is much larger than the corresponding quantities of the circular element. Therefore, the frequency response has only one maximum.

**TRANSIENT RESPONSE**

Another important property for microphone application and design is the transient response of the transducer. The membrane was excited by an electrostatic actuator using a high-voltage wide-band amplifier. Fig. 6 shows the transient response of a well damped and a lightly damped microphone with resonance frequencies at 12 and 18 kHz, respectively. Excited by rectangular signals, the well damped membrane shows a small response time \((a)\). If the carrier frequency of a sine-wave-burst signal is in the smooth
part of the transmission range, backplate electret microphones show an excellent transient response (c and d). For signal frequencies in the resonance range, which are not depicted here, the transducers show the behavior of band-pass filters.

Fig. 6: Transient response of backplate electret microphones for rectangular and sine wave burst excitation at 1 kHz carrier frequency. The upper part of each drawing shows the excitation signal.
left side: well damped microphone right side: lightly damped microphone

SUMMARY
Backplate electret microphones show the same good response to rapidly changing acoustic signals as other highly tuned transducers. It is impossible to design combinations of circular and annular ring elements which fulfill the condition of equal sensitivity and equal resonance frequency simultaneously. However, theoretical and measured results show good agreement. Circular transducers can be designed for various applications, such as in telephony and audio systems.

REFERENCES
VIBRATION AND ACOUSTIC RADIATION OF PIEZOELECTRIC TRANSDUCER

--- PFM - EQUIVALENT CIRCUIT ANALYSIS

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Introduction

Several papers about numerical calculation based on finite element method (FEM) for vibration and radiation of piezoelectric transducers were published. In this paper, the FEM-equivalent circuit analysis, in which a multimode equivalent circuit with lumped parameters calculated by FEM is applied, is developed. The computed performances of two types of underwater transducers are presented and compared with the measured.

Calculation Principle

It is assumed that the piezoelectric transducer and medium considered here are linear and without interloss. There are two electric terminals in the transducer.

Being treated with FEM, a piezoelectric transducer of arbitrary configuration is described by discrete equation

\[
\begin{bmatrix}
K_{11} & -K_{12} \\
-K_{12}^T & K_{22}
\end{bmatrix}
\begin{bmatrix}
d \\ 
\varphi
\end{bmatrix}
+ \omega^2
\begin{bmatrix}
M & 0 \\
0 & 0
\end{bmatrix}
\begin{bmatrix}
d \\ 
\varphi
\end{bmatrix}
=
\begin{bmatrix}
F_p \\
-q_p
\end{bmatrix}
\] (1)

where \(d\) and \(\varphi\) are nodal displacements and electric potentials, \(F_p\) and \(q_p\) nodal external forces and electric charges, \(K_{11}, K_{22}, K_{12}\) and \(M\) are stiffness, electric stiffness, electromechanical coupling and mass matrixes, respectively. \(\tau\) is transposition operator.

It is proved that an arbitrary \(d\) and corresponding \(\varphi\) can be expanded in the form

\[
\begin{bmatrix}
d \\ 
\varphi
\end{bmatrix}
= \sum_{i=1}^{n} q_{i_1} \begin{bmatrix}
d_i \\ 
\varphi_i
\end{bmatrix}
+ V
\begin{bmatrix}
0 \\ 
\varphi_0
\end{bmatrix}
\] (2)

where \(V\) is the voltage between two terminals, \(d_i\) and \(\varphi_i\) are ith displacement eigenmode and corresponding nodal electric potentials when \(V=0\), \(\varphi_0\) is nodal electric potentials when \(d=0\), \(q_{i_1}\) is complex coefficient of ith mode, \(n\) is dimension number of \(d\).
Substituting Eq. (2) into (1) yields the equation for modes:
\[ Z_g q = gV + F \]
\[ Q = g^T q + C_0 V \]  \hfill (3)
where impedance matrix for modes \( Z_g \) is a diagonal matrix composed of terms \( \omega_1^2 - \omega_2^2 \), \( \omega_1 \) is eigenfrequency, \( Q \) charge on the terminal, \( C_0 \) capacitance when \( d=0 \), vector \( q = [q_1 \ldots q_n]^T \), coupling vector for modes \( g = [g_1 \ldots g_n]^T \), force vector for modes \( F = [F_1 \ldots F_n]^T \) and
\[ g_i = d_i^T V \]
\[ F_i = d_i^T F \]  \hfill (4)
\hfill (5)

When the transducer radiates in infinite liquid, the external force is acoustic pressure \( p \) on surface, which is a linear function of normal vibration displacement \( w \). Treating \( p \) and \( w \) with discrete method and expanding in terms of modes, we obtain
\[ F = -Z_r q = -W^T SPq \]  \hfill (6)
where \( W \) is normal displacement matrix, 1th column of which is the discrete normal displacement of 1th mode on the surface, \( S \) is diagonal matrix composed of area of discrete surface elements, \( P \) is acoustic pressure matrix, 1th column of which is the surface pressure when the transducer vibrates with displacement of 1th mode and can be calculated by solving the Helmholtz integral equation on the surface numerically\( ^{15} \).

Substituting Eq. (6) into Eq. (3) yields the equation of transducer radiating in medium
\[ (Z_g + Z_r) q = gV \]
\[ Q = g^T q + C_0 V \]  \hfill (7)

The equivalent circuit corresponding to this equation is shown in Fig. 1, where each sub-loop is relative to a mode and the component parameters connected with transducer are independent of frequency, i.e., they are lumped. Besides, though this circuit is derived through finite element approach, according to the convergence of FEM it will express continual piezoelectric elastic structure accurately when \( n \) approaches infinity.

The vibration displacements and electric impedance of transducer can be obtained by solving Eq. (7) for \( q \) and \( Q \) when \( V \) and \( \omega \) are given. Then the acoustic field can be computed through Helmholtz formula. Since the modes that their eigenfrequencies are much greater than operating frequency have very little contribution, only a few modes need to be taken in practical calculation, which will simplify calculating process.

Practical Examples

The calculated results of a composite rod and a concave flexensional underwater transducer are shown in Tab. 1, Fig. 2 and Tab. 2, Fig. 3 respectively and conformed with the measured
Fig. 1 Multimode Equivalent Circuit

on the whole. Some differences between the calculated and the measured may be caused by the ignorance of damping in calculation and other causes.

Conclusion

The calculation method explained in this paper has some advantages of both FEM and equivalent circuit analysis. It is available for arbitrary geometry and structure and easy to be simplified as well as helpful to obtain more clear physical concepts.

References

3. Y. Kagawa & T. Yamabuchi, IEEJ SU-26, 81 (1979)

Tab.1. Eigenfrequencies of Composite Rod Transducer (kHz)

<table>
<thead>
<tr>
<th></th>
<th>f₁</th>
<th>f₂</th>
<th>f₃</th>
<th>f₄</th>
<th>f₅</th>
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<td>20.5</td>
<td>25.7</td>
<td>36.1</td>
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<tr>
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<td>0</td>
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<td>21.0</td>
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<td>36.5</td>
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Tab.2. Eigenfrequencies of Flextensional Transducer (kHz)

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<th>f₂</th>
<th>f₃</th>
<th>f₄</th>
<th>f₅</th>
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<tbody>
<tr>
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<td>6.56</td>
<td>9.20</td>
<td>11.1</td>
<td>14.1</td>
</tr>
</tbody>
</table>

a. half cross section  b. calculated displacements  c. electric conductance of modes
**Fig. 2** Performances of the Composite Rod Transducer
- calculated with 5 modes, measured, except b.

**Fig. 3** Performances of the Concave Flextensional Transducer
- calculated with 5 modes, measured, except b.
MESURE DIRECTE DES PROPRIÉTÉS ACOUSTIQUES DES MATÉRIAUX

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1 - INTRODUCTION

L'obtention de l'impédance et du coefficient d'absorption d'un matériau nécessite une installation de laboratoire si modeste soit elle. L'objectif recherché est de développer un système maniable et portable afin de simplifier le protocole de mesure et de limiter l'appareillage en vue d'une utilisation en basses fréquences pour le contrôle du suivi des caractéristiques acoustiques des matériaux. Ce système se compose d'un capteur, réalisé pour obtenir les informations utiles, auquel est associé un appareil électronique qui fournit le signal excité et traite les signaux délivrés par le capteur.

2 - PRINCIPE DE LA MESURE

On excite une cavité cylindrique fermée par un piston rigide contre lequel on peut placer ou non le matériau à tester. La méthode utilisée est une méthode relative : elle permet de déduire les caractéristiques cherchées par comparaison des résultats obtenus avec et sans matériau.

Dans l'hypothèse où les dimensions de la cavité sont suffisamment faibles pour pouvoir y admettre l'uniformité de la pression sonore, le relevé de cette pression et la connaissance du débit acoustique au niveau de l'excitation (§ 4 ci-dessous), permettent d'obtenir l'impédance du matériau à partir de la relation usuelle liant ces deux quantités dans une petite cavité.

Cependant, l'uniformité du champ dans la cavité suppose que les dimensions de celle-ci soient très inférieures à la longueur d'onde. Cette condition est facilement réalisable en ce qui concerne la profondeur de la cavité, les fréquences utilisées ici restant inférieures à 2000 Hz. Par contre, afin d'éviter certains problèmes liés à la non homogénéité des matériaux testés et à la découpe de ces matériaux, il est intéressant de pouvoir donner au rayon de la cavité des valeurs pouvant atteindre quelques centimètres ; il est alors utile d'étudier la répartition de la pression dans la cavité.
3 - ÉTUDE DE L'UNIFORMITÉ DU CHAMP DANS LA CAVITÉ

La pression acoustique dans la cavité est obtenue à partir de l'équation intégrale de Helmholtz-Huygens en choisissant une fonction de Green, satisfaisant aux limites du domaine aux conditions de Neumann, que l'on développe sur la base des fonctions propres (satisfaisant à ces mêmes conditions). L'expression de la pression est alors une double sommation dont les indices sont liés aux modes radiaux et axiaux dans la cavité. Le premier terme de la série, correspondant au mode \((0, 0)\) n'est autre que l'expression usuellement retenue pour la pression dans une petite cavité. Il est possible d'annuler tous les termes \((1, n)\) correspondant au premier mode radial (termes principalement responsables de la non uniformité du champ dans la cavité, la profondeur de celle-ci restant petite) par un choix correct de la source excitant la cavité [1], à savoir une source annulaire dont les rayons intérieur \((r_1)\) et extérieur \((r_2)\) sont choisis en fonction du rayon \(r_c\) de la cavité de manière à annuler l'expression :

\[
J_1(r_1/r_c) - J_1(r_2/r_c)
\]

\((\gamma_1\) est la première racine non nulle de \(J'_0(\gamma) = 0\)).

C'est la solution que nous avons retenue dès que le rayon \(r_c\) est supérieur à 1 cm.

Tout en assurant ainsi au mieux l'uniformité de la pression dans la cavité plusieurs types de capteurs peuvent être imaginés. En particulier nous voyons maintenant que l'excitation sonore délivrée par un haut-parleur peut être transmise à la cavité par l'intermédiaire d'un tube capillaire (ou d'une fente annulaire très fine). La propriété utile de ce tube (ou de la fente) est qu'il permet d'obtenir une fonction de transfert-vitesse particulaire en aval sur pression acoustique en amont indépendante de la charge acoustique présentée par la cavité. Le relevé des pressions en amont et en aval du tube capillaire fournit alors les informations utiles : la pression sonore au niveau du matériau et le débit acoustique excitateur dans la cavité. Nous précisons ci-dessous les conditions à satisfaire lors du choix du capillaire.

4 - ÉTUDE DE LA FONCTION DE TRANSFERT DU CAPILLAIRE

A partir des solutions ondes planes pour la pression dans un tuyau cylindrique avec amortissement viscothermique [2], on obtient la fonction de transfert :

\[
\frac{P_1}{u_0} = Z_{ch} \text{ch} \ ikl + Z_{c} \text{sh} \ ikl
\]
M. BRUNEAU et al. - MESURE DIRECTE DES PROPRIETES ACOUSTIQUES DES MATERIAUX

où $l$ est la longueur du tube capillaire et $Z_0$ son impédance acoustique au niveau de l'extrémité aval et où la constante de propagation $ik$ et l'impédance caractéristique $Z_c$ peuvent être exprimées en fonction, entre autres paramètres, du rayon $r_0$ du tube.

La méthode de mesure utilisée étant une méthode relative, le rapport $P_1/u_0$ peut être fonction de la fréquence, mais ne doit pas dépendre de $Z_0$, puisque cette impédance est liée à la charge acoustique placée en aval. Ainsi si $Z_{in}$ et $S_{in}$ sont l'impédance acoustique d'entrée et la section de la cavité de mesure, et $S_0$ la section du tube, la condition

$$Z_0 = Z_{in} S_0 / S_{in} << Z_c \text{ th ikl}$$

reliant $r_0$ et $l$ aux caractéristiques de la charge acoustique doit être satisfaite.

5 - RESULTATS ET CONCLUSION

A l'heure actuelle les résultats obtenus sont convenables pour certains matériaux mais sont très décevants pour d'autres.

Les résultats, lorsqu'ils sont mauvais, sont en général non reproduisibles et en particulier sont très sensibles au conditionnement du matériau (échantillon plus ou moins serré dans la cavité). Il est possible que les vibrations de la structure du matériau soient en grande partie à l'origine des problèmes rencontrés [3].

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A TRANSIENT METHOD FOR MEASURING THE IMPULSE AND FREQUENCY RESPONSE OF MICROPHONES IN THE FREE FIELD.

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INTRODUCTION

Traditionally, anechoic conditions have been needed to accurately determine the frequency response of an acoustic system. Anechoic chambers are very costly to install and it is difficult to eliminate stray reflections from necessary equipment inside the chamber. The availability of fast and relatively low-cost computers has thus heralded renewed interest in transient testing methods which do not rely on the use of anechoic chambers. This report describes a transient testing method which allows the free field impulse response and transfer function of a microphone to be measured in an ordinary laboratory.

The method is based on that developed by Berman and Fincham [1] to measure the frequency response of loudspeakers. A narrow voltage pulse is applied to a loudspeaker and the resulting acoustic pressure response is measured by a microphone some distance away. Examination of the time history of the microphone's output shows the loudspeaker's response to the pulse which directly propagates to the microphone clearly separated from the subsequent echoes caused by reflections from the room boundaries. The reflections may be "windowed" out to leave only the direct response which can be used as an estimate of the free field response of the loudspeaker to the pulse, assuming the microphone has a flat free field amplitude response. If the direct response is then Fourier transformed, the loudspeaker's free field frequency response is given.

If the procedure is repeated with another microphone placed in exactly the same position with respect to the loudspeaker, the difference between the measured frequency response of the second - "test" - microphone and the first - "reference" - microphone is the free field frequency response of the test microphone, again assuming the reference microphone has a flat response. The inverse Fourier transform of this difference is the response of the test microphone to a simulated delta function of pressure propagating in the free field past the microphone.

It should be noted that the detailed response of the loudspeaker to the pulse does not in principle affect the final result for the test microphone provided the direct response has decayed sufficiently before the arrival of the first room reflection. All that is assumed is that the initial pressure response of the loudspeaker is the same from experiment to experiment. In
practice, however, a reasonable signal to noise ratio must be preserved at all frequencies of interest and this determines the choice of loudspeaker and the pulse width.

RESULTS AND DISCUSSION

A computer-controlled system was used to acquire the preliminary results presented in this report. A computer triggers a pulse generator to produce a narrow rectangular pulse (10μs wide) which is amplified by a power amplifier and fed to a loudspeaker suspended in the measuring room. A microphone measures the response of the loudspeaker to the pulse. The microphone response is amplified and low-pass filtered before it is read into the computer through an analogue to digital converter. Signal averaging is employed to improve the signal to noise ratio; the reverberation time of the measuring room is such that 64 averages take a little over 1 minute to complete. Once the response has been acquired, it is saved on the computer's disk memory for future use.

The size of the measuring room determines the time scale of the impulse response and hence the frequency resolution and lower frequency limit of the final frequency response. The upper frequency limit is determined by the sampling rate of the analogue to digital converter and the frequency range of the loudspeaker. In these measurements, the frequency range of the final results is approximately 200Hz to 60kHz.

A typical microphone output signal (the reference microphone in this case) after the loudspeaker has been excited by the pulse, is shown in fig. 1(a). The direct response is followed some time later by the arrival of the first room reflection. Providing the direct response has decayed into the background noise by the time the first reflection arrives, it is possible to use a temporal gating technique to remove the reflections, leaving the true anechoic response measured by the microphone. Fig. 1(b) shows the Fourier transform (magnitude and phase) of the gated free field transient response and shows the spectrum of the free field output of the loudspeaker. Also shown is the spectrum of the background noise, obtained by repeating the experiment with the pulse generator disconnected. It can be seen that the signal to noise ratio is better than 40dB over the frequency range of interest. The reference microphone used was a B&K 1/4" 4135 free field microphone.

Once the reference and test responses have been acquired, they are gated using a half-Hanning window and Fourier transformed. The test microphone's frequency response is calculated by dividing the amplitude responses and subtracting the phase responses of the test and reference microphones. Inverse Fourier transformation then gives the desired test microphone impulse response. Fig. 2 shows the impulse response and frequency response of a B&K 1" 4145 free field microphone (with and without protection grid) produced in this way. It should be noted that small uncertainties in positioning the test and reference microphones relative to the loudspeaker inevitably give rise to linear phase shifts due to differing loudspeaker to microphone delays. Consequently, the position of the origin of the calculated impulse response is also uncertain.

Fig. 3 shows the free field correction curves (magnitude and phase) for a B&K 1" 4145 free field microphone at various angles of incidence with and without the normal protection grid. The curves represent the difference
between the microphone's pressure response and it's free field response. The free field responses were calculated as described above. The pressure response was measured in a similar way, although in this case the microphone was excited by a voltage pulse applied to an electrostatic actuator on the microphone, rather than via a loudspeaker, no gating or reference response being needed in this case. Once calculated, the correction curves may be saved on disk and are thus available for a fast and accurate calculation of the microphone's free field response at various angles of incidence.

Individual results may be acquired, processed and plotted in a matter of minutes using this method, and disk storage allows the results acquired over several months to be compared. Such comparisons of nominally identical experiments have shown the results produced by the method to be reliable and reproducible.

CONCLUSIONS

The method described in this report offers a fast, accurate and reliable method of measuring the free field impulse response and frequency response (magnitude and phase) of a microphone from about 200Hz to 60kHz without the need for an anechoic chamber.

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Fig. 1. (a) Reference microphone output signal showing the direct response followed by the first room reflection.
(b) Fourier transform of the gated reference microphone output signal showing the magnitude and phase of the output frequency response and the magnitude of the background noise spectrum.
(B&K 1/4" 4135 free field microphone)
Fig. 2 Impulse response and frequency response of B&K 1" 4145 free field mic. at 0° angle of incidence. (a) Impulse response with protection grid. (b) Impulse response without protection grid. (c) Frequency response with and without protection grid.

Fig. 3. Free field correction curves for B&K 1" 4145 free field microphone at 0° and 90° angle of incidence. (a) with normal protection grid. (b) without normal protection grid.
PIEZOMAGNETIC ALCOPER TRANSUDCERS FOR THE FREQUENCIES 22 AND 33 kHz

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Introduction

A typical magnetostrictive or piezomagnetic transducer is designed with the aid of an equivalent circuit, which is an electrical representation of differential equations that govern the motion of the transducer. The active elements of piezomagnetic transducer are typically in the shape of a thin rod, a stack of thin laminations and a bulk or laminated ring. From the magnetic reason the best form of the piezomagnetic transducers is a toroidal transducer. The simplest transducer from the mechanical point of view is a rod. If two such rods are joined together with cross shoulders it will be a window transducer (Fig. 1). This transducer has the advantage of forming a closed magnetic circuit. The result is nearly uniform excitation of the entire length of each leg (colum) and reduction of leakage flux to a low value. In this case the calculations of resonance frequencies from the one dimension only, i.e. length will not
always give a good approximation with errors depending on the
shape and the proportions of the all transducer dimensions.
Author proposed the approximate formula for narrow columns
window transducers assuming that additional mass $m_h$ of cross
pieces between columns is added to the equivalent mass of
the column $m_l$ [1]. For the symmetrical window type transduc-
ers the resonance frequency will be given [1] by

$$f_{rm} = \frac{f_H}{\sqrt{1 + \frac{2f_h}{e1} + \frac{2f^2}{e1} \left( \frac{1 - \pi}{2 - \pi} \right)}}$$

where $f_H = \sqrt{\frac{81}{E}}$ is resonant frequency at constant magnetic
field $H$ for the thin rod with the length $l$, $E$ is Young modu-
lus at constant $H$, $\epsilon$ is density, $l$ and $e$ are the length and
the width of column (leg), $h$ and $f$ are the high and the width
of shoulder and $r$ is the radius of are equal half width of
the window, i.e. $2r = f$ (see Fig.1).

Alcofer transducers

Transducer cores were made from Al-Co-Fe alloy (Alcofer)
containing 12% Al and 2% Co [2]. Laminations for cores have

![Fig.2](image-url) Impedance circles of the 33 alcofer transducer at the
static magnetic field of 50 and 500 A/m and amplitude of
dynamic magnetic field of 1 mA/m
thickness equal 0.2 mm, length of 82 or 55 mm and width of 32 or 24 mm (see Fig. 1). After cutting the sheets were annealed at 1000°C for 2 hours in H₂. Then these laminations were insulated and soldered (in the case of 22 kHz) or cemented (in the case of the 33 kHz transducer) on the face side. Physical properties of alcofer are similar to those of alfer: \( B_0 = 1.4 \, T, \lambda = 50 \times 10^{-6}, T > 500^\circ C, \sigma = 1.4 \, \mu \Omega \, m, k > 0.25 \) [3,4]. Some properties of the transducers were presented earlier [5].

Experimental results

Impedance circles of the alcofer transducers at the different bias were measured (Fig. 2). From the impedance circles

![Impedance circles](image)

Rys. 3. Dependences of the resonance and antiresonance frequencies \( f_r \) and \( f_a \) and magnetomechanical coupling coefficient \( k \) of the 22 and 33 kHz transducers on the magnetic bias

![Dependence of magnetic permeability](image)

Rys. 4. Dependences of the magnetic permeability at 15, 20, 30 and 50 kHz on the magnetic bias for the 22 and 33 kHz transducer at the temperatures of 20 and 100°C
the piezomagnetic properties of the 22 and 33 kHz Alcofer transducers for the room temperature were determined (Figs. 3 and 4). These piezomagnetic properties were examined as a function of temperature up to 100°C (Fig. 5).

Fig. 5. Temperature dependences of the resonance and antiresonance frequencies and magnetomechanical coupling coefficient for 22 and 33 kHz transducers at the magnetic bias of 0.5 and 1 kA/m

Conclusions

Generally, the changes of the piezomagnetic properties in the range of the temperature from 20 to 100°C did not exceed 30%. Maximum values of the k coefficient were higher than 0.30 but the optimum bias field was varying with temperature. Magnetomechanical coupling coefficient was greater than 0.15 at the bias field range from 100 to 1200 A/m. Relative values of the magnetic permeability were dropping from 800 at the demagnetized state to 200 at the bias corresponding to the maximum value of k.

Due to the high resistivity the Alcofer laminations may be thicker than those of nickel in transducers working at the same frequency or may be used at higher frequencies (up to 100 kHz) with higher efficiency as that of nickel transducers for 22 kHz.

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MICROPHONE SYSTEM FOR EXTREMELY LOW SOUND LEVELS

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Introduction
In some application fields, for example, hearing research, there is a need to measure very low sound pressure levels, in the order of −10 dB in third octave bands or lower. The lowest noise level microphone systems existing today, have typical inherent noise levels of 10 to 15 dB(A) or −5 to 5 dB in the third octaves between 20 Hz and 20 kHz. The noise spectrum of such a system is shown in Fig. 1, curve A. Since the lowest sound levels of interest cannot be measured by such systems directly use must be made of advanced signal processing and time consuming procedures during which the experimental conditions may change. To overcome these difficulties, experiments have been carried out to reduce the inherent noise from the internal sources of the microphone systems. These noise sources are partly in the microphone cartridge and partly in the preamplifier. While the cartridge noise contributes to the spectrum A it has been eliminated in spectrum B in Fig. 1, which shows the preamplifier noise only. For the measurement of spectrum B, the cartridge was substituted by an equivalent capacitor. As can be seen, a significant noise reduction requires minimization of cartridge noise as well as preamplifier noise.

Experimental Low Noise Preamplifier
During the last 10 – 15 years the inherent noise of preamplifiers has been significantly reduced. This is mainly due to the Field Effect Transistor which has replaced the vacuum
tube. The dominating noise sources of preamplifiers are described in the literature and shall not be discussed in this paper. However by appropriate component selection and careful optimization of the circuit, it has been possible to make experimental preamplifiers, having a 6 – 10 dB lower noise level than corresponding preamplifiers which are available today. With a source capacitance of 50 pF they have typically an A-weighted noise level of 0.7 \( \mu \text{V} \) and 1.2 \( \mu \text{V} \) for flat weighting from 20 Hz to 20 kHz; the noise spectrum of such an experimental preamplifier is shown in Fig.1, curve C.

The noise has been reduced to a level which corresponds to the noise of the input stage of modern measuring amplifiers. A low noise voltage amplifier has therefore been combined with the preamplifier to raise its output level by 20 dB which thus eliminates the influence of the measuring amplifier noise. Extra filtration of the supply voltages has been necessary to minimize hum components in the noise.

**Noise sources of Condenser Microphone Cartridges**

Analysis of transducer characteristics are often simplified by use of equivalent electrical networks. This is also relevant in connection with analysis of inherent noise in microphone cartridgels. A useful condenser microphone model is shown in Fig.2. In electrical as well as acoustical circuits Thermal Noise is produced by resistances or damping mechanisms. The noise pressure produced by an acoustical resistance is determined by the following formula:

\[
\overline{p_n^2} = \int_{f_1}^{f_2} 4 \cdot K \cdot T \cdot R_n \cdot df
\]

\( \overline{p_n^2} \): mean value of squared pressure  
\( K \): Boltzmann’s constant  
\( T \): absolute temperature  
\( R_n \): acoustical resistance  
\( f \): frequency

Inside condenser microphones there will normally be two such sources. One source belongs to the damping mechanism behind the diaphragm \( (R_1) \) and the other one to the resistance of the pressure equalisation vent \( (R_2) \).

The impedance loading the diaphragm on its outside \( (Z_a) \) has a real part \( (R_a) \) which also produces thermal noise. In principle there is a corresponding impedance \( (Z_b) \) and resistance \( (R_b) \) connected to the outside opening of the equalisation vent.

![Fig. 2. Model of Condenser Microphone Cartridge](image-url)
In the equivalent circuit the internal noise pressure generators can be treated as if they were connected to the respective acoustical input terminals. To determine the contribution of each noise generator to the overall cartridge noise, the spectrum of each generator should be multiplied by the transfer function between the respective acoustical input terminals and the electrical output terminals. Since the typical resistance values of Type 4145 are known, their noise can be calculated. The two transfer functions from the acoustical inputs to the electrical output are also known; from the diaphragm terminals it is practically equal to the pressure response, as the outside load on the diaphragm, $Z_d$, is small compared with the acoustical impedance of the diaphragm and its internal damping. The transfer function from the vent opening decreases by 20 dB/decade from the lower limiting frequency of the cartridge, 1.5 Hz to about 1000 Hz where it starts rolling off at a higher rate. From the above it can be shown that the most significant noise source in Type 4145 is $R_1$, the diaphragm damping. At frequencies below 30 Hz the vent resistance, $R_2$, is the most significant noise source.

**Experimental Low Noise Microphone Cartridge**

Since the most serious cartridge noise in Type 4145 and other existing microphones is produced by the diaphragm damping resistance, $R_1$, experiments have been carried out to minimize this effect, resulting in the development of 1" cartridges having a resistance 40 times lower than that of Type 4145.

As it has not been practically possible to reduce the diaphragm mass and stiffness proportionally, the frequency response of the microphone cartridges changes significantly; a peak appears at the diaphragm resonance (8 kHz). However, an electrical network has been developed which compensates for the microphone response so that a flat response is obtained for cartridge and network up to 13 kHz within 1 dB, see Fig. 3.

![Diagram](image_url)

The sensitivity of the cartridges is increased by 6 dB compared to Type 4145; i.e., $-20 \text{ dB re } 1 \text{ V per Pa}$.

The sensitivity increase minimized the influence of the preamplifier noise correspondingly; also the compensation network which is combined with the preamplifier contributes to the noise reduction.

The damping resistance, $R_1$, of the low noise microphone is typically $1.25 \times 10^8 \text{ Ns/m}^5$. It’s noise pressure ($1.43 \times 10^7 \text{ Pa/Hz}$ or $-43 \text{ dB SPL}$ for 1 Hz b.w.) is 16 dB lower than for $R_1$ of Type 4145. The output voltage noise spectrum is also 16 dB lower, as the transfer function of the experimental cartridge with compensating network is the same as that of Type 4145, see the third octave spectrum in Fig. 4.
Because of the significant cartridge noise reduction, the real part of the external load impedance, $R_a$, has become a dominating noise source at higher frequencies, see spectrum in Fig.4. At 10 kHz $R_a$ exceeds the internal diaphragm damping resistance, $R_1$.

$$R_a = \frac{\rho \cdot f^2 \cdot \pi \cdot K_r}{c} \quad \rho: \text{density of the air;}$$

$$f: \text{frequency;}$$

$$c: \text{speed of sound;}$$

$$R_a[10 \text{ kHz}] = 1.5 \cdot 10^6 \text{ Ns/m}^5 \quad K_r: \text{ratio between random and pressure response.}$$

The noise produced by the vent resistance, $R_2$ which is typically $6 \cdot 10^6 \text{ Ns/m}^5$ is also shown in Fig.4. Due to the transfer response from the vent terminals, this noise will contribute to the output noise at low frequencies. However, it does not play any practical role, as the preamplifier noise is dominant in that frequency range.

The noise contribution of the external vent resistance, $R_e$, is very low and can be neglected.

The acoustical noise spectra have all been calculated, while the preamplifier noise spectrum which is also shown in the figure has been measured.

The noise source analysis explains clearly the measured total noise spectrum of the system.

The A-weighted levels of the preamplifier, the internal damping, $R_1$, and the external load resistance, $R_a$, are -10 dB, -7 dB and -11 dB respectively resulting in an overall level of -4 dB(A).

Conclusion
An experimental one inch microphone system with a flat frequency response up to 13 kHz has been developed; a system which is able to detect extremely low sound pressure levels — the inherent A-weighted noise of the system is -4 dB or about 15 dB lower than the noise of any corresponding systems.

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MODE ANALYSIS OF PIEZOELECTRIC DISK TYPE TRANSDUCER

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1. Introduction

As is well-known, a disk transducer used in the thickness extensional vibration (TE) mode does not vibrate uniformly at
the flat surface. For predicting the performances such as the
directivity function and the volume velocity, the mode
analysis of disk transducers have been studied by many
researchers (1-5). Due to the lack of the measurements of the
vibration patterns, however, the relation between the actual
amplitude distribution and the theoretical one has been hardly
investigated.

The purpose of this study is to measure the vertical
velocity distributions and the frequency spectra as a function
of the diameter to thickness ratio (DTR), and to find out the
approximate boundary conditions by which the measured
frequency spectra and the measured velocity distributions can
be explained.

2. Experimental Procedure

The geometry of disk transducer is shown in Fig. 1. The
flat faces of Pb(Zr,Ti)O_3 transducer are electroded with
silver coating and poled along the axial direction. The
thickness and original diameter are 5 mm and 60 mm
respectively and their fundamental resonant frequency of TE
vibration is expected at about 420 kHz. This transducer is
reduce in diameter by about 2 mm with a conventional grinder
after the measurements of the frequency spectrum and the
vibration velocity distributions. This process is repeated
from 12 to 2 in DTR.

The frequency spectra are measured with a frequency
synthesizer (HP3325A), a gain-phase meter (HP3575A) and an X-Y
plotter. To show the degree of frequency response an
electromechanical coupling coefficient is calculated (6).

To measure the vibration velocity distributions, an
optical heterodyne technique is employed, which can measure
not only the vibration velocity amplitude but the vibration phases too.

3. Experimental Results

Figure 2 shows the product of the resonant frequency and the thickness \( f \cdot t \) as a function of DTR, where a diameter of open circle is proportional to the corresponding electromechanical coupling coefficient. Mode numbers shown in the figure are classified in accordance with vibration patterns and mode analysis mentioned later. Obviously two radial modes are observed besides a TE and a edge(E) modes(2). One radial mode, which is called R mode, can be observed throughout the frequency range used for the experiment and the other, which is called A mode after the literature(4), appears only for larger value of \( f \cdot t \) than 1.8 kHz.m.

Axial velocity distributions along diameters for the experimental points around the TE-1 in Fig.2 are presented in Fig.3. The base lines on which the patterns are drawn are proportional to \( d/l \). From this figure one can differentiate two radial mode series, because one mode series(R) has larger wave numbers in the radial direction than that of the other one (A) for the same order of vibration. Grouping of mode series in Fig.2 is made referring to these vibration patterns as well as the analysis followed later.

4. Discussion

Explanations of changes of \( f \cdot t \) and vibration patterns are tried to derive analytically assuming the disk as a finite isotropic one. Using the boundary condition that the flat surfaces are free, the wave equation gives the frequency equations whose dispersion curves are shown in Fig.4. The abscissa and the ordinate denote the product of the wave number in the radial direction \( g \) and the thickness \( l \) and normalized frequency respectively as used elsewhere(5), and a experimentally obtained value of \( \nu = 0.34 \) is used as a Poisson's ratio for this material. The numerical suffix denotes the order of root of the frequency equations, where \( \psi \) and \( \phi \) correspond to antisymmetric and symmetric modes with respect to the central plane respectively.

To examine which branches correspond to A and R modes, the third boundary condition that the tangential stress (\( \tau \)) or the extensional stress (\( \sigma \)) is zero at the cylindrical surface is applied to the frequency equations. The frequency spectra calculated applying the third boundary condition to a certain branch in Fig.4 are compared with ones in Fig.2. As results of these identifications, it is proved that \( \psi_1 \) accompanied with the boundary conditions \( \sigma = 0 \) at the cylindrical surface gives R mode series and \( \psi_2 \) with \( \tau = 0 \) gives A mode series. Vibration patterns of A mode series picked up from Fig.3 are shown in Fig.5 neglecting small ripples in the patterns and these vibration patterns agree well with ones calculated by Aggarwal (2) for the same
boundary conditions used here if the vibration phases at the center are neglected. So one can conclude that A mode series is given by the dispersion curve of $\psi_2$ with the boundary condition of $r = 0$ at the cylindrical surface.

5. Conclusion

Vibration velocity distributions in thick disk are successfully measured. In addition to the measurements, relations among frequency spectrum, boundary conditions and vibration patterns are made clear experimentally and analytically. These results may be useful for calculation both volume velocities and directivity functions of disk type transducers.

References

Fig.1. Geometry of transducer.

Fig.4. Dispersion curve for infinite plate.

Fig.5. Theoretical amplitude distribution of A mode.
Fig. 2. Experimental frequency spectrum with circles proportional to electromechanical coupling coefficient.

Fig. 3. Observed vertical velocity distribution patterns.
MEASUREMENTS OF THE DYNAMIC AND OPERATING PARAMETERS OF THE PIEZOELECTRIC TRANSDUCERS

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The transducer is always the kern of any ultrasonic equipment and to determine its operating data, mainly the efficiency and band-with it is necessary to know the dynamic parameters: the resonance frequency, the electromechanical coupling coefficient, the mechanical and electrical quality factors \( Q_m \) and \( Q_e \).

But, in case of a good ceramic transducer, the widely used method of measuring the motional loop on the complex plane is quite useless. This is explained in Fig.1

![Image](image-url)  
**Fig.1.** Admittance diagram of a piezoelectric transducer.

The dielectric loss angle

\[
\tan \delta = \frac{1}{Q_e}
\]

When the losses are reasonably small \( Q_e \gg 1 \) the loss angle \( \delta < 1^\circ \) and the motional circle is practically tangent to the imaginary axis. It may be easily shown, that

\[
\tan \beta = \frac{1}{\kappa^2 Q_m} \quad 2.
\]

This \( \kappa^2 Q_m \) value of the unloaded transducer is usually greater then about 60, and then \( \beta \) is also smaller then 1°, that is to say the center of the circle is situated very nearly on the real axis.
In this situation only the admittance resonance frequency \( f_y \) and the band-width \( Q_{me} \) may be determined, but exactly the same information may be obtained with much simpler measurement of admittance modulus \( Y \), as shown in Fig.2.

![Diagram of a ceramic transducer](image)

**Fig. 2. The diagram of a ceramic transducer**

Neither coupling coefficient nor the electrical \( Q_e \) can be found.

To solve this problem a new method is proposed, based on the comparison of the admittance and impedance moduli \( Y \) and \( Z \). This concept is shown in Fig.3.

Here the coupling factor may be calculated as it is done in some papers/from the ratio of the admittance and impedance resonance frequencies \( f_y \) and \( f_z \) as:

\[
k^2 = 1 - \left( \frac{f_y}{f_z} \right)^2
\]

The new concept lies in that the value of the product of the two moduli gives the product of the dynamic parameters \( n \), which may be called the "power transfer factor" [1];

\[
n = k^2 Q_{me} Q_e
\]
Fig. 3. The moduli of admittance and impedance

This factor is also equal to the ratio of the motional circle diameter to its distance from the imaginary axis (Fig. 1):

\[ n = Q_{ma}/Q_e \]

But, as \( Y = \frac{1}{Z} \), at admittance resonance, with the circle diameter situated on the real axis:

\[ Q_e = \frac{1}{R_{ma}} \]

and therefore:

\[ n = Q_{ma} \cdot R_{ma} \]

If needed, the electrical \( Q_e \) or dielectric loss angle at resonance may be also calculated as:

\[ Q_e = Q_{ma} \cdot R_{ma} / k^2 Q_{mo} \]

When the power transfer factor is known, the efficiency, and load line may be calculated [1] as:

\[ \eta_{ea} = \eta_{ma} \cdot (P_{n+p+1}/(P_{n+p+1})) \]

In this relation, \( p \) is the "load factor" of the loaded transducer.

\[ p = G_{mo} / G_L \]

Where \( G_L \) is the radiation conductance introduced into the circuit by the load.
With loading the circle diameter/or modulus maximum / decreases, which means also the lower value of the $Q_{mL}$.

$$\frac{Q_{ma}}{Q_{ml}} = p + 1$$  \hspace{1cm} \text{(10)}$$

When the relatively small difference between the admittance and impedance resonance frequencies is neglected, the relation /10/ may be also written as:

$$\frac{W_s}{W_e} = p + 1$$  \hspace{1cm} \text{(11)}$$

Is is now enough to measure the band-width of the loaded transducer to determine the load factor and corresponding efficiencies, and also to check whether the transducer is under or over-loaded/Fig.A/.

When the transducer is heavily loaded, as it is the case with good, high $Q_{ma}$ transducers and quarter-wave matching, the modulus curve becomes very flat and the static capacity makes the band-width measurements very inaccurate.

It is then better to measure the band-width "outside", in the radiated field with a small transducer used as a hydrophone.

But then it should be noted/Fig.2/that the transducer's $Q_m$ corresponds to the quadrantal frequencies $f$ and $f''$/Fig.1/ i.e. to the -3 dB points on the modulus, whereas for the radiation the conductance only is responsible and the quadrantal points correspond to the half value of the $G_m$.

Therefore, for determination of the $Q_{mL}$ the band-width of the loaded transducer should be measured as the 6 dB drop on the resonance curve measured in the radiated field.

THE LISSAJOUS FIGURES APPLICATION FOR THE PIEZOELECTRIC VIBRATOR UNIT MEASUREMENTS.

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Introduction.
A number of methods have been developed for measurement of the fundamental properties in piezoelectric materials [1-5]. The most widely used of them is the transmission method with a phase criterion. Here, a new, simple measuring circuit based on the transmission T-network is proposed, which with an oscilloscope as a detector only allows to determine the frequencies: \( f_m \) /at minimum impedance/, \( f_R \) /at maximum impedance/, \( f_q \) /at resonance/, \( f_a \) /at antiresonance/ of piezoelectric crystal unit. The determination of the characteristic frequencies and the low-frequency free permittivity permits to calculate the piezoelectric, dielectric and elastic coefficients of piezoelectric crystal or ceramic [2,5].

Base of experimental procedure.
The measuring circuit consists of signal generator, frequency meter, vibrator and oscilloscope as a detector. One lead from generator is connected with \( x \)-channel and the second through vibrator with \( y \)-channel of the oscilloscope. When an ac-voltage is applied the Lissajous figures can be observed on the oscilloscope screen. The shape of the Lissajous figure is a function of frequency, for the frequencies far below and far above resonance the impedance of the vibrator is capacitive and if the dielectric loss tangent of the vibrator is small the phase shift will be observed between the \( x \) and \( y \) channels voltages. When the frequency comes near resonance the impedance of the vibrator was influenced by its motional components \( /C_1, L_1, R_1/ \) and draws the circle on the impedance diagram. The voltage across the vibrator is retarded and the phase shift appears between the voltages applied to the \( x \) and \( y \) plates of cathode ray oscilloscope. At frequencies of resonance /\( f_R/ \) and antiresonance /\( f_a/ \) the impedance circle bisect the resistance axe and the voltage across the vibrator is shifted against the angle \( \pi/2 \) relatively to the voltage signal from the generator. The phase angle between the \( x \) and \( y \) channels voltages can be readily determined from the shape of the Lissajous figures.
As the voltage applied to the $x$ plates $V_x = V_0 \sin \omega t$ and the voltage applied to the $y$ plates $V_y = V_0 \sin(\omega t + \phi)$, for $x = 0$, we have $V_y = V_0 \sin \phi$. Then the angle of the phase shift $\phi = \arcsin(V_y/V_0)$ where $V_y$ is the voltage at $x = 0$ and $V_0$ is the amplitude of the voltage signal, both read from the oscilloscope screen. The $y$ channel voltage has a maximum at the frequency of minimum impedance $f_m$ and minimum at the frequency of maximum impedance $f_n$ of the vibrator. Thus, from the amplitude and phase changes of the Lissajous figures the frequencies $f_m$, $f_r$, $f_a$, $f_n$ can be determined. To calculate the piezoelectric properties of the resonator the frequency of series resonance $f_s$ and parallel resonance $f_p$ are required. From the impedance circle results that $f_m(f_s(f_r, f_a, f_n)f_p)$. Thus the frequencies $f_s$ and $f_p$ are known with the errors of $f_r-f_m$ and $f_n-f_a$ respectively.

Experiment.

The piezoelectric, elastic and dielectric properties of crystals grown from the vapour phase [7] have been investigated. As an example the temperature characteristics of dielectric constant $\varepsilon_{33}/T$ - fig.2, electromechanical coupling factor $k_{33}/T$ - fig.2, elastic compliance coefficients $s^{E}_{33}/T$ and $s^{B}_{33}/T$ - fig.3 and piezoelectric modulus $d_{33}/T$ - fig.4 of $\text{GaAs}$ single crystal/sample $0.4 \times 0.4 \times 0.5\text{mm}^3$ are shown. In fig.5 the Lissajous figures observation of the sample $\text{MgO}/K\text{NO}_3/\text{MgO}$ - fig.2 at 10°C is presented. As to results from the shape of the Lissajous figure shown in fig.5b, the frequency $f_2 = 64\text{kHz}$ stimulates the resonator to the vibrations with resonance frequency $f_r = 27.20\text{kHz}$. This observation prove that the frequency $f_2 = 64\text{kHz}$ is not a separate mode and $f_r = 27.20\text{kHz}$ is the fundamental frequency of the vibrator.

Conclusions.

The simple measuring circuit proposed here enable from the shape of the Lissajous figures the fundamental frequency of a vibrator distinguish and to determine the frequencies $f_r$ and $f_p$ with the experimental errors $f_r-f_m$ and $f_n-f_a$ adequately.

References.

Fig. 1 - Temperature dependence of the dielectric constant $\varepsilon_{33}$ /sample 3MN/

Fig. 2 - Temperature dependence of the electromechanical coupling factors $k$ in 3BS7 single crystals /1-M2, 2-M2l, 3-M2n/

Fig. 3 - Temperature dependence of the elastic compliance coefficients $s_{53}^E$ and $s_{53}^R$ /3MN/
**Fig. 4** - **Temperature dependence of the piezoelectric modulus** $d_{33}$ /sample Mn/

**Fig. 5** - **The Lissajous figures versus frequency** /sample Mg, 10°C/:
- a/ $f_a=13,64$kHz
- b/ $f_b=13,16$kHz
- c/ $f_c=25,84$kHz
- d/ $f_d=27,20$kHz
- e/ $f_e=28,73$kHz
- f/ $f_f=34,63$kHz
- g/ $f_g=42,26$kHz
SYNTHETIC DESIGN OF ACOUSTIC FILTER FOR DIGITAL MICROPHONE

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Introduction

As for an acoustic antialiasing filter, design examples for telephone transducer are firstly introduced by J.L. Flanagan in 1979 [1]. We introduce another design for digital microphone [2], intended for professional and commercial audio engineering services, after the synthetic theory of passive networks.

At the design of acoustic filter for microphones, we usually face the following physical restrictions:

(a) Acoustic elements are usually lossless.
(b) The source impedance is small and reactive for wide range of frequency.
(c) Load impedance of the filter must be very high when a transducer of condenser type is used.

In this paper, we introduce a precise design method after the synthetic theory of linear passive networks and a modification process by computer aided design method for realization of acoustic antialiasing filter.

Design Concepts

At the design of the acoustic filter, we take the following policy:

(a) Synthetic method should be used in order to determine the basic structure and element values of filter. A computer aided design for modification of element values should be utilized when the terminal condition becomes too complicated to apply the synthetic method at the higher frequency region.

(b) Some acoustic resistances must be used in the circuits in order to obtain a flat frequency response below the cut-off frequency. The number of resistances, however, should be limited to only one so that we can make the adjustment in production easy.

Synthetic Design Theory

The general structure of the acoustic filter which satisfies the physical restrictions is shown in Fig. 1. The open circuit transmission coefficient \( N_0 \) is given by the following equation:

Conception synthétique de filtre acoustique pour microphone numérique
\[ N_0 = \frac{R_1 + Z_{11}}{Z_{12}} \]  (1)

where \( R_1 \) is artificially added resistance, and \( Z_{11} \) and \( Z_{12} \) are elements of \( Z \)-matrix of the lossless network. As \( N_0 \) is a rational function of \( s \), it is generally represented by

\[ N_0^{-1} = \frac{q(s)}{q(s)} = \frac{m + n}{m_2 + n_2} \]  (2)

where \( m \) and \( n \) are even and odd components of polynomial \( g(s) \) respectively, and \( m_2 \) and \( n_2 \) are even and odd components of denominator polynomial \( q(s) \) respectively.

As \( Z_{11} \) and \( Z_{12} \) are restricted to lossless, they must be ratios of even and odd polynomials. Therefore, \( N_0^{-1} \) should be either

\[ N_0^{-1} = \frac{m}{m_2 + n_2} \quad \text{or} \quad N_0^{-1} = \frac{n}{m_2 + n_2} \]  (3)

instead of the general expression of Eq. (2).

We assume \( |N_0^{-1}|^2 \) is prescribed as a function of frequency. Returning to Eq. (2), the following Eqs. can be given.

\[ |N_0^{-1}|^2 = \frac{q(s) \cdot q(-s)}{q(s) \cdot q(-s)} = \frac{m^2 - n_2}{m_2^2 - n_2^2} \]  (4)

\[ = \frac{A_0 + A_1 s^2 + \cdots + A_p s^p}{B_0 + B_1 s^2 + \cdots + B_p s^p} = \frac{A(\omega^2)}{B(\omega^2)} \]  (5)

Comparing Eqs. (1), (3), we conclude that either

\[ Z_{12} = \frac{m}{n_2}, \quad \text{and} \quad Z_{11} = \frac{n_2}{m_2} \]  (a)

or

\[ Z_{12} = \frac{n}{m_2}, \quad \text{and} \quad Z_{11} = \frac{n_2}{m_2} \]  (b)

should be held.

Combining above Eq. (4) \& (5), we obtain following Eqs.:

\[ B(-s^2) = B(\omega^2) = m_2^2 - n_2^2 \]  (7)

and either

\[ A(-s^2) = A(\omega^2) = m^2 \]  (a)

or

\[ -A(-s^2) = A(\omega^2) = n^2 \]  (b)

Because \( m_2 + n_2 \) must be Hurwitz polynomial for the sake of realizability, its zeros are composed of zeros, which lie on the left half plane, of \( B(-s^2) \).

From Eq. (8), \( m \) and \( n \) can be obtained as

\[ m = \sqrt{A(-s^2)}, \quad \text{and} \quad n = \sqrt{-A(-s^2)} \]  (9)

for both cases of (a), and (b).
Using Eqs. (6) and (8), we can determine $Z_{11}$ and $Z_{12}$ as a rational function form.

Practical Design

As an example of design, a Butterworth lowpass filter for antialiasing has been designed. The filter is composed with an acoustic filter of order five and an electronic active filter of order four in tandem. The electronic filter suppresses the effect of radial mode resonances of the acoustic system at higher frequency region than the cut-off frequency.

The lowpass filter of order nine is expressed as,

$$\left| N_0(j\omega)^{-1} \right|^2 = \frac{1}{(1 + \omega^2)} $$

and its poles which lie on the left half of $s$-plane are given by

$s_v = e^{j(2\pi/18)v}$

$v = 5, 6, 7, 8, 9, 10, 11, 12, 13$.

Of these poles, $S_5$, $S_5$, $S_5$, and $S_5$, $S_5$ are selected as the poles of the fifth order acoustic filter.

Then, we get the following denominator polynomial

$$m_2 + n_2 = s^5 + 2.3473s^4 + 3.6946s^3 + 2.3473s + 1$$

Therefore,

$$m_2 = 2.3473s^4 + 3.6946s^2 + 1$$

$$n_2 = s^5 + 3.6946s^3 + 2.3473s$$

Accordingly, the elements $Z_{11}$ and $Z_{12}$ of open circuit impedance matrix is given by

$$Z_{11} = \frac{2.3473s^4 + 3.6946s^2 + 1}{s^5 + 3.6946s^3 + 2.3473s}$$

$$Z_{12} = \frac{1}{s^5 + 3.6946s^3 + 2.3473s}$$

Expanding $Z_{11}$ into the continued fraction form, we set the following equation.

$$Z_{11} = \frac{1}{0.46260s + \frac{1}{1.1069s + \frac{1}{1.3525s + \frac{1}{2.7566s + \frac{1}{0.5688s}}}}}$$

This result directly leads us the circuits shown in Fig. 2. The frequency response of $|N_5^{-1}|$ which is calculated from the circuit is shown by dotted line in Fig. 4.

A peak at 12 kHz is cancelled by the electronic active filter and Butterworth characteristics is finally obtained. Fig. 3 shows the structure of the acoustic filter in which the acoustic resistance 87.7( cgs) is realized by thin nylon cloth. The frequency characteristics of the acoustic filter which is depicted in Fig. 3 is shown by solid line in Fig. 4 and the
chain line shows the overall characteristics with the electronic filter.

A flat frequency response within passband under 15 kHz and more than 40 dB attenuation in stop-band are realized. Fig. 5 shows a unit of the acoustic filter mounted on a electret microphone, and the outside view of whole microphone head.

Fig. 2 Realization of normalized transfer impedance by continued fraction expansion method.

Fig. 3 Structure of the acoustic filter

Fig. 4 Frequency characteristics of the antialiasing filter

Fig. 5 A microphone unit with acoustic filter, and whole microphone head.

Conclusion

Synthetic design of acoustic antialiasing filter with one resistive element is described. Design example for an electret condenser microphone is also described and it shows flat response below 15 kHz.

We would like to express our appreciation to Mr. Kazuhiko Kajiwara, who contributed his experimental skill during the development of the acoustic filter.

References

LARGE-AMPLITUDE CHARACTERISTICS MEASUREMENT OF BOLT-CLAMPED LANGEVIN TYPE VIBRATOR

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1. Introduction

In order to design such an ultrasonic vibration system for ultrasonic welding of plastics or metals, large-amplitude characteristics of a vibrator used are essential. Such characteristics have been hardly measured due to the lack of a suitable variable mechanical load resistance and the nonlinearity of mechanical vibration loss.

In this paper a variable mechanical load resistance is proposed and large-amplitude characteristics of a bolt-clamped Langevin type vibrator (BLT) are measured employing a new testing method (1) proposed by one of the authors.

2. Principle and Experimental Procedure

Consider the case where the electric input power $W_0$, the vibration velocity $v$, and the terminal voltage $V$ of the vibrator are measured as shown in Fig.1.

Let the dielectric and the mechanical losses of the vibrator be $W_d$ and $W_m$ respectively, and $W_0$ be generally expressed as

$$W_0(v, V, r_a) = W_d(V) + W_e(v) + W_a(r_a, v)$$

(1)

where $W_d$ denotes the acoustic radiation power including mechanical loss of the radiator, $r_a$, the mechanical load resistance looked out from the mechanical terminal of the vibrator. The mechanical and the dielectric losses are assumed to depend only on the vibration velocity $v$ and the applied voltage $V$ respectively.

In order to obtain $W_d$, $W_m$ and $W_a$, three tests, clamped, no-load and load tests should be carried out. The experimental procedure is explained in orderly manner:

(1) CLAMPED TEST

As shown in Fig.2(a), two vibrators which have almost the same vibration characteristics are mechanically connected with
each other at the mechanical terminal. And they are driven at the original resonant frequency, applying the same driving signal not to build up their vibrations. The electric components $L$ and $C$ are adjusted to make the power-factor to be units.

Specifications of the bolt-clamped Langevin type vibrators used for the test are listed in Table 1.

The electric input power $W_e$ measured changing the applied voltage $V$ gives $3W_d$ itself under this condition, because the vibrations of the vibrators cancel with each other and do not build up at original resonant frequency. To obtain $W_e$, the losses of the electric components are subtracted from the value measured by the watt-meter.

(2) NO-LOAD TEST

Easily to support the vibration system at the nodal points, the mechanically connected two vibrator used for the clamped test are also employed for the no-load test as shown in Fig.2(b). In this case only one vibrator is driven electrically and the electric terminal of the other one is kept open and the electric input power are measured at the resonant frequency as a function of vibration velocity. A frequency at which the vibration amplitude shows the maximum value with constant terminal voltage is adopted as the resonant frequency. The vibration amplitude of the mechanical terminal MT is measured by a pick-up calibrated with an optical microscope.

The mechanical vibration loss per one vibrator is calculated according with the equation

$$2W_m(v) = W_e(v, V) - W_d(v). \tag{2}$$

(3) LOAD TEST

As is schematically shown in Fig.1, the electric input power $W_e$, the vibration velocity $v$, and the terminal voltage are measured. A variable mechanical load used for the load test consists of a rod-disk radiator(2) and a reverberation tank as shown in Figs.1 and 3. Varying the effective mechanical resistance $r_a$ is performed by changing the depth of the radiator immersed in the water-filled reverberation tank. The reason of using the reverberation tank is not to build up standing waves in the tank.

The mechanical output power is calculated in accordance with the equation

$$W_a(r_a, v) = W_e(v, V, r_a) - (W_d(v) + W_m(v)) \tag{3}$$

and the equivalent mechanical load resistance $r_a$ and the electroacoustic efficiency $\eta_{ea}$ are expressed as

$$r_a = \frac{W_a(r_a, v)}{v^2} \tag{4}$$

and

$$\eta_{ea} = \frac{W_a(r_a, v)}{W_e(v, V, r_a)} \times 100 \% \tag{5}$$

respectively.
3. Experimental Results and Discussion

As the result of the clamped test, the dielectric loss is proved to be proportional to $\sqrt{v}$ up to $v = 1.5$ kHz and $\tan \delta$ is estimated to be 0.04. This value is somehow a little larger than that of the piezoelectric element itself. The reason of this result may be due to the error of this method.

The mechanical vibration loss is plotted in Fig.4 as a function of the vibration velocity and is proportional to $v^2$. The proportional constant, that is, the equivalent mechanical resistance of the vibrator is about $2 \times 10^4$ dyne/cm/sec. The corresponding mechanical $Q$-factor of the vibrator is estimated to be 800.

Figure 5 shows the mechanical output power as a function of mechanical load resistance $r_m$, where the vibration velocity is used as a parameter. The maximum output power is about 800 watt in this measurement. The available output power of the vibrator is estimated to be larger, but the practical value is constrained by the power limit of the electric power amplifier.

Figure 6 shows the electroacoustic efficiency as a function of the equivalent mechanical load resistance, using the vibration velocity as a parameter.

The efficiency shows the maximum value of about 95% at a mechanical load resistance of about $7 \times 10^5$ dyne/cm/sec, and the optimum mechanical load resistance is about 35 times of the equivalent mechanical resistance of the vibrator.

4. Conclusion

Using the proposed method and the variable mechanical load, large-amplitude characteristics of the high-power ultrasonic vibrator have been measured and both the method and the variable mechanical load are proved to work successfully.

The authors wish to express their gratitude to Prof. S. Kaneko (Shibaura Institute of Technology), Mr. K. Okada, Mr. K. Ohya and Mr. M. Masuda (NGK Spark Plug Co., Ltd) for the co-operation in the experiment.

References

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TABLE 1 Specifications of B.L.T.-D4420PC-
(M=200g)

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<td>equivalent mechanical resistance $\times 10^4$ (dyne/kine)</td>
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Fig. 1 Block diagram of load test. (MT: mechanical terminal)

Fig. 2 Connections of clamped and no-load tests.

Fig. 3 Ultrasonic radiator used variable mechanical load.

Fig. 4 Dielectric loss vs. terminal voltage.

Fig. 5 Acoustic output power vs. load resistance.

Fig. 6 Electroacoustic efficiency vs. mechanical load resistance.
AIRCRAFT NOISE MEASUREMENT USING GROUND PLANE MICROPHONES

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INTRODUCTION

A microphone height of 1.2 m over the ground plane has come to be regarded as the convention in measurement of environmental noise levels and is mandatory for noise certification of aircraft (ICAO 1981). Noise measurements made above a reflecting surface are subject to either cancellation or augmentation due to interference effects between the incident and ground-reflected waves, resulting in distortion of the original spectral shape. For sound at normal incidence to the ground, interference effects occur at frequencies below approximately 1 kHz. At higher frequencies the sound-field derives from energy addition of direct and reflected waves and the resulting increase over the free-field value is dependent on the acoustic impedance of the ground surface. A number of methods to overcome this problem have been proposed. One involves the use of alternative microphone heights to shift the interference effects out of the frequency range of interest and another attempts to remove the effect of the reflected wave altogether by placing the microphone in the ground plane.

In this paper spectral sound pressure levels from several ground plane microphone configurations are compared with pressure-doubled levels derived from the 1.2 m microphone position. Experimental data have been obtained in the laboratory and in the field using a stationary sound source producing both pure tones and band limited noise.

LABORATORY MEASUREMENTS

Three microphone configurations were examined in a semi-anechoic room using a point sound source at a fixed height of approximately 1.15 m over the reflecting floor surface of the room. Pure tones were generated and swept over the range 0.1 kHz to 10 kHz. The microphone configurations examined were:

a) half-inch microphone lying on its side
b) half-inch inverted microphone - the gap between the diaphragm and the reflecting surface was varied over the range 1 mm to 10 mm
c) half-inch microphone fixed at the centre of a circular baffle with its diaphragm flush with the baffle surface.

For (a) and (b) the floor was covered with 6 mm thick hardboard, while for (c) the covering was 25 mm thick polyurethane foam. In addition to the microphone under investigation a second microphone was secured from beneath
the floor of the room with its diaphragm flush with the surface of the floor covering. The output of this microphone was assumed to represent the effect of pressure doubling on the incident sound signal.

Microphone configuration (a) showed no significant departure from pressure doubling up to a frequency of approximately 1 kHz. Above this frequency the microphone output decreased steadily, resulting in differences from pressure doubling of 2 dB at 4 kHz rising to 16 dB at 10 kHz. This decrease in output was the result of interference and was readily estimated. For a sound wave of unit amplitude incident on a perfectly reflecting surface the pressure amplitude at a point above the surface is given by:

\[ 1 + \cos \left( 2\pi f x / c \right) \]  

where \( f \) is frequency, \( c \) is velocity of sound and \( x \) is the path difference between the incident and reflected waves. At present correction factors sufficient to describe the diffraction effects associated with microphone configurations close to a reflecting surface are not available and so Equation 1 has not been modified to include these effects. Differences from pressure doubling calculated using Equation 1 assuming that the microphone is effectively one-half the diaphragm diameter above the reflecting surface were, however, within 0.4 dB of the measured differences discussed above.

Microphone configuration (b) again showed no significant departure from pressure doubling up to approximately 1 kHz. Using Equation 1 to estimate the change in microphone output at frequencies greater than 1 kHz, with \( x \) equal to twice the gap width, proved unsuccessful. At frequencies above 4 kHz for the smaller gap widths the predicted interference cancellations did not occur, in fact an increase in microphone output was observed. This increase was probably due to some form of resonance in the gaps.

For microphone configuration (c) the floor of the anechoic room was covered with sheets of polyurethane foam to provide a change of acoustic impedance between the baffle and the ground surface; in out-door measurements the baffle is likely to be placed on a grass or soil surface. The output of this microphone was compared with predictions after Taylor (1978). The microphone output oscillated about the pressure-doubled value by approximately ± 3 dB at all frequencies up to 10 kHz. The agreement between experimental and theoretical results was sufficient to validate Taylor's approach. A subsidiary experiment to examine the effect of a vertical separation between the board and the ground surface highlighted the importance of board placement, as shown by Pernet and Payne (1976). For aircraft noise measurement, the effect of diffraction must be estimated for noise signals. This considerably reduces the variation from pressure-doubling at higher frequencies. The magnitude of the oscillations is reduced still further when values of acoustic impedance relating to grass covered surfaces are used. As a general conclusion these laboratory measurements did not indicate any one configuration to be ideal but on the other hand none were shown to be completely without merit. It was decided therefore to continue by examining their performance further, in the field.

FIELD MEASUREMENTS

A loudspeaker was supported at a height of approximately 6.5 m above the ground; the ground cover was a dense layer of grass cut to about 3 cm. White noise was shaped using a multifilter in order to obtain a reasonably
flat spectrum in the far field on the axis of the loudspeaker. This signal was measured near the ground using a number of different microphone configurations which included the standard 1.2 m height, and half-inch microphones positioned close to the surface of a 0.4 m diameter metal baffle. The ground-based microphone configurations were similar to those used in the laboratory experiments described above, referred to as (a), (b) and (c). In (b) the microphone was supported by a small tripod with the gap fixed at 4 mm. For (c) a hole was bored into the ground under the baffle to accommodate the pre-amplifier body. In all cases the microphone was positioned at one-quarter radius from the edge of the baffle (diffraction effects are much less evident at this position than at the centre of a circular baffle). Two baffles were used in turn, 2 mm thick aluminium and 1 cm thick brass. Both baffles were positioned on the grass surface. In addition, for the brass baffle, a 0.4 m diameter turf was cut and the disc placed firmly on the underlying soil. A separate arrangement was also examined in which the grass surface was covered with 6 mm thick hardboard such that the measurement microphones were at least 2.5 m from a board/ grass interface.

The output from the 1.2 m microphone was used to estimate a free field spectrum, using the procedure described in the Engineering Sciences Data Unit Item Number 80038 (ESDU 1980) to allow for ground reflections. The ground surface was assumed to be represented by a flow resistance of 1000 kPas/m² (Payne 1982). As an estimate of a pressure-doubled spectrum at the surface of the baffle, 6 dB was added to the calculated free field values. The spectra resulting from the analysis of the baffle microphones were compared to this pressure doubled spectrum.

For all the experimental arrangements described above there was good agreement between the output of the three ground-based systems below approximately 3.15 kHz, all three lying within a 0.5 dB window down to 50 Hz. Above 3.15 kHz there were variations between systems due to interference and microphone directivity, similar to those observed in the laboratory. The deviation from pressure doubling for the aluminium disc data showed for all three configurations a trough between 315 Hz and 1.6 kHz with differences of approximately -2 dB and a peak between 1.5 kHz and 4 kHz of up to +2 dB. The brass disc data showed similar deviations but of slightly reduced magnitude (-1.5 dB to +1.5 dB). However when the turf was cut and the brass disc placed firmly on the underlying soil these variations were eliminated and the measured spectra became similar to those observed when the grass surface was covered with hardboard, where good agreement between ground-based microphone output and pressure doubled levels were obtained. It was first thought that the peak and trough were due to some vibration/ resonance effect associated with the light aluminium disc resting on top of the grass surface but an examination of data from the heavy brass disc experiments points more toward errors caused by a vertical separation between the disc and the effective reflecting surface.

Above 4 kHz the average departure from pressure doubling for the microphone on its side, configuration (a), was the same as that measured in the laboratory. There was however a variation between experimental arrangements such that the deviation from pressure doubling at 10 kHz, for instance, could only be predicted, using Equation 1, to within 3 dB. High frequency deviations for the inverted microphone, configuration (b), had a mean value at 10 kHz of -1.5 dB. In the laboratory a positive difference
was observed. This field result compares more favourably with a calculated value of -5 dB. The slight improvement is thought to result from the reduced effect of resonance in the gap when noise signals are used. In order to minimise possible calibration errors the same microphone, a B & K type 4133, was used for all ground-based systems. The data from configuration (c) were therefore corrected using manufacturer's pressure correction factors. This having been done the mean deviation from pressure doubling at high frequencies amounted to approximately 1.5 dB at 10 kHz.

CONCLUSIONS

When care was taken to locate the baffle with its surface firmly in the plane of the ground surface all three configurations produced reasonable estimates of the pressure-doubled spectrum (free field +6 dB) for frequencies below 3.15 kHz. Above this frequency deviations due to interference and microphone directivity occurred which could to a limited extent be predicted but further work in quantifying microphone directivity correction factors is needed. The effect of any one of the three systems on overall measures of aircraft noise (e.g. PNL) will be dependent on the distribution of energy in the noise spectrum (Payne 1982). The interference effects observed at high frequencies may therefore be important for measurement of noise levels from jet aircraft but it is thought that ground based microphone configurations may be suitable for light piston engined aircraft or helicopters where the one-third octave bands which are dominant in overall measures generally occur at somewhat lower frequencies. It is therefore intended to compare the results of these current measurements with aircraft noise spectra and to expand the present experimental program to include field trials using flyovers of light propeller aircraft.

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6.2


Oscillators. Amplifiers. Attenuators. Lines

SUPPRESSION OF HOWLING BETWEEN MICROPHONES AND MONITORING SPEAKERS - AN APPLICATION OF AN ADAPTIVE FILTER -

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1. Introduction
When a feedback monitor-speaker is placed in a satellite studio or any place where sound is picked up by a microphone and relayed to a main station, a signal loop is made as shown in Fig.1, which causes howling. In order to prevent this howling, adjustment usually must be made to the level of the monitor-speaker, the frequency characteristics of the monitoring device, and/or the directivity of the microphone. If a transmission line, which can be used specifically for monitoring, is available from the main studio to the satellite studio, the howling can be suppressed by performing a suitable mixing operation at the main studio. Such a line is, however, often unavailable in actual situations.

In this paper we propose a new method to suppress the howling caused by a feedback monitor-speaker using an adaptive filter. We constructed a howling suppresser which works in actual situations. The result of an experiment in a simulated broadcasting system is presented here.

2. Principle, algorithm and hardware
Fig.2 shows the principle of the method proposed here. The output of a microphone set at a relaying point is subtracted, after equalization by an adaptive transversal filter,

Fig. 1 Howling caused by a feedback moniter-speaker
from the output of a monitor receiver before it goes into a monitor-speaker placed at the relaying point. In the figure, \( \hat{H} \) represents the transfer function of the path from the output point of the microphone to the output point of the monitor-receiver via the transmission line and the transmitter.

The transfer function of the adaptive filter is represented by \( \tilde{H} \). The adaptive filter adjusts itself so that \( \tilde{H} \) approximates \( \hat{H} \). It is obvious that if \( \tilde{H} \) coincides with \( \hat{H} \), the signal picked up by the microphone is completely cancelled by subtraction and does not come out of the monitor-speaker, so the howling loop in Fig.1 is cut. Other signals mixed at the main studio come straight out of the monitor-receiver and can be monitored intact.

A number of algorithms have been proposed to adjust the coefficients of the adaptive filter automatically, the most widely known and used being the LMS-algorithm proposed by Widrow et al. (1). The LMS-algorithm requires a relatively small amount of computation and is easy to implement. It has, however, a rather low speed of convergence and is not suitable for our purpose, because we would like the machine to work as soon as a broadcast program goes on the air. For this reason, we adopted a faster algorithm called the Learning Method, proposed by Nagumo et al. (2). In the Learning Method, the filter coefficient vector

\[
|h|=(h^{j}_1, h^{j}_2, \ldots, h^{j}_K)^T
\]

is revised in the following manner:

\[
|h|_{j+1} = |h|_j + \alpha \cdot e_j \cdot x_j / \|x_j\|^2
\]

where

\[
x_j = (x_j, x_{j-1}, \ldots, x_{j-(n-1)})^T
\]

\[
\|x_j\|^2 = \sum_{i=0}^{n-1} x_{j-i}^2
\]

and \( \alpha \) is a parameter to control the speed of convergence. Although this algorithm requires about two times more amount of computation than the LMS-algorithm, it has favorable properties, such as the convergence is faster than the LMS-algorithm, and the monotonicity of convergence is guaranteed for \( 0 < \alpha < 2 \).

We have also tried a modified Learning Method in which the coefficient vector is divided into \( N \) sections

\[
|h| = (h^{j}_1, h^{j}_2, \ldots, h^{j}_K), \quad h^{j}_k = (h^{j}_{p+1}, h^{j}_{p+2}, \ldots, h^{j}_{p+M}), \quad M = (h^{j}_{p+1}, h^{j}_{p+2}, \ldots, h^{j}_{p+M})^T
\]

and each section is revised in turn in each sampling period as

\[
|h|_{j+1} = h|_k + \alpha \cdot e_\cdot x_{j,k} / \|x_j\|^2, \quad \text{where } x_{j,k} = (x_{j-(k-1)p+1}, x_{j-(k-1)p+2}, \ldots, x_{j-kp+1})^T
\]
By doing this, we can limit the amount of computation required within a sampling period. In other words, we increase the number of filter stages within the limited computational ability of the machine. Although there is no theoretical proof as to the convergence of this algorithm, we have found that the convergence is obtained for most common signals. The speed of convergence in this modified algorithm was found to be inversely proportional to N(3).

Fig. 3 is a block diagram of the howling suppresser we have constructed. The central part is a high-speed microprocessor (Miproc-16) which executes instructions in 250μS, and attached to it are an A/D converter, a D/A converter and a high-speed arithmetic unit which was specifically designed for the filtering operation. Multiplication and division are also performed in the high-speed arithmetic unit. The revision of the filter coefficients is performed on the side of the microprocessor so that various algorithms can be tried by programming, and the resulting new coefficients are transferred to the high-speed arithmetic unit. The precision of the A/D, D/A converters is 12-bit. In the machine, operations are performed in 16-bit fix-point arithmetic.

Through the use of this machine we can realize an 82-stage adaptive filter with the Learning Method working in real time, and a 152-stage adaptive filter with the modified Learning Method when the sampling frequency is set at 7.5kHz.

3. Experiment

Fig. 4 is a block diagram of our experiment. In the figure, the dummy transmission line is a filter having about 40μS of group delay in the lower frequency region simulating an actual cable, the transmitter is a 90MHz FM transmitter and the monitor-receiver is a commercially available FM receiver. This is a good approximation of a real situation.
In order to see the effectiveness of the howling suppressor in each frequency region, a 1/3 octave band pass filter was inserted in the howling loop and the loop gain on the verge of howling was measured for various center frequencies of the band pass filter. Fig. 5 is the result of the measurement, showing that by using the howling suppressor, the loop gain can be raised more than 10dB in most frequency regions. Measured in the full band, improvement of the loop gain was about 11dB, and the sound level of the monitor-speaker could be raised by the same amount.

Signals mixed at the main studio act as a disturbance to the adaptive filter (3), and this results in distortion of the monitoring sound. In order to prevent this distortion, the coefficients of the adaptive filter are frozen when the level of signals mixed at the main studio exceeds that of the output signal of the microphone.

4. Concluding Remarks
A howling suppressor was constructed using an adaptive filter, and the effectiveness was confirmed in an experiment using a simulated broadcasting system. Currently we are planning the following for further research:
(1) A test of the machine in a real broadcasting system.
(2) Elaboration of the decision rule of when to freeze the filter coefficients.
(3) Design and construction of a higher speed machine so that the frequency band width can be expanded and an adaptive filter with more stages can be implemented.

References
6.3

Enregistrement et reproduction
Recording and reproduction of sound
Aufnahme und Wiedergabe
LA RELATION ENTRE LA RÉPARTITION DE L'INTENSITÉ DANS L'ENTREFER ET LA CARACTÉRISTIQUE EN FRÉQUENCE CHEZ L'ENREGISTREMENT DU SIGNAL

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Introduction
En cas de l'enregistrement du signal par le faisceau de la lumière, d'électrons ou de l'agent la caractéristique en fréquence dépend de la forme de l'entrefer dans lequel l'intensité a l'influence sur le support d'enregistrement et aussi de la répartition de l'intensité dans la superficie de l'entrefer. Il en va de même en cas de lecture de l'enregistrement par la manière optique.

La méthode pour calculer la caractéristique en fréquence
On peut calculer la caractéristique en fréquence des valeurs de l'intensité par la méthode suivante.
Les points dans la superficie de l'entrefer dans lequel l'intensité a l'influence sur le support d'enregistrement, peuvent être marqués par la coordonnée $\xi$, qui est orientée dans la direction du mouvement du support d'enregistrement, et la coordonnée $\eta$ qui est perpendiculaire. Par la représentation graphique des valeurs de l'intensité sur les coordonnées $\xi$ et $\eta$ on obtient le corps de l'intensité dans lequel on peut faire les coupes parallèles avec les axes $\xi$ et $\eta$. Par exemple on fait l'enregistrement du son dans la technique cinématographique avec l'entrefer du rectangle, qu'on peut voir sur la figure 1 avec les coupes dans les axes.

La coupe $I(\eta)$ pour $\xi = \xi_i$.

Fig. 1.
Pour la calculation de la caractéristique en fréquence est décisive la répartition de l'intensité dans l'direction du mouvement du support d'enregistrement, qu'on appelle I(ξ). Nous supposons d'abord que dans la direction η sont les valeurs d'intensité constantes dans chaque coupe, alors I(η) = const. Il faut déterminer l'influence I(ξ) sur le spectre de l'enregistrement ou dans le cas de lecture sur le spectre du signal, On doit d'abord transformer la fonction I(ξ) dans l'impulsion du temps I(t) suivant le mouvement relatif entre le support de l'enregistrement et l'enterrer, qui est indiquée par \( v \). Puis on doit replacer la coordonnée ξ par le temps t. La durée de l'impulsion du temps d sur la calcule selon \( d = b/v \). La transposition est sur le figure 2.

![Fig. 2.](image)

Cette impulsion du temps sert de l'object pour la trans-
formation du Fourier, par laquelle on obtient le spectre. Les valeurs absolues du spectre en dépendance sur la fré-
quence déterminent la fonction du spectre duquel peut dé-
rivée la caractéristique en fréquence. La transformation

\[
F(j\omega) = \int_{-\infty}^{\infty} I(t) e^{-j\omega t} dt
\]

Les amplitudes du spectrum \(|F(\omega)|\) sont données par les
valeurs absolues de la fonction \( F(j\omega) \) pour les fréquen-
ces \( \omega \). Pour le calcul de ces valeurs en décibels, on ob-
tient la caractéristique en fréquence, qu'on a cherchée.

À présent on peut utiliser les ordinateurs électroniques,
pour le calcul de cette caractéristique. La fonction I(ξ)
est donnée par les valeurs discrètes et puis l'impulsion
du temps I(t) est discrète aussi, Pour calculer la trans-
formation du Fourier on utilise l'algorithme de la vite
transformation du Fourier connue comme FFT. La fonction
du spectre est aussi discrète et ses amplitudes peuvent
être calculées comme la deuxième racine de la somme des
carrées de la part réelle et imaginaire.

L'exemple le plus simple en est le cas quand les valeurs
deg l'intensité sont les mêmes dans toute la Superficie de
l'enterrer. On peut indiquer ces valeurs comme A. Puis
I(ξ) = A pour \(-b/2 < ξ < b/2\), et I(t) = A pour \(-d/2 < t < d/2\).
La transformation du Fourier dans ce cas sera

\[ F(j\omega) = \frac{\delta}{j\omega} e^{-j\omega t} = \frac{A}{-j\omega} \left[ e^{-j\omega \frac{\delta}{2}} - e^{j\omega \frac{\delta}{2}} \right] = A \frac{\sin \omega \frac{\delta}{2}}{\omega \frac{\delta}{2}} \]

\[ \omega \frac{\delta}{2} = 2\pi f \frac{b}{2} = \pi \frac{b}{\lambda} \]

\[ F(j\omega) = A \frac{\sin \frac{\pi b}{\lambda}}{\frac{\pi b}{\lambda}} \]

Ce résultat contient la fonction d’entrefil, qui était déjà connue dans la théorie de l’enregistrement du son pour les cas, quand l’intensité dans l’entrefil a les mêmes valeurs.

Cette méthode peut être élargie pour les exemples avec les valeurs différentes dans les coupes \( I(\eta) \) pour l’entrefil rectangulaire. Ce sont les cas où on peut trouver dans les cas pratiques de l’enregistrement du signal. Puis on doit d’abord calculer dans les coupes \( I(\eta_{1}, \eta_{2}, \ldots) \) les valeurs moyennes

\[ I(\xi_{1}, \xi_{2}, \ldots) = \frac{1}{a} \int_{-\frac{\delta}{2}}^{\frac{\delta}{2}} I(\eta_{1}, \eta_{2}, \ldots) d\eta \]

où pour les valeurs discrètes, quand on utilise l’ordinateur, on calcule la moyenne arithmétique. On obtient les valeurs discrètes \( I(\xi_{1}, \xi_{2}, \ldots) \) comme la fonction de \( \xi \).

C’est notre fonction \( I(\xi) \) et le calcul ultérieur est le même comme auparavant avec l’usage du procédé FFT. Le résultat de ce calcul est la caractéristique en fréquence du procédé de l’enregistrement du signal ou de lecture.

L’influence de la forme de l’entrefil

Dans quelques cas, par exemple la lecture des video-discs ou de l’enregistrement digitale du son (le système CD), l’entrefil a la forme du cercle. Pour ce cas on peut aussi appliquer notre méthode. Tout d’abord, on fait des coupes dans le corps d’intensité, qui sont perpendiculaires vers le mouvement du support de l’enregistrement et ainsi on obtient \( I(\eta_{1}, \eta_{2}, \ldots) \) pour les coordonnées \( \xi_{1}, \xi_{2}, \ldots \). Dans chaque coupe on doit calculer la valeur moyenne corrigée sur la longeur maximum de l’entrefil dans la direction parallèle avec \( \eta \). Cette longueur est indiquée comme \( d \) et puis

\[ I(\xi_{1}, \xi_{2}, \ldots) = \frac{1}{d} \int I(\eta_{1}, \eta_{2}, \ldots) d\eta \]

On ajoute ces valeurs corrigées \( I(\xi_{1}, \xi_{2}, \ldots) \), qui sont discré-
tes à $\hat{\gamma}$ et ainsi on obtient la fonction $I(\hat{\gamma})$. La calcul suivant est le même comme dans le cas précédent avec l'usage de la procédé FFT.

On peut voir, que cette méthode et les applications dans les cas de l'enregistrement du signal ou de lecture par la manière optique permet de déterminer la caractéristique en fréquence de ce procédé, qui est donnée par la répartition de l'intensité dans l'entrefer pour les différents cas pratiques.
DIGITAL SOUND REPEATER

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Introduction

Current broadcasting sound repeaters using cartridge tapes are supposed to be replaced by digital systems using solid memories, in the aspects of durability and maintainability.

We have developed a digital sound repeater which may replace the conventional analog system, using reliable magnetic bubble memories which are nonvolatile and need no battery back-up, instead of using a mechanism or magnetic tapes which will be soon worn away. The SDPCM (sectional DPCM) coding method[1] which is better than the DPCM and the likes in the sound quality, and which meets the requirement of broadcasting sound quality has been adopted into the system. Newly developed digital sound repeater (DSR) is being used as an ID signal generator for the satellite communication lines.

Data Compression by Means of SDPCM

The SDPCM coding method is a kind of linear predictive coding in which a discrete signal series is divided at certain intervals and multiplied by a scale factor $\gamma r$ ($r$ indicates the section number), and several non-linear quantization function is used for each section selectively according to the standard deviation of the residual signal of the corresponding section, in order to quantize the residual signal itself.

By using this method, a 20kHz-sampled 12-bit PCM music signal is compressed to 6 bits per sample and a 10kHz-sampled 12-bit PCM speech signal to 4 bits per sample while maintaining the broadcasting sound quality.

Fig. 1 (a) and (b) show the relationship between the number of predictors and the prediction gain. As seen from Fig. 1, four predictors are suitable for the music signal and ten predictors for the speech signal. However, four predictors have been adopted for both signals to save the processing time. The update cycle of predictive coefficients is 15.25ms for the music signal or 23.2ms for the speech signal.

Répéteur de son numérique.
according to the data format in the bubble memories. In order to enlarge the dynamic range, the quantization function for the predictive residual signal was set in four kinds instead of conventional three kinds.\(^{(1)}\) The quantization function is obtained by using the quantizing method for minum distortion.\(^{(2)}\)

Fig. 2 shows the S/Nd of a 1 kHz-sine wave signal. It is maximum 61dB for the music-mode and 52dB for the voice-mode. As seen from Fig. 2, in spite of the 6- or 4-bit coding system, reproducibility of the waveform is excellent in the low level range, and the dynamic range is equivalent to that of the 11-bit PCM system.

According to the computer simulation for music signals, the S/Nd is 38.5dB\(^{-}\)48.5dB, the entropy of the original signals when the signal level occurrences are assumed to be independent one another is 9.8 bits/sample \(\times\) 10.7 bits/sample, and that of the compression signals is 5.38 bits/sample. Hence, the data compression rate for the music signals is about 1/2. Concerning speech signals, the S/Nd is 24.9\% 28.3dB, the entropy of the original signals is 8.1 bits/sample \(\times\) 9.4 bits/sample, and that of the compression signals is 3.3 bits/sample. Hence, the data compression rate for the speech signals is about 1/3. The bit rate after data compression including control data is 134.3kbps for the music signal and 41.4kbps for the speech signal.

System Configuration and Functions of the DSR
Table 1 shows the specifications of the digital sound repeater. Fig. 3 shows the system configuration.

The data which have been compressed previously by the host computer are stored into BBM\(^{\circ}\) 3 (magnetic bubble memories). As the maximum 37ms is required to access the bubble memories, four-8KB buffers are provided to perform simultaneous and parallel playback from A and B channels in the mixed mode of music and voice by sharing the BBM data. Four-1MB bubble memories are controlled in parallel to increase the read-out speed up to 400kb/s.

The system permits playback up to 16 files. Selection of the music or message to be put out to A or B channel, and assignment of the file sequence can be programmed by operating the panel. The data file can be extended in 4Mb-units. However, with the memories of 4Mb, the maximum playing time is 31.2 seconds (for music signal only) or 95 seconds (for speech signal only).

Performance Evaluation
The ripple of frequency characteristic within the frequency bandwidth shown in the table 1, is less than 1.5dB for both music-mode and voice-mode. The distortion factor is 0.16% \(\times\) 0.22% at 0dBs output for the music-mode or 0.24% \(\times\) 0.38% for the voice-mode. The crosstalk between A and B channels is -50.6dB \(\sim\) -50.7dB.
The segmental S/Nd of the reproduced sound by the DSR which is being used as an ID signal generator for the satellite communication lines is about 31dB for the speech sound. Almost no difference is found in languages (English, French and Japanese), or speakers. These values are 2-3dB better than the simulated values. This is caused by the optimum recording level for the original speech sound.

Conclusions
We developed a digital sound repeater using the SDPCM coding system and magnetic bubble memories. A 12-bit PCM music signal was compressed to 6 bits/sample, and a speech signal was compressed to 4 bits/sample. The dynamic range was equivalent to that of the 11-bit PCM system for music or speech signal. The maximum S/Nd (1kHz sine wave) was 61dB (music mode) or 52dB (speech mode). These values met the requirement of broadcasting sound quality. The maximum playing time with 4Mb memories was 31.2 seconds (for music signal only) or 95 seconds (for speech signal only).

The system is now used as an ID signal generator for the satellite communication lines. The segmental S/Nd of the reproduced sound by the system is about 31dB in average. Almost no difference was found in languages (English, French and Japanese), or speakers.

References
(1) Y. Arai, "Sectional DPCM and its application to the speech reproduction systems", Report of the 9th ICA, R1, Madrid, Spain, 1977
**Fig. 2 (a)** Signal-to-distortion-noise ratio against sinusoidal signal levels (Music mode)

**Fig. 2 (a)** Rapport de signal-à-distorsion-bruit selon niveaux de signaux sinusoidaux (Mode musique).

**Fig. 2 (b)** Signal-to-distortion-noise ratio against sinusoidal signal levels (Voice mode).

**Fig. 2 (b)** Rapport de signal-à-distorsion-bruit selon niveaux de signaux sinusoidaux (Mode parole).

**Fig. 3** System configuration of the digital sound repeater using SDPCM

**Fig. 3** Configuration du système de répétiteur de son numérique en utilisant SDPCM

<table>
<thead>
<tr>
<th>Item</th>
<th>M mode (music)</th>
<th>V mode (voice)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Data compression coding</td>
<td>6 bit SDPCM</td>
<td>4 bit SDPCM</td>
</tr>
<tr>
<td>Sampling frequency</td>
<td>20 kHz</td>
<td>10 kHz</td>
</tr>
<tr>
<td>Playback time</td>
<td>30 seconds</td>
<td>95 seconds</td>
</tr>
<tr>
<td>Number of files</td>
<td>15ch (min. 40kHz/sec)</td>
<td>15ch (min. 1.48kHz/sec)</td>
</tr>
<tr>
<td>Frequency band width</td>
<td>8kHz ~ 8.5kHz</td>
<td>100kHz ~ 4.2kHz</td>
</tr>
<tr>
<td>S/Nd</td>
<td>more than 40dB (1 kHz)</td>
<td>more than 50dB (1 kHz)</td>
</tr>
<tr>
<td>Output</td>
<td>Simultaneous output into 2 channels, A and B, 600Ω load + 22kHz/0Ω/8kΩ</td>
<td></td>
</tr>
</tbody>
</table>

**Table 1** Specifications of the digital sound repeater

**Tableau 1** Caractéristiques du Répétiteur de son numérique
BEWEGUNGSGEGENKOPPLUNG EINES DYNAMISCHEN LAUTSPRECHERS IM ZEITMULTIPEX

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1. Prinzip und Ziel der Bewegungsgegenkopplung

Das Prinzip der Bewegungsgegenkopplung (BGK) ist in Fig.1a dargestellt. Eine der mechanischen Bewegunggröße (z.B. v) proportionale Bewegungsspannung \( U_B \) wird über einen Rückverstärker \( V_R \) in geeigneter Größe und Phasenlage der Eingangsspannung \( U_E \) hinzuaddiert; die Bewegungsspannung wird mit \( U_R \) verglichen, Abweichungen des beabsichtigten Funktionsverlaufs werden korrigiert. Diese Korrekturen erstrecken sich vor allem auf drei Aspekte der Übertragung:

a) das Einschwingverhalten der Membran. Die Membran ist ein mehr oder weniger stark bedämpfter mechanischer Schwingkreis, der auf impulsartige Signale normalerweise mit der Erregung seiner Resonanzfrequenz antwortet. Durch die BGK können die Eigenschwingungen sehr stark unterdrückt werden. Gleichzeitig erzwingt die BGK eine Memranbewegung, die dem elektrischen Eingangssignal \( U_E \) auch dann entspricht, wenn der Membrannasse plötzliche Schnelleänderungen abverlangt werden.

b) das Verzerrungsverhalten der Memran. Prinzipiell werden Verzerrungen beseitigt, die z.B. durch Nichtlinearitäten der Aufhängung oder durch Störfunktionen eines mit einem geringen Leck behafteten Gehäuses hervorgerufen werden. Nichtlinearitäten des 'inneren Wandler' (z.B. Magnetfeldverzerrungen) werden nur dann eliminiert, wenn die Bewegungsspannung durch einen separaten Wandler abgeleitet wird.


2. Bekannte Methoden zur Erzeugung der Gegenkopplungsspannung

Es sind verschiedene Schaltungen zur Gewinnung einer Bewegungsspannung bei einem dynamischen Lautsprecher angegeben worden. Sehr bekannt ist die Brückenschaltung (Fig.1b), bei der mit Hilfe einer Lautsprechernachbildung \( L_n \) des festgebremsten Sy-
stems im Brückenzweig die Bewegungsspannung direkt abgegriffen werden kann /1/. Sie wird durch das Magnetfeld des inneren Wandlers erzeugt. Die Schaltung hat den Vorteil, daß am Lautsprecher keine Modifikation vorgenommen zu werden braucht, den Nachteil, daß sehr hohe Anforderungen an die Genauigkeit der Nachbildung gestellt werden müssen. Einfacher zu handhaben sind Schaltungen, bei denen die Bewegungsspannung durch einen separaten Wandler abgegriffen wird, z.B. durch einen Beschleunigungsaufnehmer auf der Membran /2/. Dies hat den Nachteil, daß man einen Speziallautsprecher benötigt, dessen Eigenschaften bedingt durch die Bauart (Wandler, seismische Masse) - zunächst schlechter als die eines Üblichen Systems sind.

3. Zeitmultiplexverfahren

Eine weitere Möglichkeit besteht darin, zur 'Rückmeldung' der Bewegungsspannung das Zeitmultiplexverfahren anzuwenden. In Fig.2a ist das Prinzip dargestellt, das darin besteht, die Speisesspannung U_L nur getastet an die Lautsprecherklemme anzulegen und in den 'Austaststücken' die an den Lautsprecherklemmen anstehende Bewegungsspannung U_B zu messen. Die Tastung des Speise- und Rückmeldekanals erfolgt durch zwei in Antiphase betriebene Schalter, deren Schalterfrequenzen ausreichend hoch zu wählen ist. Der störungsfreie Zeitmultiplexbetrieb setzt zunächst voraus, daß die Lautsprecherlast reell ist. Nur in diesem Fall kann rechteckförmig getastet werden. Dies ist notwendig, wenn die Bewegungsspannung störungsfrei - d.h. ohne 'Übersprechen' der Speisesspannung U_L - gewonnen werden soll, Fig.2b, obere Skizze. Tatsächlich ist diese Voraussetzung nicht erfüllt. Infolge der Blindanteile (z.B. Schwingspuleninduktivität) kann der Spulenstrom nicht abrupt ausgeschaltet werden. Um die Tastung trotzdem zu ermöglichen, sind besondere Maßnahmen erforderlich, z.B. durch Erzeugung eines negativen Zusatzimpulses den Strom möglichst schnell auf Null zu bringen und die Abfrage von U_L dann erst zu vollziehen, d.h. kurz vor der erneuten Speisung, Fig.2b, untere Skizze. Die Rückführung der Bewegungsspannung erfolgt prinziell wie in Fig.1a, zusätzlich jedoch über einen Tiefpaß TP, Fig.2a.


4. Erste Meßergebnisse

Es wurde eine Zeitmultiplexschaltung aufgebaut, bei der das Signal dem Brückenzweig entnommen wurde /3/. Als Lautsprecher nachbildung diente ein festgebremster Lautsprecher gleichen Typs. Die Schaltfrequenz betrug 38 kHz. Die schnelle Stromunter-
brechung bei der Austastung wurde durch ein Differenzierglied
der größten Zeitkonstante zusätzlich bewirkt, welches dem Endver-
stärker vorgeschaltet wurde.
Die Schaltung war bei einem Betriebsbereich bis 500 Hz bis zu
einem Gegenkopplungsgrad von max. 23 dB stabil. Für den Betrieb
wurde ein Gegenkopplungsgrad von 14 dB eingestellt. Bei dieser
Einstellung konnte selbst bei einem 13 cm-Lautsprecher mit
kleiner Wandlerkonstanten ein gutes Betriebsverhalten erreicht
werden. Fig. 3 zeigt die Stoßantwort des Lautsprechers auf den
Aufprall einer 10 g schweren Kugel aus definierter Höhe. Ge-
esen wurde die Bewegungsspannung a) ohne und b) mit Gegenkop-
plung. Der Gegenkopplungsgrad betrug im Bereich der Resonanz-
frequenz (45 Hz) 14 dB. Der Kurvenverlauf zeigt bei eingeschal-
teter Gegenkopplung ein fast aperiodisches Verhalten. In Über-
einstimmung mit dem Höreindruck wird die Membraneiggenschwingung
sehr gut unterdrückt.
Fig. 5a, b zeigen die Bewegungsspannung der Membran a) ohne und
b) mit Gegenkopplung bei einem dreieckförmigen Eingangssignal
(45 Hz). Auf dem unteren Teil des Bildes ist jeweils die Spei-
sespannung U_s oszillographiert. Man sieht deutlich, daß die
abrupte Schnelleänderung der Membran durch einen entsprechen-
den Anstieg der Speisespannung erzwungen wird. Der Fig. 4 ent-
nimmt man, wie die Membranbewegung auf ein plötzliches Anstoßen
an ein Hindernis reagiert. Es wurde vor die Membran ein Blei-
stift gehalten; die ansteigende Speisespannung (untere Skizze)
sucht die sinusförmige Lautsprecherbewegung aufrecht zu erhal-
ten; die Oszillation rührt von prinzipiell vermeidbaren Resonan-
zeiten im Rückkopplungszweig her. Klirrfaktormessungen zeigten im
Bereich der Membranresonanz eine Verminderung der Verzerrungen.
Das generelle Verhalten der Bewegungsgegenkopplung ge-
genüber Verzerrungen und dem Dopplereffekt bei unterschiedli-
chen Randbedingungen bedarf jedoch noch einer weiteren Unter-
suchung. Die bisherigen erfreulichen Meßresultate rechtfertigen
die weitere Beschäftigung mit der Bewegungsgegenkopplung.

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sehen Wandlers als Sender und Empfänger durch
Zeitmultiplex, Wissenschaftliche Hausarbeit
am Institut für Technische Akustik 1981
Fig. 1a Prinzipschaltung für Bewegungsgegenkopplung
Fig. 2a Bewegungsgegenkopplung im Zeitmultiplex
Fig. 3
Fig. 4
Fig. 5a
Fig. 5b
THE NEAR - FIELD SOUND DIFFRACTION BY THE HUMAN HEAD

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1. Introduction
Circumferential sound fields are determined for human heads, an artificial head and a sphere because the sound level distribution on the surface of the latter object is quite well understood. The aim of the investigation is to get a better understanding of some problems in directional hearing, e.g. the poor performance of an artificial head to reproduce frontal sound incidence.

2. Experimental set - up
The sound pressure level is recorded at a distance of about 2 cm between the microphone M2 and the diffracting surface in an anechoic chamber. This distance is controlled by an acoustic sensor and adjusted dynamically ($\Delta X$). The microphone can be rotated from 0 - 360 degrees by means of a stepping motor and the sound field is measured every 10 degrees. The sound signal is a pseudo random noise which emitted by a loudspeaker 2 m away from the head. The linear distortions produced by the loudspeaker and the microphone are eliminated by measuring the transfer function between an identical reference microphone M1 in the free field and M2. This is performed by a hardware FFT - processor in the frequency range from 0 - 10 kHz with a resolution of 125 lines. The transfer function is calculated on the basis of averaged spectra ($N=16$). The calculated amplitudes and phases are stored on floppy disks which are connected to a PDP 11/70 minicomputer that controls the whole measurement.

At the beginning of each measurement the

Fig. 1
Apparatus for measuring the near-field sound diffraction by the human head (in an anechoic chamber) M1 - reference microphone; M2 - measuring microphone; $\Delta X$ - distance controller; sensor-acoustical distance detector; $\phi$ - stepping motor control; FFT - calculation of transfer function (M2 to M1); PDP 11/70 - processing control.
subject's head is fixed in relation to the rotational axis. The microphone M2 moves in horizontal plane which is in the same height as the ear canal entrances. Besides human beings, a wooden sphere ($\sigma = 17.5\text{cm}$) and an artificial head also serve as 'subjects'.

3. Results and Discussion

The Figures 2 - 6 show the amplitudes of the measured transfer function in a 3D-representation.

The position of the microphone M2 given by the angle $\phi$ is drawn on the x-axis, whereas the y-axis represents the frequency range $f$, starting with the low frequencies in the rear. The dB level of the amplitudes is drawn on the z-axis with arbitrary origin.

Every time the amplitudes of 36 transfer functions ($360^\circ$ in $10^\circ$ steps) have been smoothed in the frequency and angle range. Thus interference minima are flattened, but the comparison among the figures becomes easier.

For comparison the diffracted sound field of the wooden sphere with head dimensions is shown in Figure 2. The minima and maxima of the near field appear like (smooth) peaks and valleys. The zero position angle $\phi = 0^\circ$ corresponds to the direction of sound incidence. The extended, symmetrical valleys show the first interference minima in the rear ($\phi = 180^\circ$) of the sphere. Further interference minima in the form of side valleys can also be detected in frontal regions of the sphere if the frequency is increased. Throughout the Fig. 2 small waves indicate regions of in-phase and out-of-phase situations.

If these results are compared to the theoretical calculations of the sound pressure on the surface of a sphere $\sigma / 1 /$, one finds differences due to the fact that we record the sound 2cm distant from the surface.

It can be seen from the following figures that the striking difference between the sphere and human heads (and the artificial head $/ 2 /$) are the more pronounced interference valleys of the sound fields. The sound field structures of all human subjects investigated show a high degree of similarity and the results of a typical subject are presented in the Figs. 3 and 4.

Figure 3 (subject BAJ) shows the diffracted sound field in the case of frontal sound incidence. A typical feature of this direction of incidence is that the interference valleys are stronger curved in the low frequency region below 5 kHz compared to the case where the sound comes from behind ($\phi = 180^\circ$).
Fig. 3
Diffraction pattern of a human head (subject BAJ)
\( \varphi \): Mic. M2 position (nose: \( \varphi = 0^\circ \), clockwise increasing)
f: Frequency and l: Sound level 2 cm distant from the surface (resp. gain of the transfer function M2 to M1).
Sound incidence: Frontal (\( \varphi = 0^\circ \)).

Fig. 4
Diffraction pattern of a human head (subject BAJ).
Symbols see Fig. 3
Sound incidence: Backwards (\( \varphi = 180^\circ \)).

Fig. 5
Diffraction pattern of the artificial head.
Symbols see Fig. 3
Sound incidence: Frontal (\( \varphi = 0^\circ \)).

Fig. 6
Diffraction pattern of the artificial head.
Symbols see Fig. 3
Sound incidence: Backwards (\( \varphi = 180^\circ \)).
Figure 4. Here the corresponding valleys are more stretched. The sound field of an artificial head are given in Fig. 5 and Fig. 6 for frontal and backward sound incidence. It is remarkable, that the structures of both sound fields are similar only to the case for a human subject where the sound comes from rear (see Fig. 4).

In this context it is interesting, that the reproduction of the frontal incidence by an artificial head is still very difficult. A comparison of the frequency regions of preference band /3/ at the position of the ears (typically $\varphi = 95^\circ$ and $265^\circ$) yields the following results.

For backward sound localisation the frequency preference band of 1 kHz is of importance. The corresponding regions in the Figs. 4 (see BAJ) and Fig. 6 (artificial head) show a rather well similarity at 1 kHz. However, considerable differences are observed in the 4 kHz region of the Fig. 3 (s. BAJ) and of the Fig. 5 (artificial head) which is of importance for frontal sound localization.

There is a first suspicion that these differences might deteriorate the subjective reproduction of the frontal direction of incidence in artificial head stereophony.

4. Conclusions

The diffraacted sound field near the surface of several 'subject's' heads have been measured for different directions of sound incidence.

For the frontal sound incidence the sound field structures differ remarkable between the human and the artificial heads, whereas there is a greater similarity if the sound comes from the rear. This is especially obvious for frequency regions of importance for the localization of frontal or backward sound incidence /3/.

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COMMENT CARACTÉRISER UN SYSTEME ÉLECTROACOUSTIQUE DE RESTITUTION SPATIALE?

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Cette communication est destinée à être présentée sur panneaux. Le plan d'expériences et les résultats seront donc commentés oralement à partir de tableaux dont 3 seulement sont reproduits ci-après. L'ensemble des résultats sera exposé pendant la séance de présentation.

RESUME

Les systèmes de restitution spatiale fournissent une image sonore et une impression d'espace, dite effet de salle. Il est très difficile de comparer ces systèmes si l'on n'essaie pas d'interpréter l'aspect subjectif en l'objectivant au moyen de critères fournissant des valeurs numériques. Un plan d'expériences a été bâti, utilisant plusieurs systèmes de micros donnant a priori une qualité de nature très différente, et 12 sources sonores variées. L'écoute était réalisée en 9 emplacements fournissant, a priori, des impressions très différentes toutes choses égales d'ailleurs. L'étude a porté sur 10 attributs, dont l'intérêt a été mis en évidence pour caractériser les qualités du système: largeur, profondeur, excentrement en largeur et en profondeur, homogénéité, conformité, précision, qualité globale, effet de réalité, effet de salle. Des échelles graphiques ont permis d'exprimer ces caractéristiques par des nombres allant de 1 à 5. La variation des cotations en fonction de la grande diversité des conditions expérimentales représente l'enseignement principal de ces essais. Les dimensions de la scène restituée ne constituent pas un attribut majeur de la qualité, l'impression de réalité ou l'effet de salle sont beaucoup plus importants ; ce qui ouvre l'ensemble des systèmes valables en particulier à ceux que suggèrent les recherches sur l'amélioration de l'ambiance, au moyen des traitements électroacoustiques.
RÉALISATION

Moyens d'écoute  

\[ \begin{array}{cccc} 
1 & 2 & 3 \\
4 & 5 & 6 \\
7 & 8 & 9 
\end{array} \]

Caractères de la scène sonore

- Largeur \( l \)
- Profondeur \( p \)
- Homogénéité \( h \)
- Conformité \( c \)

Disposition des microphones

- Distance \( d \)
- Angle \( \theta \)

Caractères généraux

- Précision \( P \)
- Qualité d'ensemble \( Q \)
- Effet de réalité \( R \)
- Effet de salle \( S \)

Plan d'expériences

<table>
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<th>Modulations (durée 15 s)</th>
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<th>m cm</th>
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Dépouillement : Deux types d'expériences
- détermination de \( P, Q, R, S \) par notation graphique sur des échelles à 5 niveaux.
- détermination des paramètres de la scène sonore.

Deux dessins \( S \) pour chaque sujet sur un plan de la salle, représentant l'impression d'écoute. Cette opération était riche d'enseignements. Interprétation faite par l'expérimentateur sur une échelle à 5 niveaux.
### MOYENNE GENERALE

(Ttes) Toutes places 1,...,9
(Op²) Places optimales 2,5,8

- **P** précision
- **Q** qualité d'ensemble
- **R** effet de réalité
- **S** effet de salle

1, 2 - 12 : modulations.

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### MOYENNE GENERALE

(Ttes) Toutes places 1…9  
(Op+) Places optimales 2, 5, 8

- $\delta l$ : écart de centrage en largeur
- $\delta p$ : écart de centrage en profondeur

1, 2, ..., 12 = modulations

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THE INFLUENCE OF THE SHAPE AND SIZE OF THE DUMMYHEAD UPON THE THÉVENIN ACOUSTIC IMPEDANCE AND THE THÉVENIN PRESSURE

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I. INTRODUCTION
To construct the dummyhead-headphone system for acoustic measurement we should determine the specification of the standard dummyhead and the required characteristics of the equalizer. The equalizer characteristics have been expressed by the author in terms of the Thévenin acoustic impedance and Thévenin pressure at the earcanal entrance, the length of the earcanal, the impedance of the tympanic membrane, the general circuit parameters of the headphone attached to the external ear, and so on.1) Further, it has been proposed that the Thévenin acoustic impedance and the Thévenin pressure at the earcanal entrance should be adopted as acoustical parameters of evaluating the effect of the shape and size of human head having individual differences.2)

In this paper, the influences of the shape and size of the head upon the Thévenin acoustic impedance and the Thévenin pressure are examined numerically in the case of two types of model heads. Since the shape and size of the model heads are given according to the somatometrical study of human head, the computational result are useful for the discussion of the specification of the standard dummyhead.

II. MODEL HEADS FOR COMPUTATION
We consider the two types of model heads as shown in Fig.1: One is a rigid spherical shell, the other is a rigid oblate spheroidal shell, and their auricle consists of only a cylindrical concha. The shape and size of the model head is determined on the basis of the somatometrical study of human head.3,4) The incident angle $\theta$ is defined as shown in Fig.2.

![Diagram](image)

Fig.1 Two types of model heads.
Fig.2 Definition of the incident angle.
2.1 The size of the model head for the computation of the Thévenin acoustic impedance.

Three spherical heads of radius 60, 100, and 140 (mm) are used, being ranged from the minimum to the maximum of head length, head breadth, head circumference and face length).

Two oblate spheroidal heads whose generating ellipse has minor diameter 150 (mm) and major diameter 180 (mm), and 170 (mm) and 200 (mm) respectively are used.

2.2 The size of the model head for the computation of the Thévenin pressure.

Three spherical heads of radius 82.5, 89.1 and 95.7 (mm) are used. They correspond in circumference to the lower limit \( m=1.15 \), the average \( m \), and the upper limit \( m=1.15 \) with respect to 75% distribution of the head circumference), being assured that the distribution is the normal distribution with average \( m \) and variance \( \sigma \).

Nine oblate spheroidal heads are provided. These two parameters of the shape and size, i.e., the minor diameter to major diameter ratio \( \eta \) and the total arc length \( L \) of the generating ellipse of the oblate spheroid, which correspond to the cephalic index \( I \) (ratio of head breadth to head length) and the head circumference respectively, are determined as shown in Fig. 3.

( Of the somatometrical data by Itou) the average \( \overline{\eta} \) and the standard deviation \( \sigma_\eta \) of the cephalic index \( I \) and the head circumference of 807 young male adults ranged in age 21 to 23 are used: \( \overline{\eta}=3.86 \), \( \sigma_\eta=15.0 \) (mm).)

### III. EQUATIONS FOR COMPUTING THE THÉVENIN ACOUSTIC IMPEDANCE AND THE THÉVENIN PRESSURE

Assuming that a plane wave of frequency \( f \) propagates in the concha in the longitudinal direction, we obtain the relation between the Thévenin acoustic impedance \( Z_0(x=0) \) and the Thévenin pressure \( P_0(x=0) \) at the ear canal entrance and the Thévenin acoustic impedance \( Z_0 \) and the Thévenin pressure \( P_0 \) at the concha entrance as shown in Eq. (1).

\[
P_0(x=0) = \frac{P_0}{(\cos kd + j(Z_0/\rho) \sin kd)}
\]

\[
Z_0(x=0) = Z_0(x=0) + kd\tan(kd)/(Z_0 + jZ_0 \tan(kd))
\]

where \( k=2\pi f/\omega \), \( Z_0=\rho c/S \), and \( S \) is the cross-sectional area of the concha, \( d \) the depth of the concha, \( c \) the sound speed of air, and \( \rho \) the density of air.

The \( Z_0 \) of the spherical head and the \( Z_0,\rho_0 \) of the oblate spheroidal head can be calculated from the radiation impedance \( Z_{rad,0} \) of the cap set of area \( S \) in the rigid spherical baffle 5) and the one \( Z_{rad,\rho_0} \) in the rigid oblate spheroidal baffle 6), respectively.

The \( P_0,\rho_0 \) of the spherical head and the \( P_0,\rho_0 \) of the oblate spheroidal head can be calculated from the pressure \( P_0(\alpha,\beta,\gamma,k) \) on the surface of the sphere placed in the spherical sound field 7) and the one \( P_0(\eta,L,\delta,k) \) on the surface of the oblate spheroid placed in the plane sound field 8), respectively. Then, the equations for the computation are as follows:

![Table showing parameter values for the nine oblate spheroidal heads.](image)

![Diagram showing the parameter values for the nine oblate spheroidal heads.](image)
\[ z_{e,s} = \frac{z_{rad,s}}{S^2}, \quad z_{e,os} = \frac{z_{rad,os}}{S^2} \]
\[ P_{e,s} = P_{e}(a,s,R,k), \quad P_{e,os} = P_{os}(n,L,\theta,k) \]  \hspace{1cm} (2)

where \( a \) is the radius of sphere, \( \theta \) the incident angle, and \( R \) the distance from the simple source to the sphere.

IV. RESULTS AND DISCUSSIONS

The computation of the \( Z_{e} \) and \( P_{e} \) of the model head described in Chapter 2 was made up to the frequency 15 kHz.

4.1 The influence of the shape and size of the model heads upon the \( Z_{e} \)

Each ratio of the \( Z_{e} \) of the model heads described in Sec. 2.1 to the \( Z_{e} \) of the spherical head of radius 100 mm is shown in Fig. 4. It can be seen that the \( Z_{e} \) is hardly influenced by the shape and size of the model heads.

Fig. 4 Ratio of the \( Z_{e} \) of the model head to the \( Z_{e} \) of the spherical head of radius 100 mm.

(a) The \( Z_{e} \) of the spherical head, and (b) the \( Z_{e} \) of the oblate spheroidal head. Here, the area \( S \) of the concha is chosen as \( S=380 \) mm.

4.2 The influence of the shape and size of the model heads upon the \( P_{e} \)

4.2.1 Comparison between the spherical heads having the equal deviation from the average in circumference except for their signs

The comparison showed that the log magnitude and phase of the Thévenin pressure ratio \( P_{e,\theta}(a=82.5)/P_{e,\theta}(a=89.1) \) are equal to the negative ones of the ratio \( P_{e,\theta}(a=95.7)/P_{e,\theta}(a=89.1) \).

4.2.2 Comparison between the oblate spheroidal heads

The comparison of the Thévenin pressure ratio \( P_{e,os}(X)/P_{e,os}(S) \) showed the followings where \( P_{e,os}(X) \) denotes the \( P_{e,os} \) of the head "X":

1. The log magnitude and phase of the \( P_{e,os}(B)/P_{e,os}(S) \) are equal to the negative ones of the \( P_{e,os}(D)/P_{e,os}(S) \).
2. The log magnitude and phase of the \( P_{e,os}(A)/P_{e,os}(S) \) are equal to the negative ones of the \( P_{e,os}(F)/P_{e,os}(S) \).
3. The log magnitude and phase of the \( P_{e,os}(C)/P_{e,os}(S) \) are obtained from the following relation:

\[ P_{e,os}(C)/P_{e,os}(S) = (P_{e,os}(A)/P_{e,os}(S))(P_{e,os}(B)/P_{e,os}(S)) \]

The similar relation is held for the heads E, G, and H: (E)-(A,D), (G)-(B,F), and (H)-(D,F). (For above (1), (2), and (3), see Figs. 5, 6, and 7.)

Further, the first extremal value of the Thévenin pressure ratio with the incident angle changed gives the maximum in magnitude at any frequency and this value reaches to the maximum at the highest frequency in concern. And the maximum in phase of the ratio occurs at either 0° or 180° and this value increases with frequencies. (Figures are omitted here.)

From these observations and the relation mentioned above the equal magnitude and equal phase contours of the maximum of the Thévenin pressure ratio at frequencies less than 15 kHz are easily obtained as shown in Fig. 8.
Fig. 5 The comparison of \( P_{s,os}(B)/P_{s,os}(S) \) (- -) with \( P_{s,os}(D)/P_{s,os}(S) \) (-----).

Fig. 6 The comparison of \( P_{s,os}(A)/P_{s,os}(S) \) (- -) with \( P_{s,os}(F)/P_{s,os}(S) \) (-----).

Fig. 8 The equal magnitude (a) and equal phase (b) contours of the maximum of the Thévenin pressure ratio at frequencies less than 15 kHz.

4.2.3 Concluding remarks

From the results of Sec. 4.2.1 and 4.2.2, it can be said that the head average in dimension has the average acoustical properties of heads in the range of 75% distribution of the somatometrical data of cephalic index I and head circumference of human head.

REFERENCES
DISTORTION MEASURE BASED ON A MODEL FOR THE AUDITORY PERIPHERY.

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Introduction
Sound reproducing systems as e.g. hearing aids and telephones distort sound depending upon physical imperfections. These imperfections can be of a nonlinear nature or that the system has a frequency dependent gain. Conventional physical measures of distortion, e.g. harmonic distortion and intermodulation, do not give a one to one correspondence with perceived distortion, because these measures do not take into account the properties of the hearing system.

By using a model for the auditory periphery, which takes into account the masking and filtering properties of the auditory system, simulations of nonlinear networks with sinusoidal input signals have been made.

Model
The model consists of different parts which describe the meatus, the middle ear and the basilar membrane. The model for the meatus consists of a simple recursive filter giving a resonance of 10 dB at 3.6 kHz. The model of the middle ear is from Shaw (1979). Figure 1 shows an electric analog of a section of the basilar membrane model (Nilsson, 1977). 175 sections have been used in this study.

The analog electric models of middle ear and basilar membrane have been transformed to a wave digital filter (Fettweis, 1971) and programmed on computer. The resulting model is thus recursively computable, which means that the simulation is fast compared to the solution methods used earlier.

Figure 2 shows the frequency transfer function between the pressure at the eardrum and the velocity at various points of the basilar membrane with different characteristic frequency.

Model for spectral sharpening
In attempts to explain the differences in frequency selectivity between in animal experiment measured basilar membrane movements and neural activity a large number of models for a "second filter" have been suggested. Most of these imply that there is a resonance circuit after the basilar membrane.
Some authors (Zwislocki, 1974; Allen, 1977) have presented alternative models which assume interactions between outer and inner hair cells.

Figure 3 shows the model that has been used in this study to increase the frequency selectivity. The input signals, representing membrane velocities at different points, are half-wave rectified and summed to a left going and a right going inhibition signal, respectively. As the coefficients have values between zero and one, the influence of inhibition between two points of the basilar membrane will decrease the longer the distance between the points is. An input signal from the basilar membrane can only pass through the network if it is bigger than the sum of the inhibition signals. From Figure 4 it is clear that the frequency selectivity has increased markedly by the influence of inhibition.

The resulting neural levels of activity for points with various characteristic frequency are within some dB of the levels of hearing threshold according to ISO R226 if the influence of the diffraction of the head is taken into account.

Results
Experiments for determining the thresholds of nonlinear distortion with a sinusoidal input show that one is more sensible to cubic than quadratic distortion. Different authors give different values but a clear tendency is that one tolerates approximately twice as much quadratic distortion as compared to cubic.

To test if the model could explain these discrepancies, simulations with a distorted 1 kHz sinusoidal as input signal were performed.

For each degree of distortion the relation between the levels of activity of the harmonic and the fundamental was registrated. (1% cubic distortion gave a relative harmonic activity of -72 dB.)

The curve in Figure 5 gives the corresponding values of quadratic and cubic distortion which give the same relative harmonic activity. If this measure is relevant for the perception of the distortion, then the model predicts that the threshold for quadratic distortion is approximately 1.6 times the threshold for cubic distortion.

Discussion
The simulations described above is a part in experiments to obtain distortion measures with good correlation to perceived distortion for arbitrary systems and input signals.

In the case with a nonlinearly distorted sinusoidal signal the simulation shows that this is possible if one uses a model for the transformation of the acoustic signal to neural activity.

If the same measure can be used for arbitrary systems and input signals remains to be shown.
The model is also of interest when one wants to study other aspects regarding sound transmission in the auditory system e.g. for estimation of the risk of damage in the hearing system from impulsive sounds. In this case it is of course essential to have a good understanding of the actual stimulation on the basilar membrane.

References


![Figure 1](image1)

The electrical analog of one section of basilar membrane model (Nilsson, 1977).

![Figure 2](image2)

Model frequency transfer functions from pressure at eardrum to velocities at various points of the basilar membrane.
Input signals from basilar membrane

Figure 3
Model for lateral inhibition

Neural activity

Figure 4
A comparison between model mechanical and neural frequency selectivity for one point of the basilar membrane

Figure 5
The relation between levels of quadratic and cubic distortion which give the same model neural activity
LA CONSERVATION DES ENREGISTREMENTS SONORES SUR BANDES MAGNETIQUES

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Bien que relevant de techniques relativement peu anciennes par rapport à d'autres supports de la pensée, les Archives Sonores n'en constituent pas moins aujourd'hui une part importante de notre patrimoine. L'intérêt culturel reconnu est tel qu'une action est entreprise pour définir et appliquer, parfois de manière urgente, toutes mesures destinées à conserver durablement les documents sonores de nos collections publiques. C'est ainsi que, faisant suite en particulier à l'insistance de conservateurs de Phonothèques responsables, une recherche est lancée en France depuis près de 3 ans dans le cadre du C.N.R.S. et du Ministère de la Culture. Un des objectifs principaux concerne la pérennité des phonogrammes déposés à la Phonothèque Nationale (Bibliothèque Nationale) au titre du Dépôt Légal. Documents, il faut le souligner, non destinés initialement à être conservés...

Mais bien évidemment l'ensemble des collections d'Archives sonores qui se constituent et dont l'augmentation s'accélère sont concernées, ne serait-ce que dans les laboratoires d'Acoustique!

Les problèmes soulevés par la conservation des informations sonores (de tous types) sont vastes du fait de l'extrême diversité des supports (cylindres, disques aux multiples compositions, fil et ruban métallique, bandes magnétiques aux incessantes formulations, etc) et parce que l'appareillage tient une place tout aussi prépondérante tant à l'inscription du message qu'à l'accès à celui-ci. La conservation des appareils dont le standard est rare ou susceptible de changer, et des pièces de rechange maîtresses est aussi importante que celle du support.

La mise en œuvre des moyens d'étude étant très progressive, nous nous limitons pour l'heure aux recherches sur les bandes magnétiques enregistrées en mode analogique. Notre démarche consiste à se faire une certaine idée de la qualité des supports actuellement disponibles sur des critères étendus à ceux de la conservation et à percevoir les processus de détérioration des enregistrements (toute origine confondue). Phase qui prépare les possibles restaurations. Celles-ci se situent d'ailleurs essentiellement à ce niveau du traitement du signal lors d'un transfert sur un support sain : pour le meilleur et pour le pire!
Le comportement magnétique est-il en cause ?

La démagnétisation spontanée est parfois suspectée. Elle peut se produire lorsque le traitement chimique des particules d’oxyde de fer, destiné à augmenter l’aimantation rémanente et la force coercitive (dopage) confère des instabilités ioniques. Phénomène rare mais déjà rencontré. L’enregistrement des hautes fréquences à vitesse réduite favorise l’action contrariante des micro-champs interarticulaires : les moments magnétiques des particules instables tendent à se neutraliser mutuellement au cours du temps. Le simple passage dans le système de défilement de l’appareil peut avoir un effet néfaste pour des particules magnétiquement instables. Ce phénomène (magnétostriiction) atteint alors le niveau de réponse des hautes fréquences, surtout aux toutes premières lectures.

Le danger d’exposition à un champ magnétique continu ou alternatif capable d’effacer purement et simplement l’enregistrement, de le déformer de manière catastrophique n’est jamais écouté totalement, mais le niveau des champs ayant ces effets est particulièrement intense : les circonstances de tels voisinages restent tout à fait accidentelles (manipulation malheureuse sur machine, malveillance ...).

Par contre les champs magnétiques susceptibles d’altérer, même légèrement, la qualité de l’enregistrement de type analogique, peuvent n’être que de faible intensité et seront par conséquent rencontrés couramment : au cours des transports, dans les studios, les laboratoires où les systèmes électromagnétiques ne manquent pas, etc. On observe alors une augmentation du bruit de fond, de la distorsion harmonique et de l’effet de copie (anhystérétique, c’est-à-dire irréversible). La température est redoutable surtout pour certains types d’oxyde où un effet sensible sur l’induction rémanente est noté. Se conjugant avec le temps, la température constitue le principal facteur extérieur de l’effet de copie (partiellement réversible et)

Les problèmes d’ordre mécanique

La fabrication détermine bien évidemment les performances initiales mais aussi une grande partie du potentiel de bonne tenue des produits en présence. Les tolérances géométriques, la formulation du liant, les caractéristiques d’homogénéité du revêtement, l’état de surface après calendrage, ... constituent quelques exemples de points de fabrication particulièrement critiques. L’utilisation du film polyester préétreint a amélioré de manière considérable le comportement mécanique du ruban, dans la mesure toutefois où l’épaisseur n’était pas trop faible ! Les conditionnements en boîtiers fermés (cassettes, cartouches, jingles) apportent une protection efficace de la bande mais aussi autant de problèmes possibles de défilement de celle-ci, de contact sur les têtes qu’ils comportent d’organes.

L’utilisation peut pénaliser mécaniquement les bandes pendant les manipulations négligentes et les passages sur machines défectueuses. On ne souligne jamais assez l’importance des précautions à prendre par l’utilisateur, même si elles paraissent contraignantes. Ceci pour prévenir des risques d’écrasement de bord de spires, de déroulement "en perruque", de dépôts de poussière, etc. Les bobines et conteneurs doivent faire l’objet d’attention toute particulière, de même que les appareils sur les aspects mécaniques.
Pendant le stockage, les agressions mécaniques résultent essentiellement des variations thermohygrométriques. Les coefficients de dilatation du polyester actuellement utilisé sont faibles mais la longueur du ruban est telle que des contraintes importantes s'instaurent dans le bobinage pour seulement quelques degrés centigrades ou quelques % d'humidité relative d'écart. Peu à peu le fluage du polyester rend les déformations au sein du bobinage irréversibles. Lorsque les conditions climatiques ne sont pas maitrisées (elles ne le sont jamais parfaitement), la seule parade consiste à redistribuer les contraintes en procédant régulièrement à des déroulements-réenroulements à vitesses constantes (mode lecture). On recommande d'opérer ainsi au moins une fois par an. Un autre point concerne la poussière. Un filtrage des locaux de stockage s'avère nécessaire. La poussière qui s'immisce entre les spires présente le grave inconvénient d'écarter localement la bande de la tête (perte de modulation commençant par les fréquences les plus élevées). Elle risque fort en outre de rayer la surface sensible, voire les têtes...

Les phénomènes de dégradation d'origine chimique

Le liant présent le plus de risques à cet égard. Sa constitution physico-chimique peut être modifiée par l'action de facteurs internes (réaction propre des constituants, interaction entre eux-ci, et externes (souillages mécaniques, climatiques). Les cas extrêmes de dégradation se traduisent, par exemple, par un changement de consistance du revêtement qui devient poisseux ou encore par un détachement pur et simple de celui-ci.

Dans la formulation de la pâte destinée à être enduite, de nombreux additifs assurent les qualités d'enrobage, de dispersion des poudres de souplasse convenable du revêtement, etc : ils peuvent se volatiliser progressivement ou se transformer au fil des ans. De plus, les particules déclenchent ou accentuent les réactions chimiques lorsqu'elles jouent le rôle d'oxydant et de catalyseur.

Récemment, la cinétique de dégradation d'un type de liant (polyuréthanes) en fonction des conditions thermohygrométriques a fait l'objet d'études approfondies (1-2). Le maintien des bandes en climat relativement sec (20 40 % HR) constitue une protection acceptable, à température ambiante. Mais il nous faut bien reconnaître que pour la plupart des matériaux nous ignorons la formulation et l'inconnue sur leur possible évolution est entière...

Un regard général de l'ensemble de ces problèmes (3) et les observations in situ nous ont amené à entreprendre des études en vue des applications au sein des institutions responsables des collections, qui s'orientent dans 3 directions principales :

1. Définir quels types de bande présentent les meilleures garanties du double point de vue de la qualité d'enregistrement et d'une conservation correcte. Cette démarche implique la sélection et le classement régulièrement mis à jour des produits retenus disponibles sur le marché. Ceci doit être établi sur des critères spécifiques déterminés à l'aide de modes d'agression contrôlés et révélant au mieux les traumatisme subits et transposables dans des situations réelles. Naturellement les dispositions d'enregistrement doivent être très solignées : choix des matériaux, mise en place de modes opératoires rigoureux sachant que l'importance de toutes les opérations à ce niveau sont fondamentales pour la qualité du document, sa survie.
2. Préciser les conditions de stockage les plus favorables

On a bien une idée de l'origine des dangers mais les risques encourus par les documents sonores sont encore difficiles à préciser et à hiérarchiser, du fait du manque de recul dont nous disposons et de par la très grande variété des supports concernés. Très peu d'études spécifiques ont été entreprises sur ces questions. Les expériences que nous menons permettront de dégager les mesures prioritaires et de situer les "degrés de liberté" tolerables.

3. Rechercher et développer les moyens de restauration des documents détériorés.

Pour l'essentiel, les traitements concernent le signal acoustique dès lors que l'on souhaite se consacrer à l'information enregistrée pour elle-même et si le document reste lisible : la lecture correcte, c'est-à-dire le réglage optimum du lecteur constitue la première étape, bien délicate lorsqu'on ne possède ni référence, ni indication sur les points de fonctionnement de l'appareillage d'enregistrement. L'utilisation des dispositifs destinés à offrir une restitution "satisfaisante" est extrêmement difficile quand on songe à l'ambiguïté du terme satisfaisant. Dans ce domaine où les aspects techniques sont confrontés avec d'autres aspirations (historiques, esthétiques, musicologiques,...) il n'est pas question d'avoir une attitude de principe mais de proposer une solution à chaque cas de manière spécifique.

Nous abordons ces études en relation avec des organismes récemment constitués qui réunissent sur le plan national et international les compétences dans les domaines de la conservation, l'électronacoustique, la prise de son, des fabricants de bandes et de "machines", des collectionneurs, des artistes musiciens, des chercheurs travaillant sur documents sonores,... En France, depuis 1979, l'Association Française des Archives Sonores (A.F. A.S.) s'est donnée pour but, entre autres tâches, d'apporter des informations, des conseils pour assurer une bonne conservation des documents. L'Association Internationale des Archives Sonores (sigle anglais I.A.S.A.) comporte un Comité technique chargé d'étudier les dispositions à prendre au niveau des pays adhérents, facilitant ainsi par exemple les échanges d'enregistrement.

Avec des intentions comparables s'est constitué en novembre 81 aux USA (N.Y.) le Comité de Conservation des Enregistrements devenu depuis Fondation (Foundation for the Preservation of Recordings, Inc.). Elle rassemble ici aussi les compétences interdisciplinaires indispensables pour poursuivre un programme aussi vaste.


(2) BERTRAM (N.), CUDDHY (E.F.), Kinetics of the humid aging of magnetic recording tape. IEEE Trans. on Mag. vol. MAG 18, n° 5, sept. 1982, pp. 993-999

PRACTICAL RESULTS WITH "PRESSURE ZONE MICROPHONES" FACING THE SOUND SOURCE

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1. Scope of this paper
In recent years a new discussion was stirred up on "pressure zone microphones". These microphones are positioned with their diaphragms opposed but nearly flush with a hard surface, e.g. the wall or floor of a recording studio or concert hall. A new microphone facing the sound source with a flush diaphragm avoids reflections and interferences at its own surface. Because the use of these microphones making use of the boundary-layer-effect has been accompanied by diverse discussions all over the world, this paper attempts to cast a little light onto facts related to boundary-layer-microphones, also trying to contribute to a more neutral discussion on this topic.

2. Description of the transducer
A microphone of the type MKE 212 has been in use under real recording conditions for more than one year. This microphone consists of a small low noise electret capsule with a pick-up opening of only 0.5 mm ø and is mounted flush in a rigid plate of the dimensions 165 mm x 185 mm. Thus reflections and diffractions in the vicinity of the transducer are avoided. Recordings have been made with this new tool on a great number of recording sessions covering the whole repertoire of music. The results as estimated by the sound engineers are extremely favourable. In general this microphone was used as "footlight-microphone" placed directly on the floor. Although successful attempts of using it attached to a grand piano as well, with similar good results, most applications used a pair of these new microphones as main microphones resulting in a very favourable reproduction of the room and depth.

In the past discussions and recent papers application of these transducers hung freely somewhere in the concert hall or mounted on plates of different dimensions have been propagated. Because this kind of application must lead to disturbances of the frequency response it was our aim to show the influence of certain set-ups on the characteristics of those microphones.

3. Measurements
Consequently a great number of measurements have been carried out on boundary-layer-microphones of the type MKE 212 of Sennheiser electronic. The frequency response shown in the different graphs have been calculated relative to the freefield-response of the electret transducer used. Results are
shown for different angle of incidence and measuring set-ups.

4. Facts that have to be observed

It is a fundamental fact that a sound-wave incident perpendicular to a hard surface experiences a pressure increase of 6 dB. This holds true for unlimited plane surfaces only. Restrictions of this "law" have to be observed if the boundary is finite or certain set-ups for the microphone are used, as pointed out by Lipshitz (1) in his paper at the 69th AES Convention in Los Angeles, May 1981. Also all diffraction phenomena have to be taken into account as for instance shown in the paper of Olson (2) from 1938. Keeping all this in mind some meaningful results can be deducted from the following measurements.

5. Results of measurements

Fig. 1 shows the free-field-response of the boundary-layer-microphone MKE 212 for different angles of incidence in 1 m measuring distance. Large irregularities in the frequency response can be identified for the frequency of 1.3 kHz and its multiples, which can be clearly correlated to the diagonal dimension of the plate used. The different mechanisms of diffraction and interference result in an almost "ideal" frequency response for 60° with a pressure doubling above the critical frequency where the dimension of the plate are large compared to the wavelength of sound. This holds true only for this angle of incidence and a considerable ripple in the frequency response for other angles of incidence have to be observed!

Fig. 2 shows the same measurements with the electret capsule mounted asymmetrically in the same plate; the location was chosen in such a way that the distances from the transducer to the circumference of the plate allows for statistically distributed path lengths, resulting in a corresponding number or possibly "preferred resonant frequencies". All measurements carried out with this scope in mind show virtually the same behaviour as fig.1 with - of course - a certain shifting of peaks and valleys in the frequency response.

Totally different are the results of fig. 3 when the same measurements as in fig. 1 are carried out in a measuring distance of 2.7 m in the same room. Peaks and valleys in the frequency response are smoothed out and it seems impossible to find a correlation of those irregularities with the dimensions of the plate. The reason for this is the more ideal plane wavefield for the large measuring distance.

Lastly, measurements have been carried out with the transducer or a complete MKE 212 placed on a larger plate of the dimensions 575 mm x 585 mm. Fig. 4 and 5 show the results for different angles of incidence in a measuring distance of 2.85 m at an elevation angle of 18° respectively. Now an even new set of frequency responses as a function of angle of incidence can be recorded, which is only disturbed by a resonance phenomena in the vicinity of 800 Hz correlated to the square dimensions (length and width) of the plate. The ripple for higher frequencies may well be attributed to disturbances caused by the large plate in the anechoic chamber and should be left out of consideration.

6. Conclusions

In contrast to the often stipulated impossibility of measuring boundary-layer-microphones in a conventional way it has been shown that meaningful
frequency response of such microphones can be obtained with standard procedure. The results show that the properties of such microphones are far from ideal in a studio sense and are highly dependent on the dimensions of the plate used and the measuring distance; this implies that in real recording situations the frequency response is highly dependent on both the distance between microphone and sound-source and the room-acoustics of the environment. This also gives us the hint why such microphones can be advantageous under certain recording conditions because they may exhibit a very special but shaped frequency response suitable for this certain recording task.

On the other hand the parameters influencing the frequency response in a certain set-up cannot validly be predetermined and thus are highly unpredictable factors. Further research is necessary to classify the different parameters in order to lead the boundary-layer-microphone from a "subjective" transducer to an objective microphone for the sound-recording engineer.

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**Fig. 1:** Transducer in the middle of small rectangular plate

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**Fig. 2:** Transducer asymmetrically mounted in small rectangular plate
Fig. 3: Rectangular plate in greater distance from source

Fig. 4: Larger plate with transducer in greater distance

Fig. 5: Complete MKE 212 mounted on larger plate

Literature:
MEßSYSTEM ZUR KONTINUIERLICHEN ÜBERWACHUNG DER SCHLEIFENVERSTÄRKUNG IN BESCHALLUNGSANLAGEN

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I.) Einleitung

Die akustische Rückkopplung in Beschallungsanlagen kann bekanntlich störende Klangverfärbungen bis hin zum sogenannten Rückkopplungspfeifen hervorrufen, wenn sich die Schleifenverstärkung (Mikrofon-Verstärker-Lautsprecher-Raum-Mikrofon) dem kritischen Wert 1 nähert. Im allgemeinen kann zwar eine ausreichende Schallversorgung der Zuhörer ohne Rückkopplungseinfüsse durch geeignete konstruktive Maßnahmen (Richtmikrophone, Lautsprecherzeilen) erreicht werden, doch bleibt es dem Bediener der Anlage überlassen, die Verstärkung nach dem Gehöreindruck zu regulieren und so (insbesondere bei Hand- und Umhängemikrophonen) Instabilitätserscheinungen durch Änderung der Schleifenverstärkung zu verhindern.


II.) Meßverfahren

Das Meßverfahren nutzt das Sprechersignal als "Testsignal" zur Messung der Schleifenverstärkung aus. Dieses erzeugt im Mikrophonsignal (M1 in Fig. 1) einen Direktschallanteil (mit der Leistung D) und einen Raumanteil (Leistung R), der maßgeblich durch die von der Lautsprecheranlage hervorgebrufenen Schallanteile bestimmt ist. Das Verhältnis R/D (d.h. der auf die Sprecherleistung normierte Raumschallanteil) charakterisiert die Schleifenverstärkung in der Beschallungsanlage. Zur Unterscheidung beider Schallanteile werden bei dem im folgenden
vorgestellten Meßkonzept die Modulations-Übertragungseigen-
chaften des Raumes ausgenutzt. Der Raumschallanteil hat nach
Durchlaufen der Übertragungstrecke vom Lautsprecher zum Mi-
krophonort eine gegenüber dem Direktschall veränderte Hüll-
kurve. Im Normalfall wird sich das Sprechermikrophon im
Diffusfeld der Beschallungslautsprecher befinden, dann ergibt
sich die Einhüllende der Raumsignalleistung \( R \) nach der Theorie
der Modulations-Übertragungsfunktion (\( = MTF = \text{modulation-}
transfer-function} | 3) durch eine einfache Tiefpaaufschaltung aus
der Modulation der Direktschalleistung \( D \). Die Grenzfrequenz
der Tiefpaaufschaltung ist von der Nachhallzeit des Raumes
abhängig (siehe \( | 3) und liegt typischerweise im Bereich von
1 Hz bis 4 Hz, d.h. in dem Frequenzbereich, in dem auch die
Sprachmodulation sehr große Anteile aufweist. Die durch die
Tiefpaaufschaltung hervorgerufene Phasendrehung zwischen den Modu-
lationsanteilen \( D \) und \( R \) ermöglicht eine Trennung der Anteile
in einer Meßanordnung nach Fig.1, bestehend aus dem Sprecher-
mitrophon \( M1 \), einem Hilfsmikrophon \( M2 \) in einem Abstand von 10
bis 30 cm hinter \( M1 \) und einer Detektionsschaltung.
Die Leistung \( P1 \) des Mikrophonsignals an \( M1 \) ergibt sich aus
Direkt- und Raumannteil zu

\[
P_1 = D + R
\]

Am zweiten Mikrophon \( M2 \) trifft das Sprechersignal wegen des
größeren Sprecherabstandes um einen Faktor \( k \) geschwächt, aber
mit derselben Form der Hüllkurve, ein. Die Raumsignalleistung
dagegen ist an beiden Mikrophonen annähernd gleich groß, da
der Abstand der Mikrophone zueinander sehr klein gegenüber dem
gemeinsamen Abstand zum Lautsprecher ist und da sich außerdem
beide im Diffusfeld der Beschallungsanlage befinden. Somit
kann

\[
P_2 = k \cdot D + R
\]

angenommen werden, wobei der Faktor \( k \) allerdings vom Abstand
des Sprechers zu den Mikrophonen abhängt und nicht bekannt ist.
In der Analyseschaltung wird zunächst in einem 90°-Phasen-
schieber das zu \( P1 \) orthogonale Signal \( P_{10} \) gebildet (natürlich
nur für die Wechselanteile) und sodann das Produkt

\[
P = P_{10} \cdot P_2 = (D_{10} + R_{10}) \cdot (k \cdot D + R) = k \cdot D \cdot D_{10} + D_{10} \cdot R + k \cdot D \cdot R_{10} + R \cdot R_{10}
\]

(Index \( o = \text{orthogonal zum Ursprungssignal} \)

\[
(3)
\]

Das Produkt zweier orthogonalen Funktionen \((D_{10}, R_{10})\) ist de-

\[
\text{finitionsgemäß mittelwertfrei, so daß sich der zeitliche}
\]

\[
\text{Mittelwert von } P \text{ nach Gl. 4 ergibt:}
\]

\[
\bar{P} = D \cdot R + k \cdot D \cdot R_{10} = (1-k) \cdot D \cdot R
\]

\[
(4)
\]

\[
\text{(mit } D \cdot R = D \cdot R_{10}; \text{ allg.: } \bar{X} = \text{zeitlicher Mittelwert von } X)\]

Durch Normierung auf \((P_1 - P_2) \cdot \bar{P}_1\) folgt

\[
P_N = \frac{\bar{P}}{(P_1 - P_2) \cdot \bar{P}_1} = \frac{(1-k) \cdot D \cdot R}{(1-k) \cdot D \cdot (D + R)}
\]

\[
(5)
\]
Der Mittelwert des Produkts $D_0 \cdot R$ ist abhängig von $D_0$ (= $\bar{D}$), $\bar{R}$ und der Phasenverschiebung zwischen $D$ und $R$, die durch die MTP des Raumes gegeben ist, somit gilt $D_0 R \cdot D \cdot R$. In der Klammer im Nenner kann im stabilen Betriebsbereich der Anlage $\bar{R}$ gegen $D$ vernachlässigt werden.

$$P_N = \frac{\bar{D} \cdot \bar{R}}{D} = \frac{\bar{R}}{D}$$ (6)

$P_N$ charakterisiert also das zeitlich gemittelte Verhältnis von Raumschallanteil zu Direktschallanteil am Sprecher mikrophon und damit die Schleifenverstärkung in der Anlage.


III.) Meßergebnisse

Erste Meßergebnisse mit einer Versuchsschaltung gemäß Fig. 1 sind in Fig. 2 und 3 dargestellt. Die Messungen wurden in
einem großen Laborraum (Nachhallzeit 1 sec.) mit sprachbewertetem, sinusförmig moduliertem Rauschen als Testsignal durchgeführt. Die Mikrophone mit nierenförmiger Richtcharakteristik hatten dabei einen gegenseitigen Abstand von 20 cm. Fig. 2 wurde etwa 3 dB unter der Schwinggrenze aufgenommen und zeigt die Abhängigkeit der Ausgangsspannung $P$ von der Modulationsfrequenz. In guter Übereinstimmung mit theoretischen Überlegungen liegt das Empfindlichkeitsmaximum bei einer Modulationsfrequenz von etwa 3 Hz. Bei tieferen Frequenzen ist die Phasenverschiebung von Raum- und Direktschallanteil zu gering (d. h. $D_0$ und $R$ sind fast orthogonal), bei höheren Frequenzen macht sich die Tiefpasswirkung der MTF bemerkbar. Fig. 3 zeigt für eine Modulationsfrequenz von 3 Hz die Abhängigkeit der Ausgangsspannung $P$ von der auf die Schwinggrenze $V_0$ normierten Schleifenverstärkung $V$.

Die bisherigen ersten Ergebnisse lassen erwarten, daß nach dem oben beschriebenen Verfahren die Messung und Kontrolle der Schleifenverstärkung einer Beschallungsanlage ohne Beeinträchtigung der Übertragungsqualität möglich ist.

![Fig. 2](image1)

![Fig. 3](image2)

Literatur:


A SIMPLE MATHEMATICAL THEORY OF STEREOPHONIC LOCALIZATION

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1. Introduction
Previous theories found in the literature are restricted mainly to the determination of sound image direction. An exception is Toshitada Doi's paper in 1977. These theories postulate that the human brain determines the sound image direction on basis of ITD/interaural time differences and ILD/interaural level differences.

Our theory presented in this paper does not restrict the method of comparison sound signals coming from the source to the left and right ear. Our fundamental condition is the equality of sound signal at the ear/left or right/ coming from the stereophonic system with sound signal at the ear/left or right/ from a single/phantom/ source. By derivation of localization expressions is assumed that the original source/signal of which is recorded/ is identical with the phantom source of the sound image/e.g.: the phantom sound source has the directional characteristic like the original one/. Happens the stereophonic sound reproduction wether through headphones or with help of loudspeakers the shadowing effect of the head is taken into account, too. This theory results localization formulas which give the place coordinates and the pressure amplitude of the image source /if the original one is a point source/ in dependence of recording and reproducing parameters.

2. The model of investigation
Many measured values found in the literature verify the fact that ITD is proportional to the path differences/in the kHz range too/ and the ILD has the 10-15dB maximal value because of the shadowing effect of the head. With consideration to the above-mentioned we construct a model whose actual ITD is given through the path differences whilst the ILD is produced with help of directional characteristic applied to the ear.

/See Fig.1/ In order to investigate the localization problem we set the listener's head into a rectangular coordinate system. So the applied directional characteristic of the ear we can give as follows:
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\[
e_r = \left\{ \left[ (1 + \hat{a}^2 + \hat{c}^2) \frac{z(a + d) + by + cz}{\sqrt{\hat{z}^2 + \hat{y}^2 + (x + d)^2}} \right]^{1/2} \right\}
\]

\[
e_r = \left\{ \left[ \frac{\hat{a}^2 + \hat{c}^2 + \hat{z}^2}{2} \frac{a}{(x + d)} \right] \frac{1}{\sqrt{\hat{z}^2 + \hat{y}^2 + (x + d)^2}} \right\}^{1/2}
\]

\[
\rho_r = \left[ \frac{z^2 + y^2 + (x + d)^2}{\sqrt{\hat{z}^2 + \hat{y}^2 + (x + d)^2}} \right]^{1/2}
\]

Where \( f(t) \) is any periodic function depending on time; \( a, b, c \) are transformations of the center of the spherical directional characteristic with a radius equal to unity. If we take into account that the pressure wave amplitude of a point source depends reciprocally on the path then on base of our new theory the initial equations are:

\[
p_r \cdot f(t - \tau_r) = p_r \cdot f(t - \tau_r) = \frac{p_r \cdot f(t - \tau_r)}{\rho_r \cdot \rho_r} \]

and from 1.3 follows:

\[
p_r = p_r \cdot \rho_r / \rho_r \]

where \( \rho \) includes the frequency dependence of \( \rho_r \) and \( \eta \). It can be determined from base of measured values given in the literature e.g. in *Inserting now 1.4 and 1.5 into 1.4 we get:

\[
p_r^2 = \frac{p_r \cdot \rho_r}{\rho_r} \]

\[
p_r = \frac{p_r \cdot \rho_r}{\rho_r} \]

\[
\eta = \frac{\rho_r \cdot \rho_r}{\rho_r} \]

\[\tau_r = \frac{\rho_r \cdot \rho_r}{\rho_r} \]

With help of equations 1.5, 1.6, 1.7 the place coordinates and the pressure amplitude of the phantom source are expressible as follows:

\[
x = c_0^2 (\tau_r^{2} \tau_r^{2}) / 4 M C^2 \]

\[
y = \{A b + C [B (c_0^2 + b^2) - A^2]^{1/2} / (b^2 + c^2) \}
\]

\[
z = \{(A + b) [B (c_0^2 + b^2) - A^2]^{1/2} / (b^2 + c^2) \}
\]

\[
p = \{c_0^3 (p_r^{2} \tau_r^{2} - p_r^{2} \tau_r^{2}) / \rho_r^3 [G (c_0^2 \tau_r^{2} - 4 a d)] \}
\]

\[
A = \left[ \frac{c_0 G}{2 \rho_r} \left( \frac{p_r^{3} \tau_r^{2}}{4 d c_0} \right) \right] - \left[ \frac{c_0 G}{2 \rho_r} \left( \frac{p_r^{3} \tau_r^{2}}{4 d c_0} \right) - d \right] \left( \frac{p_r^{3} \tau_r^{2}}{4 d c_0} \right)
\]

\[
B = \frac{c_0^2 (c_0^2 \tau_r^{2} - c_0^2 \tau_r^{2})}{2 \rho_r^2} - (c_0^2 \tau_r^{2} - c_0^2 \tau_r^{2})^2 / 4 d^2
\]

**Kuhn, G.F.: Model for the interaural... JASA vol. 62, 1977 (107-167)**
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If we suppose that the spherical directional characteristic of the head has the symmetry \( b = c = 0 \) /low and mid frequency approximation/ then the coordinates of the image source in the horizontal plane are:

\[
X = \frac{c^2 - \frac{y^2 + x^2}{a^2}}{\kappa_x^2 - \frac{y^2 + x^2}{a^2}} X_0^2 \; ; \; Y = \left[ \frac{c^2 - \frac{y^2 + x^2}{a^2}}{\kappa_x^2 - \frac{y^2 + x^2}{a^2}} \right]^{1/2} X_0 \; ; \; Z = 0
\]

1.1

2. The input values \( p, p', \tau_l, \tau_r \) for different cases

At first we consider the case in which recording is made with an artificial head and the reproducing happens through headphones. On base of figure 2, it is clear that:

\[
p f(t) \Rightarrow p \Rightarrow p f(t-\tau) \; ; \; p f(t-c_{10}/c_0) \Rightarrow p \Rightarrow p f(t-\tau)
\]

From these above equations follows:

\[
p_l = \frac{F_{l0}^2 - 2 F_{r0} \left[ \frac{c^2 - \frac{y^2 + x^2}{a^2}}{\kappa_x^2 - \frac{y^2 + x^2}{a^2}} \right]^{1/2} \tau_l = \frac{c_{10}^2}{c_0}}{F_{r0}^2 - 2 F_{r0} \left[ \frac{c^2 - \frac{y^2 + x^2}{a^2}}{\kappa_x^2 - \frac{y^2 + x^2}{a^2}} \right]^{1/2} \tau_r = \frac{c_{10}^2}{c_0}}
\]

where:

\[
\tau_l = \frac{c_{10}^2}{c_0} \quad \tau_r = \frac{c_{10}^2}{c_0}
\]

At second we consider the case in which the recording is made with a pair of microphones and the reproducing happens by a pair of loudspeakers. On base of figure 3, we can give the outgoing equations which can be used to determine the input values:

\[
p \left[ \frac{e_{l0} e_{r0} - e_{l0} e_{r0}}{P_{r0} P_{r0}} f(t - \frac{c_{10}^2}{c_0}, \tau_l) + \frac{e_{l0} e_{r0} - e_{l0} e_{r0}}{P_{r0} P_{r0}} f(t - \frac{c_{10}^2}{c_0}, \tau_r) \right] = p f(t - \tau)
\]

\[
p \left[ \frac{e_{l0} e_{r0} - e_{l0} e_{r0}}{P_{r0} P_{r0}} f(t - \frac{c_{10}^2}{c_0}, \tau_l) + \frac{e_{l0} e_{r0} - e_{l0} e_{r0}}{P_{r0} P_{r0}} f(t - \frac{c_{10}^2}{c_0}, \tau_r) \right] = p f(t - \tau)
\]

If \( f(t) \) is periodic \( \sin, \cos \) or a trigonometric series, the addition at the left hand side is not difficult. So \( \tau_l, \tau_r \) are derivable and determinable.

3. Special results (used: artificial head and headphones)

To get the course of the phantom source depending on the moving parameters of the real source we use the expressions 1.1 and \( \beta_0^2 \beta_0^2 \) from part 2, and eliminate the variable \( \kappa^2 \). After this:

\[
y^2 + x^2 \cdot (1 - \frac{d^2}{d^2}) = d^2 - d^2 + y^2
\]

the course of the phantom source

\[
x = x \cdot \frac{d}{d^2} \; ; \; y = \sqrt{y^2 + x^2 \cdot (1 - \frac{d^2}{d^2}) + d^2 - d^2}^{1/2}
\]

the place coordinates

The different courses depending on the parameters \( d, d' \) and \( y' \) are shown in figure 4.
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\[ \text{figure 1.} \]

\[ \text{figure 2.} \]

\[ \text{figure 3.} \]

\[ \text{figure 4.} \]
ULTRASONIC VIBRATORY SYSTEM FOR A VIBRODRIVE

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One of the ways of improving tape drive and electrical playback mechanisms in magnetic recording and playback devices is related with the development of a new type of motors, i.e. vibrodribe. The principle of action of an ultrasonic vibrodribe is based on the transformation of vibration motion of a working element into the rotational motion of a rotor during their impact interaction. The device allowing to obtain the rotation of a rotor during bending vibrations of a piezoelectric rod vibrator has been proposed for the first time in 1948 (1). A number of alternative devices of ultrasonic vibrodribe have been described elsewhere (see, for example, (2-3). It was shown that applying the ultrasonic vibrodribe in recording and playback devices allows to decrease the level of a low-frequency noise and to create a low-power drive which is more effective and more simple in manufacturing.

This paper describes one of the designs of ultrasonic vibrodribe, in which the rod vibratory system is used as a working element. The designs of such a type have been used in HiFi playback devices (4). The principal scheme of a device is shown in Fig. 1. The end of a rod vibratory system (2) is pressed against the rotor (1). The system performs longitudinal ultrasonic vibrations, and its axis is inclined to the tangential plane. The amplitude of vibrational displacements of the system end is specified by means of applying the electrical voltage at frequency \( \omega \) corresponding to the mechanical resonance frequency of a vibratory system. This is accomplished by sending the electrical signal from the piezoelectrical feedback sensor (3) to the input of the electrical oscillator (4). The devices allowing to control the displacement amplitude and the applied force are provided.

Assuming that the friction coefficient does not depend on the relative velocity of the vibratory system end and
on the rotor surface, the mean value of the force moment \( \bar{M} \) during the time of contact, \( t \), was obtained as

\[
(1) \quad \bar{M} = \frac{\omega e R}{2\pi} \int_0^\infty F_n(t) dt = \frac{\omega e R t}{4\pi} F_{n_{\text{max}}}
\]

Here \( R \) is the radius of the cylindrical surface of a rotor \( e \) is the friction coefficient of the pair "the rotor—the vibratory system end"; \( F_n(t) \) is the normal component of an instantaneous force acting in the contact zone; \( F_{n_{\text{max}}} \) is the maximum value of this force during the contact. The experimental studies \((5)\) have shown that the duration of contact and the maximum value of the acting force, depend on both the applied force, \( F_0 \) and the amplitude of vibrational displacements, \( \delta_m \). This means that the main characteristics of a vibrodrive, i.e., the rotation rate and the acceleration time, also depend on both the displacement amplitude and the applied force. The character of the dependence of the rotation rate, \( n \), on the abovementioned parameters is shown in Fig. 2. The dotted line restricts the area of working values of the applied force. It was shown that the optimum value of the amplitude for a vibrodrive equals some microns. Thus, the calculation of a vibratory system with the purpose of obtaining the specified amplitude of vibrational displacements is one of the main tasks of the design.

Fig. 3 shows a vibratory rod system \((6)\). The cylindrical piezoelement \((1)\) with the height \( l \), is suppressed between two metal rods \((2, 3)\) having the heights \( l_2 \) and \( l_3 \) respectively. The end of such a composite transducer is attached to the rod \((4)\) with the length \( l_4 \), that serves as a concentrator. This rod is a working element of a vibrodrive. We introduce the following designations: \( Z_\kappa = \rho_\kappa c_\kappa S_\kappa \) where \( \rho_\kappa \) is the material density of any rod \((k = 1, 2, 3, 4); c_\kappa \) is the rod sound velocity, \( S_\kappa \) is the cross section area, \( \lambda_\kappa = \omega l_\kappa \) is the wave length of a rod, \( Q_\kappa \) is the quality factor.

The solution of a wave equation for a four-rod system, due regard for boundary conditions and a piezoeffect, yielded the following expression for the displacement amplitude at the working end:

\[
(2) \quad \delta_m = \frac{\delta_0}{\cos \lambda \nu + i(\nu \lambda / \lambda \nu + \nu e) \sin \lambda \nu}
\]

Here \( \delta_0 \) is the displacement amplitude at the transducer end, \( \nu e \) is the loading coefficient, equal to the ratio of the load resistance, \( R_e \), to \( Z_\nu \). The amplitude \( \delta_0 \) may be found from the equation

\[
(3) \quad \delta_0 = dU_m (\cos \lambda_2 - \nu_e) (\alpha, \phi)^{-1}
\]
Kasantsev... Ultrasonic vibratory system for a vibrodrive

where

\[
\Phi = \frac{1}{2} \left( \frac{Y_1}{Y_2} \cdot \sin \alpha_2 + \frac{Y_1}{Y_2} \cos \alpha_2 - \frac{1}{\nu} \cdot \frac{Y_1}{Y_2} \left( \frac{Y_1}{Y_2} \cos \alpha_2 - \frac{Y_1}{Y_2} \sin \alpha_2 \right) \right);
\]

\[
\nu_1 = \cos \alpha_1, \cos \alpha_2 - \frac{Y_1}{Y_2} \sin \alpha_1, \sin \alpha_2;
\]

\[
\nu_2 = \sin \alpha_1, \cos \alpha_2 - \frac{Y_1}{Y_2} \sin \alpha_1, \sin \alpha_2.
\]

Here $d$ is the piezoconstant, $U_m$ is the electrical voltage amplitude at the system input, $j \phi = \frac{Z_i}{Z_o}$.

The computer calculation was carried out with the purpose of parameter optimization. It has been shown that for small loading coefficients, $\frac{Y_1}{Y_2} < 0.2$ the amplitude $\frac{\Phi}{\Phi_0}$ is maximum when two conditions are satisfied, $\alpha_1 + \alpha_2 + \alpha_3 = \pi$ and $\alpha_\nu = \pi/2$. These conditions determine the resonance length of a vibrational system, which is equal to $\lambda \pi / 4$ ($\lambda$ is the sound wavelength). The length of the rod (4) is close to $\lambda / 4$. The gain $\kappa_\nu = \frac{\Phi}{\Phi_0}$ essentially depends on the load resistance value and on the losses in the rod material. The limit value is $\kappa_\nu = \frac{\pi \lambda / 4}{\Phi_0}$ when $\Phi_0 \rightarrow 0$. In our experiments the value $\kappa_\nu \approx 30\%$ has been obtained. It should be noted that the presence of the rod (4) does not change practically the character of the amplitude distribution over the length of the composite piezoelectrical transducer. This distribution remains the same as for the usual rod of halfwave length.

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Figure 1. The vibrodrive scheme. 1 - rotor; 2 - vibratory rod system; 3 - feedback element; 4 - electrical oscillator.

Figure 2. The rotation rate, $n$, depending on the applied force, $F_0$, for different displacement amplitudes.

Figure 3. The vibratory rod system design. 1 - piezoelement; 2-3 - passive metal rods; 4 - concentrator rod.
DIRECTIVITY ASPECTS OF BODY-WORN MICROPHONES

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1. Survey

Body-worn microphones are known as lapel, tie-tack or lavalier microphones. Nearly all of them are of pressure type with the human body influencing the directional characteristics. Experience was gathered with a new directional type, which showed superior results under critical acoustical conditions. The practical results are supported by measurement results showing the differences of both types worn on the human body.

2. The acoustical problem

The voice of single speakers or singers shall be picked up and the acoustic ambience shall be suppressed to a convenient amount. This affords either highly directive microphones in some meters distance or the microphone in close proximity of the sound source. Unwanted influences are reverberant sound of the room, other sources and very often acoustical feedback in case of p.a. systems in the same room.

3. Possible solutions

Line or shotgun microphones have a very high directivity but need highly skilled sound technicians because they have to be adjusted properly to the sound source. If in addition the acoustic ambience is of importance the off-axis coloration of the sound is sometimes too high.

A very short distance between microphone and mouth is achieved by wearing microphones on the body, mainly in the middle of the chest resulting in approximately 30 cm distance to the mouth. The dynamic types used formerly have mainly been replaced by electret condenser microphones which can be designed extremely small. The microphone leads between actor and mixer may be disadvantageous. This is the reason for increased use of wireless techniques. Mainly in connection with FM-modulated transmitters for wireless microphones one gets complete mobility of the actor together with high acoustic quality.
This contribution deals with body-worn microphones. The measurements were done with an omnidirectional and a directional microphone. The omnidirectional microphone is of very flat construction so that it can be fixed to the clothing without risk of extended noise due to rubbing on textile. The flat design results in the membrane facing the auditory. The raise of the frequency response at 8 kHz and above has been taken into account in the acoustical design. The unidirectional microphone is of cardioid type. The membrane points towards the mouth of the actor.

4. Measurement results

Under practical conditions the microphone output is the response to a superposition of three different sound fields. The wanted field of the actor is practically of spherical type. Ambient sound is as well directional as diffuse. The measurements were done with respect to these three types of excitation.

The curves have to be compared with the free-field frequency response of both microphones. Straight lines show the 0°-free-field response of the pressure microphone (figure 1 a) and of the pressure gradient microphone (figure 1 b). The dotted lines give the response under 90°. The omnidirectional microphone shows only above 8 kHz a clearly increase due to diffraction. The unidirectional type is nearly up to 10 kHz an ideal cardioid with a 6 dB suppression under 90°.

The response to diffuse sound was measured in a reflective chamber. The response of both microphones for the same pressure as in the free field mounted on a tripod and mounted at the human body is also given in figure 1. On a tripod the omnidirectional microphone practically answers with the free-field response. The unidirectional microphone shows the directivity as expected. Both responses have a wide maximum around 500 Hz. This may result from the width of the human chest. Frequencies above 2 kHz are partially absorbed by the body.

The influence of directive external sources was examined in the anechoic chamber. The influence of the body is highly depending on the angle of incidence (figure 2). The 0° response shows again the wide maximum at 500 Hz. With the sound waves travelling parallel to the chest (90°-incidence) the microphones behave as in the free-field at lower frequencies whereas higher frequencies are of course irregularly reflected and absorbed. The curves are referred to the free-field response of each microphone. An exact explanation of these maxima and minima cannot yet be given due to the measurement tolerances. It is planned to calculate the responses for a computer model. The literature shows results for simple geometrical shapes and non-absorbing material (1).

The main signal is originated by the wearer of the microphone. The corresponding measurements were made with an artificial voice. The manikin consists of chest and head, the loudspeaker being located in
the chest and the acoustical waves brought to the mouth via an acoustical duct. A measurement microphone straight in front of the corner of the mouth allows control to constant sound pressure level from the mouth.

Figure 3 shows the differences at three different rotating angles of the head. It can be seen that the differences measured with the cardioid microphone are not larger than those of the omnidirectional type. The reason is that even a $\pm 60^\circ$ turn of the head results in a far smaller angle referred to the axis of a microphone on the middle of the chest. To use this advantage it is necessary to place the microphone symmetrically in approximately 30 cm distance to the mouth. Too small a distance or even placement at one side of the chest increases the ripples already known from omnidirectional microphones.

5. Conclusions

The measurements confirmed the practical experience that by using cardioid microphones no audible increase of level differences occur. It was also confirmed that the directivity is not affected so that it may be used as additional gain without increasing the risk of acoustical feedback.

Literature (1)


Fig. 1 Free field and diffuse field response of omnidirectional (1a) and unidirectional (1b) microphone
Fig. 2 Body-worn microphones and angle of incidence

Fig. 3 Response to artificial voice at different angles of the head
6.4

Systèmes microphoniques. Antennes
Microphone arrays. Antennae
Mikrofonsysteme. Antennen
UTILISATION D'ANTENNES FOCALISEES POUR LA LOCALISATION DES SOURCES ACOUSTIQUES

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Introduction

Pour améliorer le pouvoir de résolution spatiale d'une antenne acoustique, on a intérêt à rapprocher l'antenne de la région-source. Cependant, dès que la région-source entre dans la zone de Fresnel de l'antenne, les ondes acoustiques ne peuvent plus être supposées planes et le traitement classique fondé sur une relation de transformée de Fourier entre le champ de pression sur l'antenne et la direction des sources ne s'applique plus. On doit alors tenir compte du caractère sphérique des ondes et procéder à un traitement d'antenne focalisée [1], [2]. De plus pour des applications en soufflerie, il faut adapter le traitement aux effets de propagation (convection et réfraction) qui modifient la géométrie des fronts d'ondes. Dans la description des performances d'une telle antenne focalisée, on se limite ici au cas d'une antenne linéaire de longueur L, composée de N capteurs équidistants de δ, les sources étant supposées se situer sur une ligne parallèle à l'antenne à la distance D de celle-ci.

Principes de base du traitement : Soit une source placée en $\vec{x}_c$, de nombre d'ondes $k$ et d'amplitude $A$. Le signal reçu sur le $i$-ème capteur à la position est :

$$p_i(\vec{y}_i) = A \exp(-ik'_{ij} - \vec{x}_i) / |\vec{y}_i - \vec{x}_i|$$

Le traitement (linéaire) consiste à choisir un point de focalisation $\vec{z}$, à multiplier $p_i(\vec{y}_i)$ par un facteur de phase $-ik'_{ij} - \vec{x}_i$ et un facteur d'amplitude, et sommer l'ensemble des N signaux ainsi corrigés. A priori, le choix de la pondération en amplitude est largement arbitraire. Une manière de lever cette ambiguïté est de poser le problème en terme d'estimation de $A$ et de $\vec{x}_c$, en supposant les signaux $p_i(\vec{y}_i)$ bruités (bruits blancs gaussiens indépendants). On obtient alors les résultats suivants. Pour un point de focalisation $\vec{z}$, la meilleure estimation de $A$ est (1)

$$\hat{A}(\vec{z}) = \frac{1}{N} \sum_{i=1}^{N} \frac{1}{|\vec{y}_i - \vec{x}_i|}$$

(1) $p_i(\vec{y}_i) = e^{-ik'_{ij} - \vec{x}_i} / |\vec{y}_i - \vec{x}_i|$ (on note qu'en l'absence de bruit, on trouve bien $\hat{A}(\vec{z}) = A$. Par contre, on n'est pas assuré que $|\hat{A}(\vec{z})|$ passe par un maximum pour $\vec{z} = \vec{x}_c$. Pour cela il faut chercher le
maximum de \( B(x) \). Il apparait donc que contrairement au cas d’une antenne afocale, les opérations pour chercher la position de la source et son amplitude ne sont pas identiques. Heureusement, les simulations numériques montrent que lorsque la source est au cœur de la zone de Fresnel de l’antenne, \( |\hat{A}(x)| \) a son maximum à proximité immédiate de \( x_0 \), de telle sorte que le traitement (1) assure les deux opérations.

Avant de poursuivre, il est intéressant de comparer le traitement (1) à un traitement afocal, de type transformée de Fourier. La figure 1 montre une simulation pour une antenne composée de \( N=21 \) capteurs, la source étant face au centre de l’antenne, à la distance \( D=L \). La fréquence d’émission est repérée par le nombre d’ondes réduit \( K=\frac{\mu}{\lambda} \). Dans ce cas, la zone de Fresnel de l’antenne sur son axe est limitee à \( K_0>L \). La source est donc dans cette zone si \( K>26 \). Il apparaît bien que le traitement afocal est inopérant pour \( K=50 \), et ne reste qu’à peine acceptable à la limite de la zone de Fresnel (\( K=20 \)).

**Performances de l’antenne focalisée**

Le pouvoir de résolution angulaire d’une antenne afocale \( \Delta(\theta) \) conduit à un pouvoir de résolution spatiale à la distance \( D \) qui s’écrit \( \Delta X = \frac{\lambda}{D} \) soit : (3) \( \Delta X = \frac{\lambda}{D L} \). À la limite de la zone de Fresnel, on montre facilement que \( \Delta X \gg L \), soit \( \Delta X \gg \lambda \) (puisque (3) suppose que \( \lambda \ll L \)). Une expression approchée de \( B(x) \) permet de montrer que (3) continue à s’appliquer pour le traitement focalisé à condition que \( D \gg L \), ce qui permet d’approcher par traitement focalisé la résolution \( \Delta X \gg \lambda \). Effectivement, les passages à zéro de \( |\hat{A}(x)|^2 \) de la figure 1 sont observés en \( \lambda/L = 0,32 \) et 0,13 pour \( K=20 \) et \( K=50 \), conformément à (3) qui prévoit pour \( \theta = 0 \) \( \Delta X \gg \lambda/L = 2\pi/\lambda \). Au-delà, c’est-à-dire pour \( L \gg D \), le traitement focalisé fournit \( A(x) \approx \sum_j (c_j \delta) \). Le pouvoir de résolution ultime est donc \( \Delta X_{\text{min}} = 0,38 \lambda \). La conséquence pratique de cette évaluation est qu’à partir d’une longueur \( L \approx D \), on ne gagne que peu en résolution à allonger la base de l’antenne.

Une des caractéristiques de l’antenne focalisée est l’existence d’une profondeur de champ, qui permet de retrouver la distance \( D \) de la ligne des sources. En définissant cette profondeur de champ par la distance \( \Delta \) telle que \( B(x) \) tombe à 25% de sa valeur nominale, un calcul approché donne : (5) \( \Delta \approx 5,2 \lambda (D/L)^2 \) pour une source placée face au centre de l’antenne.

C’est ce que confirme la simulation présentée sur la figure 2, où la source est encore à la distance \( D=L \) de l’antenne. Lorsque \( K \) est faible, la source est dans la zone de Fraunhofer et la profondeur de champ est pratiquement infinie. Pour \( K>26 \), elle est dans la zone de Fresnel et l’approximation (5) est bien vérifiée pour \( K>50 \).

De même que pour une antenne afocale, l’échantillonnage spatial des signaux de l’antenne induit un repliement spatial autour des positions définies par \( \Delta(\theta=\pi/2) \) par \( \pi/\delta \). Cependant, la réponse de l’antenne n’est pas strictement périodique car il n’existe pas deux positions de source telles que la trace des ondes sphériques qu’elles émettent soient...
rigoureusement identiques aux positions d'échantillonnage. Les remontées de B (x) aux positions de repliement sont ainsi fonction de la position de la source, de la fréquence et du nombre de capteurs. Elles restent le plus souvent très inférieures à la valeur du pic principal.

Lors d'essais en soufflerie, les effets de convection (pour une soufflerie à veine fermée) et de réfraction à la traversée de la zone de mélange du jet libre (pour une soufflerie à veine ouverte) induisent une modification du champ acoustique sur l'antenne pour une source monopolaire, dont le traitement doit tenir compte. Ces effets sont bien caractérisés, et sont modélisables par le calcul de la trajectoire des rayons sources [3]. Le traitement est alors analogue au traitement sans écoulement, au remplacement près du facteur de phase k / √(2πf z) et du facteur d'amplitude 1 / |√(2πf z)| dans (1) par les valeurs calculées à l'aide du modèle de propagation.

En pratique, les signaux de l'antenne sont enregistrés sur enregistreur magnétique analogique 28 pistes et le traitement est effectué à l'aide d'un calculateur parallèle. Typiquement, une "image" comportant 100 points de focalisation sur la ligne-source et 128 valeurs de la fréquence est calculée en 5s pour N=20 capteurs. La figure 3 montre une telle image en représentation tridimensionnelle amplitude-distance-fréquence, obtenue à l'aide d'une source électroacoustique émettant à la fréquence de 1500 Hz dans une soufflerie à veine fermée avec un écoulement de 50 m/s. On a choisi D=L, de telle sorte que le pic à la fréquence de la source s'étend sur une distance de 20 cm, soit sensiblement une longueur d'onde, conformément à (3). De même, la figure 4 est l'image équivalente monopolaire d'un bruit de jet obtenu en soufflerie à veine ouverte (V=90 m/s). Elle résulte d'une moyenne effectuée sur 40 images élémentaires.

Conclusion Les considérations précédentes donnent un aperçu des possibilités d'une antenne focalisée, dont l'intérêt principal est de permettre d'atteindre un pouvoir de résolution spatiale de l'ordre de la longueur d'onde, ainsi que la comparaison de ses performances par rapport à une antenne focale. Il faut cependant insister sur le fait que le traitement présenté ici est fondé sur une représentation monopolaire équivalente des sources, modèle dont l'utilisateur doit être conscient pour l'interprétation des images obtenues et leur correspondance avec les sources de bruit réelles.

1. $|\hat{A}(x)|^2$

Fig. 1 — $N = 21$, $D/L = 1$. Position de la source $X = 0$.
- traitement focalisé
- traitement afocal.

2. $S(x)$

Fig. 2 — Profondeur de champ $N = 21$. Position de la source. $D/L = 1$, $X = 0$.

3. Image d'une source électroacoustique en représentation fréquence-distance. Vitesse de soufflerie : 51 m/s.

LEAST-SQUARES METHOD FOR SOURCE SOUND SEPARATION IN MULTI-SOURCE ENVIRONMENT

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1. INTRODUCTION
Least-squares method is applied to multi-channel deconvolution. A concept of matrix convolution is introduced as an expression of multi-channel convolution and the generalized convolutional inverse matrix[1] is employed to get the time domain expression for least-squares multi-channel deconvolution. A practical experiment of source sound separation in a multi-source environment[2,3] is described assuming the time invariance of the transfer characteristics among the sound sources and microphones.

2. MULTI-CHANNEL CONVOLUTION
Suppose that the output $f'_i(t)$ of a linear $n$-input $m$-output system is the sum of the convolutions of the impulse responses $h_{ij}(t)$ of the channels involved and their corresponding inputs $s_j(t)$, that is

$$f'_i(t) = \sum_{j=1}^{n} h_{ij}(t) * s_j(t), \quad \text{for } i=1,2,\ldots,m$$

(1)

where $n$ and $m$ are the numbers of input and output terminals and $*$ denotes convolution integral. Here, we introduce a concept of multi-channel convolution for convenience of expression. Equation (1) can be forcibly rewritten in matrix form as

$$f(t) = h(t) \circ s(t)$$

(2)

where

$$f(t) = \begin{bmatrix} f_1(t) \\ \vdots \\ f_m(t) \end{bmatrix}, \quad h(t) = \begin{bmatrix} h_{11}(t) & \cdots & h_{1n}(t) \\ \vdots & \ddots & \vdots \\ h_{m1}(t) & \cdots & h_{mn}(t) \end{bmatrix}, \quad s(t) = \begin{bmatrix} s_1(t) \\ \vdots \\ s_n(t) \end{bmatrix},$$

and $\circ$ denotes the matrix convolution or multi-channel convolution.

3. MULTI-CHANNEL DECONVOLUTION
Multi-channel deconvolution is, then, defined as to solve eq.(2) for $s(t)$. It, however, seems hard to solve eq.(2) directly in the time domain. Then, we take the Fourier transform of eq.(1) to get the frequency domain description for it.

$$F'_i(\omega) = \sum_{j=1}^{n} H_{ij}(\omega) S_j(\omega), \quad \text{for } i=1,2,\ldots,m$$

(3)

where
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\[ P(\omega) = H(\omega) S(\omega) \]

where

\[ P(\omega) = \begin{bmatrix} P_{11}(\omega) & \cdots & P_{1n}(\omega) \\ \vdots & \ddots & \vdots \\ P_{m1}(\omega) & \cdots & P_{mn}(\omega) \end{bmatrix}, \quad H(\omega) = \begin{bmatrix} H_{11}(\omega) & \cdots & H_{1n}(\omega) \\ \vdots & \ddots & \vdots \\ H_{m1}(\omega) & \cdots & H_{mn}(\omega) \end{bmatrix}, \quad S(\omega) = \begin{bmatrix} S_{1}(\omega) \\ \vdots \\ S_{n}(\omega) \end{bmatrix}. \]

If \( m \geq n \geq \text{rank}(H(\omega)) \), there exists the least-squares solution \( \hat{S}(\omega) \) for \( S(\omega) \) and is obtained from \( H(\omega) \) and \( P(\omega) \) as

\[ \hat{S}(\omega) = H^+ (\omega) P(\omega), \]

where \( H^+ (\omega) \) is the generalized inverse of \( H(\omega) \). Then, the time domain solution is expressed as

\[ \hat{s}(t) = F^{-1}[H^+ (\omega) P(\omega)] \]

where \( F^{-1} \) denotes the Fourier inverse transform.

To formulate the deconvolution as a direct time domain expression, a generalized convolutional inverse matrix \( h^\oplus(t) \) should be defined as

\[ \hat{s}(t) = h^\oplus(t) \ast f(t) \]

where

\[ h^\oplus(t) = F^{-1}[H^+ (\omega)]. \]

This is regarded as the formal expression of the least-squares multi-channel deconvolution.

It would be suggestive to note the properties of the generalized convolutional inverse matrix. The generalized convolutional inverse matrix of our type satisfies the following equation:

\[ h^\oplus_n(t) \ast h(t) = \delta^{(r)} (t) \]

or

\[ \sum_{k=1}^{m} h^\oplus_{ik}(t) \ast h_{kj}(t) = \delta^{(r)}_{ji}(t) \]

\[ \delta(t) \quad i=j \leq r \]

\[ 0 \quad \text{otherwise}. \]

This is easily obtained from the Fourier inverse transform of the frequency domain relation

\[ H^+ (\omega) H(\omega) = I^{(r)}_n = \begin{bmatrix} I^p & 0 \\ 0 & 0 \end{bmatrix} \]

where \( I^p \) is the unit matrix of order \( p \).

Equation (10) shows that \( h^\oplus_n(t) \) can be calculated by successive time domain computation directly from \( h_{ij}(t) \), but a simple method in the frequency domain based on eqs. (11) and (8) is more efficient.

4. SOURCE SOUND SEPARATION

The above-mentioned multi-channel deconvolution by using the generalized convolutional inverse matrix is applied to source signal estimation or sound separation in a multi-source environment.

The Model: Suppose that \( s_j(t) \) is the waveform of source \( S_j \), \( f_k(t) \) is the waveform received by microphone \( M_k \) and that \( h_{ij}(t) \) is the impulse response of the sound propagation path from \( S_j \) to \( M_i \). Here, the locations of the
sound sources and microphones are assumed to be fixed and the field is assumed to be linear and time invariant.

**The Problem:** To separate the source sounds one by one or to get the least-squares estimate of each source signal.

**Preparatory Processing:** Before the separation process, the following procedures are required in advance:

(i) In order to get \( \mathbf{H}^i(t) \) for executing eq.(7), the impulse response \( h_{ii}^i(t) \) or its Fourier transform \( H_{ii}^i(\omega) \) for each pair of input and output terminals is to be measured beforehand. \( H_{ij}^i(\omega) \) is obtained as

\[
H_{ij}^i(\omega) = \frac{F[f_i^i(t)]}{F[s_{ji}^i(t)]} \quad \left| s_{ji}^i(t)=0, \ k \neq j \right.
\]

and is stored in a three-dimensional complex array. Impulse-like sounds would be preferable for \( s_{ji}^i(t) \) in eq.(12).

(ii) Calculate \( \mathbf{H}^i(\omega) \) for each sample point on the frequency axis and store it again in the 3-D array used in (i).
In most cases, rank[\( \mathbf{H}(\omega) \)] = \( n \) holds and

\[
\mathbf{H}^i(\omega) = [\mathbf{H}(\omega) \mathbf{H}(\omega)]^{-1} \mathbf{H}(\omega)
\]

where * denotes the conjugate transpose. In case of large \( m \) and \( n \), an efficient computation algorithm is available[4].

(iii) Calculate \( h_{ji}^i(\omega) \) as

\[
h_{ji}^i(t) = F^{-1} [h_{ji}^i(\omega)]
\]

for desired \( j \) corresponding to \( s_j^i(t) \) to be estimated.

**Separation Process:** According to eq.(7), the least-squares estimate for \( s_j^i(t) \) is obtained as

\[
s_j^i(t) \approx \sum_{l=1}^{m} h_{ji}^l(t) * f_i^l(t) \quad \forall i, j
\]

The whole computation process is depicted in Fig.1.

5. **EXPERIMENT**

The experiment was conducted for two-source three-microphone case. It was carried out in an anechoic chamber in order to avoid environmental noise, but there were brought many reflecting boards to simulate an ordinary room situation. The two loudspeakers and three microphones were situated asymmetrically. Impulse-like sounds were used as \( s_j^i(t) \) in eq. (12) and pre-recorded vowels /a/ and /i/ were used as the source sounds to be separated. The performance of source sound separation is shown in Fig.2, where (a) shows the received waveforms and their frequency spectra and (b), the estimated waveforms and their spectra, and (c), the spectra of the sources.

6. **PRACTICAL MEANS TO IMPROVE THE SEPARATION PERFORMANCE**

Some possible causes that degrade the performance of the proposed method can be thought of, and the following seems dominant among them.

(i) The method can be regarded as a statistical processing because it is based on the least-squares criterion. However, especially in case of
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Fig. 2 Separation of mixed vowel speech /a/ and /i/.

(a) The waveforms received by the three microphones and their frequency spectra.

(b) Estimated waveforms and their frequency spectra.

(c) Frequency spectra of the source sounds.

Taking those matters into account, it may be possible to improve the separation performance by selective use of microphones and/or averaging the transfer characteristics over several times of impulse response measurements. Figure 3 shows the results of improvement for the previous experiment by averaging 120 times of the impulse response measurements. Comparing $S_2(\omega)$ in Fig. 3 with that in Fig. 2(b), a remarkable improvement is recognized in superiority of $F_1$ and $F_2$ of $S_2(\omega)$ over $F_1$, $F_2$, and $F_3$ of $S_1(\omega)$. If much more microphones are available, selective use of microphones would be effective.

Quantitative and perceptual evaluation for this method is under investigation.

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REFERENCES
LEAST-SQUARES METHOD FOR SOUND-IMAGE LOCALIZATION IN MULTI-SPEAKER MULTI-LISTENER CONFIGURATION

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1. INTRODUCTION
Recently, new approaches for sound-image localization[1,2] have been developed based on a model of hearing. However, they are applicable only to two-loudspeaker one-listener case. The method proposed here is their extension to multi-loudspeaker multi-listener case. Although our method cannot offer the optimal configuration of loudspeakers and listeners, it provides the least-squares solution for the transfer characteristics of the channels to the loudspeakers for given configuration of loudspeakers and listeners. This method is one of the applications of the generalized inverse matrices to acoustic signal processing[3].

2. THEORY of SOUND-IMAGE LOCALIZATION for MULTI-LOUDSPEAKER MULTI-LISTENER CASE
The signals at the ears of listeners are characterized by acoustic transfer functions associated with the propagation paths from the sound sources to the ears as illustrated in Fig.1, where S denotes the source characteristics.

The sound pressure \( P_L \) and \( P_R \) produced acoustically at the left and right ears are expressed as

\[
\begin{align*}
P_L &= H_L S \\
P_R &= H_R S.
\end{align*}
\]  

(1)

The acoustic transfer functions between the loudspeakers \( S_j (j=1,2,...,n) \) and both ears of the listeners \( L_i (i=1,2,...,m) \) are shown in Fig.2, where \( T_{ijL} \) and \( T_{ijR} \) denote the transfer functions of the sound propagation paths from \( S_j \) to the left and right ears of \( L_i \), respectively.

Multi-loudspeaker multi-listener sound-image localization can be realized by inserting channel compensation network \( C_{ij} \) into each signal channel to the loudspeaker \( S_j \).

Fig.1 A model of hearing.

\( S \) : Sound source to be localized
\( L \) : Listener
\( H_L, H_R \) : Required transfer function
\( P_L, P_R \) : Sound pressure characteristics to be perceived.
The sound pressure $P'_L$ and $P'_R$ produced at the left and right ears are the summation of the products of the voltages supplied to the loudspeakers and the transfer functions of the sound propagation paths as follows:

\[
P'_L = \sum_{j=1}^{n} T_{i,jL} C_j S, \quad P'_R = \sum_{j=1}^{n} T_{i,jR} C_j S \quad \text{for } i=1,2,\ldots,m \tag{2}
\]

where $S$ denotes the source characteristics.

In order to obtain exact sound-image localization, the following relation should hold:

\[
\begin{align*}
P'_L &= P_L \
P'_R &= P_R
\end{align*}
\]

(3)

or

\[
\begin{align*}
\sum_{j=1}^{n} T_{i,jL} C_j &= H_L \
\sum_{j=1}^{n} T_{i,jR} C_j &= H_R
\end{align*}
\]

(4)

for $i=1,2,\ldots,m$.

With matrix notation, eq. (4) can be rewritten as

\[
T C = H, \quad 2m \times n, n \times 1, 2m \times 1
\]

(5)

where

\[
T = \begin{bmatrix}
T_{11L} & T_{12L} & & & & T_{1nL} \\
& & & & & \\
& & & & & \\
T_{m1L} & T_{m2L} & & & & T_{mnL} \\
& & & & & \\
& & & & & \\
T_{11R} & T_{12R} & & & & T_{1nR} \\
& & & & & \\
& & & & & \\
T_{m1R} & T_{m2R} & & & & T_{mnR}
\end{bmatrix}
\]

\[
C = \begin{bmatrix}
C_1 \\
\vdots \\
C_n
\end{bmatrix}
\]

and

\[
H = \begin{bmatrix}
H_L \\
\vdots \\
H_R
\end{bmatrix}
\]

(6)

Fig. 2 The scheme of sound-image localization for $n$-loudspeaker $m$-listener case.

\[
T_{i,jL/R} : \text{Transfer function from loudspeaker } L_j^i \text{ to the left/right ear of listener } R_j^i.
C_j : \text{Compensation network for channel } j.
S_i : \text{Desired sound-image.}
P'_L/R : \text{Sound pressure characteristics perceived by the left/right ear of each listener.}
\]
\[ \hat{C} = T^+ H \] (7)

where \( T^+ \) denotes the generalized inverse of \( T \), and \( \hat{C} \) means the l.s. estimate for \( C \). If different sound-image localization is required among listeners, the definition of \( H \) should be modified as

\[ H = [ H_{1L} \ldots \ H_{mL} \ H_{1R} \ldots \ H_{mR} ]^T, \] (8)

where \( H_{iL} \) and \( H_{iR} \) correspond to \( H_L \) and \( H_R \) for listener \( L_i \).

3. EXACT SOLUTION FOR SOME SPECIAL CASES

The proposed method gives only the l.s. solution for the characteristics of the channel compensation network \( C \), to be inserted into the channel to loudspeaker \( S_j \). That means the proposed method cannot realize the desired sound field exactly. The sound field generated by the proposed method is merely the approximated one under the l.s. criterion. There, however, can be some cases in which the exact solution exists. Those are the cases as follows:

(A) Two-loudspeaker one-listener case

For this case, \( 2m=n \) holds and eq.\,(7) yields the exact solution

\[ C = \begin{bmatrix} T_{11L} & T_{12L} \\ T_{11R} & T_{12R} \end{bmatrix}^{-1} \begin{bmatrix} H_L \\ H_R \end{bmatrix}, \] (9)

provided that the inverse matrix exists for any frequency point under consideration. By locating the two loudspeakers at the symmetrical positions concerning the median plane of the listener, the following approximation holds:

\[ T_{11L} \approx T_{12R} \quad ( = T_S ), \quad T_{11R} \approx T_{12L} \quad ( = T_C ) \] (10)

where \( T_S \) and \( T_C \) denote the transfer characteristics of the channels from each loudspeaker to the ear of the same side and that to the opposite side, respectively. Under this approximation, eq.\,(9) can be simplified as

\[ C = \begin{bmatrix} T_S & T_C \\ T_C & T_S \end{bmatrix}^{-1} \begin{bmatrix} H_L \\ H_R \end{bmatrix}. \] (11)

(B) Multi-listener case with Head-sets

For this case, we can set \( m=2 \) and \( n=1 \) even if the actual number of listeners is more than one provided that the transfer characteristics or the coupling characteristics between each head-set and the corresponding ear are same and that the cross-talk between the left and right channel is negligible. Under these assumption, eq.\,(5) yields

\[ \begin{bmatrix} T_{11L} & 0 \\ 0 & T_{12R} \end{bmatrix} \begin{bmatrix} 0 \\ 0 \end{bmatrix} C = \begin{bmatrix} H_L \\ H_R \end{bmatrix} \] (12)

and we get

\[ C = \begin{bmatrix} H_L/T_{11L} \\ H_R/T_{12R} \end{bmatrix}. \] (13)

Further, \( T_{11L} \) and \( T_{12R} \) can be assumed to be identical to each other for most of the listeners, and eq.\,(13) might be simplified to
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\[ c = \frac{1}{T_H} \begin{bmatrix} H_L \\ H_R \end{bmatrix} \]  \hspace{1cm} (14)

where \[ T_H \triangleq T_{1L} = T_{12R} \].  \hspace{1cm} (15)

4. DISCUSSIONS
In practical applications, the most wanted is the method to determine the number and the configuration of loudspeakers for a given number and configuration of listeners. The proposed method gives the least-squares solution for a given configuration of loudspeakers and listeners, but it does not give any information about the desirable configuration at all. It, however, is possible to calculate the least-squares error for the given numbers and configuration of loudspeakers and listeners, therefore we can evaluate the degree of goodness of the numbers and the configuration of them. Consequently, some conventional search technique in a multi-dimensional space can be employed to find the optimal solution for the numbers and the configuration of loudspeakers and listeners.

As a practical way to obtain the optimal number and configuration of loudspeakers for a given number and configuration of listeners, the following approach would suffice:

**Step I** Assume a number \( n \) somewhat larger than the given number \( m \) of the listeners as the initial number of necessary loudspeakers.

**Step II** Assume an initial configuration for loudspeakers and evaluate the least-squares error.

**Step III** Move the loudspeakers one after another by appropriate width and evaluate the least-squares error for each modified configuration.

**Step IV** If the least-squares error can be reduced by modifying the loudspeaker configuration, renew the configuration and repeat Step III, else select among the following three cases:
- a. Restart from a different initial configuration in Step II.
- b. Increase \( n \) and restart Step II.
- c. Stop (Assume the present configuration is optimal).

5. CONCLUDING REMARKS
At present, any practical experiment has not been conducted yet, but the proposed method gives the exact solution for a two-loudspeaker one-listener case. Therefore, the method seems to be the most relevant way for realizing multi-loudspeaker multi-listener sound reproduction scheme so far as the least-squares criterion has a sense in sound perception.

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6.5

Analyse et traitement du signal. Imagerie, intensimétrie. Holographie

Analysis and signal processing. Statistics. Dosimetry

Signalanalyse- und Aufbereitung. (Sichtbarmachung) Optische Verfahren. Intensitätsmessung. Holographie
TRANSDUCERS FOR INTENSITY MEASUREMENTS

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The use of the two pressure microphone method for acoustic intensity measurements is setting new demands to calibration methods and accuracy. The development of new configurations optimizing the symmetry of the cosine directional pattern is important for the measurement of the true direction of the intensity vector especially useful in source location and very important for intensity measurements in the presence of very reactive acoustic field. The importance of proper $\Delta r$ and phase calibration in order to obtain correct intensity readings in reactive fields is seen from fig. 1, from which the actual phase difference at any given frequency and for a given $\Delta r$ between two pressure microphones may be deduced as a function of the ratio between intensity and pressure.

Fig. 1: Intensity-Pressure-Phase-$\Delta r$-Frequency relationship
Furthermore, the ratio between the intensity level and the pressure level, within which one may expect to be able to measure with a given possible phase error, may be deduced by adding the $\pm$ phase tolerance to the actual phase difference. The importance of small phase difference and of selecting the proper $\Delta$ for optimum performance at the most important frequencies to be measured is then clearly seen.

The acoustic calibration technique used for low frequency calibration of condenser microphones is important. The frequent use of standing wave tubes, fig. 2, or small cavities for calibration leads to large errors due to improper exposure of the total microphone including the air equalization hole from the volume behind the diaphragm of the microphone. The air equalization should not be exposed to a pressure gradient much different from that of the gradient to be sensed by the two diaphragms.

Fig. 2: Calibration of intensity microphones in ducts is wrong and should be avoided.

The design of a typical condenser microphone used for intensity measurements is seen in fig. 3.

The phase shift between input pressure and output voltage at low frequencies of a condenser microphone is mainly determined by the R.C. time constant of the internal volume behind the diaphragm, and the air resistance of the air equalization hole provided in order to equalize the atmospheric pressure. The performance varies greatly whether the air equalization hole is exposed to the pressure variations or not, as seen from fig. 4.

Phase calibration should therefore take place exposing the complete sound intensity probe including the effective air equalization path to the sound field and the $\Delta r$ between the air vent exits should be kept as small as possible.

A good preamplifier configuration using guard ring shielding will mini-
Fig. 4: Low frequency response of a condenser microphone with static pressure equalization vent outside (curve A) and inside (curve B)

mize the significance of any phase shift from the microphone capacity vs. input impedance. Typically, a variation at 20 Hz of from 6 pF to 3 pF causes less than 0.2° of change.

The stability of the relative phase between a matched pair of 1/2" condenser microphones proves to be very good which is to be expected from the mechanical arrangement (fig. 5). 1/4" microphones are more critical regarding long term stability.

The equivalent volume of a 1/2" microphone, type 4165 is 40 mm³, and the resistive leak is a slot 20μ high, 2 mm long and approximately 200μ wide. The equivalent volume of a 1/4" microphone 4135 is 0.6 mm³, and the resistance of the leak thus 66 times higher, and it is correspondingly more difficult to ensure the mechanical stability.

Long term variation in the relative phase between two 1/4" microphones is in general less than 0.5° at 20 Hz and represents therefore no specific problem although 1/2" microphones are recommended when good accuracy and performance at low frequencies (below 200 Hz) is very important. A typical calibration set-up and the resulting calibration chart are shown in fig. 6 and 7.

The free field phase response is of great importance at higher frequen-
Fig. 6: Set-up for measurement of phase match

Fig. 7: Typical phase calibration chart for matched pair of condenser microphone cartridge

...cies especially where diffraction takes place. It is very important that the two microphones are identical and rotationally symmetrical as seen from the impinging sound waves. The relative phase between the two microphones should vary as little as possible. Compensation for phase differences in the digital programming of analyzers requires the use of correct relative phase information. In general, matched microphones used in the proper configuration do not require or benefit from compensation at discrete frequencies.
IMAGERIE DES BRUITS DE VEHICULES ET ECOULEMENT

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Introduction

L'étude entreprise depuis quelques années a montré que l'imagerie acoustique était possible pour des véhicules ou des profils aérodynamiques placés dans un écoulement (H. Arbey, A. Hellion, B. Escudié, 1979). Le présent travail cherche à décrire les effets perturbateurs dus aux écoulements moyens (couches limites, couches de cisaillement décollées, zones de mélange ...) lors de l'imagerie acoustique d'un véhicule placé à des distances relativement faibles. A cet effet différents obstacles, équipés de hauts-parleurs émettant un son pur, sont placés dans un écoulement et la localisation de la source sonore est effectuée à l'aide de la technique d'imagerie acoustique.

Conditions expérimentales des essais entrepris

Les essais sont effectués dans une soufflerie à veine ouverte de l'Ecole Centrale de Lyon (fig. 1). Différents obstacles bidimensionnels sont placés dans le cône à potentiel d'un jet, issu d'une buse carrée de 0,3 m de côté, à 0,3 m du plan de sortie de la buse (P. Dufourcq et al., 1982). Les obstacles utilisés, représentés sur la figure 2, fournissent des configurations d'écoulement très différentes : l'obstacle profilé (1), semblable à ceux utilisés lors des essais antérieurs (Arbey et al., 1979), permet d'obtenir des couches limites adhérentes aux parois, alors que l'obstacle non profilé (4) présente des couches limites décollées, comme c'est le cas sur certaines parties d'un véhicule réel. Enfin, les obstacles (2) et (3) assurent des conditions d'écoulement intermédiaires entre les deux précédents cas. Ces obstacles sont équipés de hauts-parleurs, implantés au milieu d'une face latérale, émettant un son pur à la fréquence de 2 kHz. Le choix de cette fréquence a pour but d'éviter la dispersion du son par la turbulence du jet.
Localisation de la source par imagerie acoustique

La localisation de la source est entreprise à l'aide d'une antenne à réseau gracieux de 13 pas, placée à 4 m de l'obstacle et orientée de telle façon que son axe ($\alpha = 0$), coïncide avec le centre du haut-parleur (fig. 1). L'angle de gisement $\alpha$ désigne l'angle (osS'), où S' est la source réelle, et O le point d'observation centre de l'antenne, et S' la source apparente. Un exemple d'imagerie angulaire du champ sonore, obtenue pour l'obstacle profilé (1) et une vitesse d'écoulement de 40 m/s, est donné sur la figure 3.

Résultats obtenus

Les résultats obtenus pour les différents obstacles et une vitesse d'écoulement variant de 0 à 40 m/s, sont portés sur la figure 4, où l'on remarque, lorsque la vitesse de l'écoulement augmente, une migration de la source apparente vers l'aval du jet. Cette migration est moins importante pour les obstacles non profilés (3) et (4), que pour les obstacles profilés (1) et (2). Et, dans le cas de l'obstacle profilé (1), le déplacement de la source apparente est supérieur à 1° dès que la vitesse de l'écoulement dépasse 20 m/s. Puisque ces essais ont lieu dans une soufflerie à veine ouverte, il est nécessaire d'évaluer l'influence de la frontière du jet sur la localisation de la source.
Réfraction des ondes sonores par la frontière du jet

L'utilisation du modèle d'Amiet (Dufourcq et al., 1982) permet de prédire l'influence de la frontière du jet sur la réfraction du champ sonore. Pour nos conditions expérimentales, ce modèle prédit, lorsque la vitesse de l'écoulement croît, une migration de la source apparente vers l'aval du jet. L'ordre de grandeur du déplacement est comparable à la mesure dans le cas de l'obstacle non profilé (4) (fig. 4). Par contre, pour l'obstacle profilé (1), un écart très important, entre la mesure et la prédiction, est observé dès que le nombre de Reynolds de l'écoulement dépasse 10^5. On est donc conduit à relier ce phénomène à la transition de la couche limite se développant sur les faces latérales de l'obstacle. Cette hypothèse est confirmée par des essais préliminaires au cours desquels la source est placée sur la paroi d'un jet pariétal (Dufourcq et al., 1983).

Fig. 3 - Imagerie angulaire (obstacle profilé (1), \( U = 30 \) m/s).

Fig. 4 - Evolution, avec la vitesse de l'écoulement, de la déviation angulaire.

Conclusion

Pour étudier les effets perturbateurs dus à l'écoulement moyen lors de l'imagerie acoustique de véhicules, différents obstacles équipés de hauts-parleurs émettant un son pur ont été placés dans un jet. La localisation de la source sonore est entreprise à l'aide de la technique d'imagerie acoustique, pour les différents obstacles et une vitesse d'écoulement variant de 0 à 40 m/s. Les résultats obtenus montrent que le déplacement
de la source apparente, observé dans le cas d'un obstacle non profilé, est correctement prédit par une méthode de calcul de "réfraction à la frontière de l'écoulement". Au contraire, dans le cas d'un obstacle profilé, l'expérience montre que la migration apparente est forte. On peut la relier à la transition de la couche limite se développant sur les parois de l'obstacle.

Références

APPLICATION OF SOUND INTENSITY MEASURING TECHNIQUE TO SOUND INSULATION MEASUREMENT

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1. INTRODUCTION
Recently, the sound (acoustic) intensity measuring technique based on the measurements of sound pressures at two closely spaced points have become popular, and it is being applied to various kinds of acoustic measurements such as the measurement of the sound power emitted by machines.

In this paper, some experimental results of applications of the sound intensity technique to sound insulation measurements in architectural acoustics are presented.

2. MEASURING SYSTEM
Fig.1 shows the sound intensity measuring system used in this study. In this system, the sound intensity is calculated from the sound pressures measured at two closely spaced points, based on the cross-spectrum method proposed by Fahy and Chung. As well as the sound intensity, the averaged sound pressure level of the two measuring points is also obtained in this measuring system.

As shown in Fig.2, two closely spaced microphones (1/2 inch in diameter) were used in the face to face configuration, and the gap between them was set 12mm in this study for the measuring frequency range from 250Hz to 8kHz.

3. APPLICATIONS TO SOUND INSULATION MEASUREMENT
(i) Transmission Loss Measurement
As the first experiment, the transmission loss measurement by the sound intensity technique was tried. This examination was carried out by scale model experiment.

The total sound power \( W \) transmitted through a material from a reverberant room of diffused sound field is expressed as,

\[
W = \frac{p^2}{4\rho c} \cdot S_m \cdot \eta
\]

Fig.1 Sound intensity measuring system

Fig.2 Microphone configuration

(1)
where \( p \) is the mean sound pressure in the reverberant room, \( S_m \) is the area of the material and \( \tau \) is the transmission coefficient of the material. The total sound power (W) can be obtained as the sum of the products of sound intensity (\( I_i \)) and the area (\( S_i \)) in each divided section, as follows:

\[
W = \sum_{i} I_i \cdot S_i
\]

(2)

Consequently, the transmission loss (TL) is expressed as,

\[
TL = 10 \log \frac{1}{\tau} = 10 \log \frac{p^2 \cdot S_m}{4 \rho c \sum I_i \cdot S_i}
\]

(3)

Fig. 3 shows the experimental arrangement using a scale model reverberant chamber of 0.2 m\(^3\) air volume. As the specimen, aluminum panel of 0.3 mm thick was fixed to the opening (30 cm \( \times \) 45 cm) in the wall of the reverberant chamber. The measuring surface was chosen 3 cm away from the surface of the specimen. On the center point of each divided area, both of the sound intensity levels and the sound pressure levels were measured at the same time. The sound pressure levels in the reverberant chamber were measured at five points, and they were averaged. Fig. 4 shows the comparison between the transmission loss values calculated by equation (3) from the measured sound intensity and those measured by the ordinary method using two reverberant chambers. As a result, it can be seen that these two kinds of values are in fairly good agreement.

In Fig. 4, the broken line represents the transmission loss values calculated by equation (4) from the measured sound pressure, under the assumption that the sound intensity could be related to the sound pressure near the specimen as,

\[
I = \frac{p^2(E)}{\rho c}
\]

(4)

(This relation holds true in the far field.) As a result, these values are 1 to 3 dB lower than the values measured by the two methods mentioned above.
(2) Sound Insulation between Two Rooms

As the second experiment, the sound power flow through a partition wall between adjacent two rooms was measured by the sound intensity technique. Fig.5 shows the arrangement of this experiment, using a 1/2 scale model. The sound source room was made reflective and the receiving room was made in two conditions of reflective and absorptive. Aluminum panel of 0.3mm thick was fixed to the opening of 30cm x 45cm in the partition wall made of gypsum boards of 12mm thick. The measuring surface was set 5cm away from the surface of the partition wall, and it was divided as shown in Fig.6. On the center point of each divided area, both of sound intensity levels and the sound pressure levels were measured.

As an example of the measured results, Fig.6 shows the values of sound intensity levels and sound pressure levels of 1kHz in 1/3 octave band measured under the condition that the receiving room was reflective.

Provided that the sound field in the receiving room is perfectly diffused, the relation between the total sound power (W) transmitted into the receiving room from the source room and the sound energy density (E) in the receiving room is expressed as,

$$ W = \frac{C}{4} \frac{E}{A} \quad (5) $$

where A is the total sound absorbing power of the receiving room. In this equation, the total power flow (W) can be known by measuring the sound intensity near the partition wall, and the absorption power (A) can be obtained by measuring the reverberation time in the receiving room.

![Fig.5 Sound insulation measurement using a 1/2 scale model](image)

![Fig.6 Measuring points (5 cm apart from the wall) and measured sound intensity levels (upper) and sound pressure levels (lower) of 1kHz in 1/3 octave band](image)
Based on the relation of $\bar{x} = \frac{p}{c}$, the sound pressure ($p$) in the receiving room is expressed as follows.

$$p = \left( \frac{4 \rho c}{A} \cdot \sqrt{W} \right)^\alpha = \left( \frac{4 \rho c}{A} \cdot \frac{1}{\Sigma \frac{t_i}{S_i}} \right)^\alpha \quad (6)$$

Fig.7 shows the comparison between the sound pressure levels calculated by equation (6) using the measured values of sound intensity and the averaged values of the sound pressure levels practically measured at six points in the receiving room. In this result, it can be seen that these two kinds of values are in fairly good agreement in both cases that the receiving room is reflective and absorptive, except for low frequencies.

The discrepancy in low frequencies might be attributed to the invalidity of equation (6) that is valid when the sound field is perfectly diffused.

Assuming that the sound intensity would be expressed as $p/c$, even in the field near the partition wall, the sound power flow was calculated from the sound pressures measured in the measuring surface mentioned above, and the mean sound pressure level in the receiving room was calculated according to equation (6).

As a result, the differences of 8 to 11 dB between the calculated values and measured values are observed.

4. CONCLUSIONS

In this study, sound insulation experiments were carried out by using the sound intensity measuring technique. As a result, it was found that this technique is markedly effective for this kind of acoustic measurement, by measuring directly the sound power flow through the partition walls.

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ABOUT THE ALTERNATIVE MEANING OF THE INTENSITY

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All the present theories of continuous medium describe the energy flow by means of a vector, the so called vector density of energy flow or intensity vector. This vector consists of two parts according to the two basic physical phenomena i.e. convection and conduction. The convective part of intensity characterizes the direct energy transport when moving particles carry the energy from one place to another. The direction of the energy flow is the direction of the particle velocity. The other part of the intensity vector characterizes the conductive flow of energy. But the conduction is extremely different from the convective flow. The energy spreads to all directions and therefore it can’t be described by a vector.

There are some special cases when one can assume that the energy flows in a given direction, but the careful examination would find some initial condition which distinguishes the mentioned direction. E.g. a plane source with linear dimensions much larger than the wavelength will distinguish the direction normal to the plane. But a little disturbance always spreads in every direction.

The phenomenon of spread can be characterized by a tensor instead of a vector. Our aim is to find the correct form of this tensor.

In a former communication we supposed that the energy spread can be described by a third-rank tensor \[ [1] \]. It is well-known from the tensor calculus that the flux of a vector field generated by the second-rank tensor \( A_{ij} \) is the first scalar invariant of the tensor \( \Phi / A = A_{ij} \).

The balance equation of energy in a medium without any source and loss is described by

\[
\int \partial_i w \, dV + \int \overline{I} \, dS = 0
\]

where \( w \) is the energy density and \( \overline{I} \) the intensity vector. The integration is completed over the closed volume \( V \) bounded by the surface \( S \). If the surface element is small enough, the intensity which depends only on the derivatives of the displacement can be regarded as constant over the surface element. But it will depend on the orientation of the surface. Therefore we assume that the normal component of the intensity vector is rather a flux of a second-rank tensor:

\[
\overline{I} \mathbf{n} = \frac{1}{2} \overline{J}_{ij} \mathbf{n} (\mathbf{r})
\]
because only the half flux will propagate outwards through the surface. But in this case the intensity vector can be replaced by a third-rank tensor:

\[ I_k = \frac{1}{2} \mathbf{j}_{iik} \]

from which

\[ \mathbf{i}_n = \frac{1}{2} \mathbf{j}_{iik} n_k \]

Generally this tensor depends on the orientation of the surface element. The simplest assumption is that the tensor of the energy flow is a fourth-rank tensor and hence

\[ \mathbf{j}_{ijk} = \mathbf{j}_{iern} n_e \]

\[ \mathbf{j}_{ij} (\mathbf{n}) = \mathbf{j}_{ij} (\mathbf{n}) n_k n_k \]

The \( \mathbf{j}_{ij} \mathbf{n} / \mathbf{n} \) second-rank tensor describes the conductive energy flow on a surface element with a given orientation. The directions and magnitude of flow can be represented by the tensor ellipsoid. The absolute value of energy flow in a direction \( \mathbf{e} \) is as follows:

\[ j_e = \mathbf{j}_{ij} (\mathbf{n}) \mathbf{e}_i \mathbf{e}_j \]

The energy flow is normal to the surface, therefore

\[ j_n = \mathbf{j}_{ij} (\mathbf{n}) n_i n_j \]

will generally differ from the total outward flow, described by \( \mathbf{i}_n \).

Let us see the form of the energy flow tensor in isotropic media. In this case the energy flow doesn't depend on the orientation of the surface, that is the tensor \( \mathbf{j}_{ijkn} \mathbf{n}_k \mathbf{n}_i \) is independent of \( \mathbf{n} \). The simplest assumption is that

\[ \mathbf{j}_{ij11} = \mathbf{j}_{ij22} = \mathbf{j}_{ij33} \]

and the other components are zero.

We suppose to get the conductive energy flow tensor as the product of a sound velocity tensor and the energy density. The simplest form of such a tensor is

\[ \mathbf{j}_{ij} = w C_{ij} + w C^x \]

where

\[ w = \frac{1}{2} \mathbf{u}_i u_i \quad \text{and} \quad w = \mathbf{u}_i u_i - \mathbf{u} \]

is the energy density tensor.

By substituting /10/ to /9/ we get:

\[ \mathbf{j}_{ij} = w (C_{ij} + C^x) + \mathbf{u}_i u_i C^x \]

The second-rank tensor belonging to the direction \( \mathbf{n} \) will be

\[ \mathbf{j}_{ij} (\mathbf{n}) = w (C_{ij} n_k n_i - C^x n_k n_i) + \mathbf{u}_i u_i C^x n_k n_k \]

In isotropic medium \( \mathbf{C}_{i11} = \mathbf{C}_{i22} = \mathbf{C}_{i33} = \mathbf{C} \) and \( C^x = C^x = C^x = C \), so

\[ \mathbf{j}_{ij} (\mathbf{n}) = \mathbf{d}_{ij} (\mathbf{C} - \mathbf{C}^x) \mathbf{w} + \mathbf{u}_i u_i \mathbf{C}^x \]

The energy-flow tensor of /13/ is independent of the direction. The value of the energy-flow in an arbitrary direction \( \mathbf{e} \) will be:

\[ j_e = \mathbf{j}_{ij} \mathbf{e}_i \mathbf{e}_j = (C - C^x) \mathbf{w} + \mathbf{u} (u \mathbf{e})^2 C^x \]
Separate the velocity vector \( \vec{u} \) in two parts, one parallel and one perpendicular to \( \vec{e} \) ! Denote the parallel part with \( u_{\text{long}} \), the perpendicular part with \( u_{tr} \). Then

\[
\dot{\vec{u}} = (C - C^x) \frac{4}{3} \varrho \left( u_{\text{long}}^2 + u_{tr}^2 \right) + C^x \varrho \ u_{\text{long}}^2 /15/
\]

Hence

\[
\dot{\vec{u}} = (C + C^x) \frac{4}{3} \varrho \ u_{\text{long}}^2 + (C - C^x) \frac{4}{3} \varrho \ u_{tr}^2 /16/
\]

The sound velocity of the longitudinal waves is \( C + C^x \), the velocity of transversal waves is \( C - C^x \).

In the case of fluids and gases \( C^x = 0 \), and

\[
\bar{\kappa} = C \cdot \varrho /17/
\]

from which the value of energy flow in every direction will be

\[
\dot{j} = C \cdot \varrho /18/
\]

We can suppose that the \( w \)-dependent part in /9/ will be independent of the direction in anisotropic medium, too. Therefore let be

\[
C_{ijkl} = C_{ijkl}^x /19/
\]

and by substituting /19/ to /12/ we get:

\[
\bar{\kappa}_{ijkl} = \varrho \left( C_{ijkl}^x - C_{ijkl}^x \right) + g u_{i} u_{j} C_{ijkl}^x /20/
\]

from which

\[
\bar{\kappa}_{ijkl} (\vec{n}) = \varrho \left[ C_{ijkl}^x - C_{ijkl}^x \right] + g u_{i} u_{j} C_{ijkl}^x (\vec{n}) /21/
\]

where

\[
C^x (\vec{n}) = C_{kl}^x n_k n_l /22/
\]

The first part of /19/ is independent of the local orientation and it can be represented everywhere in the medium by the same tensor-ellipsoid. The second part makes contribution only in the direction of particle velocity. Therefore this second part describes only the longitudinal propagation, while the first can give account of the spread. The maximum number of the independent sound velocity components is 2x9=18.

According to our opinion this model gives a more realistic description of the propagation of energy in arbitrary deformable media that the intensity vector. The introduced assumptions are in good agreement with the well-known facts of sound propagation. The components of the two sound velocity tensors can be easily determined by sound velocity measurements.

There are some interesting conclusions of this model. Two different physical processes seem to determine the energy flow in a medium, the spread and the linear propagation parallel to the particle velocity. The latter occurs only in solids. The spread always contains "longitudinal" and "transversal" processes, but it is better to say "compressional" and "shear" processes. In fluids and gases there are neither shear, nor linear waves. The only remaining component is the compressional wave.

The intensity vector of the new model can be calculated by substituting /11/ and /19/ to /3/:

\[
\bar{\kappa}_{ijkl} (\vec{n}) = \frac{1}{2} \varrho \left( C_{ijkl}^x n_k n_l + \frac{4}{3} \varrho \right) /22/
\]
Miklós András - About the alternative meaning of the intensity

from which in isotropic case we get
\[ I_k = \frac{3}{2} C w n_k + \frac{1}{2} C' w n_k \]
and the normal component of the intensity is as follows:
\[ \overline{I} n = w \frac{\frac{3}{2} C + C'}{2} \]
In general case
\[ \overline{I} \hat{n} = w \frac{C + C' \hat{n}}{2} \]
In the case of fluids and gases
\[ \overline{I} \hat{n} = \frac{3}{2} C w \]
The total energy flow through the surface is one and a half times greater than the normal energy flow. We must take a strict distinction between the density of energy flow in a given direction, and the total amount of energy passing through the surface. According to /1/ and /2/ the power radiated by the source is in a close connection with the latter one. Therefore the amount of total energy flow must be regarded as intensity. The defined intensity is always a scalar quantity. The intensity vector in /22/ does not give correct information about the energy propagation, therefore its application is not recommended. On the other hand, the density of energy flow from a surface is always a vector, showing in the direction of propagation. The radiation and absorption across a surface can be characterized by the intensity, while the propagation of energy can be characterized by the vector density of energy flow, i.e.
\[ \overline{I} (\hat{e}, \hat{n}) = \sum_{ijkl} e_j n_k n_l \]
The introduced energy flow tensor has another remarkable property, too. It can be seen from /14/, that the energy flow in a given direction is always positive. It has zero value only in the case when the sound totally diminishes. It is a very important thing. We can characterize the sources with positive and the absorbers with negative energy flow. In the first case the components of the sound velocity tensors are positive, and in the second case they are negative.

In a complex sound field the different intensities from different sources must be added on the considered surface. The sign of the intensities will depend on the position of the source compared to the surface. If the angle between the surface normal and the position vector of the source is greater than 90°, the sign is positive, and in other cases it is negative. Therefore the resultant intensity of more sources can be zero.

Reference


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L'ECHANTILLONNAGE DES CHAMPS D'INTENSITE ACOUSTIQUE

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SUMMARY

This paper deals with the sampling of acoustical fields where sound power is determined using intensity measurements. The theoretical considerations that have given rise to the hypothesis of an optimal distance between the sound source and the measurement surface are recalled. The hypothesis is verified by means of a computer simulation of the acoustic field around a dipole with and without background noise perturbation. The variance on the sound power is plotted versus distance from the source and shows the expected minimum.

INTRODUCTION

La mesure d'intensité acoustique permet la détermination de la puissance acoustique rayonnée par une machine située dans un environnement industriel même si le bruit de fond domine le bruit propre de la machine. Dans ce dernier cas une hypothèse formulée précédemment [1] est qu'il existerait une distance optimale entre la machine et la surface où sont échantillonnées les valeurs de l'intensité normale. Cette hypothèse et les considérations théoriques sur lesquelles elle est basée font l'objet d'une première partie. Une deuxième partie concerne la vérification de cette hypothèse dans le cas d'une simulation sur ordinateur de la mesure d'un dipôle dont le champ est perturbé par du bruit de fond.

JUSTIFICATIONS THEORIQUES DE L'HYPOTHESE SUR L'EXISTENCE D'UNE DISTANCE OPTIMALE DE MESURE

La précision sur la valeur de la puissance acoustique W peut être caractérisée par sa variance. L'intensité I étant considérée comme la somme de l'intensité IS due à la source mesurée et de l'intensité IB due au bruit de fond, il est possible de décomposer la variance sur la puissance acoustique en fonction des variances de ces deux intensités:

$$\frac{\text{var}(W)}{W^2} = \frac{1}{N} \left( \frac{\text{var}(IS)}{IS^2} + \frac{WB}{W} \frac{\text{var}(IB)}{IB^2} \right)$$

où $WB = \int |IB| \, ds$
Champ direct de la source (terme α). Dans tout le champ créé par un ensemble de sources non cohérentes ou dans le champ lointain créé par des sources cohérentes, le rotationnel de l'intensité active est nul [2]. Par contre, dans le champ proche créé par un ensemble de sources cohérentes ce rotationnel n'est pas nul, les lignes de champ auront donc tendance à boucler et la distribution spatiale du vecteur intensité active sera perturbée. La variance sur l'intensité sera d'autant plus grande que l'on sera proche de l'ensemble de sources cohérentes.

Champ perturbateur (terme β) Il intervient par un produit de deux termes : variance et rapport bruit/signal (\(\frac{WB}{W}\)). Du fait que l'on restera généralement dans le champ lointain d'autres sources, la variance changera peu avec la distance de mesure. Par contre le rapport bruit/signal augmentera proportionnellement au carré de la distance de mesure (pour un bruit de fond uniforme). La somme des termes α et β est fonction de la distance et doit présenter un minimum [1]. Le terme 1/N intervient indépendamment de la distance ; le nombre de points de mesure peut être augmenté pour améliorer la précision.

SIMULATION D'UNE MESURE

Un cas simple d'ensemble de sources cohérentes a été choisi, il s'agit d'un dipôle constitué de deux monopoles en opposition de phase à la distance d. La valeur de kd choisie est de 10E-5. La surface de mesure est une sphère centrée sur le point milieu du dipôle.

Le champ direct de la source dipolaire présente bien une variance qui décroît avec la distance (figure 1). Il y a décroissance rapide dans la zone du champ proche où l'intensité est très perturbée comme l'indique le tracé de l'intensité normale à la sphère de rayon 2d (figure 2). La zone du champ lointain est caractérisée par la directivité en cosθ de l'intensité rayonnée par le dipôle comme le montre le tracé sur la sphère de rayon 1000d (figure 3).

Le champ perturbateur choisi est celui d'un dipôle situé à grande distance (300d) pour uniformisé le champ. Il est considéré comme non cohérent avec le dipôle mesuré, c'est à dire que les intensités s'ajoutent. Sa variance est assez constante quel que soit la distance de mesure.

Lorsque les deux champs sont superposés, la variance sur W présente le minimum attendu (figure 4) : dans le champ proche la source mesurée domine le bruit de fond et la variance présente la même décroissance qu'en l'absence de bruit de fond, la remontée de la variance aux grandes distances est due à l'augmentation du rapport bruit de fond/signal avec l'éloignement de la source mesurée.

CONCLUSION

L'hypothèse de l'existence d'une distance optimale est basée sur la nature irrotationnelle du champ d'intensité active proche des sources étendue et l'augmentation du rapport bruit/signal avec l'éloignement. Cette hypothèse a été vérifiée dans le cas d'une source dipolaire par simulation sur ordinateur.
REFERENCES BIBLIOGRAPHIQUES


Figure 1 - Champ du dipôle seul - Précision de la détermination de W en fonction de la distance

Figure 2 - Intensité dans le champ proche du dipôle ( sphère de rayon 2d)
Figure 3 - Intensité dans le champ lointain du dipôle (sphère de rayon 1000d)

Figure 4 - Champ du dipôle + champ perturbateur - Précision de la détermination de W en fonction de la distance.
MÉTHODES DE TRAITEMENT DU SIGNAL APPLIQUÉES À LA MESURE DE L'ISOLEMENT ACoustique PAR EXCITATION IMPulsionnelle (13)

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I - Introduction

La détermination de l'isolement acoustique des locaux d'habitation se ramène au point de vue du traitement du signal à la mesure de la fonction de transfert du quadripôle équivalent suivant :

\[ \begin{align*}
&\text{p}_1 : \text{pression acoustique local émission} \\
&\text{p}_2 : \text{pression acoustique local réception} \\
&\text{R}_2 : \text{absorption acoustique local réception}
\end{align*} \]

La mesure de cette fonction de transfert peut être envisagée de deux façons :
- en régime permanent, le signal d'excitation est aléatoire (généralement bruit blanc ou bruit rose)
- en régime transitoire, le signal d'excitation est alors une impulsion.

La mesure en régime permanent largement utilisée, reste inadaptée dans certains cas particuliers où l'on a besoin d'une source portable ou d'un niveau de pression acoustique important devant un bruit de fond élevé. La nature des deux signaux nous conduit dans le premier cas à une réponse aléatoire du système qui ne peut être caractérisée que par une moyenne d'ensemble alors que la réponse impulsionnelle donne par définition une caractéristique déterministe du même système (21).

Dans un cas comme dans l'autre, on peut s'affranchir de la variance de la pression quadratique moyenne en faisant l'hypothèse du champ diffus, hypothèse vérifiée dès lors que les fréquences étudiées sont supérieures à la fréquence de Schroeder (3) (43).

II - Rappels des relations utilisées

On cherche à connaître le gain du quadripôle pour une valeur normalisée de la charge \( R_2 \).
- à partir de \( p_1 \) et \( p_2 \) on peut connaître le gain de la fonction de transfert du quadripôle chargé par \( R_2 \); c'est l'isolement brut.
- en injectant un flux de vitesse connu à travers \( R_2 \) et en mesurant \( p_2 \) on déduit la valeur \( R_2 \); c'est la détermination de l'aire d'absorption acoustique du local réception.
2.1. Régime permanent

2.1.1. Transmission à travers la paroi

On définit le facteur de transmission $\tau(\omega)$ de la paroi par :

1. $\tau(\omega) = \frac{W_2(\omega)}{W_1(\omega)}$

où $W_1(\omega)$ et $W_2(\omega)$ sont respectivement les densités spectrales de puissance incidentes à la paroi et transmises par celle-ci.

On peut relier ces grandeurs aux densités spectrales de puissance de $p_1$ et $p_2$ (pressions acoustiques captées par les microphones émission et réception) :

2. $W_1(\omega) = \frac{|P_1(\omega)|^2 S}{4 \pi c}$

3. $W_2(\omega) = \frac{|P_2(\omega)|^2 R_2}{4 \pi c}$

où $S$ : surface de la paroi de séparation

$R_2$ : aire d'absorption du local de réception.

2.1.2. Absorption du local réception

La relation 3 permet, en utilisant une source de puissance acoustique connue, de déduire $R_2$ par simple mesure de $P_2$.

2.2 Régime transitoire

2.2.1. Transmission à travers la paroi

Le signal étant à énergie finie, il suffit dans les relations 1, 2, 3 de remplacer $W_1(\omega), W_2(\omega), |P_1(\omega)|^2, |P_2(\omega)|^2$ par $E_x(\omega), E_x(\omega), |P_1(\omega)|^2, |P_2(\omega)|^2$ qui représentent des densités spectrales d'énergie.

2.2.2. Absorption du local de réception

La réponse impulsionnelle du local de réception permet d'éviter l'emploi d'une source étalon. En effet, compte tenu de la brièveté de l'impulsion, l'onde directe et les ondes réfléchies qui présentent un retard dû à leurs réflexions, seront séparées dans le temps. Il ne nous reste plus qu'à obtenir une relation qui nous permet d'obtenir l'aide d'absorption acoustique du local de réception.

Champ direct

Dans le cas d'une source ponctuelle, omnidirectionnelle, située à une distance $d$ du microphone, la propagation sphérique nous permet de décrire :

4. $|P_d(\omega)|^2/4 \pi c = E(\omega)/4 \pi d^2$

où $|P_d(\omega)|^2$ et $E(\omega)$ sont respectivement les densités spectrales d'énergie de la pression de l'onde directe et de la source.

Champ réverbéré

La relation 3 s'écrit ici :

5. $|P_r(\omega)|^2/4 \pi c = 4 \pi E(\omega)/R_2(\omega)$

où $|P_r(\omega)|$ est la densité spectrale d'énergie du champ réverbéré.

Les deux expressions 4 et 5 nous donnent la relation cherchée :
III - Vérification expérimentale

Les essais de transmission ont été effectués entre deux locaux de volume 114 m3. La surface de la paroi testée était égale à 19,5 m2. La source d'excitation était un pistolet à amorces donnant un niveau de pression crête égal à 137 dB, supposée ponctuelle et omnidirectionnelle. La source et le microphone étaient situés sur une diagonale de la pièce à un tiers de celle-ci et à deux mètres au dessus du sol. La distance de la source au microphone était égale à 3 mètres ; la première onde réfléchie présentait un retard de 6,8 ms sur l’onde directe. Les signaux de sortie étaient échantillonnés à une fréquence de 15625 Hz, puis stockés sur cassette numérique à des fins de traitement.

3.1. Traitement des signaux par transformée rapide de FOURIER

Le traitement numérique était effectué à l'aide d'un calculateur de bureau HP 9825 et d'un analyseur de signal HP 5420 effectuant le calcul de la transformée de FOURIER sur 512 points. La capacité mémoire du calculateur de 22 000 octets limitait le traitement à 10 000 échantillons pour chaque voie. Pour chaque voie, le calculateur échantillonne la source à la fréquence de 640 ms. Les échantillons ont donc été rangés dans 40 fichiers de 512 valeurs, puis aiguillés vers l’analyseur.

La valeur cherchée étant l'énergie du signal par bande d'octave, nous avons calculé la somme des densités spectrales d'énergie des spectres consécutifs et nous avons regroupés les raies pondérées par la fonction de transfert du filtre d'octave normalisé.

Les résultats obtenus sont présentés sur le tableau récapitulatif.

3.2. Traitement des signaux par filtrage numérique

Comme précédemment le traitement portait sur un enregistrement temporel de 640 ms pour chaque voie. Nous avons synthétisé les filtres numériques à partir du filtre analogique passé bande de TCHEBYSCHEFF dont les principales caractéristiques sont les suivantes :
- filtre du 6ème ordre
- ondulation de 0,5 dB sur la largeur de bande
- largeur de bande relative de 60 % pour les octaves centrés sur 125, 250, 500, 1000 Hz, de 80 % pour l'octave centré sur 2000 Hz et de 100 % pour l'octave centré sur 4000 Hz.

Chaque filtre a été simulé sur table traçante pour vérifier le respect des conditions imposées par le gabarit normalisé. Après avoir mis la fonction de transfert du filtre numérique sous forme canonique, nous déduisons une relation de récurrence qui permet de traiter en un seul passage les 10 000 échantillons correspondant à chaque voie. L'énergie par octave est accumulée dans une mémoire tampon après chaque passage d'échantillon.

Les résultats correspondants à ce traitement sont regroupés avec les précédents et apparaissent dans le tableau récapitulatif.

3.3. Résultats

Pour la mesure en transmission, les signaux traités étaient le signal
Mesure de l'isolement acoustique par excitation impulsionnelle

pression à l'émission et le signal pression à la réception.
Dans le cas de la mesure d'absorption les signaux utilisés étaient l'onde directe et les ondes réfléchies.

<table>
<thead>
<tr>
<th>Fréquences Hz</th>
<th>125</th>
<th>250</th>
<th>500</th>
<th>1000</th>
<th>2000</th>
<th>4000</th>
</tr>
</thead>
<tbody>
<tr>
<td>Régime permanent</td>
<td>24,1</td>
<td>28,7</td>
<td>35</td>
<td>38,8</td>
<td>43,7</td>
<td>47,3</td>
</tr>
<tr>
<td>Régime transitoire F.F.T.</td>
<td>24,2</td>
<td>27,33</td>
<td>35,6</td>
<td>38,4</td>
<td>42,3</td>
<td>44,9</td>
</tr>
<tr>
<td>Régime transitoire Filtrage numérique</td>
<td>25,4</td>
<td>31,35</td>
<td>37,9</td>
<td>41,9</td>
<td>45</td>
<td>45,7</td>
</tr>
</tbody>
</table>

ISOLEMENT BRUT dB

ABSORPTION m² SABINE

**TABLEAU RECAPITULATIF**

**CONCLUSIONS**

La méthode de mesure des caractéristiques acoustiques des locaux d'habitation en régime transitoire pose des problèmes métrologiques liés à la nature même de l'excitation.
En effet, les dispositifs de mesure doivent être équipés d'un facteur de crête important rarement suffisant sur les appareils de type analogique.
Cependant, le traitement numérique du signal permet de suppléer cette carence technologique.
Les résultats expérimentaux obtenus par la mesure en transmission sont en bon accord avec ceux donnés par la méthode traditionnelle.
Les résultats concernant la mesure d'absorption présentent une dispersion importante que nous attribuons à la directivité engendrée par le canon du pistolet à amorces utilisé comme source impulsion.
Il convient donc de compléter la relation (6) par un coefficient qui tiendra compte de la directivité de la source pour chaque octave.
Cette technique semble prometteuse, compte tenu de l'arrivée récente sur le marché de microprocesseurs spécialement adaptés au traitement numérique du signal.

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METHODE NUMERIQUE DE TRAITEMENT DU SIGNAL ACOUSTIQUE UTILISABLE EN INTEN-
SIMETRIE

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Introduction

La méthode de mesure de l'intensité Acoustique utilisant deux micro-
phones, est très influencée par les déphasages parasites sur les deux voies.
Seul le traitement numérique permet de s'affranchir des déphasages dus à
l'instrument de mesure.

Dans notre étude, deux méthodes numériques de traitement du signal
sont envisagées :

- Analyse par FFT en bicanal, où l'intensité acoustique est calculée
  par l'interspectre de puissance.

- Filtrage numérique par bande d'octave.

Ces deux techniques seront décrites, les avantages et inconvénients
des deux méthodes seront étudiés en vue de la réalisation d'un instrument
effectuant en temps réel la mesure de l'intensité acoustique.

Un exemple de réalisation d'un appareil, à l'aide d'un processeur
spécifique au traitement du signal sera présenté.

Nous concluons notre étude par des résultats permettant de chiffrer
la précision de la méthode.

I - Rappels et définitions

L'intensité acoustique suivant une direction données et en un point
donné de l'espace, est l'énergie sonore qui traverse pendant l'unité de
temp une surface unité, qui serait normale à cette direction. Elle se
exprime en W/m². L'intensité en un point s'écrit :

$I = p \cdot \bar{v}$ (p : pression en ce point ; $\bar{v}$ : vitesse instantanée).

La valeur moyenne de I est donnée par : $\overline{I} = \frac{1}{T} \int_{P} \overline{\mathbf{v}} \, dt ; (1)$

En utilisant la méthode des deux microphones placés côte à côte (voir
fig. 1), on approche le gradient par une différence finie $\overline{P_1} - \overline{P_2}$
, et la
pression sonore au point de mesure M par $\overline{P_1} \cdot \overline{P_2} \cdot \overline{R_1} \cdot \overline{R_2}$ étant des pressions
recueillies respectivement par les microphones 1 et 2.

Avec ces deux approximations, l'intensité acoustique peut s'écrire[1]:
SIDKI M. Méthode numérique de traitement du signal acoustique utilisant en intensité

\[ I = \int \frac{1}{\rho} \frac{\partial}{\partial x} \rho_2 dt \quad (\rho: 	ext{masse volumique de l'air}) \]

L'erreur la plus importante lorsqu'on estime l'intensité par l'expression ci-dessus, est celle introduite par le déphasage entre les deux voies.

La réalisation d'un intensimètre analogique [3], a montré que la plus grande source de déphasage entre les deux voies était la dispersion sur la rotation de phase entre les filtres utilisés pour l'amélioration du rapport signal sur bruit, et pour une analyse par bande d'octave ou tiers d'octave.

Toutes ces raisons conduisent à envisager un traitement numérique du signal pression.

II - Méthodes de l'interspectre

En utilisant les transformées de FOUER de \( p \) et \( u \), on a montré [2], que la densité spectrale d'intensité suivant une direction Ox s'écrit:

\[ I = \frac{1}{\rho_2} \frac{\partial}{\partial x} G_{12}, \]

où \( G_{12} \) est l'interspectre entre les deux signaux \( \rho_1 \) et \( \rho_2 \), avec la restriction \( k_d < 1 \) (k : nombre d'onde).

Nous avons vérifié la validité de la méthode à l'aide du dispositif expérimental suivant (fig. 2)

![Diagramme de dispositif expérimental](image)

Les signaux issus des deux microphones sont appliqués sur les deux canaux de l'analyseur 'HP. 5420'. Il calcule leur interspectre et le divise par \( j \omega \). Les données sont ensuite envoyées à une calculatrice 'HP. 9825 A', où s'effectue la sommation des composantes intérieures à chaque octave (ou tiers d'octave). Les résultats sont enfin renvoyés à l'analyseur pour visualiser l'intensité acoustique.

Notons que dans une analyse par F.F.T., on est limité par la durée de chaque échantillon de signal utilisé pour le calcul de la transformée de FOUER ; on sera par conséquent loin d'une analyse en temps réel. En outre, c'est une opération lourde et coûteuse.

III - Méthode par filtrage numérique

On réalise une chaîne de mesure conforme à la relation (2) suivant le schéma de la fig. 3.

Cette méthode traite respectivement sur les deux voies, les deux signaux de pression \( \rho_1 \) et \( \rho_2 \), par filtrage numérique en bande d'octave normalisée. Les deux grandeurs numériques \( x_i \) et \( y_i \) sont multipliées et stockées en mémoire, pour en calculer la valeur moyenne de l'intensité, selon un temps d'intégration prédéterminé.
ICI L'OBJECTIF DE L'EXÉCUTION EN TEMPS RÉEL, N'EST CONCEvable QUE DANS LA MESURE, OÙ L'ON PEUT SYNTHÉTISER SIMULTANÉMENT TOUS LES FILTRES D'OCTAve.

UN TEL OBJECTIF NOUS CONDUIRA CEPTAULD AU CHOIX D'UN MICROPROCESSEUR SPÉCIFIQUE AU TRAITEMENT DU SIGNAL, SUIVI D'UN AUTRE POUR EFFECTuer LE TRAITEMENT ARITHMÉTIQUE DES DONNées.

LE GRAND AVANTAGE DE CETTE MÉTHODE RÉSIDE DANS LE FAIT, QUE LA TECHNIQUE DU FILTRAGE NUMÉRIQUE PERMET DE S'AFFRANCHIR DU DÉPHASAGE RELATIF ENTRE LES DEUX VOIES DE MESURE ET DONC DE GARANTIR LA PRÉCISION SUR LA MESURE DE L'INTENSITé.

LA RÉALISATION D'UN APPAREIL UTILISANT CETTE MÉTHODE A ÊTÉ EFFECTUÉE À L'AIDE D'UN MICROPROCESSEUR SPÉCIFIQUE "INTEL 2920" [4]

III - 1. BANC D'ESSAI ET RÉSULTATS

LE SYNOPTIQUE, (FIG. 4) FAIT APPARAIR L'INSTRUMENT DE TRAITEMENT DU SIGNAL RÉALISÉ, SUIVI D'UNE CALCULATRICE "HP 9825", POUR LE TRAITEMENT DES DONNées.

Chacune des deux voies traite le signal de pression par filtrage en bande d'octave. La sortie est effectuée en numérique sur 8 bits dont 1 bit de signe.

Le filtrage numérique a été synthétisé [5] à partir de la fonction de transfert d'un filtre passe-bande de Tchebycheff [6] :
- 6ème ordre, - ondulation 0,5 dB en bande passante, - largeur de bande 100 %.

Sur chaque voie la fréquence d'échantillonnage est fixée à quatre fois la fréquence centrale des filtres d'octave centrés sur : 4khz, 2, 1, 0.5, 0.25, 0.125. L'analyse par bande d'octave est effectuée séquentiellement, grâce à une commutation programmable de l'horloge, selon un temps d'intégration spécifique pour chaque octave (125 ms pour l'octave 125 hz).

Les courbes de réponses en fréquence obtenue, sont représentées ci-dessous ainsi que la courbe de linéarité correspondante.
Conclusion

Les différents tests sur le 2920 nous ont permis de faire une évaluation sur ces capacités de traitement du signal. Nous espérons disposer dans un avenir proche d'un instrument complet permettant la mesure de l'intensité acoustique.

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HIGH-SPEED HOLOGRAPHY AND SPECKLE PHOTOGRAPHY OF SHOCK WAVE PROPAGATION
IN AIR

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1. Introduction

Small-scale ultrasound models are a convenient way to study sound propaga-
tion in natural environments. A short shock wave often serves as acoustic
probing signal. In the study of this wave, optical techniques are of advan-
tage as compared with acoustical methods. They do not disturb the
acoustic field, which is the case in microphone scanning of the field.
Furthermore, the spatial configuration of the wave front is much easier to
determine in optical imaging.

In the present case, the model environment that is studied as to its
influence on sound propagation, introduces an optically diffuse background.
Thus some classical methods of sound visualization like Schlieren or dark-
field are difficult to employ. In this investigation we present the
application of two methods suitable for the study of shock waves in the
small-scale model: speckle photography and holographic interferometry /1/.

2. Acoustic properties of the shock wave

The shock waves used in the present investigation are created by electric P
 discharge in a spark gap. After
transmission through a PVC tube, the
original signals steepen by nonlinear
effects. The typical pressure pulse
at a distance of 1 cm behind the tube
opening was recorded with a micro-
phone and is shown in Fig. 1.
Typically, the pressure rises to
about 145 dB within some 2 - 3 µsec.
Several additional smaller peaks
follow, position and amplitude
depending on the geometrical con-
figuration of spark-chamber and
tube employed.

Fig. 1: Record of sound pressure
vs time
3. Holographic interferometry of the shock wave

Optically, the raised pressure in the shock wave changes the optical index of refraction and thus imposes a phase shift upon an illuminating coherent light wave. Under present conditions the order of magnitude of this shift is only $0.05 - 2n$. To record the change in phase, the well-known setup for optical holography is used /2/.

![Holographic setup diagram](image)

**Fig. 2**: Setup for optical holography

A three-stage Q-switched ruby laser serves as light source. Its output beam of approximately 1 J per pulse is split into an object beam, that illuminates the white background screen via a beam expanding lens and a ground-glass diffuser, and a reference beam for illumination of the holographic recording plate. The shock wave is recorded in front of the white back ground. The holographic plate is exposed twice with a time separation of 20 s, one exposure is made without, the other in the presence of a shock wave.

The shock-induced phase changes $\Delta \phi$ produce interference fringes in the reconstructed holographic image, the light intensity $I$ following the relation

$$I = 4I_0 \cos^2 \frac{\Delta \phi}{2}$$

where $I_0$ is the intensity of each of the two interfering waves. Due to the cosine dependence there is very little sensitivity as to small phase changes, thus a phase bias of $\Delta \phi = \pi/2$ is introduced to shift the curves into the most sensitive region. This bias can be achieved by an appropriate micropositioning of one of the mirrors in Fig. 2. In many cases, however, when there are no special precautions against vibrations, accidental readjustments impose phase shifts of the required order of magnitude.
Fig. 3: Pictures taken from two holograms of the shock wave. The right picture shows reflection of the wave at an aluminium plate.

The power of the method is illustrated in two more elementary examples of Fig. 3. They demonstrate clearly the possibility to visualize shock fronts. For a quantitative analysis the recorded images need scanning by a microdensitometer. Instead, we have applied the following method of speckle photography.

4. Speckle photography of the shock wave

In the method of double exposure speckle photography \( /3/ \), the process under study imposes a translational shift between two superimposed speckle patterns. Magnitude and direction of the shift are determined from a Young's fringes analysis in the far-field diffraction pattern of the speckle image. In our case, the displacement of the speckle patterns originates from the refraction of light by the shock-wave produced gradient of the refractive index. The optical setup is shown in Fig. 4, where two superimposed images of the shock front region are taken.

Fig. 4: Optical setup for speckle photography

The background diffusor is displaced between exposures a small amount, and the shock wave is present during one exposure only. In the evaluation of the processed image transparency by illumination with focused laser light, the background displacement results in Young's fringes of a corresponding spacing. In positions occupied by the shock wave, the spacing has been altered by the additional displacement induced by the shock wave pressure gradient. Locations of the wave front and information about peak pressure or shock front width can be obtained from this analysis. Results agree well with data obtained from microphone measurements.

In this paper we show a more qualitative but quite illustrative technique of evaluation (Fig. 5).
Fig. 5: Setup for qualitative analysis of specklegrams

When the complete double exposure specklegram is illuminated normally and viewed at an angle, the light intensity observed at different points of the transparency depends on the speckle displacement at this position. Thus, the shock front may be seen as either a brighter or darker region in front of a medium bright background. In Fig. 6, images of a shock front are shown that were obtained in this way.

Fig. 6: Pictures of the shock wave taken by speckle photography

5. Conclusion

The examples presented in this paper illustrate the principal properties of the methods proposed for shock front studies in front of an optically diffuse background. The application considered is the study of acoustic propagation in small-scale ultrasound models of environmental conditions. Work on such a model is in progress.

References

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Fortschritte der Akustik, FASE/DAGA '82, p. 459

/2/ M. Vest: Holographic Interferometry, Wiley, New York 1979

SOURCE ENERGY ALLOCATION

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ABSTRACT

Frequency response and coherence functions can be used to analyze vibration and noise spectra and to identify their possible sources. Based on measurements near each expected source, the output noise energy can be allocated and quantified to the respective sources. The results are independent of the amplitude or transducer type at the various measurement locations and the transmission path between inputs and outputs. Coherence methods will be proven to be a simplification of a more general frequency response based, multiple input-output system approach. More details can be found in ref. 7.

1. MULTIPLE INPUT-OUTPUT SYSTEMS.

In linear systems analysis, a mechano-acoustical situation is described by various interrelationships (transfer functions $H$) between inputs ($X$) and outputs ($Y$). In addition to the inputs also unrelated sources of noise and interference contaminate the outputs.

\[ Y = \sum_{i=1}^{n} H_i x_i + N = U + N \]

where $U$ is defined as the coherent part (related to the defined inputs) of the output. These input-output relationship also can be expressed as function of the output powerspectrum:

\[ S_{yy} = \sum_{i=1}^{n} H_i^* S_{ii} \]

For annotation problems we will limit further derivations to three input-single output systems.

2. COHERENCES

The coherence function states a causal relationship between two signals when the phase difference between identical spectral lines remains constant over several measurement averages. The measurement result is independent of the power levels at two points and the transmission gain between them.
The ordinary coherence gives the proportion of the total power that is related to both signals:

\begin{equation}
\gamma_{yy} = \frac{S_{yy}^2}{S_{yy} S_{yy}^*}
\end{equation}

Multiple coherence is defined as the correlation coefficient describing the possible causal relationship between an output and all known inputs:

\begin{equation}
\gamma_{m} = \frac{S_{uu} S_{yy}^*}{S_{yy}} = S_{yy} - S_{nn} / S_{yy}
\end{equation}

Therefor multiple coherence can be used to evaluate the importance of unknown contributions such as measurement noise, non-linearities or unknown inputs to each output.

The partial coherence function approach is based on residual spectral density. Residual represents the spectral contribution of the desired input (x3) to an output (y) when the linear effects of other inputs (x1 and x2) are removed both from x3 and y. So can be defined:

\begin{equation}
\gamma_{3y,12} = \frac{S_{3y,12}^2}{S_{33,12} S_{yy,12}}
\end{equation}

as the residual output power due to input x3. The partial coherence itself is:

\begin{equation}
\gamma_{3y,12} = \frac{S_{3y,12}^2}{S_{33,12} S_{yy,12}}
\end{equation}

The residual power spectra can be defined following Bendat (1):

\begin{equation}
S_{ij,r} = S_{ij} (r-1)! - L_{ij} S_{ii} (r-1)! \quad \text{and} \quad L_{ij} = \frac{s_{ij} (r-1)!}{s_{rr} (r-1)!}
\end{equation}

An iterative matrix method for this solution was defined by Mahalingam (4) and Desanghere (5). A more general matrix based definition of partial coherence was defined by Potter (3), but can be proven to be identical to the iterative Bendat approach (5).

3. NOISE ENERGY ALLOCATION.

A fundamental limitation of coherence methods implies the fact that different contributing sources amplify or diminish each other in building up the considered output noise energy. Therefore a more general transferfunction method is presented to divide up energy to different sources, to quantify their relative relationships and to visualize their vectorial interrelation. For a three input-one output system the output energy can be written as a combination of inputs and their respective transferfunctions. This can be expressed using both input-output cross spectra (eq.8) or only input spectra (eq.9). In both cases a number of partial power coefficients can be given:

\begin{equation}
S_{yy} = H_1 S_{1y} + H_2 S_{2y} + H_3 S_{3y} = (A_1 + A_2 + A_3). S_{yy}
\end{equation}

\begin{equation}
S_{yy} = H_1^2 S_{11} + H_2^2 S_{22} + H_3^2 S_{33} + 2. Re(H_1 H_2 S_{12} + H_1 H_3 S_{13} + H_2 H_3 S_{23}) + (a_1 + a_2 + a_3 + a_{12} + a_{13} + a_{23}). S_{yy}
\end{equation}

The coefficients of eq. 8 yield both amplitude and phase values (3) where the coefficients (6) of eq. 9 only are scalars.
4. COMPARAISON OF COHERENCE AND ENERGY ALLOCATION COEFFICIENTS

Underneath it will be shown that ordinary, multiple and partial coherence functions are only special combinations of partial power coefficients:

\[ \gamma_{1y} S_{yy} = (a_1 + \gamma_{12} a_2 + \gamma_{13} a_3 + a_{12} + a_{13}) S_{yy} \]

\[ \gamma_{1y,23} S_{yy,23} = a_1 (1 - \gamma_{13}) S_{yy} \]

Thus the ordinary coherence overestimates the contribution of a specific source to the output; the partial coherence gives an underestimation. For a two input system with sources of equal strength a crosstalk of 10% results in a coherent power estimation of 75% (ordinary) and 25% (partial). Of course it has to be considered that the more powerful energy allocation procedures require the knowledge of the transfer functions which can require some additional measurements.

5. Example: A TWO SPEAKER SYSTEM

To illustrate the use of the energy allocation procedure a simple experiment was carried out: two loudspeakers emitted identical noise with a variable phase difference. In table 1. the partial power coefficients are given for the noise output. For a phase variation of 90° a 3dB summation, for in phase radiation a 6dB augmentation is found, where the out of phase radiation showed a noise cancelation of 30dB.

<table>
<thead>
<tr>
<th>phase difference</th>
<th>( a_1 )</th>
<th>( a_2 )</th>
<th>( a_{12} )</th>
<th>total</th>
</tr>
</thead>
<tbody>
<tr>
<td>0°</td>
<td>83.7</td>
<td>83.6</td>
<td>86.7</td>
<td>89.7</td>
</tr>
<tr>
<td>90°</td>
<td>83.7</td>
<td>83.5</td>
<td>-77.4</td>
<td>86.0</td>
</tr>
<tr>
<td>180°</td>
<td>83.6</td>
<td>83.8</td>
<td>-86.4</td>
<td>58.3</td>
</tr>
</tbody>
</table>

6. MECHANICAL NOISE IN AN AUDIO CASSETTE DECK

About the mechanical noise of a cassette deck three possible sources were expected: electrical engine, push-bar, friction drive (fig. 2). Characteristic signals at the suspected sources were obtained using accelerometers and acoustic probes. Analysis of the noise spectrum (fig. 1) yield four important frequency regions:

<table>
<thead>
<tr>
<th>freq. range</th>
<th>mult. coh.</th>
<th>ordinary (partial) coh. at source:</th>
</tr>
</thead>
<tbody>
<tr>
<td>420 Hz.</td>
<td>76.4%</td>
<td>1 [73.7 (46.7) 17.2 (.20) 15.8 (.7)]</td>
</tr>
<tr>
<td>550</td>
<td>66.4</td>
<td>2 [15.3 (.97) 45.7 (20.5) 13.3 (15.3)]</td>
</tr>
<tr>
<td>700</td>
<td>97.2</td>
<td>3 [71.0 (0.0) 13.1 (.1) 97.1 (25.4)]</td>
</tr>
<tr>
<td>1090</td>
<td>90.9</td>
<td>4 [90.1 (86.5) .7 (.1) 3.9 (.6)]</td>
</tr>
</tbody>
</table>

Coherence analysis could identify the peaks around 700 Hz. (friction drive) and 1090 Hz. (engine). Energy allocation analysis was necessary around 420 and 550 Hz., fig.3. Around 420 Hz. two uncorrelated sources are found: engine and friction drive. Around 550 Hz. the friction drive was shown to be the major source, where the engine vibration gave the tendency to reduce the noise level (noise cancelation).
7. BIBLIOGRAPHY


Fig. 1. Noise Spectrum of a Cassette deck

Fig. 2. Backside View of a Cassette deck-Drive

<table>
<thead>
<tr>
<th>Inputs:</th>
<th>Energy Allocation</th>
<th>420 Hz.</th>
<th>550 Hz.</th>
</tr>
</thead>
<tbody>
<tr>
<td>1. ENGINE</td>
<td>$a_1$</td>
<td>35.0</td>
<td>31.3</td>
</tr>
<tr>
<td>2. PUSH BAR</td>
<td>$a_{12}$ $a_2$</td>
<td>1.4</td>
<td>15.3</td>
</tr>
<tr>
<td>3. FRICTION DRIVE</td>
<td>$a_{13}$ $a_{23}$ $a_3$</td>
<td>4.7</td>
<td>19.2</td>
</tr>
<tr>
<td>Output:</td>
<td>$a_3$</td>
<td>22.8</td>
<td>38.1</td>
</tr>
</tbody>
</table>

Fig. 3. Source Energy Allocation of a Cassette deck.
TRANSMISSION LOSS OF AIRCRAFT STRUCTURES DETERMINED USING ACOUSTIC INTENSITY TECHNIQUE

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INTRODUCTION
The results of laboratory studies on the transmission of sound through four areas of an airplane fuselage sidewall were studied. The airplane fuselage was studied in both reverberant and anechoic spaces, since the exterior sound fields for these two idealized cases are well known. A loudspeaker created the sound fields. The acoustic intensity transmitted into the cabin interior through the different fuselage areas was determined with the new two-microphone acoustic intensity technique[1-3]. The transmission loss of the different fuselage areas was measured and compared with theoretical predictions in order to discover any difficulties which may be encountered with this approach.

BACKGROUND
One check on the accuracy of the acoustic intensity measurement approach for transmission path identification, is to determine the transmission loss (TL) of the airplane panels, since experimental and theoretical knowledge of the TL of such panels already exists. The transmission loss can be defined as:

\[ TL = 10 \log_{10} \left( \frac{\langle I_i \rangle_S}{\langle I_t \rangle_S} \right) \]  

where \( \langle I_i \rangle_S \) and \( \langle I_t \rangle_S \) are the incident and transmitted intensities averaged over the panel area \( S \).

EXPERIMENTAL SET-UP
The fuselage was suspended in the semi-anechoic chamber as shown in Fig. 1. Fiberglass sheets were placed on the floor beneath the fuselage to eliminate floor reflections. The transmitted acoustic intensity, \( \langle I_t \rangle_S \), for each panel was measured by sweeping as close as possible the two-microphone array over the interior of each panel of interest. The incident acoustic intensity, \( \langle I_i \rangle_S \), was also measured using the two-microphone technique. To do this, the fuselage was removed and the acoustic intensity was measured over the same
areas as was previously occupied by the fuselage panels under investigation.

COMPARISON WITH MASS LAW THEORY

The tests in the semi-anechoic chamber were performed when the fuselage interior was made reasonably anechoic and when the major flanking paths were blocked with lead-backed foam. To obtain an anechoic interior, the side opposite to the starboard measurement side, was filled with fiberglass about 0.25m thick. Lead-backed foam was attached to the exterior of the fuselage surface to eliminate the major flanking paths. Flanking paths were considered to be those paths, not including the panel of interest, that could transmit a considerable amount of acoustic intensity. For example, if the TL of the back passenger panel was to be measured, then the back passenger window and the entire door along with the windshield and dashboard, was covered.

With these two modifications to the fuselage, the TL of the panel areas were measured and calculated. The narrow band measurement of TL for the back passenger window is compared with the theoretical mass law in Fig. 2. The theoretical mass law for the TL of a flat homogeneous panel is [4]:

\[
TL(\theta) = 10 \log \left[ 1 + \left( \frac{\omega \rho_s}{(2 \rho_c)} \right)^2 \cos^2 \theta \right] \text{ dB.} \quad (2)
\]

In this case, the angle of incidence \( \theta \) is normal to the surface, so \( \theta = 0 \). The theoretical prediction of the TL seems to agree very well with that found experimentally (Figure 3). The magnitude and slope of the experimental and theoretical TL curves above 500 Hz are nearly the same.

EFFECTS OF FLANKING TRANSMISSION

Tests were also performed to find the effects of flanking noise on the measurements of the TL of the panels. For this, all the exterior lead-backed foam was removed to expose the flanking paths. In principle, if the transmission loss of panel A was being measured, the effect of the transmitted intensity radiated from panels B and C, should not change the measured values of the TL of Panel A provided this interior panel surface can be assumed non-absorbing.

A dramatic difference is seen in Fig. 3, where the TL of the back passenger panel (aluminum panel with trim) was measured. For this, the horn was pointed at a nearby window and aluminum panel, and not pointed at the panel under investigation. In Fig. 3, the TL for the case of blocked flanking paths, Curve 1, is higher than the TL with unblocked paths, Curve 2, in all but a few of the frequency bands. For frequencies above 1000 Hz, the curves begin to diverge, varying up to 30 dB at 5000 Hz. As in this case when the source is pointed at a panel with a low TL such as a window, then the transmission through the window will dominate particularly in the high frequency range. Thus the TL of the panel is dramatically underestimated in the high frequency range where
CROCKER, Malcolm J., Transmission Loss Using Acoustic Intensity

The space-averaged interior cabin pressure level was quite well predicted from the measured transmitted intensities of different panels using a simple theoretical model.

REFERENCES

ACKNOWLEDGMENTS
The authors would like to thank the National Aeronautics and Space Administration (NASA Langley), Hampton, Virginia for financial support under Grant number NAG-1-58.

Fig. 1 Airplane in semi-anechoic chamber.

Fig. 2 TL of plexiglass window; Measured --- , Mass law prediction ----.

Fig. 3 TL of wall panel; Flanking paths: 1) Blocked, 2) Unblocked.

Fig. 4 Fuselage Noise Reduction; Measured o---o---o, Predicted o--o--o.
the measurement of the intensity transmitted through the high TL panel is severely contaminated by the intensity transmitted through the low TL window panel.

Fig. 3 shows that in order to obtain accurate estimates of the TL of a panel, flanking paths must be blocked when the flanking is very strong. In the low frequency range, the contamination due to flanking paths is not so drastic since all the panels have TL values of the same magnitude.

THEORETICAL PREDICTION OF FUSELAGE NOISE REDUCTION

A room equation was used to predict the noise reduction, NR:

\[
NR = L_p - L_{pi} - 10 \log_{10} \left( \frac{Q}{4\pi r^2} + \frac{4}{R} \right),
\]

where \( L_p \) and \( L_{pi} \) are the exterior and interior sound pressure levels, respectively, \( L_w \) is the level of the transmitted sound power, \( Q \) is the directivity factor (assumed unity), \( r \) is the distance from the fuselage wall to the receiver, and \( R \) is the room constant. The room constant, \( R \), was obtained from the reverberation time measurements. The acoustic pressure, \( L_{pi} \), at one chosen location in the cabin was calculated by using Eq. (3) for one panel. Assuming the acoustic pressures transmitted through the panels are incoherent, the overall noise reduction was obtained by summing the mean-square acoustic pressures radiated from each of the panels comprising the fuselage walls. In this investigation, altogether ten panels were involved in the theoretical prediction of NR. The noise reduction, NR, at a location near the pilot's head was predicted from Eq. (3) and is shown in Fig. 4 as a dashed line.

The noise reduction was measured by subtracting the spatially averaged sound pressure level outside the fuselage from the sound pressure level measured inside the cabin while the airplane was hung in the reverberation room. Measured results averaged over one-third octave frequency bands were also plotted in Figure 4 as a solid line. Fairly good agreement between the predicted and measured values suggests that Eq. (3) can be used to predict interior cabin sound pressure levels.

Discrepancies between the predicted and the measured noise reductions in the frequency range below 400 Hz, are shown in Fig. 4. These are believed to be caused by the inaccurate measured values of the interior room constant, as too few points were chosen for a space average.

CONCLUSIONS

It has been shown in this paper that the two-microphone acoustic intensity approach can be utilized successfully to determine the dominant paths of sound transmission into an airplane cabin. It appears that if there are strong flanking paths present, inaccurate estimates are obtained of the sound intensity transmitted by the primary sound transmission path.
CARACTERISATION D'UNE METHODE DE MESURE DE REPONSE IMPULSIONNELLE
EN ACOUSTIQUE DES SALLES.

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I. INTRODUCTION

La mesure des réponses impulsionnelles liées à un phénomène acoustique est
le point de départ pour traiter aussi bien les problèmes de
caractérisation, de modélisation et de reproduction de ce phénomène
acoustique [1]. Un compromis entre la qualité acoustique et la puissance
des transducteurs amène à travailler avec de mauvais rapports signal sur
bruit. La mesure décrite ici pallie cet inconvénient. Nous avons procédé à
une étude expérimentale systématique pour vérifier les performances
théoriques de cette mesure. Les systèmes mesurés étaient principalement des
haut-parleurs et des filtres analogiques.

II. PRINCIPE ET REALISATION

Une séquence périodique générée par un registre à décalage [2] traverse le
système linéaire étudié. L'intercorrélation de la sortie du système avec
 cette même séquence donne la réponse impulsionnelle de ce système [3].
Soit $N = 2^m - 1$ la longueur de la séquence $x$ générée par un registre à $m$
etages. La propriété fondamentale de $x$ est que sa fonction
d'autocorrélation vaut :

$$X(n) = N \delta(n) - 1$$

$\delta$ impulsion de Dirac.

Si $h$ est la réponse impulsionnelle du système, on a :

$$y = h \ast x + b$$

sortie du système ; $b$ : bruit additif de mesure.

On trouve facilement que le résultat de l'intercorrélation entre $y$ et $x$
vaut :

$$r = Nh - \overline{h} + b \ast x,$$

où $\overline{h}$ : composante continue de $h$.

$r$ est donc le bruit additif près $b \ast x$ une mesure de la réponse
impulsionnelle du système moins sa composante continue, laquelle est
toujours éliminée en pratique par l'appareillage.

Un processeur spécialisé [4] réalise les conversions analogique-numérique
(12 bits) et génère la séquence pseudo-aléatoire (jusqu'à $m = 16$).

Un microcalculateur LSI 11-03 assure la gestion du processeur et recueille
la sortie du système pour l'enregistrer sur un disque RK 05. Cet ensemble
permet de faire des mesures jusqu'à la fréquence d'échantillonnage de 36
kHz. Une chaîne analogique d'amplification et filtrage (anti-aliasing)
complète ce dispositif.
Calcul de l'intercorrélation : N n'étant pas une puissance de 2, le calcul par FFT nécessite une interpolation [5] et un calcul sur autant de points (jusqu'à 64 K) introduit un bruit non négligeable. Pour l'éviter nous avons calculé directement l'intercorrélation entre y et x grâce au processeur utilisé pour la mesure qui accumule sur 32 bits des produits d'entiers de 16 bits. Le calcul de r ne produit donc aucun bruit de calcul.

III. BRUIT ADDITIF

Ce bruit est dû principalement au souffle des amplis et filtres, au bruit acoustique et au bruit de quantification.

1) Prévisions théoriques.
En supposant que b n'est pas corrélate avec h * x on exprime les énergies :

\[ E(y) = NE(h) + E(b) \quad E(r) = N^2E(h) + NE(b). \]

On définit les grandeurs suivantes ramenées à l'échantillon :

- variance du bruit additif sur y : \[ \sigma_y^2 = E(b)/N \]
- sur r : \[ \sigma_r^2 = E(b) \]

\[ y_e \text{ valeur efficace du signal utile sur y: } y_e^2 = E(h) \]
\[ r_e \text{ sur r : } r_e^2 = NE(h) \]

Ainsi le rapport signal sur bruit usuel est le même sur y et sur r :

\[ S/B = 10 \log y_e^2/\sigma_y^2 = 10 \log r_e^2/\sigma_r^2 = 10 \log NE(h)/E(b) \]

par contre \[ r_\text{utile} = Nh \text{ et } \sigma_r = \sqrt{N} \sigma_y \]

D'autre part on observe qu'à 1 ou 2 dB près le facteur de crête sur y reste de l'ordre de 12 dB : \( E(h) = y_e^2 \) est fixée à une valeur constante égale à l'appareillage (limite de saturation du convertisseur) ; de même \( \sigma_y \) ne dépend que des conditions expérimentales. Le gain en dynamique est égal à la différence des facteurs de crête de r et de y : pour \( m=14 \), le gain est typiquement de 30 à 50 dB sur les réponses impulsionnelles mesurées, ce qui fait toute la différence entre une mesure exploitable ou non.

2) Résultats expérimentaux.

a) soit \( \sigma_{y_e}^2 \) la variance du bruit électrique mesuré à l'entrée du convertisseur analogique-numérique et \( \sigma_{y_e}^2 \) la variance du bruit additif sur y calculée par comparaison de deux périodes successives de y. On observe une bonne correspondance entre \( \sigma_{y_e} \) varie systématiquement et \( \sigma_{y_e} \) sauf pour les petits niveaux de bruit électrique où la valeur calculée \( \sigma_{y_e} \) se maintient à un niveau plancher supérieur au bruit de quantification. Ce bruit supplémentaire est analysé plus loin.

b) Transmission du bruit additif sur la réponse impulsionnelle.
Le bruit additif sur r est évalué en comparant deux mesures successives sur une portion de la réponse impulsionnelle où le signal est faible. Soit \( \sigma_{r_e}^2 \) la variance de ce bruit calculée.

Le tableau 1 montre qu'aux approximations près dues à l'estimation de \( \sigma_{r_e} \) sur un petit nombre de mesures (3 à 10), \( \sigma_{r_e} = \sqrt{N \sigma_{y_e}} \) est une bonne estimation du bruit additif sur r.
Tableau 1 : variations de $\sigma_{re}/\sigma_{rt}$ en fonction de $\sigma_{ye}$.
Mesures d'un haut-parleur Philips, $m = 14$, $0$ dB $= y_e$, $r$ max. $= 1687$.

<table>
<thead>
<tr>
<th>$\sigma_{ye}$ dB</th>
<th>$\sigma_{rt}$</th>
<th>$\sigma_{re}/\sigma_{rt}$: min.</th>
<th>max.</th>
<th>moyen</th>
</tr>
</thead>
<tbody>
<tr>
<td>6</td>
<td>9.866</td>
<td>0.81</td>
<td>1.37</td>
<td>1.04</td>
</tr>
<tr>
<td>0</td>
<td>11.146</td>
<td>0.91</td>
<td>1.36</td>
<td>1.09</td>
</tr>
<tr>
<td>-6</td>
<td>7.602</td>
<td>0.94</td>
<td>1.29</td>
<td>1.07</td>
</tr>
<tr>
<td>-12</td>
<td>4.318</td>
<td>1.09</td>
<td>1.58</td>
<td>1.34</td>
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<td>-18</td>
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<td>0.54</td>
<td>0.94</td>
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<td>-30</td>
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<td>-42</td>
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<td>0.73</td>
<td>1.19</td>
<td>0.93</td>
</tr>
<tr>
<td>-48</td>
<td>0.087</td>
<td>0.48</td>
<td>0.86</td>
<td>0.70</td>
</tr>
<tr>
<td>-54</td>
<td>0.043</td>
<td>0.60</td>
<td>0.97</td>
<td>0.74</td>
</tr>
<tr>
<td>-60</td>
<td>0.027</td>
<td>0.67</td>
<td>0.71</td>
<td>0.70</td>
</tr>
<tr>
<td>-63</td>
<td>0.027</td>
<td>0.57</td>
<td>1.95</td>
<td>1.00</td>
</tr>
</tbody>
</table>

IV. BRUIT DE PHASE

La comparaison de mesures successives de la même réponse impulsionnelle nous a montré d'une part que les perturbations d'une mesure à l'autre sont beaucoup plus importantes que celles dues au bruit additif, d'autre part que ces perturbations sont correctement représentées par un effet de dilatation. Sans vouloir expliquer cet effet nous l'avons évalué et pu le corriger en grande partie.

1) Définition d'outils mathématiques appropriés.
Soit $y = r_1 - r_2$ différence entre deux mesures, $r' = dr/dt$ vecteur dérivée $r'' = t dr/dt$ vecteur dilatation de $r$.
On définit : indice de déphasage $I_p = d r'/ \sqrt{d^2 r''}$
Si $I_p = 1$, $r_1$ et $r_2$ sont décalés d'une fraction $T_0$ de la période $T = 1/f_0$
indice de dilatation $I_d = d r'' / \sqrt{d^2 r''}$
Si $I_d = 1$, $r_2$ est échantillonné à $f_0$ et $f'$ est $= df/f_0$.
Ces deux indices s'interprètent comme les cosinus respectifs des angles que fait $d$ avec $r'$ et $r''$
On appellera décalage optimal le décalage qui annule $I_p$. Si $I_p = 1$, le décalage optimal est obtenu pour $T_0 = n_0 \frac{\pi}{2}$, où $n_0 = r'' / r''$.
L'indice $I_d$ est alors recalculé par rapport à $r'' = (n_0) / r''$.

2) Résultats expérimentaux.
Tableau 2 : évaluation de $I_p$, $I_d$ et évolution de ces indices : $I_p^{(2)}$

<table>
<thead>
<tr>
<th>$E(r)$</th>
<th>$n_0 \frac{\pi}{2} \times 10^5$</th>
<th>$E(d)$</th>
<th>$I_p$</th>
<th>$I_d$</th>
<th>$I_p^{(2)}$</th>
<th>$I_d^{(2)}$</th>
<th>$E(d)^{(2)}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>haut-parleur Philips en chambre sourde à 2 m du micro, $m = 14$, $f_e = 32$ KHz valeur maximale $r : 1500$, $r_0 = 0.05$</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>66 dB</td>
<td>207</td>
<td>31 dB</td>
<td>0.99</td>
<td>0.99</td>
<td>0.11</td>
<td>0.36</td>
<td>14 dB</td>
</tr>
<tr>
<td>66 dB</td>
<td>207</td>
<td>15 dB</td>
<td>0.99</td>
<td>0.99</td>
<td>0.16 $10^{-4}$</td>
<td>0.50 $15$ dB</td>
<td></td>
</tr>
<tr>
<td>66 dB</td>
<td>207</td>
<td>20 dB</td>
<td>0.99</td>
<td>1.00</td>
<td>0.74 $10^{-5}$</td>
<td>0.64 17 dB</td>
<td></td>
</tr>
</tbody>
</table>

filtre analogique, $m=16$, $f_e = 32$ KHz, valeur maximale $r : 6200$, $\sigma_r = 0.008$

| 80 dB  | 7     | 8     | 12 dB | 0.93  | 0.95  | 0.28 $10^{-4}$ | 0.53 3 dB  |
|--------|-------|-------|-------|-------|-------|-------------|-------------|--------------|
| 80 dB  | 7     | 8     | 12 dB | 0.96  | 0.97  | 0.32 $10^{-4}$ | 0.52 0 dB  |             |
| 80 dB  | 7     | 4     | 5 dB  | 0.92  | 0.92  | 0.65 $10^{-4}$ | 0.29 -3 dB |             |
Cet effet de dilatation corrigé, il reste un effet systématique qu'il est encore possible de réduire pour ne laisser subsister que le bruit additif. Remarquons que l'effet de dilatation agit dans le même sens pendant des périodes longues (plusieurs heures). Cela interdit absolument tout moyen de laisser augmenter le bruit sans connaître la dilatation précédente.

3) Prévision du bruit de phase à partir de l'étude de $y$.
Si deux mesures sont faites avec des fréquences d'échantillonnage $f_e$ et $f_e + df$, la première sera dilatée par rapport à la seconde de $\Xi = df/f_e$. On peut montrer qu'une dilatation sur $y$ provoquera une dilatation sur $r$ du même ordre. Il est donc possible de définir un indice $I_d$ et une dilatation $\Xi$ sur l'enregistrement $y$ de la même façon que pour $r$.
Tableau 3: résultats de ce type d'analyse sur $y$. On a repris les séries de mesures dont il était question dans le tableau 2.

<table>
<thead>
<tr>
<th>S/B</th>
<th>$\sigma_{r_\delta}$</th>
<th>$I_d$</th>
<th>$\Xi \times 10^5$</th>
</tr>
</thead>
<tbody>
<tr>
<td>52 dB</td>
<td>0.077</td>
<td>0.68</td>
<td>1.52</td>
</tr>
<tr>
<td>49 dB</td>
<td>0.090</td>
<td>0.18</td>
<td>0.39</td>
</tr>
<tr>
<td>48 dB</td>
<td>0.090</td>
<td>0.25</td>
<td>0.68</td>
</tr>
<tr>
<td>51 dB</td>
<td>0.089</td>
<td>0.81</td>
<td>2.34</td>
</tr>
<tr>
<td>54 dB</td>
<td>0.080</td>
<td>0.24</td>
<td>8.99</td>
</tr>
<tr>
<td>55 dB</td>
<td>0.070</td>
<td>0.22</td>
<td>7.22</td>
</tr>
<tr>
<td>55 dB</td>
<td>0.071</td>
<td>0.27</td>
<td>8.78</td>
</tr>
<tr>
<td>59 dB</td>
<td>0.046</td>
<td>0.30</td>
<td>6.46</td>
</tr>
</tbody>
</table>

Ainsi seules les mesures avec un registre à 16 étages ont pu donner une bonne prévision du bruit de phase. Cela s'explique bien puisque l'importance du bruit de phase est très liée d'après nos observations à la durée et à l'espacement des mesures (effet de dérive).

V. CONCLUSION

Nous avons présenté une évaluation expérimentale d'une méthode de mesure acoustique prometteuse. Les résultats obtenus confirment les performances théoriquement attendues, mais on doit prendre en compte des effets systématiques qui s'ajoutent au bruit additif. On doit aussi prendre en compte le recouvrement du à une fenêtre de mesure limitée, ainsi que des impulsions parasites que peut produire le système d'acquisition.
Nous utilisons cette méthode de mesure pour l'étude des systèmes acoustiques de prise et de restitution du son.

BIBLIOGRAPHIE
EQUIVALENT BANDWIDTH OF A GENERAL CLASS OF POLYNOMIAL SMOOTHERS

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This paper presents a comprehensive account of the detailed investigation of the properties of a general class of polynomial Least-Mean-Square-Fit (LMSF) smoothers in the presence of a white or color input sequence. Results of the investigation show that the LMSF can be described as a low-pass filter, the frequency response characteristics of which can be evaluated exactly. A particularly useful and simple result derived from the frequency response characteristics is the LMSF equivalent bandwidth. Results of the analysis are verified by simulation and the usefulness of the LMSF equivalent noise bandwidth is demonstrated.

INTRODUCTION

This study was motivated by the performance parameter (bias and variance) evaluation of sonar bearing trackers from experimental measured data. The bearing tracker works on the following principle. A split beam is formed from a sensor array. Time delay between half-beam phase center is measured via a generalized cross-correlator (GCC) [1]. Fig. 1 shows a closed-loop implementation of a typical bearing tracker. The evaluation procedure consists of processing a finite error sequence obtained by comparing tracker bearings from the corresponding known reference bearings. An LMSF procedure (Fig. 2) is employed to estimate the random and the slowly time-varying mean components of the tracker bearing errors.

In applying an LMSF for the purpose just stated, three issues require immediate attention: (1) the form and the order of the LMSF, (2) the appropriate correction factor for the variance estimate, and (3) the variance reduction on the mean estimate.

A key parameter needed to resolve these issues is given by the LMSF equivalent noise bandwidth. To demonstrate its usefulness, a basic result of the investigation is first discussed, then followed by a rigorous derivation.

Let \( \{x(i)\} \) be the input sequence with variance \( \text{Var}(x) \) and bandwidth \( B_i \), \( |H(\omega)|^2 \) be the normalized LMSF frequency response function with equivalent bandwidth \( B_0 \), then the variance of the LMSF output sequence \( \{y(i)\} \), \( \text{Var}(y) \), is related to the input variance by (Fig. 3) \( \text{Var}(y) = (1-B_0/B_i)\text{Var}(x) \).
In many applications \( \{y(i)\} \) is nearly a stationary sequence whose variance can be estimated using standard technique, thus, the unbiased variance estimate of the input process can be obtained.

**PROBLEM FORMULATION**

Let \( \{x(i)\}, i = 0, \ldots, N-1 \) be a time-sampled sequence of \( N \) observations from a random process, the mean and variance of which are given by

\[
E[x(i)] = m(i), \quad E[(x(i)-m(i))(x(j)-m(j))] = V R(i,j); \quad j = 0, \ldots, N-1,
\]

where \( E[\cdot] \) is the expectation operator and \( R(i,j) \) is the correlation coefficient. It is assumed that \( m(i) \) can be modeled by a \( k \)-th order polynomial;

\[
m(i) = \sum_{n=0}^{K} a(n) t(i)^n, \quad t(i) = iT
\]

(1)

where \( T \) is the sampling interval. The sampled sequence can be written concisely in matrix form as

\[
x = Ha + w
\]

(2)

where \( x = [x(o), x(1), \ldots, x(N-1)]^T \), \( a = [a(o), a(1), \ldots, a(K)]^T \), \( w = [w(o), w(1), \ldots, w(N-1)]^T \) and \( H \) is an \( NxK \) matrix, the \( i \)-th row of which is given by \( H(i) = [1, t(i), t(i)^2, \ldots, t(i)^K] \). The random noise vector \( w \) has mean \( E[w] = 0 \) and covariance \( E[ww]^T = V R \). The parameters of interest are the unknown coefficient vectors \( a \). A least mean square estimate of \( a \) yields [2]

\[
\hat{a} = (H^T H)^{-1} H^T x
\]

(3)

Let \( y = [y(o), y(1), \ldots, y(N-1)]^T \) be the LMSF output vector, then one can write \( y = Lx \) where the linear operator \( L = H(H^T H)^{-1} H^T \) is an \( NxN \) symmetric matrix and \( LL^T = L \). This describes the input/output relation of the LMSF and from which the equivalent bandwidth and the spectral response can be calculated.

**LMSF EQUIVALENT BANDWIDTH**

Let \( \{X(\ell)\} \) and \( \{Y(\ell)\}, \ell = 0, 1, \ldots, N-1 \), be the Discrete Fourier Transform (DFT) of the input and output sequences, respectively. Then the equivalent input \( \{B_i\} \) and output \( \{B_o\} \) bandwidths can be written as [3]

\[
B_i = \sum_{\ell=0}^{N-1} \frac{E[|X(\ell)|^2]}{E[|X(0)|^2]} \Delta f; \quad B_o = \sum_{\ell=0}^{N-1} \frac{E[|Y(\ell)|^2]}{E[|Y(0)|^2]} \Delta f
\]

(4)
where $\Delta f = 1/NT$ is the discrete frequency spacing. But, by definition one can write

$$E\{|X(0)|^2\} = V \text{SUM}\{R\}; \quad E\{|Y(0)|^2\} = V \text{SUM}\{LRL^T\} \tag{5}$$

where $\text{SUM}\{\}$ denotes the element sum of a matrix and $L$ is the LMSF operator defined earlier. Now, applying the Parseval relation yields

$$\sum_{k=0}^{N-1} E\{|X(k)|^2\} = V N \text{tr}\{R\}; \quad \sum_{k=0}^{N-1} E\{|Y(k)|^2\} = V N \text{tr}\{LR\} \tag{6}$$

where $\text{tr}\{\}$ denotes the trace of a matrix. Therefore, using (5) and (6) in (4) yields

$$\frac{B_0}{B_i} = \frac{\text{tr}\{LR\}}{\text{tr}\{R\}} \frac{\text{SUM}\{R\}}{\text{SUM}\{LRL^T\}} = \text{tr}\{H(H^TH)^{-1}H^TR\}/N \tag{7}$$

where the general identity $\text{SUM}\{R\} = \text{SUM}\{LRL^T\}$ and $\text{tr}\{R\} = N$ have been used [3]. Thus, (7) relates the LMSF equivalent bandwidth to the input bandwidth. For the case of white input sequence, i.e., $R = I$, (7) reduces to $B_0 = Bw \Delta (K+1)/NT$. If the input sequence is correlated but $B_i >> Bw$, then the foregoing is still an excellent approximation since $\text{tr}\{H(H^TH)^{-1}H^TH\} = K+1$ and $B_i = 1/T$. Note that the LMSF bandwidth can be stated in terms of the triplet $(K,N,T)$, where $K$ is the order of the fit. $N$ is the sample size, and $T$ is the sampling time. The above simple and handy relation can be used to determine the triplet for a given desired LMSF bandwidth.

**UNBIASED VARIANCE ESTIMATE**

We have demonstrated earlier the important role played by the LMSF equivalent bandwidth in obtaining an unbiased variance estimate. Here, a rigorous derivation is presented. Let the error residual vector be defined as (Fig. 2) $\mathbf{e} = \mathbf{x} - \mathbf{y} = \mathbf{H}(\mathbf{a} - \mathbf{a}) + \mathbf{w}$. Now the standard estimate of the variance gives $\text{Var}(s) = \mathbf{e}^T \mathbf{e} / N = \text{tr}\{\mathbf{e} \mathbf{e}^T\} / N$ and the expected value of this estimate yields

$$E\{\text{Var}(s)\} = V \text{tr}\{LRL^T - LR - RL^T + R\} / N = (1 - \text{tr}\{LR\} / N)V \tag{8}$$

Where the relations $\text{tr}\{R\} = N$, $\text{tr}\{LR\} = \text{tr}\{RL^T\}$ and $\text{tr}\{LRL^T\} = \text{tr}\{LR\}$ have been used.

Using (7) in (8), the unbiased variance estimate is

$$\hat{V} = \text{Var}(s) / (1 - B_0 / B_i) \tag{9}$$

This expression is identical to the one obtained via intuitive argument. If $Bw >> B_i$, then (9) can be approximated by $V = \text{Var}(s) / [(1 - (K+1)/(NTB_i))$, which is a function of the triplet $(K,N,T)$ and the equivalent bandwidth of the input sequence.
COMPUTER ANALYSES

Extensive computer analyses and simulations were performed to verify the derived properties of the LMSF procedure.

Assuming white input sequence, Fig. 4 shows the resulting discrete power spectra for an LMSF filter for 100 samples (N) sampled at a 1-second rate (T) and the order of the fit (k) equal to 2, 4, 6. Note that the order of the polynomial fit varies directly with the pass band of the LMSF filter.

With identical conditions as above, Fig. (5) shows the theoretical discrete power spectrum for a Fourier LMSF (K=6). Note that the equivalent two-sided bandwidth is obviously given by \((K+1)/(NT)\), which is the same as the polynomial LMSF filter. However, the rolloff is infinite. Note that all LMSF filters have identical equivalent bandwidths; however, the fine structure depends on the choice of the fitting functions.

The discrete power transfer function for the LMSF filter is, in general, dependent on the correlation of the input sequence. Fig. 5 shows the comparison between the power transfer for an input white sequence and input correlated noise sequence for a 6th order polynomial fit.

In Fig. 6, the ratio of the equivalent bandwidth of the LMSF to the equivalent bandwidth of the input sequence (r) was equal to .25. It is apparent that the power transfer function for the white and correlated input sequences are practically identical until approximately -14dB. Not shown but interesting is the fact that the zero-order polynomial LMSF and all Fourier LMSF filters are not dependent on the correlation of the input sequence.

Fig. 7 compares the theoretical percentage error in the standard correction factor \(1/(1-(K+1)/N)\) and the approximate correction factor \(1/(1-(K+1)/(NT^2))\) versus the parameter r. The number of samples was set at 100, the sampling time was 1 second and the order of the polynomial LMSF was 6. Fig. 7 shows that the approximate correction factor is significantly better than the standard correction factor in the useful region where \(r \leq .25\).

SUMMARY AND CONCLUSIONS

The LMSF can be modeled and analyzed as a low-pass filter acting on a white or correlated noise input sequence. The equivalent bandwidth of the LMSF can be calculated exactly.

The calculation of variance based on LMSF residuals is known to be biased and an appropriate correction factor is needed. For the case of uncorrelated data, the correction factor is known and is determined by the order of the fit, the number of samples and the sampling interval. It was shown in this study that the correction factor can be equated to the equivalent bandwidth of the LMSF when treated and analyzed as a filter.
An excellent approximation, which depends on the estimate of the input equivalent two-sided bandwidth, has been derived which corrects the sum of squared residuals of the LMSF to an unbiased estimate of the desired input variance.

REFERENCES


**Fig. 1. Basic Block Diagram of Bearing Tracker**

**Fig. 2. Simplified Diagram, LMSF Data Analysis Loop**
Fig. 3. LMSF Equivalent Transfer Function

Fig. 4. LMSF Power vs Frequency, Polynomial Fit, 100 Samples

Fig. 5. LMSF Power vs Frequency 4th Order SIN / COS Fit, 100 Samples

Fig. 6. LMSF Power vs Frequency 8th Order Polynomial Fit, 100 Samples

Fig. 7. % Error vs Output / Input Bandwidth Ratio 8th Order Polynomial Fit, 100 Samples
AN ANALOG INTENSITY ANALYZER BASED ON SWITCHED-CAPACITOR FILTERS

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ABSTRACT

This paper describes a project of a portable, self-sufficient, analog sound intensity meter, which is suitable for measurements under field conditions and which can display the intensity spectrum in one band at a time (octave or third octave). The filtering section of this analyzer is based on sampled-data devices.

INTRODUCTION

In acoustics there is often a need of a portable, battery-operated sound intensity meter, which is suitable for measurements under field conditions and which is as easily operated as hand-held sound pressure meters.

Considering the expected application in industrial environments, small analog intensity meters which only give the result in a linear or A-weighted channel are not suitable. Experiments performed in the Acoustics Laboratory have indicated that measurements of sound intensity have to be taken in frequency bands to minimize various errors which inevitably occur.

The two-microphone technique employed to measure acoustic intensity is very sensitive for both amplitude and phase mismatch in the two channels as many authors have pointed out. The problem involved in matching analog filters have prevented the development of such analyzers.

Digital signal processing has solved the matching problem [1]. Unfortunately, sound intensity real-time analyzers based on FFT processors or digital filters do not as yet meet the foregoing requirements because of their bulky size and heavy power consumption.

With the advent of Charge Transfer Devices, the problem of matching of analog filters has largely been overcome.

This paper will describe briefly an analog/discrete-time implementation of a portable sound intensity meter based on a new filter technological innovation - the sampled-data or switched-capacitor filter.
SWITCHED-CAPACITOR FILTERS

Principle

Resistors are completely absent in a switched-capacitor filter and are replaced with the switched elements seen in Fig.1

\[
\begin{align*}
\text{Fig.1: A switched-capacitor equivalent to the analog resistor applies when } f_c, \text{ the switch rate, is high.}
\end{align*}
\]

Fig.2 shows the switched-capacitor network building block: an integrator which replaces a resistor with a capacitance ratio \( C_2/C_1 \) and a toggle switch frequency \( f_c \).

\[
\begin{align*}
\text{Fig.2: Comparison between a standard active analog integrator and a switched-capacitor integrator.}
\end{align*}
\]

Active filter functions traditionally implemented with discrete components can be fully integrated with the switched-capacitor technology [2].

Phase matching

An important property of the switched-capacitor filters is that the capacitor ratios which determine the filter characteristics can be made very precisely with LSI photolithography and any temperature or voltage coefficients for the capacitors are correspondingly reduced. The ratio accuracy can easily reach \( 1\% \) with \( 0.1\% \) - a practical upper limit.

On the basis of this property, some measurements have been performed to check the possibility of matching such devices. It appears that it is quite simple to select a pair of well phase-matched 1/1- or 1/3-octave bandpass filters and, furthermore, it can be expected that a very accurate phase matching will be possible on the same substrate.
HYBRID IMPLEMENTATION

Block diagram

Fig. 3 shows the simplified block diagram of an intensity meter implemented by continuous time and sampled-data techniques. The incoming pressure signals are directly bandpass filtered, the pressure sum and integrated pressure difference signals are multiplied and then time averaged.

Filtering section

The filtering section consists of two 1/1- or 1/3-octave switched-capacitor filters (6-pole Chebyshev). The devices used are carefully selected by matching their phase responses. The measured phase mismatch does not exceed 0.5 degree in the range from 20 Hz to 12 kHz. The center frequency of the filters is tunable by a clock frequency.

As in all sampled-data systems, signals above half the sampling frequency will be aliased and may appear in the band of interest. An antialiasing filter is required. It is a 2-pole Butterworth voltage-controlled lowpass filter which gives a maximally flat passband and has cut-off frequency is proportional to the sampling frequency.

The clock residue may affect system performance. A clock suppression filter must be added to the filter output. This filter is of the same type as the antialiasing filter.

Filter unit location

It can be shown that intensity measurements based on the two-microphone technique are much more sensitive to differences between the phase responses of the two microphone channels than to phase deviations in the velocity channel. Therefore, the two pressure channels must be better matched than the 'p' and 'v' channels.

In practice, electronic filtering often causes significant phase mismatch between the channels.

For the reason stated before, the bandpass filters are usually located in the 'p' and 'v' channels. Unfortunately, matched highpass filters have to be included in the two pressure channels. Such filters are essential to reject unwanted signals from the circuit. These signals can be 1/f- and low frequency noise as well as noise, produced by air flows in the area of the microphones. Without these filters the differences of unwanted noise will become greater than those of the acoustical signals [3].

In the implementation described here, the bandpass filters can be located in the two pressure channels because of their accurate phase matching. This eliminates the need of highpass filtering.

The foregoing modification has the advantages of increasing the signal-to-noise ratio especially at the lower frequency end of the measurement range and, of preventing velocity signal overloading which can occur particularly at low frequencies and in windy conditions.
CONCLUSIONS

Switched-capacitor filters offer precise, stable performance in small space. In contrast to all digital approaches, these analog devices present significant power, size and cost savings with a modest trade-off in their performance capabilities. Matching possibility of such devices has been checked.

In conclusion, the feasibility to build the filter unit of a portable sound intensity analyzer with switched-capacitor filters is shown.

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INTENSITY MEASUREMENTS IN STRUCTURES

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The measurement of acoustic intensity has become a well established technique. The precision and reproducibility when proper microphone and processing technique is used are well documented and described for a large number of applications. The use of intensity techniques in measurement of energy flow in structures has been devoted much less attention although it may be of the greatest benefit to the engineer working with the dynamic behaviour of mechanical structures.

All mechanical systems have to be supplied with power in order to vibrate. The flow of this power within the system can be described by the intensity vector, giving the magnitude and direction of the power flow. By measuring the direction of the intensity vector, it is possible to trace the path of the power flow and hereby locate the sources delivering the power necessary for the vibrations. Different sources can be ranked by measuring the power delivered by each source. Mapping the magnitude of the vibration intensity vector gives information about where to position damping materials.

**Theory:** The structure borne wave intensity vector, \( \vec{\mathcal{I}} \), describes the net flow of mechanical energy per unit width of a plate at a given position. The vibration intensity, \( \mathcal{I} \), in a given direction is the projection of the intensity vector, \( \vec{\mathcal{I}} \), on this direction.

The vibration intensity in the x-direction within a small cross-section of a plate is transported by the work done by two moments, \( M_{xy} \) and \( M_x \), and one force \( Q_x \).

The work can be found by multiplying the force and moments with their corresponding velocity and angular velocities. The angular velocities \( \dot{\theta}_x \) and \( \dot{\theta}_y \) are given by the following equations:

\[
\dot{\theta}_x = \frac{\partial^2 \xi}{\partial\xi x} \quad \dot{\theta}_y = \frac{\partial^2 \xi}{\partial\xi y}
\]

(1) (2)

where \( \xi \) is the displacement of the plate in the z-direction. The total vibration intensity in the x-direction is:

\[
\mathcal{I}_x = \langle Q_x \xi \rangle_t + \langle M_{xy} \dot{\theta}_y \rangle_t + \langle M_x \dot{\theta}_x \rangle_t
\]

(3)

where \( \langle \cdot \rangle_t \) denotes a time averaging.
As shown by Noiseux [1], the two moment contributions in many cases can be approximated by a modified moment component \( w_{xm} \) given by
\[
\begin{align*}
\dot{w}_{xm} &= \left\langle (M_x + M_y)/(1+\mu) \right\rangle \cdot \dot{\theta} \quad \mu \text{ is Poisson's ratio} \\
\end{align*}
\]

For the sum of the two moments we have:
\[
\begin{align*}
M_x + M_y &= -B(1+\mu) \left( \frac{\partial^2 \xi}{\partial x^2} + \frac{\partial^2 \xi}{\partial y^2} \right) \\
B &= \text{plate stiffness} \\
B &= \frac{Eh^3}{12(1-\mu^2)}
\end{align*}
\]
\[E \text{ is Young's modulus and } h \text{ the thickness of the plate. For plane waves in a free field:} \]
\[
\begin{align*}
\frac{\partial^2 \xi}{\partial x^2} + \frac{\partial^2 \xi}{\partial y^2} &= k^2 \xi \\
\text{the wave number } k &= \sqrt{\frac{\omega^2 m}{B}} \\
\omega &= \text{the angular frequency and } m = \text{the mass per unit area.}
\end{align*}
\]

Under the given conditions, the force intensity component, \( w_f \), equals the moment component \( \dot{w}_{mx} \):
\[
\dot{w}_f = \dot{w}_{mx} = \frac{\dot{w}_x}{2} \quad \text{which gives: } \quad \dot{w}_x = 2 \dot{w}_{mx} = 2 \dot{w}_{mx}
\]

From equations (4), (5), (7) and (10) we have:
\[
\dot{w}_x = < 2Bk^2 \xi \cdot \ddot{\theta} > = < 2Bk^2 \ddot{\theta} \int \xi \ d^2 t >
\]

For pure sine excitations this becomes:
\[
\dot{w}_x = < \frac{2\sqrt{\pi m}}{\omega} \dot{\theta} \cdot \xi >
\]

According to equation (12) the intensity in the \( x \)-axis can be determined by simultaneously measuring the angular velocity \( \dot{\theta}_x \) and acceleration \( \ddot{\xi}_x \). A similar equation for one dimensional waves has been derived by Pavic [2].

**Measuring technique:** The transducer configuration shown in fig. 1 has been used for measuring angular velocity and acceleration.

![Fig. 1: Vibration intensity transducer](image-url)
The acceleration \( \ddot{\xi} \) at the point of measuring is the mean value of the accelerations of the two accelerometers:

\[
\ddot{\xi} = \frac{1}{2} (\ddot{\xi}_1 + \ddot{\xi}_2)
\]

(13)

The angular velocity \( \ddot{\theta} \) is calculated from the difference between the accelerations:

\[
\ddot{\theta} = \int \left( \frac{\ddot{\xi}_2 - \ddot{\xi}_1}{\Delta r} \right) dt
\]

(14)

where \( \Delta r \) is the distance between the accelerometers.

From equations (12), (13) and (14) we have:

\[
w_x = \frac{2\sqrt{\varepsilon_m}}{\omega} < \frac{1}{2} (\ddot{\xi}_1 + \ddot{\xi}_2) \int \left( \frac{\ddot{\xi}_2 - \ddot{\xi}_1}{\Delta r} \right) dt >
\]

(15)

In a sound intensity analyzer, the sound intensity, \( I_p \), is determined from two measured sound pressures, \( p_1 \) and \( p_2 \):

\[
I_p = \left( \frac{p_1 + p_2}{2} \right)^2 \frac{1}{p_0^2} \int \left( \frac{p_2 - p_1}{\Delta r} \right) dt >
\]

(16)

where \( p_0 \) is the density of air and \( \Delta r \) is the chosen microphone distance. When measuring vibration intensity, each channel of the intensity analyzer is calibrated to measure the acceleration in \( \text{dB re } 20 \cdot 10^{-6}\text{ms}^{-2} \).

The vibration intensity, \( w_x \), in \( \text{dB re } 10^{-12}\text{W/m} \) can be obtained from the reading \( w'_x \) on the intensity analyzer using the correction:

\[
w_x = w'_x + 10 \log_{10} \left( \frac{\sqrt{\varepsilon_m} \Delta r}{\omega \Delta r} \right)
\]

(17)

This correction is valid for pure sine tones. However, using 1/3 octave digital filtering, the maximum error is 0.6 dB to either side of the center frequency of each 1/3 octave band.

The power delivered by a source can be determined by linearly integrating the intensity in a number of points around the source. Is the intensity integrated in \( N \) points on a circle with radius \( R \), the delivered power \( P \) is:

\[
P = w'_x + 10 \log_{10} \left( \frac{\sqrt{\varepsilon_m} \Delta r}{\omega \Delta r} \right) + 10 \log_{10} \left( \frac{2\pi R}{1m} \right) - 10 \log_{10} (N)
\]

(18)

Here \( w'_x \) is the level read from the analyzer and \( P \) is in \( \text{dB re } 10^{-12}\text{W} \).

Practical measurements: The vibration intensity in an iron plate, loaded with two viscous dampers, has been measured using the set-up, shown on fig. 2. The iron plate had the following data:

- Length \( L = 800 \text{ mm} \)
- Width \( b = 500 \text{ mm} \)
- Thickness \( h = 1,25 \text{ mm} \)
- Young's modulus \( E = 215 \cdot 10^9 \text{ N/m}^2 \)
- Poisson's ratio \( \mu = 0,3 \)
- Plate stiffness \( B = 38,4 \text{ N/m} \)
- Density \( \rho = 7850 \text{ kg/m}^3 \)
Fig. 2: Set-up for vibration intensity measurements

**Power measurements:** For pure sine tones, with the frequency, $f$, the vibration power $P_e$, delivered by the vibration exciter can be determined by simultaneously measuring the amplitude of the force, $F$, acting on the plate, the resulting acceleration, $a$, and the phase $\phi$ between the force and acceleration, at the point where the vibration energy enters the plate:

$$P_e = \frac{a}{2\pi f} \cdot F \sin \phi$$  \hspace{1cm} (19)

In table 1, the result of measurements made according to equation (19) using a force transducer and an accelerometer are compared to the results obtained by measuring according to equation (18), using the vibration intensity transducer, fig. 1. Furthermore, the sum $P_d$ of the powers absorbed by the two viscous dampers has been measured using the vibration intensity transducer.

<table>
<thead>
<tr>
<th>$f$: Hz</th>
<th>$P_e$: dB re $10^{-12}W$</th>
<th>$P$: dB re $10^{-12}W$</th>
<th>$P_d$: dB re $10^{-12}W$</th>
</tr>
</thead>
<tbody>
<tr>
<td>80</td>
<td>97,2</td>
<td>99,6</td>
<td>99,6</td>
</tr>
<tr>
<td>100</td>
<td>99,6</td>
<td>100,0</td>
<td>99,2</td>
</tr>
<tr>
<td>125</td>
<td>97,8</td>
<td>99,6</td>
<td>100,4</td>
</tr>
<tr>
<td>160</td>
<td>104,0</td>
<td>104,0</td>
<td>101,2</td>
</tr>
<tr>
<td>315</td>
<td>102,9</td>
<td>101,2</td>
<td>99,6</td>
</tr>
<tr>
<td>630</td>
<td>97,1</td>
<td>94,3</td>
<td>91,8</td>
</tr>
</tbody>
</table>

Table 1: $P_e$ and $P = $ power delivered by vibration exciter, $P_d = $ power absorbed by dampers

**Literature:**


A FAST HADAMARD TRANSFORM METHOD FOR THE EVALUATION OF MEASUREMENTS USING PSEUDORANDOM TEST SIGNALS

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High precision measurements of linear system impulse responses are carried out using pseudorandom test signals. In this manner a wide frequency band response with a high signal-to-noise ratio can be obtained in a single and short measurement [1]. The use of pseudorandom noise instead of a single pulse requires digital signal processing to recover the desired impulse response. Because of the properties of the test signal this can be done by very simple methods. In the case of maximum-length sequences an efficient algorithm is possible.

Pseudorandom Noise Sequences

A pseudorandom noise sequence is a periodic sequence \( m_k \) of integers. In the case of binary sequences the \( m_k \) are restricted to two values only, say +1 and -1. The most important property of pseudorandom noise sequences is their two-valued autocorrelation function. For a sequence with period length \( L \) this reads as:

\[
    r_k = \sum_{j=0}^{L-1} m_{j+k} \delta_{k, j+\Delta} = \begin{cases} 
    L & \text{for } k \equiv 0 \mod L; \Delta = \sum_{j=0}^{L-1} m_j \\
    \frac{\Delta^2 - L}{L-1} & \text{else}
    \end{cases}
\]

For many types of binary pseudorandom noise sequences \( \Delta^2 \) has the value one [2]. Examples are (some) Barker-Codes, Legendre sequences and maximum-length sequences. The latter are very simply generated by a digital shift register. Several stages of the shift register (which have values of 0 or 1) are added modulo 2 and feed back to the first register stage. In other words a shift register of \( n \) stages produces a sequence \( a_n \), described by the recursion:

\[
    a_k = \sum_{j=1}^{n} g_j a_{k-j}
\]

where addition is modulo 2 and the maximum-length sequence \( m_k \) is given by \( m_k = -2a_k + 1 \). With a properly chosen set of \( g_i \) the shift register cycles through all binary states (without the all-zeros state) and the sequence has a length of \( L = 2^n - 1 \). The properties of various types of pseudorandom sequences differ if higher order statistics are considered. For example, even two maximum-length sequences of the same length differ, if the third moments of tuple-weight distributions are regarded [3]. These properties come into interest if low-pass filtered or distorted sequences appear (as
it may result from overloading of the system under test).

Impulse response measurements on linear systems use, instead of a single pulse, a periodic sequence of $L$ pulses with signs corresponding to a pseudorandom noise sequence $m_k$. Thus, the energy of the test signal is enhanced by a factor $L$, which results in an improvement of the signal-to-noise ratio by $\sqrt{(L+1)/2}$. If the duration of the single impulse response $h(t)$ is shorter than one period of the sequence, the system response $y(t)$ is given by a circular convolution. For convenience we assume $y(t)$ to be sampled at the clock rate of the sequence. Then the sampled system response $y_k$ is given by

$$y_k = \sum_{j=0}^{L-1} h_j m_{k-j}$$

From the two-valued autocorrelation function it is easy to derive that the impulse response $h_k$ may be computed by a crosscorrelation of the signal $y_k$ and the sequence $\overline{m_k} = m_k + 1$ ($\Delta = -1$ assumed). As all non-zero elements of $\overline{m_k}$ equal 2 this gives a very simple method for the computation of $h$. Figure 1 shows an example for a sequence of length 7. The 'deconvolution' is simply done by adding four shifted versions of the recorded signal. The time delays reflect the positions of the +1 in the pseudorandom sequence $m_k$ shown at the top of Fig. 1. However, this method only works for short sequences because the computing time grows proportional to $L^2$.

Fast Hadamard Transform Algorithm

If a maximum-length sequence is used as test signal the 'deconvolution' of the system response can be done in a much more efficient way. First we write eq. (1) in matrix form,

$$\begin{align*}
\begin{bmatrix}
y_0 \\
y_1 \\
\vdots \\
y_{L-1}
\end{bmatrix}
= M
\begin{bmatrix}
h_0 \\
h_1 \\
\vdots \\
h_{L-1}
\end{bmatrix}
\end{align*}$$

with $y = (y_0, y_1, \ldots, y_{L-1})$; $h = (h_0, h_1, \ldots, h_2, h_1)$; $M_{k} = m_{k+l}$.

$M$ is a left circular matrix of order $L = 2^N - 1$ and all cyclic shifts of the sequence are found in the rows as well as in the columns of $M$. For convenience we assume that $M$ is built from the idempotent sequence $m_k$ (i.e. $m_k = 2^{2k}$ for all $k$). Now $h$ is computed according to

$$h = 1/(L+1) \left( \begin{bmatrix} \overline{y_0} \\ \overline{y_1} \\ \overline{y_{L-1}} \end{bmatrix} \right)$$

From Lempel [4] it is known that the 'M-sequence transform' in eq. (2) can be substituted by a Hadamard transform. With an additional row and column of ones the matrix $M$ reads as

$$\begin{bmatrix}
1 \\
\vdots \\
1
\end{bmatrix}$$

where $P$ is a permutation matrix of order $L+1$. The matrix $H_n$ is a Silvester-type Hadamard matrix. The Hadamard matrix can be constructed recursively via $H_1 = (1)$ and
\[ H_{2r} = \begin{pmatrix} H_r & H_r \\ H_r & -H_r \end{pmatrix} \]

From this construction a fast algorithm for the Hadamard transform is easily derived. This algorithm takes only \( L \cdot \log_2(L) \) operations. Since only add and subtract operations are used it can be run very efficiently even on a microcomputer system.

For the construction of the permutation matrix \( P \) we give a recipe in the following:

1) write the sequence in idempotent order, let (modulo 2):
   \[ a_0 = n \]
   \[ a_k = g_k(k+n) \sum_{j=1}^{n} g_j a_{k-j} \quad k = 1, \ldots, n-1 \]
   \[ a_k = \sum_{j=1}^{n} g_j a_{k-j} \quad k = n, \ldots \]

2) look up \( n \) indices \( \lambda_i \) so that
   \[ a_{\lambda_i} = 1 \]
   \[ a_{\lambda_i + \lambda_j} = \delta_{ij} \]
   (i.e. find a trace-orthogonal basis of \( GF(2^n) \))

3) construct the permutation from
   \[ p_0 = 0 \]
   \[ p_i = \sum_{j=0}^{\lambda_i} 2^j \quad p_{ij} = \delta_{ij} \]

From our experience this method works quite efficient and is easy to program on a computer. As steps 1) and 3) are straightforward the permutation may be computed from the knowledge of only few numbers, the \( g_i \) and \( \lambda_i \).

Now the \( H \)-sequence transform in eq. (2) is reduced to first sorting the vector \((0, y)\) according to the permutation \( P \), the performing a fast Hadamard transform of order \( n \), and finally sorting the resulting vector according to \( F^T \). The final result is the vector \((\gamma, h)\).

Measurements

The method is applied for the measurement of room acoustic impulse responses. A sequence of length \( L = 2^{15} - 1 = 32767 \) is generated by a digital shift register at a clock rate of 25 kHz. This test signal is radiated on the stage of a concert hall and recordings are made at several places. On a second track of the tape recorder the clock signal of the shift register is recorded. Thus it is possible to have perfect commensurate clock rates for the generation of the test signal and the sampling of the system response. Figure 2 shows an impulse response from the Glockensaal in Bremen. The signal-to-noise ratio is about 45 dB and the response has a bandwidth of 10 kHz. Figure 3 shows the impulse response of a loudspeaker recorded in an absorbent environment. The signal-to-noise ratio exceeds 80 dB. A comparison with a 'deconvolution' done by means of FFT signal processing [1] shows an improvement of the signal-to-noise ratio by more than 10 dB.

Figure 3 also exhibits some peculiar peaks, for example at 170 and 310 milliseconds. We found that they are due to a nonlinear distortion of the pseudorandom signal produced by the shift register generator. It is typical for a maximum-length sequence that a nonlinear distortion causes replicas of the impulse response appearing at some distinct delays which are related
to the characteristic polynomial of the sequence. If, for example, a Legendre sequence is used, these errors are at a lower amplitude, but the total energy of the errors is the same in both cases.

![Fig. 2. Impulse response of a concert hall measured with a maximum length sequence. Only 0.7 sec of 1.3 sec are shown](image1)

![Fig. 3. Impulse response of a loudspeaker standing in an absorbent environment. Observe change of vertical scale.](image2)

**Conclusion**

Pseudorandom sequences are well suited for generally all measurements on impulse responses of linear systems. Since the measuring time must be at least as long as the impulse response, it is clear that for optimum signal-to-noise ratio the test signal should last for the same time. Binary pseudorandom sequences are a good choice because they have a flat power spectrum and also a low peak factor, i.e. maximum energy for a given amplitude range.

For measurements using pseudorandom test signals a simple method for the computation of the impulse response has been demonstrated. For maximum-length sequences, which are very simply generated, a fast algorithm for the deconvolution has been explained in detail. As only very basic arithmetic operations are used, this algorithm may be run even on 'dumb' computer systems. As an example for the application of our method measurements of room acoustic impulse responses are presented.


A SIMPLE METHOD TO ANALYZE FAST FLUCTUATING SIGNAL

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1. INTRODUCTION

We usually detect a machine trouble by distinguishing the slight difference of the radiated sound quality (timbre, tone color) by ears. Then it is worth-while to study what way of signal processing can distinguish the difference of the sound quality. The difference sometimes appears in its time-averaged spectral levels and pattern (tonal or random, high-frequency-dominant or low-frequency-dominant, etc.), and sometimes in the time-fluctuating pattern of the spectrum. In many cases, time-averaged levels such as root-mean-square levels of band-passed signals are only monitored, because of its easiness. But this method cannot distinguish the fluctuation of the spectrum.

In this paper, how to distinguish the difference of the time-fluctuating pattern of the spectrum and how to quantify it are studied. And a simple method is found to be useful which quantifies the fluctuating pattern by the magnitude of the level-fluctuation of band-passed signal and by the intensity of its periodic component. Air conditioner noises are studied as examples.

2. AIR CONDITIONER NOISE

Air conditioner noise has various kinds of sound quality shown in Table 1. When pure tones and fast fluctuating sounds are loud and audible, the air conditioner is considered to be a little abnormal. Concerning to pure tones, the ordinary time-averaged power-spectral analysis is enough to distinguish whether they are audible or not. But concerning to the fast fluctuating sounds, it is necessary to analyze the time-fluctuating pattern of the sound spectrum.

They are analyzed by the sound spectrograph as shown in Fig 1. The levels of band-passed signals varying in time are also shown in Fig 1. It is found that the level-fluctuating pattern of band-passed signal varies with the sound quality. But these analyses cannot determine the difference of sound quality quantitatively.
Table 1  Various Kinds of Sound Quality in Air Conditioner Noise

<table>
<thead>
<tr>
<th>Stable Sound</th>
<th>Fast Fluctuating Sound</th>
</tr>
</thead>
<tbody>
<tr>
<td>Broad-Band Random Noise</td>
<td>Pure Tones</td>
</tr>
<tr>
<td>Fan Noise</td>
<td>Singular Sound Fluctuating Rapidly</td>
</tr>
<tr>
<td>Flow Noise etc.</td>
<td>Clatter</td>
</tr>
<tr>
<td></td>
<td>Wind-Breath</td>
</tr>
<tr>
<td></td>
<td>Capillary Tube Noise</td>
</tr>
<tr>
<td></td>
<td>Bearing Noise etc.</td>
</tr>
</tbody>
</table>

(a) Spectrograms

(b) Levels of band-passed signals varying in time

Fig. 1 Analyses of Air Conditioner Noise

3. LEVEL FLUCTUATION ANALYSIS

In order to quantify the characteristics of the fast fluctuating sounds, we tried to get the auto-power spectrum of band-passed and rectified signal which is called Level Fluctuation Spectrum. The procedure of signal processing is shown in Fig. 2. A signal passes through a band-pass filter and a rectifier to be transformed to magnitude signal. And then an auto-power spectrum is calculated. The level fluctuation spectrum is normalized as follows.
\[ L_{fl}(f') = 10 \log \left( \frac{F(|V'|) \cdot F^*(|V'|)}{V^2_{\text{rms}}} \right) \tag{1} \]

Where

- \( L_{fl}(f') \) : level fluctuation spectrum
- \( f' \) : fluctuation frequency
- \(|V'|\) : magnitude signal
- \( V^2_{\text{rms}} \) : long-period mean square of band-passed signal
- \( F(|V'|) \) : Fourier-Transform of \(|V'|\)
- \( F^*(|V'|) \) : complex-conjugate of \( F(|V'|) \)

This method was applied to analyze various test sounds. Fig. 3 shows the examples of the results. It is found that a randomly fluctuating sound such as wind-breath has large component in lower fluctuation frequency, and that a periodically fluctuating sound such as clutter has spectral peaks. Moreover it is found that the louder the fluctuating sound, the higher the fluctuation levels and the peak levels. Where \( g \) denotes the audibility factor of the fluctuating sound obtained by a hearing test. (see Fig. 5)
4. MAGNITUDE OF LEVEL FLUCTUATION AND INTENSITY OF PERIODICITY

It is necessary to find out a physical value which is representative of abnormal sound, in order to apply these methods to detect machine troubles. We defined Magnitude of Level Fluctuation \([\Delta L_{fl}]_R\) and Intensity of Periodicity \([\Delta L_{fl}]_P\) as follows. (see Fig. 4)

\[
[\Delta L_{fl}]_R = 10 \log \frac{\int_{f_{max}}^{f_{max}'} \frac{L_{fl}}{10} df'}{\int_{f_{max}}^{f_{max}'} \frac{L_{fl}}{10} df'}
\]

\([\Delta L_{fl}]_P = L_{flP} - L_{flB}\)

Where \(f_{max}'\) : specified fluctuation frequency

\(L_{fl0}\) : standard level fluctuation spectrum

\(L_{flP}\) : peak level of the spectrum

\(L_{flB}\) : base level of the spectrum

Fig. 4 Definition of \([\Delta L_{fl}]_R\) and \([\Delta L_{fl}]_P\) \(\Delta L_{flR}\) and \([\Delta L_{fl}]_P\) are obtained in each sound frequency band. Therefore, as the representatives of these values, the maximum values \((\Delta L_{fl})_{R_{max}}\) and \((\Delta L_{fl})_{P_{max}}\) are defined. Fig. 5 shows the relation between these values and the audibility factors of the fast fluctuating sounds. It is found that these values are well correlated and that periodically fluctuating sounds are well audible even if their magnitude of level fluctuation is small.

5. CONCLUSION

In order to detect machine troubles by sound, how to distinguish the difference of the radiated sound quality by signal processing was studied. And a simple method was found to be useful which quantifies the time-fluctuating pattern of sound by the magnitude of level fluctuation and by the intensity of its periodic component.

Fig. 5 Relation between \([\Delta L_{fl}]_{R_{max}}, ([\Delta L_{fl}]_{P_{max}}\) and \(g\)

\(.\) : clatter, \(\Delta\) : wind-breath,

\(\times\) : capillary tube noise
EFFETS DE LA DIFFRACTION SUR UNE SONDE INTENSIMÉTRIQUE
POSSIBILITÉS D'OPTIMISATION

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INTRODUCTION
La mesure de l'intensité acoustique par la technique du gradient de pression nécessite un appareillage qui permet le traitement simultané de deux signaux. La précision de la mesure dépend essentiellement de la parfaite maîtrise de la phase entre ces deux signaux dans toute la chaîne de mesure. À cet égard, la sonde (corps compact dans lequel sont intégrés les deux microphones) se trouve dans une situation particulière puisque le déphasage qu'elle introduit est dû à deux mécanismes différents :

- le premier est lié aux réponses en phase des microphones. Le déphasage résultant peut être éliminé dans une mesure, soit par le choix de microphones appariés, soit par une correction ultérieure, [17].
- le deuxième est lié à la perturbation du champ acoustique par la présence de la sonde. Cette perturbation peut être décrite en terme de diffraction dépendant de la nature du champ, des dimensions et de la forme de la sonde et de son orientation dans le champ.

Les sondes intensimétriques présentées ces dernières années par différents laboratoires et fabricants de matériel ont des profils très variés (microphones côte à côte, dos à dos, face à face, ...) mais à part quelques exceptions [1], [2] aucune information sur les effets de la diffraction n'est donnée pour ce matériel. Il nous a donc paru utile de revoir la littérature existante sous l'aspect spécifique du déphasage et d'effectuer des essais complémentaires sur des profils particuliers [3].

ANALYSE BIBLIOGRAPHIQUE
Depuis les travaux de Rayleigh, Lamb, Clebsch et Lorenz un nombre très important de publications est apparu sur les divers aspects de la diffraction (en Optique, Acoustique, Ondes électro-magnétiques) dont on trouve un très bon classement chez Meinel [4]. Les travaux sur la diffraction d'ondes acoustiques apparaissent à partir des années 40 avec notamment les publications de Schwarz [5], Morse [6], Lax et al. [7], Schoch [8] etc ...

Les premières approches consistent à représenter les solutions par des séries de fonctions propres essentiellement pour des corps simples tels que la sphère et le cylindre infini dont les surfaces sont considérées comme rigides ou souples et plus rarement d'impédance complexe.
Une autre approche [9 à 11] utilise des solutions type intégrale soit pour améliorer la convergence des séries, soit pour obtenir des représentations analytiques plus accessibles à l'interprétation des phénomènes physiques, ou encore pour traiter des corps de forme quelconque. (cylindre fini, cube, ...).

La plupart des travaux présente les résultats en terme de pression (total ou diffractée) et/ou de vitesse acoustique soit en champ lointain, soit à la surface du corps et les publications fournissant des renseignements explicites sur l'évolution de la phase sont relativement peu nombreuses, par exemple [11 à 13].

RESULTATS

Parmi les travaux qui ont déjà été effectués en vue d'optimiser des transducteurs acoustiques, [10, 11, 14, 15] on rencontre surtout un grand nombre de calculs numériques et d'analyses expérimentale pour le cylindre court rigide. Cette forme correspond aussi à peu près à la configuration des sondes "dos à dos". La figure 1 [11] montre à titre d'exemple l'augmentation de la pression (par rapport à la configuration sans obstacle) ainsi que l'évolution de la phase sur les faces AV et AR d'un cylindre avec L/D (longueur/diamètre) = 0,25. En haute fréquence, on observe de fortes perturbations qui sont localisées surtout à proximité des bords. Le dépassement entre les centres des faces AV et AR est plus fort qu'en l'absence d'obstacle, phénomène qui est observé dans une moindre mesure aussi sur la sphère.

Une certaine amélioration peut être obtenue par un choix différent du rapport L/D, l'optimum étant d'après nos mesures de 0,5 environ. Nous avons déterminé l'évolution de la phase autour d'un cylindre de cette proportion pour deux valeurs kr (r : rayon) dans la figure 2. Des perturbations - bien que atténuées - subsistent toujours près des bords.

Un lissage supplémentaire des courbes est obtenu en arrondissant les bords du corps diffractant : la figure 3 [10] montre à titre d'illustration les pressions autour d'un cylindre avec L/D = 0,5 et d'une sphéroïde aplatie de même rapport L/D pour laquelle la courbe s'avère nettement plus plate.

Une autre technique pour réduire les effets de bord consiste à reporter sur les faces AV et AR des calottes hémisphériques (et de loger par exemple les microphones dans des interstices sur la partie cylindrique). Les mesures que nous avons effectuées confirment effectivement que la phase mesurée sur une génératrice de ce corps est nettement moins perturbée qu'avec le cylindre à bords vifs : figure 4.

Une forme similaire est donnée par la sphéroïde allongée. Elle a l'avantage de conduire à des solutions qui s'expriment au moyen de fonctions propres pour lesquelles des programmes de calcul sont disponibles [16]. Il est alors aisé d'optimiser d'abord le profil par une simulation numérique (variation de la fréquence, de l'angle d'incidence, etc ...) avant de réaliser le corps définitif de la sonde. Les recherches menées au CETIM en ce sens sont prometteuses et montrent qu'une optimisation est encore possible.
RESUME
La diffraction sur les sondes intensimétriques qui engendre des déphasages peut être minimisée par un choix approprié du profil des sondes. A cet égard le cylindre court nécessite des précautions particulières à cause des effets de bord néfastes. Une forme moins perturbatrice est constituée par le cylindre à calotte hémisphérique ou mieux encore par la sphéroïde allongée. Sur cette dernière, des travaux d'optimisation par simulation numérique devraient apporter encore une amélioration notable.

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M. BOCKHOFF - B. DUPONT - Diffraction sur sonde intensimétrique

**FIGURE 1**: pression et phase sur les faces AV et AR d'un cylindre \( \frac{L}{D} = 0.25 \), [11].

**FIGURE 2**: phase autour d'un cylindre \( \frac{L}{D} = 0.5 \), [3].

**FIGURE 3**: pression pour deux corps différents [10].

**FIGURE 4**: phase le long d'un corps cylindrique arrondi [3].
UTILISATION DE NOUVEAUX "DESCRIPTEURS" DU CHAMP ACOUSTIQUE POUR L'ÉTUDE D'UN TURBOALTERNATEUR. AVANTAGE DE L'INTENSIMÉTRIE PAR F.F.T.

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INTRODUCTION

L'intensimétrie acoustique a été axée jusqu'alors sur la mesure presque exclusive de la seule intensité active $|I|$, qui est utilisée aujourd'hui essentiellement dans 2 domaines :
- localisation de sources sonores,
- détermination de la puissance acoustique, même dans des milieux perturbés

À côté de l'intensité active d'autres "descripteurs" du champ acoustique existent tels que l'intensité réactive, vitesse acoustique, densité d'énergie etc ... qui, à part quelques travaux sur la densité d'énergie [2], ne sont pas encore utilisés couramment. Pourtant leur détermination à l'aide d'intensimètres utilisant la technique F.F.T ne constitue pas un problème majeur à condition de disposer de sondes 3D.
Les résultats obtenus sur un turboalternateur de grande puissance illustrent l'intérêt de ces "descripteurs" pour l'analyse des champs complexes que l'on rencontre à proximité de sources étendues reliées entre elles et fortement couplées sur le plan acoustique et vibratoire.

MESURES

Les mesures ont été effectuées dans une Centrale Nucléaire sur un turboalternateur de 970 MW monté en ligne avec des corps haute et basse pression. L'alternateur (15 x 5 x 5 m) et les turbines sont fixés sur des massifs en béton à 16 m au-dessus du sol. L'émission du bruit est de nature essentiellement bande étroite et se produit à 100 Hz (et aux harmoniques), fréquence à laquelle le pouvoir absorbant des parois est très faible ce qui engendre des ondes stationnaires dans l'ensemble du hall (180 x 54 x 25 m).

Mesures sur un contour
Neuf doublons constitués de micros à condensateur (B & K type 4133, 1/2") avec un écartement de 90 mm ont été fixés à 0,5 m de la machine sur un portique mobile qui a été déplacé par pas de 1 m, ce qui a permis d'enregistrer 126 spectra d'intensité et de pression. Pour commuter les sondes, nous avons utilisé un multiplexeur HP 3497 ; l'analyse et le traitement des signaux ont été effectués sur un analyseur bi-canal HP 3582 relié à un calculateur HP 9835. Tous les résultats (inter-spectres en bandes fines)
ont été stockés pour les dépouillements ultérieurs. Exploration 3D du champ. Une analyse plus fine du champ acoustique a été effectuée dans un plan horizontal proche de l'alternateur - figure 1 - à l'aide de 5 sondes tridimensionnelles montées sur un support léger. Chaque sonde est composée de 4 microphones formant un tétraèdre ($\Delta x = \Delta y = \Delta z = 96$ mm) qui permet de mesurer les trois composantes du vecteur intensité. Avec un maillage de 0,68 m qui assure une résolution suffisamment fine de l'analyse (20% de la longueur d'onde), nous avons exploré une surface de 6 x 9 m correspondant à 130 points de mesure.

**RESULTS**

La figure 2 montre quelques résultats typiques des mesures réalisées sur le contour de l'alternateur. Sur la face supérieure, les vecteurs sont orientés vers l'extérieur mais sur les côtés, ils sont souvent dirigés vers l'alternateur, ce qui signifie un rayonnement important de sources situées à l'extérieur du contour. Le calcul des puissances partielles face par face révèle une émission qui est plus marquée pour la face supérieure que pour les faces latérales dont les puissances restent tout de même positives. Les spectres de la puissance totale sont tracés sur la figure 3.

**Figure 2 :**
Composantes normales de l'intensité à 100 Hz.

**Figure 3 :**
Puissance totale calculée à partir des intensités (traits pleins) et à partir des pressions (pointillés).

Pour la fréquence fondamentale 100 Hz et les harmoniques, l'écart entre les valeurs obtenues à partir de la pression et de l'intensité est de 5 à 7 dB,
ordre de grandeur qu'on rencontre souvent sur des grandes sources. Dans les bandes intermédiaires où l'émission de l'alternateur est négligeable l'écart atteint 13 dB, signifiant une excellente compensation des bruits extérieurs par l'intensimétrie.

Le champ vectoriel de l'intensité active est tracé figure 4. On constate que dans la zone explorée, il n'y a pas d'émission de l'alternateur, par contre on observe un flux d'énergie très marquée en provenance du corps BP. Au-dessus des caillebotis, les directions des vecteurs sont plus aléatoires à cause des contributions multiples de diverses sources réparties aux étages inférieurs.

**Figure 4 : intensité active à 100 Hz**

Les figures suivantes tracées pour la même zone que figure 4 permettent de comparer différentes grandeurs obtenues à partir des mêmes données.

Le champ de pression $p^2/\rho c$ - figure 5 - est caractérisée par des alternances très marquées de zones nodales et ventrales qui ne correspondent pas à des ondes stationnaires élémentaires mais sont le résultat d'interférences plus complexes. Le module $I$ de l'intensité active - figure 6 - montre des variations nettement plus faibles avec une certaine concentration à proximité de l'alternateur. La vitesse acoustique $\rho c \sqrt{2}$ - figure 7 - guère plus fluctuante - est caractérisée par des niveaux nettement plus élevés que la pression, les écarts pouvant atteindre 10 à 15 dB.

L'addition de ces deux grandeurs conduit à la densité d'énergie

$$D = D_{pot} + D_{cin} = \left(\frac{p^2}{\rho c} + \rho c \sqrt{2}\right)/2c$$

**Figure 5 :** Dynamique : 95 / 105 dB

Pression acoustique à 100 Hz

**Figure 6 :** Dynamique : 95 / 105 dB

Module d'intensité active à 100 Hz
Dans une onde stationnaire, l'énergie totale $D$ bascule suivant l'endroit entre l'énergie cinétique $D_{\text{cin}}$ et l'énergie potentielle $D_{\text{pot}}$. Par contre, dans une onde simple (sphérique par exemple), les deux termes de $D$ sont identiques en champ lointain et $D_{\text{pot}} = \rho \left( \frac{1}{k} \right)^2$ prédomine seulement en champ proche : $D_{\text{cin}} / D_{\text{pot}} = 1 + (1/kr)^2$. Dans la figure 9 nous avons représenté $D_{\text{cin}} / D_{\text{pot}}$ et on constate que le champ "proche" d'un grand ensemble peut être très étendu en basse fréquence. Si l'on admet l'analogie avec des sources simples, les zones où $D_{\text{cin}} / D_{\text{pot}} \gg 1$ devraient diminuer lorsque la fréquence augmente. La figure 10, tracée pour 200 Hz, confirme effectivement cette tendance. Par ailleurs, elle laisse apparaître une oscillation de $D_{\text{cin}} / D_{\text{pot}}$ autour de 1 qui est caractéristique pour le champ stationnaire à établissant loin des sources. La grandeur $D_{\text{cin}} / D_{\text{pot}}$ constitue apparemment un descripteur utile du champ proche, même lorsque les sources forment un ensemble assez complexe.
MÉTHODE ÉLECTROSTATIQUE POUR LA CALIBRATION DES INTENSIMÈTRES F.F.T.

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INTRODUCTION

Les performances des intensimètres acoustiques à gradient de pression se trouvent limitées par l'erreur de phase relative entre microphones. Cette erreur provoque une distorsion de la directivité de la sonde qui est mise en évidence par la différence de sensibilité qui apparaît quand elle pivote de 180° dans le champ sonore. L'influence de cette erreur de phase sur l'intensité est d'autant plus importante que la mesure s'effectue dans le bas de la gamme de fréquence.

Plusieurs méthodes, utilisables avec les intensimètres basés sur l'analyse F.F.T., permettent une calibration en phase de l'ensemble de deux voies de mesure. Parmi celles-ci, la méthode de l'excitation électrostatique des microphones présente l'avantage d'être facile à mettre en œuvre et, parce qu'elle ne dépend pas de la qualité d'un champ sonore, d'être applicable sur une gamme de fréquence étendue. Toutefois, une correction complémentaire doit être effectuée sur la fonction de transfert obtenue pour prendre en compte l'influence des trous d'égalisation de pression atmosphérique des microphones.

PRECISION EN PHASE ET BANDE PASSANTE

Il a été montré [1] que la bande passante des intensimètres dépend seulement de l'erreur de phase relative. Ainsi, pour une onde progressive se propageant dans la direction de l'alignement des microphones, l'étendue en octaves de la bande passante à -3 dB correspond au logarithme base 2 de l'inverse de l'incertitude résiduelle sur la phase relative |Δφ| exprimée en radians (voir figure 1 et tableau 1).

Un facteur N, déterminé à partir des figures 3 et 8 de la référence [1], définit les réductions à opérer sur cette bande passante pour des précisions supérieures.
J.C. PASCAL - Calibration des intensimètres

\[ BW(\varepsilon) = \log_2(1/|\Delta\phi|) - N(\varepsilon) - \log_2 K \]

<table>
<thead>
<tr>
<th>(\varepsilon) : précision (dB)</th>
<th>-3</th>
<th>±1</th>
<th>±0,5</th>
<th>±0,25</th>
</tr>
</thead>
<tbody>
<tr>
<td>N (octaves)</td>
<td>0</td>
<td>2</td>
<td>3,4</td>
<td>4,8</td>
</tr>
</tbody>
</table>

Dans des champs sonores plus complexes (réactif, semi-diffus, ...) l'influence de l'erreur de phase se trouve accentuée et un facteur supplémentaire \(\log_2 K\) doit être introduit pour connaître la bande passante effective. Quelques exemples du paramètre \(K\), pour plusieurs types de champs sonores sont fournis dans le tableau I.

**TABLEAU I**

| Onde progressive avec une incidence \(\alpha\) | \(K = \frac{1}{\cos \alpha}\) |
| Teaux d'onde stationnaire | \(K = \frac{1}{(1-\tau)}\) |
| Rapport champ diffus/champ direct | \(K = 1 + R\) |

En considérant que les besoins industriels nécessitent une bande passante à \(±1\) dB de l'ordre de 6 octaves, on déduit des renseignements précédents que l'incertitude sur la phase relative entre les deux voies de mesure ne doit pas dépasser \(2/10\) de degré dans les fréquences basses.

**OBTENTION DES FONCTIONS DE TRANSFERT**

La fonction de transfert sert à corriger l'interspectre mesuré. Elle est obtenue en créant à l'aide d'une paire d'excitateurs électrostatiques une même force en phase sur les membranes des microphones. Le potentiel électrique appliqué est constitué par un bruit blanc superposé à une polarisation continue.

\[ H(f) = \frac{S_{21}(f)}{|S_{21}(f)|} \]

**Figure 2 - Détermination de la fonction de transfert.**

On a observé expérimentalement pour certaines paires de microphones que des écarts de phase persistaient après cette correction. D'autres auteurs \(|2|\) ont fait la même constatation en comparant cette méthode avec la méthode de correction en fond de tube à ondes stationnaires. Cette différence peut s'expliquer par la présence des trous d'égalisation de la pression atmosphérique des microphones qui ont un comportement différent quand le microphone est plongé dans un champ sonore ou quand sa membrane est excitée par un champ électrique. Toutefois, une procédure pour corriger cet effet peut être envisagée.

...
REPRISE EN PHASE DES MICROPHONES A CONDENSATEUR

La bande passante utile des microphones est limitée dans les hautes fréquences par la résonance de la membrane pour laquelle la réponse en phase est de 90°. En dessous, elle est pratiquement contrôlée par la raideur. Le mouvement de la membrane dépend de la différence des pressions Δp entre ses deux faces. Pour que le capteur soit sensible aux fluctuations de pression acoustique, celles-ci doivent être appliquées sur une seule face de la membrane, l'autre étant fermée sur une cavité dont l'impédance acoustique est Zv = ρc²/jωv. La forme de cette impédance permet de faire une analogie électrique avec une capacité Cm = V/ρc². L'inverse de la raideur de la membrane (la complaisance) peut également être représentée par une capacité Cm'. Comme le montre la figure 3, bien en-dessous de la résonance, le microphone dont le volume interne est clos présente une réponse en fréquence uniforme.

Cependant, dans des conditions pratiques de mesure, des perturbations très basses fréquences (< 10 Hz) due à des variations brusques de pression ambiantes risquent de provoquer des saturations du signal de pression acoustique. Il est important de filtrer ces très basses fréquences de réaliser une égalisation de pression entre les deux faces de la membrane du microphone par un tube capillaire. Dans ce tube, les forces d'amortissement visqueux sont prépondérantes devant celles dues à l'inertie ou à la raideur de l'air |3|, ce qui permet de compléter le modèle de la figure 3 par la résistance R. Quand la membrane est seule excitée par une pression électrostatique, la pression acoustique en sortie du tube est nulle (fig. 4A : potentiel p = 0). Quand l'ensemble de la capsule est dans le champ, la sortie du tube est soumise à la même pression acoustique que la membrane (figure 4B : potentiel p). Dans ce dernier cas, la réponse de la capsule microphonique présente une fréquence de coupure f_c qui n'apparaît pas pour une excitation électrostatique et qui provoque une rotation de phase de 90°. La différence de phase d'une réponse par rapport à l'autre peut s'exprimer par (1 + ζ) f_c/f.

CORRECTION DE LA FONCTION DE TRANSFERT

L'écart de phase entre les deux microphones qui n'est pas pris en compte par la calibration électrostatique est fonction de la différence des fréquences de coupure des capsules :

\[ \Delta \phi_{21} = (1 + \zeta) \frac{f_{02} - f_{01}}{f} \]

mais aussi du rapport ζ de la raideur ajoutée par l'air de la cavité sur la raideur de la membrane (normalement compris entre 10 et 20 Ω). Pour les microphones utilisés, le fabricant garantit des fréquences de coupure comprises entre 1 et 2 Hz. Dans ces conditions, à 100 Hz, limite basse fréquence généralement choisie pour l'intensimètre, il peut se pro-
duire un écart maximal non-corrigé de l'ordre de 0,65°, bien supérieur à l'incertitude admise (< 0,2°).

Pour réaliser une correction de la fonction de transfert obtenue selon la procédure illustrée par la figure 2, on doit déterminer la fréquence de coupure des capsules microphoniques à l'aide d'un système de calibration basse fréquence [4].

\[
\text{fonction de transfert corrigée: } H_{\text{cor}}(f) = H(f) \exp \left[ j(1 + \xi) \frac{f_{c1} - f_{c2}}{f} \right]
\]

On conserve une précision suffisante en prenant \( \xi = 0,15 \). Il est encore possible de le déterminer à partir de l'accroissement de sensibilité \( \Delta L \) de la courbe de réponse à une excitation électrostatique pour la fréquence de coupure \( f_c \) (voir figure 4).

Autrement, la correction supplémentaire peut être évitée en choisissant des capsules microphoniques ayant des fréquences de coupure très voisines (pour \( \Delta \phi < 0,1^\circ \) à 100 Hz, il faut \( \Delta f_c < 0,15 \) Hz).

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UNE NOUVELLE CONCEPTION DES INSTRUMENTS DE MESURE D'ENVIRONNEMENT

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I - EVOLUTION DES METHODES

Les méthodes de mesures utilisées dans le contrôle de l'environnement ont depuis quelques années déjà, abandonné la représentation du phénomène acoustique par des valeurs quasi instantanées (niveau de pression acoustique acoustique) pour ne garder comme grandeur représentative, que les unités obtenues après réduction de la quantité d'information contenues dans l'évolution de $L(t)$.

Parmi tous les moyens utilisables pour réduire cette quantité d'information, un consensus s'est établi autour du niveau continu équivalent (Leq) obtenu sur des périodes temporelles qui restent caractéristiques du secteur d'activité.

C'est ainsi, que dans le monde du travail, on calcule le Leq sur la journée de 8 heures ou la semaine de 40 heures, alors que les études des nuisances engendrées par les bruits de circulation font référence au Leq diurne 8h/20h, ou au Leq nocturne 22h/6h.

La mise en application de ces nouvelles unités a semblé des l'origine satisfaisantes et des corrélations ont pu être établies entre les effets et les causes.

Cependant, dans un grand nombre de situations on a rapidement remarqué que la réduction d'information effectuée était trop forte et que la représentation par un nombre unique d'un phénomène acoustique fluctuant, étudié sur une durée importante, pouvait être dénuée de sens et conduire à des conclusions contradictoires sur les effets prévus.

Aussi, a-t'on assisté, ces dernières années à un retour en arrière dans les méthodes d'exploration de l'environnement acoustique. Une des dernières démarches d'ordre technologique a consisté à améliorer la collecte des informations traditionnelles L. A l'aide d'une instrumentation automatique, à grande dynamique, on a pu remplacer l'opérateur humain pour recueillir l'information horodée ($L(t)$).

L'avantage d'un tel système réside dans la mise en mémoire d'un phénomène avec une bonne résolution ce qui permet à posteriori d'ajuster un traitement aux caractéristiques statistiques du signal par exemple, et dans tous les cas, de pouvoir "rajouter la partie"et rechercher les informations qu'un traitement arbitraire aurait pu masquer.
C. AZAIS: Une nouvelle conception des instruments de mesure d'environnement

Une attitude récente consiste à effectuer dès sa collecte un prétraitement de l'information par intégration énergétique sur des durées de 10s, 20s, 10mn par exemple. Ceci conduit à la technique dite des "Leq courts". Cette méthode qui va partiellement à l'encontre des avantages exposés précédemment, permet de faire face - à capacité de mémoire donnée - à des durées de mesure plus importantes ou bien à la collecte d'informations recueillies simultanément à plusieurs points.

La technologie moderne permet de réaliser ces appareils avec de fortes capacités de mémoire qui leur permettent de suivre les phénomènes acoustiques pendant des durées importantes. Cependant, il n'est pas encore possible de les réduire à des formats portables permettant à l'utilisateur d'exploiter simplement sur le terrain les informations obtenues.

La multiplication des mesures d'environnement a montré la nécessité de disposer d'un appareil fournissant en plus d'une grande intégrée, une caractéristique concernant la variabilité du phénomène.

C'est sur ce problème que notre Laboratoire s'est penché en travaillant ici en étroite collaboration avec des spécialistes du traitement acoustique des bruits industriels.

Pour les raisons technologiques évoquées plus haut, il ne nous a pas paru possible de conserver dans l'appareillage proposé l'information "date" que nous avons abandonnée au bénéfice de l'information "durée" prise sous la forme de l'histogramme des niveaux de pression acoustique.

II - APPORT DE L'HISTOGRAMME DANS LE TRAITEMENT DES MESURES DE BRUIT.

Nous examinerons ici quelques exemples types d'exploitation de résultats relevés dans le monde du travail. Le but poursuivi dans cette étude est d'examiner les situations rencontrées dans un atelier de chaudronnerie lourde et de proposer des stratégies d'amélioration de situation.
II.1 - Poste moulage

L'histogramme relevé à ce poste (fig. 1) est de forme régulière. Le Leq mesuré se situe dans la partie supérieure de la classe la plus probable, l'apport d'information apporté par l'histogramme est faible. Dans ce cas particulier, le Leq est une grandeur suffisamment pertinente pour caractériser la situation.

II.2 - Zone centrale

L'histogramme observé ici (Fig. 2) apporte une information très importante. On remarquera en effet, que le Leq s'écarte de la classe de niveau la plus probable. Pour étudier l'incidence de cette dispersion, nous tracerons le tableau donnant l'évolution du Leq lorsque l'histogramme des niveaux s'étend des classes de faibles niveaux vers des classes de niveaux de plus en plus élevés, par exemple Leq/7 représente le Leq du phénomène dans l'hypothèse où tout l'histogramme est compris entre les classes 80 et 85. On peut ainsi observer l'évolution du Leq lorsque l'on effectue un traitement (idéal) éliminant les niveaux de pression acoustique les plus élevés.

<table>
<thead>
<tr>
<th>Classes</th>
<th>80</th>
<th>85</th>
<th>90</th>
<th>95</th>
<th>100</th>
<th>105</th>
<th>110</th>
<th>115</th>
<th>120</th>
</tr>
</thead>
<tbody>
<tr>
<td>%</td>
<td>4</td>
<td>15</td>
<td>46</td>
<td>17,5</td>
<td>6,5</td>
<td>5,5</td>
<td>5</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>Leq/7</td>
<td>80</td>
<td>85,4</td>
<td>90,2</td>
<td>92,3</td>
<td>94,2</td>
<td>96,4</td>
<td>98,7</td>
<td>99,7</td>
<td>99,7</td>
</tr>
</tbody>
</table>

L'examen du tableau nous montre que le traitement des sources bruyantes, classes 110 et 115 de taux d'occupation très faible, pourrait abaisser le Leq. Cependant, l'examen de l'histogramme montre que le niveau sonore est pour environ 50 % du temps de l'ordre de 90 dB. On peut craindre alors que les sources très bruyantes (de durée d'action brève) soient psychologiquement ignorées des occupants de l'atelier (ceci a été vérifié par démarche participative).

Dans ce cas, une réduction du bruit sur les sources les plus bruyantes, sera certes bénéfique à la protection auditive des occupants, mais ne sera pas perçue par ceux-ci comme une amélioration ; alors qu'un traitement acoustique général qui diminuerait de quelques dB les niveaux sonores des sources les plus fréquentes sans toucher aux sources les plus bruyantes ne fera pas diminuer le niveau équivalent, mais donnera aux occupants une sensation d'amélioration.

II.3 - Atelier de rectification

On observe sur la figure 3, un histogramme largement distribué avec une émergence notable dans une classe de niveau élevé, l'origine du bruit est ici l'opération de "rectification". Il est manifeste que dans ce cas, le Leq rend mal compte de la situation de cet atelier qui est en réalité, durant 80 % du temps un atelier "presque silencieux". Ici, un traitement (amélioration de l'outillage) s'impose autour de la source de bruit incriminée.
III - APPAREILLAGE

L'instrument utilisé pour ces expériences, est un produit issu des recherches du Laboratoire et fabriqué par une société Française. C'est un appareil portable de volume réduit, fournissant le niveau continu équivalent et les résultats de l'histogramme par appel sur des touches spécialisées d'un clavier (1 touche par grandeur ou classe). Il fournit en outre, le niveau de pression acoustique pondéré A et C (mesures simultanées) ainsi que les maxima de ces grandeurs.

Le domaine de mesure est divisé en deux gammes 40-95 dB et 75-130 dB, se dynamique utile est donc de 55 dB. Cependant, l'organisation du classement des niveaux sonores permet d'explorer des phénomènes évoluant sur des dynamiques supérieures.

Ce classement est effectué dans 8 classes, parmi celles-ci, 7 ont une largeur égale à 5 dB, alors que les classes situées respectivement aux limites inférieure et supérieure sont ouvertes. On peut ainsi en faire travailler successivement l'appareil aux bas niveaux et aux forts niveaux vérifier la stationnarité du phénomène en faisant plusieurs expériences et, lorsque cette hypothèse est vérifiée, composer les résultats pour obtenir un histogramme étendu, comme celui exposé Fig. 3.

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ÉTUDE D’UN GÉNÉRATEUR NUMÉRIQUE DE SIGNAUX ALEATOIRES SIMULANT LE PHÉNOMÈNE PRESSION ACOUSTIQUE EN VUE DU TEST DES LEQ-METRES.

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INTRODUCTION
Le but que nous nous proposons d’atteindre peut se résumer en ceci : être capable de reproduire en laboratoire un signal pression acoustique dans le cas de bruits d’environnement. Ce signal bien que considéré comme étant aléatoire, doit être reproductible avec des caractéristiques probabilistes données et connues mais, surtout, toujours identiques, même si le phénomène aléatoire considéré n’est que localement stationnaire. De plus ce signal doit être similaire à certains types de bruit d’environnement tant en probabilité qu’en caractéristiques spectrales.

Deux études ont été entreprises au L.A.M.I, dans ce but, La première consistait, bien sûr, à déterminer par une Analyse statistique et spectrale les caractéristiques d’un certain nombre de bruits d’environnement réels cf. publi. de MM. CRASMUCK et GUILHOT, session n° 6.

L’autre devrait envisager les moyens de produire un signal électrique analogique aléatoire de caractéristiques statistiques et spectrales imposées mais surtout toujours identiques, même dans le cadre de la stationnarité locale.

Les deux études devaient concourir, in fine, à la réalisation d’un générateur de signal aléatoire de laboratoire de caractéristique conforme à celle de divers signaux "bruits d’environnement" réels.

CHOIX DE LA MÉTHODE
Après avoir étudié les diverses méthodes possibles de génération de signaux électriques aléatoires, nous avons choisi d’adopter une méthode de génération numérique d’un processus aléatoire, quitte à opérer une conversion numérique-analogique pour obtenir le signal électrique final. La raison de ce choix réside en la souplesse du procédé permettant de programmer n’importe quelle caractéristique probabiliste ou spectrale du signal de sortie. Cela était d’autant plus important que cette programmation ne pouvait être faite qu’après obtention du résultat de l’étude d’analyse et de modélisation menée en parallèle. De plus cela rend possible à tout instant et à postériori d’ajouter des signaux aléatoires divers correspondant à d’autres types de bruit d’environnement.
PRINCIPE DE LA MÉTHODE
Soit une variable aléatoire équiprobable y indépendante en probabilité
\[ P_y(y) = k = \text{cte} \]
pour obtenir une variable aléatoire x ayant une loi de probabilité \( P_x(x) \)
il faut que
\[ \frac{dy}{dx} = \frac{P_x(x)}{P_y(y)} = \frac{1}{k} P_x(x) \]
où en intégrant
\[ \gamma(x) = \frac{1}{k} \int_{-\infty}^{x} P_x(u) du \]
\[ \gamma(x) = \frac{1}{k} \cdot F_x(x) \]
où \( F_x(x) \) est la fonction de répartition de la loi de probabilité recherchée.
On peut donc calculer x en fonction de y
\[ x(y) = F^{-1}_x (k y) \]
si y est indépendante en probabilité, x le sera aussi.

Il est donc possible à partir d'une table de variables aléatoires équiprobables indépendantes y de calculer au moyen d'un tableau de la fonction inverse de répartition d'une loi de probabilité \( P_x(x) \) des variables aléatoires x ayant cette loi de probabilité. Notons que l'on peut aussi utiliser un algorithme de calcul en place des tableaux. Le choix entre ces deux moyens étant édicté par le souci d'obtenir une cadence de sortie des échantillons relativement élevée.
Remarquons également que l'on peut obtenir des variables aléatoires quasi-gaussiennes et ayant un coefficient d'autocorrélation non nul par combinaison linéaire de variables successives y.

CASY DU BRUIT NON STATIONNAIRE
L'étude sus-évoquée a montré que l'on pouvait considérer le phénomène pression acoustique dans le cas de bruit de circulation urbaine comme étant de la forme.
\[ P(t) = \left[ 1 + 3(t) \right] \cdot x(t) \]
on \( x(t) \) est un phénomène aléatoire gaussien stationnaire centré blanc jusqu'à 200 Hz. Modulé par un phénomène aléatoire \( 1 + z(t) \) non centré gaussien stationnaire blanc jusqu'à 0,01 Hz suivi d'une décroissance du 1er ordre.
Le fait que le phénomène modulant \( 1 + z(t) \) soit lentement variable appelle, à considérer le phénomène \( p(t) \) comme localement stationnaire et à le synthétiser par multiplication analogique après obtention des deux signaux \( x(t) \) et \( 1 + z(t) \) cf. schéma. Un signal stationnaire peut être obtenu, bien sûr, par commande de programmation (entrée C) du générateur en faisant \( z(t) = 0 \).

FORMATION DES SPECTRES
Les filtrages requis pour les deux phénomènes aléatoires peuvent être obtenus de deux manières différentes, on peut introduire des filtres analogiques aux points A et B. L'on peut aussi opérer un filtrage numérique par des modules de programme, sur les échantillons fournis par le système.
E. BARANI  Générateur numérique de signaux aléatoires acoustiques.
CONCLUSION
Que l'on ne s'y méprenne pas, l'apparence informatique de cette description résulte uniquement de l'outil utilisé, le microprocesseur. Aussi bien dans ces apparences physiques que dans son utilisation, ce générateur de signaux acoustiques aléatoires contrepant des signaux réels n'a rien d'un ordinateur. Il comporte des commutateurs (bruit de circulation, bruit de moteur électrique, etc...) des attenuateurs etc... qui lui donne l'aspect d'un appareil de mesure classique. Cependant, son aptitude à être commandé électriquement par le canal C lui permet de s'intégrer à un système automatisé de test pour instrument de mesure acoustique.

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ETUDE STATISTIQUE DE SIGNAUX ACOUSTIQUES REELS EN VUE DE LA SYNTHESE DE SIGNAUX TYPES

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Introduction
Notre étude se propose de chercher des paramètres caractéristiques des bruits réels et ce avec une double finalité :
- on a besoin de connaître ces paramètres si l'on recherche une corrélation avec la sensation, la gêne ou le risque de perte d'audition provoquée par ces bruits,
- les appareils de mesure étant imparfaits, l'erreur commise en les utilisant dépend des propriétés des signaux étudiés. Il peut par exemple être utile de chiffrer l'erreur due à l'utilisation d'un sonomètre intégrateur dans le cas de signaux correspondant à des utilisations-types.

Nous allons tenter de définir les propriétés statistiques de deux types de bruits usuels :
- bruits stationnaires (bruit d'atelier)
- bruit évolutif (bruit de circulation urbaine)

Nous en déduirons une méthode de synthèse nous permettant de disposer de signaux-types dans la plupart des utilisations prévues.

1. Principe de l'étude
1.1. Bruits stationnaires
Les bruits stationnaires peuvent être caractérisés
- au premier ordre : par leur densité de probabilité
- au second ordre par leur densité spectrale de puissance
Un signal acoustique peut être considéré comme une réalisation d'un processus aléatoire stationnaire et ergodique à valeur moyenne nulle.

1.2. Bruits évolutifs
On peut considérer une tranche temporelle de bruit évolutif comme une réalisation d'un processus aléatoire non stationnaire. Cette non stationnarité porte forcément sur les moments du second ordre, la valeur moyenne du signal pression acoustique étant nulle.
Cette non-stationnarité peut porter sur la valeur efficace (signal modulé en amplitude) ou sur la densité spectrale de puissance. Pour simplifier l'étude nous avons fait l'hypothèse d'une non stationnarité de la valeur efficace.
L'étude des signaux non stationnaires n'est possible que dans deux cas très particuliers:
- le signal est reproductible. On dispose alors d'un ensemble de réalisations sur lequel il est possible d'effectuer une étude statistique. Ceci suppose la connaissance de l'origine des temps pour chacun des enregistrements.
- le signal est localement stationnaire : c'est le cas de signaux modulés lentement en amplitude. Il est alors possible de séparer les deux composantes du signal. Un seul enregistrement peut alors servir pour l'étude. C'est cette hypothèse que nous allons utiliser ici.

Considérons une réalisation quelconque du signal \( \{ p(t) \} \) comme le résultat d'une modulation où porteuse et signal modulant seraient remplacés par des phénomènes aléatoires stationnaires, gaussiens, centrés :

\[
\{ p(t) \} = (1 + \{ x(t) \}) \cdot \{ y(t) \}
\]

où les densités spectrales de puissance de \( \{ x(t) \} \) et \( \{ y(t) \} \) sont disjointes.

On réalise un filtrage du processus \( \{ p(t) \} \) par calcul de valeurs quadratiques moyennes sur des tranches de temps \( T \) telles que l'estimateur de la variance de \( \{ y(t) \} \) soit convergent vers la vraie valeur \( \sigma_y^2 \) alors que le modulant \( \{ x(t) \} \) est supposé constant.

On peut écrire que

\[
\frac{p(t)^2}{T} = \sigma_y^2 \cdot (1 + x(t))^2
\]

\[
\frac{p(t)}{\sigma_y} = \frac{1}{\sigma_y} \cdot \sigma_x^2 + 1
\]

\[
\text{var} \left[ \frac{p(t)}{\sigma_y} \right] = 2 \cdot \sigma_y^4 \left[ (1 + \sigma_x^2)^2 - 1 \right]
\]

\[
\text{var} \left[ \frac{\sigma^2(t)}{\sigma_y^2} \right] / \sigma^2 = 2 \cdot \left[ 1 - 1 / \left( 1 + \sigma_x^2 \right)^2 \right]
\]

Cette relation nous permet d'accéder à la variance du signal modulant et donc de monter à la variance du signal modulé.

Connaisant cette variance il est possible de normaliser le phénomène global sur des tranches temporelles de valeur \( T \):

\[
p(t) / \left[ \left( \frac{p(t)}{\sigma_y} \right)^T \right]^{1/2}
\]

De cette dernière expression on peut déduire les propriétés du signal normal modulé.

2. Étude de quelques bruits types

Le signal pression acoustique était enregistré in-situ sur bande magnétique puis numérisé et stocké sur une mémoire de masse (disquette magnétique). L'exploitation des données se faisait en temps différé à partir de cette mémoire de masse.

2.1. Bruit stationnaire

On a étudié un bruit d'atelier, stable, sans contenu impulsionnel. La densité de probabilité de la pression est normale et son spectre plat jusqu'à une fréquence de 3 kHz ce qui correspond à un bruit blanc.
filtré passe-bas.

2.2. Bruit non stationnaire
Il s'agit d'un bruit fluctuant : bruit de circulation urbaine dense, fluide, pulsée.
L'étude de la variance de la pression quadratique moyenne calculée sur une durée \( T \) fait apparaître une zone d'invariance qui correspond à la durée de stationnarité locale (cf. figure ci-dessous). Cette durée est comprise entre 200 et 400 ms.

\[
\text{Ceci vérifie bien l'hypothèse du signal localement stationnaire.}
\]
On vérifie que les deux processus pouvaient être considérés comme normaux et de spectre blanc filtré par un filtre passe bas du premier ordre de fréquence de coupure égale à 200 Hz pour le signal modulé et 10 mHz pour le signal modulant.

**CONCLUSION**

Notre étude nous a montré qu'il était possible, par méthodes numériques, de stocker et de traiter un grand nombre d'échantillons du signal pression.
Des hypothèses statistiques sur ces signaux nous permettent d'obtenir un modèle satisfaisant. La connaissance expérimentale des paramètres fondamentaux permet d'envisager la synthèse de signaux types, de caractéristiques bien connues utilisables à des fins météorologiques.

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TRAITEMENT RAPIDE EN TEMPS DIFFEREE DES MESURES D'ENVIRONNEMENT ACOUSTIQUE

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FINALITE : LUTTE CONTRES LES NUISANCES ACOUSTIQUES

Une des méthodes possibles pour analyser l'environnement acoustique consiste à stocker en mémoire la suite temporelle des mesures de niveaux de bruit au moyen d'un système d'acquisition implanté temporairement in situ, puis à relever régulièrement cette moisson de mesures pour la traiter, ensuite, au laboratoire et enfin, l'archiver aux fins de validation légale des résultats obtenus ou de traitements ultérieurs.
Cette méthode présente l'avantage de permettre l'optimisation du traitement nécessaire pour affiner correctement l'intervalle de confiance des estimateurs du phénomène aléatoire "Niveau de bruit" que constituent les indices de bruit.

LES MOYENS TECHNOLOGIQUES

Sans prétendre être exhaustif, examinons les qualités et les défauts des différentes sortes de mémoire utilisables pour effectuer le stockage des données in situ.

Les mémoires vives en technologie CMOS rendues non volatiles à court terme par adjonction d'une pile de sauvegarde, ont un temps d'accès très court mais sont onéreuses et surtout volatiles à long terme, donc difficilement archivables.

Les mémoires à bulles magnétiques possèdent un temps d'accès encore acceptable pour ne pas trop pénaliser la rapidité de traitement en laboratoire, mais leur prix actuel ne permet d'envisager leur archivage.

Les mémoires magnétiques longues à déplacement cinématique : bandes magnétiques, cassettes ou cartouches numériques sont séquentielles par essence même et de ce fait bien adaptées à l'enregistrement de mesures à cadence fixe. Par corollaire, leur caractère séquentiel n'autorise des temps d'accès moyens compatibles avec la vitesse requise de traitement au laboratoire que si l'on effectue la recherche des données en défilement rapide, ce qui est préjudiciable à la longévité du support et par suite, de l'information.

Les disques magnétiques souples par leur structure superficielle par
opposition à la structure linéique des précédents supports, permettent un accès semi indexé, donc plus rapide.
Il en résulte des temps de recherche plus courts et par suite des temps moyens d'accès très en rapport avec les temps de traitement, sans préjudice pour la sécurité de l'information. Leur coût rendrait l'archivage direct parfaitement possible, malheureusement, l'astreinte d'autonomie des systèmes d'acquisition rend le stockage direct in situ sur ce support inenvisageable en raison de la consommation électrique élevée des cinématiques et surtout parce que la rotation du disque à vitesse angulaire constante est obtenue au moyen de moteurs électriques synchrones au secteur électrique. Il n'est, cependant, pas interdit d'espérer que les progrès de la technique rendent possible le stockage direct sur disque souple.

LA SOLUTION ACTUELLE

Le processus actuel d'acquisition-analyse est parfaitement compatible avec cette éventualité.
Le scénario est le suivant :
Quelque soit le support choisi pour enregistrer les mesures in situ, l'information est relue, convertie et reportée sur disque souple. Une copie de sauvegarde du disque est prise sur une autre disquette et archivée.
Le traitement immédiat est, alors, entrepris en utilisant toujours la disquette de travail. La disquette d'archive n'est employée que pour refaire une disquette de travail en cas de dégradation de celle-ci.

LES ETAPES DE L'EXPERIMENTATION

Un premier procédé consistant à enregistrer in situ sur cassette numérique puis à traiter au moyen d'un calculateur scientifique, a fait ressortir la nécessité d'atteindre à une plus grande rapidité de traitement.
La lenteur du procédé était due à deux causes distinctes : l'excès de précision du calculateur et le temps d'accès moyen inhérent à l'utilisation de la cassette comme support d'information strictement séquentiel, lors du traitement.

La nécessité constatée d'obtenir une plus grande rapidité de traitement nous a conduit à étudier, à réaliser et à expérimenter un deuxième système de traitement.
Il comprenait un calculateur à vocation essentiellement acoustique bâti autour d'un microprocesseur à huit bits et un système de recherche rapide de l'information sur la cassette numérique par défilement avant et arrière rapide. Les programmes de calcul étaient implantés en mémoire morte. Nous obtenions ainsi une vitesse de traitement parfaitement acceptable, mais les à-coups que faisait subir à la bande le dispositif de recherche rapide diminuaient très vite sa fiabilité.
Telles sont les raisons qui nous ont conduit à adopter la méthode actuelle décrite sommairement plus haut.

DESCRIPTION DU SYSTÈME ACTUEL

Il est constitué par un mini-ordinateur dit "EXORCISER" dont la vocation originelle est le développement de microsystèmes, entouré d'un certain nombre de périphériques :
-console opérateur acousticien (clavier et visualisation sur tube cathodique)
-lecteur de cassettes numériques
-enregistrateur lecteur de disques souples (2 cinématiques en simple face et simple densité)
-imprimante rapide par ligne.

Le rôle de ce système étant double, nous avons développé, en fait, deux programmes très différents par leur taille.

Un premier programme a pour rôle de relire et de vérifier le contenu de la cassette. Il transcode, ensuite, en binaire le contenu de la cassette qui est codé en code ASCII (NF Z-62-010) comme l'impose la norme AFNOR NF Z-64-141 pour l'enregistrement des cassettes numériques.
Enfin, il crée un fichier sur disque qui contient environ une journée de mesure, ainsi qu'un secteur d'identification comprenant tous les paramètres physiques et métrologiques des mesures de cette cassette.
Notons que l'horodatage séquentiel des cassettes est reporté sur la disquette.
La recopie en sauvegarde de ce fichier est réalisé par un programme utilitaire du système original.

Le deuxième programme, de très loin le plus volumineux, a pour rôle global le traitement de l'information reportée sur disquette.
Nous l'avons baptisé COMPAC (pour computer acoustique). Il serait hors de propos dans le cadre de ces quelques lignes, de prétendre en faire l'analyse complète.
Nous allons essayer d'en donner les caractéristiques les plus importantes.
L'optique primordiale qui a présidé à son élaboration se résume en un seul mot: interactivité.
En effet, à tout moment, l'opérateur doit pouvoir, au vu des résultats partiels déjà acquis, suspendre l'exécution du traitement en cours pour:
-accéder à des renseignements complémentaires
-revoir sur l'écran des résultats partiels antérieurs
-obtenir un certain nombre de courbes sur l'imprimante
-modifier certaines consignes du traitement en cours
-reprendre le traitement en cours
ou
-interrompre définitivement ce traitement
pour
-commencer un traitement d'analyse différente.

COMPAC, en son état actuel, permet les traitements d'analyse suivants, par tranches temporelles paramétrables, depuis une date donnée jusqu'à une autre date donnée:
-analyse statistique des niveaux de bruit
-tracé des histogrammes par tranche temporelle
-tracé des probabilités cumulées par tranche
-calcul de l'évolution du LQ
-calcul de l'évolution des Lx
-tracé de l'évolution des ces indices
-etc.....
E. BARANI

Traitement rapide en temps différé des mesures d'environnement acoustique.

COMPAC permet, en sus, de voir les caractéristiques métrologiques de l'acquisition, de tracer simplement l'évolution temporelle du niveau de bruit dans une tranche de temps et de trouver les périodes d'étalonnage du système d'acquisition.

Il est à noter que l'utilisation de la programmation modulaire structurée qui a été scrupuleusement respectée tout au long de l'étude autorise l'adjonction de modules supplémentaires par superposition avec les modules existant en mémoire, ce qui présente l'intérêt de ne pas faire perdre les résultats intermédiaires déjà acquis. Ajoutons, pour les informaticiens, que le langage utilisé a été le PL1 pour le programme principal et plus rarement l'assembler pour certains modules.

CONCLUSION

Nous avons essayé, tout au long de cette étude, de tendre vers un certain pôle : permettre au physicien-acousticien d'oublier l'outil informatique et essayer de lui donner la possibilité de faire jouer la partie "Évaluation de l'ambiance acoustique d'une journée" en un temps largement plus court que le temps réel, cela autant de fois qu'il le désire et à la discrétion de son analyse.

Nous ne saurions, étant juge et partie, dire dans quelle "MESURE" nous avons approché cette asymptote.

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Système de traitement numérique rapide en temps différé de mesures acoustiques d'environnement (Doctorat-
STATISTICAL METHODS OF NOISE ASSESSMENT

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Introduction

When a number of noise levels of a varying noise situation are to be represented by a single number the question arises whether to calculate the mean of the noise levels or the mean of their corresponding sound intensities. Moreover it is often required to give limits for an interval within which the mean value is expected with a certain probability. It is our object here to discuss this problem using parametric and nonparametric statistical methods. Nonparametric methods - on the contrary to parametric ones - do not assume that the varying values - either levels or intensities - follow a certain distribution.

Mean Values

It has been noted that a lot of variates do not follow a normal distribution. An obvious explanation for that is often found to be the inability of the variate to take values under or over a certain limit. So the variability is inhibited in that direction. Specially when the distribution is limited to positive values or zero a transformation of the original variates to their logarithms is often a suitable method to obtain approximately a normal distribution. This well known fact from statistics explains why the distribution of sound levels rather than their corresponding intensities usually follows a normal one. This has been a temptation to carry the statistics on the transformed variate (sound level) rather than on the original variate (sound intensity).

But as $L_{eq}$ has now been almost accepted on international basis to represent - as a single number - a varying noise situation, it is the arithmetic mean of the sound intensities and not that of the sound levels which is to be calculated. In fact there is a certain relation between $L_{eq}$ and the arithmetic mean of the noise levels $L$ which is rigorously true for normally distributed sound levels. This relation is given in equation 1.
\[ L_{eq} = \bar{L} + 0.115 s_L^2 \]

\( s_L \) is the standard deviation of the noise levels.

It might be interesting for those who possess measured data of various noise situations to try to compare calculated \( L_{eq} \) from equation 1 with the true \( L_{eq} \). The following three distributions of noise levels are given in Table 1 as an example for the application of equation 1. Distribution Nr. 1 is approximately normal, nr. 2 has a positive skew and nr. 3 has a negative skew.

**Table 1** Distributions of noise levels

<table>
<thead>
<tr>
<th>Noise level (db)</th>
<th>45</th>
<th>50</th>
<th>55</th>
<th>60</th>
<th>65</th>
<th>70</th>
<th>75</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dist. nr.</td>
<td>Frequency</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>1</td>
<td>4</td>
<td>54</td>
<td>242</td>
<td>399</td>
<td>242</td>
<td>54</td>
<td>4</td>
</tr>
<tr>
<td>2</td>
<td>4</td>
<td>399</td>
<td>300</td>
<td>180</td>
<td>80</td>
<td>20</td>
<td>4</td>
</tr>
<tr>
<td>3</td>
<td>4</td>
<td>20</td>
<td>80</td>
<td>180</td>
<td>300</td>
<td>399</td>
<td>4</td>
</tr>
</tbody>
</table>

The values for \( \bar{L} \), true \( L_{eq} \), \( L_{eq} \) calculated from equation 1 and the difference \( D = L_{eq} \) (eq. 1) - true \( L_{eq} \) are given for the above 3 distributions in Table 2.

**Table 2** Mean values for the distributions in Table 1

<table>
<thead>
<tr>
<th>Dist. nr.</th>
<th>( \bar{L} )</th>
<th>true ( L_{eq} )</th>
<th>( L_{eq} ) (eq. 1)</th>
<th>( D )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>60</td>
<td>62.82</td>
<td>62.85</td>
<td>0.03</td>
</tr>
<tr>
<td>2</td>
<td>55</td>
<td>59.57</td>
<td>58.43</td>
<td>-1.14</td>
</tr>
<tr>
<td>3</td>
<td>65</td>
<td>67.28</td>
<td>60.34</td>
<td>1.06</td>
</tr>
</tbody>
</table>

A general rule of statistics which is always valid irrespective of distribution is that the arithmetic mean of a sample is greater than or equal to its geometric mean. The equality holds only if all sample values are identical. Since \( L_{eq} \) corresponds to the arithmetic mean of the intensities whereas \( \bar{L} \) corresponds to the geometric mean of the intensities it is always true that \( L_{eq} \geq \bar{L} \).

The proof is simple and obvious from the following equations.

Arithmetic mean = \( \bar{x} = \frac{1}{n} \sum_{i=1}^{n} x_i \)

Geometric mean = \( GM = \left( \prod_{i=1}^{n} x_i \right)^{1/n} = (x_1 \times x_2 \times \ldots \times x_n)^{1/n} \quad x_i > 0 \)

\[ \lg(GM) = \frac{1}{n} \left( \lg x_1 + \lg x_2 + \ldots + \lg x_n \right) = \frac{1}{n} \sum_{i=1}^{n} \lg x_i \]

\[ L = 10 \lg (I/I_0) \quad L = \text{Level} \quad I = \text{Intensity} \quad I_0 = \text{Reference Intens.} \]

Putting \( x_i = \frac{I_i}{I_0} = 10^{L_i/10} \)

\[ L_{eq} = 10 \lg \left[ \frac{1}{n} \sum_{i=1}^{n} 10^{L_i/10} \right] = 10 \lg \left[ \frac{1}{n} \sum_{i=1}^{n} x_i \right] = 10 \lg \bar{x} \]

\[ \bar{L} = \frac{1}{n} \sum_{i=1}^{n} 10 \lg x_i = \frac{10}{n} \sum_{i=1}^{n} \lg x_i = 10 \lg (GM) \]
Confidence Interval

ISO 2602 /1/ gives the following equation 2 for calculating the confidence interval for the arithmetic mean of a normal population.

\[ \bar{x} - t s / \sqrt{n} < \mu < \bar{x} + t s / \sqrt{n} \]

n: total number of results
m: population mean
\( \bar{x} \): arithmetic mean of the n results
s: sample standard deviation
t: value of Student's t-distribution at a certain confidence level

The validity of equation 2 and the accuracy of determining the confidence interval depend on how near the population distribution is to the normal one. According to Draft VDI 3723 /2/ L\(_{eq}\) of the population and its confidence interval can be calculated using equation 2 assuming a normal distribution of the sound intensity. Even if the distribution is not known it is permitted according to Draft VDI 3723 to estimate L\(_{eq}\) if the number of measured noise levels n \( \geq 5 \).

It must be kept in mind that equation 2 is only suitable to determine the confidence interval if the variate (in our case the sound intensity) belongs to a population which is normally distributed. But it has been already stated that the distribution of sound intensity is inhibited at the lower end by the minimum value zero. To illustrate how detrimental the effect can be of estimating the confidence interval of L\(_{eq}\) according to VDI 3723 examples will be given in table 3 for which the confidence interval of L\(_{eq}\) will be estimated both parametrically according to Draft VDI 3723 and by using the nonparametric method. The nonparametric method for estimating the confidence interval of L\(_{eq}\) does not yield a simple equation according to which a straight forward solution can be obtained. It needs much more calculations which was certainly a drawback at times when a computer was not easily available. It is out of scope here to explain the nonparametric method.

In the following examples (table 3) the two-sided confidence interval for L\(_{eq}\) is estimated at the confidence level 80% to comply with VDI 3723. This means that the corresponding confidence level for the one-sided confidence interval is 90%. In other words with a probability of 90% L\(_{eq}\) lies under the upper limit of the confidence interval or over the lower limit of that interval.
Table 3  Examples using variable noise levels in db

<table>
<thead>
<tr>
<th>Nr.</th>
<th>Variable levels</th>
<th>$L_{eq}$</th>
<th>Confidence interval</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td></td>
<td>Parametric</td>
</tr>
<tr>
<td>1</td>
<td>55 57 59 61 63</td>
<td>59.9</td>
<td>57.1 - 61.6</td>
</tr>
<tr>
<td>2</td>
<td>52 54 55 57 63</td>
<td>58.1</td>
<td>50.8 - 60.7</td>
</tr>
<tr>
<td>3</td>
<td>52 54 55 57 64</td>
<td>58.7</td>
<td>48.2 - 61.6</td>
</tr>
<tr>
<td>4</td>
<td>52 54 55 57 65</td>
<td>59.4</td>
<td>- 62.4</td>
</tr>
</tbody>
</table>

As to the upper limit of the confidence interval there is a good agreement between the two methods for the examples in table 3. For example nr.1 there is even a good agreement for the lower limit. But considering the examples 2 to 4 a deterioration is obvious in the agreement between the lower limits as estimated from the two methods. The reason lies in the assumption of the parametric method that the population of sound intensity is a normal one. The normal distribution is a symmetrical one. As the highest value of observed noise levels increases, the symmetry of the normal distribution dictates a decreasing lower limit of the confidence interval until a situation arises as in example 4 where the estimated lower limit of the intensity is negative. In this case Draft VDI 3723 gives up and states that a lower value for the confidence interval cannot be given (logarithm of a negative value). The nonparametric lower limit of the confidence interval remains constant for examples 2 to 4. Acoustically speaking this is reasonable: a set of 4 values of noise level remains constant and only the highest fifth value increases in two steps of 1 db each leading to an increase in the upper limit of the confidence interval but leaving the lower limit unchanged.

The superiority of the nonparametric method of estimating the confidence interval for $L_{eq}$ is evident.

Literature

1/ ISO 2602-1980, Statistical interpretation of test results—
   Estimation of the mean—confidence interval.

UNIT FOR MEASUREMENT AND STORAGE OF ENVIRONMENTAL NOISE

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1. Introduction

A portable and low consumption measurement station is developed for sampling and recording sound pressure levels, for later statistical processing of environmental noise.

With a sampling rate of one Hertz, measurements are accomplished over ranges limited between 30 and 120 dB, with 2 dB of resolution.

The dB(A) compensated analog signal is transformed to discrete binary values by an analog-to-digital converter and then stored into a memory.

Data accumulated every four minutes are transferred in block into one channel of a stereo tape recorder. Each information block includes date, hour of occurrence and measurement range. In the other channel samples of analog signals for monitoring are recorded.

An inexpensive commercial recorder, with closed tape, is used and permits with a 180-minute cassette length, recordings for measurements of about 100 hours.

Certain prefixed critical values for meteorological phenomena like rain and wind, put the unit temporarily out of service.

The registers obtained from digitalized blocks are formatted in a normalized mode for numerical processing by a micro computer system and permits the obtention of diverse statistical parameters like L eq, L 10, L 50, etc. for periods of the required length.

2. Description of the instrument (Fig. 1)

2.1 Analog section

The signal coming from a preamplifier-microphone goes into a low noise attenuator-amplifier with a 50 dB dynamic range.

In order to adapt to the different variation levels of
environmental noise, the minimum level can be selected from 30 dB to 70 dB in 10 dB steps, thus obtaining five measurement ranges of 50 dB each. The output of this attenuator-amplifier is connected to a circuit board of a low-cost, commercial SLM. Moreover, this signal can be stored in one of the channels of a cassette stereo recorder. The SLM delivers a DC voltage of 40 mV for each dB in sound level changes.

Due to the 2 dB resolution specified, each measurement range will contain 25 channels. Besides, two extreme channels are considered for the levels above and below the range selected. This is why the A/D converter utilized is of 5 bits. The series output of the converter is connected to the display control in order to visualize the active measurement channel.

2.2 Digital section

The 5 bits parallel output of the analog-to-digital converter are used to address the memory. In this way, each channel corresponds to a definite location in the memory, storing the number of times a sound level is repeated during 4-minute periods. For operative purposes two sectors are differentiated in the memory: A and B. A sector contains the values accumulated during the present measuring period, whereas the values stored at the previous four minutes are maintained in B sector. Thus the measuring process is not interrupted while the data accumulated during the previous period are transferred from B sector to the magnetic tape recorder. The address generator is used during transfer of the accumulated data from A to B.

Every time a channel is addressed by the A/D converter the memory places into the data bus the value so far accumulated. To it the arithmetic unit immediately adds one. Once again, through the temporary data register, the new value is stored in the same location.

An hour clock and running-day calendar are implemented. To each data block accumulated are added range, time and date when the measurement was carried out. These additional values can be pre-adjusted by the operator and visualized in a seven-segment display.

Series-to-parallel converters are shift registers. Before starting a block recording they are loaded with present range, time and date and after this, sequentially, all 27 measurement channels located in the memory's B sector. The data block is preceded by the FF hexadecimal number. The bits strip feeds a frequency modulator (FSK) in order to be recorded in the other channel of the above mentioned recorder or to be transmitted through a telephone line.

2.3 Logic Control

This sequential circuit manages all the activity of the instrument. Control signals generated at the suitable moment constitute the control bus. Fig. 2 (a) shows a diagram of the operation sequence. Once it is powered on the unit remains in
stand-by, until the order to measure is issued. Range, hour
and date values can be preadjusted in this state. Once the
measuring order is issued, its execution is delayed until the
onset of the next minute. When this happens, the counter which
determines the four-minute periods, is enabled. This is done
in order that the first measuring block should be the same
length as the remainder. After this the memory locations of A
sector corresponding to the 27 measuring channels are cleared.
The one-second signal determines the time when a sample of the
analog signal must be taken. The analog-digital conversion
takes place at once. After this comes the reading of the accu-
mulated value in the addressed channel; one is added and it is
taken back to its original location. If the four minute signal
have not yet gone by and there is no recording in progress on
the magnetic tape once again the next second is awaited and
the cycle is repeated. If the 4 minutes have passed, the order
of preparation for the data transference is given, and the re-
corder set in motion (REC MODE). At this time, moreover, the
data accumulated during the last four minutes are moved from A
sector to B sector in the memory. Finally, a return is made to
(2) (see Fig. 2 (a)). Once in REC MODE logic control directs
alternately the transmission of data towards FSK and the acqui-
sition and conversion of new values. The transferring process
to cassette of one data block takes 7 seconds altogether.

Asynchronically, (Fig. 2 (b)) the measuring process of the
instrument can be stopped in an adjustment or stopping operat-
on is begun, or if some adverse atmospheric conditions sensor
is activated. An additional starting order must be executed in
order to restart the measurement.

2.4 Outputs

Analog and FSK output are sent simultaneously during REC
MODE into a stereo cassette recorder.

Digital serial output and status bits can be connected to a
microcomputer for real time processing.

Telephone line output permits remote reception and proces-
sing of the information.

3. Conclusions

This portable instrument is an efficient working tool for
the study and analysis of environmental noise during long time
periods.

Due to its continuous 4-day operation autonomy registering
4-minute data blocks (360 blocks a day), an appreciable amount
of information is available for processing.

Its wide measuring range, from 30 dB to 120 dB, and its
2 dB resolution make it suitable for statistical analyses in
order to determine the various influences of noise on man.
RAMOS Oscar - UNIT FOR MEASUREMENT AND STORAGE OF ENVIRONMENTAL NOISE

**Fig. 1**

- ATTEN./AMPLIFIER
- S.L.M.
- A/D CONVERTER
- ADDRESS GENERATOR
- MULTIPLEXER
- MEMORY
- PARALLEL/SERIE CONVERTERS
- RANGE
- DATE
- CLOCK
- FRONT PANEL CONTROL
- CLOCK GENERATOR/DIVIDER
- DISPLAY CONTROL
- LOGIC CONTROL
- ARITHMETIC UNIT
- DATA REG
- F.S.K.
- Digital serial output
- to telephone line
- to tape recorder
- Analog output

**Flowchart**

1. **Power on**
   - MED? NO
   - YES

2. **Clear Channels (sector)**
   - sec.? NO
   - YES

3. **Take a Sample**
   - Perform Analog / Digital Conversion
   - Add one to active channel

4. **Set REC MOD Move (sector) into (sector one)**
   - YES
   - NO

5. **Record**
   - End? NO
   - YES

6. **Reset REC MOD**
   - YES
   - NO

7. **GO TO**
   - (2)
   - (3)
   - (4)
IEC 651  A NEW MEASUREMENT CONCEPT

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Traditionally there has been a big difference in both price and performance between precision (IEC 179) and industrial (IEC 123) sound level meters. However, since these two specifications came into being technology has moved to a position where the electronics, formerly the most expensive part of the meter, are now the least costly. The micro-chip revolution has meant that the number and cost of parts has reduced in constant value terms, while the cost of the mechanical parts has risen at least in line with general inflation.

Thus, when IEC 651 was published in 1979 manufacturers were faced with a dilemma. Either they could simply introduce new models to meet the new general purpose Grade 2 or ignore this grade entirely or perhaps they could rethink their whole rationale. Table 1 shows clearly the relationship between the new and old specifications and illustrates the dilemma.

<table>
<thead>
<tr>
<th>Grade</th>
<th>Description</th>
<th>Old UK</th>
<th>Old US</th>
<th>Old Int.</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>Laboratory</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>1</td>
<td>Precision</td>
<td>BS4197</td>
<td>ANSI 1</td>
<td>IEC 179</td>
</tr>
<tr>
<td>2</td>
<td>Gen purpose</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>3</td>
<td>Survey</td>
<td>BS3489</td>
<td>ANSI 2</td>
<td>IEC 123</td>
</tr>
</tbody>
</table>

In the event, all 3 routes were chosen. Manufacturers with 'old' electronics for their IEC 123 models mostly pulled out of the development race. Others introduced totally new models all dedicated to a single grade, which meant a big proliferation of new models on the market.

The 3rd route, a total rethink produced 2 main conclusions:

1. Intelligent use of microprocessors could revolutionise the whole field of acoustics.
2. Integrated circuits meant that the main difference between a Grade 1 and a Grade 3 meter is now the microphone; the electronics being mainly the same.
The microprocessor approach naturally appealed mainly to university based groups and several designs came out quite quickly. However, as in many 1st generation units, vital parameters were missing. Mainly, little thought was given to the back-up in terms of ancillary equipment which the larger commercial operations could provide, and prices were usually pitched at a level near to the market leaders, with disastrous results in terms of sales. The market clearly preferred the combination of more traditional technology and worldwide back-up. Some commercial companies, particularly Americans who had produced microprocessor based units well in advance of IEC 651 now produced 2nd generation units which did not fall into the same trap.

The second new approach was the grade convertible system. This concept was simply based upon the logic of acoustic measurement. In all such measurements there are several near essential requirements as shown in Fig. 2.

These basic requirements do not change whether the final block, log/readout is a simple meter or a micro computer, the principle is the same. If the elements within the block A to E are able to give a basic accuracy of say 3% overall, - of the order of 0.3dB - , it is clear that the whole unit accuracy depends on the microphone before A and the log-readout after E.

Initially such units called convertible grade meters were made with normal fixed meter readouts but with a microphone input system based on the LJM interface. Now, however, this approach has refined into a convertible grade system where the function, display and accuracy are all under user control.

Essentially, a convertible grade system has all the elements of the 'basic' meter with several very important additions.

The most vital additions to the basic unit consist of electronic switches - normally closed - which can break the signal paths at the low level ac and high level dc points together with accessible outputs from each block in the system. Fig. 3 shows this clearly.
The next addition is a logic controlled "hold" circuit which can function either as a "max hold" or as an "automatic memory" to choice. This permits the sound or vibration level at any time to be locked into the display on external demand.

The 3rd addition is the provision of a highly accurate voltage reference supply not only to add stability to the system but to act as the calibration source for external units. This needs to be accurate to about 0.1%.

The final addition is a DP15 bus connector which carries all the inputs and outputs except the microphone. Each pin being specified in terms of sensitivity, level, impedance and function. Once this standard has been established, the realisation inside each block becomes unimportant. For example, blocks 1 and 2 the input amplifier/attenuator can be autoranged, or block 6 the readout can be a standard microcomputer, with almost any programme subject to the programmer's imagination.

The mechanical form of the units while specified in terms of the distance of the fixing bush - a standard tripod mount - is not vital to the system provided that a mechanical form chosen does not break the capacity rules for the system. If these rules are followed units from different manufacturers can be connected together and function within the specified parameters. This allows the smaller manufacturers or universities to build small volume specialised units to compete effectively with both the original design group - Cirrus Research - and the other companies making units in the system.

What is the advantage to the user of the convertible grade system over dedicated instruments? There are 2 main user advantages due to the flexibility of the convertible concept,
The first is the advantage to the industrial user who often has no clear idea of his measuring needs and has today severe financial restrictions. With a convertible grade system, the minimum purchase of a body and a Grade 3 microphone not only limits the investment but when ideas are crystallised any function is immediately available by plug-in, including the obvious octave band filter and Leq etc.

The second advantage is in the laboratory, where the unique ability to access the circuit at all points via the DP 15 with a clean and specified interface, means that many of the traditional restraints on the use of sound level meters - particularly relating to their flexibility - now disappears.

The revolution of the microprocessor fits particularly well with the L3M - DP15 logic. 3 main outputs are available:

1. High level ac after filtering
2. r.m.s. dc at 1V for 40dB
3. log r.m.s. dc at 1V/decade

If Leq is the main requirement, the ac output can be fed for example to a squarer and charge balancing integrator and an input to the CPU proportional to (pressure)^2 is available. If "binning" is required for say L10, L90 the log r.m.s. can be fed to multiple I/O ports on the CPU to perform the A/D conversion. If Sound Pressure or Vibration Level is the prime requirement the r.m.s. output is fed to the I/O ports.

Finally the readout can be fed back to the display on the main frame or as it is available on the DP15 it can be fed to a digital readout.

The final question relating to the L3M - DP15 must relate to its adaptibility to the well established de facto international standard. Because both the DP15 and the L3M interfaces are so highly specified, a single conversion often passive is usually sufficient to change systems. For example, the MV 180B 200V generation pre-amplifier takes the L3M direct to the standard half-inch microphone system and thus the vast range of existing knowledge and equipment based on this standard is at once available to L3M users.

It would seem that the L3M - DP15 bus system is capable of all the parameters currently in use to any grade of IEC 651 other than '0'. Its innate flexibility would seem to ensure that any new requirement can be met with a minimum of change by many possible suppliers.

Ref: La Commission Electro-technique Internationale (IEC 651)
DISTRIBUTION OF POWER SPECTRA ESTIMATES

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Introduction

In various acoustical applications of spectrum analysis it is important to know the accuracy of the results. The best way to characterise the accuracy is to specify the probability density function (pdf) of spectrum estimates. In this paper an algorithm is presented for the calculation of pdf $\phi(\gamma, BT, \delta)$ of the results $\gamma$ of power spectrum density (psd) determination by means of band pass filtering - squaring - averaging type af an analyzer with sharp-edged band pass, ideal squarer and true integrator. $B, f_0, T$ are filter's bandwidth, its central frequency and averager's integration time respectively. Stationary Gaussian noise with slow varying psd is assumed at the analyzer's input. All possible values of $BT$ and two extreme values of $\delta=B/f_0$, $\delta<<1$ and $\delta=2$, are considered. Small $\delta$ corresponds to a high resolution spectrum analysis while the maximum one, $\delta=2$, describes the situation when total sound intensity is measured.

The relation of $\phi(\gamma, BT, \delta)$ to $\chi^2$-distribution is also discussed here. The details of derivation and the results of (successful) experimental verification of the mathematical model are given elsewhere [1].

Algorithm

The pdf of psd estimates normalized to unity mean, $\gamma$, is given as follows:

$$
\phi(\gamma, BT, \delta) = \int_{-\infty}^{\infty} \delta(z, BT, \delta) e^{-i2\pi\gamma z} \, dz,
$$

$$
\delta(z, BT, 0) = \sum_{k=0}^{N(BT, 0)-1} \left[ 1 - i2\pi\mu_k(BT, 0)z \right]^{-1},
$$

$$
\delta(z, BT, 2) = \sum_{k=0}^{N(BT, 2)-1} \left[ 1 - i2\pi\mu_k(BT, 2)z \right]^{-1/2},
$$

$$
\mu_k(BT, 0) = \frac{\lambda_k(\pi BT/2)}{BT}, \quad \mu_k(BT, 2) = \frac{\lambda_k(\pi BT)}{BT}, \quad k=0, 1, 2, \ldots
$$

(1)
Here $\delta=0$ and $2$ stand for $\delta<1$ and $\delta=2$ respectively, while $\lambda_k(c)$ are eigenvalues associated with the prolate spheroidal wave functions $\psi_k(c,x)$:

\[
\int \frac{sinc(x-y)}{\pi(x-y)} \psi_k(c,x)dx = \lambda_k(c)\psi_k(c,y), \quad k=0,1,2,\ldots,
\]

and $N(BT,\delta)$ is given in the table.

\[
N(BT,\delta) = \begin{cases} 
2[C+\ln(8nBT)]/\pi^2+BT+1/2, & \delta=0, \\
2[C+\ln(16nBT)]/\pi^2+2BT+1/2, & \delta=2,
\end{cases}
\]

$C=0.57721\ldots$ is Euler constant.

Two approximations are usable for the calculation of $\lambda_k(c)$, $k=0,1,2,\ldots$:

\[
\lambda_k(c) = \frac{2}{\pi} \frac{2^k}{(k!)^{3/2}} \frac{\Gamma(2k+1)}{\Gamma(k+1)^2} \left[1 - \frac{(2k+1)c^2}{(2k-1)(2k+3)^2}\right].
\]

\[
\sigma(k,c) = \frac{\pi^2/4}{C+\ln(16c)} \left(\frac{k-2c}{\pi} + \frac{1}{2}\right),
\]

Eq. (4) is to be used with (2) in the following cases:

\[\begin{align*}
&[(\delta=0, BT<1.0) \text{ and } (\delta=2, BT<0.5)] \text{ for all } k, \\
&[(\delta=0, 1.0<BT<1.8) \text{ and } (\delta=2, 0.5<BT<0.9)] \text{ for } k=2;
\end{align*}\]

in all other cases (5) with (2) will be more appropriate.

The pdf $\phi(y,BT,\delta)$, $\delta=0$ or 2, follows from $\phi(z,BT,\delta)$ via Fourier transformation (1). It has to be calculated only for $y>0$ since

\[
\phi(y,BT,\delta)=0, \quad y<0,
\]

holds. Fast Fourier transform is an appropriate technique although there are some analytical solutions as listed bellow ($y>0$):

\[
\phi(y,BT,0)=e^{-2y}, \quad 0<BT<0.2, \quad (6)
\]

\[
\phi(y,BT,2)=e^{-y/2}/\sqrt{2\pi}, \quad 0<BT<0.1, \quad (7)
\]

\[
\phi(y,BT,2) = \frac{1}{\sqrt{2\pi}} e^{-\frac{1}{2}y} I_0 \left[\frac{1}{2} \left(1-\frac{2}{\mu}\right)\right], \quad (8)
\]

where

\[
\mu=2 \left\{ \frac{(\pi BT)^2}{3} \right\} \left[1-3 \left\{ \frac{(\pi BT)^2}{5} \right\} \right], \quad 0.1<BT<0.3,
\]

\[
\phi(y,BT,0) = \sum_{k=0}^{N-1} \frac{1}{\mu_k} \frac{1}{\mu_k} e^{-y/\mu_k} \prod_{k=0}^{N-1} \left(1-\frac{\mu_k}{\mu_{k'}}\right), \quad (8)
\]
\[ \phi(\gamma, \beta T, 0) = \frac{N-1}{\mu_0} \left( \frac{\mu_0}{\mu_k} \right)^{N-1} \left[ e^{-\gamma/\mu_k} - e^{-\gamma/\mu_0} \right] \left[ \sum_{j=0}^{M-1} \frac{1}{j!} \left( \frac{\mu_0}{\mu_k} \right)^j \right]. \] (9)

Here \( N \) and \( \mu_k \) stand for \( N(\beta T, 0) \) and \( \mu_k(\beta T, 0) \), respectively, and \( I_0(z) \) is the zero order modified Bessel function. When \( 0.2 < \beta T \leq 1.8 \) (8) has to be used, while both (8) and (9) are applicable when \( \beta T > 1.8 \). The former should be used if all \( \mu_k, k=0,1,2,\ldots, N \), are different and the later if first \( M>1 \) among the coefficients are equal.

Illustrative review of pdf \( \phi(\gamma, \beta T, \delta) \), \( \delta = 0 \) and 2, are given in Fig.1, 2.

Relation to \( \chi^2 \)-distribution

\( \chi^2 \)-distribution is closely related to pdf of psd estimates. The approximations (6) and (7) which are valid in the low-\( \beta T \) range are \( \chi^2 \)-distributions themselves. The detailed analysis has shown that \( \phi(\gamma, \beta T, \delta) \), \( \delta = 0, 2 \), tend to \( \chi^2 \)-distribution only when \( \beta T > 0 \) and when \( \beta T \) is as high as necessary for distribution to approach the normal one.

Some middle-\( \beta T \) functions are shown in Fig.3.

Fig.3 - Comparison of (a)-\( \chi^2 \)-distribution functions with 2\( \beta T \) degrees of freedom in the variable 2\( \beta T \gamma \), and (b)-correct pdf \( \phi(\gamma, \beta T, \delta) \) with \( \delta = 2 \) and three values of \( \beta T \).

References

PEAK SOUND PRESSURE LEVEL AND SOUND EXPOSURE LEVEL OF A SONIC BOOM

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Some theory

Peak flat sound pressure (level) at a designated place has been the usually quoted measure of a sonic boom. But the peak is only a small part of the information inherent in such a transient. Sound exposure (level) is a summary descriptor of all the pressure-time information of the transient. Sound exposure level spectra can be obtained readily with a digital spectrum analyzer. The process is here described in terms of sound pressure and time, even though the instrument processes analogous voltages. As Eq.(1) shows, sound exposure \( E \) is the integral of squared (possibly frequency-weighted) sound pressure over time \( T \); \( E \) is also equal to the product of time-mean-square sound pressure \( <p^2> \) and averaging time \( T \). The squared instantaneous sound pressure is treated equally throughout the integration; there is no Hanning or exponential time weighting.

\[
E = \int_0^T p^2(t)dt = <p^2> \cdot T, \quad (1)
\]

\[
L_E = 10 \lg(E/E_0); \quad E_0 = (20 \mu Pa)^2 \cdot (1 s), \quad (2)
\]

\[
L_E = L_P + 10 \lg(T/1 s), \quad (3)
\]

\[
L_{Ecal} = L_{Pcal} + 10 \lg(t_{cal}/1 s) = L_{ESys} + K. \quad (4)
\]

If 1024 samples during time window \( T \),

\[
T = 1/b = 400/B = 1024/2.56B, \quad (5)
\]

\[
L_{P1Hz} = L_{Pb} - 10 \lg(b/1 Hz) = L_EB, \quad (6)
\]

\[
L_{EB} = L_{Pb} + 10 \lg(T/1 s) = L_{PB} - 10 \lg(B/400 \text{ Hz}), \quad (7)
\]

\[
L_{ELHz} = L_{Pb} + 10 \lg(T/1 s) - 10 \lg(b/1 Hz), \quad (8a)
\]

\[
L_{ELHz} = L_{Pb} - 20 \lg(B/400 \text{ Hz}), \quad (8b)
\]

At integer-\( n \)-multiples of characteristic frequency \( f_C \):

\[
L_{pk} = L_{ELHz} + 20 \lg(n f_C/0.450 \text{ Hz}); \quad \sqrt{8/2\pi} = 0.450. \quad (9)
\]
Equation (2) defines sound exposure level $L_e$ and reference exposure $E_0$. Equation (3), which follows directly from Eq. (1), is the first of two basic relations here involved. The correction $K_e$, to be added to a system reading $L_{sys}$ to get a correct sound exposure level $L_{ecal}$, can be calculated by Eq. (4) with a calibration sound pressure level $L_{p,cal}$ or its voltage equivalent, for a calibration time period $t_{cal}$.

In many digital analyzers, a signal is captured by 1024 samples during time window $T$. The instrument measures time-average sound pressure level $L_{p,avg}$ during $T$, in 400 contiguous "bins". The frequency width of each bin is $B$; that of the complete analysis band is $B = 400B$; the widths are related to time window $T$ by Eq. (5).

Per Eq. (6), average one-hertz band sound pressure level $L_{p,1Hz}$ (sound pressure spectrum level) during time $T$, is obtained by subtracting ten times the logarithm of the ratio of the bin width to one hertz, from a bin sound pressure level $L_{p,bin}$. This is the second basic relation.

The first form of Eq. (7) comes from Eq. (3); the second follows from Eq. (5). Equation (8a) contains both the time-integration function and the spectral density function. Equation (8b) is often a more convenient form.

Sonic boom and measuring instruments

The sonic boom here analyzed was generated by an F-15 aircraft flying Mach 1.1 at an altitude of 6100 m (20 000 ft) above the desert floor. The receiving microphone was in a foam windball, 18-cm in diameter, resting on a rocky peak 600 m (2000 ft) above the desert floor. The lateral range from aircraft to microphone was about 35 km (19 nautical miles).

Fig. 1 (20:0850). Sonic boom waveform of F-15 aircraft.
The pressure microphone was Brüel & Kjær Type 4147 No. 33782, connected to B&K Type 2631 Microphone Carrier System No. 353742, lower limiting frequency 0.01 Hz. The dc signal was recorded on RACAL Store 4D tape recorder No. 2606; tape speed 30 in/s. On playback, the analyzed signal was displayed as voltage level re 1 µV and analyzed by Spectral Dynamics 345-1-44 Spectroscopy III No. 257.

The overall system was calibrated at six frequencies, by GenRad 1986 Omical Calibrator No. 00539. With settings of controls as used, the pressure-voltage sensitivity of the system on the average at 125, 250, 500, 1000 Hz was 126 Pa/V; the correction \( K \) to be added to a bin voltage level to obtain bin sound pressure level was 16.2 dB; for analysis band sound pressure levels, \( K = 18.0 \) dB, in consideration of a design error in SD345. Adjustments to obtain a sound exposure level in indicated bands from a time-average sound pressure level, are given in Eqs.(7) and (8).

Waveform and sound exposure level spectra

The recorded waveform of the sonic boom is shown in Fig.(1). The duration of the quasi N-wave is 150 ms, so the characteristic frequency is \( f_c = 1/0.150 \text{ s} = 6.6 \text{ Hz} \). The spectrum shape is roughly that of an ideal sonic boom (N-wave), for which peak flat sound pressure level \( \text{PKT} \) is related by Eq.(9) to respective one-hertz band flat sound exposure levels (1-Hz TSEL) at integer-multiples of \( f_c \). This relation follows from recognition of the fact that in the usual engineering sense of positive frequency only sound exposure spectral density (here called one-hertz sound exposure) of a transient is twice the square of the Fourier transform, and that the Fourier transform of an N-wave at integer multiples of the characteristic frequency simplifies to Eq.(9). The process is shown graphically in Fig. 2. The PKT extrapolated back to 0.45 Hz for \( n = 1,2,3 \) are

![Fig.2(20:0850). Flat sound exposure level spectrum of F-15 aircraft.](image-url)
126.9, 129.8, 130.6 dB. The average PKT = 129.4 dB is in good agreement with 128.6 dB determined independently in Fig.1.

The A-weighted sound exposure level spectrum is given in Fig. 3. The broad maximum is around 50-500 Hz, instead of near 5 Hz for flat. The 1-kHz analysis band A-weighted sound exposure level is 82.9 dB, compared with 116.1 flat.

Wide-band flat sound exposure level of an N-wave of 150 ms duration theoretically exceeds peak flat sound pressure level by 10 lg(0.150/3) = -13.0 dB. This is near the experimental 116.1 - 128.6 = -12.5 dB from Figs. 1 and 2. From Fig. 3 for A-weighted sound exposure level, the corresponding difference is 82.9 - 128.6 = -45.7 dB.

In the past, predictions of possible vibration and damage to buildings, and audible annoyance of people, have been largely based on peak flat sound pressure levels of sonic booms. The spectrum in Fig. 2, relative to peak flat sound pressure level, is potentially useful in identifying parts of the spectrum particularly relevant to resonances in building structures. The difference of -45.7 dB between A-weighted sound exposure level and peak flat sound pressure level is potentially useful in relating past predictions based on "peak" to future predictions of long-term average A-weighted sound levels in communities where there are sonic booms from distant F-15 aircraft.

MICROCOMPUTER MEASUREMENTS OF IMPULSE NOISE

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Laajaniityntie 1, 01620 Vantaa 62 - Finland

Introduction

Within a few years microcomputers have become applicable for field measurements of acoustical signals because of advances made in the speed of processing and diminishing dimensions. A tape recorder usually has limitations due to errors of reproduction and its dynamic range, especially in the measurement of transient signals. For the purpose of evaluating noise-induced hearing loss, impulsiveness is analyzed, as the difference between the peak levels (L_Ap) and rms levels (L_As) of the noise. Nowadays hearing protectors are commonly used, but their efficiency in protecting against industrial impulse noise is unknown. For this reason the peak levels and rms levels were measured both on the outside and the inside the earmuff during shipyard work.

Instrumentation

The personal exposure to noise was measured by two microphones simultaneously. The miniature microphone inside the earmuff was a Knowles 1802 (size 8 mm x 6 mm x 2 mm), and the one outside was of the same type but with a 1/2" cylindrical housing. A stationary 1/4" microphone was used with the precision sound level meter. The telemetric system allowed the worker freedom to move within a distance of 40 meters. The new definition of impulsiveness (see lecture in session 88) is based on the detected L_Ap and L_As levels, which are recorded on the tape recorder and processed by the microcomputer. The microcomputer is equipped with an analog to digital converter (ADC) with a 60 dB dynamic range. The ADC is able to sample the four input signals 500 times per second with basic language.

The microcomputer was programmed with assembly language to increase the speed to 15000 samples per second. The assembly language was then compiled to machine language. The telemetric unit reveals the transmission disturbances with an error pulse. Unfortunately the frequency of error pulses was too high for them to be sampled even with the machine language. Thus a schmitt-trigger, which inhibited the sampling for 100 ms from the first error pulse, had to be installed.
Fig. 1. a) The telemetric system with an earborne microphone; b) The sound level meter with a stationary microphone

During the measurement the microcomputer displayed the four instantaneous $L_{AP}$- and $L_{AS}$-values and collected them to a memory address according to their absolute voltage values. After a measuring period of 10 minutes the cumulative distribution functions were calculated and saved, with a measurement index, within the memory diskettes.

The whole measurement system was tested in a shipyard by installing the three microphones to the same tripod. Different types of impulse noise were used for comparison in the time domain. In the measurements of exposure to noise the miniature microphone was attached beside the ear canal by tape. The 1/2" microphone was attached to the helmet outside the earmuff.

Results

Ten sledge impulses were averaged in seven ten-minute comparison measurements with the three microphones on the same tripod (Table I). The $L_{AP}$, $L_{AS}$, and $I$ were estimated from the corresponding impulses in the level recorder paper. During the measurements a normal microphone was changed to a miniature microphone in the dosimeter number 2 to ensure that the dimensions of the microphone did not affect the results.

The average value and the standard deviation of the difference between pair-wise measurements of 40 sledge impulses measured with the normal 1/2" microphone and the 1/4" microphone were

$L_{AP} = (-0.8 \pm 1.5) \text{ dB and } L_{AS} = (-1.7 \pm 1.3) \text{ dB}$

The corresponding differences between the 1/2" microphone and the miniature microphone were

$L_{AP} = (0.3 \pm 1.4) \text{ dB and } L_{AS} = (0.9 \pm 1.7) \text{ dB}$
Table I. The average levels in the comparison measurements:
1 = outside microphone; 2 = inside microphone; 3 = sound level meter;
U = 1/2" microphone in 2; and M = miniature microphone in 2

<table>
<thead>
<tr>
<th></th>
<th>\textit{L}_\text{AP}</th>
<th>\textit{L}_\text{AS}</th>
<th>I</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>1</td>
<td>2</td>
<td>3</td>
</tr>
<tr>
<td>1 U</td>
<td>134</td>
<td>131</td>
<td>133</td>
</tr>
<tr>
<td>2 U</td>
<td>109</td>
<td>109</td>
<td>111</td>
</tr>
<tr>
<td>3 U</td>
<td>102</td>
<td>101</td>
<td>-</td>
</tr>
<tr>
<td>4 M</td>
<td>118</td>
<td>116</td>
<td>119</td>
</tr>
<tr>
<td>5 M</td>
<td>114</td>
<td>114</td>
<td>116</td>
</tr>
<tr>
<td>6 M</td>
<td>106</td>
<td>107</td>
<td>109</td>
</tr>
<tr>
<td>7 M</td>
<td>129</td>
<td>127</td>
<td>-</td>
</tr>
</tbody>
</table>

The actual exposure to noise from a sledge, chiselling, and grinding was measured from both sides of the earmuff (Table II). The averages of the same impulses outside and inside the earmuff were calculated for the \textit{L}_\text{AP}, \textit{L}_\text{AS}, and I values of ten impulses.

Table II. The average levels measured on either side of the earmuff:
S = sledge impulses; C = chiselling; G = grinding; and A = attenuation of the earmuff

<table>
<thead>
<tr>
<th></th>
<th>\textit{L}_\text{AP}</th>
<th>\textit{L}_\text{AS}</th>
<th>I</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>1</td>
<td>2</td>
<td>A</td>
</tr>
<tr>
<td>1 S</td>
<td>129</td>
<td>101</td>
<td>28</td>
</tr>
<tr>
<td>2 S</td>
<td>120</td>
<td>99</td>
<td>21</td>
</tr>
<tr>
<td>3 C</td>
<td>121</td>
<td>96</td>
<td>25</td>
</tr>
<tr>
<td>4 C</td>
<td>125</td>
<td>105</td>
<td>20</td>
</tr>
<tr>
<td>5 C</td>
<td>122</td>
<td>101</td>
<td>21</td>
</tr>
<tr>
<td>6 C</td>
<td>119</td>
<td>94</td>
<td>25</td>
</tr>
<tr>
<td>7 C</td>
<td>123</td>
<td>96</td>
<td>27</td>
</tr>
<tr>
<td>8 G</td>
<td>113</td>
<td>84</td>
<td>29</td>
</tr>
<tr>
<td>9 G</td>
<td>111</td>
<td>81</td>
<td>30</td>
</tr>
</tbody>
</table>

The differences between the levels measured with microphone inside and outside the earmuff reflect the efficiency of the earmuff to attenuate the energy and impulsiveness of the noise. The \textit{L}_\text{AS}(2) levels inside the earmuff show whether the level of 85 dB was exceeded. That level is considered as the limit after which hearing loss can occur; if it is exceeded inside the earmuff, then sufficient hearing protection is not provided. In measurements 4C and 5C the level of 85 dB was exceeded by 6 dB and 4 dB, respectively. The excess was mainly due to the high levels of noise outside the earmuffs (\textit{L}_\text{AS} = 110 dB).

On average, the earmuff decreased the peak level by 25 dB and the rms level by 23 dB. The cumulative distribution function of the difference between the \textit{L}_\text{AP} and \textit{L}_\text{AS} levels was analyzed with the digital signal analyzer (HP 5420A) and the microcomputer (ABC-80). The microcomputer was used at the measurement site, whereas the digital signal analyzer
analyzed the tape recorded signal in the laboratory. The cumulative probability distribution curves were calculated for the comparison measurements with the microphones on the same tripod (Fig. 2). The vertical axis is the percentage of the total time of measurement when the corresponding impulsiveness was exceeded on the horizontal axis.

Fig. 2. The cumulative probability distribution of the difference $L_{AP} - L_{AS}$ defining the impulsiveness analyzed with: a) a digital signal analyzer; b) a microcomputer

Conclusion

The microcomputer was found to be applicable in field measurements of acoustical signals. The comparison of the noise characteristics inside and outside the earmuffs suggests that both the equivalent levels and the impulsiveness was attenuated by the earmuffs. In noisy environments where the level of 85 dB was exceeded inside the earmuff, the simultaneous use of earplugs and earmuffs could be recommended. The straight processing of the most important parameters without the time-consuming tape-recording and manual analysis, will enable more reliable evaluation of hearing loss in relation to the exposure to industrial impulse noise.

References


COMPARATIVE STUDY OF METHODS FOR THE ESTIMATION OF IMPULSE RESPONSE

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Introduction

It is important for acoustic signal processing to estimate exact impulse response of an acoustic transmission system. And it seems that the methods for estimation using continuous signal such as the white noise or using impulse as the source signal are already established. But, they are not always the best method in the viewpoint of computation time and the accuracy.

This paper proposes first an index for the evaluation of accuracy of the estimated impulse response. And four methods using white noise, burst noise, rectangular pulse and definite short duration signals are experimentally compared with one another in the viewpoint of computation time and accuracy.

Experimental Set-up

Experiments were carried out using a small room (6 x 4 x 3 m³) whose reverberation time is about 0.5 second. Fig. 1 shows the experimental set-up. The source signal is supplied to the power amplifier and the radiated sound is picked up by a condenser microphone. The signal at the driving point of the amplifier and that at the output terminal of microphone amplifier are fed into AD converter, through low pass filters. The cut off frequency of the filters is 2.8 kHz and the sampling period is 128 us. The output of the AD converter is fed to an electronic computer by which the impulse response is estimated and the accuracy of the impulse response is evaluated.

Fig. 1 Experimental set-up
Method for Evaluation

The white noise is used as the source signal for the evaluation of accuracy of the estimation methods, and the following index \( P \) is defined.

\[
P = 10 \cdot \log_{10} \left[ \frac{(y(p) - x(p) \cdot h(p))^2}{(y(p))^2} \right] \tag{1}
\]

where \( x(p) \): sampled sequence of source signal  
\( h(p) \): sampled sequence of estimated impulse response  
\( y(p) \): sampled sequence of observed response  
\( * \): operator for convolution  

\( P \) will be \( \infty \) dB if the estimation and the computation are perfect.

Estimation Methods

(1) Cross Spectral Technique Using White Noise

In this method, computational error is generated by the effect of circular convolution and the shape of time window. The error can be decreased using very long time window.

(2) Cross Spectral Technique Using Burst Noise

The length of time window can be shortened using burst noise of short duration as the source signal. The principle is easily seen in Fig. 2. The response to the source signal does not continue more than \( T_d + T_r + T_s \) after the ending point of source signal where \( T_d \) is delay due to propagation and \( T_r \) is the duration of impulse response. And the enough length of time window is \( T_s + T_d + T_r \) where \( T_s \) is the duration of source signal.

The rectangular window is the best in this case as the circular convolution not come into question and the rectangular window does not change the wave forms.

A number of the computation of DFT is necessary for improving the signal to noise ratio. And the development of some method for reducing the quantity of computation is desired.

(3) Use of Rectangular Pulse as the Source Signal

The number of the DFT computation can be decreased to zero if a unit impulse is used as the source signal. The rectangular pulse can be used as the source signal instead of unit pulse if the duration is shorter than the sampling period.

But, the time lag of sampling point from the center of the rectangular pulse is not allowed because the lag causes the phase rotation proportional
to the frequency in the spectrum.

Signal to noise ratio is improved by the accumulation of every sample of the response tuning up the initial point. But many times of

![Graph](image)

**Fig. 3** Power spectrum of two kinds of M-sequence burst and their sum

![Graph](image)

**Fig. 4** Changes of evaluation of P (cancellation effect) with length of impulse response used for computation
accumulation is necessary as the pulse has a little energy.

This estimation method is very simple and practical so long as the rectangular pulse can be used as the source signal. But, the acoustical rectangular pulse can not be realized and the use of this method is limited to the special case where the driving point is in the electrical circuit.

(4) Cross Spectral Technique Using Intermittent Definit Signals

The burst signal with some duration has much more energy than a rectangular pulse. And a high signal to noise ratio is easily achieved by the accumulation of response truing up the initial point, if a definite same signal is used everytime. A burst of M-sequence is suitable for this purpose. But, the power spectrum of the burst has zeros as shown in Fig. 3-a. The transfer function can not be evaluated if the source signal has zeros in its power spectrum. This problem is solved using two kinds of the burst Fig. 3-b shows the power spectrum of a burst of the other M-sequence used in our experiment. The sum of the two power spectra has no zero as shown in Fig. 3-c.

Comparison Between the Method

Fig. 4 shows the changes in value of P by the length of impulse response for three methods. This figure indicates the superiority of the proposed method using two kinds of definite signal as the source signal.

The following table shows the special features of those methods.

<table>
<thead>
<tr>
<th>Method</th>
<th>(1)</th>
<th>(2)</th>
<th>(3)</th>
<th>(4)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Source signal</td>
<td>white noise</td>
<td>burst noise</td>
<td>rectangular pulse</td>
<td>burst of 2 definit signal</td>
</tr>
<tr>
<td>Time window</td>
<td>long (Parzen)</td>
<td>short (rectangular)</td>
<td>shortest (rectangular)</td>
<td>short (rectangular)</td>
</tr>
<tr>
<td>No. of FFT</td>
<td>F1</td>
<td>F2</td>
<td>0</td>
<td>5</td>
</tr>
<tr>
<td>for the same S/N</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>No. of time domain accumulation</td>
<td>0</td>
<td>0</td>
<td>M1</td>
<td>M2</td>
</tr>
<tr>
<td>for the same S/N</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Computation</td>
<td>C1</td>
<td>C2</td>
<td>C3</td>
<td>C4</td>
</tr>
<tr>
<td>Time for estimation</td>
<td>T1</td>
<td>T2</td>
<td>T3</td>
<td>T4</td>
</tr>
</tbody>
</table>

Apart from the above discussion, the method (1) has a practical merit that speech or music can be used as the source signal if the observation for long time is allowed.

Conclusion

Four methods for estimating impulse response of an acoustic transmission system with reverberation are compared. The proposed method using two kinds of definite signal is considered to be the best in the view point of less computation and less measuring time.
CANCELLATION OF SIGNAL PICKED UP IN A ROOM BY ESTIMATED SIGNAL

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Introduction

Effective sound reinforcement system will be realized if the loop gain of the sound reinforcement system through the microphone, amplifier, loudspeaker and acoustic transmission path (room) can be decreased by some method. The cancellation of the signal radiated from the loudspeaker and picked up by the microphone by the computed signal is considered to be one of the most effective method for reducing the loop gain.

We have realized a real time system for cancellation at the output terminal of the microphone. In the system, the convolution between the signal at the driving point of the loudspeaker and the impulse response of the transmission path from that point to the output terminal of the microphone is subtracted from the output of the microphone. The attenuation more than 20 dB is achieved under the condition that the room is kept unchanged from the time of measurement of the impulse response. The gain of the sound reinforcement system is increased stably by 6 dB even if there are moving men around the microphone.

Principle

Fig. 1 shows the schematic diagram of the sound reinforcement system equipped with the cancellation system. Before the use of the system, the

![Diagram of sound reinforcement system equipped with signal cancellation system](image-url)
impulse response \( h(t) \) of the transmission path from the driving point of the loudspeaker to the output terminal of the microphone is measured. When the sound reinforcement system is used, the convolution \( v(t) \) between the signal \( x(t) \) at the driving point of the loudspeaker and the impulse response \( h(t) \) is computed by high speed digital convolver.

\[
v(t) = x(t) * h(t)
\]  (1)

\( v(t) \) is subtracted from the output of the microphone \( y(t) \) which is the sum of two signals produced by the signal radiated from the loudspeaker and the external signal \( s(t) \) to be reinforced. The remaining signal after the subtraction is \( s(t) \) if the computation of \( v(t) \) is exact. And the feedback loop of the sound reinforcement system is cut off.

Preliminary Experiments

Computer simulations and preliminary experiments were carried out before the design of the real time cancellation system. The effect of the duration of the impulse response used for computing convolution on the cancellation is first investigated. The following index is defined for the evaluation of the cancellation effect.

\[
P = 10 \log_{10}[\frac{\{y(t)\}^2}{\{y(t)-v(t)\}^2}]
\]  (2)

That is, the index \( P \) denotes the ratio of the power of the signal picked up by the microphone to the power of cancelled signal.

The effect of length of impulse response used for computing convolution on the cancellation is first investigated. Fig. 2 shows the changes in the index \( P \) by the length of the impulse response. The reverberation time is about 0.52 sec. (4,096 samples). The experimental results agree well with those of computer simulation, and the possibility

Fig. 2 Effect of the length of impulse response on the cancellation of signal picked up by the microphone in the room
of cancellation more than 20 dB is shown. Fig. 3 shows an example of the spectrum of signal picked up by the microphone and that of the cancelled signal where the loudspeaker is driven by white noise. The results shown in Fig. 2 are obtained under the condition where the room is never changed during the period of experiment. But, in a real room there should be moving men and others.

Then, the experiments were carried out bringing cartons into the room after measuring the impulse response. Fig. 4 shows the effect of cartons on the cancellation. From those results, the length of the impulse response to be used for the real time canceller is determined to 1024.

![Fig. 3 Power spectrum of the signal picked up by the microphone and that of the cancelled signal](image1)

![Fig. 4 Effect of cartons brought into the room on the cancellation of signal](image2)

Cancellation System

For the measurement of the impulse response, a train of rectangular pulse is fed to the driving point of the loudspeaker (input terminal of power amplifier) and the signal picked up by the microphone is amplified and is digitalized. Every sample of the picked up signal is accumulated truing up the initial point of the response which should be at the center of the rectangular pulse as shown in Fig. 5. In our system, the sampling period is 128 micro second, the cut off frequency of the low pass filters used for AD and DA conversion is 2.8 kHz and the length of the impulse response to be used for computation is 0.131 sec. (1022 points).

For the cancellation, the convolution between the signal at the driving point of loudspeaker and the impulse response is computed and is subtracted from the output signal of the microphone. Every sample of the convolution should be computed within 128 micro second, that is, 1022 times of multiplication and addition should be carried out every 128 micro second. Two sets of high speed multiplying accumulator are used. All the system is controlled by a micro computer.
Fig. 5 Source signal (rectangular pulse) and the response picked up by the microphone, every sample of which is accumulated truing up the initial point.

Performance of the System

The cancellation more than 20 dB is achieved in a room where the reverberation time is less than 0.3 second and the index P decreases in a room of longer reverberation time. When someone enters the room and/or the door is opened after measuring the impulse response, the value P decreases largely. But it is easy to maintain the value more than 6 dB even if some persons move around the microphone and a man approaches the microphone by 30 cm so long as the position of the microphone is unchanged.

When P decreases by a little displacement of the microphone, the value P can be some how recovered by shifting the impulse response along the time axis. But the big compensation can not be expected.

A sound reinforcement system is equipped with the cancellation system. The increase of more than 6 dB in the amplifier gain is stably obtained under the condition that a man moves in front of the microphone.

Conclusion

The principle, the outline of the experimental system and the experimental results of a signal cancellation system are presented. With this system, the signal picked up in a room is cancelled by the estimated signal. The system is applied to the sound reinforcement system to avoid the howling due to acoustic feedback and a good result is obtained in spite of the changes in impulse response by moving men in the room. This cancellation system will be more effectively used for the conference between two rooms coupled by telephone network and equipped with hand free telephones. The increase in the amplifier gain more than 12 dB is expected for the cancellation systems can be set both sides.

The use of this cancellation system for picking up the signal in noisy circumstance is expected. But, from what signal the signal for cancellation is estimated is a big problem. And some achievement is expected if the main noise is transmitted into the room through narrow channels.
EFFECT OF TIME WINDOW ON ACCURACY OF ESTIMATION OF IMPULSE RESPONSE USING WHITE NOISE

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Introduction

Cross spectral technique is an available method for estimating an impulse response of a system if both input and output signals can be obtained. The power spectrum of an input signal should not be zero at every frequency to obtain an accurate impulse response. Therefore, if an arbitrary signal can be used as the test signal, stable white noise is used frequently. However, the accuracy of the estimated impulse response is decreased so long as the length of the time window is finite even if there is no external noise.

The present paper describes the reason why the accuracy of estimated impulse response is decreased and recommends the type of time window appropriate for the estimation of impulse response using white noise as a test signal.

The reason why the accuracy of estimated impulse response is decreased

The impulse response is estimated correctly using cross spectral technique if both input signal and its whole response are observed. However, the whole response can not always be observed since the length of time window is finite.

First, it is assumed that the input signal is an impulse, \( h(t) \) the impulse response, \( T_w \) the length of input time window and \( T_o \) the output time window. In the case (a) shown in Fig. 1-a, the ideal response should be zero in the output time window in order to obtain a correct impulse response. However, the shaded part of the response remains in the output time window and is assumed to be noise.
In the case (b) shown in Fig. 1-b, the shaded part of the response is shortened and also assumed to be noise. That is, the output signal is assumed to be the sum of the ideal response and the noise.

In the case (c) shown in Fig. 1-c, the ideal response is modified by the output time window. Thus, the shaded part, that is, the difference between the ideal response and the modified response is assumed to be noise.

Here, an arbitrary signal $x(t)$ is used as an input signal instead of an impulse. In the case (a), the response for the signal $x(t)$ outside the input time window is assumed to be a noise and represented as follows:

$$
n_a(t) = \begin{cases} 
\int_{-\infty}^{0} x(T) h(t-T) w_o(t) dT, & 0 \leq t \leq T_w_0 \\
0, & \text{otherwise.}
\end{cases} \tag{1}
$$

Whereas, the ideal response $s_a(t)=0$.

In the case (b), a part of the response to the signal $x(t)$ involved in the input time window overflows the output time window. The output signal is also assumed to be the sum of the ideal response $s_b(t)$ and the noise $n_b(t)$:

$$
s_b(t) = \begin{cases} 
\int_{-\infty}^{T} x(T) h(t-T) w_i(T) w_o(T) dT, & T_w_0 < t < \infty \\
0, & \text{otherwise.}
\end{cases} \tag{2}
$$

$$
n_b(t) = -s_b(t). \tag{3}
$$

In the case (c), the response for the signal $x(t)$ is modified by the output time window. The modified response is assumed to be the sum of the ideal response $s_c(t)$ and the noise $n_c(t)$:

$$
s_c(t) = \begin{cases} 
\int_{-\infty}^{T} x(T) h(t-T) w_i(T) w_o(T) dT, & 0 \leq t \leq T_w_0 \\
0, & \text{otherwise.}
\end{cases} \tag{4}
$$

$$
n_c(t) = \begin{cases} 
\int_{-\infty}^{T} x(T) h(t-T) w_i(T)[w_o(t)-w_o(T)] dT, & 0 \leq t \leq T_w_0 \\
0, & \text{otherwise.}
\end{cases} \tag{5}
$$
The signal to noise ratio is decreased due to one of the above-mentioned three reasons, and the accuracy of the estimated impulse response is decreased.

The signal to noise ratio S/N is easily calculated from the shape of time window when the input signal is white noise and the impulse response is represented by a delayed delta function or by an exponentially decaying white noise. However, it is clear that the higher the signal to noise ratio, the more accurate the estimated impulse response is.

The accuracy of estimated impulse response

Here, the index Ac is introduced for expressing the accuracy of the estimated impulse response:

\[ Ac = 10 \log_{10} \left[ \frac{\int_{-\infty}^{\infty} h^2(t) dt}{\int_{-\infty}^{\infty} \{ \tilde{h}(t)-h(t) \}^2 dt} \right]. \] (6)

where \( h(t) \) is the true impulse response and \( \tilde{h}(t) \) is the estimated one. The greater the value \( Ac \), the higher the accuracy is. The type and the length of the time windows are the same for input and output signals for the computation of the accuracy \( Ac \). The accuracy \( Ac \) is calculated for 8 time windows.

For the case where the impulse response is expressed by a delayed delta function, the mean value and the variance of the accuracy \( Ac \) are computed using 10 different white noise as the input signal. 100 cross spectra are averaged and the impulse response is estimated for every case. The length of time window \( T_w = 128 \) and the delay time \( T_d = 2, 4, \ldots, 64 \). Table 1-a shows the results.

For the case where the impulse response is represented by an exponentially decaying white noise, 10 different impulse responses are generated using 10 different white noise so that the envelope is expressed as

\[ h(p) = 10^{-2p/127}, \quad p=0,1,\ldots,127. \] (7)

For a time window whose length \( T_w = 8, 16, 32, \ldots, 2048 \), the mean value and the variance of the accuracy \( Ac \) are computed using 10 different white noise as the input signal. 100 cross spectra are averaged for every length of time window. Table 1-b shows the results.

It is clear that Rectangular window is the best for the case where the impulse response is represented by a delayed delta function and the time delay is longer than a quarter of the length of time window. The Rectangular window is also the best for the case where the impulse response is represented by an exponentially decaying white noise if the length of the impulse response is longer than that of time window. But, Parzen window is the best if the time window longer than that of the impulse response can be used.

Conclusion

The effect of time window on the estimation of impulse response by cross spectral technique is investigated. According to the results, Parzen window is the best except for the specially short time window in the view
point of the accuracy of estimated impulse response.

Reference


Table 1-a

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Table 1-b

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1*: Type of time window  2*: \( T_d/T_w \)
3*: \( T_w/T_1 \)  \( T_d \): Delay time
(\*): Length of time window \( T_1 \): Length of impulse response

(unit: dB)
Mesures de puissance et rayonnement
Power and radiation measurement
Leistungs- und Strahlungsmessungen
EFFECT OF THE SOUND FIELD STATISTICS IN ENCLOSURES WITH A ROTATING DIFFUSER ON THE SOUND POWER MEASUREMENTS

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U.S.A.

Introduction. The characterization of a noise source by its radiated sound power has now been widely adopted in many standards. The newly developed intensity method is suitable for large sources. However, the reverberation room technique can be more economical when measuring small sources, particularly on a repetitive basis.

The precision of the reverberation method can be greatly improved by a rotating diffuser when the source radiates pure tones. Although rotating diffusers have been used for many decades, their qualitative effect on the sound field has been explained only recently (Lubman, Tichy, Waterhouse) and quantitative theory based on the statistical theory of its modulation effects has been formulated and verified in 1980 (Elko, Tichy). This recent work can be particularly useful to determine the efficiency of a diffuser.

Modulation effects and criteria of diffuser efficiency. A rotating diffuser is causing both the frequency and the amplitude modulation of the sound field. As both modulation effects are derived by the same periodicity of the rotation, the resulting AM and FM spectra have their lines at the same frequency. When measuring the spatial average of the $p^2$ in the room, needed to determine the sound power, the microphone signal is rectified and time averaged either at every microphone position or along a continuous path. The resulting $p^2$ amplitudes at each microphone location are then averaged to determine the sound power. When measuring in $N$ uncorrelated points (or averaging along one-half wavelength long continuous path) the variance of $p^2$ average (above the room critical frequency) is $1/N$. The effect of a rotating diffuser consists in reducing this variance to $1/N.M$, where $M$ is Figure of Merit of the rotating diffuser. The determination of $M$ cannot be done with high precision (Elko).

The new approach to the rotating diffuser evaluation is based on statistical analysis of the modulation envelope of the AM spectra at different spatial locations. Both the magnitude and phase of each spectral line vary from point to point and the signal at a random room position can be expressed as a summation of many randomly phased discrete frequency components. It was experimentally verified that the process is narrow band. The statistical theory of the rotating diffuser leads also to the conclusion on maximum possible precision improvement.
Statistical Distribution of the Pressure Envelope. The amplitude of the sound pressure in a diffuse sound field has a Rayleigh distribution while the \( p^2 \) is distributed exponentially. When a rotating diffuser modulates the sound field the envelope function (Fig.1) of the AM varies with location. Its statistical distribution can be derived using a random process model based on random flights as developed by Kluiver. He found that a magnitude variable \( r \) of \( N \) components with different amplitudes \( a_m \) has a cumulative distribution \( F_n(r) \)

\[
F_n(r) = \frac{r^j}{j} \frac{J_1(rt)}{0} \prod_{m=1}^{N} \frac{J(a_{m}t)}{0} \\
(1)
\]

where \( J_n \) are the Bessel functions of the order \( n \). Approximate solution of the integral in Eq. (1) was found by Rayleigh, Slack and Lyon while Bénet found the solution for all components equal. Elko worked out the solution for unequal components and obtained

\[
F_n(r) = 1 - \frac{2}{A} \sum_{n=1}^{N} a_n \frac{J_n(a_{i}j/\pi)}{i=1} \frac{J_n(a_{m}j/\pi)}{m=1} \frac{1}{n} \frac{J_n(a_{i}j/\pi)}{i=1} \frac{J_n(a_{m}j/\pi)}{m=1} \\
(2) \quad \frac{A}{\pi} = \sum_{m=1}^{N} a_m
\]

where \( j_n \) is the \( n \)-th root of the Bessel function of zero order and \( A \) represents the sum of all the components. The amplitude of the components of the AM spectra vary from point to point and for each frequency from an ensemble. To apply the above distribution on the rotating diffuser we will represent its modulation effects by the spatial average of its spectral frequency components. An example of such an averaged spectrum is in Fig.2 for the carrier frequency 4kHz and the diffuser rotating at 5 rps. (The measurements were performed on a scale model). The cumulative distribution of the pressure was calculated from Eq.(2) substituting for the components \( a \) the space averaged amplitude of the spectra. Fig.3 shows both the theoretically predicted and the measured values of \( F_n(p) \). The agreement is very good. Measurements and calculations using different rotational speed resulted in generally good agreement.

When using the rotating diffuser it is desirable to achieve as deep AM modulation as possible, because the time averaged pressure amplitude will then vary less with microphone location.

Finding the limit from Eq.(2) for a narrow band process and an infinite number of spectral components the cumulative distribution converges to the Rayleigh, which has always a greater variance (0.273) than a distribution from a limited number of components. This means that the diffuser design goal should be to generate as many spectral lines as possible, however once the diffuser offers significant numbers of freedom the distribution converges quickly to the Rayleigh.

The experimental evidence reveals that except for the frequency intervals the amplitude envelope spectra are independent of the diffuser rotational speed for the same carrier frequency. This means that increasing the speed does not result in better diffuser performance and the choice of the diffuser speed is effected mainly by practical considerations such as mechanical stability, noise generated by air flow and the time needed for data acquisition because at each point the sound pressure has to be averaged over the entire diffuser revolution.
Relationship between the normalized Pressure-squared Variance and the Rotating Diffuser Speed. One of the important effects for applications is the reduction of the spatial $p^2$ variance by a rotating diffuser. Both the theoretical predictions and experimental results show that the major effect of the diffuser speed is in changing the sideband spacing. The key factor to determine whether the diffuser speed has some effect on the variance magnitude is the correlation between the frequency lines of the sidebands. Schroeder derived the expression for the frequency correlation $\rho$ between two tones with frequencies $f_j$ and $f_k$ exciting a room with fixed boundaries

$$\rho(f_j - f_k) = \left[ 1 + (0.455 T_{60} (f_j - f_k))^2 \right]^{-1}$$

where $T_{60}$ is the room reverberation time.

The statistical model presented above assumes that the sidebands are independent. Because the diffuser causes time varying perturbation of the room geometry it invalidates the time-invariant assumption of Schroeder. Generally the normalized spatial variance of a multimodal sound field in a room is given by

$$\sigma^2 = \frac{N}{N} \sum_{j=1}^{N} \sum_{k=1}^{N} \sum_{m=1}^{m} a_j a_k \rho(f_j - f_k) / (\sum_{m=1}^{m} a)^2$$

where in our case $a$ is the space averaged amplitude of the pressure squared of the $m$-th sideband and $\rho(f_j - f_k)$ is the frequency correlation function given by Eq. 3.

Fig. (4) shows the results of calculations of $\sigma^2$ using Eq. 4 and experimentally determined variance from the measurements in 25 uncorrelated points. Curve b represents the variance prediction assuming $\rho=1$ for $f_j=f_k$ and $\rho=0$ if $f_j \neq f_k$, which means that the spectrum lines are considered uncorrelated. The agreement with the measured data (curve a) is very good. On the other hand, if the variance is calculated by substituting into Eq. 4 the values of $\rho$ obtained from Eq. 3 (curve a) the variance becomes speed dependent as the correlation increases with decreasing speed because the frequency intervals between the spectral lines are becoming smaller. This proves that the lines are uncorrelated which is another verification of the assumptions used to develop the statistical model of the rotating diffuser.

In a limiting case if the diffuser is stopped the multimodal field will converge into single excitation frequency and Eq. 4 yields $\sigma^2 = 1$. The theoretical curve b therefore exhibits a discontinuity for 0 fps as it corresponds to the qualitative change of the boundary conditions.

Conclusion. The developed theory allows the comparison of different diffusers and determination to what degree they achieved the limits of efficiency. We can conclude from our other research that the diffuser size (measured in wavelengths) is essential for a deep modulation, however, once a certain size is achieved further increase may cause more mechanical complications without substantial additional acoustical benefits.

Fig. 1  Sound pressure variation with diffuser rotation

Fig. 2  Measured spatially averaged spectrum at 4 kHz and rotating diffuser rotating at 4 rps

Fig. 3  Cumulative distribution function of the pressure envelope for rotating diffuser rotating at 5 rps
   a) Predicted from Eq. 2
   b) Measured
   c) Rayleigh distribution

Fig. 4  Normalized $p^2$ variance as a function of rotating diffuser speed
   a) Measured
   b) Predicted for sidebands uncorrelated
   c) Predicted with sidebands correlated
SCHALLDIFFUSITÄT UND IHR EINFLUß AUF DIE MESSUNG DES ABSORPTIONSSTADES MIT DER HALLRAUMMETHODE.

ANDERT Tomas

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Summary
Values of the absorption coefficient $\alpha$ obtained in different reverberation chambers are not very reproducible and sometimes even larger than unity. Using a directional microphone with a parabolic mirror, we found differences between the directional distribution of sound in an empty reverberation chamber and in one with some absorbers as well as a lack of diffusivity in both cases. This means that the evaluation of $\alpha$-value is usually based on a comparison of reverberation times which are not only affected by different absorption, but also by different energy storage capabilities of the reverberation chamber. Thus, this approach becomes questionable. An improvement is possible if the reverberation times determined with a known absorber and with the absorber under investigation are compared directly.

Stationäre Schallrichtungsverteilung im Hallraum
SRV-en in Fig. 1 and 2 /2,3/ zeigen, daß der Hallraum im angeregten Zustand nicht diffus ist. Jedes Maximum in der SRV läßt sich auf einen nicht genügend gestreuten, den geometrischen Reflexionen folgenden, Schallstrahl zurückführen.
Diese Erkenntnisse wurden auch im Hallraum des Forschungsinstituts für Aufnahme- und Wiedergabetechnik in Prag (beschrieben in /1/) bestätigt. Einen einfachen Weg zur Verbesserung der SRV im angeregten Zustand stellt die Verwendung von mehreren Quellen dar. Die Experimente in Prag zeigten, daß

Fig. 1 und 2  SRV im leeren bzw. mit 11m$^2$ Absorber belegten Hallraum gemessen mit einem Parabolspiegellichtrichtmikrofon
die bei mehreren Quellen resultierende SRV diffuser ist als diejenige der einzelnen Quellen.
Schallrichtungsverteilung im Nachhall
Die stationäre SRV wird wegen der ungenügend gestreuten Schallstrahlen stark von der Quelle aus geprägt, dagegen wird die SRV im Nachhall von der Quelle wenig beeinflusst, da die Direktschallanteile rasch abklingen /37/. Ein Absorber stellt eine Energiesenke dar, die auch während des Nachhalls die SRV beeinflusst.

Fig. 3 und 4 Schallintensität im Nachhall mit 11,25 m² geschlossen am Boden bzw. auf fünf Wände verteilt angebrachten Absorber
Die Fig. 3 und 4 zeigen die Intensitäten im Nachhall, die in einem gedachten Würfel aus den Richtungen der 6 Flächen und der 8 Ecken einfallen, für zwei Extremfälle der Anbringung der Absorber. Zuerst wurde der Absorber (11,25 m²) geschlossen am Boden angebracht. Der geringere Schalleinfall aus der Richtung des Absorbers in Fig. 3 blieb während des gesamten Nachhalls vorhanden. Der Anstieg des Schalleinfalls aus der Richtung der vertikalen Wände ist einerseits auf die geringe Absorption der tangential zum Absorber verlaufenden Wellen zurückzuführen. Andererseits hingen aus Sicherheitsgründen in dieser Ebene keine Diffusoren, die für die Streuung der wenig gedämpften Moden gesorgt hatten. In Fig. 4 wurden 15 Absorber (ά 0,75m²) auf alle Wände und den Boden verteilt. Hier wird rascher endgültige Verteilung des Nachhalls erreicht (nach ca. 0,3 sek.), die weniger von den stationären Intensitäten bei t=0 abweicht. Ein auffälliger Anstieg aus den Wandrichtungen ist ebenfalls vorhanden, er ist jedoch wegen der Verteilung der Absorptionsfläche nicht so stark ausgeprägt. Die Nachhallzeiten in den Wandrichtungen liegen über denjenigen in anderen Richtungen.

Diffusität und Absorptionsgrad
Fig. 5 gibt die bei unterschiedlicher Diffusität gemessenen Absorptionsgrade η wieder. Die kleinste Diffusität im stationären Zustand wurde bei der Kurve 1 beim stark ausgeprägten Direktschall, wenn die Quelle im Sichtkontakt zum Absorber stand, realisiert. Die Kurve 1 zeigt einen ungleichmäßigem Verlauf, der auf die häufig auftretenden Knickel in den Nachhallkurven zurückzuführen ist. Hier wurden die realistischsten η-Werte gemessen. Dies ist ersichtlich aus ihrem Vergleich mit Kurve 4, die für senkrechten Schalleinfall im reflektionsfreien Raum bestimmt wurde. Bei verbesserter Diffusität hatte die Quelle keinen Sichtkontakt zum Absorber und es wurden zwei verschiedene Anordnungen der Diffusoren untersucht. Zunächst wurden sie nach allgemein an-
Bedingungen der Diffusität erkannten Empfehlungen (u.a. DIN 52212) angebracht. Die α-Werte in Kurve 2 verlaufen trotz der starken Abweichung der gemessenen SRV von idealer Diffusität gleichmäßig. Sie liegen zum Teil über Eins, was auch den Herstellerangaben entspricht. Anschließend wurde die Lage und die Anzahl der Diffusoren so verändert, daß am Boden eine gegenüber dem vorherigen Zustand bessere SRV entstand. Das Verhältnis der maximalen zur minimalen Intensität wurde von 1:0,10 (Fig. 1) auf 1:0,25 verbessert. Trotzdem sind die damit gewonnenen α-Werte (Kurve 3) wenig realistisch und zeigen einen deutlichen Abfall für Frequenzen über 2 kHz. Dies widerspricht den physikalischen Eigenschaften eines faserigen Absorbers. Das α eines offensichtlichen Absorbers bleibt in diesem Frequenzbereich, wie auch die α-Messung für senkrechten Schalleinfall bestätigt, konstant. Der Abfall bei Frequenzen über 2 kHz ist auf die Ausbildung von wenig gedämpften Moden im Bereich über dem Absorber zurückzuführen, wie bereits bei der Messung der Schallintensitäten im Nachhall gezeigt wurde. Die ungünstige Meßanordnung führte zu einem relativ langsamen Abbau der Energie, zu einer "zu langem" Nachhallzeit und einem "zu kleinen" nicht realistischen α. Auch die Berücksichtigung des Kannten-effektes verringert nicht den Streubereich der α-Werte.


Angenommen sei ein ebenses Modell nach Fig. 6, in dem eine Wand gegenüber den anderen Wänden stark absorbierend ausgebildet ist. Herrscht in diesem Modell eine ungleichmäßige SRV wie in Fig. 6 Fall a, in der die absorbierende Wand bevorzugt beschaltet wird, so wird die resultierende Nachhallzeit kürzer sein als im Fall c, da im Fall a die gespeicherte Energie schneller
abgebaut wird als im Fall c. So ist es möglich, im Falle einer ungleichmäßigen SRV ein zu geringes, zu großes oder auch das der Definition entsprechende $\alpha$ zu messen. Es wird an dieser Stelle absichtlich von der SRV und nicht von der Diffusität gesprochen, da es auf die Orientierung der SRV gegenüber dem Absorber ankommt.

Für die $\alpha$-Bestimmung läßt sich ein elektrisches Ersatzschaltbild in Fig. 7 aufstellen, wobei die Widerstände $R_a$ und $R_b$ aus den Zeitkonstanten bestimmt werden. Wenn die Schaltung nach Fig. 7 falsche Ergebnisse liefert, und dies ist für $\alpha > 1$ der Fall, dann liegt es entweder an falscher Messung der Zeitkonstanten oder daran, daß die Kapazität nicht konstant ist. Die Kapazität $C$ ist im Modell das Speichersedium, und sie entspricht in Wirklichkeit dem Speichervermögen des Hallraums. Für den Hallraum wird in beiden Messungen das gleiche Speichervermögen vorausgesetzt in der Annahme, daß zu jedem Zeitpunkt ein diffuses Schallfeld herrscht.

Bei der Hallraummethode wird ein konstantes Speichervermögen des Hallraumes auch dann erreicht, wenn in beiden Fällen die gleiche SRV herrscht. Man sollte also nicht versuchen den Hallraum unabhängig vom untersuchten Gegenstand diffus zu halten, was für stark absorbierebene Materialien per Prinzip nicht möglich ist. Stattdessen seien man den Hallraum mit einem bekannten in gleicher Grobeneinheit liegenden Absorber. Diese Methode kommt für große $\alpha$-Werte auch dem Sabineschen Modell sehr nahe, in dem der Absorber mit einem offenen Fenster verglichen werden. Zur Eichung können die Ergebnisse der aufwendigen Ringversuche /1/ sehr gut verwertet werden. Der im Ringversuch untersuchte Absorber kann, nachdem sein genaues $\alpha$ z.B. mit der Reflexionsmethode bestimmt wurde, als Eichabsorber für die Hallräume herangezogen werden. Weitere Vergleichsmessungen würden dann, wenn man $\alpha$ aus dem Vergleich mit dem Eichabsorber bestimmt, reproduzierbare Werte ergeben.

Zusammenfassung
Für eine gute Diffusität im stationären Zustand soll der Hallraum über eine ausreichende Anzahl von gekrümmten Diffusoren verfügen, die die Sichtverbindung zwischen der Quelle und jedes der untersuchten Orte verhindern. Die Wände des Hallraumes sollen konvex schallstreuend wirken.

Die relative Schallintensität ist während des Nachhalls großen Änderungen unterworfen und starke Absorber verursachen auch während des Nachhallsverzögerung Einbrüche in der SRV. Raumrichtungen, die nicht durch Diffusoren oder andere Maßnahmen aufgelockert werden, verzerrn nachhaltig das Schallfeld und wirken verlängernd auf die Nachhallzeit.

Die in Hallräumen erzielbaren Verbesserungen der stationären Diffusität reichen nicht für eine zuverlässige $\alpha$-Messung aus. Unzureichende Diffusität ist die Ursache für eine geringe Reproduzierbarkeit der in verschiedenen Hallräumen bestimmten $\alpha$-Werte. Die nicht erfüllende Bedingung, daß im Hallraum zu jedem Zeitpunkt ein diffuses Schallfeld herrschen muß, kann geändert werden. Es muß bei beiden Messungen, die der Berechnung des $\alpha$-Wertes dienen, die gleiche SRV herrschen.

A METHOD OF DETERMINING SOUND ENERGY RADIATED BY IMPULSIVE SOUND SOURCES

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1. INTRODUCTION

Concerning the sound outputs of stationary sound sources, the term of sound power level is defined and its measuring methods have been established. (For example, ISO 3740-3746) While, the way of specifying the sound energy emitted by impulsive (or intermittent) sound sources is not physically clarified, and, therefore, its measuring methods have not been investigated.

So, in this paper, a way of quantitatively specifying the sound energy emitted by impulsive sound sources is proposed, and the measuring methods are investigated.

2. THE SOUND ENERGY LEVEL

For the stationary sound sources, the sound power level is defined as,

$$L_W = 10 \log \frac{W}{W_0}$$  \hspace{1cm} (1)

where, \( W \) : the sound power of the sound source (in watts)
\( W_0 \): the reference sound power (10^{-12} W)

While, as for the output of the impulsive sound source, the sound power which has the concept of time averaging is no longer applicable and the total sound energy should be considered.

Here, by introducing the concept of the instantaneous sound power \( w(t) \), the total sound energy \( E_{total} \) is expressed as,

$$E_{total} = \int_0^\infty w(t) dt \hspace{1cm} (2)$$

In accordance with the sound power level, the total sound energy of an impulsive sound source can be expressed in decibels as follows. In this paper, this quantity is termed the sound energy level.

$$L_E = 10 \log \frac{E_{total}}{E_0}$$  \hspace{1cm} (3)

or

$$L_E = 10 \log \left[ \frac{1}{T_0} \int_0^\infty w(t) dt / W_0 \right]$$  \hspace{1cm} (4)

where, \( E_0 \) : the reference sound energy
\( T_0 \) : the reference time (1s)

Fig. 1 Total sound energy emitted by an impulsive sound source.
The sound energy level represents the level of the total sound energy of a single event emitted by an impulsive sound source, and, as shown in Fig. 1, it corresponds to the sound power level of the stationary sound source of 1 second duration time which has the same total energy as the impulsive sound source. Like the sound power level, the sound energy level can be also defined in each 1/3 or 1/1 octave band, or as the A-weighted value.

3. MEASURING METHODS OF THE SOUND ENERGY LEVEL
In order to measure the sound energy level above defined, there are two methods as follows, like the sound power level measurement.

(1) Free Field Method
When an omnidirectional impulsive sound source is set in the free sound field, the relation between the instantaneous sound power \( w(t) \) of the sound source and the sound pressure \( p(t) \) or the sound intensity \( I(t) \) at a point of \( r \) meters away from the sound source is expressed as follows.

\[
\frac{w(t)}{4\pi r^2} = I(t - \frac{r}{c}) = \frac{p^2(t - \frac{r}{c})}{\rho c}
\]

(5)

where, \( c \) :the sound velocity

accordingly,

\[
10 \log \left[ \frac{1}{4\pi r^2} \int_0^\infty \frac{w(t)}{w_0} \, dt \right] = 10 \log \left[ \frac{1}{T_0} \int_0^\infty \frac{p^2(t)}{P_0^2} \, dt \right]
\]

(6)

where, \( p_0 \) :the reference sound pressure (2 \times 10^{-5} \text{ Pa})

Following the definition of the sound exposure level (ISO 1996/1), by representing the right hand of equation (6) by \( L_{pE} \),

\[
L_{pE} = 10 \log \left[ \frac{1}{T_0} \int_0^\infty \frac{p^2(t)}{P_0^2} \, dt \right]
\]

(7)

Then,

\[
L_E = L_{pE} + 20 \log r + 11
\]

(8)

Equation (8) means that the sound energy level is obtained by measuring \( L_{pE} \). (In case that the measurement is carried out in the semi-free field, or the sound source has directivity, proper corrections should be made in the same manner like in the case of the sound power level measurements.)

(2) Diffused Sound Field Method
When the impulsive sound source is set in the diffused sound field, the relation between the instantaneous sound power \( w(t) \) of the sound source and the sound energy density of the sound field \( E_d \) is expressed as,

\[
w(t) = V \cdot \frac{d E_d(t)}{dt} + \frac{c \cdot E_d(t)}{4} A
\]

(9)

or

\[
\int_0^t w(t) \, dt = V \cdot E_d(t) + \frac{c \cdot A}{4} \int_0^t E_d(t) \, dt
\]

(10)

where, \( V \) :the air volume of the sound field (reverberant room)

\( A \) :the total sound absorption power of the sound field

Since \( E_d(t) \) becomes 0 when \( t \) becomes infinity,

\[
\int_0^\infty w(t) \, dt = \frac{c \cdot A}{4} \int_0^\infty E_d(t) \, dt
\]

(11)
H. TACHIBANA:
A Method of Determining Sound Energy Radiated by Impulsive Sound Sources.

In the diffused sound field, there exists the following relation between the sound energy density \( E_d \) and the sound pressure \( p \),
\[
E_d = \frac{p^2}{\rho c^2}
\]  
(12)

then,
\[
\int_0^{\infty} w(t) \, dt = \frac{A}{4 \rho c} \int_0^{\infty} p^2(t) \, dt
\]  
(13)

By changing this equation into level expression, the following equation can be derived,
\[
L_{PE} = L_{PE} + 10 \log \frac{A}{4}
\]  
(14)

That is, the sound energy level of an impulsive sound source can be obtained by measuring \( L_{PE} \) values in the diffused sound field.

4. PRELIMINARY EXPERIMENTS
In order to confirm the measuring principles mentioned above, the following two preliminary experiments were carried out.

As the first experiment, the relation expressed by equation (14) was examined. In this experiment, 1 octave band noise burst signals with various duration time \( \Delta T \) were radiated in a reverberant room of 200m\(^3\) volume by use of an omnidirectional loud-speaker system composed of twelve loud-speakers, and, by changing \( \Delta T \) from 0.125s to 1s, \( L_{PE} \) values were measured at five points in the reverberant room. (In the measurement, a square-integrating circuit was used.) The measured results at five points were averaged for each duration time.

As an example of the experimental results, Fig.-2 shows the differences between \( L_{PE} \) and \( L_p \) in 1/1 octave band centered on 1kHz plotted against the duration time \( \Delta T \) of the burst signal. Here, \( L_p \) represents the sound pressure level of the stationary band noise which has the same amplitude as the band noise burst signals mentioned above.

This result confirms that \( L_{PE} \) increases by 3dB every doubling \( \Delta T \), and it agrees with \( L_p \) when \( \Delta T = 1s \).

As the second experiment, the sound energy level of an artificial impulsive sound source was measured by using both of the free field method and the diffused sound field method. As the sound source, the impulse-responses of 1/1 octave band pass filters were radiated from the omnidirectional loud-speaker system in an anechoic room and in the reverberant room, and \( L_{PE} \) values were measured at five points in each room, respectively. The results measured in each room were averaged, and \( L_E \) values were calculated from these \( L_{PE} \) values according to equation (8) and (14).
Fig. 3 shows the comparison between the sound energy level of the same artificial impulsive sound source obtained by the two measuring methods, and it can be seen that these two kinds of results are in fairly good agreement.

5. EXAMPLES OF THE SOUND ENERGY LEVEL MEASUREMENTS

Following the basic investigations mentioned above, the sound energy levels of the actual impulsive sound sources were measured practically by the diffused sound field method, by using the reverberant room.

Fig. 4 shows the measured results of five kinds of impulsive sound sources. In these measurements, the burst sounds received in the reverberant room were once recorded on a magnetic tape recorder, and, then, 1/1 octave band spectra and the over all values (A and C weighted) were analyzed.

![Diagram showing sound energy levels of five kinds of impulsive sound sources.](image)

Fig. 4

The sound energy levels of five kinds of impulsive sound sources, measured by the diffused sound field method.
ESTIMATION "IN SITU" DE LA PUISSANCE ACOUSTIQUE

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1° Introduction

La prédiction des niveaux de bruits produits par des machines évoluant en site extérieur requiert la connaissance de différents paramètres relatifs aux sources de bruit.

La directivité et la puissance acoustique émise lors de conditions réelles de travail sont deux informations de première importance.

En ce qui concerne la puissance acoustique des machines, c'est une donnée qui est très souvent connue pour des cycles de fonctionnement spéciaux mais, rarement pour des conditions de travail réel.

Quant à la directivité, là, on peut raisonnablement affirmer que sa connaissance est loin d'être parfaite et précise.

Cet article décrit une méthode d'estimation de la puissance acoustique émise par une source placée sur un site réel et exploite le phénomène d'interférence acoustique résultant de la réflexion des ondes sonores sur le sol.

2° Influence du sol

Le modèle simple de décroissance du niveau sonore

\[ L_p = 10 \log \frac{QW}{4\pi r^2} \]  (I)

où \( L_p \) = niveau de pression;
\( W \) = puissance de la source;
\( r \) = distance de la source,

ne tient pas compte des phénomènes d'interférence entre onde directe et onde réfléchie sur le sol.
La figure 1 illustre ce phénomène.

![Figure 1](image)

Et, une formulation plus correcte conduit à l'expression suivante :

\[ L_p = L_w + 10 \log \left( \left(1/r^2\right) + \left(\left|C_r\right|^2/r^2\right) + 2\left(\left|C_r\right|/r'\right) \cos(k(r-r') + \phi_r) \right) - 11 \]  

(2)

Ce modèle conduit à des courbes valables pour des sons purs. Dans la pratique, les mesures s'effectuent dans des bandes de fréquences relativement larges (1/3 d'octave ou octave). Le modèle (2) peut être directement étendu à des bruits larges bandes. Ainsi, l'expression (3) est valable pour des bruits dont le spectre fréquentiel est compris entre 2 fréquences \( f_1 \) et \( f_2 \).

\[ L_p = L_w + 10 \log \left[ \left(1/r^2\right) + \left(\left|C_r\right|^2/r'^2\right) + \left(\left|C_r\right|/r'^{\prime}\right)(f_2 - f_1) \right] \]

\[ (\sin(k_2(r-r') + \phi_2) - \sin(k_1(r-r') + \phi_1)) \]  

- 11

(3)

La courbe de la figure 3 résulte de cette formulation.

3° Calcul d'un indicateur de la puissance acoustique

Le dispositif utilisé pour l'estimation de la puissance est schématisé à la figure 2.

![Figure 2](image)
A partir des 2 niveaux de pression mesurés $L_{P1}$ et $L_{P2}$ respectivement à des distances de $d_1$ et $d_2$; on obtient un indicateur de la puissance en ajustant aux sens de moindres carrés, la courbe de décroissance calculée correspondant à la situation envisagée.

La figure 3 montre l'ajustement de la courbe à 2 points de mesure.

![Figure 3](image)

L'analyse des résultats numériques correspondant à cette fréquence montre la qualité de l'ajustement.

$f = 500$ Hz; $L_{P1} = 65.6$ dB; $L_{P2} = 57$ dB; $L_w = 82.3$; $S_m = 1.9$ dB².

En effet, le facteur $S_m$ qui représente le critère à minimiser lors de l'ajustement (somme des carrés des différences) reste à une valeur très modérée.

4° Précision de l'estimation

Le test a été répété sur un certain nombre de machines réelles et les valeurs de puissance comparées avec des valeurs obtenues par une méthode normalisée (ISO 3744) présentent des différences ne dépassant pas 2 dBA.

5° Remarque sur la directivité des sources

La présente méthode amène à formuler les remarques suivantes :

- La directivité d'une source est un paramètre tridimensionnel; il existe une directivité dans le plan vertical et dans le plan horizontal;

- La directivité dans le plan vertical est surtout influencée par les réflexions sur le sol et est prise en compte dans notre approche, quant à la directivité dans le plan horizontal, on peut l'estimer, en réalisant des mesures suivant plusieurs directions autour de la source.
6° Conclusions

La méthode développée permet de mesurer sur place un estimateur de la puissance acoustique des machines et, permet ainsi de résoudre le problème de la connaissance de paramètres acoustiques des sources lorsque l'on est confronté à des problèmes de prédiction de niveaux sonores en espace ouvert.

Références


UTILISATION D'UN SIGNAL TRANSITOIRE DE DETONATION A LA
DETERMINATION DE L'EFFICACITE D'UN ECRAN EN PLACE.

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Dans le fascicule du Guide du Bruit relatif aux ouvrages de protection [1], des recommandations relatives aux caractéristiques acoustiques (transmission et absorption) ont été faites sur la base de valeurs déterminées à partir d'essais effectués en laboratoire. Ces procédures d'essais plus représentatives des conditions de mise en place des matériaux dans les bâtiments donnent des indications utiles lors de la conception du projet. Cependant, on note parfois des écarts significatifs entre les caractéristiques mesurées in-situ et les valeurs mesurées en laboratoire. Pour la transparence, les causes essentielles en sont les suivantes :

- influence de la dimension de l’échantillon et des conditions de montage
- mauvaise mise en œuvre
- défaut de jonction entre module ou entre éléments constitutifs des modules.

En ce qui concerne l'absorption, il nous semble que les procédures actuelles aboutissent à des résultats dont on cerne mal la relation (\( \alpha_{KUNDT}, \alpha_{SABINE} \)) et qui, dans leur principe pour les écrans, sont assez peu représentatifs de la réalité.

Ces diverses raisons nous ont conduit à étudier une méthode de mesure in-situ des caractéristiques des écrans.

Nous avons choisi de développer une méthode basée sur le traitement de signaux transitoires de détonation; ce choix a été fondé sur les bons résultats obtenus dans les comparaisons théorie/expérience [2] et sur sa facilité de mise en œuvre.

La source de bruit est un pistolet d’alarme de calibre 8 mm muni d’un filtre acoustique ("silencieux") à deux cavités remplies d'absorbant fibreuse et traversé par un tube perforé. Les caractéristiques principales de cette source sont les suivantes :

- durée de l'impulsion de l'ordre de 1 ms
- niveau crête de l'ordre de 130 dB à 1 m
bonne reproductibilité (± 2 dB entre 200 et 3 kHz)
- énergie importante entre 200 Hz et 3 kHz
- omnidirectionnalité à 2 dB dans le plan perpendiculaire au canon.

Description de la méthode

La source impulsionnelle ponctuelle placée en S émet un signal transitoire. Le microphone M₁ reçoit une onde directe p₁(t) et une onde réfléchie pᵣ(t) ; le microphone M₂ reçoit une onde transmise pᵣᵣ(t) et les ondes diffractées par les bords de la structure testée.

Chaque trajet élémentaire de propagation peut être assimilé à un filtre linéaire causal à une seule entrée et une seule sortie.

La durée du signal étant fixée, deux éléments sont à prendre en compte pour permettre la séparation des signaux d'entrée et de sortie par filtrage temporel :
- les distances source, microphone, structure
- les réponses impulsionnelles des filtres élémentaires.

Mesure de coefficient d'absorption

Dans le cas où l'on peut assimiler les surfaces d'ondes à des plans et si le matériau est homogène, le module du coefficient de réflexion est donné par [3].
DELANNE Yves, BERENGIER Michel, Utilisation d'un signal transitoire de détonation à la détermination de l'efficacité d'un écran en place

$$|R(\omega)| = \left| \frac{S_p(\omega)}{S_i(\omega)} \right| \phi(\delta)$$ et $\alpha^2 = 1 - R^2$

avec :

$\phi(\delta)$ : correction de divergence sphérique

$S_p(\omega)$ : transformation de Fourier de l'onde réfléchie

$S_i(\omega)$ : transformation de Fourier de l'onde incidence.

A partir du principe d'Huygens, un calcul démontre qu'en incidence oblique, si la durée du signal réfléchi est $T_2$, la surface "active" est l'intersection entre un ellipsoïde dont les foyers sont la source ($S$) et le microphone ($M_i$) avec le plan absorbant. Pour une incidence normale, lorsqu'il n'y a pas réflexion sur le support de l'absorbant, la surface vue est un cercle de rayon $rm$.

Les ordres de grandeur sont les suivants :

<table>
<thead>
<tr>
<th>(a) incidence oblique 45°</th>
<th>(b) incidence normale</th>
</tr>
</thead>
<tbody>
<tr>
<td>$T_1 = 1.5$ : 2a = 2.20 m; 2b = 1.60 m</td>
<td>$rm = 0,726$ m</td>
</tr>
<tr>
<td>$T_2 = 1$ ms : 2a = 1.75 m; 2b = 1.16 m</td>
<td>$rm = 0,572$ m</td>
</tr>
</tbody>
</table>

La comparaison des valeurs de $|R(\omega)|$ mesurées en incidence normale sur des absorbants fibres avec celle obtenue au tube à ondes stationnaires montre une bonne concordance des résultats $|4|$.

Mesure de la transparence acoustique

Les caractéristiques en transmission sont déduites de la fonction de transfert mesurée entre le signal transitoire transmis et le signal incident, corrigée de la divergence. Les comparaisons théorie/expérience que nous avons faites sur des structures simples ont donné de très bons résultats. C'est le cas notamment pour les valeurs prédites par la formulation de CREMER et celles mesurées sur une plaque d'aggloméré de bois de 2 m x 2 m environ.

Cette méthode s'est avérée très efficace pour détecter les défauts de jonction entre module.
Conclusion

La méthode mise au point a été testée sur 18 écrans routiers en place. Elle s'est avérée parfaitement opérationnelle. Elle permet donc :

- de déterminer la valeur du coefficient d'absorption pour différents indices
- de caractériser la "transparence" intrinsèque des différents éléments constitutifs des modules des écrans
- de caractériser la qualité des jonctions de ces éléments et des modules entre eux.

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6.8

Etalonnage. Sources de référence.
Contrôle

Calibration. References sources.
Monitoring

Eichung. Referenzquellen.
Kontrolle
TENTATIVE D'EXPLICATION DE L'ÉCART SYSTÉMATIQUE OBSERVE ENTRE LES PUISSANCES ACOUSTIQUES MESURÉES EN SALLE ANÉCHOÏQUE ET EN SALLE RÉVERBERANTE

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Introduction
L'objet de ce travail a été de rechercher la cause de l'écart systématique et de signe fixe (ΔK<sub>VA</sub> = ΔK<sub>VE</sub>) existant entre les niveaux de puissance sonore d'une même source, mesurés en salle anéchoïque et en salle réverbérante.

L'étude en salle réverbérante repose sur l'hypothèse selon laquelle la propagation du son dans des locaux clos est correctement décrite, à partir d'une certaine fréquence, par l'acoustique géométrique statistique.

De tous les paramètres physiques en jeu dans la théorie de la réverbération, le plus discutable par son mode d'évaluation est "le libre parcours moyen" (1pm). Un mode nouveau de calcul du 1pm qui consiste à appliquer les lois de la réflexion aux rayons sonores émis par la source, est proposé. Avec la valeur ainsi obtenue, on a calculé la puissance acoustique d'une source à large bande en salle réverbérante, préalablement étalonnée en salle anéchoïque selon la méthode décrite dans la norme AFNOR 31-026, en mesurant le temps de réverbération et les niveaux de pression sonore dans le champ réverbéré.

I - Libre parcours moyen (1pm)
Son calcul a été effectué sur la base de sa définition classique (distance moyenne parcourue par un rayon sonore entre deux réflexions successives sur les surfaces limitant la salle). Pour cela, un parallèle est fait avec l'optique géométrique dans l'application des lois de la propagation et de la réflexion : les ondes sonores sont représentées par des rayons, les parois sont parfaitement réfléchissantes, le rayon incident, la normale à la paroi et le rayon réfléchi sont dans le même plan.

Soit une chambre parallélépipédique (Lx, Ly, Lz) contenant une source sonore placée dans un coin et en fonctionnement depuis un temps suffisamment long pour que l'équilibre entre l'énergie acoustique qu'elle émet et celle qui est absorbée par les parois soit établi. On calcule les distances parcourues par les rayons entre la 1<sup>ère</sup> et 2<sup>ème</sup> réflexion, aucun ordre n'étant privilégié. "1pm" est la valeur moyenne de ces distances. Trois cas possibles de réflexions sont rencontrés :
1) Réflexions sur le plafond P<sub>plp</sub>.
2) Réflexions sur la paroi en bout P<sub>prp</sub> (x = Lx)
3) Réflexions sur la paroi latérale P<sub>plp</sub> (y = Ly).
Lorsqu'on considère les réflexions sur le plafond (figure 1), on identi-
fie quatre parcours possibles :

a : rayons qui ont subi leur 1re réflexion sur le plafond et la 2ème sur
le plancher ; on les appelle $R_{a II}$

$a'$ : rayons qui ont subi leur 1re réflexion sur le plafond et la 2ème sur
la paroi $y = Ly$ ; on les appelle $R_{a III}$

Ces deux familles de rayons sont contenues dans des plans qui font avec
l'axe $x$ un angle $\Phi_x$, tel que : $0 \leq \Phi_x \leq \text{arcctg} \frac{Ly}{Lx} \Rightarrow \Phi_x = \frac{1}{2} \text{arcctg} \frac{Ly}{Lx}$

Les parcours moyens pour ces cas sont :

$$R_{a II} = \frac{Lz}{\cos \theta_z} \text{ avec } \theta_z = \frac{1}{2} \text{arcctg} \frac{Lx}{2L_z \cos \Phi_x}$$

et

$$R_{a III} = \frac{Lx}{\cos \Phi_x \sin \theta_z} + \frac{Lz}{\cos \theta_z} \text{ avec } \theta_z = \frac{1}{2} \left[ \text{arcctg} \frac{Lx}{2L_z \cos \Phi_x} + \text{arcctg} \frac{Ly}{Lx} \right]$$

b : rayons qui ont subi leur 1re réflexion sur le plafond et la 2ème sur
le plancher ; on les appelle $R_{b II}$

$b'$ : rayons qui ont subi leur 1re réflexion sur le plafond et la 2ème sur
la paroi $x = Lx$ ; on les appelle $R'_{b II}$

Ces deux familles de rayons sont contenues dans des plans qui font avec
l'axe $x$ un angle $\Phi_x$, tel que : $\text{arcctg} \frac{Ly}{Lx} \leq \Phi_x \leq \frac{\pi}{2} \Rightarrow \Phi_x = \frac{1}{2} \left[ \text{arcctg} \frac{Ly}{Lx} \right]$.

Les parcours moyens pour ces deux cas sont :

$$R_{b II} = \frac{Lz}{\cos \theta_z} \text{ avec } \theta_z = \frac{1}{2} \text{arcctg} \frac{Ly}{2L_z \sin \Phi_x}$$

et

$$R'_{b II} = \frac{Lx}{\sin \Phi_x \sin \theta_z} + \frac{Lz}{\cos \theta_z} \text{ avec } \theta_z = \frac{1}{2} \left[ \text{arcctg} \frac{Ly}{2L_z \sin \Phi_x} + \text{arcctg} \frac{Ly}{Lx \sin \Phi_x} \right]$$

On en déduit le parcours moyen pour les rayons réfléchis sur le plafond :

$$\overline{R}_{II} = \frac{1}{4} \left( R_{a II} + R'_{a II} + R_{b II} + R'_{b II} \right)$$

On fait le même raisonnement pour les réflexions sur les parois $x = Lx$ et
$y = Ly$, ce qui conduit à :

$$I_{pm} = \frac{S_{PL} R_{II}^{PL} + S_{PHR} R_{II}^{PHR} + S_{LH} R_{II}^{LH}}{S_{PL} + S_{PHR} + S_{LH}}$$

(1)

II - Puissance sonore d'une source mesurée en salle réverbérante

La puissance sonore mesurée en salle réverbérante est, en accord avec
l'acoustique statistique :

$$W = \frac{P_{\text{REV}}}{4 \rho_c}$$

A est l'aire d'absorption équivalente qui s'exprime en fonction du
temps de réverbération $T_R$ et de la valeur du $I_{pm}$ fournie par l'équation (1)
AULETTA Nélida - PUissance acoustique et libre parcours moyen

\[ A = S_{\text{TOT}} \left[ - \ln (1-\alpha) \right] = 0.04 \frac{1}{T_a} \frac{R_m}{T_a} S_{\text{TOT}} \quad (3) \]

\( S_{\text{TOT}} \) étant la surface totale des parois du local.

Des équations (2) et (3), on tire :

\[ W = \frac{P_{\text{REV}}}{4 \rho C} \times 0.04 \frac{1}{T_a} \frac{R_m}{T_a} S_{\text{TOT}} \quad (4) \]
d'où :

\[ L_w = 10 \log \left( \frac{S_{\text{TOT}}}{V_0} \right) + 10 \log \left( \frac{R_m}{T_a} \right) - 10 \log \left( \frac{T_a}{T_0} \right) - 10 \log \left( \frac{C}{C_0} \right) + 10 \log \left( \frac{1 + \lambda S_{\text{TOT}}}{B V} \right) - 20 \quad (5) \]

avec \( V_0 = 1 \text{ m}^3 \), \( T_0 = 1 \text{ sec} \), 10 \( \log (1 + \lambda \frac{S_{\text{TOT}}}{B V}) \) étant le terme correctif de Waterhouse (Norme AFNOR 31-022).

Le tableau 1 résume les résultats expérimentaux de mesure de la puissance acoustique d'une source en salle anéchoïque (Norme AFNOR 31-026) et en salle réverbérante pour le libre parcours moyen préconisé par la Norme AFNOR 31-022 et pour celui fourni par l'équation (1).

III - Conclusion

Les résultats obtenus permettent de conclure qu'à partir de la nouvelle expression proposée pour le libre parcours moyen, on réduit l'écart systématique traditionnellement observé entre les puissances acoustiques mesurées en salle anéchoïque et en salle réverbérante. De plus, cet écart est maintenant de signe variable avec la bande de fréquences considérée. Ceci donne à penser qu'un biais introduit par le libre parcours moyen préconisé par la norme a été éliminé.

Bibliographie
1 - BARKECHL M. - Régime sonore d'une salle après l'extinction de la source - Acustica (1951), 59-74.
2 - GOMPETS C. - Do the classical reverberation formula still have a right for existence ? Acustica 16 (1965-1966), 255-268.
Figure 1 : Illustration des quatre familles de rayons sonores subissant une première réflexion par le plafond

<table>
<thead>
<tr>
<th>Fréquences centrales (Hz)</th>
<th>$L_{W_{\text{RNECH,}}}^{\text{dB}}$</th>
<th>$L_{W_{\text{pm}}}^{\text{(dB)}}$</th>
<th>$L_{W_{\text{norm.}}}^{\text{(dB)}}$</th>
<th>$L_{W_{\text{pm}}}^{\text{(dB)}} - L_{W_{\text{RNECH,}}}^{\text{dB}}$</th>
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Tableau 1 : Niveaux de puissance sonore moyens de la source SSR-2000 par bandes de tiers d'octave
THREE DIMENSIONAL ACOUSTIC INTENSITY MEASUREMENTS WITH INDUSTRIAL ROBOTS

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1 INTRODUCTION

The determination of the sound intensity in the near field is based on the simultaneous estimation of sound pressure and particle velocity in one point, the product of both quantities yielding the acoustic intensity. Early attempts to physically measure the particle velocity were only partially successful due to instrumentation problems. However the availability of reliable digital FFT analysers renewed the interest in near field intensity measurements since the intensity can be formulated in terms of cross-spectral densities. This idea was first introduced by Pahy (ref. 1) and Chung (ref. 2) and provides the base of the work presented here. The present paper deals with automated three dimensional acoustic intensity measurement procedures. Two systems based on industrial robots will be presented, one oriented towards research experiments, and one developed for the estimation of large intensity patterns such as those radiated by vehicles.

2 PRINCIPLES

For our applications the acoustic intensity approach has been retained over the surface intensity method for its superior flexibility and since it allows intensity measurements as well in the far as in the near field.

The measurement procedure is similar to the one presented by Chung and Pope (ref. 2) and is based on the measurement of the cross-spectrum between two closely spaced microphones. Indeed it can be proven (ref. 3) that the acoustic intensity is proportional to the imaginary part of the mentioned cross-spectrum. The circuit switch Chung proposes has been replaced by a calibration procedure whereby each measured cross-spectrum is on line phase corrected with a calibration curve stored in the memory of the FFT analyser. 1/4 inch condenser microphones have been selected as transducers. The microphone spacing is adjustable in accordance with the frequency range of interest.

The cross spectrum calculations and all numerical manipulations are carried out on a four channel Fast Fourier Analyser with post processing capability (HP 5451C). The software implemented on the analyser includes a single frequency option that enables, for example the estimation of the intensity levels for only the resonance frequencies of a structure. The software can handle up to 750 cross spectra continuously, allowing the calculation of the spatial intensity vector in 250 measurement points, for up to 30 single
frequencies. Dedicated graphic software has been developed for representing those intensity vectors yielding animated intensity patterns or intensity contour plots.

When conducting acoustic intensity measurements one has to be aware of the limits involved. Indeed it has been shown that instrumentation (phase shift) and processing errors (finite difference approximation error) impose respectively a lower and upper bound to the operational frequency range of intensity measurements. As mentioned, the instrumentation error can be considerably reduced if one applies the on line phase correction. The approximation error on the contrary, which is due to the finite separation distance between both microphones, cannot be avoided and induces a distance dependant limit on the operational frequency range.

3 THREE DIMENSIONAL INTENSITY MEASUREMENTS

The acoustic intensity is a spatial vector, and can be represented threedimensionally. This representation results in a better understanding of acoustic phenomena such as hydrodynamical short circuit, radiation efficiency and radiation directivity. The experimental visualization of the intensity pattern introduced here is a method based upon the measurement of three perpendicular intensity components in a series of points. The measurement of the perpendicular components can be accelerated using a three dimensional microphone probe. The three dimensional probe consists of four identical 1/4 inch condensor microphones, geometrically arranged in such a way to yield three perpendicular intensity components (fig.1)

![Diagram of three dimensional microphone probe](image)

Fig 1 Lay-out of the three dimensional intensity probe

As shown by fig.1, the x component of the intensity is estimated by microphone combination (1-2), the y component by combination (1-3) and the z component by combination (1-4). The resulting intensity components consequently do not relate to one single point but only approximate the intensity components for that point. The corresponding approximation error is dependant on the acoustic field, and has been derived for a quadrupole source lay-out. The resulting error levels shown in fig.2 prove that the accuracy loss can be admitted, however, one has to bear in mind that those
levels are only indicative, since the acoustic fields that occur in reality are too complex to be analytically characterised.

Fig. 2. Approximation error due to the three dimensional probe

4. INTENSITY MEASUREMENT ROBOTS

The benefits of automated intensity measurements are evident if one considers the large number of measurements necessary to cover the intensity patterns of real life sources. Even for a relatively simple source, such as a flat plate, up to 50 measurement points are necessary to describe the intensity pattern related to the fifth resonance frequency.

Fig. 3 Measurement robot and intensity pattern above a randomly excited plate (50x50x2cm 245Hz)

The first robot that was modified for acoustic intensity measurements was originally designed for assembling Hi-Fi installations. It is a so-called rectangular coordinate robot with five degrees of freedom: x, y, z translation and two rotations of the wrist (fig.3). The robot is controlled by the processor of the FFT analyser in such a way that robot control and
measurements can be managed by the same program. The robot can be equipped with the three- or with the one dimensional probe as shown in fig.3. If the latter is used the probe will be successively oriented by the robot in three perpendicular directions. 

Due to its limited x,y,z range (100x80x50cm), this robot is mainly intended for research applications. For example the resonant near field patterns of various free/free plates, randomly excited with white noise, have been measured with the robot and compared to the intensity patterns derived by an acoustic radiation model (ref.3). The near field of various loudspeaker boxes have been compared and the radiation behaviour of combustion engine components such as the oil carter have been analysed using this robot. As an illustration of the potential use of the robot the near field intensity pattern of one resonance frequency of a free/free plate is shown in fig.3. The second robot (fig.4) was specially designed for the analysis of the intensity pattern of larger sources. It is a rectangular coordinate robot with three degrees of freedom (x,y,z translation) intended to scan large surfaces such as the body of a car or of a machine tool. The three dimensional probe will always be used since no rotation of the probe is possible. 

This measurement robot has been successfully used in industry to analyse the near field of the noise radiated by a passenger car running under full load on a test bench. One of the resulting patterns is shown in fig.4 and illustrates again the power of the method.

---

Fig.4 Measurement robot and intensity field of the side of a passenger car.

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Research granted by Belgian Government funds as part of the project TEST EN DESIGntechnIEKEN TER OPTIMALISATIE VAN HET DYNAMISCH GEDEG VAN MECHANISCHE STRUKTUREN. Conc. Actie 80-B5 Wetenschapsbeleid.
INTERCOMPARISON OF SOUND PRESSURE STANDARDS

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Introduction

By its nature the standard of sound pressure has to be realised indirectly, e.g. in terms of the output of a microphone of known sensitivity. Consequently there is continuing demand for calibration of reference microphones and there is need to establish comparability between standardising laboratories. During 1963/64 an intercomparison of pressure sensitivity levels of one-inch microphones was carried out under the auspices of WG 13 of IEC TC29 in which laboratories in Denmark, France, Germany, Japan, UK, USA and USSR took part. From this, residual differences in measured sensitivity level of up to 0.8 dB were revealed. Such discrepancies are large compared with the uncertainty of 0.1 to 0.2 dB which is essential to meet the requirements of EEC Directives. The publication of an IEC Standard (1) should have led to an improvement in the agreement between those laboratories able to implement them fully, but it remained to demonstrate comparability between national standardising laboratories within the Community.

The purpose of the work described in this paper was to compare the results for the pressure sensitivity of laboratory standard microphones obtained by different primary calibration laboratories and to resolve any significant differences. In order to reduce the time required, a radial rather than a round-robin procedure was adopted. The central laboratory calibrated a number of microphones, of which two were sent to each participating laboratory to be calibrated; these microphones were finally returned for recalibration by the central laboratory. The microphone type selected was the Type 4160 manufactured by Brüel and Kjær A/S as this is specially designed to serve as a stable laboratory standard. It has an integral front cavity, making it suitable for direct use with the recommended coupler configurations (1).

The National Physical Laboratory (UK) acted as central Reporting Laboratory and the other Contributing Laboratories were: Danmarks Tekniske Højskole (DTU), Denmark; Instituto Elettrotecnico Nazionale (IEN), Italy; Institut National de Métrologie (INM) and Laboratoire National d’Essais (LNE), France; Physikalisch-Technische Bundesanstalt (PTB), Federal Republic of Germany. A detailed report of this intercomparison is
available (2) and only a brief summary of the most significant results is given here.

Calibration technique

For a given frequency the pressure sensitivity of a microphone is expressed as the ratio of the open-circuit voltage at the output of the microphone to the sound pressure uniformly applied over the whole surface of its diaphragm. All participating laboratories employed the reciprocity calibration technique. The basic procedures are given in (1) and considered in more detail in (3). If two linear, passive, reciprocal microphones having pressure sensitivities \( M_1 \) and \( M_2 \) are coupled together acoustically, one being driven by an electrical current \( i_1 \) and producing an open-circuit voltage \( U_2 \) at the output of the other, then

\[
M_1 M_2 = Y_{12} U_2 / i_1
\]

where \( Y_{12} \) is the acoustical transfer admittance of the coupler/microphone system, i.e. the ratio of the short circuit volume velocity of microphone 1 to the pressure generated at microphone 2. The electrical quantities \( U_2 \) and \( i_1 \) can be measured directly whilst for certain microphone/coupler configurations \( Y_{12} \) can be evaluated theoretically in terms of the dimensions of the coupling cavity and the physical properties of the gas within it, thus giving a value for the sensitivity product. Appropriate corrections are required for the shunting effect of the capillary tubes and for heat-conduction to the walls of the cavity. Using pair-wise combinations of three microphones, and determining values of \( Y / i \) corresponding to each of these sensitivity products, allows the individual sensitivities to be determined.

Microphone parameters

When evaluating \( Y_{12} \) it is necessary to take into account not only the impedance of the cavity but also the impedance of the microphones. For the highest calibration accuracy it is necessary to determine individual values for the actual microphones used.

Participating laboratories used various methods to determine the magnitudes of the lumped parameter elements of the equivalent acoustical circuit representation of the microphone. In the case of acoustical compliance, reported estimates of uncertainty ranged from 2% to 9%, and one laboratory reported values which were significantly high. In the case of microphone resonance frequency, where the estimated uncertainty varied from 15 Hz to 30 Hz, one laboratory reported results which were very significantly low (2.4%). Combining results for acoustic compliance and resonance frequency to yield the acoustic mass of the diaphragm, the estimated uncertainty ranged from 3.5% to 5% (note that two laboratories did not determine the mass of the diaphragm - one employed the manufacturer's nominal value and another the NPL value). One laboratory reported results which were approximately 8% lower than would be expected, though such a discrepancy affects the values of derived sensitivity only at higher frequencies and then only to a small extent. In the case of
acoustic resistance, estimates of uncertainty ranged from 1% to 5% and the largest discrepancy was less than 5%; again, an error in resistance of such a magnitude would be expected to have a fairly small effect on the derived sensitivity.

All but one laboratory determined individual values for the depth of the microphone front cavity using optical techniques, the estimated systematic uncertainty ranging from 1 to 10 μm. In general there was good agreement between the results obtained by the contributing laboratories and those obtained by NPL. Various methods were used to determine the front cavity volume plus equivalent volume of the individual microphones at one or more frequencies below 1 kHz, with estimated systematic uncertainties between 2 and 5 mm³.

Results for sensitivity level

Two different coupler configurations are in general use for reciprocity calibration - the "20 cm³" coupler and the "plane-wave" coupler. As a minimum all laboratories made determinations of sensitivity level at the preferred octave frequencies using the 20 cm³ coupler from 250 Hz to 1 kHz inclusive and using the plane-wave coupler from 1 kHz to 8 kHz inclusive, with the results shown in Fig. 7a.

Considering first the 20 cm³ coupler results and taking NPL values as baseline, there appears to be a systematic difference between laboratories ranging from +0.027 dB to -0.069 dB, the latter difference being significant. Detailed consideration indicated three factors which might account for such a discrepancy: the crosstalk level, the value adopted for the ratio of specific heats of the air/water-vapour mixture inside the coupler, and the polarising voltage applied to the microphones (particularly the receiver).

The differences in the plane-wave coupler results at 1 kHz are similar to those just considered for the 20 cm³ coupler and the sources of the discrepancies are probably similar, although errors in determining the effective volume coupling the microphones will be nearly six times more important because the volume of the plane-wave coupler is only approximately 3.4 cm³. Additional likely sources of error include the use of different numerical values for the various parameters of the microphones being calibrated and inadequate correction of the measured sensitivity levels to the reference ambient conditions of 101,325 kPa and 23.0°C, for it has been shown (3) that both the temperature and pressure coefficients of sensitivity level are highly frequency dependent.

Additional measurements

Following completion of the intercomparison and consideration of the draft synthesis report, representatives of the calibrating laboratories agreed that the value of the exercise would be enhanced if those laboratories which had reassessed their calibration techniques were given an opportunity to demonstrate that discrepancies were indeed thereby reduced (all but one laboratory did so). Accordingly, one microphone from each pair was returned to the corresponding laboratory. Changes in the
Figure 1. Sensitivity levels reported by Contributing Laboratories relative to NPL values (a) initial results, (b) final results calibration technique effected by the various laboratories included a reappraisal of the capillary correction (4), reduction of stray capacitance, reduction of crosstalk, improved background noise levels, and greater care with grounding. Results derived from these additional studies are shown in Fig. 1b.

Conclusions

As can be seen, the spread of data is significantly reduced as a result of using the improved measuring techniques and evaluation procedures. Thus mutual confidence in the realisation of sound pressure in the various standardising laboratories within the Community is increased. Although well within the estimated systematic uncertainties, it can be seen that differences still exist towards high frequencies. A significant source of such differences resides in the values adopted for the lumped parameter elements used to derive the acoustic impedance of the microphones, and this aspect of the comparison requires further work.

References

6.9

Mesures d’impédance, absorption, transmission, réverberation
Impedance of measurements, absorption, transmission, reverberation
Impedanzmessung, Absorption, Transmission, Nachhall
presentation and analysis of interregional interactions...
INFORMATION CONTENT OF FFT ANALYSIS OF ROOM IMPULSE RESPONSE

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Introduction

Let a spherical impulse sound wave be emitted from a source in the room, and both the emitted impulse \( p_e(\mathbf{r},t) \) and the response \( p_r(\mathbf{r},t) \) are recorded digitally. If the emitted impulse is short inasmuch as the delay of the response after the emitted impulse is longer than the total time duration of the emitted impulse then the Fourier transform of \( p_e \) and \( p_r \) can be computed from the digital record using the FFT technique. Let's compute the "Quadratic Spectra" of the emitted impulse and the response as

\[
Q_e(\mathbf{r},f_k) = \left| \frac{1}{f_s} \sum_{n=0}^{N-1} p_e(\mathbf{r},t_n) \exp(-j2\pi nk/N) \right|^2,
\]

\[
Q_r(\mathbf{r},f_k) = \left| \frac{1}{f_s} \sum_{n=0}^{N-1} p_r(\mathbf{r},t_n) \exp(-j2\pi nk/N) \right|^2,
\]

where \( f_s \) is the sampling frequency of FFT analyser so that

\[
t_n = n t_1 = \frac{n}{f_s},
\]

\[
f_k = k f_1 = k \frac{f_s}{N}.
\]

The holder of information is the normed spectrum of the response determined according to formula (see ref./1/)

\[
e(\mathbf{r},f_k) = \frac{S}{16\pi r^2} \frac{Q_r(\mathbf{r},f_k)}{Q_e(\mathbf{r},f_k)},
\]

where \( S \) is the total surface of the room, and \( r \) is the distance between the source and the microphone.
Information about sound absorptivity

The impulse response method of sound absorptivity measurement was presented in previous paper /1/. This measurement consists of following digital procedures. At first the normed spectra of the responses are determined according to eqs. (1) - (4) in a number of microphone positions. Then the band acoustic coefficient is approximated as

\[ \bar{\varepsilon}_B = \frac{1}{m} \sum_{i=1}^{m} \frac{1}{\Delta k} \sum_{k=k_d}^{k_h} e(f_i, f_k), \]

where \( m \) is the number of microphone positions, and integers \( k_d, k_h \) correspond to frequency limits \( f_d, f_h \) by approximative relations

\[ k_d = \frac{f_d}{f_1}, \quad k_h = \frac{f_h}{f_1}. \]

The number of FFT values of \( e \) within the frequency band is approximately

\[ \Delta k = \frac{\Delta f}{f_1}. \]

The band absorption coefficient is then

\[ \bar{a}_B = \frac{1}{1 + \bar{\varepsilon}_B}, \]

and the standard deviation of this approximation is

\[ \sigma_a = \frac{\bar{a}_B(1 - \bar{a}_B)}{\sqrt{\Delta k m}}. \]

The experimental verification of this measuring method is presented in paper /1/.

Information about sound diffusivity

Information about sound diffusivity of a room in the given frequency band can be read from the empirical distribution function of FFT normed spectra values lying within this band. For a diffusive room this distribution was found to be exponential one (ref./2/). The cumulative distribution function is then

\[ F(e) = 1 - \exp\left(-\frac{e}{\bar{e}_B}\right), \]

where \( \bar{e}_B \) is the ensemble average of \( e \) within the frequency band. The standard deviation of this distribution is
\[ \sigma = \epsilon B \]  

The empirical distribution function can be determined after rearrangement of the FFT normen spectra values within the frequency band into the increasing sequence \( \{ \epsilon_v \} \). Then

\[ F(\epsilon_v) = \frac{v}{\Delta k} \]

\[ v = 1, 2, \ldots, \Delta k \]  

The defect of diffusivity is then indicated by deviations between the distributions \( \{ 12 \} \), and \( \{ 10 \} \). Information about the degree and character of diffusivity defect gives the normalised standard deviation of FFT values of \( \epsilon \) lying within the examined band

\[ \delta = \frac{\sigma}{\epsilon B} \]  

The diagnosis of diffusivity is illustrated in graphs. As an experimental enclosure the room was used in which the absorptivity measurement described in paper /1/ was done. The investigated bandwidth is \( \Delta f = 500 \text{ Hz} \), the number of FFT values of \( \epsilon \) within each band is \( \Delta k = 82 \). The empirical distribution functions for several frequency bands identified in the fig. 1 are plotted at fig. 2. The ideal exponential distribution is plotted, too.

The following situations can occur:

\( \delta = 1 \), good diffusivity - uniform statistical distribution of reflected sound energy into all directions

\[ \delta = 0,96 \]

\( \delta > 1 \), defect of diffusivity - dominant single standing wave

\[ \delta = 4,87 \]

\( \delta < 1 \), defect of diffusivity - dominant single progressive wave

\[ \delta = 0,75 \]

References:

/1/ Kyncl, Z., Impulse Response Method of Sound Absorptivity Measurement in Enclosures, 8th Collog. on Acoustics, Budapest 1982, 181 - 186

/2/ Kyncl, Z., Spectral Statistical Distribution of Room Impulse Response, FASE/DAGA'82, 231 - 234
Fig. 1. FFT normed spectrum of room impulse response determined for one microphone position - see ref. /1/
\[ V = 80 \text{ m}^3, S = 126 \text{ m}^2, f_B = 50 \text{ kHz}, f_1 = 6.1 \text{ Hz} \]

Fig. 2. Empirical cumulative distribution functions of $\varepsilon$ for frequency bands identified in fig. 1
MEASUREMENT OF THE ACOUSTICAL PROPERTIES OF POROUS ABSORBING MATERIALS AT ELEVATED PRESSURE

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Introduction

This paper describes a piece of apparatus that has been designed and constructed in order to measure the complex propagation constant and characteristic acoustic impedance of porous absorbing materials at both atmospheric and elevated ambient pressure, some of the results obtained with this apparatus in the pressure range one to three atmospheres, and an attempt to codify these results using the empirical approach of Delany and Bazley (1). It is hoped that further study of the way the acoustical properties of such materials vary with ambient pressure, a well-controlled variable, will shed some light on the relative correctness of some of the various theories available for their prediction.

Principles of measurement method

It can be shown (2) that the ratio of the complex acoustic pressure at a distance \( l_1 \) in front of a rigid termination, \( p_1 \), and the complex acoustic pressure at that termination, \( p_0 \), is equal to the hyperbolic cosine of the product of \( l_1 \) and the propagation constant, i.e.

\[
p_1/p_0 = \cosh(\gamma l_1) \quad \text{where } \gamma = \alpha - jk \text{ is the propagation constant}
\]

The reflection coefficient at an interface may be shown to be given by

\[
r = \frac{1 - (p_1/p_2) \exp(jk_0 l_2)}{1 - (p_1/p_2) \exp(-jk_0 l_2)}
\]

where \( k_0 \) is the wavenumber in the region in front of the interface, \( p_1 \) is the complex pressure at the interface, and \( p_2 \) is the complex pressure a distance \( l_2 \) in front of the interface. Since

\[
r = \frac{Z_{\text{face}} - Z_0}{Z_{\text{face}} + Z_0}
\]

where \( Z_0 \) is the characteristic impedance of the medium in front of the interface and \( Z_{\text{face}} = Z_c \coth l_1 \), knowledge of \( \gamma l_1 \) (from above), \( p_1 \) and \( p_2 \) enables \( Z_{\text{face}}/Z_0 \), the relative characteristic impedance of the sample, to be determined.
Experiment

The experimental apparatus for containing the specimen under test is shown in Figure 1. The sample is placed directly in front of a rigid plate that blocks one end of a cylindrical tube, at the other end of which is a loudspeaker. Two different tubes were used for the work reported here, one 12.7 cm in diameter for use up to around 1 kHz, the other 3.2 cm diameter for use up to about 4 kHz. The holes in the sidewall allow the insertion of Bruel and Kjaer 1-inch microphones, one at the back of the sample immediately in front of the rigid plate, one at the sample surface, and one at a distance in front of the sample.

The electrical arrangement is shown in Figure 2. The whole experiment is controlled by the microcomputer which, on instruction from the experimenter, calculates the digital samples that will, when fed to the D/A converter, generate the desired signal (a windowed linearly-frequency-swept sinewave is usually used). After the responses from the microphones have been captured they are Fourier transformed and, after applying calibration corrections (previously calculated from an experiment in which the microphones were placed adjacent to each other in a calibration chamber), used in the equations explained in the previous section to calculate \( a \), \( k \) and \( Z_c/Z_0 \).

For experiments at elevated pressure the tube containing the specimen, together with the microphones and loudspeaker, was placed inside a pressure vessel. Note that for these experiments it was also necessary to carry out a microphone calibration experiment at each of the pressures that were to be used.

In order to compare the results obtained with the Delany and Bazley empirical approach, the flow resistance of the materials used was also measured (at atmospheric pressure only).

Results

Some results on log-log plots are shown in Figures 3 and 4. They are of \( R-1 \) where \( R \) is the real part of \( Z_c/Z_0 \), and \( X \), the imaginary part of \( Z_c/Z_0 \), both plotted against the non-dimensional parameter \( pf/\sigma \), where \( \rho \) is the density of air (which of course changes with ambient pressure), \( f \) is the frequency, and \( \sigma \) is the specific flow resistance per unit thickness. The materials used to obtain the plots shown were Fibroce of densities 48, 96, and 144 kg/m² (with a range of flow resistances from \( 7 \times 10^8 \) to \( 3 \times 10^4 \) rayls/m). Pressures from 1 to 3 atmospheres and frequencies from 400 to 3300 Hz were used in the experiment. As can be seen reasonable straight line fits were obtained in this regime. A summary of the results is:

<table>
<thead>
<tr>
<th>Property</th>
<th>Delany and Bazley</th>
<th>Present work</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>gradient</td>
<td>intercept</td>
</tr>
<tr>
<td>( a/k_0 )</td>
<td>-0.595</td>
<td>0.187</td>
</tr>
<tr>
<td>( (k/k_0)-1 )</td>
<td>-0.700</td>
<td>0.097</td>
</tr>
<tr>
<td>( R-1 )</td>
<td>-0.754</td>
<td>0.057</td>
</tr>
<tr>
<td>( X )</td>
<td>-0.732</td>
<td>0.086</td>
</tr>
</tbody>
</table>

As can be seen the power laws obtained are of the same order as those obtained by Delany and Bazley for a wider range of materials, but not of course including the effect of changing the ambient pressure.
Acknowledgment
Part of the work described here was carried out with support from the British Gas Corporation Engineering Research Station.

References

![Figure 1](attachment:figure1.png)

*Figure 1*

![Figure 2](attachment:figure2.png)

*Figure 2*
UTILISATION D'UN ÉGALISEUR ASSERVI
POUR LA MESURE SIMPLE ET RAPIDE
DES ISOLEMENTS ACOUSTIQUES IN SITU

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De nombreux pays manifestent actuellement le besoin d'une méthode d'évaluation simple, rapide et peu onéreuse des isola"etons aux bruits aériens dans les constructions [1]. La complexité des méthodes actuelles [2] est en effet difficilement compatible avec le développement intensif d'un contrôle de qualité susceptible d'être utilisé aussi bien par les constructeurs, architectes, que par les organismes de contrôle publics ou privés. C'est pourquoi, en réponse à une demande formulée par le Ministère chargé des problèmes de la Construction, le laboratoire d'acoustique de l'établissement public Télédiffusion de France a mis au point une méthode et un appareillage adaptés aux problèmes des contrôles in situ.

Figure 1: Mise en oeuvre du matériel

La figure 1 montre la mise en oeuvre du matériel nécessaire dans le cas d'une mesure d'isolement à l'intérieur d'une construction. L'appareil effectuant l'essentiel du traitement des signaux : l'isolomètre se connecte au sonomètre utilisé côté réception et reçoit par liaison hertzienne le signal du microphone situé côté émission. Dans le cas d'une mesure d'isolement vis-à-vis des bruits extérieurs à la construction notre méthode permet d'utiliser le bruit existant (navigation aérienne, circulation routière et/ou ferrovière) pour obtenir directement un résultat normalisé.
La méthode et l'appareillage que nous avons mis au point sont décrits dans une précédente publication [3]. Rappelons brièvement le principe de fonctionnement de l'Isolmètre (figure 2)

Figure 2: Synoptique simplifié de l'Isolmètre.

À l'émission, le bruit produit est reçu par un microphone. Après transmission, son spectre et son niveau sont comparés au spectre et au niveau déterminés par la réglementation selon la mesure effectuée et choisis parmi les spectres et les niveaux disponibles dans l'appareil.

L'erreur constatée est corrigée grâce à un égaliseur dont l'action asservie modifie automatiquement, par bandes de fréquences, le signal électrique correspondant au bruit reçu par le microphone du sonomètre placé à la réception.

Le facteur de correction issu de la mesure des temps de réverbération du local de réception est introduit à l'aide d'un correcteur de niveau, pour les mêmes bandes de fréquences.

Le signal resultant est restitué au sonomètre et le niveau affiché par ce dernier est celui que l'on aurait effectivement obtenu, dans le local normalisé, si le bruit à l'émission avait eu le spectre et le niveau de référence.

Le niveau lu sur le sonomètre est donc un niveau transmis normalisé dont la valeur absolue peut être comparée directement à une limite réglementaire ou bien encore être retranchée du niveau de référence programmé si l'on désire connaître la valeur de l'isolement proprement dit.

L'adaptation de l'Isolmètre à une réglementation particulière est immédiate dès l'instant où celle-ci est basée sur le respect, soit d'un isolement minimum, soit d'un niveau transmis maximum (vis à vis d'un bruit émission de référence). Il suffit de mettre en mémoire, dans le premier cas, le spectre du bruit de référence avec un niveau global arbitraire, dans le second cas, le spectre du bruit de référence avec le niveau absolu imposé par la réglementation.

Dans le but d'estimer les possibilités réelles d'un égaliseur asservi tel que celui que nous venons d'évoquer, nous nous sommes livrés à un test de comparaison en laboratoire en nous limitant volontairement aux signaux électriques habituellement traités par les appareils de mesure, dans des conditions normales d'utilisation.
La figure 3 représente l'ensemble des appareils de mesure utilisés autour de l'Isolomètre associé à un sonomètre de classe 1.

Figure 3: Test de validation de l'Isolomètre

L'isolement acoustique est simulé à l'aide d'un metteur en forme de spectre (B.K. type 5587) suivi d'une ligne d'affaiblissement (L.E.A. 0/11 dB). Le signal émission peut être choisi parmi quatre signaux. Les signaux "a" et "c" ont un spectre très proche du bruit de référence ROSE ou ROUTE décrits dans la réglementation française [4]. Le signal "b" correspond à ce qu'il est possible d'obtenir dans un petit local d'émission avec une source quelconque. Le signal "d" est l'enregistrement d'un bruit recueilli en bordure d'une voie de circulation à un moment où le passage des véhicules est assez faible et ne comporte pas de camion. Ces signaux ont une allure spectrale assez différente de l'allure idéale de manière à éprouver les capacités de correction de l'Isolomètre.

Nous avons retenu, parmi de nombreux résultats de laboratoire, sept isolements: A, B, C, D, E, F, G qui tous présentent une difficulté de mesure à cause d'une pente élevée et/ou d'accidents en fonction de la fréquence. Chaque isolement est mesuré successivement avec les signaux a, b, c, d dénommés respectivement Rose référence, Rose quelconque, Route référence, Route réelle.

La mesure dite "traditionnelle" consiste à utiliser un analyseur numérique de fréquence temps réel (B.K. type 2131) de manière à obtenir une évaluation globale en dB(A) soit:

- par différence directe des niveaux émission et réception (uniquement dans le cas des signaux a et c), l'intégration est linéaire sur 32 s.
- par analyse 1/3 d'octave des signaux (intégration 64 s) puis calcul.
- par analyse octave des mêmes signaux (intégration 32 s) puis calcul.

La mesure dite "Isolomètre" consiste à utiliser celui-ci en liaison avec un sonomètre ou un analyseur statistique (B.K. type 2209 et type 4426).
Toutes les mesures ont été effectuées en respectant une équivalence électrique/acoustique de 50 mV représentant 94 dB (réf: 2.10-5 Pa). Les résultats obtenus sont donnés figure 4. Pour simplifier les comparaisons ils sont tous exprimés en isolements bien que ce ne soit pas, comme nous l'avons expliqué, les chiffres indiqués directement par l'Isolmètre.

<table>
<thead>
<tr>
<th>ROSE</th>
<th>RÉFÉRENCE (A)</th>
<th>QUELCONQUE (B)</th>
<th>ROUTE</th>
<th>RÉFÉRENCE (C)</th>
<th>RÉEL (D)</th>
</tr>
</thead>
<tbody>
<tr>
<td>illage</td>
<td>%</td>
<td>%</td>
<td>%</td>
<td>%</td>
<td>%</td>
</tr>
<tr>
<td>54.1</td>
<td>53.2</td>
<td>52.7</td>
<td>54.2</td>
<td>53.7</td>
<td>53.5</td>
</tr>
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<td>42.5</td>
<td>43.6</td>
<td>44</td>
<td>43.5</td>
<td>43.2</td>
<td>43</td>
</tr>
<tr>
<td>53.3</td>
<td>53.5</td>
<td>52</td>
<td>53</td>
<td>55.5</td>
<td>53.2</td>
</tr>
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<td>32.9</td>
<td>32.8</td>
<td>32.9</td>
<td>32.8</td>
<td>32.5</td>
</tr>
<tr>
<td>28.6</td>
<td>28.9</td>
<td>28.9</td>
<td>29.7</td>
<td>28.8</td>
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</tr>
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<td>48.7</td>
<td>48.5</td>
<td>49.2</td>
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</tr>
<tr>
<td>28.4</td>
<td>28.5</td>
<td>28.5</td>
<td>28.2</td>
<td>28.3</td>
<td>28.6</td>
</tr>
</tbody>
</table>

Figure 4: Tableau comparatif des résultats.

Comme on peut le constater les valeurs diffèrent assez peu entre elles. Dans certains cas le résultat fourni par l'Isolmètre est même plus proche du résultat obtenu par analyse et calcul 1/3 d'octave que par analyse et calcul octave et ceci malgré l'emploi de filtres d'octave sur le prototype que nous avons réalisé.

L'explication réside dans le fait que le sonomètre utilisé applique la pondération "A" de manière continue en fréquence ce qui ne peut être obtenu par calcul dès lors que la densité spectrale des bruits n'est connue que par bandes de fréquences.

En attendant que les progrès technologiques nous permettent d'effectuer facilement et rapidement in situ ce que l'on fait actuellement en laboratoire (grâce notamment à une miniaturisation du matériel) nous pensons que l'utilisation d'un égaliseur asservi est actuellement, parmi les méthodes ayant un niveau de précision suffisant, l'une des plus simples et des plus rapides.

BIBLIOGRAPHIE
THE MEASUREMENT OF REFLECTION COEFFICIENTS USING CEPSTRAL TECHNIQUES.

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1. INTRODUCTION

In view of the limitations of existing techniques of acoustic impedance determination, especially for field measurements, we have been motivated to devise a novel method based on cepstral processing. We have previously validated the technique using electrical analogues of the acoustic reflection process [1,2] and this paper describes the extension to acoustical measurements under laboratory conditions.

Two acoustic foam samples, one being film-faced, have been measured in an anechoic chamber. Following signal processing, the impulse responses of the reflectors have been extracted from the cepstra, and complex reflection coefficients and impedances calculated.

The results obtained cover a frequency range of approximately 200 Hz to 20 kHz and are in good agreement with impedance tube measurements where these are available. Additionally, the results, especially the impulse responses, provide interesting qualitative comparisons of the dissipation and propagation mechanisms in the two foam samples.

2. EXPERIMENTAL DETAILS

The two samples to be measured were the two possible orientations, film up and down, of a commercial 25 mm thick, impermeable film-faced, acoustic grade foam (Harrison and Jones 29/19 FR) placed on a 25 mm thick chipboard backing. A large sample area (3.7 m by 4.2 m) was chosen to minimize edge effects and the samples were installed in the I.S.V.R. 300 cubic metre anechoic chamber to reduce the effects of spurious reflections. A B. & K. 4165 microphone was mounted at grazing incidence, .51 m above the sample and a Sonn Audax HD 13 D 37 midrange driver was suspended 2.5 m above the sample directly above the microphone’s diaphragm.

The signal selected was a 10 ms duration, 0 Hz to 20 kHz swept-sine burst initially generated in software and produced with the aid of an EPROM storage device having a 12 bit digital-to-analog output capability. Bursts could be triggered externally and this permitted an
automated scheme to be devised for performing time domain averaging. The signals were amplified by a DC-coupled 150 W amplifier (Amcron DC 300) and fed to the loudspeaker.

The microphone signals were amplified using a B. & K. 2607 measuring amplifier before being acquired by the I.S.W.R. Data Analysis Centre's PDP 11/50 computer, after being passed through a Kemo 48 dB/Octave anti-aliasing filter. In the light of previous experience of cepstral processing [1,2] the sampling frequency was set to 40 kHz and the filter cutoff frequency to 20 kHz. Each signal used as input to the power cepstrum calculation was the result of averaging 100 individual bursts in the time domain. Three measurements were done in all: one for each of the samples and an additional measurement with the sample and backing removed for background subtraction purposes [2].

The cepstral processing has substantially followed previously devised procedures [2] except that the extraction delay has been estimated by cross-correlation with a unit impulse function rather than a theoretical impulse response. The zooming-decimation procedure yields a phase uncertainty at 20 kHz of only ±5°, equivalent to locating the front surface of the reflector to better than ±0.25 mm. The extracted impulse responses have been Fourier transformed to yield the reflection coefficients from which the surface normal plane wave impedances for normal incidence have been calculated; the results are valid to approximately 19 kHz.

3. RESULTS

The impulse responses for the foam and film-faced samples are displayed in figures 1 and 2 respectively. The impulse responses are distinctly different for the two samples and some qualitative conclusions can be drawn about the mechanisms involved in each case. The foam shows an initial reflection from the front surface followed by a reflection from the rigid backing around 0.25 ms later. The estimated wave speed in the foam is then approximately 180 m/s, consistent with an airborne wave slowed by inertial effects in the porous medium. The film-faced sample has a larger initial reflection due to the impermeable film facing. The back reflection occurs around 1.25 ms later, much delayed by comparison with the foam sample; the inferred wave speed is now only approximately 40 m/s. The striking difference suggests a different wave propagation mechanism, most probably via elastic frame motion of the foam rather than air motion.

The surface normal impedances are presented in figures 3 and 4. The foam is seen to behave as a conventional finite depth, rigid porous material with the real part of the impedance tending to a constant value (approximately 2) at high frequencies and the imaginary part tending to zero with increasing frequency. There is some suggestion of frame interaction effects, especially below 8 kHz. The impedance of the film-faced foam is strongly marked by layer resonance effects, closely spaced in frequency due to the slow wave speed in this case. It is also seen that the only significant energy dissipation is below 2 kHz.
4. CONCLUSIONS

The techniques developed have allowed accurate laboratory measurements of surface normal impedance of standard acoustical materials to be made over a very wide frequency band. Physically the experiments are easy to conduct but there are substantial demands on signal processing capabilities. Work in hand is directed towards development of hardware to allow these measurements to be conducted in the field.

The acoustic foam measurements presented show interesting differences in the behaviour when the foam is open and when faced with a flexible, impervious layer. Open foam essentially behaves as a rigid porous material but in contrast, the faced foam shows clear signs of elastic frame motion.

REFERENCES


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Figure 1  Figure 2
GOLD-BOLTON, CEPSTRAL MEASUREMENT OF REFLECTION COEFFICIENTS.

Figure 3

(Imedance normalized with respect to $\rho_0 c_0$)

Figure 4

(Imedance normalized with respect to $\rho_0 c_0$)
TWO CHANNEL FFT ANALYSIS APPLIED TO TRANSMISSION MEASUREMENTS

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Introduction

Dual channel FFT analysers offer a number of effective analysis techniques for building acoustics and noise control measurements. The primary dual channel method today is the determination of acoustic intensity. In transmission measurements of structures this method allows the special reverberant receiving room of the traditional transmission suite to be rejected [1].

In this paper, some extensions are presented to the two-transducer cross-spectral method, and a few preliminary experiments are described. The basic modification given here has been shown [2,3] to yield directly the specific acoustic impedance, the radiation efficiency and, under some assumptions, the absorption coefficient. Moreover, the extensions will provide a step towards the removing of the remaining reverberation room in transmission loss measurements.

The (one-sided) real intensity spectral density is basically given by

\[ I(f) = \text{Re}(G_{vp}(f)) \]  \hspace{0.5cm} (1)

where \( G_{vp} \) is the (one-sided) cross-spectral density of velocity \( v(t) \) and pressure \( p(t) \) signals. The intensity is in practice estimated from the pressure and acceleration \( a(t) \) signals of the surface intensity technique or the two pressure signals \( p_1(t) \) and \( p_2(t) \) by the familiar expressions

\[ I(f) = -\frac{1}{\omega} \text{Im}(G_{vp}(f)) = -\frac{1}{\omega cd} \text{Im}(G_{12}(f)) \]  \hspace{0.5cm} (2)

Direct measurement of acoustic impedance

In many cases, however, the absolute intensity and power are not the desired information. Often one wishes to determine relative quantities such as the radiation efficiency or the absorption coefficient. The main contributing factor in these is the specific acoustic impedance, and, indeed, the same two-transducer and dual-channel-FFT instrumentation can be used for measuring impedance as well. The impedance \( Z(f) \) is defined, and for deterministic signals with no measurement noise directly given, by the ratio \( P(f)/V(f) \) of the Fourier transforms of pressure and velocity. However, in the more general case of stochastic signals impedance is estimated by

\[ Z(f) = \frac{G_{vp}(f)}{G_{vp}(f)} = \frac{E[V*P]}{E[V*V]} \]  \hspace{0.5cm} (3)
which is the frequency response function of a linear system with \( v(t) \) as input and \( p(t) \) as output. In analogy to Eq (2), the surface intensity and acoustic intensity instrumentation gives

\[
\hat{z}(f) = j\omega H_{ap}(f) = -\frac{\omega pd}{2} H_{DS}(f)
\]

(4)

where

\[
H_{DS}(f) = \frac{B(P_2^*P_1^*)}{B(P_2^*P_1^*)} = \frac{G_{22} - G_{11}}{G_{11} + G_{22}} - \frac{j2 \text{ Im}(G_{12})}{2 \text{ Re}(G_{12})}
\]

(5)

The impedance determined in this way is related to the particle velocity component in the sensing direction of the transducer pair. Hence, perpendicular orientation at or near the boundary gives the radiation impedance and the point impedance which are of interest here because the normal velocity alone contributes to the power transmitted through a boundary.

**Determination of absorption coefficient and incident intensity**

Assuming a plane wave striking a structure with an angle of incidence \( \theta \), the absorption coefficient of the boundary may be expressed as [4]

\[
\alpha(\theta, f) = \frac{4 \text{ Re}(a(f))}{|x(f) + z_p(f)|^2}
\]

(6)

where \( a(f) \) is the normalised point impedance \( z(f)/\rho c \) of the boundary from Eq (4) and \( z_p(f) \) is a normalised "field impedance", a function of, e.g., \( \theta \), see [4]. For an infinite homogeneous plane boundary \( z_p = 1/\cos \theta \). A statistical absorption \( \alpha_{\text{stat}}(f) \) for the diffuse field is given by an integral of \( \alpha(\theta, f) \). For practical calculations \( z_p \) is tabulated for some common configurations and the integral can be evaluated by a simple sum [5]. Hence, the incident intensity needed for the transmission loss can be obtained using \( \alpha(\theta, f) \) or \( \alpha_{\text{stat}}(f) \), if the transmitted power is determined via intensity on the incident side of the structure. Eq (6) can be used directly in measurements with one approximate plane wave, with the exciting loudspeaker effectively in free field, either spatially (in an anechoic room or outside a façade) or temporally (pulse excitation and suitable data windowing).

**Direct measurement of radiation efficiency**

The radiation efficiency of a radiating surface may be defined [2] by

\[
\sigma(f) = \frac{1}{\rho c} \langle I_c(f) \rangle_s = \frac{1}{\rho c} \langle I_{up}(f) \rangle_s
\]

(7)

where \( \langle \rangle_s \) denotes spatial average over the radiating surface and \( I_c \) is the complex intensity. \( \sigma(f) \) has been estimated by taking separate measurements for the average intensity and the average velocity [6,7]. However, it is readily apparent that the direct methods of impedance measurement in Eq (4) can be applied here. In practical analyses the expectation is approximated by ensemble averaging over sample F-transforms, denoted by \( \langle \rangle_e \). Hence,

\[
\sigma(f) = \frac{1}{\rho c} \langle \langle v \phi \rangle \rangle_e_s = \frac{1}{\rho c} \hat{H}_{up}(f)
\]

(8)

where the caret denotes an estimate whereby the analyser is used to collect the averages both over sample records and over the surface. Thus,

\[
\sigma(f) = \frac{j\omega}{\rho c} \hat{H}_{ap}(f) = \frac{j\omega}{2a} \hat{H}_{DS}(f) = \frac{j\omega}{2a} \left( \frac{G_{22} - G_{11}}{G_{11} + G_{22}} - \frac{j2 \text{ Im}(G_{12})}{2 \text{ Re}(G_{12})} \right)
\]

(9)
Transmission loss and frequency response

More detailed information on the transmission than what contained in the transmission loss (TL) can be obtained by the frequency response function $H_{xy}(f)$. This may be determined between an incident plane wave $s(t)$, or more practically a spherical wave from a loudspeaker, and pressure $y(t)$ at one point on the receiving side. Another approach to estimating TL is provided by an average over $H_{xy}$ with different directions for $x$ and positions for $y$.

In standard FFT analysers the data records used in computation are, however, usually too short compared with the impulse response length of the system for acceptable estimation of $H_{xy}$ [8]. This is also why another spectral method, time delay spectrometry, fails in the TL estimation of common structures [9]. This problem may be solved by zoom FFT analysis, but perhaps more effectively by using pulse excitation, allowing synchronisation of data windows, and $n$ short contiguously delayed segments in the $G_{xy}$ estimate as

$$G_{xy}(f) = 2E[x^* \sum_{k=1}^{n} Y_k]$$  \hspace{1cm} (11)$$

Analysis examples and comments

To illustrate the use of the presented methods, transmission through a double-glazed window was analysed. Data were acquired on both sides of the window installed in an ordinary transmission suite and using the surface intensity technique. The statistical absorption coefficient in Fig 1(c) is a rough approximation calculated with a mean $s(f)$ over the window surface and $\bar{s}(f) = 1/\cos \theta$, using Eq (18) of [5]. The difference between the intensities on the front and back sides in Fig 2(a) corresponds to internal losses of the window. The TL curves in Fig 3 estimated by standard and FFT methods exhibit some resemblance. It is emphasized that the early results shown here are the outcome of mere familiarization and qualitative demonstration of the methods (e.g. including severe undersampling in space). Further work aimed to answer quantitative questions is now in progress.

The two transducer technique together with dual channel FFT methods are seen to give directly several important acoustical quantities in situ in an arbitrary environment, which could formerly be obtained only in special facilities and/or by indirect methods. Another potential field of application
Fig 2. (a) mean intensity on the window surface, upper curve: transmitting, lower: receiving room, (b) real radiation efficiency of the window.

Fig 3. Transmission loss of the window, calculated from diffuse field pressure (transmitting room) and radiated intensity, from point impedance, absorption coefficient and intensities on both sides, measured by standard transmission suite method; integration into 1/3 octave levels.

for the impedance method presented here may be acoustic modal analysis which also has so far suffered from indirect or restricted information of acoustic impedance [10].

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A NEW METHOD OF MEASURING REVERBERATION TIME

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1. Introduction

Crosscorrelation method using maximum-length sequence is effective to get impulse response of a room with high signal-to-noise ratio[1]. We can get ensemble average of decay curves excited by white noise from the squared values of impulse response, through Schroeder's integration formula[2]. Sagayama introduced a crosscorrelation method using white noise modulated by a maximum-length sequence as a sound source. The method directly gives impulse power response, but the result showed poor signal-to-noise ratio[3]. DATE et al. improved Sagayama's method by introducing two relatively prime pseudorandom sequences and decimation technique which result computational saving and increased signal to noise ratio[4]. This paper describes the more refined version, the PR-method, which gives impulse response instead of impulse power response with higher signal-to-noise ratio and smaller observation time than the preceding method.

2. Principles

Fig.1 shows the block diagram of this method. A modified pseudorandom sequence \( M'(n) \), whose autocorrelation is an impulse train with exactly zero rested value, is driven by a clock \( T_{c} \) high enough to ensure sufficient power spectra within the frequency region under observation. The other pseudorandom sequence \( M(m) \), whose value is either 1 or -1, is driven by \( d \)-times slower clock, switches \( M'(n) \) to get input signal \( x(n) \) to a loudspeaker, that is

\[
x(n) = \frac{1}{2} (M(m)+1)M'(n)
\]

where \( m = [n/d] \). The symbol \([a]\) denotes the maximum integer that does not exceed \( a \). We choose the values of \( N \) and \( N' \), which are periods of the sequences \( M(m) \) and \( M'(n) \) respectively, are relatively prime. Furthermore, \( d \) is an integer which is prime to \( N' \). The output of the microphone is

\[
y(n) = \sum_{k=0}^{d} x(k)r(n-k)+N(n)
\]

where \( r(n) \) is the impulse response of the room and \( N(n) \) is an external noise. (Note that \( r(n) \) can be of infinite duration in this method).

The output \( y(n) \) can be decimated by an integer factor \( d \) if the bandwidth of \( r(n) \) is equal to or less than \( 1/T_{c} \). The decimated sequence is

\[
y(md) = \frac{1}{2} \sum_{k=0}^{d} [M(m+[\frac{-k}{d}])+1]M'(md-k)r(n)+N(md)
\]

We define a new sequence called PR-sequence given by
Pr(m)=M(m)M'(md) (4)

The crosscorrelation $\phi_{PrV}(i)$, $0<i<N$, between $Pr(m)$ and $y(md)$ can be divided into two parts

$$\phi_{PrV}(i)=\phi_{Sy}(i)+\phi_{N}(i)$$ (5)

where $\phi_{Sy}(i)$ is the purely signal-originated component and $\phi_{N}(i)$ is the component influenced by external noise.

As for the signal component, the following equation holds

$$\phi_{S}(i)=(1/2Np)\sum_{m=0}^{Np-1} M(m+i)\sum_{m=0}^{Np-1} M'(m) M'(md-k)m(k)$$ (6)

where $Np=NNd$. Because $N$ and $N'$ are relatively prime, $M(m)$ and $M'(n)$ are statistically independent.

Therefore, the eq(6) can be written as:

$$\phi_{S}(i)=(1/2Np)\sum_{m=0}^{Np-1} \sum_{m=0}^{Np-1} M(m+i)M'(md-k)M'(md-id)m(k)$$ (7)

Utilizing the following property of the modified pseudorandom sequence

$$\sum_{m=0}^{Np-1} M'(md-k)M'(md-id)=M'(n)$$ (8)

where $\delta(n)$ represents unit sample sequence and $M'(n)$ is the power of $M'(n)$, we obtain

$$\phi_{S}(i)=(1/2)\sum_{m=0}^{Np-1} M(m+i)M'(md-id)m(k)$$ (9)

Furthermore using the following property of pseudorandom binary sequence

$$\sum_{m=0}^{Np-1} M(m+i)M'(md-id)=0$$ (10)

we finally obtain

$$\phi_{S}(i)=(1/2Np)M'(i)$$ (11)

This final equation means $\phi_{S}(i)$ equals the decimated value of the impulse response without time-aliasing which usually originates from finite length $N$ of modulating sequence.

Signal-to-Noise Ratio

The noise component $\phi_{N}(i)$ of the crosscorrelation, is given by

$$\phi_{N}(i)=(1/2Np)\sum_{m=0}^{Np-1} N(md)M'(md-id)$$ (12)

Because of statistical independence between $N(n)$, $N(m)$ and $M'(n)$, eq(12) can be written as

$$\phi_{N}(i)=(1/2Np)\sum_{m=0}^{Np-1} M(m+i)M'(md-id)$$ (12)

Substituting the following relations

$$\sum_{m=0}^{Np-1} M(m)=0, \quad \sum_{m=0}^{Np-1} M(m+i)=N$$ (13)

we obtain

$$\phi_{N}(i)=(1/2Np)\sum_{m=0}^{Np-1} \sum_{m=0}^{Np-1} M'(md-id)=0$$ (14)
where \( C \) is a proportional constant, into eq(13), we obtain

\[
\phi_N(1) = C \left( \frac{N'_0}{N_0} \frac{N_p}{N_p} \right)^{1/2}
\]  

(15)

Combining eq(11) and (15) and assuming \( r(n) = \delta(n) \) conservatively, we define the signal-to-noise ratio of this PR-method as

\[
\text{SNR} = \frac{\sigma_S(0)}{\sigma_N(0)} = \left( \frac{N'_0}{N_0} \right)^{1/2} \frac{N_p}{N_p}^{1/2}/2C
\]  

(16)

Clearly SNR is proportional to the SNR \( \left( \frac{N'_0}{N_0} \right)^{1/2} \) in the original sound field and square root of observation time counted by the clock \( T_c \). In this method, decimation does not serve to increase of SNR as it does in the preceding case[4].

Observation Time

The observation time necessary for each impulse response is \( N_p/T_c = N N'_d/T_c \). If \( N \) is a multiple of \( d \), \( N_p \) can be replaced by \( N N'_d \). Table 1 shows an example of selected values of parameters. The product of decimation factor \( d \) and clock time \( T_c \) is kept constant irrespective of maximum frequency range which is inversely proportional to clock time \( T_c \). The observation time is 44 seconds for effective duration 630ms of impulse response in this case.

Amount of Computation

In the computation of crosscorrelation, indicated in eq(5), multiplication is unnecessary because the value of \( Pr(m) \) can be either 1 or -1. Therefore,
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Table 1  An example of selected values of parameters for the PR-method applied to a room whose effective duration of impulse response is less than 630ms. The observation time for each case is 44 seconds.

<table>
<thead>
<tr>
<th>decimation (d)</th>
<th>clock for n (Tc)</th>
<th>max. freq. range (f_{max})</th>
<th>bandwidth (Δf)</th>
<th>Period of M(n), (N)</th>
<th>Period of N'(n)'(N')</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>10 ms</td>
<td>100 Hz</td>
<td>100 Hz</td>
<td>67</td>
<td>63</td>
</tr>
<tr>
<td>3</td>
<td>3</td>
<td>330</td>
<td>100</td>
<td>67</td>
<td>63</td>
</tr>
<tr>
<td>7</td>
<td>1.4</td>
<td>720</td>
<td>100</td>
<td>67</td>
<td>63</td>
</tr>
<tr>
<td>11</td>
<td>0.9</td>
<td>1100</td>
<td>100</td>
<td>67</td>
<td>63</td>
</tr>
<tr>
<td>65</td>
<td>0.15</td>
<td>6600</td>
<td>100</td>
<td>67</td>
<td>63</td>
</tr>
</tbody>
</table>

\((N_p-N)N\)-times additions are only necessary for each observation.

3. Experiments

Simulation results are shown in Fig.2 and 3, whose conditions differ in the amount of noise only. The horizontal axes represent "m", namely, decimated sample number. Dynamic range more than 50dB is guaranteed for decay curve of no external noise. For the case of same level for signal and noise shown in Fig.3, the dynamic range of the decay curve is about 25 dB which equals the predicted value of SNR if we simply put \(C=1.5\) in eq. (16).

Fig. 4 is an example of observed reverberation curve of a room. (1/3 octave, 500Hz center frequency, \(N=63\), \(N'=67\), \(d=11\), clock interval for \(n=1.4\)ms, clock interval for \(m=9.6\)ms and duration of observation =44s.)

4. Summary

The PR-method, which measures impulse response by using two relatively prime pseudorandom sequences and decimation, is introduced. It has merits of shorter observation time than the preceding method, which gives impulse power response, high signal-to-noise ratio and decreased computational amount.

References

MEASUREMENT OF WAVE PROPAGATION IN BEAMS AND PIPES

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1. INTRODUCTION

Pipes form essential components of many industrial, aerospace and marine vehicle systems, conveying mass, energy and momentum from one part of a system to another. In so doing, they frequently transport unwanted energy in the form of fluid and structural vibration between connected regions. In seeking to minimise vibration transmission a designer needs to understand the nature of the waves carried by a pipe, and the fluid which it contains, so that he can optimise the design of control elements incorporated into the system. Within the audio frequency range a number of dispersive forms of wave, many involving cross sectional deformations of a pipe, may propagate [1]. In conjunction with theoretical analyses, a programme is in progress to develop experimental techniques to evaluate the propagation characteristics of high frequency dispersive waves in fluid-filled pipes, and also the reflection and transmission behaviour of various forms of discontinuity in a pipeline, for which wave amplitude detection is necessary. This paper summarises the results of attempts to investigate dispersive wave behaviour by means of correlation and wave field transformation techniques applied to bounded systems in which boundary reflections can cause problems. The results reported herein were obtained by measurements on a simple uniform beam system carrying bending waves; the beam cross section dimensions were 6 mm x 50 mm and it was 6.27 m in length. In the lecture results obtained on circular section pipes will be presented.

2. CORRELATION TECHNIQUES

2.1 M-Sequence Technique

The conventional technique of time lag cross correlation between broadband vibration signals from two or more positions on a waveguide is not useful in the case of dispersive waves. Various M-sequence techniques have been reported by Kashima [2]: the input consists of band limited random noise which is modulated by an on-off pseudo-random binary sequence (prbs); the envelope of the response signal is produced by squaring, followed by low pass filtering and correlation with the modulation signal. In principle the correlation peaks take the form of the prbs auto-correlation, which can be made almost perfectly triangular by using long sequences. This technique was not found to be satisfactory in the case of strong dis-
persion and high group velocities, and a variation was tried, in which
integral numbers of complete cycles of a single frequency carrier wave were
released according to a prbs: the idea was to generate a deterministic
signal with a strong narrow spectral peak, providing group velocity infor-
mation, and widely spread, low amplitude side band components. The deter-
mministic nature of the carrier wave allowed heterodyning, followed by low
pass filtering, to be used for envelope detection.

2.2 Envelope Extraction by Heterodyning

The degree of effectiveness of the technique was found to be very
variable: at some frequencies the correlation peaks were very clean and
sharp (Figure 1a), whereas at other frequencies the correlation curve ex-
hibited strong oscillation (Figure 1b). The problem appeared to be caused
by the generation of strong spectral peaks at frequencies other than the
carrier wave, by side band excitation of resonances; these frequencies
produced interference components in the correlation curve. This possibility
was confirmed theoretically.

2.3 Envelope Extraction by Hilbert Transform

The envelope of a signal \( y(t) \) may be expressed as \( |z(t)| \) where

\[
z(t) = y(t) + j \hat{y}(t) \quad \text{and} \quad \hat{y}(t) = \frac{1}{\pi} \int_{-\infty}^{\infty} \frac{y(\xi)}{t - \xi} \, d\xi.
\]

Simulation of the effects of dispersion on the M-sequence correlation using
a single frequency carrier wave produced the results shown in Figures 2a
and 2b. It was found by simulation, and by experiment, that dispersion
prevented the relative amplitudes of the various wave group reflections
from being correctly evaluated, thereby making wave energy reflection and
transmission coefficient determination impossible.

2.4 Envelope of the Cross Correlation

An attempt was made to employ the envelope of the cross correlation of
the response signal \( y(t) \) and the prbs \( E(t) \). This process is equivalent to
cross correlating \( E(t) \) with \( y(t) \), and with \( \hat{y}(t) \), and taking the modulus of
the complex cross correlation thus obtained. This method is, in principle,
numerically equivalent to the heterodyning procedure. A typical corre-
lation plot is shown in Figure 3; again, the results were not consistently
acceptable.

2.5 Cross Correlation of the Envelope of Response Signals at Two Points

The envelopes of modulated carrier wave signals from two points near
a discontinuity in a pipe were extracted by Hilbert transform and the
envelopes were then cross correlated to give the group delay and amplitude
ratios. This method has the advantage of minimising dispersion of the
waves by a suitable choice of separation between the measurement points.
Some results are shown in Figures 4a and 4b. Unfortunately side band
excited resonances appear to distort the correlation curves and diminish
the accuracy of estimation of the wave group energies.
3. WAVE FIELD TRANSFORM TECHNIQUES

At any point in a wave-bearing system the response variable may be expressed as

$$v(x,t) = \int \int X(\omega, k) \exp(j(\omega t - kr)) \, dk \, d\omega,$$  \hspace{1cm} (1)

where $X(\omega)$ is the Fourier transform of the excitation and $H(\omega, k)$ is the two dimensional transfer function: for a lossless system $H(\omega, k) = \delta(k - \omega/c)$. By performing a 'slant-stack' operation [3] on the response variable, thus:

$$u(\tau, c_0) = \int v(r, \tau + r/c_0) \, dr,$$

$$= \int \int \int X(\omega, k) \exp(j\omega \tau) \exp(j(\omega/c_0 - k)r) \, r \, dr \, dk \, d\omega$$

$$= \int X(\omega) \cdot H(\omega, \omega/c_0) \exp(j\omega \tau) \, d\omega. \hspace{1cm} (2)$$

The physical significance of this operation is that the component of the response which travels with phase velocity $c_0$ is extracted. Therefore the corresponding spectrum

$$U(\omega, c_0) = X(\omega) \cdot H(\omega, \omega/c_0)$$  \hspace{1cm} (3)

is obtained.

For a lossless dispersive wave process where $H(k, \omega) = \delta(k - \omega/c)$, $U(\omega, c_0)$ is equal to $X(\omega)$ for $k = \omega/c_0$, and vanishes elsewhere. Theoretical and experimental results have been obtained for flexural waves in the beam, with pulse excitation; typical results are presented in Figure 5 in which the outgoing component alone is shown. Similar results could be obtained for reflected waves.

In the case of waves in pipes modal extraction is performed in the circumferential sense, in addition to the slant stack operation on axially spaced samples. Results obtained on pipes will be presented at the meeting.

4. ACKNOWLEDGEMENT

This research is supported by the Ministry of Defence.

REFERENCES


Figure 1.

\[ R_{E,A}(\tau) \] 2 KHz

\[ R_{E,A}(\tau) \] 4 KHz

\( \tau \) (ms)

Figure 2.

\[ R_{E,IZ}(\tau) \] non-dispersive

\[ R_{E,IZ}(\tau) \] dispersive

\( \tau \) (ms)

Figure 3.

\[ R_{E,A}(\tau) \]

\[ R_{E,Z}(\tau) \]

\( \tau \) (ms)

Figure 4.

\[ R_{v_1,y_2}(\tau) \]

\[ R_{v_1,y_2}(\tau) \]

\( \tau \) (ms)

Figure 5.

\[ \Delta \rho = 4 \times 10^{-4} \]
SPARK PULSE SOURCE FOR SCALE MODEL STUDIES

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INTRODUCTION

Spark pulses are usual test signals on scale model studies. This work deals with the generation of spark pulses at low charged voltages (≤500 v). The electric process is analysed with regard to the spark production. The acoustical features of the associated sound pulse are studied and related to the electric characteristics. Some considerations concerning the control of the spectrum are also made.

ELECTRIC PROCESS

The basis of our source of detonations is a capacitor bank previously charged. The discharge of the electric energy, controled by means of a thyristor, occurs across a graphitum resistor placed between the electrodes. Simultaneously an arc is generated and an impulsive sound follows.

The charged voltage of the capacitor is from 350 v to 500 v (d.c.) and the capacitance is from 40 μF to 100 μF. The graphitum piece is the responsible for the production of the arc to so high electrode gap (up to 15 mm and more) and low voltages (classic woks use charged voltages ranged between 6 and 15 kv /1/). It facilitates the ionisation of the surrounding fluid when crossed by a high electric current.

The analysis of the electrical process of discharge was carried out by monitoring the dischargin voltage and current waveforms for many different values of charged voltage, capacit, form of the electrode support and graphitum length.

In all cases we have found the same generic forms for the time dependence of voltages. Similar results hold for the electric currents. Figure 1 shows the voltage and current waveforms for charged voltage = 350 v, capacitance = 60 μF and electrode form = . (graphitum length = 6 mm).

The set of curves v(t) and i(t) show three stages :
the first stage 1 is quite similar to a typical capacitor discharge across a resistor and we have named it ionization stage because during it the ionization of the surrounding fluid occurs; the second stage 2 is characterised by the production of the arc, by a instantaneous decrease of the voltage and an increase of the electric current. Finally the third stage 3 corresponds to the capacitor discharge beginning from the residual charge. (See Figure 1).

The decrease of the voltage in the second stage is related with the amount of thermal energy transferred to the fluid, hence with the energy of the acoustic impulse. Unfortunately the accuracy of the measured values of that decrease were rather low.

A theoretic model of the complete electrical process was designed. In accordance with this model the effective resistance undergoes a slow decrease during the first stage, a instantaneous decrease in the second stage and a progressive increase up to reach the initial value in the third stage. That model has proved very efficient for it produces figures for \(v(t)\) and \(i(t)\) nearly coincident with the experimental results of the actual cases analysed. The recursive algorithm used is

\[
v(t_s) = v(t_{s-1}) \cdot \exp\left\{ -\frac{\Delta t}{R(t_s) \cdot C} \right\}
\]

\[
\Delta t = t_s - t_{s-1} \leq 10 \mu s
\]

\[
R(t_s) = \begin{cases} 
  cte & t_s \leq \text{ionization time} \\
  \approx \arctan(t) & (\text{matching extreme values})
\end{cases}
\]

The intensity results simply \(i(t_s) = \frac{v(t_s)}{R(t_s)}\)

**ACOUSTICAL IMPULSES**

The electric arc causes a cessation of calorific energy to the surrounding fluid and it is the responsible for the acoustic detonation /2/. Once the calorific energy is transferred to the fluid the elastic and thermal coefficients control the acoustic wave.

The acoustic waveform produced by this process is a N wave as it is shown in Figure 1 (the levels are in arbitrary units). The acoustic wave properties (peak level and spectrum) have been analysed with regard to the electrical characteristics and to the shape and size of the graphite resistor and support of the electrodes. The acoustic impulses were produced inside an anechoic room. The peak levels were measured on an oscilloscope display and the spectra by using a FFT computing spectrum analyzer.
We have observed the peak levels to grow with the charged voltage and with the capacitance. The influence of the charged voltages on peak levels exceeds the influence of the capacitance, mainly for values over 400V. Under all circumstances the peak levels reach the maximum values asintotically as the charged voltage or the capacitance increases.

The spectrum shape depends upon the charged voltage and the capacitance only when the size of the electrodes and support are reduced enough as for avoiding significant reflections (case of the electrode-support system type 2, from Figure 2). Supports of some 3 cm in size (electrode-support system type 1, Figure 2) mask the influence of the above mentioned electrical characteristics.

Figure 3 compares eleven spectra corresponding to several voltages and capacitances for the electrode-support syste m type 1. We can observe in that figure, a spectrum rise with frequency at a rate of 5.3 dB/octave when the levels are measured in frequency bands of constant lineal width (8.3 dB/octave for levels measured in bands of constant logarithmic width). Similar results have been measured by other authors for gun shots /3/. The maximum of the spectrum is reached at about 8 kHz., indicating the adequacy as test signal on model studies with a scale factor of 10.

The length of the graphite piece has a secondary effect on the peak levels and on the spectra. That facts supports the hypothesis of Weber /2/, according which the process of energy transfert to the surrounding fluid corresponds to a thermal expansion wave, the equivalent sphere radius depending mainly of the total energy transferred.

An interesting question is how to translate the sound spectrum in the frequency domain, mainly towards the low frequencies, in order to facilitate the use of other scale factors in experimental model studies. With this purpose in mind we have analysed the possibilities offered by the superposition of several impulses of that nature.

In a first stage we studied the superposition of two sound impulses produced by two different sources of that type when both pairs of electrodes were placed near enough one from the other in order to ensure neglecting variations of the resulting signal in different directions. Technical and economical problems increase enormously if more than two impulses are to be superposed. On Figure 4, two experimental sound impulses ans the resulting experimental superposition are shown. A time delay of 50 µs between both impulses was got by controlling the duration of the ionization stage by means of appropriate values of the charged voltage, capacitance and appropriate nature of the graphite piece /4/. That choice makes possible time delays between both impulses up to 70 µs, well adapted to that particular impulses.
That particular superposition translates the frequency maximum towards the low frequencies (=6 kHz) and changes the slope of the spectrum to about 3 dB/octave (levels in bands of constant lineal width).

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MESURE DE L'ABSORPTION ET DE L'ISOLEMENT ACOUSTIQUE AVEC UNE SOURCE DE REFERENCE

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Les mesures de l'absorption en salle réverbérante ou de l'isolement acoustique exigent généralement la production d'un son diffus dans le local où se trouve la source, de manière à provoquer soit une décroissance du son aussi régulièrement possible à toutes fréquences, soit une intensité du son réverbéré aussi constante que possible dans une grande partie du local où sont placés les microphones. Les dispositions adoptées pour produire cette diffusion sont très souvent l'installation de panneaux orientés de manière diverse dans le local. Mais l'introduction de ces éléments modifie les conditions de propagation du son dans le local et notamment le libre parcours moyen du son, si bien que les formules liant la durée de réverbération ou l'intensité du son réverbéré à l'absorption du local doivent être modifiées pour tenir compte de ce fait. Mais cette modification ne peut être précisée de manière facile. Nous avons cherché à produire la diffusion d'une autre manière pour éviter l'emploi de panneaux diffusants qui, de toute façon, ne peuvent être utilisés de manière pratique dans les mesures sur chantier.

L'idée retenue est celle de l'emploi d'une source de faible encombrement, néanmoins puissante, omnidirectionnelle aux fréquences d'essai et qui puisse se déplacer pendant qu'elle émet.

1.- LES CARACTERISTIQUES DE LA SOURCE DE REFERENCE

L'enceinte acoustique de forme sphérique (diamètre 30 cm) contient 4 haut-parleurs élémentaire. Son niveau de puissance est de 110 dB. Elle peut être placée - soit en position fixe, sur un pied constitué de tubes vissés directement sur le boîtier électronique servant de base au pied,
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- soit sur un dispositif tournant donnant à la source un déplacement complexe qui la met dans des dispositions diverses dans le local.

Afin de pouvoir faire servir la source à des mesures simplifiées de l'isolement acoustique servant au contrôle réglementaire de l'isolation, son électronique a été élaborée pour qu'elle puisse produire des signaux avec 4 types de spectre :

1) un bruit rose en champ libre
2) un bruit type routier
3) un bruit qui sert à la mesure globale du facteur de correction 10 lg (2T) d'un isolement normalisé
4) un bruit permettant d'obtenir, en moyenne, un spectre de bruit rose dans le champ réverbéré des pièces d'habitation.

Cette source se prête à des mesures acoustiques diverses. Nous examinerons ici son emploi dans 2 méthodes de mesure simple de l'absorption, puis de l'isolement acoustique.

2 - MESURE DE L'ABSORPTION ACOUSTIQUE

La source peut tout d'abord servir dans l'application de la norme ISO 354 pour la mesure en salle réverbérante, mais sans l'emploi des panneaux diffusants. En utilisant 2 salles réverbérantes très différentes de volume (200 et 50 m²) et de forme on montre l'utilité de l'emploi de la relation

\[ T = 6 l_{m} / c \lg(1 - \alpha) \]  (1)

où \( T \) est la durée de réverbération

- \( l_{m} \) le libre parcours moyen
- \( c \) la célérité du son
- \( \alpha \) le coefficient d'absorption moyen dans la salle.

\( l_{m} \) ne peut être pris égal à 4 V/S ; il doit être calculé en fonction du rapport des dimensions. Le même matériau absorbant placé successivement dans les 2 salles de formes différentes a la même absorption si on prend cette précaution. Si on se contente de faire \( l_{m} = 4 \) V/S, il n'y a plus égalité des absorptions dans les 2 cas. Nous avons déjà indiqué dans des articles passés les différentes méthodes d'évaluation du libre parcours moyen [1].

La deuxième méthode de mesure de l'absorption est faite avec la source étalonnée. En s'appuyant sur la relation

\[ I_{R} = W(1 - \alpha) / \pi l_{me}^{2} \alpha \]  (2)
on peut déduire le coefficient d'absorption moyen $\bar{\alpha}$ du rapport $W/I_R$ et de $I_{\text{me}}$

$W$ est la puissance acoustique de la source donnée par un éta-

lonnage

$I_R$ est l'intensité du son réverbéré

$I_{\text{me}}$ est le libre parcours moyen énergétique [1]

Il faut aussi tenir compte de la concentration d'énergie sur

les parois si bien que la relation (2) est ainsi modifiée

$$\frac{W}{I_R} = n I_{\text{me}} \bar{\alpha} \frac{(4 + 5 \lambda a V)}{(4 - \bar{\alpha})}$$  

(3)

$S$ est la surface totale des parois

$V$ est le volume de la salle

$\lambda$ est la longueur d'onde du son émis par la source correspon-
dant à la fréquence médiane pour une bande de fréquences.

Le dispositif de déplacement de la source fait aussi tourner
en même temps le microphone de telle sorte que la distance de
la source à ce dernier reste constante. On peut ainsi calculer
l'intensité du son direct et la retrancher de l'intensité tota-
le pour connaître exactement l'intensité du son réverbéré.

Si on fait toujours l'essai avec une source soigneusement éta-
lonnée et de puissance bien stable on peut faire correspondre
directement le coefficient d'absorption moyen $\bar{\alpha}$ au niveau de
l'intensité du son. Le coefficient cherché est alors :

$$\alpha_a = (\bar{\alpha} - \alpha_o) S / S_a + \alpha_o$$  

(4)

$S_1$ surface du matériau en essai

$\alpha_o$ coefficient d'absorption de la salle nue qui peut être dé-
terminée une fois pour toutes si on prend soin de faire la
mesure dans la salle nue toujours dans le même état et à la mê-
me température.

Les valeurs trouvées dans les 2 méthodes sont très voisines. La
deuxième est plus rapide puisqu'il suffit de lire le niveau de
pression acoustique sur un cycle complet de rotation du dispo-
sitif tournant pour obtenir directement le coefficient d'absor-
tion moyen $\bar{\alpha}$. Mais sa précision est moins grande que celle
de la première méthode

3 - MESURE DE L'ISOLEMENT ACOUSTIQUE

La recherche d'une méthode de mesure très simplifiée de l'iso-
lement acoustique nous a conduit à faire servir la source de
référence à la fois pour la mesure de l'isolement brut et pour
celle du facteur de correction. Pour simplifier beaucoup la
méthode, il faut éviter de faire une analyse par bande d'octave
mais il est alors impossible de mesurer d'une façon correcte la
durée de réverbération globale dans le local de réception. Nous
l'avons démontré dans un article passé [2]. Il est donc néces-
saire de mesurer l'absorption dans le local de réception comme
nous l'avons fait précédemment en employant la 2ème méthode dé-
crite au chapitre précédent.
Dans une première phase de l'opération, on mesure l'isolation brut entre les 2 locaux. À la suite d'une étude des durées de réverbération dans les locaux d'habitation vides, nous avons pu déterminer la forme à donner au spectre de puissance de la source pour obtenir un bruit à peu près rose dans le champ réverbéré du local d'émission, entre les fréquences 100 et 5000 Hz des bandes 1/3 octave extrêmes. Dans une deuxième phase des essais la source est transportée dans le local de réception et l'on mesure le niveau de pression dans le champ réverbéré de ce local. Le niveau de puissance est repéré simultanément à l'aide d'un microphone étalonné placé en position fixe au voisinage immédiat de la source. Des relations (1) et (2) on peut déduire le facteur de correction 10 log (T/0,5) entrant dans l'évaluation de l'isolation normalisée entre 2 locaux.

\[ 10 \log (2T) = L_{PA} - L_{WA} + 10 \log \left( \frac{n}{n_{me}} \cdot \frac{\alpha}{\alpha} \cdot \frac{L_m}{c} \right) \]  

Nous avons supposé pour établir cette relation que le facteur d'absorption moyen était suffisamment simple pour écrire les relations (1) et (2)  

\[ I = 15,3 L_m / c \alpha \quad I_f = W / n_{me} \alpha \]  

S'il ne l'était pas il faudrait évidemment faire le calcul exact.

Pour les pièces d'habitation courantes dont les rapports de dimension ne sont pas trop grands, on peut confondre les 2 types de libre parcours moyen et écrire la relation (5) d'une manière encore plus simple

\[ 10 \log (2T) = L_{PA} - L_{WA} + 30 \log L_m - G \]  

Nous avons montré dans l'article de référence [2] que les valeurs trouvées avec la relation (5) ou (6) étaient très voisines de celles qui pouvaient être mesurées en appliquant la méthode de la norme ISO 140, alors que si on se servait des relations classiques  

\[ T = 0,16 V / S \]  

et  

\[ I = \frac{W(V - 1)}{S} \]  

conduisant à un facteur de correction 10 log (T/0,5) = L_{PA} - L_{WA} + 10 \log V - 11  

on trouvait un écart systématique de 1 dB entre les 2 méthodes. Quand la mesure des niveaux pondérés A, L_{PA} et L_{WA} est faite globalement il faut avoir soin d'insérer dans la chaîne de mesure un filtre passe bande de 88 à 5656 Hz.

La mesure est simple et rapide. Son degré de précision dépend de la forme de la courbe d'isolation. Car le spectre de fréquence utilisé pour la mesure du facteur de correction est celui d'un bruit rose ayant traversé une paroi dont la pente de la courbe d'isolation est de 5 dB par octave.

4 - CONCLUSION - La source de référence mobile se révèle un outil très intéressant pour la création d'un champ diffus. Sa facilité de transport la rend propres à des essais acoustiques aussi bien en laboratoire que sur chantier.


MAN-COMPUTER DIFFERENCES IN REVERBERATION TIME MEASUREMENT

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Introduction

The reverberation time measurement represents an important measuring procedure in the room acoustics. Initiated by the level recorder application, it is nowadays performed by the computer use, too.

In literature the values of the reverberation time, measured by both methods, i.e. by the application of level recorder and of the computer are found. That is why the question whether there is the difference in the values of reverberation time obtained on the basis of these two procedures gets in significance.

In the first method, with the level recorder application, the determination of the decay curve slope is made by a naked eye, which means by the estimation of the man performing the measurements. In the numerical method this slope determination is carried out by an approximation to the straight line, by the method of the least squares. The differences in values of reverberation time, which might appear, are the consequences of the differences in methods of approximations of the decay curve slope, i.e. the difference between the man’s eye and the computer algorithm.

This paper presents the results of comparison of these two methods by comparing the values of reverberation time of a large number of decay curves.

The Method

In order to compare these two methods about 300 various decay curves were compared. The signals picked up during the sound decay in rooms were recorded by a tape recorder. The recording was carried out in different acoustic conditions, from small dwelling rooms to concert halls. The reverberation time range of these premises was from about 0.5 to 5 sec.
On the basis of the signals from the tape recorder, the records were made on the level recorder. The paper speed for all records was 30 mm/sec, and the range 50 dB. These records were xeroxed and distributed to ten acousticians having the previous experience with such reverberation time measurements. Every one of them made the determination of the reverberation time for all decay curves. The reverberation times measured "by man" were obtained for each decay curve by averaging the values obtained for these ten acousticians.

The same tape recorded signals were used for determining the reverberation times by means of a computer. The signal was discretized by 12 bit A/D converter. The determination of the decay curve slope was made by fitting a straight line using the method of least squares.

Both methods of the slope determination were carried out in intervals from -5 to -35 dB with respect to the maximum level. The particular attention was taken that these records do not contain the decay curves with double slopes.

On the basis of so obtained values of reverberation time, the comparison of measuring methods "by man" and "by computer" was made.

Results

For every decay curve the difference in percentage was formed:

$$\Delta T = \frac{T_m - T_c}{T_m} \times 100 \ [\%]$$

Here: $T_m$ is the reverberation time obtained "by man".

$T_c$ is the reverberation time obtained "by computer".

The averaging with respect to values $T_m$ was performed and this curve is illustrated in Fig. 1. The standard deviation showing the scattering of results with respect to the mean value is also drawn in. It is noticed that $\Delta T$ is negative, which means that in average, the reverberation time values obtained by computer are somewhat higher than those obtained by the level recorder. Naturally, the magnitude of the standard deviation indicates that the scattering of singular results with respect to the mean value is big. The cases were observed where the differences man-computer were over 20%.

For every decay curve the quantity of fluctuations at the sound decay was determined as a mean value of the deviation of the decay curve from the straight line set by computer. The dependences of magnitude $\Delta T$, according to (1), in the function of these fluctuations is shown in Fig. 2. This figure shows also the standard deviation as a measure of the scattering of particular results.
It is noticed that the differences are higher if the fluctuations are higher. This is the anticipated result because it is clear that it is difficult to determine the decay curve slope by a naked eye in the presence of higher fluctuations.

Conclusion

On the basis of the presented results, it can be seen that between the values of reverberation time measured by both the level recorder and computer, there is a certain difference. In some cases the difference may reach 20%. In average, somewhat higher values are obtained by computer, which is seen in Fig. 1. The deviation mean value reaches about 5% for lower values of reverberation time. With the increase of the reverberation time, deviations are smaller. The standard deviation drawn in Fig. 1 shows that from case to case considerable differences appear with respect to the mean value. Also on Fig. 2 it was shown that the differences between man and computer are higher if the fluctuations of levels during the decay are higher.

The differences are higher for small values of the reverberation time. It is obvious that in these cases the determination of the decay curve slope by a naked eye is more subject to error. It should be mentioned that the use of smaller number of bits at conversion would introduce a greater error into the result calculated by computer /1/.

The phenomenon that by means of a computer somewhat higher values of the reverberation time are obtained can be explained by the fact that the computer strictly respects the boundaries -5 to -35dB, which often can include the beginning of the curvature of the decay curve where it occurs /2/. The man is, probably, biased to neglect this phenomenon in the lower part of the decay curve at the appreciation of the slopes. Finally, the results presented indicate that when explaining the results of measuring the reverberation time the differences obtained should be taken care of.

References

MEASURING THE REVERBERATION TIME WITH MUSIC.

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1. Introduction
The different methods for measuring the reverberation time involve normalized signals with well defined envelopes: a step function or a pulse. They cannot be extended to natural signals such as music. This remark led Houtgast and Steeneken /1/ to consider other types of quadratic envelopes for their test signals. They described the transmission of the modulation of the envelopes with the Modulation Transfer Function (MTF). The use of the MTF allowed them to successfully predict, speech intelligibility in a room. Besides, the relation between the complex MTF (CMTF) and the impulse response of a linear transmission system /2/ gives a new method for calculating the reverberation time, and can be adapted to any signal. The experimental validity of this method is here investigated.

2. Procedures
Let first consider the transmission of a modulated noise signal s(t), n(t) by a passive linear system h(t). Schroeder /2/ derived a relation between the quadratic expectations of test s(t), n(t) and received r(t) signals:

\[ \langle r^2(t) \rangle = \int h^2(t-u) \langle (s(u), n(u))^2 \rangle \, du \]  

(1)
The expectations are taken on the set of all noise signals n(t) of unity power, undergoing the modulation s(t). For a periodic modulation with \( s(t) = s(t+T) \), this set of noises can be advantageously replaced by the set of all the delayed versions of each signal for which the delay is a multiple of the period T. For example, the quadratic expectation of the received signal will be defined as:

\[ \langle r^2(t) \rangle \approx \frac{1}{N} \sum_{k=1}^{N} r^2(t+kT) \]

The resolution of equation (1) in the frequency domain gives the CMTF for frequencies \( f_k = k/T \). Besides, the CMTF is interesting at low modulation frequencies only, so the signal \( r^2(t) \) can be low pass filtered to reduce the computation time of the CMTF.

In accordance with equation (1), Schroeder /2/ proved that the CMTF is the Fourier transform of the quadratic impulse response \( h^2(t) \) of a passive linear system. In this paper, this CMTF will be referred to as the the "theoretical CMTF".
Let now consider any test signal \( s(t) \). A set of events for taking expectations exists no longer. The concept of "Modulation" must be introduced as low pass filtered quadratic value, e.g.:

\[
\overline{r^2}(t) = \int r^2(u) \, g(t-u) \, du
\]

(2)

where \( g(t) \) is a positive weighting function with appropriate low pass characteristics. The calculation of the CMTF must be generalized by the filtering method: the "empirical CMTF" is defined as the transfer function of the linear filter which predicts \( \overline{r^2}(t) \) out of \( \overline{s^2}(t) \) with the least square error. This calculation uses the cross correlation function of both modulations:

\[
\phi_{\overline{r^2}, \overline{s^2}}(\tau) = \int \int h(u) \, h(v) \, s^2(t-u) \, \overline{s^2}(t-\tau) \, \phi_{s}(u-v)_{t-u} \, du \, dv
\]

where:

\[
\phi_{s}(\tau) = \int s(u) \, s(u+\tau) \, g(t-u) \, du / s^2(t)
\]

is the normalized short-time correlation function of the test signal \( s(t) \). If the temporal mean values of \( \phi_{s}(\tau) \) vanish for \( \tau = 0 \), it remains:

\[
\phi_{\overline{r^2}, \overline{s^2}}(\tau) = \int h^2(u) \, \phi_{\overline{s^2}}(t-u) \, du
\]

(3)

In this case, empirical and theoretical CMTF coincide. They can be calculated from the Fourier transforms of the correlation functions. Beside the hypothesis on \( \phi_{s}(\tau) \), the stationarity for the modulation \( s^2(t) \) is required in equation (3). This property is not satisfied by a music signal in general. For the sake of calculating the CMTF, the local stationarity of the modulations was assumed. This introduces into the correlations functions a compensation for the louder and weaker parts of the music, that reduces the variations of dynamics.

The calculation of the CMTF involves two steps. The first one is devoted to the modulations. Each signal is sent through an octave band pass filter (700-1400 Hz) into an anologic/digital converter. This digital signal is squared and low pass filtered. The resulting modulation is then resampled at 100 Hz rate. The second step calculate the CMTF itself. For each pair of test and received signal, their power and crosspower spectra are calculated, using FFT of about 10 sec signal duration and weighted averages over 80 overlapping segments. The empirical CMTF is determined by division of the power spectra. About 240 sec signal duration is processed.

3. Experimental results
Fig. 1 shows modules and phases of the CMTF for a model calculation considering a reverberation filter /3/. The calculations involve a music signal free of reverberation /4/ and a noise signal with the same envelope. A comparison of theoretical and empirical CMTF shows a good agreement, especially at low modulation frequencies.

In the lecture hall, for two transmission paths (loudspeaker-directional microphone, about 5m, and loudspeaker-omnidirectional microphone, about 9m) recordings were made with a pulse, a noise signal with periodic pseudorandom envelope, and the two signals mentioned above. The theoretical MTF for the two paths are shown in fig.2. In good agreement are the results from the pseudorandom modulated noise (crosses in fig. 2). For the directional microphone, the importance of the direct sound appears in the horizontal
Fig. 1: Modules and phases of CMIF for reverberation filter; theoretical CMIF (left), empirical CMIF from music modulated noise (center) and from music (right).

Fig. 2: Theoretical MTF for directional (left) and omnidirectional (right) microphones. Crosses: empirical CMIF from pseudorandom modulated noise.

profile of the MTF above 3Hz. For the omnidirectional microphone, the decreasing curve is typical of reverberation. Fig. 3 shows the results of an excitation with the music signal. The empirical MTF were calculated for the signal pairs loudspeaker/directional microphone, loudspeaker/omnidirectional microphone and directional/omnidirectional microphone. These MTF agree with the theoretical one, especially at low modulation frequencies. Big systematic deviations appear at frequencies where the modulation power spectra show sharp maxima that correspond to the rhythmic structure of the music. From the initial parts of the MTF, for the signal pairs loudspeaker/omnidirectional microphone and directional/omnidirectional microphone, accurate estimations of the reverberation time can be obtained. This agreement of the estimations is explained by the very similar aspects of all modulation power spectra.
Fig. 3: Empirical MTF from music for loudspeaker/directional microphone (above left), loudspeaker/omnidirectional microphone (above right) and directional/omnidirectional microphone (beneath).

Fig. 4 compares the estimated decay curve for the signal pair directional omnidirectional microphone, with the reverberation curve for the omnidirectional microphone. It can be seen that the estimated reverberation time agrees with the early decay time (EDT).

4. Conclusion
The experiments point out the general validity of equation (3) and the relation between the CMTF and the quadratic impulse response of a passive linear system, although the linear filtering poorly predicts the transmission of the modulation.

Fig. 4: Reverberation curve of omnidirectional microphone. Dashed line: estimated decay curve from music for signal pair: directional/omnidirectional microphone.


ACOUSTICAL MEASUREMENTS WITH DYNAMICAL SIGNALS
- THEORETICAL CONSIDERATION (Part 1)

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ABSTRACT

The measurement of vibration time, transmission loss and other parameters can more efficiently, more precisely and with less acoustical power be done with dynamical signals. By using narrow frequency bands of noise, which are amplitude modulated by square-waves, the measurement can be carried out with enhanced accuracy, if long time averaging takes place. Taking into account time differences in transmitter and receiver facilities, the theoretical wave form of the energy flow provides information as to the acoustical parameters which are to be tested.

1. INTRODUCTION

It is common practice to measure acoustical parameters such as transmission loss, body-borne sound and reverberation time by stationary or quasi-stationary noises:

![Diagram](image)

Figure 1

T  transmitting phase
R  receiving phase
\( \tau \)  phase duration time
Reverberation Time

For measuring reverberation times of large volumina, very high acoustical power is needed which might in the case of industry halls be many hundred Watts.

Although the measurement by itself is very simple (measuring the time of decay of a sound) it is very troublesome because of the high power needed of the instruments which should be transportable.

2. MEASURING REVERBERATION TIMES WITH DYNAMICAL SIGNALS

The measurement of reverberation time can be done in real time in the following manner: Let us transmit a 1,000 Hz Terz frequency band noise and interrupt it in a rhythm of 1 second period according to figure 1. Half of the time $t$ the sound will be transmitted (0,5 seconds) and half of the time there will be no transmitting, so-called receiving time.

If one integrates the sound power in the transmitting phase (and gets the so-called transmitting energy) and compares it to the integrated sound power in the reception phase (so-called reception energy) one gets a relation which depends on the reverberation time. In the reception time the sound power decays according to formula 1, while

1. is the sound power
2. $I_0$ is the initial sound power
3. $T$ is the reverberation time in seconds.

$$ I = I_0 \cdot \exp(-at) \quad (1) $$

The integrated reception energy is according to formula 2. $t$ is the integrating time, which is equal to the length of the reception phase and to the length of the transmission phase.

$$ \Delta E = E_R - t \cdot I_0 \cdot \exp(-at) \quad (2) $$

The reception energy can be calculated according to formula 2a:

$$ \Delta E = (I_0/a) \cdot [1 - \exp(-at)] - t \cdot I_0 \cdot \exp(-at) \quad (2a) $$

In the transmission phase the energy can be shown according to formula 3:

$$ E'_T = I_0 \cdot t - \Delta E \quad (3) $$

- $E'_T$ is the transmitting energy corrected for background sound energy
- $E_R$ is the reception energy corrected for background sound energy

$$ E'_R = E_R + E_0 \quad (4) $$

$$ E'_T = E_T + E_0 \quad (5) $$

$E'_R$, $E'_T$ are measured energies including background noise

$E_0$ is the background energy

$$ E_0 = I_{BG} \cdot t $$
The reception energy can be taken from formula 1a:

\[ E_R = \left( I_0 / a \right) + \left( 1 - \exp(-\alpha t) \right) \]

(1a)

The difference between the reception energy and the energy gained by multiplying the minimum sound pressure level by the integrating time \( t \) can be taken from the formulas 2 - 2 b:

\[ \Delta E = E_R - t \cdot I_0 \exp(-\alpha t) \]

(2)

\[ \Delta E = \left( I_0 / a \right) \cdot \left[ 1 - (1 + \tau) \cdot \exp(-\alpha t) \right] \]

(2b)

The difference \( \Delta E \) is equal to the difference between the transmitting energy and the multiplication of maximum pressure level and the integration time \( \tau \). Thus the transmitting energy can be found according to formula 3:

\[ E_T = I_0 \cdot \tau - \Delta E \]

(3)

\[ E_T = \left( I_0 / a \right) \cdot [\alpha t - 1 + (1+at) \cdot \exp(-at)] \]

(3a)

This enables us to find the relation between the quotient describing the reverberation time in terms of relation of transmission and receiving energy, see formula 6:

\[ \frac{E_R}{E_T} = \frac{[1 - \exp(-at)]}{[\alpha t - 1 + (1+at) \cdot \exp(-at)]} \]

(6)

Expanding the term \( \exp(-at) \) in formula 7

\[ \exp(-at) = 1 - at + (at)^2 / 2 - (at)^3 / 6 + (at)^4 / 24 \]

(7)

This enables us to make approximations for small \( a t \), which leads to the practical formulas for the reverberation time, No. 8 and 9:

\[ \frac{E_R}{E_T} \sim 1 - (at)^2 / 6 \]

(8)

\[ a = 13.8 / T \]

\[ \tau = 1/2 \cdot f_m \]

\( T \) is reverberation time

\( f_m \) is the modulation frequency of the test signal

\[ T = 2.82 / \left( f_m \cdot \sqrt{1 - \frac{E_R}{E_T}} \right) \]

(9)
These relations are only valid when two main conditions can be held:

a. The measurement is done out of the reverberation radius of the loudspeaker and compensated for the propagation time of the sound from loudspeaker to microphone.

b. The reverberation time of the sound producing loudspeaker can be neglected. (This phenomenon can in many times not be neglected, see MANTEL (1982)).

3. LITERATURE

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